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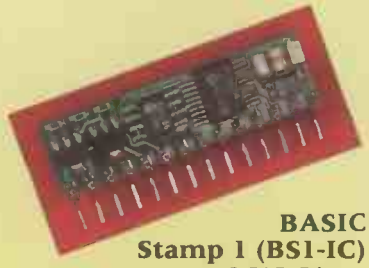

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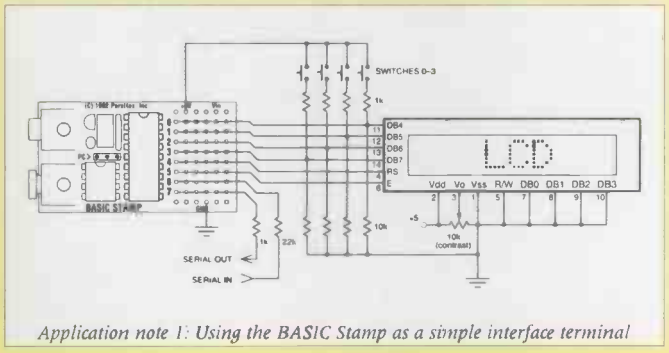

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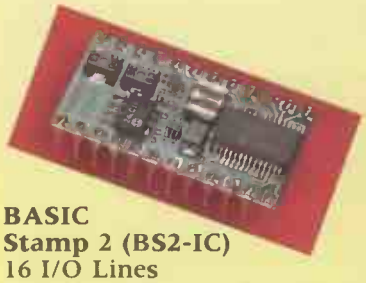

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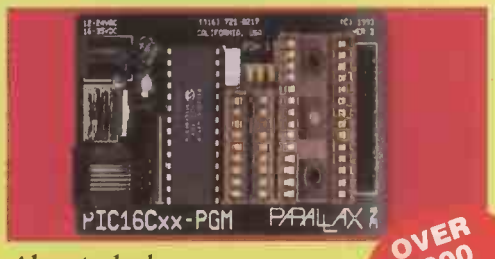

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Contents

Cover – Hashim Akib

830 MAGNETIC SENSING APPLICATIONS
Inclination detection, vehicle counting and a jam jar for hams – three applications for a three-terminal magnetic sensor.

836 NO-CONTACT CURRENT MEASUREMENT
Steve Winder shows how measuring dc via a magnetic sensor gets round the problems associated with shunts.

842 GAIN FROM PASSIVE
Ian Hickman investigates how gain can be obtained in unusual ways – using only passive components.

847 THE ROUTE TO PCB CAD
Thinking of using cad for your pcbs? Rod Cooper discusses autorouters and reviews Circuitmaker and Traxmaker.

853 FEEDING THE OFF-CENTRE DIPOLE
Richard Formato challenges traditional thinking about where the off-centre fed dipole should be tapped.

858 NIGHT THOUGHTS ON CROSSOVER DISTORTION
When designing a high-performance Class-B audio power amplifier, removing crossover distortion is one of the most problematic areas. Doug Self shares his experience.

864 HANDS-ON INTERNET
Cyril Bateman looks at circuit-design software on the net – including packages specially for rf engineering.

872 Interfacing with C – new book, and special offer for listings disk

873 AUDIO PROCESSING ON THE PC
Howard Hutchings discusses the C code needed to process audio on the PC, for effects such as flanging and chorus.

892 PRECISE FREQUENCY GENERATION
Nick Wheeler shows how to derive almost any frequency with quartz precision.

897 HYBRID POWER AMPLIFIER
Valves, transistors and ICs combine in Wim van der Jager's hybrid power amplifier, capable of delivering up to 40W.

Regulars

819 COMMENT
Media megalomania

820 NEWS
1000 times more data on a cd? Missed opportunity, Superconductor breakthrough.

823 RESEARCH NOTES
Chips that will not heat, World's smallest wire, 3D display, Safer driving

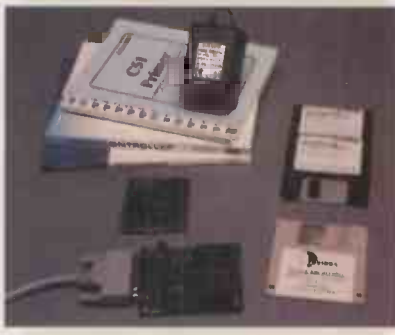
868 LETTERS
Do-it-yourself pth and an answer to the motional feedback headphone problem.

881 CIRCUIT IDEAS
Electric whisker, Deglitcher for smps, Very simple charger for NiCds, Voltage division.

887 NEW PRODUCTS
Pick of the month – classified for convenience.

EW reader discount

This 10Hz to 1MHz audio signal generator is available to *Electronics World* readers at 20% discount – page 855.



Software *and* hardware for developing Atmel Flash 8051microcontrollers in C or assemler – available exclusively to *Electronics World* readers at the special price of £184.48, pager 828.

3D all round: with this three dimensional display cube, a coloured image can float in all three axes – and be viewed from any angle – page 825.

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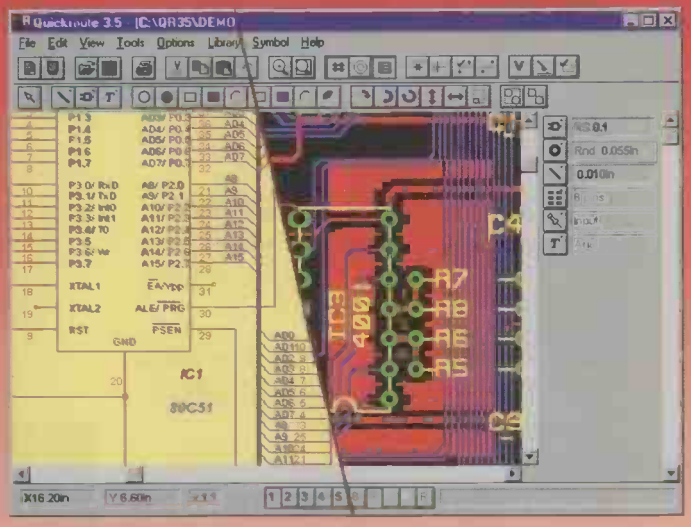

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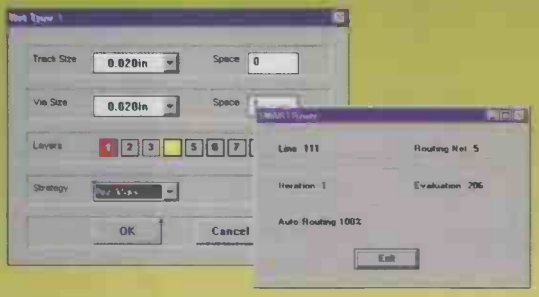

SMARTRoute 1.0 is a new 32 bit auto-router that offers amazing flexibility & power at an affordable price! Compatible with Windows 3.1/95/NT, SMARTRoute gives you total control over routing strategies including layers used, track & via sizes, design rules, etc.

SMARTRoute is completely compatible with Quickroute 3.5 and offers improved completion rates compared with Quickroute' s built In auto-router (ask for details) SMARTRoute is available for £149 plus P&P and V.A.T. Special bundle pricing for Quickroute and SMARTRoute when purchased together.

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REED BUSINESS PUBLISHING

Media megalomania

Sometimes one wonders what drives industrial moguls. Some build business empires which destroy their families – like Aristotle Onassis; some build companies which get taken over soon after they leave – like Charlie Forte; some pursue a vision of a better life – like Walt Disney; some want to pioneer technology – like the founders of Intel; but in the media world the usual driving force behind the industry's moguls is the pursuit of influence.

From William Randolph Hearst to Lord Northcliffe the motivation of media tycoons has been to amass and exercise power. Wealth came to them as a by-product. In Northcliffe's case the pursuit of power led to raving megolamania and a madman's death.

It was not just political power which Northcliffe relished. It was any power. He was said to enjoy appointing two people to the same job just for the pleasure of seeing them fight it out to see who survived.

But it was political power that caused the most concern. Even then. And in those days, because of the limitations of the technology, the voters which a media tycoon could hope to influence would be limited to one country.

Nowadays technology means that a global media tycoon can influence the voters of any country on the planet. And the advent of seamless digital technology in all the main information delivery vehicles - terrestial broadcasting, satellite broadcasting, cable, wireless telecommunications/datacommunications - means that a tycoon can access the global citizenry not just on a mass basis but, literally, on an individual basis.

Take, for instance, the apparently innocuous declared intention of BSkyB to provide Internet access. This is from a parent company – News Corp – which has terrestial and satellite broadcasting facilities, linked to cable and wireless communications interests, backed by film and news creation capabilities.

Imagine the power of all that combined with Internet access! The company will be able to monitor what you view, who you e-mail, what web sites you visit, what sort of information you access. With a bit of experience of you, the provider will be able to suggest new services, products, or ideas which you might accept.

"Nowadays technology means that a global media tycoon can influence the voters of any country on the planet."

And what that information worth to a government seeking to monitor its citizens or to influence them in the run-up to elections? One would think it worth a lot. And one would think that media tycoons could ask a lot for it. Not in money perhaps, but in concessions allowing further extensions of their power. Internet access via a big media company should carry a public health warning!

Potentially more dangerous than the open pipeline into your wallet, is the pipe into your head, particularly when it comes from a company that also creates entertainment and news. For how can entertainment and news be separated in one company? If the entertainment arm of the company makes a film about, say, aliens, and the news-dissemination arm of the company puts out stories about aliens to coincide with the launch of the film, no one is the wiser but many are attracted to see the film on the grounds of its supposed topicality.

More dangerously, if a film is made which has a recognisable characterisation of a politician seeking re-election, and then news stories are printed which support that characterisation – many will be influenced. The ability to infect and sway the zeitgeist will be immense.

Governments in the Western World have done a pretty lousy job of separating the ownership of the various media-types. It may already be too late for them to grasp back the power to do so by breaking up the big media companies. But with the digitisation of all media – and the power which that adds to media tycoons – we should all be aware and, more particularly, wary, of the megalomaniac in the business suit. ■

David Manners

Electronics World is published monthly. By post, current issue £2.35, back issues (if available £2.50. Orders, payments and general correspondence to **L333, Electronics World, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS**. Tlx:892984 REED BP G. Cheques should be made payable to Reed Business Publishing Group
Newstrade: Distributed by Marketforce (UK) Ltd, 247 Tottenham Court Road London W1P OAU 0171 261-5108.
Subscriptions: Quadrant Subscription Services, Oakfield House Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 01444 445566. Please notify change of address. Subscription rates 1 year £30 UK 2 years £48.00 3 years £70.00. Surface mail 1 year £35.00 2 years £56 00 3 years £80.00 Air mail Europe/Eu 1 year £43.00 2 years £68.00 ROW 1 year £52.00 2 years £83.00

Overseas advertising agents: France and Belgium: Pierre Mussard, 18-20 Place de la Madeleine, Paris 75008. United States of America: Ray Barnes, Reed Business Publishing Ltd, 475 Park Avenue South, 2nd Fl New York, NY 10016 Tel; (212) 679 8888 Fax; (212) 679 9455
USA mailing agents: Mercury Airfreight International Ltd Inc, 10(b) Englehard Ave, Avenel NJ 07001. 2nd class postage paid at Rahway NJ Postmaster. Send address changes to above.
Printed by BPCC Magazines (Carlisle) Ltd, Newtown Trading Estate Carlisle. Cumbria, CA2 7NR
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UK loses plastic led lead

The UK appears to have lost its chance to become the first country in the world to manufacture in volume light emitting polymer (LEP)-displays. This follows Cambridge Display Technology's (CDT) announcement that it has, for the time being, abandoned its plans to start volume manufacturing its pioneering LEP technology with Xyratex, a UK manufacturer. Instead, CDT has opted to license its technology with the consumer giant Philips being the first taker. "You can't bring technology to market by yourself. This agreement with Philips is purely a licensing. agreement.

We are talking to other people on aspects further than that," said Danny Chapchal, CDT's CEO. "We are talking to too many large Companies from East and West to ignore it." Philips will now undertake further development of the LEP technology and incorporate it into consumer products, although it did not specify when.

CDT is planning to make a second similar announcement with another European giant within six weeks. Agreements with Far Eastern companies will follow. "If you are serious about this technology then you have to look East. We know the number and kind of companies we want and in which geographical regions," said Chapchal. Xyratex, formerly IBM's Havant-based disc drive manufacturer, is aware of CDT's plans and is in full agreement. "These early licensing opportunities.make sense," said David Martin, technical director of Xyratex.

LEPs will provide an efficient, low-power, low-cost replacement for lcds and leds currently used m most consumer electronics products. CDT's own target is to have LEP based consumer products on the market by the decade end. Meanwhile it hopes to retain the image of a centre for excellence for this technology, and will continue to develop it and prime it for transfer.

Micromirror devices for evaluation

Digital micromirror devices (dmds) from Texas Instruments are to be offered evaluation kits for the first time.

The company says this move comes as a result of "thousands of enquiries". The digital light processing evaluation kits will include all the necessary components to interface to a pc.

However, the resolution of the kits is limited to VGA: 640 x 480 pixels. The actual dmds themselves can work up to 1280-by-1024 pixels, but these are only available in oem versions of the kits. Digital micromirros devices are not available in component form.

New magnets are attractive for chips

Lucent Technologies' Bell Labs has discovered a magnetic effect, characteristic of a group of superconducting materials, which could lead to the development of advanced chips.

Bell Labs' scientists claim to have found a way to control magnetic fields which impede or even destroy a superconducting state.

Caused by the flow of electrons within superconductive material or by electrical devices, ubiquitous magnetic fields have limited the degree of superconductivity possible thus far. Normally, when a superconductor is placed in a strong magnetic field, the magnetic field lines create electronic vortices which impede the flow of electrons in the material.

But chemists have discovered a single-crystal compound which resists external magnetic fields while retaining its superconductivity.

Called ErNi2B2C, the compound allows the scientists to find new patterns of magnetic lines which they hope to control. The patterns suggest the lines could be pinned without using expensive dopants. The result could be the advent of practical superconducting devices sooner than expected. However, the researchers warn that while practical applications are being explored, applying this discovery commercially will take time.

Aston students trial Smartcard ID

Aston University students are to trial the use of smartcards as part of the university's project to develop the technology. The project, started in August, is being funded with a £200,000 government grant. The cards will give students access to premises and services such as libraries, as well as access to the Internet. To extend the range of services available to student card owners, project manager Tony Bell said that the university was "aiming to work with banks, travel companies, and other organisations".

1000 times more data on a standard cd?

Compact disk media may experience a major advance in the amount of data they can store. Scientists from the University of Buffalo have announced a storage technique that allows 1000 times more data to be crammed onto cds.

The design puts the data in layers, similar to the pages of a book, on a disc made of new polymer-based photonic materials made using inexpensive plastic and new dyes.

To read the stacked data, the disc scans laterally across, similar to conventional cds, but when it reaches the end of a layer it refocuses the read beam onto the next layer, running across the disc again.

The new technology is called two-photon absorption where a molecule absorbs two photons of light simultaneously if the light beam has enough intensity. Since conventional plastics are only capable of weak light absorption, they are useless for such applications.

By coating plastic with new dyes, the material shows the strong two-photon absorption needed to tightly focus the laser beam.

Surf with confidence

A new book shows Internet users how to access confidential information about other people on the Internet. The items include driving records, address, license plate number, genealogical data, insurance claims history and other information which was once confidential.

Much of the information is US based, but it shows how such data is increasingly available over the Internet. The book, by Bob Villa and John LeCarre, is called "NetSpy: How You Can Access the Facts and Cover Your Tracks Using the Internet and Online Services." It shows how the Internet can be used to find people through myriad web sites or check people's credit histories. **http://www.ypn.com**

Growth in contract manufacture still keen

The worldwide market for electronics contract manufacturing is expected to equal $59.3bn in 1996, continuing the recent trend of sharp growth, according to a recent report by Californian-based Technology Forecasters.

The report, Contract Manufacturing: 1996 State-of-the-Industry, revealed that the 31 leading worldwide contract manufacturers accounted for one third of the global contract manufacturing market in 1995, and that the sum of their revenues increased 51 per cent between 1994 and 1995.

UK-based Design to Distribution (D2D) was mentioned in the report as one the few companies to more than double their contract manufacturing revenues.

Commenting on the report, Brian Haken, executive director of the UK's Printed Circuit Interconnection Federation, said: "Fewer electronics companies are making their own pcbs so we can expect this growth to continue." Haken, who describes Technology Forecasters as "one of the best sources of information available on the contract manufacturing market", explained that electronics companies are increasingly concentrating on core competencies, leaving the production of electronic assemblies to others.

An average of 63 per cent of outsourced pcb assemblies involved surface mount technology. This is estimated to rise to 82 per cent by the year 2000.

Mobile 'phones: "no evidence of risk"

There is no existing evidence that a health threat exists for millions of UK mobile phone users, according to the chairman of a body set up to initiate research into possible health effects related to mobile telephony.

Alastair McKinlay, who chairs an 'Expert Group' set up by the European Commission (EC) earlier this year, said: "The group is quite clear that there is no existing scientific evidence of a cancer risk."

But the group, which is to deliver a report to the EC at the end of this month, has identified that gaps do exist in knowledge of this area.

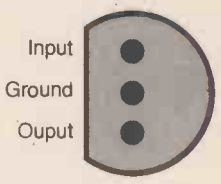
Most of the existing biological and epidemiological research that has been conducted, has been to do with power frequencies of 50Hz. "What is now required," said McKinley, "is a lot more research in the microwave regions of the electromagnetic spectrum. It is in this region, 1 to 2GHz, that mobile phones operate."

McKinley stressed that this was not because there was any concern into health effects, but because the explosion in the use of mobile phones was quite recent, and that such research makes sense to quell any public concern. ■

On this month's cover – Zetex 78L05 regulator

Zetex's ZR78L05C 5V regulator – free on this month's cover* – is a high-performance three-terminal device which is similar to the industry-standard 78L05, except that it has a quiescent current of around 350μA as opposed to 2-3mA. This makes it ideal for battery-power applications. In addition, the ZR78L05C has double the output current – at 200mA – and improved line and load regulation.

*UK readers only



Pin out of the ZR78L05C three-terminal regulator.

Absolute maximum ratings

Input voltage	**20V**
TO92 package dissipation	**0.6W**
Output current	**200mA**
Operating temperature	**–55 to 125°C**
Storage temperature	**–65 to 150°C**

Electrical characteristics

Symbol	parameter	Conditions	min.	typ.	max.	units
V_o	output voltage		4.875	5	5.125	V
		I_o=1 to 200mA T_j=55 to 125°C	4.8		5.2	V
		V_{in}=7 to 20V I_o=1 to 100mA T_j=55 to 125°C	4.8		5.2	V
ΔV_o	line regulation	V_{in}=7 to 20V		10	40	mV
ΔV_o	load regulation	I_o=1 to 200mA		5	25	mV
		I_o=1 to 100mA		2		mV
I_a	quiescent current	T_j=55 to 125°C		350	600	μA
ΔI_a	quiescent current	I_o=1 to 200mA			50	μA
	change	V_{in}=7 to 20V			100	μA
V_n	output noise voltage	f=10Hz to 10kHz		75		μV rms
$\Delta V_{in}/\Delta V_o$	ripple rejection	V_{in}=8 to 18V, f=120Hz	48	62		dB
V_{in}	input voltage required to maintain regulation		7	6.7		V
$\Delta V_o/\Delta T$	average temperature coefficient of V_o	Io=5.0mA T_j=–55 to 125°C		0.1		mV/°C

Test conditions, unless otherwise stated, T_j is 25°C, I_o is 100mA and V_{in} is 9V.

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Micro Video Cameras

Following last Month's Readers Offer for the 721-S Micro Camera many readers have contacted us asking about other items in our range of Micro Cameras and Security Surveillance equipment.

We are SOLE AUTHORISED IMPORTERS of the entire range of Cameras and Video Surveillance equipment produced by the world's leading manufacturer. ALL items in the range carry a full 12 Months Guarantee. If you would like to receive our comprehensive catalogue of Cameras and associated equipment please send a large SAE with 48p postage, marked "Camera Catalogue"

Here is a sample of the available stock.

A-721-S Micro Camera 32mm x 32mm ... £85
A-721-P Micro PIN-HOLE Camera ... 32mm x 32mm ... £85
A-921-S Camera with AUDIO ... 30mm x 30mm ... £95
A-1211 C/CS Mount Camera ... 110mm x 60mm x 60mm ... £110
A-521 Micro Cased Camera 43mm x 48mm x 58mm ... metal cased ...£120
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Auto Record Controllers ... Allow NORMAL VHS Videos to operate like professional Time Lapse or Security Recorders ... £75
QUAD-1 Multi Vision Processors ... Digital Freeze ... Quad Pictures etc £275

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Government Surplus Electronics Equipment on Special Offer This Month

TIME Electronics 404N/1021 Voltage/Current Calibrators ... 0.05% accuracy ... ONLY £275
FRANKLIN Wavetek 3600 Power Line Disturbance Monitor + Printer ... LAST 2 NOW ONLY £350
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RESEARCH NOTES

Jonathan Campbell

Staircase effect gives lift to nanodevices

Development of molecule-sized components that could be used to build more powerful computers and miniaturised electrical devices may have come a step closer thanks to work being carried out at Purdue University in the US.

The Purdue team has shown that single electrons can tunnel through a layer of ultra-small gold clusters – one to two nanometres in diameter – at room temperature, passing electrical current in a stair-step fashion that largely eliminates the problem of heat build-up found in current electrical devices.

Instead of having current go through the device continuously, the device operates by a series of discrete single electron transfers, explains Clifford Kubiak, professor of chemistry and part of the interdisciplinary team at Purdue.

The achievement marks the first time a miniature unit based on single electron tunnelling has been able to run at room temperature, and it could provide a prototype for constructing molecule-sized electronic components.

Efforts to build such small electronic components have produced active elements on computer memory chips as small as 500nm. But attempts to build smaller units have been hindered because current processes such as photolithography cannot create structures that small, and because of the heat buildup that occurs when electrical current passes through such small structures.

The structure developed at Purdue – made of gold clusters attached to a gold substrate by organic molecules – side-steps this problem by acting as a sort of "turnstile" to limit the amount of current that passes through the module, allowing just one electron at a time to cross.

The Purdue structure was designed using an approach called self-assembly, a method that allows scientists to produce a structure atom-by-atom. The scientists first produced a set of molecules shaped like a barbell with a sulphur atom on each end. When exposed to a flat gold surface, the barbells stood on end, with one sulphur atom firmly adhering to the surface and the other sulphur atom exposed. Preformed crystallites, containing 100 to 200 gold atoms, were then attached to the exposed ends.

Using a scanning tunnelling microscope the group was able to image the attached clusters and measure the relationship between current and voltage as electrons passed through the structure.

At room temperature, the current-voltage data showed the desired staircase behaviour.

Up to now the staircase effect had previously only been seen in small structures at temperatures near absolute zero. But the Purdue structure, producing the effect at room temperature, could provide a model for designing components tens to hundreds of times smaller than those currently in use today.

For more information contact: Clifford Kubiak Purdue University, West Lafayette, Indiana. email cliff@chem.purdue.edu

Dryden's digital contribution: *Nasa's Dryden Flight Research Center in Edwards, California, celebrates its fiftieth birthday. In 1946 a small number of engineers formed a group in the Mojave Desert to research the sound barrier using the X-1 research aircraft. This year, at the same site, saw the first supersonic yaw vectoring flight of the F-15 active thrust vectoring research aircraft, a technology that allows a plane to sit up in the air like a cobra rearing its head.*

In the intervening years, the Center has been the home of many steps forward in aircraft design and avionics.

For example, in 1972 the Center flew the world's first purely digital fly-by-wire aircraft, contributing to the creation of McDonnell Douglas' F-18 Hornet, General Dynamics' F-16 C/D Falcon fighters, and even aircraft such as Boeing's new 777 digital fly-by-wire airliner.

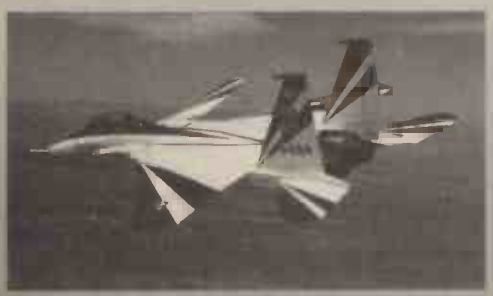
Dryden researchers also helped manufacturers explore new engine designs and integrated engine and flight control systems made possible by computer technology. The Digital Electronic Engine Control flight research project led Pratt & Whitney to commit to a digitally controlled production engine, which since then has been integrated into aircraft ranging from the McDonnell Douglas F-15 to the MD-11 and the Boeing 757.

A more advanced concept, integrating digital flight and engine controls, showed the potential of a fighter aircraft having a "self-repairing" control system, in which the aircraft would automatically use engine power to compensate for damage to an engine or flight control surface. After reading about one of several crashes resulting from the loss of flight controls because of hydraulic failures, a Dryden researcher then adapted that integrated flight control and engine concept into a potential Propulsion Controlled Aircraft (PCA) system.

A PCA system would provide a pilot with a computerised system to land an aircraft with only engine controls in the event of a catastrophic hydraulic system failure.

Although the feat was considered impossible by many engineers, Dryden nevertheless completed successful automatic PCA landings with both a McDonnell Douglas F-15 fighter in 1993 and an MD-11 airliner in 1995.

Happy birthday guys.



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Smallest wires await a connection

Creation of the world's smallest wires and encasement in a plastic polymer is being heralded as an accomplishment that could find widespread electrical and optical uses at the nanometre scale – though no-one is quite sure what at the moment.

The wires, only 6 angstroms in diameter, or just several atoms wide, could be kept separate or bunched together to make cables inside a polymer matrix, depending on the intended purpose, say the researchers at Cornell University.

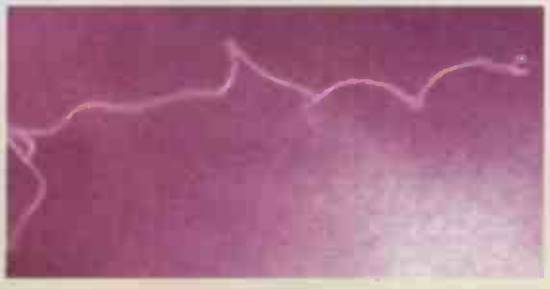
Length can be up to at least 10,000 angstroms in length. But as Francis DiSalvo, Cornell professor of chemistry who is leading the work, says: "No one has ever made wires this small before, so we're not sure what all the uses are going to be".

The wires were formed by taking atoms of molybdenum and selenium separated by lithium. By putting them in a solvent of ethylene carbonate – which polymerises into polyvinylene carbonate – the lithium was separated out, leaving long strings of the metals. An agent was then quickly added to make the polymer, so that the organic polymers gelled before the wires had a chance to clump together.

According to DiSalvo, the process can be described as: "like trapping a small, skinny sausage in a big bowl of spaghetti."

The result is a plastic block laced with sub-nanometre-sized wires. To make cables of more than one wire held together, the researchers simply have to increase the amount of metallic grains.

Now that they have shown it is possible to make such materials, the researchers are turning their attention to what they can do with them. For example, chemists are trying to use the new structures as membranes, in which the wires act as a solid-state catalyst.



A molecular wire of molybdenum selenide embedded in a polymeric matrix. The thickness of a single wire is approximately three atoms in diameter, with the length about 110 atoms. Approximate magnification is 1 million times.

Other possibilities include, anti-static polymeric materials for microelectronics, such as in the packaging of chips or for computer housings, and anti-static agents for film. In many cases, static discharges can destroy sensitive electronic equipment or leave a blotch on film.

Part of the problem is in the basic science, says DiSalvo.

"We can make these perfect wires 6 angstroms in diameter. How do you make electrical contacts for wire that thick?"

For more information, contact: Francis DiSalvo, Cornell University, Ithaca, New York, USA.

3-D display that heralds new era of cubism

Many techniques have been developed to produce three dimensional image effects using two dimensional displays. Now scientists at Stanford have gone one better, developing a three dimensional display cube within which a coloured image can float in all three axis – and be viewed from any angle.

The fluorescent glass display is based on the principle of 'up-conversion', where certain atoms emit visible light when struck in rapid succession by two infrared laser beams of slightly different wavelengths. Different kinds of atoms emit different colours of light when stimulated in this fashion. By moving the intersection of two infrared laser beams around within a cube of glass that has been doped with suitable rare earth elements, the Stanford team can trace out an image that actually exists in three dimensions.

Over the years, researchers have developed a number of different ways of producing three-dimensional images – from stereo pairs; to stacking two-dimensional images on different planes such as in a cat scan; to holography, where three-dimensional information is stored in invisible patterns on a film.

But as Stanford's Elizabeth Downing, developer of the display points out, this technology doesn't just "create an image that appears to be three dimensional, it actually produces an image that is drawn in three dimensions."

As a result, there are few restrictions on the viewing angle and a number of people can view the images at once. Also, the images are emissive – they glow – rather than reflect, so they can be seen easily in room light.

The concept of displaying three-dimensional objects in fluorescent glass dates back at least to the mid-1960s. But the materials problems involved have only now been solved.

For her display, Downing used surplus scanners from optical disk players to scan the two laser beams vertically, horizontally, and backward and forward through the volume of the cube. In this fashion she has successfully created three-dimensional wire figures, surfaces and simple solid shapes.

Surfaces formed, however, are transparent, not opaque like those of most common objects. This could be a drawback for some applications, but an advantage for others.

The technology can generate colour images by mixing atoms that create red, green and blue into the glass in separate layers close together. When the laser beams stimulate adjacent layers at nearly the same time, the different colours fuse into a single coloured dot.

The current prototype device consists of three relatively thick layers, one for each colour. An actual display, however, would consist of thousands of groupings of red, green and blue layers so that 3-D objects of any colour could be created.

Prototype fluorescent cube video display enables colour images in three dimensions.

Downing considers medical imaging to be the most natural application for the new display technology and calculates that it would cost about $80,000 to make a prototype 10-inch display of this type.

Elizabeth Downing, Mechanical Engineering, Stanford University, Stanford, CA 94305, USA. e-mail: 3dlabs@pipeline.com

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Template for practical steering

Barely a month seems to go by without news of a fatal accident on the hard shoulder of a motorway, where a fast moving vehicle has ploughed into a stationary one. Would some kind of audible alert in the out-of-control vehicle, warning that it is veering out of its lane, save lives? Researchers at Carnegie Mellon university hope they have a system that one day could do just that. In fact their eventual aim is to produce an automatic steering controller that would be fitted to a car in much the same way that a cruise controller is today, to handle all steering – without intervention of the driver.

To road test their technology, Dean Pomerleau and Todd Jochem, researchers at Carnegie Mellon University, have fitted a 1990 Pontiac Trans Sport with a video camera and stand-alone hardware platform integrating a range of navigation and control technologies. As the vehicle moves along, the camera mounted just below the rear-view mirror reads the roadway, imaging information including lane markings, oil spots, curbs and even ruts made in snow by car wheels. The camera sends the image to a portable computer between the car's front seats that processes the data and instructs an electric motor on the steering wheel to turn right or left.

Simple, commercially available components have been used in the hardware, which because it is designed to be used in a unaltered passenger vehicle, has no special power or cooling requirements.

But the development that puts the Carnegie team well in the fore of automatic steering technology is its adaptive image analysis algorithm.

Navstar 5 – a 1990 Pontiac Trans Sport – has completed a 3000mile journey (almost entirely) by steering itself.

Simple, commercially available components have been used in the hardware, which is designed to be used in a unaltered passenger vehicle.

A portable computer between the car's front seats processes data from the camera mounted below the rear-view mirror and instructs an electric motor on the steering wheel to turn right or left.

In the Carnegie system, steering is decomposed into three steps: sampling the image, determining the road curvature, and assessing the lateral offset of the vehicle relative to the lane centre.

Up to now, research on automatic steering has tended to focus on machine-vision techniques that detect particular features in video images of the road. Unfortunately, where road markings vary such systems can suffer.

Other approaches combine machine-vision and machine-learning techniques. For example, in a previous Jochem and Pomerleau system, an artificial neural network was used to learn the characteristic features of particular roads under specific conditions. But "retraining" for change takes several minutes and invariably requires human intervention.

What Pomerleau and Todd's latest system does is to resample a trapezoid-shaped area in the video image to create a 30 x 32 pixel image where important features such as lane markings (which converge toward the top of the original image) appear parallel.

By comparing this current appearance with the appearance of a template created when the vehicle was centred in the lane, the system can estimate the vehicle's current lateral offset.

As the road changes character, several different strategies can be adopted to ensure the template is altered and the vehicle still correctly steered. The most sophisticated strategy, which can handle abrupt scene changes, is to create a new rapidly adapting template based on the appearance of the road far ahead (typically 70 to 100m) of the vehicle, while the road in the foreground is used to determine the current lateral offset and the curvature of the road.

The most impressive test of the system so far has been a 2850-mile drive from Washington to San Diego, completed at an average speed of 63mile/h. Results show that the system was able to steer the vehicle autonomously for 98.1% of the trip.

The researchers hope that simplicity of the algorithm should make a custom hardware implementation feasible, promising dramatic reductions in both size and cost of subsequent versions. Their goal is to build a system that is small enough to fit behind the rear-view mirror and inexpensive enough to sell as an option on passenger cars. Initially, such a system would simply warn that a driver is drifting off the road. But, in time, a system might assume at least partial control, relieving the driver of the monotonous task of steering – just as standard cruise control has done for maintaining vehicle speed. ■

For more information contact Dean Pomerleau at the Robotics Institute, Carnegie Mellon University, Pittsburgh, PA 15213. email pomerleau@cs.cmu.edu or deanp@assistware.com

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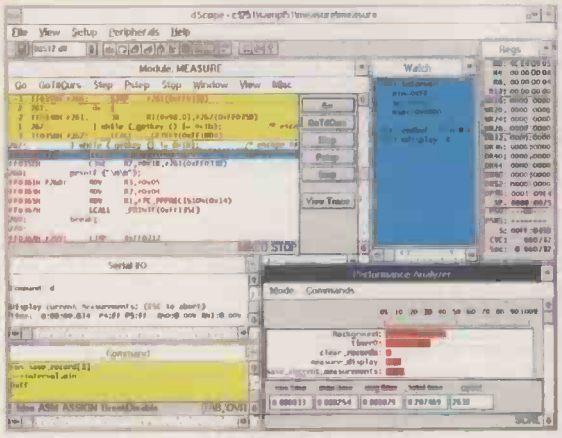

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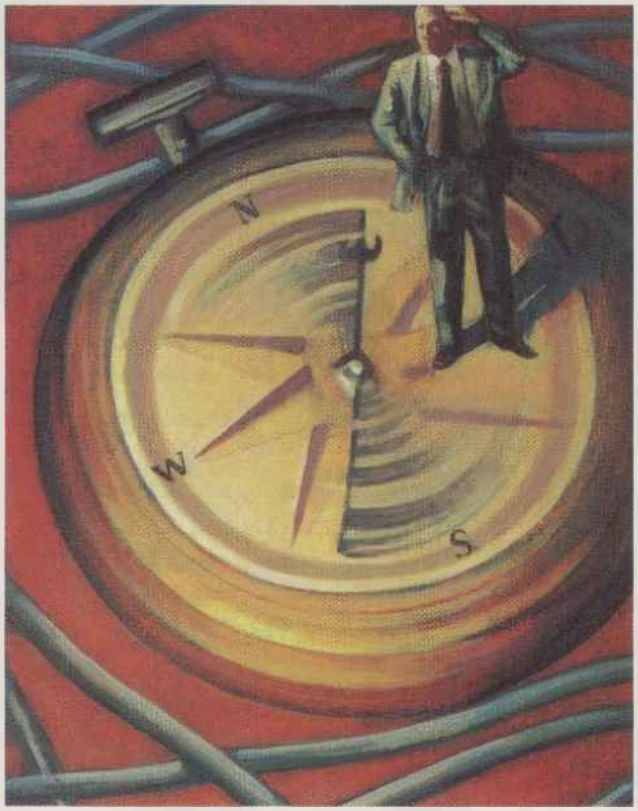
Magnetic sensing applications

There exist three-terminal magnetic sensors that are sensitive enough to detect changes in Earth's magnetic field, yet are low cost. Richard Noble describes how these sensors can be used in a number of applications – including inclination measurement and vehicle detection.

Field range of the *FGM3* magnetic sensor is 0.5 oersted, or 50 tesla. Having three leads – namely ground, 5V supply and output – the sensor produces rectangular pulses whose period is proportional to the field strength along the device axis. The *FGM3* resolves down to 10 nanoteslas, and has a variety of applications, some of which are described here. Further applications were detailed in the November 1995 issue of *Electronics World*.



Vehicle detection

Since the original article on the *FGM3* magnetic sensor, SC Ltd has designed a support IC that provides most of the functions needed to convert a stationary *FGM-3* into a vehicle detector.

Two modes of operation are provided for. One allows detection of stationary vehicles in storage or parking situations and the other detection of moving vehicle for traffic counting, etc. Both modes can be used in temporary or portable applications, as well as permanent installations since the system is fitted with automatic power-up set-up and calibration features.

In dynamic mode, the *SCL002* vehicle detection IC continuously tracks and averages background magnetic field over a short period. This averaging provides a reference level from which the brief anomalies caused by passing vehicles can be measured.

Short-term averaging is useful in that it allows automatic start-up soon after the device is switched on – or moved. It also allows the effect of a large anomaly appearing unexpectedly after set-up to be removed automatically. This can occur if, for example, a lorry chooses to stop close to the sensor during a vehicle census operation. After a brief period, its effect will be cancelled out, simply because it remains stationary.

In the static mode, selected by changing the level on one pin of the IC, the background tracking process is stopped and replaced by an initial, fixed background determination on power-up or on demand at any time by using the reset pin. This type of use includes for example the determination of slot occupancy in car parks or as a theft alarm in private garages, where a large threshold can be set to establish the presence of the vehicle parked

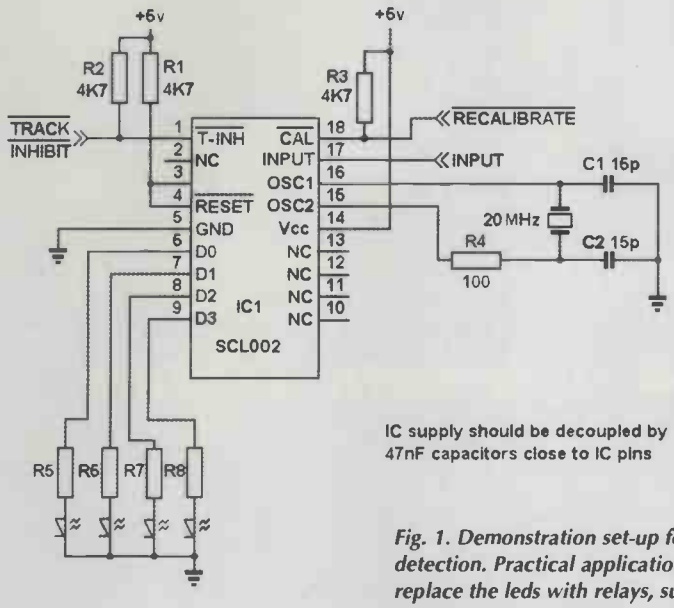


Fig. 1. Demonstration set-up for vehicle detection. Practical applications might replace the leds with relays, suitable for alarms, or connection to a data-collecting computer.

directly over or beside the sensor.

A minimum number of external components are required to produce a detection system to the level of providing four different threshold levels arranged in approximately logarithmic scale on four separate output pins. These levels are ttl compatible and can also provide sufficient power to directly drive high intensity led indicators with up to 10mA each, if required.

Pins 15 and 16 are used to connect an external crystal to give the IC a timing reference for the measurement of the field sensor's period. It is also used to set the background average timing. The sensor's output can be connected directly to the IC's input pin 17.

Pin 18 is used as a calibration input, forcing a new background average cycle whenever it is taken low. This is normally used in the static operation, but does function in dynamic mode too.

Pin 1 is an input pin which permits background tracking when high, inhibiting it when taken low, thereby providing the distinction between static and dynamic modes of operation. Pin 4 is normally taken high but will act as a master reset when taken low and then high again, forcing the chip to repeat its initialising sequence as on normal power-up.

Pins 6 to 9 are the output pins in order of increasing threshold, pin 6 being the most sensitive, with a response triggered by a change in field strength of approximately 50 nanoteslas over a one tenth second period. This is close to the minimum that can be used at this speed without detecting the normal continuously occurring micro fluctuations of the earth's field itself.

A typical application circuit is shown in Fig. 1. Track inhibit and recalibrate can be taken high through resistors to +5V if these features are not required, or used in conjunction with push buttons or switches to ground if needed.

Remember that this system is not actually a vehicle detector, but rather a magnetic field fluctuation detector and some interpretation of the results is required in practice. Vehicles do not all have the same magnetic moment and occasionally seem to have none at all. Some also seem to have multiple magnetic moments and for example an aluminium-bodied bus may produce two outputs in rapid succession as the axles pass the sensor.

Also, a magnetic moment produces field strengths which vary inversely as the cube of the distance from it, which means that the apparent sensitivity of the system tends to fall off rapidly with distance for small vehicles, less rapidly for the larger ones. Combinations of sensors on both sides of the road may be needed to resolve some of these interpretations, depending on the requirements of the operation.

Another effect which can occur is caused by the slow passage of a large vehicle close to the sensor. This can have a sufficiently large influence on the running average of the background readings to make the low threshold output persist for a longer period than normal as the average adjusts itself again. This effect is a function of the number of readings used in the average determination.

When a large number is used the effect more or less disappears, but the time taken to settle initially – or after a disturbance – increases. The performance is thus a compromise between sensitivity and settling or tracking time. For this reason the chip is made available in several versions which differ only in the number of readings used in the initial and running background averages. For example the *SCL002/64* uses 64 readings in its averaging process.

The effect described above does not necessarily mean that signals will be missed, but calls for a more sophisticated interpretation of the outputs. If a large vehicle produces the low threshold persistence and is followed rapidly by a smaller one, the low threshold output will still reverse as the vehicle passes. In other words an erroneously high output will go low if a small change in field does occur.

High sensitivity gradiometer

Another new support IC, the *SCL007*, allows two-sensor type gradiometers to be constructed.

While small magnetic field changes can be readily detected with a single static sensor, as in the case of an earth field magnetometer, the presence of the large earth field presents an immediate problem if the sensor is to be moved. A small change in angle will give a signal which is likely to be many times larger than the field magnitude variations that are being looked for as anomalies.

The gradiometer principle is based on the fact that, in a uniform field, two identical and perfectly aligned sensors will give identical outputs which can be subtracted from one another to give a zero output, effectively eliminating the apparent presence of the field. Provided the sensors remain solidly fixed in relation to one another, the whole assembly can be rotated in space without producing any orientational output.

If, however, there is a superimposed small field gradient as well as the uniform field, the output of the subtracted sensor combination will change as a function of the magnitude and direction of that gradient. Such gradients arise from the presence of anomalous magnetic moments within the capture range of the gradiometer.

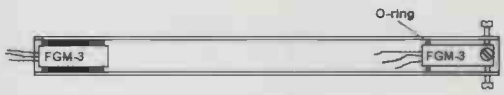


Fig. 2. Simple and flexible gradiometer construction technique, permitting a range of sizes suitable for everything from weapon detection to wreck finding.

These anomalies may arise from a great many causes, varying from the tiny firing pin of a plastic land-mine buried only a few inches under the surface, to a large marine wreck on the sea bed. The apparent capture range also varies enormously since it depends on the magnetic moment of the anomaly.

A small pole strength coupled with a very large object can produce a large magnetic moment, giving a correspondingly large capture range. Conversely even a large pole strength in a very small object can give rise to a very limited capture range. A typical example of the latter is the modern flat ceramic type of magnet which is magnetised through is thickness rather than along its larger dimension. Such magnets seem very powerful in their grip but produce very small fields at a

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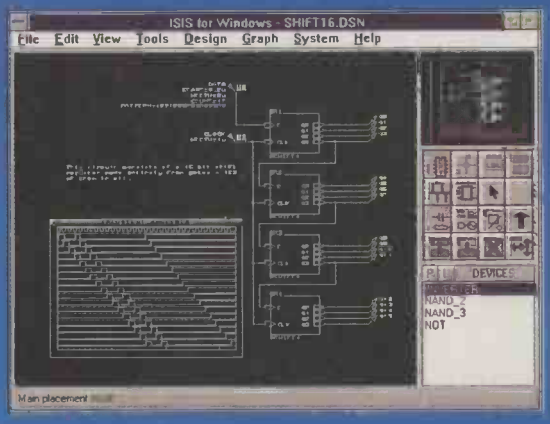

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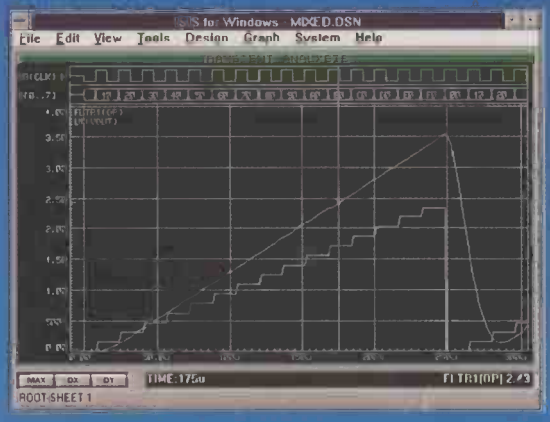

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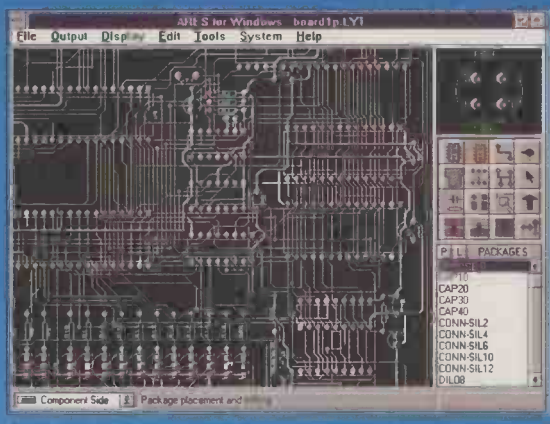

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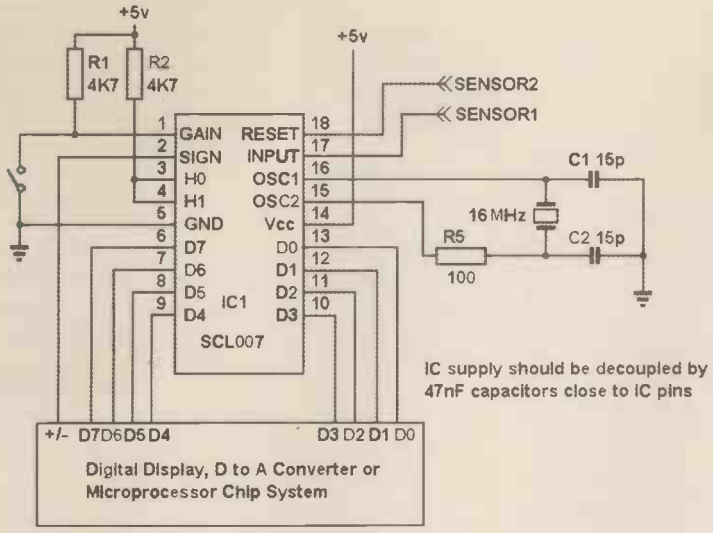


Fig. 3. Only a small number of external components are needed to provide an interface to a display system or microprocess or. Avoid meters – they have magnets in them!

distance.

In practice no two sensors are ever identical and measures must be taken to eliminate their zero-field offsets and to match their sensitivities. In practice this is not too difficult to do electronically, if an initial calibration routine is adopted on switch-on. This can be semi-automatic and requires only a simple manual manipulation.

It is more difficult however to guarantee the identical alignment of the sensor axes in the mechanical sense and for accurate instruments some adjusting mechanism will be required. This type of adjustment should fortunately not be necessary at every start-up and should remain accurate if the gradiometer is constructed from stable materials.

One method of arranging for this alignment is to build the gradiometer in a tube of diameter somewhat larger than that of the sensors. One sensor is fitted permanently into one end of the tube with appropriate packing to hold it securely. The other sensor is fitted to the opposite end of the tube but only held at one of its ends by some sort of flexible mount such as a snugly fitting o-ring. Four adjusting, non-magnetic screws can the be fitted at right angles around a circumference of the tube to force the non-clamped end of the sensor to tilt slightly in the required direction, Fig. 2.

The position of the set screws can only be determined by experiment, the objective being to reduce to a minimum the variation of output – after electronic calibration - observed when the gradiometer is rotated freely in space.

One technique for doing this is to place the gradiometer tube in mechanical engineers' vee-blocks in an approximately horizontal east-west direction and gently rotate the tube about its axis. Since the earth's field should be at right angles to both sensors in this configuration any misalignment of the sensors should result in a sinusoidal variation of output with rotation, giving a clue as to the required direction of adjustment.

A further source of potential error is the possibility that the sensors may not have identical non-linearities. This is less easy to overcome but an improvement in performance is possible in most cases by adopting an appropriate usage technique. It will vary with the application but consists basically of trying always to hold the gradiometer in the same orientation when making measurements.

For accurate measurements, where speed is not the prime requirement one good way of doing this is to suspend the tube vertically from a simple pivot allowing gravity to guarantee the repeatable alignment. In this way the gradiometer can be moved over a large grid, for example, to allow the plotting of contours of gradient in a search for underground anomalies.

Earth anomalies usually show up best in the vertical orientation, which is probably why oil companies and archaeologists make use of the vertical vector in their studies.

For simpler less accurate systems used with short ranges, for example metal detectors, it may be enough to simply maintain a constant orientation by hand and eye coordination.

Pin layout is shown in Fig. 3. Pin 1 is an input giving two different sensitivities when set either high or low. The two sensitivities, controlled by pin 1 differ by a factor of eight to provide a range for larger field anomalies. Pin 2 is an output pin which provides a polarity signal as part of the output, which should therefore be regarded as a signed magnitude, rather than the usual twos complement. This gives an extra bit of precision to the reading by effectively making the output a total of nine bits.

Pins 17 and 18 are the sensor inputs and accept the 5 volt output pulses directly. Pins 15 and 16 are for a crystal circuit to give a stable reference to measure the sensor period variations against. The remaining pins are mostly the digital output bits, D0 to D7 for use by external equipment or displays.

during the first ten to twenty seconds after switch on, the system performs an auto calibration during which it expects to see the maximum and minimum value of the Earth's field. The best way to do this is to hold the gradiometer in a north-south orientation pointing upwards at about the angle of the field's inclination, which in the UK is about 67° to the horizontal in the north/south direction. Next, switch on and rotate the gradiometer through 180° to directly reverse its direction, during the ten seconds after switch on. It is best not to do this in a hurry.

After this the system will determine the sensitivity and the zero offset for each sensor separately and correct for the errors, which would arise through sensor differences, during the signal subtraction process. It should then be possible to rotate the gradiometer slowly in any direction without getting too much output if there are no field anomalies at the location. It should be done slowly because the sensors are time multiplexed and rapid movement will beat the system to some extent.

A little practice at this technique will soon get the best cancellation and the process can be repeated as often as necessary to optimise the balance. Also for the best observations it is obviously advantageous to always have the sensor in the same orientation during the taking of readings.

You can test the success of the set-up by approaching the gradiometer with a permanent magnet as the local anomaly. The size of the anomaly will be a function of the moment of the magnet which is a function of its magnetic length.

One other point is worth mentioning. If you adopt the suggestion of hanging the gradiometer vertically, then it is an advantage to hang it from the end that makes the wires from each sensor emerge in the downward direction.

Jam jars and Earth field studies

A single axis, sensitive magnetometer for Earth field studies is relatively easy to construct using an *FGM* sensor and *SCL006A* signal conditioning IC.

This design is intended as a robust replacement for the classic – but somewhat delicate – jam-jar magnetometer, popular with radio amateurs for confirming propagation experiments. For this reason it duplicates the same type of output, namely the small angular fluctuations which occur in the direction of the earth's horizontal field component. In this way it should correlate correctly with other measurements taken by other amateurs in different locations.

A version of the *FGM-3* field sensor with higher sensitivity is used here, distinguished by an h suffix. Signal from this sensor feeds the integrated circuit, which performs all the functions required to the level of providing a digital output on eight parallel lines mirroring the tiny field fluctuations. Being digital, this output is easily input directly to a computer for data storage or digital display. Readers wishing to use a meter or chart recorder can add a *ZN429* for digital-to-analogue conversion.

The IC converts the period variations to an eight bit digital output, but only after consid-

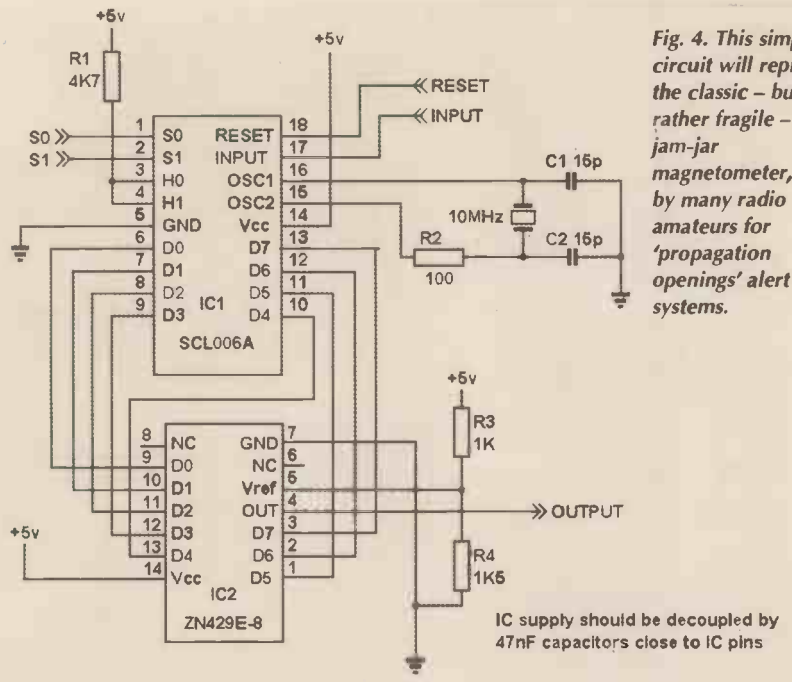


Fig. 4. This simple circuit will replace the classic – but rather fragile – jam-jar magnetometer, used by many radio amateurs for 'propagation openings' alert systems.

erable amplification and comparison to a chosen zero reference, set by changing the input to one of the IC pins, by switch or by software control from a computer output port.

Sensitivity or full-scale range can be coded by programming the levels on two input pins on the IC, either by switches or software control through a computer output port, $S_{0,1}$. Each increment in this coded input from 00_2 to 11_2 reduces the sensitivity by a factor of two.

The circuit of **Fig. 4** is for use either with a computer or chart recorder. Additional parts required for the meter or chart recorder are shown in **Fig. 5**.

If a computer is used the *ZN429* can be omitted and the lines D_{0-7} are taken directly to the input port instead. Alternatively for those with a built in analogue converter in their computer, such as the BBC computer has, the *ZN429* output can be used to input the data via this channel.

If a meter or chart recorder is used, remember that the *ZN429* has an inherent output impedance of 10kΩ and so can only provide a maximum of 100μA per volt of output. This calls for a meter with a sensitivity of 50 to 150μA to ensure a full scale deflection capability. Voltage sensitive chart recorders should have in input impedance greater than 10kΩ or a full scale range of around one volt.

In the case of a meter or current sensitive chart recorder a series 10kΩ variable resistor will permit adjustment of the scale readings. The same can be done for a voltage sensitive chart recorder by using a larger variable resistor in potentiometer configuration, Fig. 5.

If direct digital input is chosen, it should be noted that there is no strobe, interrupt or handshaking facility available from the chip. This gives rise to the risk of data lines changing during input, causing incorrect reading.

The software should take two readings in rapid succession and only accept data if they are identical, taking a third reading if necessary to obtain this identity. This problem does not arise in the case of a computer using an internal analogue to digital converter. The *SCL006a* output changes once per second and the computer input scan rate need only be slightly higher than this to collect all the available data.

This rate is probably still too high for most applications and the computer can be used to average the readings over a longer period before storing or displaying the results. Typically, a plotted point once every three or four minutes will produce a 24 hour recording across the screen, depending on the screen resolution in use.

In normal use the sensor will be mounted horizontally with its long axis lying on an east-west line. On power-up the output will automatically move to half-scale and all subsequent field variations will be measured up and down from there. At any later time the output can be set again to the centre-scale position by applying a ground level to the reset pin for slightly longer than one second. This can be done by switch or computer control and allows the user to remove any apparent bias in the reading range, caused by switching on by chance on a peak fluctuation.

The switch arrangements in **Fig. 6** will provide the control signals needed by the chart recorder type systems.

If you want to calibrate the output, it is possible to achieve a reasonably good relative calibration with simple equipment. A single layer close wound coil into which the sensor can be inserted is all that is needed in the way of additional hardware. It can be wound on a cardboard or plastic tube using any wire available from enamelled to plastic covered link-up wire.

Ideally the coil should be about ten times as long as its diameter, so it is advantageous to use the smallest diameter tube that the sensor will slide into. The sensor should be placed in the centre during calibration but exact positioning is not critical.

The information needed for calibration is the length of the winding in millimetres and the total number of turns. Multiply the number of turns by 1000 and divide by the length to obtain the number of turns per metre, *N*. If a current of i amperes is passed through this coil the magnetic field at its centre, in units of amperes/metre is just $N \times i$. To convert this to oersted $N \times i$ should be divided by 80 . To convert to gamma multiply by 1250.

Field in such a coil will be uniform over the length of the sensor to better than 0.5%. To correct for the finite length of the coil multiply the figure obtained above by 0.992 for best results.

The current required is small. Typical plastic coated wire, for example close winds to about 835 turns/metre and 0.5mA will produce just over 500 gamma. A simple series 10kΩ potentiometer and milliameter connected to the 5V power supply will suffice.

During calibration the sensor should be set in the east-west position away from potential

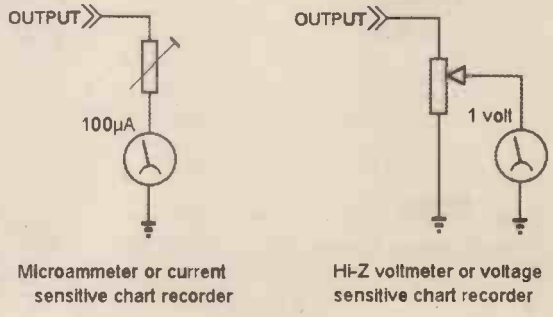


Fig. 5. Additional circuitry needed to connect output of the magnetometer to a chart recorder or voltmeter.

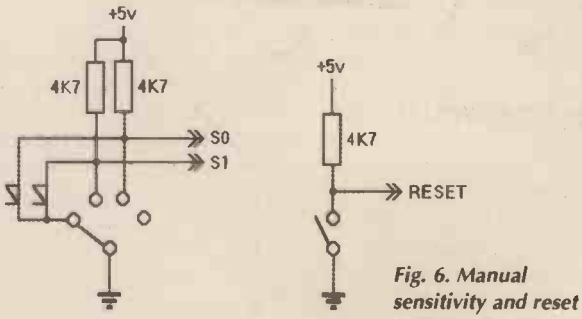
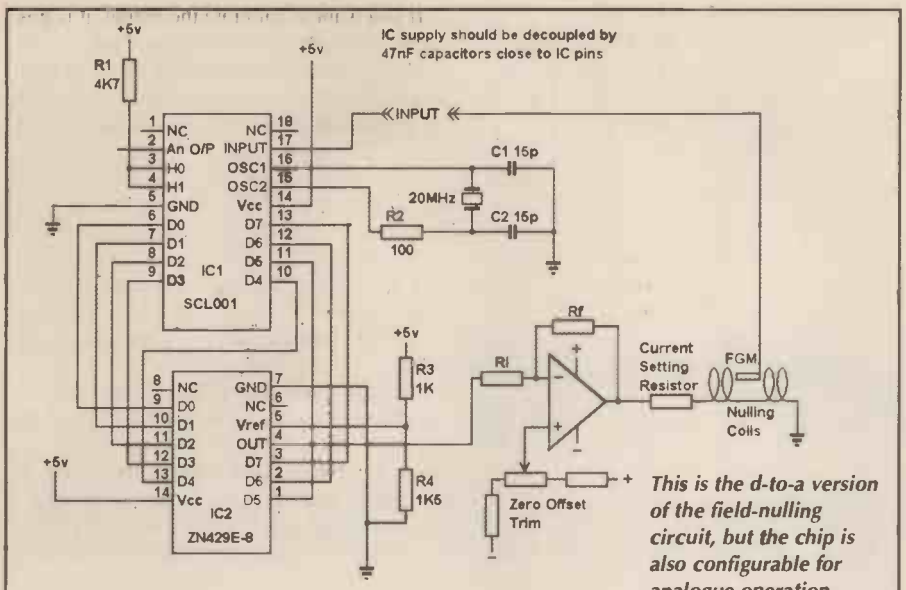


Fig. 6. Manual sensitivity and reset switches for the magnetometer.



This is the d-to-a version of the field-nulling circuit, but the chip is also configurable for analogue operation.

Magnetic field nulling

A further application-specific integrated circuit is available to provide most of the functions required to provide automatic cancellation of low level magnetic field interference. The technique employs a closed loop containing a sensor to measure the local field and a magnetic field generating coil system to provide the cancellation. A typical application is the reduction of interfering fields near the neck of crt display tubes.

Further information on this applications, FGM sensors and the ICs mentioned here is available from Speake & Co, Elvicta Estate, Crickhowell, Powys NP8 1DF, tel 01873 811281, fax 01873 810958.

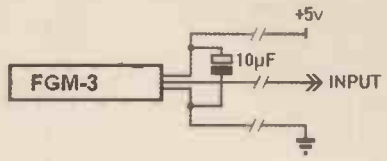


Fig. 7. Connecting a capacitor close to the sensor body minimises interference.

sources of interfering magnetic field.

Output impedance of the sensor is 330Ω and the leads to it can be augmented by quite long lengths of cable without much effect on the rectangular pulse output. This permits the sensor to be located remotely from the rest of the equipment, something that is normally necessary to avoid local field anomalies caused by ferrous metal objects being moved around and electronic equipment or electrical cables.

Moving vehicles, for example can be detected at distances of 4 to 5 metres. Burying the sensor at the bottom of the garden away from the road or mounting it on the roof are possible strategies to resolve this.

Although the sensor has been designed to minimise radiation, if it is used in conjunction with high gain radio receivers, it is obviously advisable to screen the long cable to the sensor and also to decouple the supplies as close to the sensor as possible to avoid potential harmonic pickup. The sensor connections are shown in Fig. 7. ■

No-contact current measurement

Current measurement usign a shunt has three disadvantages – the circuit under test has to be broken, the shunt resistor affects the reading, and the measurement system is not galvanically isolated. A magnetic sensor circumvents all of these, as Steve Winder explains.

Measuring current using a voltmeter relies on the voltage drop across a resistor that has been inserted in series with a circuit, Fig. 1. This method has several disadvantages. In particular, the series resistor introduces a voltage drop that in turn causes a reduction in the current flow. This causes a measurement error. Additionally, the measuring system is directly coupled to the circuit being measured and the circuit has to be broken before a measurement can be made.

Using a moving coil ammeter to measure current flow involves having its movement coil inserted in series with the circuit being measured. This means that resistance is introduced by the wire used in the coil, again reducing current flow and causing an error in the reading.

Multimeters usually have a separate resistors for each current measurement range. Low current ranges require a higher resistor value in order to produce sufficient voltage drop to be measured by the multimeter.

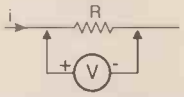


Fig. 1. Current is usually read by measuring the voltage over a shunt resistance, but inserting the shunt usually alters current flow.

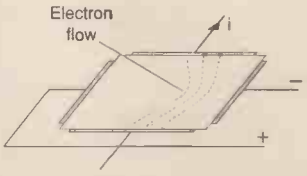


Fig. 2. In a Hall effect device, current is made to flow across a semiconductor. Voltage at the sheet sides varies with magnetic field.

Isolation between the circuit being measured and the test equipment is not usually a problem for low voltage circuits. It is more likely to be important where high voltages are used however. Operational amplifiers with an isolated output are available. This isolation is achieved by chopping the direct-current input signal, passing it through capacitive coupling circuits – or exceptionally through a transformer – and then rectifying the signal to produce a dc output.

The disadvantage of the chopping method is that the output may need filtering to remove the switching transients produced by the chopping process. This slows down the response of the circuit.

Measuring the current in a wire without breaking the circuit can also be important in some applications. It may not be possible to break a circuit, perhaps for safety reasons, for example in a fire alarm system. Alternatively, it may be that the equipment cannot be turned off since it is continuous use. Production line machinery is an example of this.

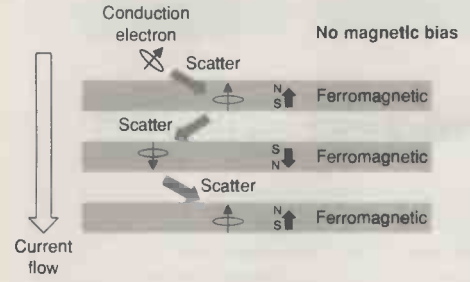


Fig. 3. Giant magnetoresistive devices comprise alternating sheets of magnetic and non-magnetic thin films. This diagram shows conditions when no magnetic field is present.

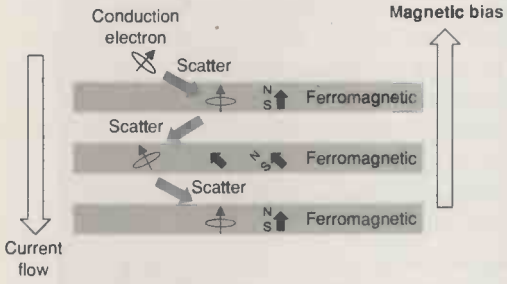


Fig. 4. Resistance of a giant magnetoresistive device reduces when magnetic field is applied.

Magnetic field sensors

Current flow causes a voltage drop when passing through a resistor. It also causes a magnetic field to be generated. There are three possible methods of sensing this field.

One method is to use a magnetometer. This device has a series of coils around a non-linear magnetic core. Alternating currents through the coils generate a flux that either aids or opposes the flux being measured. The difference between the energy supplied to one coil, compared with another, gives a measure of the flux being measured.

Magnetometers are quite large – approximately the size of a fat cigar – and are very sensitive. These devices are sometimes used for sensing the Earth's magnetic field.

Another method involves the Hall effect. Both linear and digital output Hall-effect devices are now available in TO-92 and eight-pin dual-in-line packages[1]. In a Hall-effect device, a sheet of semiconductor material forms the sensor.

Current is made to flow across the semiconductor sheet while voltage difference across the sides of the sheet are measured. When magnetic flux passes through the sheet, in the perpendicular direction, electrons are deflected on their path through the sheet. Because the electron density on one side of the sheet is higher than the other side, a voltage is developed across the sheet, at right angles to the current flow, **Fig. 2**.

Hall effect devices are not very sensitive; an analogue sensor has a typical working range of 40mT. Hall devices are more readily available in their digital form. Digital Hall devices comprise a Hall sensor and a comparator. This addition gives a logic output dependent on the flux – typically switching at 20mT. Hall devices are sometimes used in conjunction with permanent magnets to sense the movement of ferrous materials.

The magnetoresistive option

The third option involves the magnetoresistive effect. There are two magnetoresistive effects; anisotropic and giant. Anisotropic means that the effect is different dependent on the direction of the flux. The magnetoresistance effect is produced using thin-film layers of magnetic conductors and anisotropic magnetoresistive devices have a sensitivity similar to Hall effect devices.

Being discovered in 1988, giant magnetoresistance, or gmr, is a relatively recent discovery. It has been implemented by alternating magnetic and non-magnetic layers of thin-film conductors[2]. Ferromagnetic materials have spin-polarised conduction electrons.

Adjacent magnetic layers in a gmr device couple together to produce anti-parallel electron spins. As current passes through layers of magnetic material, the electrons are scattered by the alternating magnetic spin of the conduction electrons in each layer. Electrical resistance is produced when electrons are scattered, **Fig. 3**.

When an external magnetic field is applied, the conduction electron's spin in the different layers begin to align with each other and the conducting electrons are scattered less. In this way, the resistance is reduced, **Fig. 4**.

Recently, magnetic field sensors with sufficient sensitivity have been made available in IC form. US company Nonvolatile Electronics Inc has produced devices in surface mount SO-8 packages with full scale sensitivity down to 15 oersted. These are available from Rhopoint in the UK. The method used was to place four resistors in a bridge arrangement with a magnetic shield covering two of the resistors, **Fig. 5**.

The magnetic shield performs two functions: it shields resistors R_1 and R_2; and it concentrates the magnetic flux through resistors R_3 and R_4. In this way, an external field affects the resistance of resistors R_3 and R_4, but leaves resistors R_1 and R_2 unaffected and therefore suitable for use as a reference. A supply voltage is applied across two nodes of the bridge. The remaining two nodes are used to sense a voltage difference.

When a magnetic field is applied, resistors R_3 and R_4 reduce in value, making the node joining R_2 and R_3 more positive than the node joining R_1 and R_4; this is illustrated by the bridge circuit in **Fig. 5**.

When a current flows through a wire a magnetic field is produced. The magnetic flux density is given by the equation:

$$B = \frac{\mu_o I}{2\pi r} \text{ tesla}$$

Alternatively, since one tesla is 10^4 gauss,

$$B = \frac{10^4 \mu_o I}{2\pi r} \text{ gauss}$$

If the dielectric medium is air – or other non-magnetic material – then you can define the magnetic field as:

$$H = \frac{I}{2\pi r} \text{ amp/m}$$

Alternatively, since 1A/m = $4\pi.10^{-3}$ oersted:

$$H = \frac{2.10^{-3} I}{r} \text{ oersted}$$

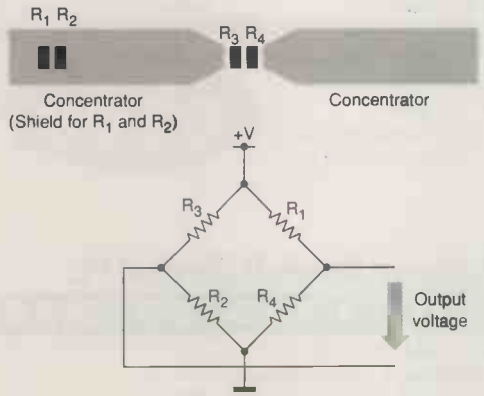


Fig. 5. Giant magnetoresistive sensor layout. In addition to screening $R_{1,2}$, the shield concentrates flux through $R_{3,4}$.

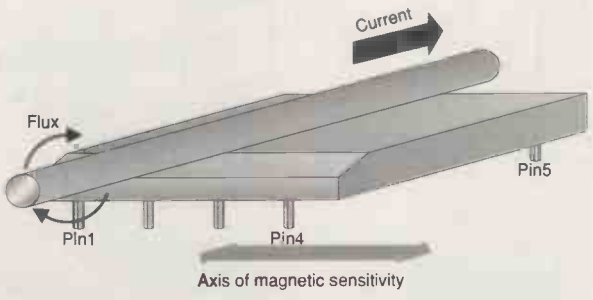


Fig. 6. Placing a conductor a precise distance from the gmr sensor allows current to be measured without calibration.

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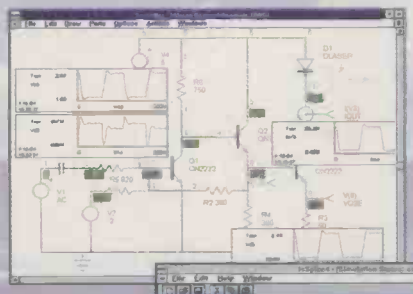

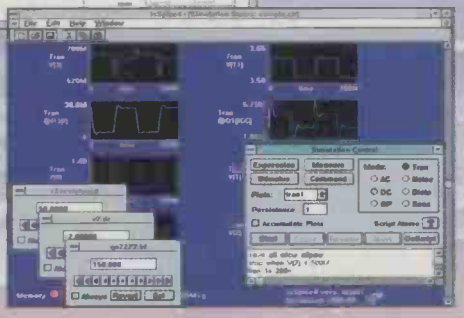

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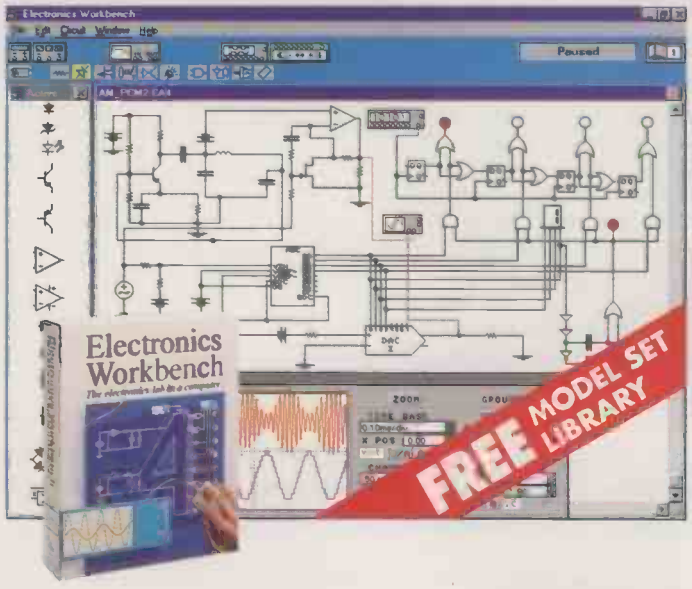

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Using gmr for current sensing

The giant magnetoresistive sensor is approximately in the centre of the SO-8 package, or about 0.65mm below the top surface. Placing a 0.7mm wire across the top of the gmr sensor gives a distance from the conductor centre to the gmr sensor of 1mm. This is convenient for solving the equations given to determine the magnetic field.

With a current of 500mA through the wire, the magnetic field at the sensor is 1 oersted. The device is sensitive to fields along the axis of the IC, as illustrated by **Fig. 6**, which shows how the trial current sensor was used.

To test the gmr sensor, I built a a differential amplifier, **Fig. 7**. This amplifier had a gain of 20, set by the 1kΩ resistor placed between the inverting inputs of the two input op-amps. Low noise op-amps were used. The input stages incorporate a dual *TLC2202* c-mos device while the output stage is a bipolar *TLE2027*. With hindsight the, *2027* limited the output voltage range and a c-mos device may have been a better choice.

I connected the gmr sensor IC across the power supply and into the differential amplifier. A 0.7mm diameter wire was glued in place across the IC and used to carry current to be sensed by the device. The *NVS5B15* gmr sensor uses 5kΩ sensing resistors and has a full scale sensitivity of 15 oersted.

Output voltage of the gmr device is proportional to the supply voltage. The *5B15* gave 50mV output per volt of supply at a field strength of 15 oersted.

I used a 10V supply, so the maximum output was 500mV. With a field of 1 oersted, relating to 500mA through the sense wire, the gmr device produced an increase in output of 500mV/15 or 33.33mV. Output of the differential amplifier increased by 666mV due to the gain of 20.

Interpreting the output

The reason for describing the output as an increase in voltage rather than an exact value is that, with no magnetic field present, the differential amplifier produced a voltage output of 0.65V. This was due to offset voltages across the gmr device.

The offset voltage was temperature dependent. A reading of 0.65V at room temperature, 20°C, rose to 0.75V at 35°C. This offset and temperature dependence could pose a problem when detecting small currents. The offset is easily removed by biasing the differential amplifier. To reduce or remove the temperature dependence however would require a more sophisticated circuit and knowledge of the temperature coefficient. This information is not supplied in either the application notes or data sheet.

Sensitivity of the current sensor can be reduced by increasing the distance between the centre of the conductor and the gmr sensor. If the distance between the conductor's centre and the gmr device is increased to 2mm, a current of 1A would produce a field strength of 1 oersted. The surface of the gmr device package would have to be 1.35mm from the centre of the conductor. This can be achieved by using a larger diameter wire, of 2.7mm, or by having insulating material between the copper wire and the IC package.

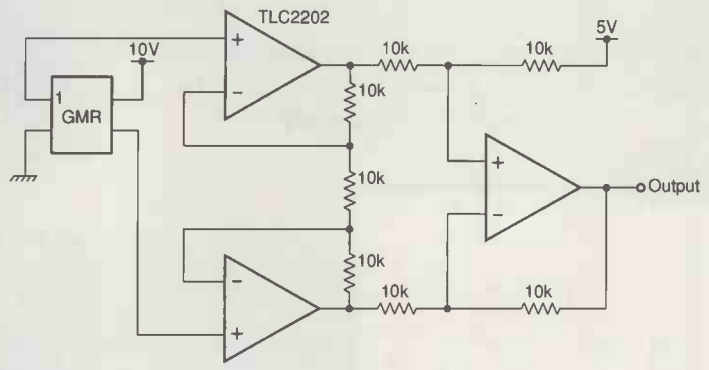


Fig. 7. Giant magnetoresistive device amplifier. Current of 500mA though the sensing wire causes a field of 1 oersted, increasing the differential amplifier output by 666mV. Power supply is 10V.

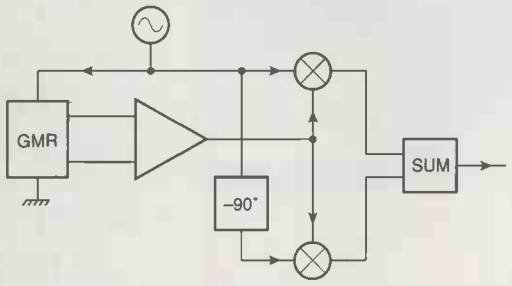


Fig. 8. Giant magnetoresistive sensors suffer form noise. Modulation could be used to remove this.

Magnetic field produced by current flowing in a wire can be calculated by considering all the current to be flowing in the centre of the conductor. To find the field strength at a point, the distance should be measured from the centre of the conductor.

The gmr sensor suffers from noise. This is $1/f$ noise and is about ten times the noise voltage of thin-film resistors. The $1/f$ noise dominates up to about 10kHz, above which thermal noise is dominant. Sensitivity of the device can be improved by applying an ac signal with a frequency greater than 10kHz across the gmr device, instead of using a dc bias. Output of the gmr device will then be an amplitude modulated carrier.

Amplifying this signal and then demodulating it, using the same ac signal, produces a dc signal proportional to the magnetic field. Any low-frequency noise is frequency shifted to high frequency and can be filtered without unduly slowing the response of the circuit.

Figure 8 shows an implementation of this system. The carrier signal is input to the mixers in quadrature and the outputs are summed so that the phase of the signal from the gmr device and amplifier do not affect the amplitude of the output.

In summary

The body of the SO-8 package used by the gmr sensor would provide some electrical isolation between the sensor and the circuit being measured. Using insulated wire would enable a much greater isolation voltage to be achieved.

Gluing the sensor to an existing wire would enable measurement of current without interrupting the circuit. No resistance is introduced into the circuit being measured, thus no power losses occur and the circuit's current is not affected by the measurement.

Development work is being carried out by NVE to find more sensitive materials and to find out the cause of the $1/f$ noise. If improvements are possible there will be many more applications for these devices. ■

References

1 Martin Eccles, Applying Hall to Good Effect, *EW+WW* July 1994.
2 Rhopoint Components, GMR Sensor Application Notes, Sept 1995.

Rhopoint, UK distributor of the *NVS5B15*, is at Holland Road, Hurst Green, Oxted, Surrey RH8 9BB, tel 01883 717988, fax 01883

Gain from passives?

Better than knowing? Ian Hickman's latest article looks at circuits whereby gain is obtained in unusual ways.

Some time ago, browsing through one of my files of cuttings, I came across an article by that guru of analogue design, Bob Pease, of National Semiconductor. It described, among other things, a passive circuit composed of just three capacitors and three resistors – not a transformer or tuned circuit in sight – which gives a voltage gain in excess of unity at one frequency.

Now this intrigued me, as it goes against one's instincts of what is possible in electronics, not to say what is decent. Of course, you can analyse the circuit to find the transfer function, although the algebra gets rather lengthy. You get a third-order equation in f, the frequency, or rather in $j\omega$, where $\omega=2\pi f$. Separating out the real and imaginary parts, and equating the latter to zero will enable you to solve for the frequency at which the phase shift through the circuit is zero.

You can then substitute this value back into the transfer function and arrive at the gain at the zero phase shift frequency. You can also differentiate the modulus of the transfer function with respect to frequency, and set the result equal to zero. Solving this equation will give you the frequency at which the gain is a maximum – not necessarily the same as the frequency of zero phase shift.

I contemplated doing just this, but decided against, being not only rather lazy, but also notoriously liable to make algebraic errors. But a more cogent reason for not doing it, is that at the end of the day, I would know what the circuit does, but would not really understand the way that it does it. And understanding is much better than just knowing.

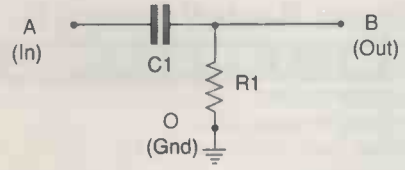
Look – no maths!

So the circuit analysis was undertaken with the aid of graphical constructions known as circle diagrams, handy things that were touched on in reference 1. There, they were used to illustrate a passive lag, or low-pass (top cut) *CR* circuit.

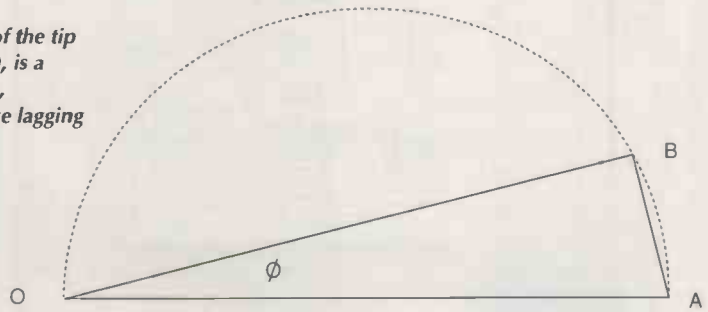
Now the passive circuit with gain mentioned above is best worked up to bit by bit, so let's start with a single section of it, a passive lead or high-pass *CR* section, **Fig. 1**, and assume that the circuit is driven from a zero impedance source and its output monitored with an oscilloscope with an infinite (all right – very high) input impedance.

At very low frequencies, very little current

Fig. 1a) Simple passive lead (high-pass) circuit. b) Circle diagram, left, showing that at any frequency the locus of the tip of the vector OB, representing the output voltage at point B in a), is a semicircle. Note, the same current flows through both C_1 and R_1, producing an in-phase voltage across R_1, but a quadrature voltage lagging the current by 90° across C_1, so angle OBA always equals 90°.



a)



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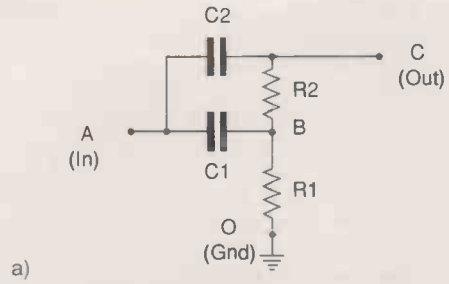


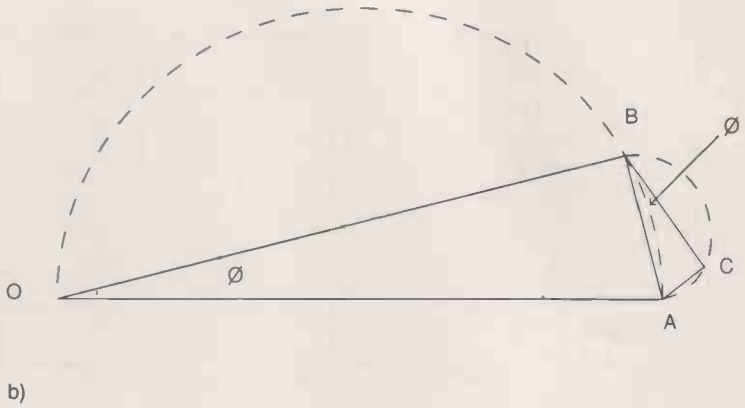


Fig. 2a) Network with two capacitors and two resistors. b) In this case, the circle diagram looks something like this (with provisos, see text). Note that output vector OC can exceed the input OA.

flows (since the reactance of C_1, Fig. 1a), is very much higher than R), and what current does flow, leads the applied voltage by very nearly 90°. Consequently, the tip of the vector OB in the vector diagram of Fig. 1b), representing the output voltage at point B, will be slightly above the ground reference point O, lying on the dotted semicircle.

The angle ϕ will be almost 90°, while, if you remember your school geometry, the angle OBA is always exactly 90° (angle subtended by a diameter, at the circumference). As the frequency increases, the angle ϕ diminishes, so that when $f=1/(2\pi CR)$, the point B lies at the highest point of the semicircle, $\phi=45°$ and the voltage across the resistor OB equals the voltage BA across the capacitor.

As the frequency is made higher still, phi tends to zero and the voltage BA across the capacitor tends to zero, this is the case illustrated in Fig. 1b). But not until the frequency becomes infinite does the output voltage equal the input voltage and ϕ become exactly zero.

A passive circuit with 'gain'...

Take a look at the circuit of **Fig. 2a)**, which strangely for such a simple circuit, is the subject of a patent[2]. In **Fig. 2b)**, the input voltage OA and output voltage OB can be represented by a circle diagram as before, with again ϕ being shown as much less than 45°, indicating a frequency well above $f=1/(2\pi CR)$. The voltage drop across C_1 is applied to the passive lead circuit C_2R_2, so the output voltage at point C, represented by vector OC, is clearly larger than the input OA. As the frequency tends to infinity, ϕ tends to zero, and the vector OC coincides with the vector OA. So there is a range of frequencies where the gain is greater than unity, but the phaseshift is always positive (or zero). This is shown as a Bode plot in **Fig. 3**.

However, a couple of niggling provisos are needed to clarify Fig. 2b). Firstly, the circle diagram OBA is as in Fig. 1, provided the loading of C_2R_2 on the voltage across C_1 is negligible. This can be ensured by making say, $R_1=R_2/1000$ and $C_1=1000C_2$, keeping the CR products the same. This was the condition assumed in the analysis in Fig. 3, though in truth there is little difference if the two Rs and the two Cs are the same. For at frequencies where the peak occurs, well above $1/(2\pi CR)$,

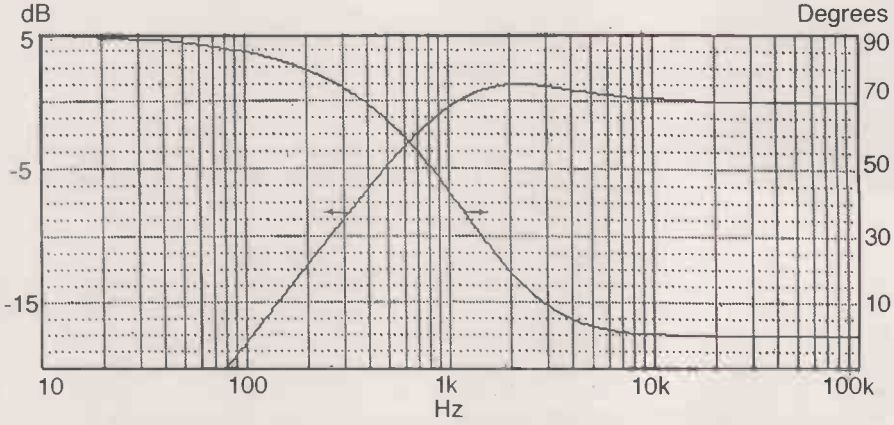


Fig. 3. Bode plot for the circuit of Fig. 2a), for the case where $R_1=100\Omega$, $C_1=1\mu F$, $R_2=100k\Omega$, $C_2=1nF$. Thus the loading of C_2R_2 on the C_1R_1 circuit is negligible. Note: if $R_1=R_2$ and $C_1=C_2$, the plot is almost the same except that the peak gain falls from +1.25dB to +0.65dB.

the reactance of C_1 is much lower than that of C_2 and R_2 in series. The only significant difference is that with equal values, peak gain is only about +0.6dB against +1.25dB in Fig. 3.

Secondly, while the circle diagram OBA is accurate enough, the circle diagram BCA shows what the voltage BC across R_2 does as the frequency varies, assuming the magnitude of the vector BA remains constant. But of course it doesn't; it too varies with frequency.

The semicircle BCA must thus be regarded as notional, since it varies in size. In fact, the locus of the tip of the output vector OC, representing how the output at C varies with frequency, is not a semicircle. It starts out at zero frequency following the circle OBA from the point O, but gradually diverges from it, becoming a little larger. Finally, it tucks back under to the left, approaching point A from the right.

...and one with zero phase shift

To make a sinewave oscillator, you need a frequency selective circuit to determine the frequency.

If the maintaining amplifier has unity gain, the frequency selective circuit must also have a gain of at least unity, at the frequency at which it provides a phase shift of zero° (or 180° if an inverting amplifier). But the circuit of Fig. 2 provides a lead at all finite frequencies: the phase shift is not zero until you get to infinite frequency, by which time the gain is back at unity. However, the gain of an opamp, with its output tied back to the inverting

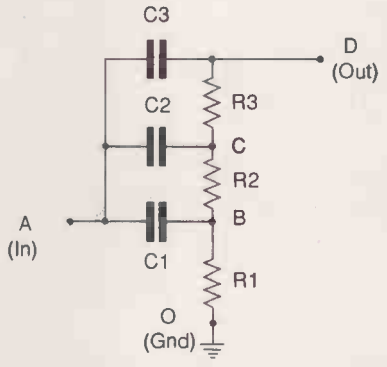


Fig. 4. Circuit giving a gain greater than unity at a finite frequency at which the phase shift is zero.

input as a unity gain buffer, is slightly less than unity, in fact $A/(1+A)$, where A is the opamp's open-loop gain. So with the circuit of Fig. 2 between its output and its non-inverting input, it can't oscillate, can it? What is needed for an oscillator using a unity gain maintaining amplifier, is a circuit with zero phase shift and a gain just in excess of unity, at a finite frequency. Such a circuit is shown in **Fig. 4**.

I haven't drawn a circle diagram for it, but you can see how it goes. Note that in Fig. 2b), the semicircle and lines BCA are a smaller scale version of OBC. Now sketch in a smaller version still, CDA, constructed upon CA as diameter. As phi tends to zero, the point D will meet and cross the horizontal axis, pro-

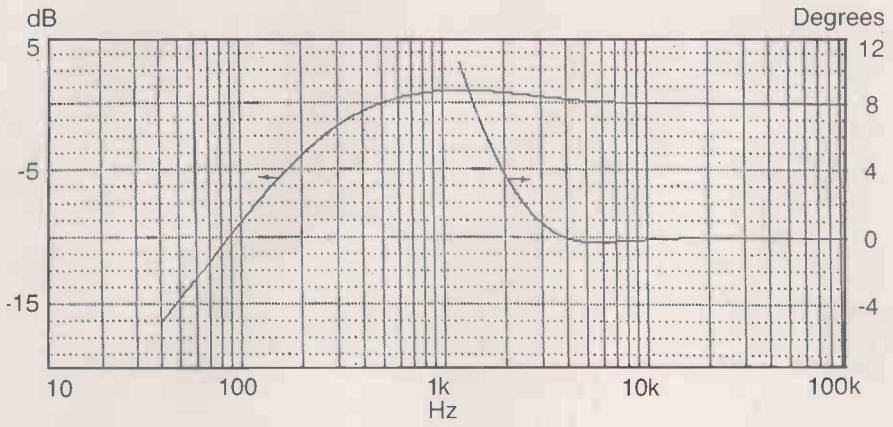


Fig. 5. Bode plot for the circuit of Fig. 4, where all Cs are 1nF and all Rs=100kΩ. Gain peaks at +0.9dB at 1.2kHz, but is still in excess of unity at the zero phase-shift frequency of 3.9kHz.

viding a zero and even negative phase shift, whilst the vector OD representing the output at D in Fig. 4 is still greater than the input OA.

In fact, where the locus of the tip of the output vector in Fig. 2b) approaches point A at infinite frequency from the right, the locus of the point D curls round back under and finally up, approaching point A from directly below. This represents an ultimate phase shift (internal to the circuit, but not appearing at the output) of 90° more than the second order circuit of Fig. 2, and 180° more than the first order circuit of Fig. 1. This is just what you would expect in fact from a third order circuit.

(Note that the Bode plot of the gain and phase shift versus frequency in **Fig. 5** is for the case where all three capacitors are 1nF and all three resistors are 100kΩ.)

An awful oscillator

If the circuit of Fig. 4 is used as the feedback network around a unity gain amplifier, **Fig. 6a)**, an oscillator results. The gain of the *TL071* buffer stage is very close to unity, while Fig. 5 shows that at the frequency of zero phase shift through the *CR* network, it still has about 0.3dB voltage gain. Consequently the loop gain exceeds unity by about this amount, and the amplitude of oscillation builds up until there is heavy clipping, **Fig. 6b)**, upper trace.

At under 3.5kHz, the frequency is less than the 3.9kHz predicted by the zero phase-shift frequency in Figure 5, but this is the usual experience when an erstwhile sinewave oscillator (without an *LC* tank circuit) has excess loop gain. When driven into saturation at each voltage extreme, the opamp's internal gain stages take time to recover; in fact, the circuit verges on a relaxation type of oscillator.

Inserting a 10kΩ preset pot at the point X in Fig. 6a), and tweaking judiciously as required, resulted in a near sinewave, as shown in the lower trace. Since the *CR* network has a capacitive component of input impedance, this has resulted in the introduction of some extra lag into the loop, and the frequency of oscillation has consequently adjusted itself to about 2.6kHz, where the network provides a compensating lead of about 2°.

Figure 7 compares the performance of the circuit with a *TL071* (upper trace) and with a *CA3130* opamp, lower trace. The lower amplitude is due to the maximum ±8V supply rating of the latter, compared with the ±15V used with the *TL071*. The waveform is better, possibly due to the lower slew rate of the *CA3130* when compensated for unity gain, but the reason for showing both traces is to highlight one of the unsatisfactory aspects of the circuit. Although it is not measurable in Fig. 7, the frequencies were 2592Hz (*TL071*) and 2642Hz (*CA3130*), a difference of 2%.

In a good oscillator circuit, the actual frequency should be much more independent of minor differences in the performance of the maintaining amplifier. To prove the point, the same two opamps were tested in a Wien bridge oscillator circuit, using two of the 100kΩ resistors and two of the 1nF capacitors from the Fig. 6a) circuit. The theoretical oscillation frequency is 1592Hz, and the actual fre-

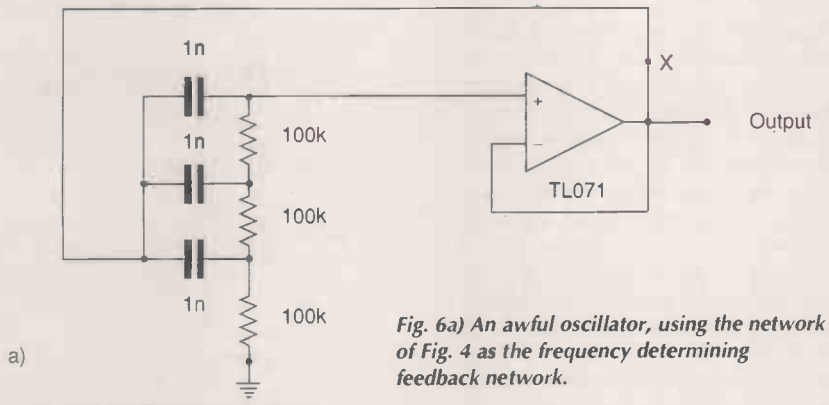
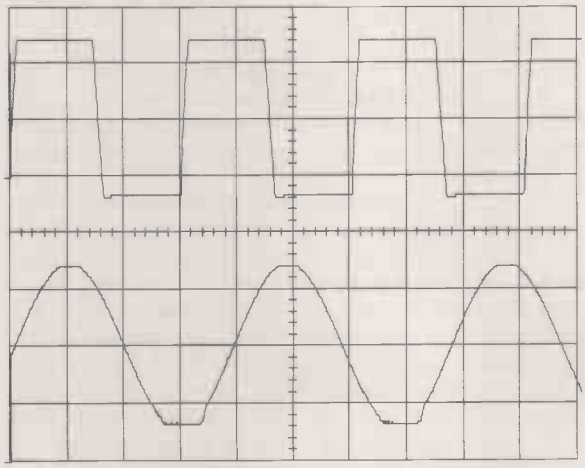
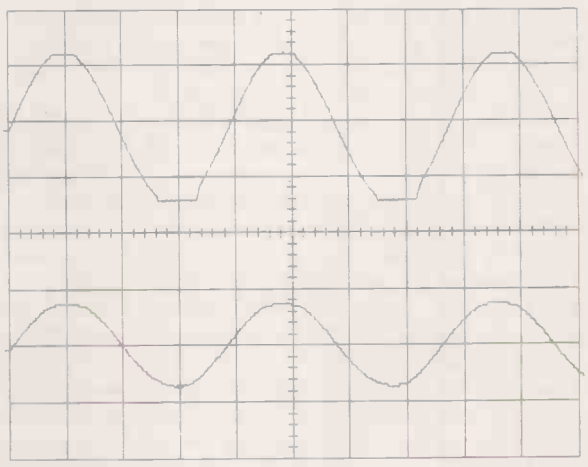


Fig. 6a) An awful oscillator, using the network of Fig. 4 as the frequency determining feedback network.

Fig. 7. Output of the circuit of Fig. 6 – including the 10kΩ potentiometer – with a TL071 (±15V supplies, upper trace) compared with the performance with a CA3130 (±8V supplies, lower trace). Frequency differs by 2%. Both traces are 10V/div., 100μs/div.



b) (upper trace) The output of circuit a). (lower trace) With the addition of a 10K preset at point X, critically adjusted. (both at 10V/div., 100μs/div.)



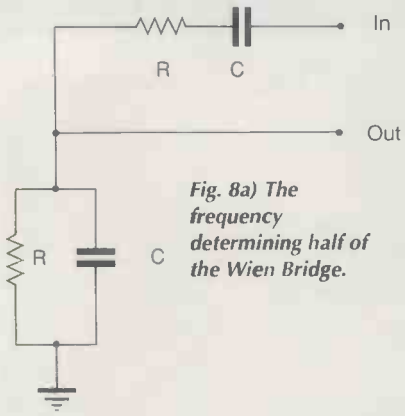
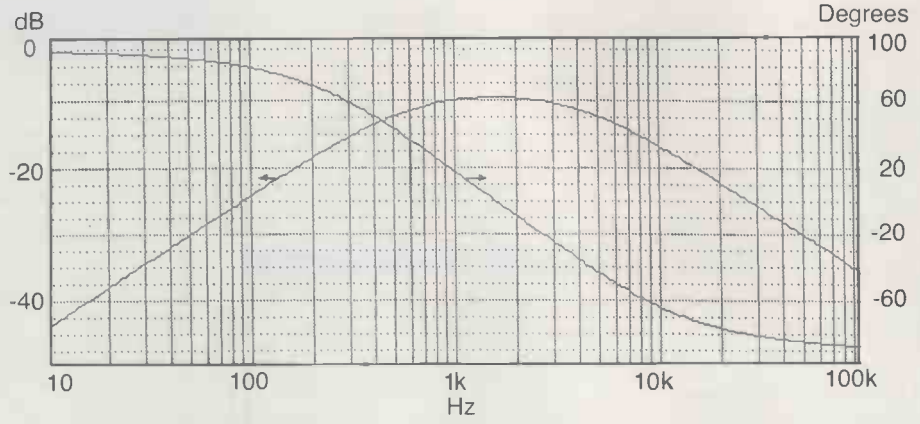


Fig. 8a) The frequency determining half of the Wien Bridge.



b) Phase and amplitude response (Bode plot) for a). Note the much more rapid change of phase with frequency in the region of maximum output, compared with Fig. 5.

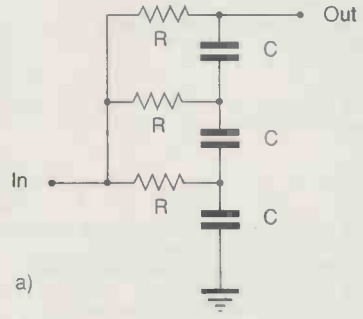
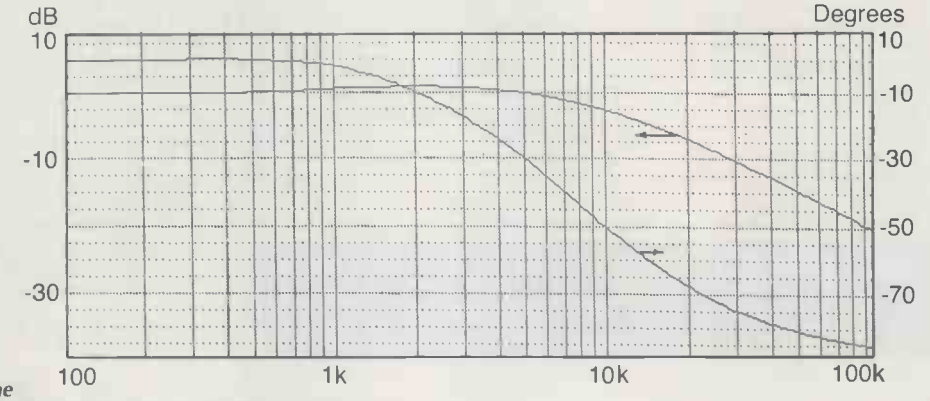


Fig. 9a) Low-pass dual of the circuit of the Figure 5 high-pass circuit.



b) Bode plot for the circuit at a).

quency was 1552Hz with either opamp; changing the *TL071* supply volts from ±8V to ±15V making no difference whatever.

The reason is the much greater rate of change of phase with change of frequency in the Wien bridge circuit, as shown in **Fig. 8**. (But it has a minmum attenuation of –9.5dB or a gain of $^1/_3$, so that a maintaining amplifier with a gain of ×3 is required.) This much greater 'phase slope' means that any slight variation in phase through the maintaining amplifier will result in a much smaller compensating frequency change. Of course, a high Q tuned circuit provides an even greater phase slope and hence less susceptibility still to the vagaries of a maintaining amplifier, and a quartz crystal a much much higher phase slope even than that.

Another disadvantage of the network of Fig. 4 as the basis of an oscillator, is the fact that it is basically a highpass filter. As such, it offers no attenuation of any harmonic distortion produced in the amplifier. The *RC* network in the Wien bridge, however, is a bandpass circuit, and as such, even with its low Q of $^1/_3$, does provide some attenuation of harmonics.

Every circuit has its dual

Every circuit has its dual, and Fig. 4 is no exception. The dual is shown in **Fig. 9a)**, and as you can see it is a low-pass circuit. Despite being also a third order circuit like Fig. 4, the stopband roll-off is, like the earlier circuit, at 6dB per octave. Figure **9b)** shows its frequency response in the form of a Bode plot, for the case where all three capacitors equal 1nF and all the resistors 100kΩ. At zero hertz, or dc, the gain is unity and the phase shift zero, the phase exhibiting a small positive value of a degree or so in the region of 300-500Hz. The amplitude also peaks, by about 0.9dB, the maximum occurring at about 2kHz.

As the frequency determining network in an oscillator, it would have the advantage over Fig. 4 of providing some attenuation at harmonics. But it cannot be used in practice, since the attenuation at dc is zero. This means that if you connect it in the circuit of Figure 6a), you simply get a bistable – or do you? Except for a 100kΩ resistor in series with the non-inverting input, as far as dc is concerned, the two opamp inputs are shorted together. I haven't tried it, but it doesn't look to be a useful circuit.

Tailpiece

For the with-it readers who spotted it, and more for those who didn't, I stated earlier that the gain of an op-amp connected as a unity gain non-inverting buffer is slightly less than unity, and that the phase shift in the circuit of Fig. 2 does not reach zero until infinite frequency, by which time the gain is back at unity (both true).

And that therefore when the overall loop gain (network plus opamp in a Fig. 6a) type circuit but with the two *C* – two R network of Fig. 2) was unity the phase shift was not zero and when the phase shift was zero the gain was less than unity.

This is true in the case of an ideal opamp, but in practice, the buffer stage will start to exhibit a little phase lag long before its gain falls appreciably. This more than compensates the residual lead in the network of Figure 2, so that if the two *C* – two *R* network is used in the circuit of Figure 6a), the circuit will in fact oscillate. With a *TL071*, it produced a heavily clipped sinewave – virtually a squarewave – at about 10kHz. But with the frequency determined by the point at which two tiny phase shifts, both changing slowly with frequency, cancel each other, the exact frequency is anybody's guess. ■

References

1. Ian Hickman, In the Picture, *Electronics World* July/August 1996 pp 558–561.
2. U.S. Patent No. 2 730 679, 1951.

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CIRCLE NO. 122 ON REPLY CARD

The route to pcb cad

Rod Cooper explains what to watch out for with autorouters and takes a look at Circuitmaker and Traxmaker design software.

PCB CAD review subjects

This review, which began in the September issue and continues next month, covers the following ten products.

PCB Designer: Niche Software Ltd, tel. UK 01432 355414. £49 inclusive (see September issue).
PIA: AW Software, tel. Germany +49 89 6915352. PIA std 99DM: extended 171DM 32bit 286DM inc tax (see September issue).
Easytrax: Protel International pty, tel. Australia 408 437 7771, UK PDSL, tel. 01892 663298 (see September issue). £6 copying charge.
Ranger2: Seetrax CAE Ltd. 01705 591037, (see October issue) £150 exc £10 p+P and VAT.
Electronics Workbench: Interactive Image Technologies Ltd (Canada), tel. 0014169775550. UK Robinson Marshall, tel. 01203 233 216, (see October issue) £199 exc p+p and VAT.
CircuitMaker: MicroCode Engineering (USA) UK agent Labvolt, tel 01480 300695. Circuitmaker and Traxmaker cost £199 each excluding vat and p&p.
Quickroute 3.5 Pro+: Quickroute Systems Ltd.
Propak: Labcenter Electronics.
Proteus: Labcenter Electronics.
EasyPC Pro XM: Number 1 Systems.

Note that although it started last month with a couple of smaller packages, this review is not in any order of complexity or competence.

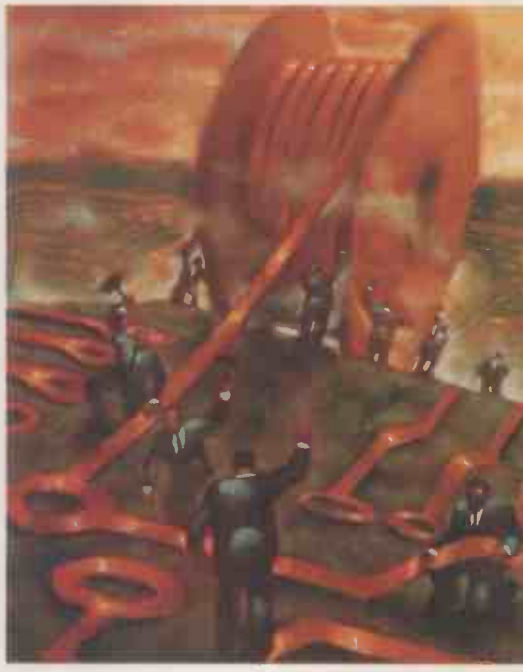
To the uninitiated, autorouters may seem like the perfect answer to the drudgery of manually routeing a board, but they have their own well-hidden snags.

Few low-cost autorouters can route a large or dense board 100% without difficulties of one sort or another. Some only reach 70% completion, leaving the undone tracks for a little 'interactive routeing'. You may think that manually routeing the other 30% would be easy, but you would be wrong. The autorouter will have done the easiest routes, leaving you to do the hardest. Moreover, in doing the easiest tracks on the board, it will have selfishly blocked off the spaces a human would have left for following tracks, so you will be faced with undoing the autorouted tracks just to get in the remaining manual tracks.

Undoing and re-routeing tracks that your expensive autorouter has already routed must surely rank as one of the most foolish and time-wasting CAD activities ever devised. Not only that, but you will probably be carrying out the final routeing by what will seem, at first, a truly awful technique called 'rubber banding' as described in the glossary in the September issue. It follows that if you want an autorouter, you must go for one that can route 100% or near. These have tended to be the most expensive, but there is now a handful of lower cost versions.

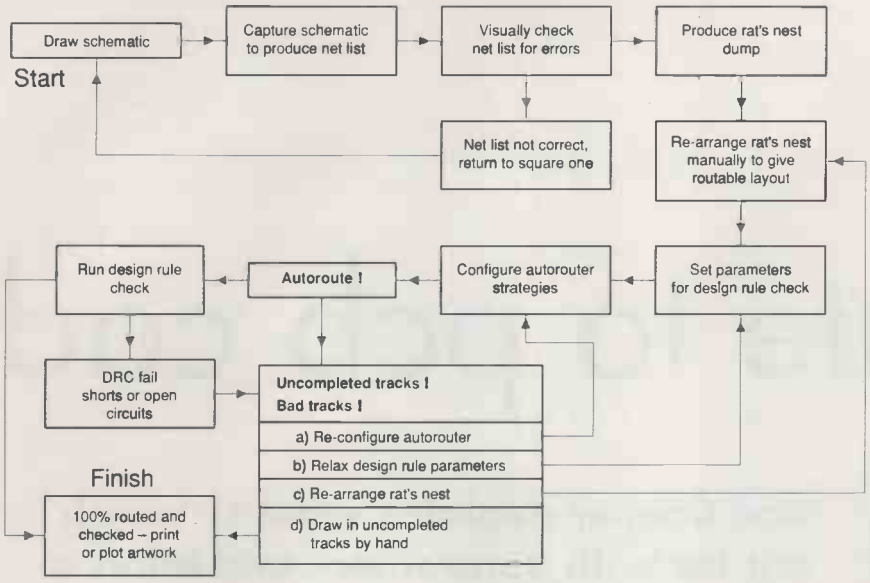
Most autorouters – especially the low-cost alternatives – balk at single-sided boards. You will not often see a CAD program claim to do a good single-sided board. I have only seen one such claim, and that was in a system where the autorouter alone cost £5300.

On the whole, autorouters do not produce tracks that compete with manually-laid tracks produced by a person with average competence. The routeing sometimes seems illogical,



the overall result can be aesthetically displeasing, and board functionality and manufacturability is often not up to standard.

Sales literature often refers to autorouter speed. If the autorouter is going to leave uncompleted tracks, then you want to know about it as soon as possible so that you can remake the rat's nest or alter the prerun configuration. For the best completion rates, an autorouter should have rip-up-and-retry and push and shove features. The speed of the autorouter is less important if you are running Windows, provided that you have sufficient resources to leave the autorouter running in the background while you get on with something else.



This chart shows the typical steps from start to finish for a schematic capture and autorouter system.

With an autorouter that you know is weak, then a short run time is a prime requirement. You will notice that all the better autorouters have high pc requirements – especially in the memory department – to improve both run times and success rates. There is clearly a trade-off between the length of run-times, success rates and the power of your pc.

Gridded versus Gridless

Gridded autorouters rely heavily on memory, and using a fine grid to emulate a gridless autorouter puts a large memory load on the pc, so it slows down. In addition, a finer grid can be counter-productive as Figs 1a,b) show. In the first case, Fig. 1a), with the coarse grid, the autorouter can put a track between pads A and B without infringing the design clearance distance 'd'. In the second, with the finer grid, the autorouter cannot put a track in because the clearance with pad A or pad B is less than 'd'. The autorouter would throw this out as an uncompleted track, even though a human operator could see that it could easily insert the track and satisfy 'd'. You would have to make the grid finer still to get back to the situation where the track was routable, extending the routeing time considerably.

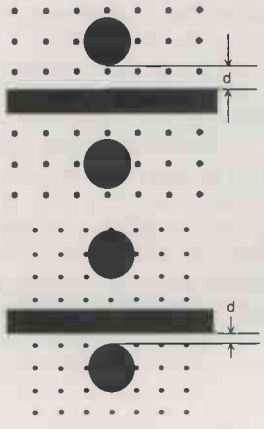


Fig. 1. Finer grids are not always better. In a) a coarse grid allows the design clearance between the track and pad to be achieved, but b), using a fine grid, does not.

One of the claims for gridless autorouters is that, by using alternative methods like the geometric-shape technique, the load on the pc's memory is lower. But this does not necessarily mean the router will be quicker. A lot depends on the routeing strategies being used.

The pre-run configuration of the autorouter has a significant effect on the artwork produced, so it is highly desirable to be able to alter the configuration of the autorouter. Such optimisation can help the result meet the demands of the pcb specification, it can reduce the number of vias to a minimum to reduce cost, and it can help make the pcb easy to manufacture by maximising clearances.

The configuration will dictate which strategies are used by the autorouter. For example, it is pointless to run the memory strategy if there are no areas of regular, repeating memory-type tracks.

Configuration capabilities should allow control of the lengths of track, allow or prohibit some via holes, and allow a decision as to whether the autorouter is permitted to autoneck or autoshave, and so on. Generally speaking, the more configurable an autorouter is, the better the results.

Current autorouters rely heavily on the operator to produce a viable rat's nest before routeing. Rotating just one component can alter the autorouter success rate dramatically, but knowing which component to rotate is a bit of a black art. With a weak autorouter, you can end up spending more time arranging and running the rat's nests than it would take you to manually route the board.

If you don't mind routeing the board manually, one of the simple pcb programs that just provides you with the pad symbols and component outlines is the best choice. That way, you get many of the computer's advantages without getting bogged down in the morass of steep learning curves, rubber-banding, and autorouters that don't route properly.

One-sided offerings

Nearly all the autorouters reviewed here are intended primarily for making large double-sided boards. Not one is aimed at the single-sided board user. This may give the impression that the single-sided board is *passé*, in some way inferior, but if you look at the photograph, **Fig. 2**, you will see that as usual the Japanese have a very different view. This board was designed by two manufacturers in Japan for mass production of office equipment. It is a microprocessor-based dc controller. It is not too small for a double-sided design, but nevertheless it is single-sided. The important thing to note is that the designers have placed more jumpers than components in their desire to remain single-sided.

Clearly it more economical to do things this way. But why? The industry standard used to be that if you had more than 50 components or half-a-dozen jumpers it was time to go double-sided. Obviously, this has now been superseded. The widespread use of robotic component insertion machines means that it is cheaper to insert more fixed-length jumpers than make a double-sided pcb – even though it is easier and quicker to design a double-sided board.

Not only is single sided board material cheaper, but the cost of removing copper is greatly reduced and the cost of making vias is of course nil. Plus which, solderability is said to be much better, and there is only need for one solder mask, one copper master etc. Naturally there are no production difficulties aligning the copper track on one side with the other. Prototyping is far easier and quicker with a single-sided board. Also, where boards are to be repaired rather than scrapped if proven faulty at the manufacturing stage or later during service, it is much easier to repair a single-sided board. Eight good reasons!

Another technique to avoid double-sided boards often seen in consumer and office equipment of Japanese design is to make mother\daughterboard arrangements with two or more single-sided boards. Sometimes these are hard-wired together to save the cost of a plug and socket.

Double-sided boards are rarely essential. With some notable exceptions like computers and the miltary, there are few applications where there is a pressing physical need or proven technical advantage in using them.

One trick used with weak autorouters to produce a single-sided board is to run the autorouter twice. First run a single-sided layout. This may well produce some undone tracks. Next, run a double-sided layout on the uncompleted tracks only. With appropriate autorouter configuration, this will often result in 100% completion with just a few tracks on the top side. These can then be turned into jumpers.

There is a large and obvious unfilled gap in the market for an affordable single-sided autorouter that does the job properly. I predict that such an autorouter would enjoy instant success, and it would have this part of the market to itself as it stands at present.

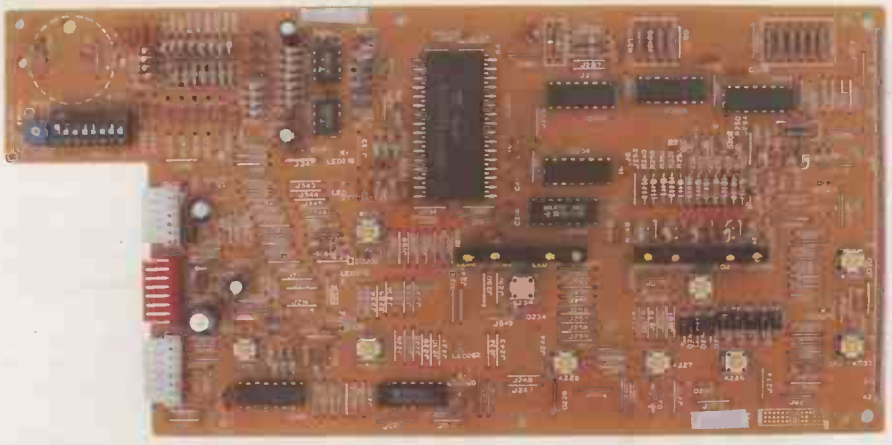


Fig. 2. Microprocessor-based dc controller of Japanese manufacture illustrates that it may wise to consider single-sided designs even for fairly complex

Autorouter testing

I discovered that it was difficult to devise a standard test that could produce a meaningful result for all the autorouters under review. This was because they could all route double-sided boards, without producing reject routes, until a large board size was reached.

The main difference between the various autorouters lay in how many vias were produced and the total length of track used to make up the design. Such factors as component density and variation in configuration of strategies unfairly influenced the results from some autorouters. Eventually I decided that a small single-sided board gave the best indication of the power of the autorouters. Such a board made it easy to fabricate a deliberate difficulty to show how the better autorouters could overcome the problem, and how the weaker ones could not.

I divided the autorouters into four categories by means of this test. Those in category A were able to complete the test, those in category B could complete it with a small relaxation of the design rules, those in category C could not complete it. Autorouters in category D could not attempt a single-sided board or did not work from a rat's nest. The **Table** presents a comparison of the autorouters reviewed.

The board design I used does not bear much relationship to a real board because I made generous space allowances everywhere so as not to inhibit certain routeing strategies. This would have made the test unfair.

Finding exactly matching component outlines proved to be problematic, so these vary a little in each program tested. The time taken for each router is also given, but this is on a relative scale only, in order to allow for differences in pc speed. As a rough indication, if a router is rated at 1 on this scale, then it would take a minute or less in running time to give the result shown on a 386 with a co-processor running at 20MHz with 16Mbyte of ram. A router rated at ten would take ten minutes, and so on.

Although routeing power and speed are important, it should be pointed out that each autorouter has its own set of attributes some of which may make a particular autorouter attractive even though it may be comparatively slow, or low on routeing power. ■

Autorouter comparisons

Category A – able to complete the test circuit (relative time taken in brackets)

Specctra (2)	from	Range 2
MultiRouter (2)	from	Easy-PC
386 Rip-up (10)	from	Ranger2
ARESIII (2)	from	Proteus
AR3 (5)	from	Quickroute 3.5

Category B – able to complete the test circuit with slight relaxation of design rules

Ares (2)	from	Propak

Category C – unable to route the test circuit completely

Range2 Standard (1)
Traxmaker (1)
Quickroute 3.5 Standard (3)

Category D – unsuitable for use with the test circuit

P.I.A
EasyTrax

Review 1 – *Circuitmaker*

This program is a schematic drawing and capture product for Windows with digital, analogue and mixed-mode simulation provided all in one package. That is, neither the simulations nor the main libraries are sold separately as in the No1 System arrangement, but are provided as part and parcel of the product. This makes *CircuitMaker* very good value for money.

There is also a pcb drafting program complete with autorouter, available as an extra, called *Traxmaker*. Curiously this runs under dos and as you may suppose, the exit from *CircuitMaker* into *Traxmaker* is not effortless. You may find the plunge from an easy-to-use intuitive Windows program to a menu-driven dos program disconcerting.

CircuitMaker needs at least a386 pc with 4Mb of ram, and a co-processor for the analogue simulations, which are based on Spice. You will need Windows 3.1 for the 16 bit version, which is the one I tested. A 32-bit version is available which requires Windows NT or 95. A well-written and comprehensive manual is provided and there is good on-line help and a separate help directory, plus a very good tutorial. The level is pitched just right for an practical introduction to CAD. Users will detect a slight bias towards the educational

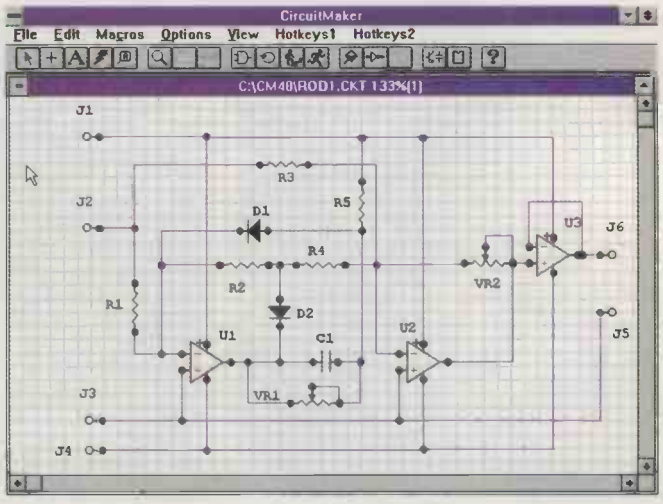


Fig. 14. Schematic in CircuitMaker. Note the lattice grid and pin-out on 741.

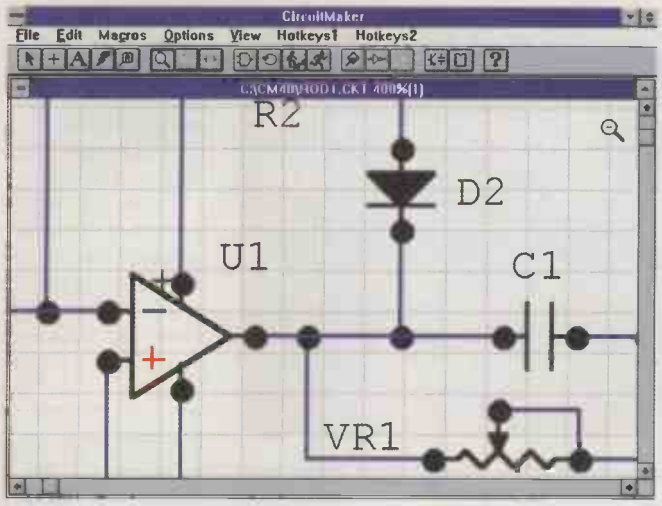


Fig. 15. Zoom view of schematic, showing quality of graphics. Compare this with some DOS screens from next month's reviews.

field, but this should not stop professional engineers from using it

The schematic drawing part of *Circuitmaker* is excellent. A lattice grid is used instead of a dot matrix, akin to that used in *Quickroute* and the available drawing area on a 14in monitor is fair at about 9 by 5.5in. The full drawing area is several times more than this. There is no support for multi-sheet schematics.

Double use is made of the title bar. It is used for displaying button bar information when any of the buttons are selected. Little things like this show that a lot of thought has been put into *CircuitMaker*'s display area to keep the drawing area from becoming cluttered.

Selecting and placing components is easy and the libraries are extensive – 1500 components, most of which carry simulation information. Access to such a large library is necessarily a little slow, but *Circuitmaker* gets round this with the hotkey concept, whereby frequently used components can be called up by one key from a parts bin.

I prefer having a large library and slow access to having fast access and a small library. The library is well organised and useful. Symbols not in the 'hot-key' parts bins are selected one at a time from the library and sent straight to the drawing area; there is no parts bin for these symbols, so this process is a little slow overall.

Parts can be automatically annotated, and they can be rotated at the selection stage with the right mouse button, which is a convenient method, or later on using the rotate tool. Component text, such as R2, 100k and BC108, stays upright during component rotation and may be moved independently to any position.

Placing multiple symbols of the same type is speeded up by using the repeat function to copy existing symbols. Should you need a component not in the library, you can make up your own functioning model as *CircuitMaker* is fully expandable. A new component can be cross-referenced with package outlines in the pcb program if required.

Drawing of lines is orthogonal, and a long cursor line is used to assist placement. There is inhibition of lines that don't make contact with pins, and as an extra aid, a small red box appears when the cursor is within striking range of a pin. This system, called SmartWire, makes *CircuitMaker* one of the easiest and quickest schematic drawing programs of this review. You are unlikely to create lines without connectivity with this system and your netlists will be sound.

An automatic router similar to Propak's *WAR* is also provided. It can put in an orthogonal line for you if you click on the two pins you want to connect. This can operate in two modes – simple or intelligent. Simple takes the shortest orthogonal path, intelligent skirts round obstacles if possible. Like *WAR* in *Propak*, some editing of routes may be required to avoid the occasional odd effect, but the main advantage is speed.

Panning is carried out with the standard scroll bar concept. There is no map to locate the drawing, but it is debatable whether one is needed. When placing a device in the drawing area, autopanning is performed. There are ten levels of zoom, plus a zoom control which makes your circuit fit the screen fully – a feature well worth having.

Overall quality of schematic drawings is good, but some devices – such as the *741* shown in the test circuit – do not show all active pins. This is a sign that *Circuitmaker* has its roots in simulation rather than pcb production. But it is possible to edit devices and draw in the missing active pins.

Converting the circuit into a netlist for export to *Traxmaker* was fairly easy, the only snag being that it is necessary to specify a *Traxmaker* component outline for any schematic symbol that doesn't have an allocated outline already. To do this, the component outline information has to be extracted from *Traxmaker*, and ferried it over to *Circuitmaker*. This means leaving *Circuitmaker*, making an exit from Windows into dos, starting up *Traxmaker*, finding the outline, then reversing these steps to get back to where you began to insert the data.

This illustrates one of the penalties for having two different formats in a system.

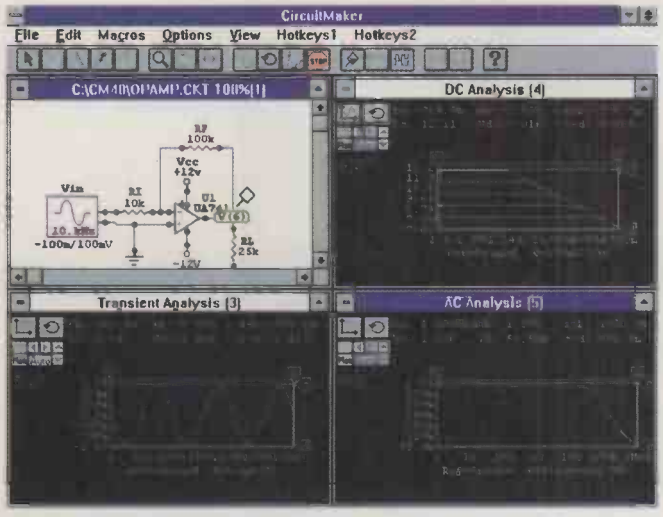


Fig. 16. Typical analogue simulation of a circuit imported into Circuitmaker from Propak using SPICE netlist transfer. Note how the sig gen is connected and the test probe location.

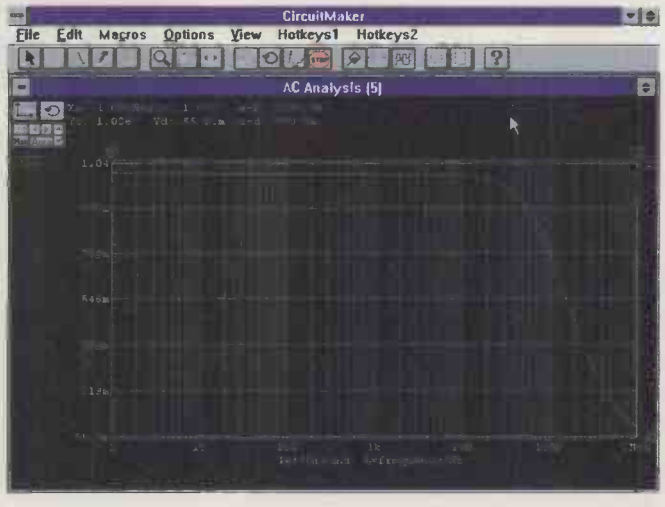


Fig. 17. Any analysis from the above simulation can be run full-screen for making accurate measurement.

However, to be fair, the majority of *Circuitmaker* components do have outlines allocated so you are not likely to do this frequently. But if you hit a similar snag – a net list problem springs to mind – you would have to go through this rigmarole to correct it. With a system like this, co-operation between the two parts, such as the automatic forward/reverse annotation as found in fully integrated packages like *Propak*, is difficult to implement. This problem is not unique to *Circuitmaker*, it is found in all set-ups where a third-party pcb package is added on to schematic drawing/capture program without full integration. Having one in dos and the other in Windows compounds the problem.

After compiling the net list, the next step is to make a note of the file name of the net list to be transferred and quit *CircuitMaker* and Windows. The rest of the process of producing pcb artwork is carried out in *TraxMaker*. To complete the netlist transfer, start *Traxmaker* from dos, enter the drawing part of the program, and import the file using a full dos path command.

Of the combined schematic-capture and autorouter products reviewed, this was one of the most long-winded netlist transfers, but I did not find it difficult.

Net list outputs in Spice and pcb format and Windows metafiles can be generated. The pcb net list raises the possibility of exporting the schematic to a third-party autorouter, perhaps a Windows-based product, but the netlist compatibility would have to be carefully checked.

CircuitMaker can import a Spice2 or 3 net list. This could be useful if you already have a pcb artwork program with Spice export and wanted to add on an inexpensive simulator. I tried the Spice netlist from *Propak* into *CircuitMaker* with complete success.

Simulators are included with *CircuitMaker* and I am sure these will be one of the deciding factors influencing would-be purchasers. The simulations in CircuitMaker are as easy to use and intuitive as the schematic drawing program.

As already mentioned, *CircuitMaker* will be of interest to educationalists. There is a section that allows deliberate faults to be put into circuits that could be very useful to teachers, if not to designers using *CircuitMaker* in earnest. Also, numerous circuit diagrams already set up and ready to demonstrate are included with *Circuitmaker*.

Review 2 – *Traxmaker*

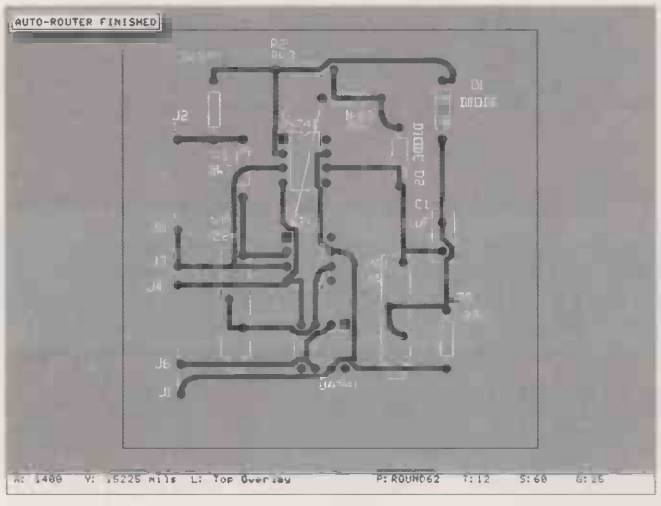


Fig. 12. The test circuit was autorouted with Traxmaker. One net is incomplete, putting this autorouter in category C.

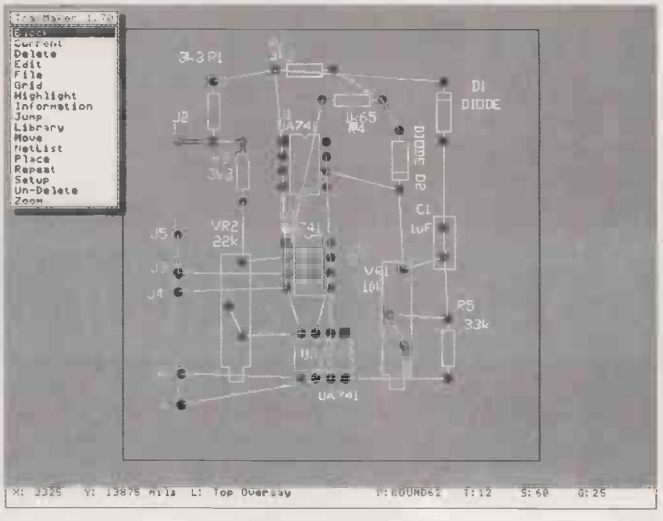


Fig. 13. Rat's nest in Traxmaker after interactive editing, showing a typical Traxmaker menu.

Traxmaker needs a 386 or higher pc and 640K of ram. Expanded memory is supported, and it is advisable to provide some to avoid 'out-of-memory' messages.

The package comprises a manual drawing package, a semi-automatic pad-to-pad router and an autorouter, and it can make single-layer, double-sided or multilayer boards. Traxmaker could be used as a stand-alone package, but here it is specifically tailored to take a netlist from *CircuitMaker*.

This is a mature menu-driven dos product and is well-developed. Users will notice an immediate similarity with a product already reviewed, *Easytrax* by *Protel*, and will guess its origin. There is a good manual, re-written for use with *Circuitmaker*.

Like *Easytrax*, *Traxmaker* has a 32in by 32in board area and a drawing area on a 14in monitor of about 9.5in by 6.5in. It has many useful features, such as adjustable autosave, and a good library of component outlines, which is text only.

There is an autopan facility but no map showing where you are, a feature beneficial to any program that incorporates autopan. To compensate for that, there is a 'jump' feature, allowing the screen to jump to a particular component.

Seven pre-set zoom levels are available and there is a zoom function for making the circuit fit the screen. The basic manual drawing program is unremarkable and can produce good results with a little diligence.

Importing a net list from *Circuitmaker* has already been discussed. Creating a rat's nest for the autorouter in *Traxmaker* presents two unusually good features. Firstly, the program has a function called 'auto-space' which can deposit the components in a relatively ordered fashion around the chosen board area. This may not be the exact arrangement you want, but it is a big improvement on having the components placed in a linear array or in a pile one on top of the other as in some other programs.

Having the components in an ordered form gives you a head start towards creating the desired rat's nest. Also, it is possible to configure auto placement to optimise the way it places the components inside the board area. This gives you an even better start. Secondly, there is a function within auto placement which can put all the components onto the chosen grid. In this way, the grid-type autorouter can route without difficulty. This feature is useful if you have been maneouvering components about and have not exactly located them back on the grid. Combined, these two features save a lot of time and effort.

There is the ability for creating tracks manually from a rat's nest but the rat lines stay in place, which can lead to a very complicated-looking piece of artwork. However, a facility is

available for removing each rat line interactively as you progress, or removing them all from sight at one go. This is not as good as the *Propak* system of interactive routeing, but is better than purely manual routeing from scratch.

There are two other interactive routeing possibilities, firstly rubber-banding the rat lines to produce tracks, or secondly using the pad-to-pad router which works in a very similar way to the one in *Easytrax*. The rubber-banding system is easier to use than some, as it does not use the keyboard to insert corners.

As already mentioned, the autorouter is of the gridded variety and although it offers plenty of choice in pre-run configuration it is not very sophisticated or powerful. There is no rip-up-and-retry strategy for example. It can be configured to route single-sided but the results are typical of an autorouter of this type, with incomplete tracks frequently being reported. The test circuit circuit could not be completed and this puts it in category C.

Traxmaker made a better job of small to medium sized double-sided boards, but as you would expect, considerably more effort was needed to arrange the rat's nest to achieve 100% completion in comparison to category A and B autorouters.

In summary

Firstly, *CircuitMaker*. The schematic drawing program can be recommended. It is easy to learn and user-friendly. With the simulations and large component library, it is very good value for money, and this will make it attractive as a quick and easy simulator as well as a way of generating net list for a pcb routeing program.

Traxmaker has a learning curve of medium steepness. It has logical system of working, some very useful features not found in other programs, and not many of those cryptic dos commands. As a result, it is reasonably pleasant to use.

For manual drawing the package is capable of producing good artwork and is versatile. With its medium-power autorouter in category C, *Traxmaker* is only fair value at around £200 compared to similar products reviewed here. *Ranger2* for example is only £200 for a complete integrated schematic/autorouter system with an autorouter (the 386 rip-up) in category A. Even so, if your interest is only in double-sided or multilayer boards, or if you wish to route manually, then *Traxmaker* may be an attractive proposition.

The snag with *Traxmaker* is that it does not share the same Windows format as *CircuitMaker*. In order to make it a complete schematic/simulation/pcb artwork package, what *Circuitmaker* really needs is a thoroughly integrated Windows based pcb program containing a rip-up-and-retry autorouter to put it on par with other autorouter systems. Then it would have a winning formula.

It is perhaps inevitable that *CircuitMaker* will be compared with *Electronic Workbench*. The latter probably has slightly more realism in its simulations, but both products are such an improvement in user-friendliness on previous budget simulators that it seems churlish to pick this out.

Electronic Workbench has no zoom feature, no library of connectors, and no integrated autorouter as yet; *Circuitmaker* can offer all three. Price may influence purchasers, because in *Electronics Workbench*, the larger library, net list export and Spice in/out are all charged extra. In *Circuitmaker* they are all in the basic price of £199. ■

Feeding the off-centre dipole

It is widely acknowledged that the best place for the off-centre-fed dipole's feed is a third the way along. Richard Formato explains this is not a universal the rule.

Theoretical data suggest that the commonly used feed point for the off-centre-fed dipole, or ocfd, may not be the best. The off-centre-fed dipole is an attractive multiband antenna because it is simple, inexpensive, and requires no antenna tuner. Improving its performance simply by moving the feed point makes the antenna even more attractive. This note illustrates how the feed point influences antenna performance by analysing computer-modelled standing-wave ratio data for three different feed point locations.

The ocfd, shown schematically in Fig. 1, consists of a single wire radiator of length L fed off-centre a distance D from one end. The usual implementation uses a '$^1/_3$-feed', that is, the rf source is located one-third of the way from the end, so that $D \cong L/3$. Why the feed point should be located there is not exactly clear. The 9th edition of the Antenna Book[1], for example, observes that there is not much theoretical justification for this choice. Nevertheless, the $^1/_3$-feed is accepted practice for building an ocfd.

Design details for a $^1/_3$-feed three-band off-centre-fed dipole (80-40-20 meters) appear in the 17th edition of the Antenna Book[2]. A 4:1 current balun at the feed point matches this antenna to any length of 50Ω coax. More recently, Bill Wright, G0FAH[3], described a four-band 1/3-feed off-centre-fed dipole (40-20-15-10 meters) fed with 300Ω ladder line. Matching 50Ω coax requires a 4:1 balun on 40-20-10 meters and a 1:1 balun on 15 meters.

Four band operation therefore requires switching baluns. Another minor limitation is that the ladder line length can be only an odd multiple of the wavelength at 21MHz because the line is used as an impedance transformer. A simpler approach to achieving four-band operation is to feed the off-centre-fed dipole at a different point along its length.

A 21.03m (69ft) long, 0.2053cm diameter (#12 AWG) off-centre-fed dipole was computer-modelled in free space. The dimensions are the same as those in the G0FAH design. Free-space results are a good approximation for antennas high enough above the ground (typically a significant fraction of a wavelength). The band-centre standing-wave ratio was computed on 40-20-15-10 meters at the antenna input terminals for a feed system impedance, Z_o, of 200Ω.

The theoretical values of input resistance and reactance were used to calculate standing-wave ratio: the antenna was not assumed to be tuned. Because Z_o is 200Ω, a 4:1 balun is required to feed the antenna with 50Ω coax. The results for three different feed points appear in Figs 2, 3 and 4.

Figure 2 plots standing-wave ratio at the antenna terminals for the conventional $^1/_3$-feed where D is 6.98m. The 40 and 10 meter values are slightly over 2:1, while the 20 meter standing-wave ratio is about 1.75. In marked con-

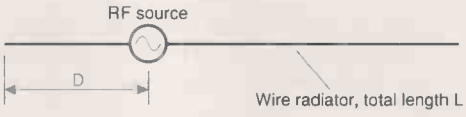


Fig. 1. Commonly, dipole feed distance D is a third of L, but this may not be the best choice.

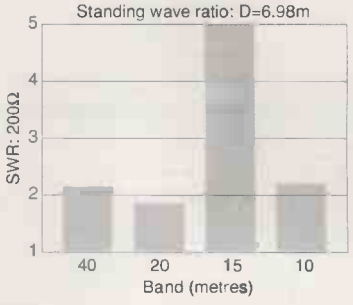


Fig. 2.Standing-wave ratio at the antenna terminals for a standard 1/3-feed dipole.

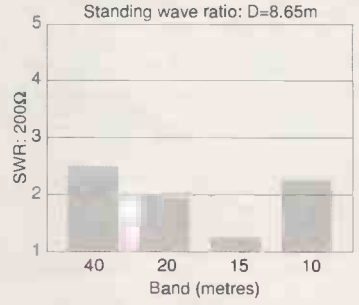


Fig. 3.When the off-centre-fed dipole feed is located at D=6.98m, this standing-wave ratio plot results.

trast, the 15 meter standing-wave ratio is off the scale (actual value >20). It is this behaviour that makes a special feed system necessary on 15 meters, a complication which can be avoided by moving the feed point.

Figure 3 plots standing-wave ratio when the off-centre-fed dipole feed is located 8.65m from one end. The 40-20-10 meter standing-wave ratios are somewhat higher than they are with the $^1/_3$ feed, but the 15 meter ratio is very low at around 1.2. Moving the feed point 1.67m closer to the centre of the antenna results in a much better average standing-wave ratio. And, more importantly, special matching is not required to achieve a standing-wave ratio of 2.5 or less at the antenna terminals on all bands. Balun and coaxial cable losses, which are inevitable, reduce the standing-wave ratio at the coax input to even lower levels. For most installations, it is probably reasonable to expect standing-wave ratio at the transmitter to be less than 2:1 on all bands.

One more example of the effect of feed point location appears in Fig. 4, which plots standing-wave ratio for a feed point 3.65m from one end.

The values on 40-20-15 meters are excellent. The 40m standing-wave ratio is only slightly above 2, and the 20 and 15 meter standing-wave ratios are below 2:1. The highest value occurs on 10m, where it is approxi-

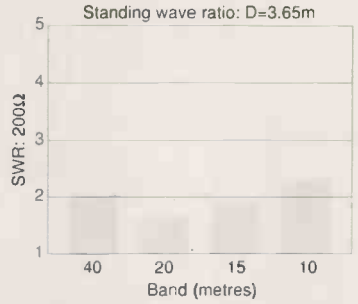


Fig. 4. A further example of how moving the feed point affects standing-wave ratio.

mately 2.4. Since the standing-wave ratio is reduced by feed system losses, it will be less than 2.4 at the coax input. And, because balun and cable losses increase with frequency, the standing-wave ratio reduction will be greatest on 10m where it is needed most. Feeding the antenna 3.65m from one end may well provide the best overall four-band performance.

In a specific implementation, the off-centre-fed dipole, like any antenna, must be 'tweaked' for optimum standing-wave ratio. This is accomplished by adjusting the feed point location. Other antennas, nearby metallic objects, and the earth are typical factors that influence antenna performance. Since these factors are not included in the computer model, they must be dealt with empirically by adjusting the antenna on-site. The data presented here provide a starting point for experimenting with different feed points.

Depending on the total antenna length L, height above ground, earth electrical parameters, and feed system Z_o, it should be possible to operate a single off-centre-fed dipole on four or more bands without an antenna tuner or special feed arrangement.

It is apparent is that the ofcd's standing-wave ratio varies dramatically as the feed point is moved, and that the commonly used $^1/_3$-feed is not necessarily the best. Other feed points may therefore produce a better antenna. ■

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3. Bill Wright, Four Bands, Off Centre, *QST Magazine* (American Radio Relay League, Inc., Newington, CT, USA), February 1996, page 65.

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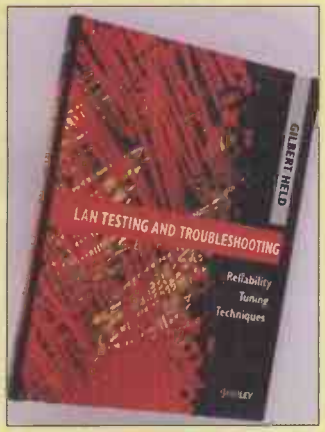

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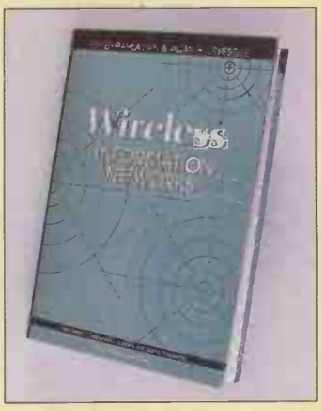

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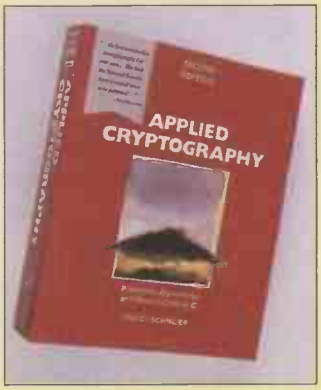

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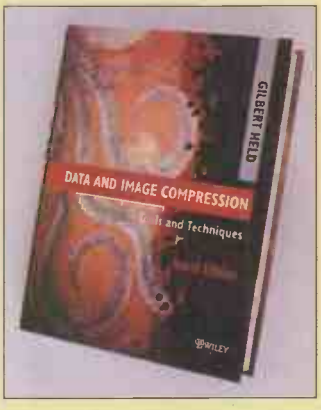

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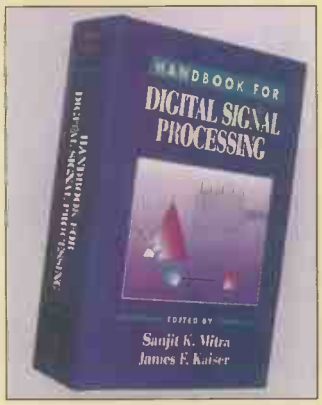

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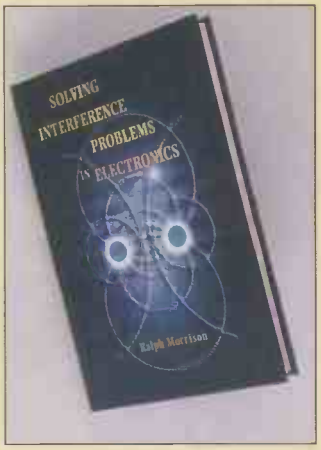

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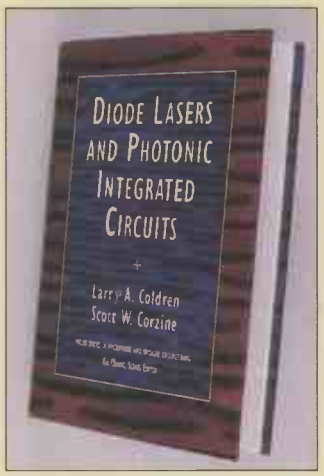

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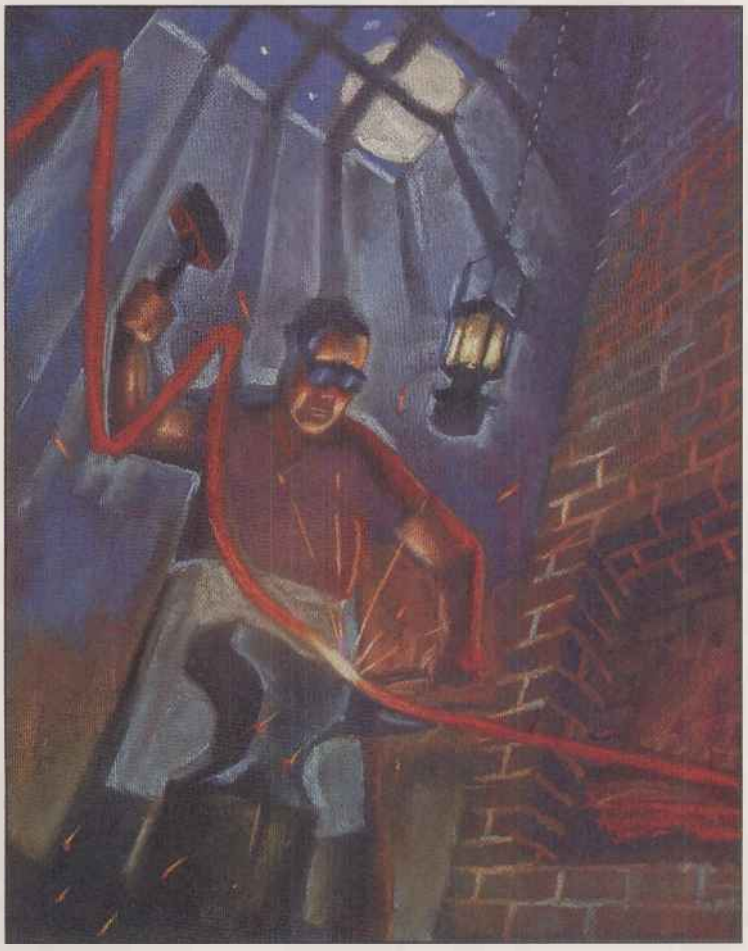
Night thoughts on crossover distortion

Douglas Self

investigates the biggest factor affecting signal purity in Class-B audio power amplifiers – crossover distortion.



"Weigh everything; and trust no man." Vannoccio Biringuccio, pioneer metallurgist, c. 1520 AD.

It is universally acknowledged that crossover distortion is the worst problem afflicting Class-B power amplifiers.

Those who have followed my investigations into amplifier distortion will recall my concept of a 'Blameless' Class-B amplifier – one so designed that the easily correctable distortions are reduced to negligible levels. This yields an amplifier where crossover distortion is the only non-linearity detectable when driving an impedance of 8Ω or greater, and even this is only measurable above 2kHz or so.

Such an amplifier typically gives a total-harmonic-distortion plot such as **Fig. 1**, where the thd is less than 0.001% from 10Hz to 1kHz, and only reaches 0.005% at 10kHz. The thd rises at 6dB/octave and emerges from the noise floor around 1kHz, firstly because the global negative feedback (nfb) has been made to fall at 6dB/octave for high-frequency stability, and secondly because crossover distortion is high-order, and so its harmonics are at high frequencies where the negative feedback factor is small.

The state of Blamelessness (an inelegant term perhaps, but no-one has yet come up with a better one) holds for 8Ω loads, but 4Ω loading introduces an extra third-order distortion due to current-dependent beta in the output devices[1].

The pernicious nature of crossover distortion is partly due to the fact that it occurs over a small part of the signal swing, and so generates high-order harmonics. Worse still, the small range over which it occurs is at the zero-crossing point. Not only is it present at all levels and all but the lightest loads, but is generally believed to increase as output level falls, having the potential to cause very poor linearity at the modest listening powers that most people use.

Seeing is believing

Being an untrusting person, I first looked to see if crossover distortion really did increase with decreasing output level in a Blameless amplifier. The problem is that a Blameless amplifier has

such a low level of distortion at 1kHz – 0.001% or less – that the crossover artefacts are barely visible in circuit noise. This holds even if low-noise techniques are used[2].

The measured percentage level of the noise-plus-distortion residual is bound to rise with falling output, because the noise voltage remains constant; this is the lowest line in **Fig. 2**. To circumvent this, the amplifier was deliberately underbiased by varying amounts to generate ample crossover spikes, on the assumption that any correctly adjusted amplifier should be less barbarous than this.

The answer from Fig. 2 is that the thd percentage does increase as level falls, but relatively slowly. Both emitter-follower and complementary-feedback-pair output stages give similar diagrams to Fig. 2. Whatever the degree of underbias, thd increases by about 1.6 times as the output voltage is halved. In other words, reducing the output power from 25W to 250mW, which is pretty drastic, only increases thd percentage by six times. This makes it clear that the absolute, as opposed to percentage, thd level in fact falls slowly with amplitude, and therefore probably remains imperceptible. This is something of a relief; but crossover distortion remains a bad thing to have.

Distortion versus level was also investigated at high frequencies, ie above 1kHz where there is more thd to measure, and optimal biasing can be used. **Figure 3** shows the variation of thd with level for the emitter-follower stage at a selection of frequencies; **Fig. 4** shows the same for the complementary feedback pair. Neither shows a significant rise in percentage thd with falling level, though it is noticeable that the emitter follower gives a good deal less distortion at lower power levels around 1W. This is an unexpected observation, and possibly a new one.

As a final look at the nature of the beast, **Fig. 5** shows that high-frequency distortion is markedly reduced by increasing the load resistance. This provides further confirmation that almost all the 8Ω distortion originates as crossover in the output stage.

Minimising crossover distortion

Unlike some more benign kinds of signal-warping, crossover distortion seems to be unanimously agreed to be something any amplifier could well do without. I therefore scrutinised some output stages to find ways to reduce its production.

The amount of crossover distortion produced depends strongly on optimal quiescent adjustment, so the thermal compensation used to stabilise this against changes in temperature and power dissipation must be accurate. The first part of this article deals with the crossover region and its quiescent conditions, the second with temperature effects on these conditions. Both reveal surprises...

The output stage examined

Fig. 6 shows the two most common types of output stage: the emitter follower and the complementary-feedback-pair configurations. The manifold types of output stage based on

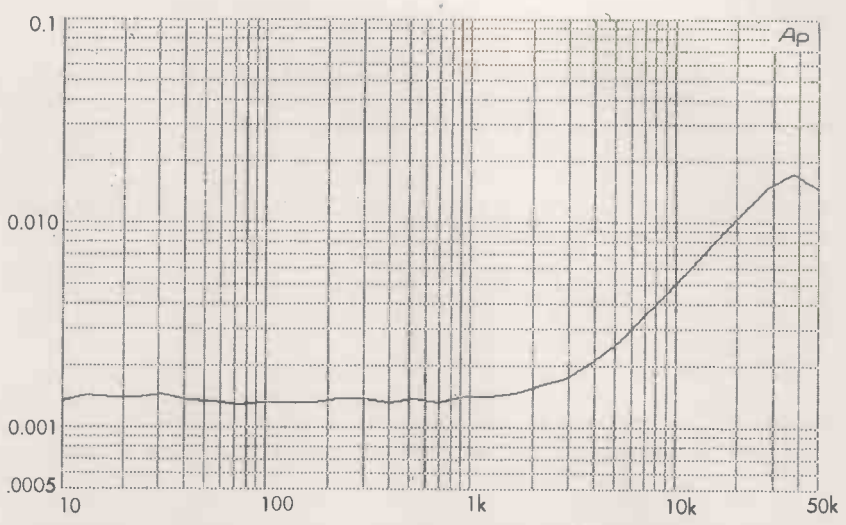


Fig 1. Distortion performance, percentage thd, of a typical Blameless Class-B amplifier at 25W/8Ω; the noise floor is at the 0.0008% level. Bandwidth is 80kHz.

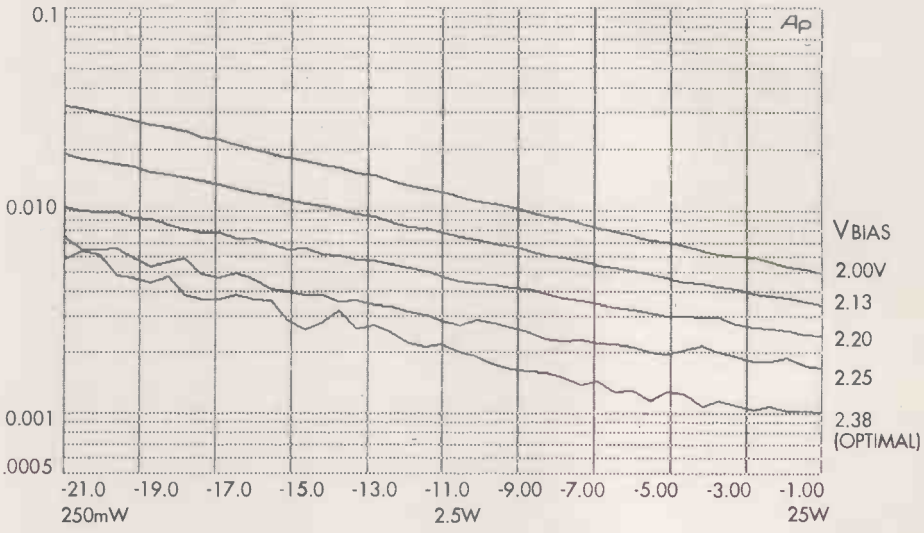


Fig 2. Showing how crossover distortion rises slowly as output power is reduced from 25W to 250mW (8Ω) for optimal bias and increasingly severe underbias (upper lines) This is an emitter-follower type output stage. Measurement bandwidth 22kHz.

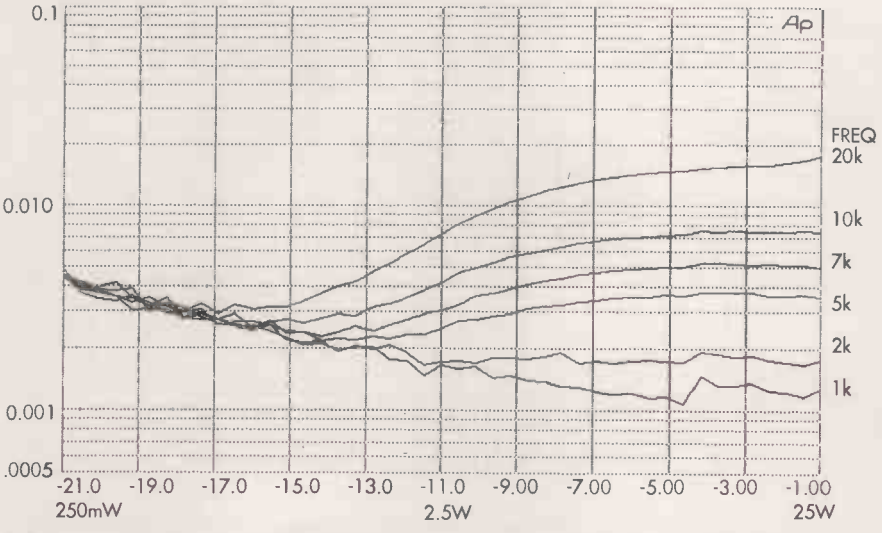


Fig 3. Variation of crossover distortion with output level for higher frequencies. Optimally biased emitter-follower output stage. Bandwidth 80kHz.

triples will have to be set aside for the moment. The two circuits shown have few components, and there are equally few variables to explore in attempting to reduce crossover distortion.

To get the terminology straight: here, as in my previous writings, V_{bias} refers to the voltage set up across the driver bases by the V_{be}-multiplier bias generator. For Class-B operation, V_{bias} is in the range 1-3V. Voltage V_q is the quiescent voltage across the two emitter resistors (hereafter R_e) alone, and is between 5 and 50mV, depending on the configuration chosen. Quiescent current I_q refers only to that flowing in the output devices, and does not include driver standing currents.

I have already shown that the two most common output configurations are quite different in behaviour, with the complementary feedback pair being superior on most criteria. **Table 1** shows that crossover gain variation for the emitter-follower stage is smoother, – being some 20 times wider – but of four times higher amplitude than for the complementary feedback pair version. It is not immediately obvious from this which stage will generate the least high-frequency thd, bearing in mind that the negative feedback factor falls with frequency.

Table 1 also emphasises that a little-known drawback of the emitter follower is that its quiescent dissipation may be significant.

An experiment with crossover

Looking hard at the two output stage circuit diagrams, intuition suggests that the value of emitter resistor R_e is worth experimenting with. Since these two resistors are placed between the output devices, and alternately pass the full load current, it seems possible that their value could be critical in mediating the hand over of output control from one device to the other. Resistor R_e was therefore stepped from 0.1 to 0.47Ω, which covers the practical range. Voltage V_{bias} was re-optimised at each step, though the changes were very small, especially for the complementary feedback pair version.

Figure 7 shows the resulting gain variations in the crossover region for the emitter-follower stage, while **Fig. 8** shows the same for the complementary feedback configuration. **Table 2** summarises some numerical results for the emitter-follower stage, and **Table 3** for the complementary feedback.

There are some obvious features; firstly R_e is clearly not critical in value as the gain changes in the crossover region are relatively minor. Reducing R_e allows the average gain to approach unity more closely, with a consequent advantage in output power capability[3]. Similarly, reducing R_e widens the crossover region for a constant load resistance, because more current must pass through one R_e to generate enough voltage-drop to turn off the other output device.

This implies that as R_e is reduced, the crossover products become lower-order and so of lower frequency. They should be better linearised by the frequency-dependent global negative feedback, and so overall closed-loop high-frequency thd should be lower.

The simulated crossover distortion experiment described in reference 4 showed that as the crossover region was made narrower, the distortion energy became more evenly spread over higher harmonics. A wider crossover region implies energy more concentrated in the lower harmonics, which will receive the benefit of more negative feedback. However, if the region is made wider, but retains the same amount of gain deviation, it seems likely that the total harmonic energy is greater. Consequently, there are two opposing effects to be considered.

This is partly confirmed by reference 2, where measurements show that the thd reaches a very shallow minimum for R_e=0.22Ω, at any rate for that particular configuration, level, and load; this is consistent with two opposing effects. While the variation of thd with R_e appears to be real, it is small, and I conclude that selecting R_e=0.1Ω for maximum efficiency is probably the over-riding consideration. This has the additional benefit that if the stage is erroneously over-biased into Class AB, the resulting g_m-doubling distortion will only be half as bad as if the more usual 0.22Ω values had been used for R_e[3].

Never assume

It would be easy to assume that higher values of R_e must be more linear, because of a vague feeling that 'there is more local feedback'. But this cannot be true as an emitter-follower already has 100% voltage feedback to its emitter, by definition. Changing the value of R_e alters slightly the total resistive load seen by the emitter itself, and this does seem to have a small but measurable effect on linearity.

The first surprise from this experiment is that in the typical Class-B output stage, qui-

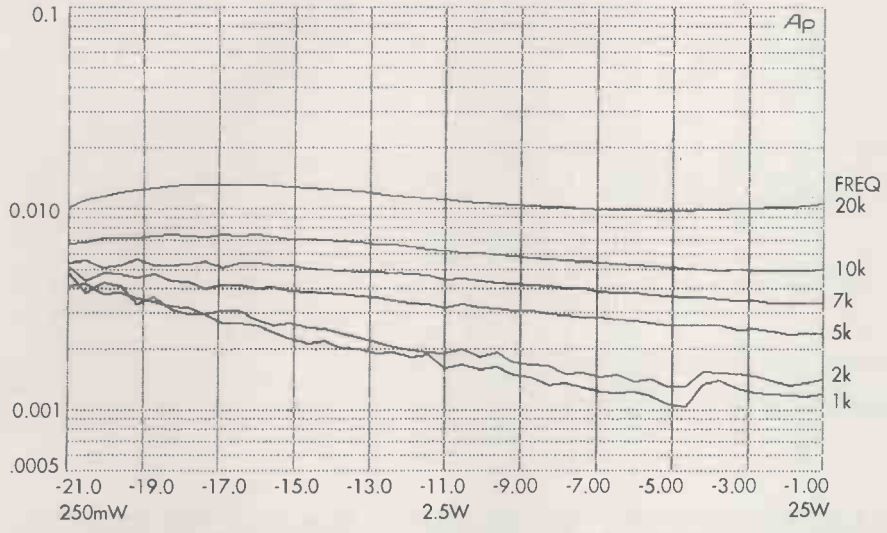


Fig 4. Variation of distortion with level for higher frequencies. Optimally biased CFP output stage. Bandwidth 80kHz.

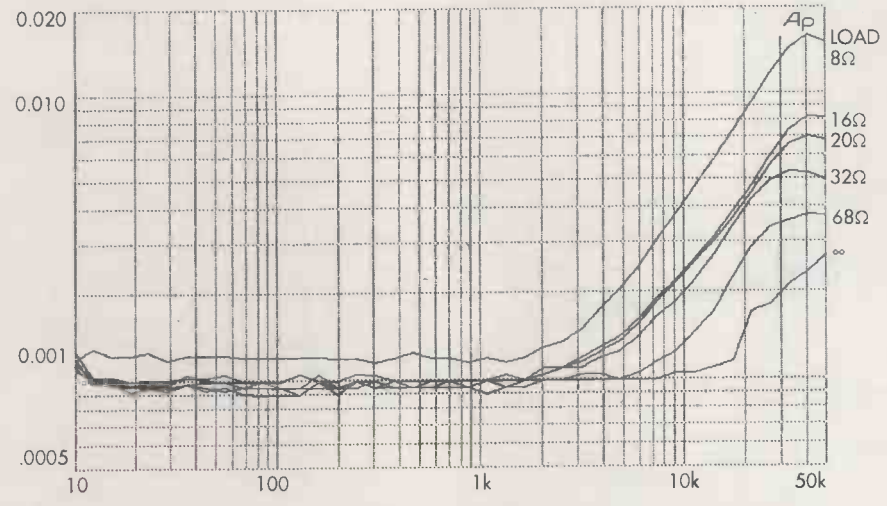


Fig 5. How crossover distortion is reduced with increasing load resistance. Power is 20W into 8Ω and bandwidth is 80kHz.

Table 1. Crossover gain variation for the emitter follower is wider, thus smoother.

	Emitter-follower	CFP
V_{bias}	2.930V	1.297V
V_q	50mV	5mV
I_q	114mA	11mA
P_q (per o/p device)	4.6W	0.44W
Average Gain	0.968	0.971
Peak gain deviation from average	0.48%	0.13%
Crossover width*	±12V	±0.6V

(for R_e=0R22, 8Ω load and ±40V supply rails)

* Crossover-width is the central region of the output voltage range over which crossover effects are significant; I have rather arbitrarily defined it as the ± output range over which the incremental gain curves diverge by more than .0005 when V_{bias} is altered around the optimum value. This is evaluated here for an 8Ω load only.

Table 2. Characteristics of the emitter-follower stage (Type 1).

Data for 8Ω load and emitter-follower o/p stage OUTEF2G.CIR (See HC #2299-304)

R_e	Optimal V_{bias}	Optimal V_q	I_q mA	X-Width	Average Gain
Ω	V	mV	mA	V	ratio
0.1	2.86	42.6	215	18	0.982
0.22	2.87	46.2	107	12	0.968
0.33	2.89	47.6	74	9	0.955
0.47	2.93	54.8	59	7	0.939

As R_e is varied, V_q varies by only 29%, while I_q varies by 365%

Table 3. Complementary feedback pair characteristics.

Data for 8Ω load and cfp o/p stage OUTPUT4G.CIR (See HC #2293-8)

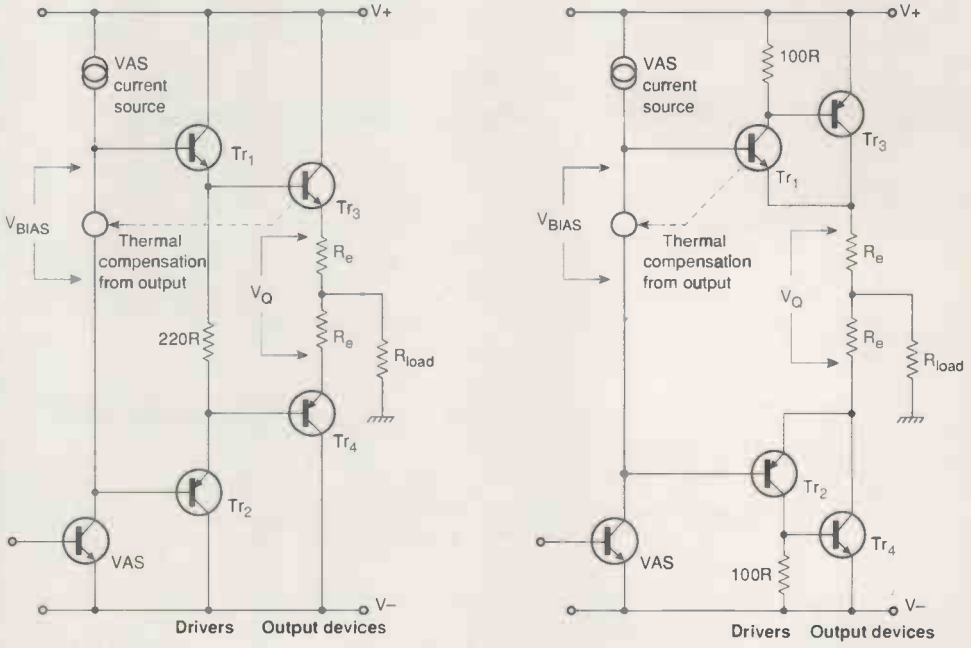
R_e	Optimal V_{bias}	Optimal V_q	I_q	X-Width	Average Gain
Ω	V	mV	mA	V	ratio
0.1	1.297	3.06	15.3	1.0	0.983
0.22	1.297	4.62	11.5	0.62	0.971
0.33	1.297	5.64	8.54	0.40	0.956
0.47	1.298	7.18	7.64	0.29	0.941

As R_e is varied, V_q varies by 230% while I_q varies by 85%. However the absolute V_q change is only 4mV, while the sum of V_{be}s varies by only 0.23%. This makes it pretty plain that the voltage domain is what counts, rather than the absolute value of I_q.

Table 4. Tolerance of Vbias for 8Ω loading.

		Follower o/p	CFP output
Crossover spikes obvious	Underbias	2.25V	1.242V
Spikes just visible	Underbias	2.29	1.258
Optimal residual	Optimal	2.38	1.283
g_m-doubling just visible	Overbias	2.50	1.291
g_m-doubling obvious	Overbias	2.76	1.330

Fig 6. The two most popular kinds of output stage: the emitter-follower, left, and complementary feedback pair. V_{bias} and V_q are identified.



escent current as such does not matter a great deal. This may be hard to believe, particularly after my repeated statements that quiescent conditions are critical in Class-B, but both assertions are true. The data for both the emitter follower and complementary feedback pair output stages show that changing R_e alters the I_q considerably, but the optimal value of V_{bias} and V_q barely change.

Voltage across the transistor base-emitter junctions and emitter resistors seems to be what counts, and the actual value of current flowing as a result is not in itself of much interest. However, the V_{bias} setting remains critical for minimum distortion; once the R_e value is settled at the design stage, the adjustment procedure for optimal crossover is just as before.

The irrelevance of quiescent current was confirmed in the Trimodal amplifier[3], which was actually designed after the work in this article was done, and where I found that changing the output emitter resistor value R_e over a 5:1 range required no alteration in V_{bias} to maintain optimal crossover conditions.

The critical factor is therefore the voltages across the various components in the output stage. Output stages get hot. When junction temperatures change, both experiment and simulation show that if V_{bias} is altered to maintain optimal crossover, V_q remains virtually constant.

This confirms the task of thermal compensation is solely to cancel out the V_{be} changes in the transistors; this may appear to be a blindingly obvious, but it was worth checking as there is no inherent reason why the optimal V_q should not be a function of device temperature. Fortunately it isn't, for thermal compensation that also dealt with a need for V_q to change with temperature might be a good deal more complex.

The recognition that V_q is the critical parameter has some interesting implications. Can we

immediately start setting up amplifiers for optimal crossover with a cheap digital voltmeter rather than an expensive thd analyser? Setting up quiescent current with a milliammeter has often been advocated, but the direct measurement of this current is not easy. It requires breaking the output circuit so a meter can be inserted, and not all amplifiers react favourably to so rude an intrusion. The amplifier must also have near-zero dc offset voltage to get any accuracy.

Measuring the total amplifier consumption is not acceptable because the standing-current taken by the small-signal and driver sections will, in the complementary feedback pair case at least, swamp the quiescent current. It is possible to determine quiescent current indirectly from the V_q drop across the emitter resistors – still assuming zero dc offset – but this can never give a very accurate current reading as the tolerance of a low-value for R_e is unlikely to be better than ±10%.

However, if V_q is the real quantity we need to get at, then R_e tolerances can be blissfully ignored. This does not make thd analysers obsolete overnight. It would be first necessary to show that V_q was always a reliable indicator of crossover setting, no matter what variations occurred in driver or output transistor parameters. This would be a sizable undertaking.

There is also the difficulty that real-life dc offsets are not zero, though this could possibly be side-stepped by measuring V_q with the no load. A final objection is that without thd analysis and visual examination of the residual, you can never be sure an amplifier is free from parasitic oscillations and working properly.

I have previously demonstrated that the distortion behaviour of a typical amplifier is quite different when driving 4Ω rather than 8Ω loads. This is because with the heavier load, the output stage gain-behaviour tends to be dominated by beta-loss in the output devices at higher currents, and consequent extra loading on the drivers, giving third-harmonic distortion. If this is to be reduced, which may be well worthwhile as many loudspeaker loads have serious impedance dips, then it will need to be tackled in a completely different way from crossover distortion.

It is disappointing to find that no manipulation of output-stage component values appears to significantly improve crossover distortion, but apart from this one small piece of (negative) information gained, we have in addition determined that:

- Quiescent current as such does not matter; V_q is the vital quantity.

- A perfect thermal compensation scheme, that was able to maintain V_q at exactly the correct value, requires no more information than the junction temperatures of the driver and output devices. Regrettably none of these temperatures are actually accessible, but at least we know what to aim for.

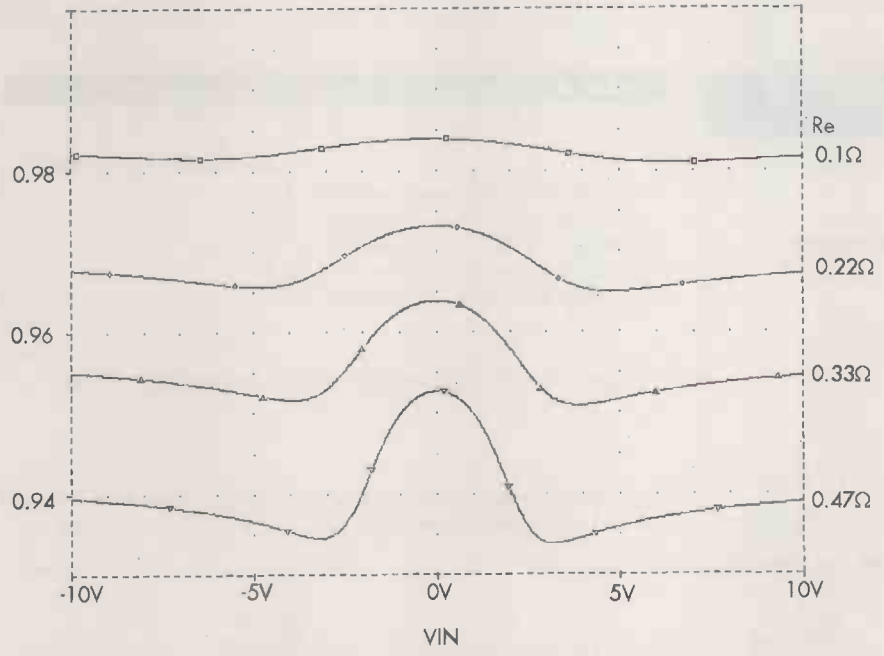


Fig 7. Output linearity of emitter-follower output stage for R_e between 0.1 and 0.47Ω.

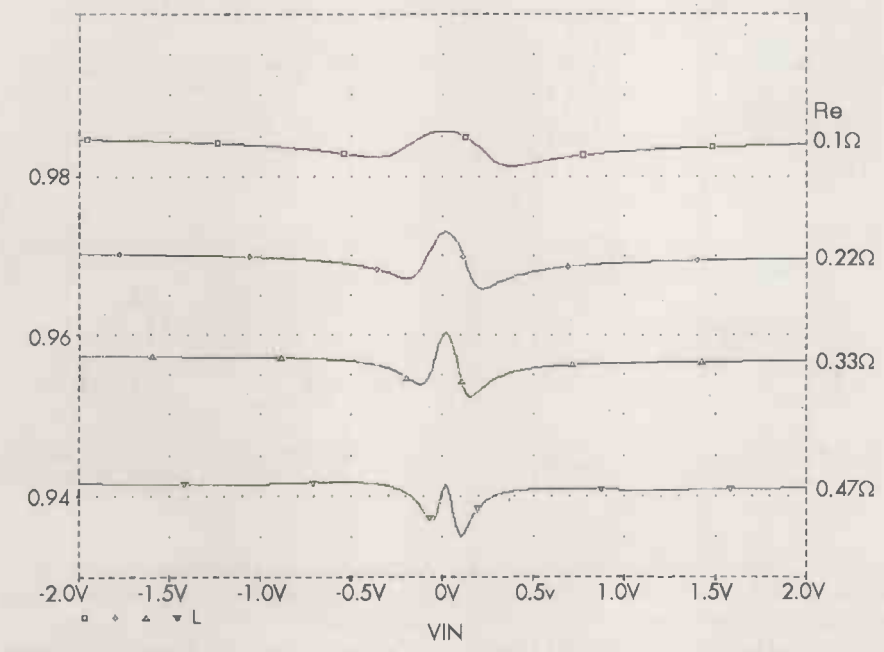


Fig 8. Output linearity of the cfp output stage for emitter-resistance R_e between 0.1 and 0.47Ω.

Thermal issues

Quiescent condition stability depends on two main factors. The first is the stability of the V_{bias} generator in the face of external perturbations, such as supply voltage variations. The second – and more important – is the effect of temperature changes in the drivers and output devices, and the accuracy with which V_{bias} can cancel them out.

From the above investigations, and given a fixed R_e, V_{bias} must cancel out temperature-induced changes in the voltage across the transistor base-emitter junctions, so that V_q remains constant. From the limited viewpoint of thermal compensation this is very much the same as the traditional criterion that the quiescent current must remain constant, and no relaxation in exactitude is permissible.

I have at last reached some conclusions on how accurate the V_{bias} setting must be for minimal distortion, after many hours squinting at furry green scope traces. The results are approximate, depending partly on visual assessment of a noisy residual signal, and will probably change slightly with transistor type.

Nonetheless, **Table 4** gives a starting point.

From these, er, subjective measurements, we can take the permissible error band for the emitter-follower stage as about ±100mV, and for the complementary-feedback pair as about ±10mV. This goes some way to explaining why the emitter-follower stage can give satisfactory quiescent stability despite its dependence on the V_{be} of hot power transistors.

Simulation

Returning to the PSpice simulator, and taking R_e=0.1Ω, a quick check on how the various transistor junction temperatures affect V_q yields:

- The emitter-follower output stage has a V_q of 42mV, with a V_q sensitivity of –2mV/°C to driver temperature, and –2mV/°C to output junction temperature. No surprises here.

- The complementary-feedback pair stage has a much smaller V_q of 3.1mV. Sensitivity of V_q is –2mV/°C to driver temperature, and only –0.1mV/°C to output device temperature. This confirms that local negative feedback in the stage makes V_q relatively independent of output device temperature, which is just as well as **Table 4** shows it needs to be about ten times more accurate.

The complementary feedback pair output devices are about 20 times less sensitive to junction temperature, but the V_q across R_e is something like 10 times less; hence the actual relationship between output junction temperature and crossover distortion is not so very different for the two configurations, indicating that as regards temperature stability the complementary feedback pair is only twice as good as the emitter follower, and not vastly better, which is perhaps the common assumption.

In real life, with a continuously varying power output, the situation is complicated by the different dissipation characteristics of the drivers as output varies. See **Fig. 9**, which shows that the complementary feedback pair driver dissipation is more variable with output, but on average runs cooler.

For both configurations, driver temperature is equally important, but the emitter-follower driver dissipation does not vary much with output power. Initial drift at switch-on is however greater, as the standing dissipation is higher. This, combined with the two times greater sensitivity to output device temperature and the greater self-heating of the emitter-follower output devices, may be the real reason why most designers have a general feeling that the emitter-follower version has inferior quiescent stability.

Having assimilated this, we can speculate on the ideal thermal compensation system for the two output configurations. The emitter-follower stage has V_q set by the subtraction of four dissimilar base-emitter junctions from V_{bias}, all having an equal say, and so all four junction temperatures ought to be factored into the final result. This would certainly be comprehensive, but four temperature-sensors per channel is perhaps overdoing it. For the complementary feedback pair stage, we can ignore the output device temperatures and only sense the drivers, which simplifies things and works well in practice.

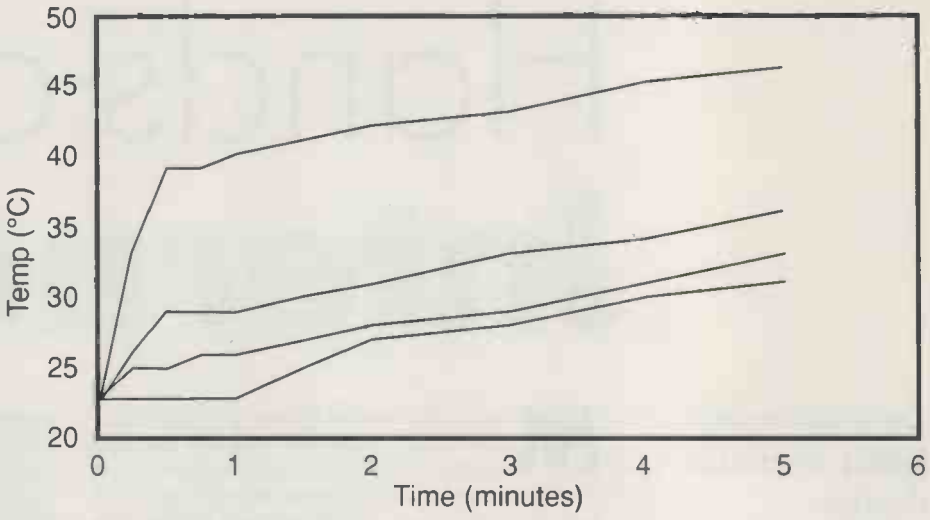


Fig 9. Driver dissipation versus output level. In all variations on the emitter-follower configuration, power dissipation varies little with output; complementary-feedback-pair driver power however varies by a factor of two or more. (This is Fig. 1 taken from Reference 5.)

If you assume that the drivers and outputs come in complementary pairs with similar V_{be} behaviour, then symmetry prevails and we need only consider one half of the output stage, so long as V_{bias} is halved to suit. This assumes that the audio signal is symmetrical over time scales of seconds to minutes, so that equal dissipations and temperature rises occur in the top and bottom halves of the output stage. This seems a safe bet, but the unaccompanied human voice has positive and negative peak values that may differ by up to 8dB, so prolonged *a cappella* performances have at least the potential to mislead any compensator that assumes symmetry.

In practice

Practical amplifiers of whatever output configuration almost invariably simplify matters to the ultimate by using only one sensor to establish V_{bias}, usually in a V_{be}-multiplier circuit. Temperature sensed is therefore at best a compromise, and the best sensor position depends crucially on the configuration chosen.

For the emitter-follower, both drivers and outputs have an equal influence on quiescent V_q, but the output devices normally get much hotter than the drivers, and their dissipation varies much more with output level. In this case the sensor goes on or near one of the output devices, thermally close to the output junction.

It has already been shown experimentally that the top of the *TO3* can is the best place to put it[5]. Recent experiments have confirmed that this holds true also for the *TO3P* package, (a large plastic package like an enlarged *TO220*, and nothing like *TO3*) which can easily get 20° hotter on its upper plastic surface than does the underlying heatsink.

In the complementary feedback pair, the drivers have most effect and the output devices, although still hot, have only one-twentieth the influence. Driver dissipation is also much more variable, so now the correct place to put the thermal sensor is as near to the driver junction as you can get it.

The temperature sensors discussed here are physically distant from the driver junction, so thermal attenuation and delay errors complicate the situation considerably. In a future article I hope to show how these errors can be determined, and markedly reduced, by improving the thermal compensation system. ■

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Hands-on Internet

Cyril Bateman

looks at circuit design simulators – including alternatives for rf engineering.

Much of the value of Internet lies in its explosive growth combined with the ease and speed at which pages can be updated. Unlike the delays implicit with conventional publication, it is quite feasible to draft a Web page and publish it on Internet – all within one day. Hence site content and its essential 'URL' address frequently change.

With the exponentially growing number of Web pages now accessible on Internet being matched by the equally rapid introduction of new search facilities, structured search patterns are desired. This need is shared equally by electronic designers and students, so it should come as no surprise that some excellently coordinated search facilities are found at university (.edu) locations.

The University of Nebraska-Lincoln[1] Electronics Shop Web Page reveals a wealth of information targeted to its students, but almost equally useful to designer engineers. Two documents should be printed out for reference, 'Electronic Design Software' and 'E.E. Internet Info Sites', in total fifteen pages giving access to preferred search engines, design software reviews, FTP and information resources.

As a contrast in styles, within two printed pages, CECI[2] publish their famed 'Electronics Search FAQ' which provides links to nine Net search engines – eight not previously covered in this series, All-in-One, Find-it, Gower.Net, Internet Sleuth, nLightn, Rice University, Search.com, Use It, Yahoo, **Fig. 1**.

This month 'All-in-One'[3], which clearly demonstrates the volume of search methods available, is chosen as a bookmark site. Its search page is conveniently organised into eleven major search categories. These present links to more than 200 search engines, all using a consistent user interface, **Fig. 2**.

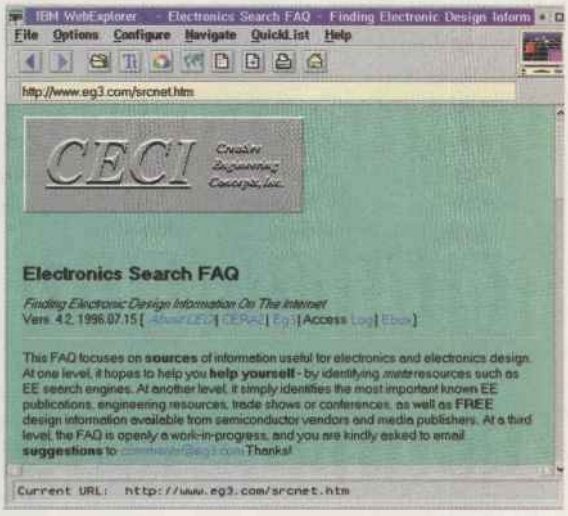

Fig. 1. Creative Engineering provides links to nine search tools. For the present this is still supplied as a no-cost service.

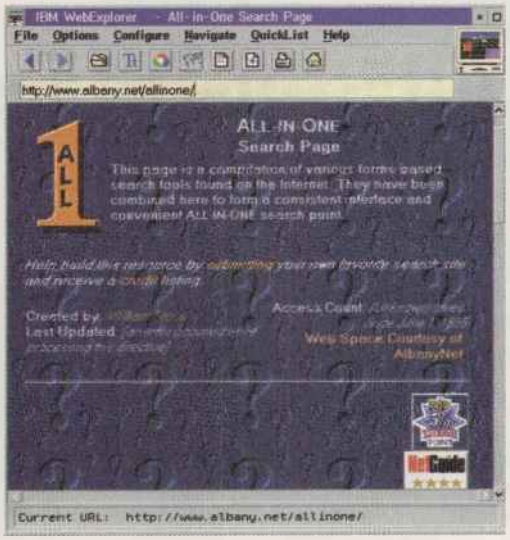

Fig. 2. All-in-One provides easy access to over 200 search engines. These are subdivided for convenience into major sub-headings, all with a consistent user interface. You can even track a UPS parcel delivery using this source.

While 'Archie' remains the preferred Internet system for locating and downloading software programs using the many FTP sites, Archie accesses Unix based archives, so cannot find all files. A special search engine 'FTPSearch'[4] located at Trondheim, Norway can often be more effective, but it is better restricted to simply finding files, rather than finding and downloading.

With the variety of search engines now identified, Web users will adopt their personal favourite search site list. To save much searching for and typing in of frequently incomprehensible URL addresses, ones personal selection can be stored or 'Bookmarked' within the Web browser to be immediately available, saving much on-line time.

Simulation software

With the prevailing 'time to market' pressures, most development circuits need simulation before committing to a breadboard.

Circuit simulation software for low frequencies, also for digital circuits, is dominated by Spice based systems working principally in 'Time Domain'. In the first of this series, March 1996, I demonstrated, using FTP, how to download an evaluation version of the popular

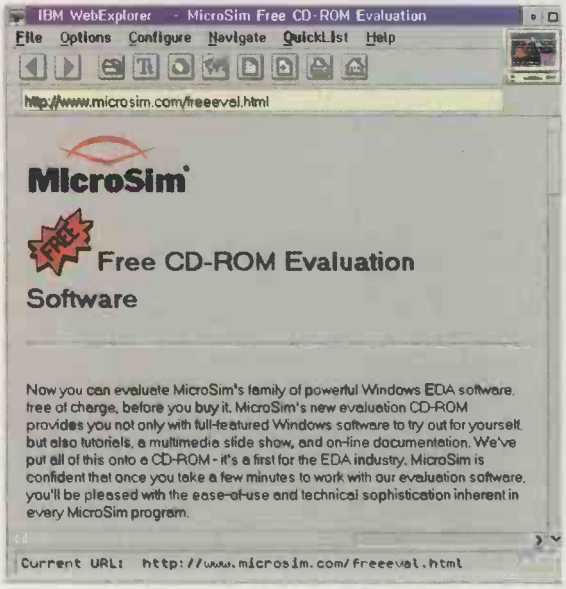

Fig. 3. The lowest cost way to evaluate the PSpice software. Can also be ordered by phone or fax.

Fig. 4. Why not try out the demo for this low cost alternative simulator? This package is currently on a very special offer.

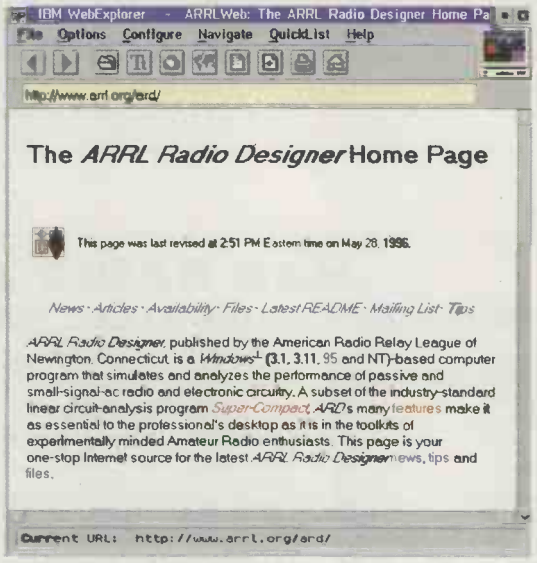

Fig. 5. This excellent low-cost introduction to rf simulators is available to non-members. My copy, ordered by phone and credit card arrived 16 July – eight weeks after ordering.

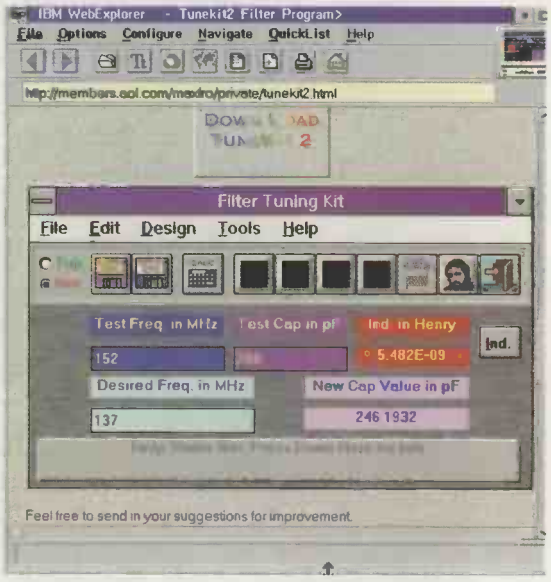

Fig. 6. The latest ToolKit v2.5 offers extra features but download v2 first. Offers easy Net-List generation for ARD, provided it matches your needs.

PSpice simulator. The current Windows evaluation package totals almost 12Mbyte, so Microsim[5], now offer this on cd rom, in addition to their FTP download site. This cd can be ordered from their Web page while on-line, by E-mail or telephone, **Fig. 3**.

While PSpice is an extremely popular package, it is expensive, so it might be beneficial to also evaluate demonstration versions of some of the many less expensive offerings. One such, highlighted in the 'Electronic Design Software' paper, is 'TurboSim'[6], which presently is on special offer for $99. It has a 1.1Mbyte demonstration which is easily downloaded, **Fig. 4**.

At radio frequencies, frequency-domain simulation dominates, with 'Touchstone'[7], and 'Super Compact'[8], both being professionally accepted. The American Radio Relay League[9], now offers *The ARRL Designer Software* v1.5 – a sub-set of *Super-Compact* at the extremely attractive price of $150 including an excellent 400 page manual and model libraries. Allow two months for surface shipment, **Fig. 5**.

A shareware 'front end' add on for *ARRL Designer*, called *TuneKit*[10], designed to generate Net Lists and expedite design of signal handling filters, has been written by Max Froding, using Visual Basic. This too can be downloaded, see **Fig. 6**.

Another low-cost frequency-domain simulator, from the 'Electronic Design Software' paper, is the Academic Technologies NSW Australia *RF* system[11]. A demonstration version is available for your evaluation and the full package lists for only $99 US.

Competing head-on with *Touchstone* and *SuperStar Professional*[12], Optotek Ltd offers *Mmicad* v2.0[13] – a mid-cost frequency domain system with increased accuracy for ceramic multilayer capacitor simulation. Incorporating the *CapCad* software enhancement from Dielectric Laboratories Inc. it provides true distributed capacitor models. A demonstration cd, and textual comparisons with *Touchstone* and *SuperStar*, can be ordered on-line for this system.

Designed to ease the problem of adequately modelling ceramic multilayer capacitors in Spice simulations, *Spicap* is an interactive on-line tool on the AVX Corporation Web

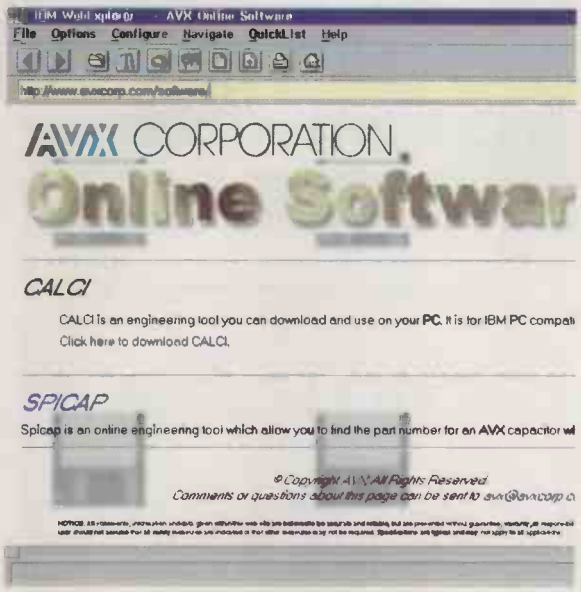

Fig. 7. SpiCap provides easy interpretation of this maker's published capacitor data. Ensures improved accuracy of capacitor modelling using Spice simulators.

page[14]. It provides users with those Spice equivalent circuit parameters of impedance, esr, series inductance, series resonant frequency and effective capacitance – all as functions of applied frequency, temperature and dc voltage. Given either a single frequency or a restricted frequency-band simulation, Spice based simulators can now make allowance for frequency dependent capacitor variables, thus offering more realistic simulations. These same parameters could also be used with frequency-domain simulators for rf, thus reducing the need for 'S Parameter' information, Fig. 7.

Readers interested in further exploring the merits of time domain versus frequency domain simulators, will find a wealth of unbiased and detailed discussion papers, including the latest *Circuit Envelope* simulator which combines both techniques, at the HP *EEsof* home page[7] – the home of the Touchstone rf software simulation system. ■

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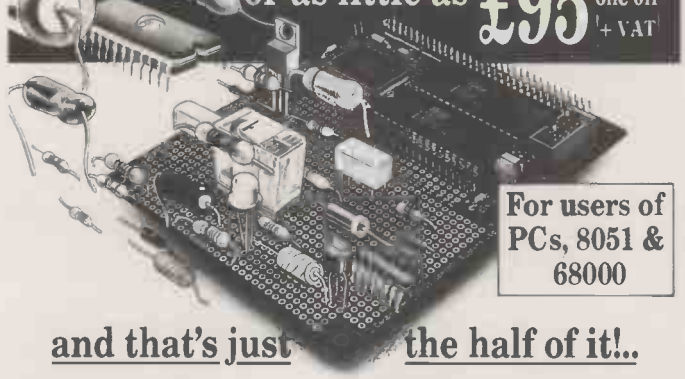

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LETTERS

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

Motional feedback headphones explained

Ian Hickman's motional-feedback headphones – Music in Mind, *EW*, October '96 – are an ingenious piece of design work, but it was inevitable that, for a variety of reasons, they would fail to produce the desired results.

There are three basic cues available in a sound signal to indicate the source direction. As Ian indicates, one of these is the disparity in time of arrival at the two ears – which will be zero when the source lies on the mid line. However, this cue is only accessible in sounds which possess distinct transients; most natural sounds do, but Ian's choice of a continuous sinewave for initial testing would have eliminated this potential source of information.

What would have been available was the phase difference between the ears, with one sampling 'further up the wave' than the other. Fortunately Ian had selected a rather low frequency test tone, so he heard the desired effect. As the wavelength shortens to about head size the phase difference represents an ambiguous cue and we seem to have evolved so as to be insensitive to interaural phase differences in frequencies above about 1500 Hz. At shorter wavelengths, diffraction around the head no longer occurs. Instead the head starts to become a significant obstacle in the wave path, so that the more distant ear receives a lower amplitude wave. The interaural intensity difference thus provides a useful direction cue in the region of the spectrum where phase differences can no longer be utilised. In the borderline region of frequencies we are not very effective direction judges, a fact capitalised upon by evolution. For example, ground-living, vulnerable pheasant chicks emit chirps of a frequency which their small-headed mother can easily locate. The broader headed fox in contrast has great difficulty in tracking down the sound source.

If appropriate phase and intensity cues are provided in headphone-delivered sounds a sense of direction is achieved, although, as Ian Hickman indicates, the sounds tend to be 'lateralised' (move from side-to-side, but still in the head) rather than 'localised' (perceived as being external). However, he was misinformed when told that the significant factor in externalisation was the effect of head movement. It is certainly important, but even more so are the effects of the pinnae (the outer ear flaps).

The most notable feature of ears is the intricate pattern of folds. They are not there to provide rigidity – they serve to modify the sound. Rather than delivering a single version of each wave, the folds in our ears produce multiple reflections, each delayed by a very brief interval. The delay is far too short to be perceived as an echo; instead the time delayed replications cause interference in the wave, with diffraction reducing the amplitude of some frequencies and enhancing others. The pinnae thus behave as comb filters, but with notch frequencies which depend upon the angle of incidence of the source. The colouration of the sound (equivalent to the rainbow colours produced by interference in an oil film) gives a unique indication of direction and a sense of 'outsideness'. As long as the sound source is broadband the brain is so effective at interpreting the colourations that a one-eared listener can make quite good direction judgements – computing the signal difference between the two ears is not essential in this task.

Most broadcast stereo material does not attempt to preserve phase differences between the channels (unless the programme is labelled 'binaural'). The direction cues are carried entirely by amplitude differences between the channels. This works, because although in real life there is negligible head-shadowing at the lower frequencies, the brain will use artificially induced amplitude differences to compute source direction. Ian's attempts to turn mono signals into stereo by introducing phase lag were reasonably successful, because the two channels retained equal amplitudes. With a stereo broadcast he was adding phase differences to signals which had conflicting amplitude differences. Under some circumstances time/intensity trading can take place: the brain can set loudness cues against phase difference and decide that the sound source is not displaced at all. However, if the pitch, delay and intensity are not carefully defined it is impossible to predict what the perception will be.

Moreover, if Ian was listening to an orchestra on a wide stage, some of the instruments should have started with phase differences, which aught to have decreased to zero as he turned his head to face them. By treating all the sounds equally he was compressing the entire orchestra into a heap in the middle of the platform – all very messy! Above all, his electronics didn't have ears; the pinnae must be modelled to achieve convincing effects.

I believe what Ian Hickman has attempted will become feasible, but it will require some fearsome dsp chips and a good deal of expense – not least because we don't all have the same shaped heads and ears. Personalisation of the transfer functions will be required.

Dr P L N Naish
Department of Psychology
The Open University
Milton Keynes

Valves – in defence

Responding to Mr Linsley-Hood's rather strident summary dismissal of valve technology, I wonder if he would give us some distortion curves of comparably powered solid state and valve amps below the 2W level. In my experience the solid state – excepting mosfets – amps always have a characteristic rise in distortion in the range below two watts. And since that is where Mr Linsley-Hood rightly observes most music reproduction occurs, isn't that a rather interesting difference between push-pull power amps using valves and solid state?

There is something about valve amplifiers that pleases many people – some very distinguished amplifier designers.†.

The author's sidebar includes mis-statements. My experience with a batch of *2N3055*s and that of the author is totally different. They are by no means identical in gain and vary far more widely than a batch of new valves. While it is true that valves deteriorate in use, we now live with far better voltage regulation and ancillary components than we had in the best days of the valve in the 50s and 60s.

Valve technology can and will be aided by not only voltage regulation but by computer technology for achieving dynamic balance in push-pull pairs and possibly to offset some of the effects of ageing.

To complain that valves can be overdriven with resulting damage is scarcely a fault exclusive to valves. I know of almost no component that does not suffer from misuse.

As to problems with high voltages relating to capacitors, the catalogues I see offer much more reliable capacitors and at higher voltages.

I hope Mr. Linsley-Hood has not missed the fact that valve technology flourishes in the former Soviet Union and that the USSR's entire aircraft and space technology was managed with valves. Further, their valves were not mere copies of those produced in the West. Russia's current space program involving the Mir space station is run entirely by valves.

Edward T. Dell, Jr
Audio Amateur Corporation
Peterborough,
USA

Hamm, Russel O., "Tubes Versus Transistors, Is There an Audible Difference?", *Journal of the Audio Engineering Society*, May 1973, Vol. 21, No. 4, pp. 267-273.

Think and measure and...

I dislike arguing in public but Douglas Self's reply to my letter in the May issue cannot be ignored.

First he misunderstands what I was "challenging" him about and then goes on to be – in my opinion – quite rude, based on these misunderstandings.

While I hold Douglas's technical articles in awe for his obvious understanding of theoretical research and development, it seems he cannot grant others similar respect if their ideas exceed his paradigms.

Contrary to his understanding, I hold no flag for the concept of "10mV diodes in copper wire"; but I observed a curious phenomenon and thought it would be interesting – and easy – for Douglas to *try* it himself, and to have his response. And if not "10mV diodes", then what?

That he doesn't try it is his business and in my opinion a loss of a chance to learn, but to suggest that slight variations on the effective

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dumping on the speaker could make such a huge difference to the perceived *distortion*, and yet say that my new book (The SuperCables CookBook.) "should preferably contain facts" is laughable.

I would never presume to judge Mr Self's power amplifier designs without actually building them and using them to listen to music over a considerable period, but I can say – as I have spent some 20 years successfully manufacturing speciality preamps – that his design in the July/August issue is a little behind the times. The use of large value electrolytics for input and stage coupling is unbelievable, and his choice of the *5532*, however quiet, shows that he cannot have done any serious listening. This IC has an unpleasant sonic signature that caused even budget mixing desk manufacturers to drop it years ago!

In my opinion, Douglas's other sparring partner Ben Duncan has written the book on IC based preamps; and his creations can make music, as do most valve designs.

To end, I'd therefore caution readers to look a little further before spending the time and money to build Mr. Self's latest design, at least if they are seeking musically satisfying result.

Allen Wright
Vacuum State Electronics
Munich

Q&A

Plating through for prototype boards

In answer to the question from Ian Tran regarding a plated-through-hole process for pcbs; I have a booklet which does a good job in describing the process and even goes so far as to give a list of chemicals and equipment needed.

I have used the process on a few projects which turned out quite well. The title of the work is "The PTH Process For Homebrewers". I no longer know the source of the booklet, but basically, the process goes like this:

Remember at all times – wear safety goggles & safety gloves and follow safe chemistry procedures. You will be dealing with very dangerous chemicals

Start with a double-sided pcb blank with about 3/4in extra on one side. Mark and drill all holes. Drill a couple of 1/4in holes in the long end (used to hang the board in the electroplating solution).

Using 200 grit sandpaper, remove all burrs from the holes (important). Clean the board thoroughly with steel wool. Degrease the board with a sodium hydroxide solution (Lye).

Rinse well with clean water. Pre-etch the board with an ammonium persulphate etching solution for about 30 seconds. Rinse well with clean water.

Acid treat for 5 minutes in 10% sulphuric acid solution (reagent grade – *not* battery acid). Rinse well with clean water.

Acid treat for 5 minutes in 33% hydrochloric acid solution. Rinse well with clean water.

Sensitise the board in solution C for 10 minutes. Rinse well with clean water. Activate the board in Solution D for 5 minutes. Rinse in deionised water for 1 minute.

Plate the board in electroless plating solution A/B for 10 minutes. Rinse well with clean water. Acid treat in 10% sulphuric acid for 1 minute. Electroplate in copper electrolyte solution mixture heated to 37°C at a current density of ASF = (L×W×2)/144×30. Suspend the board from the 1/4" holes drilled for this purpose. Do not let the plater connections and hardware used to suspend the board get into the solution. Do not use more than 6V to generate the plating current. Use two phosphorised copper rods (0.02-0.08% phosphorus) 3/4×6in as the anodes. Pure copper is fine, but phosphorised rods give better results.

Plating rate will be about 0.001in per 36 minutes at 30A/ft^2 (ASF).

Rinse well with clean water. Air dry the board (use a hair drier). From this point, use your regular photo process to sensitise, expose, develop and etch the final board pattern. When applying the photo resist, make sure you spray it inside the holes so the plating doesn't get etched away. When making the artwork, use component pads and vias without holes; this ensures the holes don't get etched out.

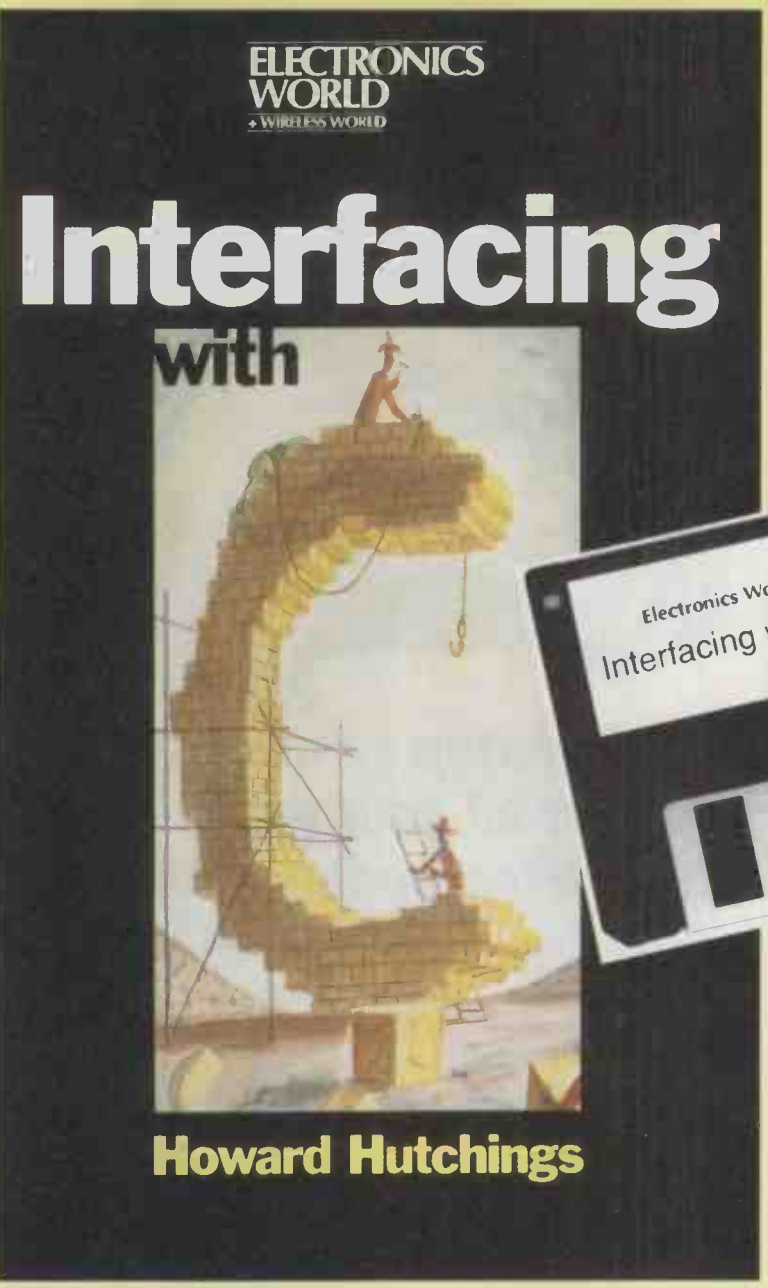
Supplies: Solutions A, B, C, D, and copper electrolyte solution may be obtained from Transcene Company, Inc., Route 1, Rowley, MA 01969 (USA), (617) 948-2501.

Explain what you are trying to do; they will know exactly what you need.

Phosphorised copper rod may be obtained from a well-supplied machine shop.

David Mason
USA

Interfacing with C



Without an engineering degree, a pile of money, or an infinite amount of time, the revised 289-page Interfacing With C is worth serious consideration by anyone interested in controlling equipment via the PC. Featuring extra chapters on Z transforms, audio processing and standard programming structures, the new Interfacing with C will be especially useful to students and engineers interested in ports, transducer interfacing, analogue-to-digital conversion, convolution, digital filters, Fourier transforms and Kalman filtering. Full of tried and tested interfacing routines.
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Audio processing ON THE PC

Howard Hutchings

discusses the C code needed to carry out audio processing functions on the pc – including flanging and chorus effect.

There are in mathematics a handful of principles which look so simple as to be worthless, but yet in practice are of the utmost importance and value. One such example is the principle of superposition – the unique relationship that connects sinusoidal signals and linear systems can be traced back to the property of frequency preservation.

Any signal represented in terms of its component frequencies will be processed by the system in a very simple way; only the amplitude and phase of the input components will be modified. This describes why the response of a linear system to a steady-state sinusoidal input is itself a sinusoid at the same frequency as the input.

This is an important observation which will be developed in this article to produce a range of frequency-selective pc-based digital filters, designed to generate echo, phase flanging, reverberation and all-pass effects.

The software discussed in this article has been written in Turbo C++. The listings are too long to publish here, but they are available; details later. Discussed in this article are the spectral performance of z-plane poles and zeros and their relevance to practical design.

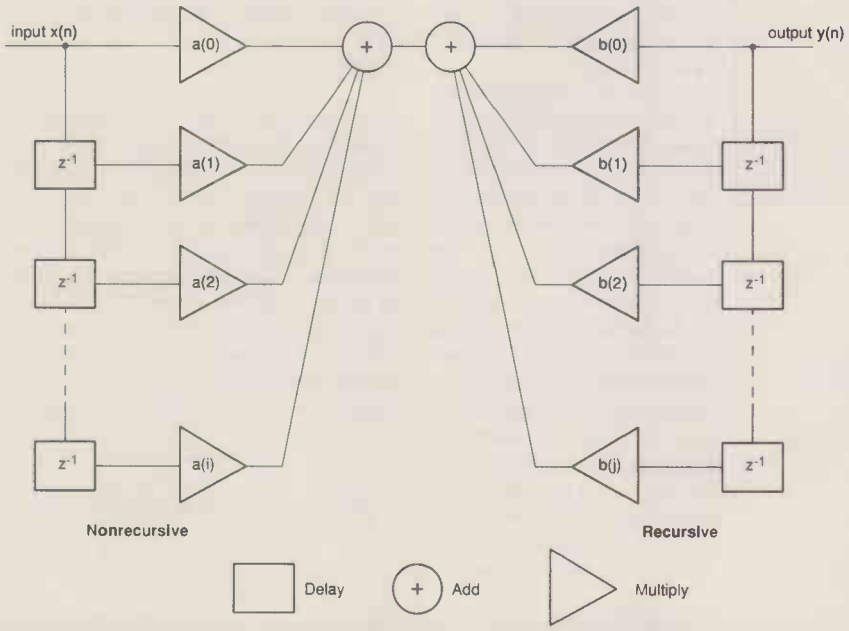


Fig. 1. The system diagram of a general digital filter combines a non-recursive section on the left and a recursive section on the right, in which the output y(n) is multiplied by a series of coefficients b(j) and added to the scaled and delayed input.

Rational functions revisited

For analytical work and conceptual purposes it is often useful to transform the time-domain model of a signal or system into a function, composed of a ratio of polynomials; which may be subsequently investigated by examination of the locations of the poles and zeros. Evidently, there is no mathematical distinction between the transforms of signals and systems. Many signals of practical interest have rational transforms.

A systematic inspection of the table of Laplace transforms of commonly-used functions, for example, 'An Introduction to the Analysis and Processing of Signals', P. A. Lynn, 1989, pp. 251, reveals only one function ($\cos \omega t$) characterised in terms of both poles and zeros. Repeating this exercise, for a similar set of functions described by z-transforms, indicates an altogether different picture. Every function in Lynn's table except one, is characterised in terms of both poles and zeros.

Despite evidence of poles and zeros arising quite naturally in the transfer function development of analogue systems, the concept of impedance excludes a digital development. Unless the notion of poles, and particular relevance of zeros, in discrete lti systems is developed properly it will return to plague the thoughtful person.

Dismantling a digital filter

In this section, the intention is to develop a general difference equation, applicable to both

recursive and non-recursive designs, into a digital transfer function. This will help to establish expressions for both the amplitude ratio and phase response.

To understand the behaviour of discrete lti systems, it is useful to start by considering a general difference equation for modelling weighted delays written as,

$$b_0y(n)+b_1y(n-1)+...+b_Ny(n-N)$$
$$=a_0x(n)+a_1x(n-1)+...+a_Nx(n-M)$$

which has $M+1$ arbitrary coefficients scaling the input values, and $N+1$ arbitrary coefficients scaling the output values. It is customary to describe the computational realisation of the process, explicitly, in terms of the scaled, and delayed, input and output sequences

$$y(n)=\frac{1}{b_0}\sum_{i=0}^{M}a_ix(n-i)-\frac{1}{b_0}\sum_{j=1}^{N}b_jy(n-j)$$

where a_i and b_j are constants for all values of i and j, and b_0 is non-zero. Such an algorithm will be described as causal and may be used to implement a digital filter in real-time, **Fig. 1**.

If N is greater than zero, the difference equation is 'recursive', since previous values of $y(n)$ are used in calculating the current output.

Table 1. ADC-42 Port map and programming model.

Address	Function
Base + 0	End of conversion flag B7 going high.
Base + 1	Upper 4 bits of data word B0–B3. Four MSB B4–B7 set to zero.
Base + 2	Reading this address will start conversion. Also contains the low byte of the data word.
Base + 3	D-to-a strobe, outputs 12-bit word.
Base + 4	D-to-a low-byte register A.
Base + 5	D-to-a high-byte register A.
Base + 6	D-to-a low-byte register B.
Base + 7	D-to-a high-byte register B.
Base + 8	Port A, digital I/O.
Base + 9	Port B, digital I/O.
Base + 10	Port C, digital I/O.
Base + 11	*8255* control register.
Base + 12	Multiplexer channel select B0–B4.
Base + 13	Programmable interrupt source control.

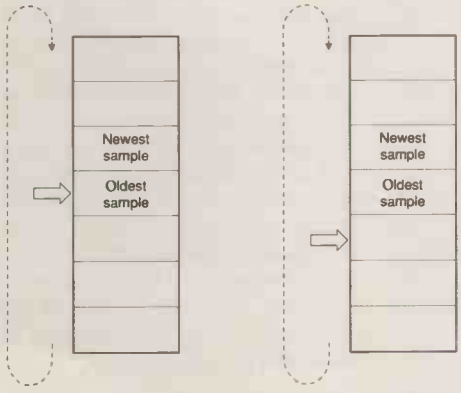


Fig. 3. Circular buffer both before (a) and after (b) a sampling interval. The oldest sample is T.N seconds old.

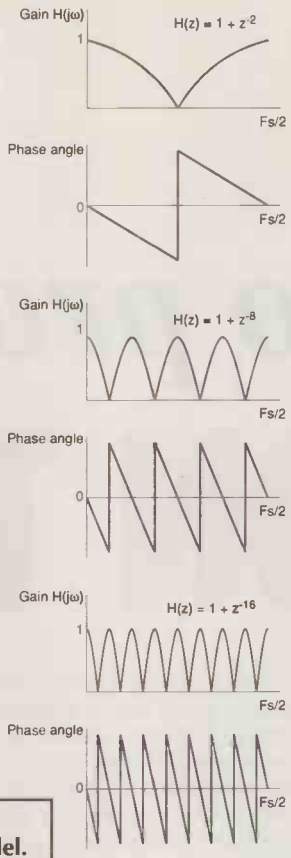


Fig. 2. Frequency response caused by phasing for different time delays. Reconsider this performance with the assistance of zeros on the circumference of unit circle.

When $N=0$, the difference equation is non-recursive and the processed output is composed exclusively of scaled input samples.

A chronic problem with abstract mathematical expressions of this form is that they tend, initially, to discourage rather than encourage further reading. Certain authors appear so completely out of touch with the real world to have 'forgotten' why digital filters exist; not for the purpose of mathematical manipulation but to remove certain frequencies and allow others through. Examine the index of your favourite dsp text. Carefully consider how the concepts of low-pass, high-pass, band-pass and band-stop are developed and cross-referenced. Are you convinced?

Frequency-selective properties of zeros

Unlike certain of their analogue counterparts, the concept of z-plane zeros figures prominently in the design of many selected discrete-time linear processors. A good grasp of the spectral properties and time-domain performance will greatly increase awareness of the special features and possibilities for design.

This is a useful opportunity to outline the characteristics of a family of non-recursive filters, formed by combining the current input to the filter with selected past inputs. Conceptually, this type of linear signal processing system is easy to understand in the time domain, and particularly revealing of the unique features of digital filtering in the frequency domain.

Discussions in this section will review the analytical signal processing principles and practical design parameters related to the performance of selected comb-filters. The natural emergence of 'notch' filtering will be shown.

To generate echo, $y(n)$, it is necessary to record or store a weighted signal, $Bx(n-k)$ before releasing it a fraction of a second later, together with the present scaled input, $Ax(n)$.

$$y(n)=Ax(n)+Bx(n-k)$$

If a complex signal is fed through a delay loop, all frequencies in the signal, whether low or high, are delayed by the same time. The effect of having the same time delay for all frequencies creates a different phase delay for different frequencies, since different frequencies have different wavelengths. For example, later discussions in terms of zeros, will show how a time delay of 2.5ms will cause a 360° phase difference in a 400Hz sinewave, but only 180° phase difference in a sinewave of 200Hz. Thus a cancellation will occur at 200Hz but not at 400Hz. This would correspond to an attenuation at 200Hz in the frequency spectrum of the signal.

The phase lag ϕ for any frequency f, for any given time delay τ, is given by

$$\phi=2\pi f\tau$$

When two identical waves separated by a phase lag are added together, their sum is a wave whose amplitude depends on the phase lag. For example, if two waves, each of amplitude A, are added together in phase (ie $\phi=N\pi$, where $N=0, 2, 4, 6,...$), the sum or resulting amplitude is $2A$. If the two waves are 180° out of phase (ie $\phi=N\pi$, where $N=1, 3, 5...$), they cancel out, so the resulting amplitude is zero. For angles between 0° and 180°, the resultant amplitude is between $2A$ and 0. For a given delay ϕ between two waves of amplitude A,

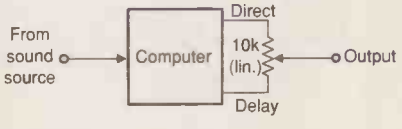


Fig. 4. Use two output ports and d-to-as to reconstitute the processed signal in the desired proportions, without software-based multiplication.

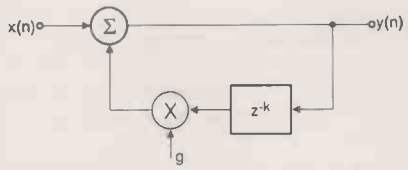


Fig. 5(a) Prototype reverberation generator, with weighted feedback.

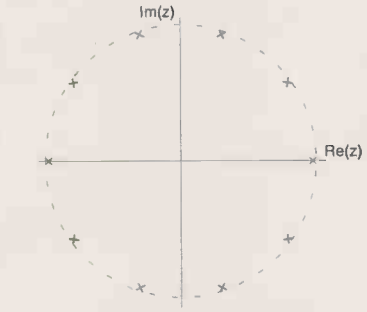


Fig. 5(b) Pole-zero diagram.

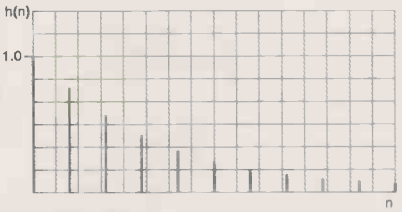


Fig. 5(c) Impulse response.

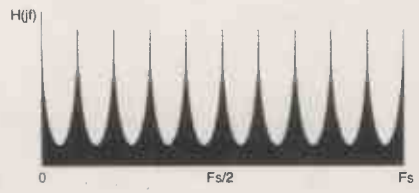


Fig. 5(d) Amplitude-ratio.

the resultant amplitude A_R of the waves summed is,

$$A_R=|2\cos(\phi/2)|A$$

The absolute value of $2\cos(\phi/2)$ is taken because the amplitude is always positive.

Combining the two equations stated before gives,

$$A_R=|2\cos(\pi f\tau)|A$$

This produces the frequency response as a function of frequency and time delay. The expression shows that the frequency response takes the same shape as a simple sine wave, with the negative peaks inverted. For long time delays, the distance between peaks on this sine curve is small. The opposite is true for short time delays.

Frequency responses

Figure 2 shows the frequency response curves for several different time delays, obtained using the software available on disk. Frequency is plotted on a linear scale to show the regularity in the pattern of the characteristic response. The effects of the zeros positioned on the circumference of the unit circle are to generate variable rejection frequencies.

The characteristic frequency response of this linear processing system is that of a comb-filter, with multiple peaks and troughs distributed regularly throughout the frequency spectrum. This is due to the effects of zeros, located on the circumference of the unit circle, completely attenuating each frequency whose period is an integral multiple of twice the time delay.

For example, the principle of echo, obtained by adding the present input $x(n)$, together with the signal captured two sampling intervals previously, can be described by the difference equation,

$$y(n)=x(n)+x(n-2)$$

and represented by the transfer function

$$H(z)=Y(z)/X(z)=1+z^{-2}$$

Equating the numerator to zero gives the position of the characteristic zeros.

$$(z+j)(z-j)=0$$

It is not difficult to show how the pair of conjugate zeros, at $z=\pm j$, are located on the circumference of the unit circle, in the z-plane, at frequencies of $f_s/4$ and $3f_s/4$ respectively. Before considering this design further, it is interesting to confirm how this difference equation, with f_s=800Hz, will provide the frequency selective properties discussed previously.

Producing echo

The use of a numerical processor with memory to simulate the effect of a 'cave-like' environment allows considerable control over the parameters of the system. For example, to reproduce a delay of 0.2048s in a system sampling at 20kHz will require a delay line composed of 4096 cells (listing 2 on disk).

The historical record will consist of digitised audio stored in an array, configured as a circular buffer made up of 4096 cells. In other words, the design should set aside 4K of memory, which will contain the oldest sample captured, 4096.T, seconds previously, (T is the sampling interval) and each subsequent sample, up to and including the present input, **Fig. 3**.

In mathematical terms the procedure is to address the buffer modulo its size. Let the buffer have N cells, and allow the real-time program to write into the buffer once every T seconds, so that the cell of current interest was last addressed $T.N$ seconds ago. The data stored there will be $T.N$ seconds old.

Examine how the buffer was first initialised using a do-while construct. This ensures a delay of 4096 sampling intervals between the newest and oldest samples. What is needed to dynamically address each member of the array, is a software structure which will increment a counter modulo the length of the buffer.

In this case, $k=k+1$ was used to increment, followed by the modulo operation $x=k\%4096$. Because the do-while construct executes the main body of the loop at least once before performing the test, the current and delayed data

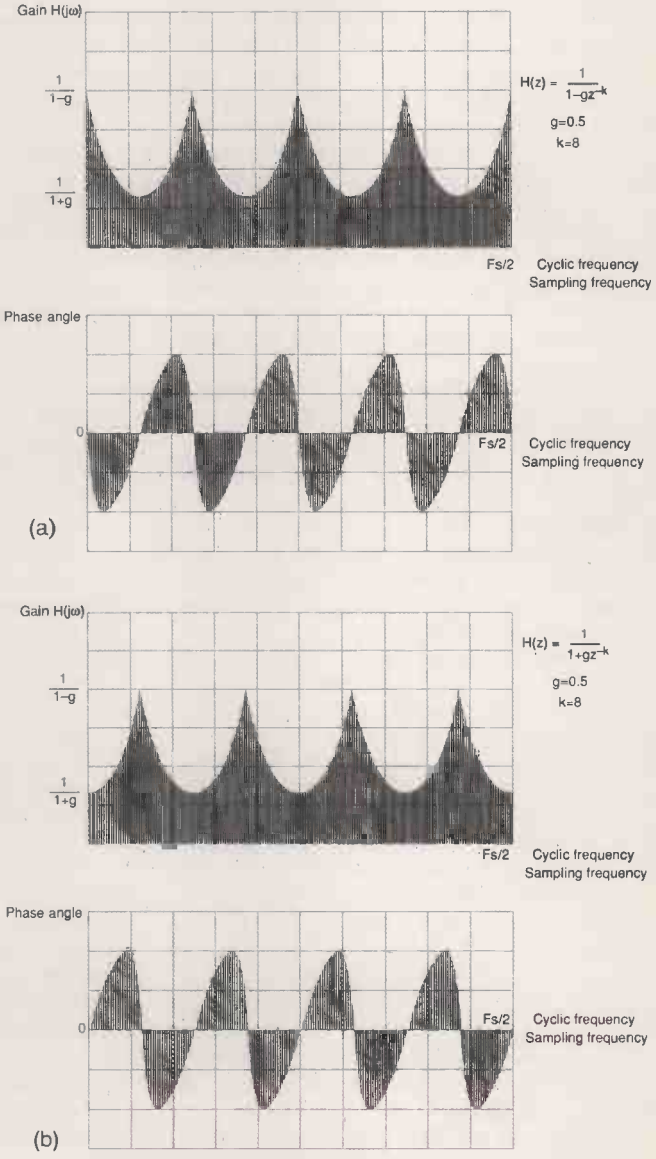


Fig. 6. The spectral performance of the reverberator. (a) Scaling factor g=–0.5. (b) Scaling factor g=0.5.

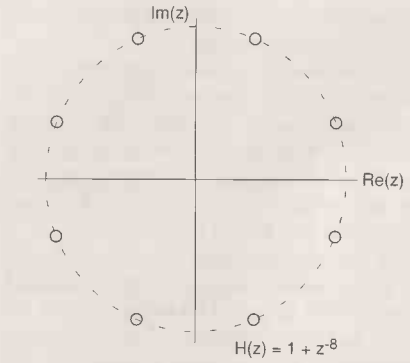


Fig. 7(a) Visualising the performance of a phase flanging processor using a pole-zero diagram.

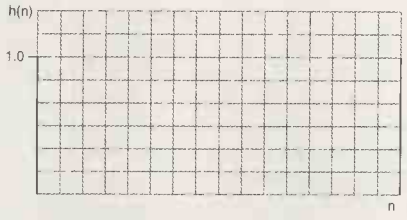


Fig. 7(b) Impulse response.

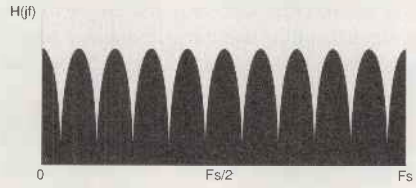


Fig. 7(c) Spectral performance.

are respectively stored and output in a single pass of the loop. This offers a real-time overhead of less than 10µs per computation, (listing 2 on disk).

The success of such a system in audio signal processing depends upon its ability to operate at ultrasonic speeds; that is, the rate at which the audio signal is digitized, processed, and output, must be well above the upper limit of the audible spectrum. Thus the speed of each of these procedures is critical. The operation of each will now be considered individually.

Together with the time-critical, synchronised high-speed programmed i/o and real-time processing controlled by the personal computer, the sampling rate of the a-to-d converter is also a most time-significant parameter. It must be fast, a conversion time of 50µs or less is necessary to support successful audio processing.

The commercially available data acquisition board, Blue Chip model *ACM-44*, used here for the purpose of description, employs the Analog Devices *AD7820*, an eight-bit half-flash converter with a 1.56µs conversion time., The present author recommends such a unit, or any industrial equivalent. It is probably faster than present requirements, but it need not be replaced when pc-based systems get faster, as they certainly will.

Slower peripherals used in conjunction with a faster processor will perform satisfactorily. For example, recently I have been using a *PCL-818* data acquisition card which has a 10µs conversion time 12-bit a-to-d converter, and a 5µs settling 12-bit d-to-a, with successful results on a 486 50MHz pc.

Returning to the original problem of echo generation. It is not difficult to see how the processed output is the scaled combination of the present and delayed inputs. To avoid the real-time overhead of software-based multiplication, necessary on account of scaling, it was decided to output the processed data through two separate d-to-a converters. The plan is to sample and process the input signal before generating two separate channels of audio output, **Fig. 4**.

By mixing the 'direct' and 'delayed' signals together across a potentiometer, it will be possible to reconstitute the two signals in any desired proportions; that is, the original signal with pronounced echo, or no delay at all, depending on the position of the wiper.

Reverberation

It has already been outlined, using zeros, how the time-domain performance of an echo generator can be modelled by

$$y(n)=Ax(n)+Bx(n-k)$$

Thus, if a single pluck of a guitar string, a unit pulse, generates the system response TWANG... TWANG, that is identified as an echo. Applying feedback to a delay loop, to be followed by subsequent developments into closed form will now introduce the relevance of poles, in order to described an elementary, but important, modification to generate multiple echoes, reverberation. The unit pulse will generate the impulse response TWANG, TWANg, TWAng,... twang, which may be modelled by

$$y(n)=Ax(n)+By(n-k)$$

The amount and quality of reverberation that occurs in a natural environment is influenced by: the volume and dimensions of the space; and the type, shape and number of surfaces that the sound encounters. The amplitude of any sound is reduced by an amount that is inversely proportional to the distance it travels; therefore the reflected sounds not only arrive later, but they have smaller amplitudes than the direct sound. This means that the impulse response will have a decaying envelope, as will be shown.

It is generally accepted that characterisation of reverberation is particularly difficult, because the quality of reverberation cannot be quantified objectively. Four parameters which may be correlated with the perceived quality of reverberation are: the reverberation time, the frequency dependence of the reverberation time, the time delay between the arrival of the direct sound and the first reflected sound, and the rate of build up of echo density.

The reverberation time indicates the amount of time required for a sound to die away to 1/1000 (–60dB) of its amplitude after the source is removed. The choice of –60dB represents a convenience inherited from early researchers of room acoustics.

The relationship between reverberation time

Listing 1. All-pass filter designed to generate chorus effect. It samples i/p through a-to-d o/p through dual d-to-a converters. Loop time is 0.2048s, reverberation time is 1.89s.

```
#include<stdio.h>
#include<conio.h>
#include<dos.h>
#define BASE        768
#define M           4096
void main(void)
{
char key;
unsigned int i, k = 0, x = 0;
int contents, output, temp;
int input_data[M];
int output_data[M];
textmode(C80);
textbackground(1);
textcolor(14);
clrscr();
gotoxy(6,4);
cprintf("Digital Filter designed
to generate Reverberation");
gotoxy(6,6);
cprintf("y[n] = g.y[n - k] + x[n
- k] - g.x[n]");
for(i = 0; i <= M; i++)
   {
    input_data[i] = 0; /* Flush
buffers */
   output_data[i] = 0;
   }
outportb(BASE,1);
/* Select I/P Channel */
for(;;)
   {
    do
     {
     outportb(BASE + 2,0);
     /* Start conversion */
      contents = inportb(BASE +
2);
            output   =   0.7   *
(output_data[x]  -  contents)  +
input_data[x];
     outportb(BASE + 4, output);
     /* Write to port DIRECT */
     input_data[x] = contents;
     output_data[x] = output;
     /* Store incoming data in
        a circular buffer */
     k ++;
     x = k % M;
     }
     while(k < M);
               outportb(BASE   +
5,input_data[x]);
     /* Write to port DELAY */
     if(k & 8192)
       {
       k = M;
       }
   }
/* End of main */
}
```

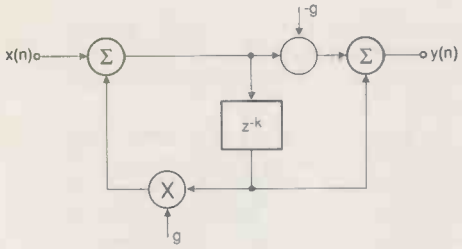
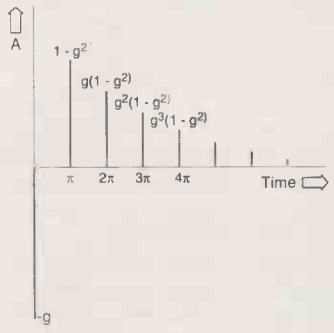


Fig. 8(a) All-pass filter flow diagram.



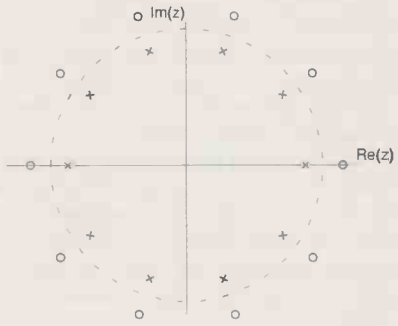


Fig. 8(c) Pole-zero configuration shown for k=10.

and frequency will be described later.

Defining delay times

The delay time is the amount of time that elapses between receiving a direct sound and its first reflection. A long delay of 50ms or more can result in distinct echoes, whereas a very short delay of 5ms or less can contribute to the listener's perception that the space is small. A delay in the range 10 to 20ms is found in many concert halls.

Following the initial reflection, the rate at which the echoes reach the listener begins to increase rapidly. A listener can distinguish differences in echo density of up to a density of one echo/ms. The amount of time required to reach this threshold is typically 100ms. This time is approximately proportional to the square root of the volume of the room, so that small spaces are characterised by a rapid build up of echo density.

In a recursive comb-filter, **Fig. 5a**, the input signal enters a delay line. When it reaches the output, it is fed back to the input after being multiplied by the scaling factor g. The time that it takes to circulate once through the delay line is identified as the loop-time. The loop-time is the product of the delay introduced by the delay line, k, and the sampling interval T. Listing 3 on the disk helps visualisation of the pole-zero configuration, impulse response and amplitude-ratio for various delays, k, input by the user, Fig. 5(b–d).

Consider the performance of the prototype reverberator,

$$H(z)=1/1-gz^{-k}$$

When a unit pulse is applied to the input, the impulse begins to propagate in the delay line. The output of the filter is zero until, after kT seconds, the impulse emerges from the delay line. At this time the output of the filter is the impulse with unit amplitude. Meanwhile, the impulse is multiplied by the scaling factor g and fed back into the delay line with amplitude g. The process continues; a pulse is output every kT seconds, and each pulse has an amplitude that is a factor of g times that of the preceding pulse. The modulus of the scaling factor must be less than unity for the filter to be stable, typically written as $|g|<1$, which places the poles inside the unit circle.

Fantasia revisited?

Recursive designs generally make use of poles situated close to the unit circle. The use of limited word lengths may result in small errors in the coefficients of the time domain recurrence relationship, effectively causing the system poles to move outside the unit circle. Do you remember the fate of the sorcerer's apprentice who experimented with feedback using the magic broom, sweeping water from the magic pump, which he was unable to control? A more judicious choice of pole locations might have resolved the problem.

The impulse response decays exponentially as determined by the parameters chosen for the loop time and g. Values of g closest to unity give the longest delay times. To obtain a desired reverberation time (T_r), g can be calculated, given the loop time (τ), from the relationship,

$$T_r/\tau=-3/\log_{10}(g)$$

This parameter is probably one of the most important in characterising the performance of the acoustic environment. It is of some interest to note a few practical figures. A cathedral, for example, might have a reverberation time of about 5s; a typical home living room less than 1s. For the purpose of this discussion it will be sufficient to note that a reverberation time of 1.9s is considered suitable for the best concert halls.

I will now detail the relationship between the parameter g and the frequency response of the discrete LTI system $H(z)$ (listing 3 on disk). Two cases must be considered.

Case A:

$$H(z)=1/1-gz^{-k}$$

The location of the poles is given by the roots of the equation:

$$z=g^{1/k}\exp(j2\pi m/k)$$
$$m=0,1,...k-1$$

Expressed as a power series in z^{-k}, this can be written as the unipolar impulse response,

$$H(z)=1+gz^{-k}+g^2z^{-2k}+g^3z^{-3k}+...$$

Case B:

$$H(z)=1/1+gz^{-k}$$

The location of the poles will be given by the roots of the equation:

$$z=g^{1/k}\exp(j(2m+1)\pi/k)$$

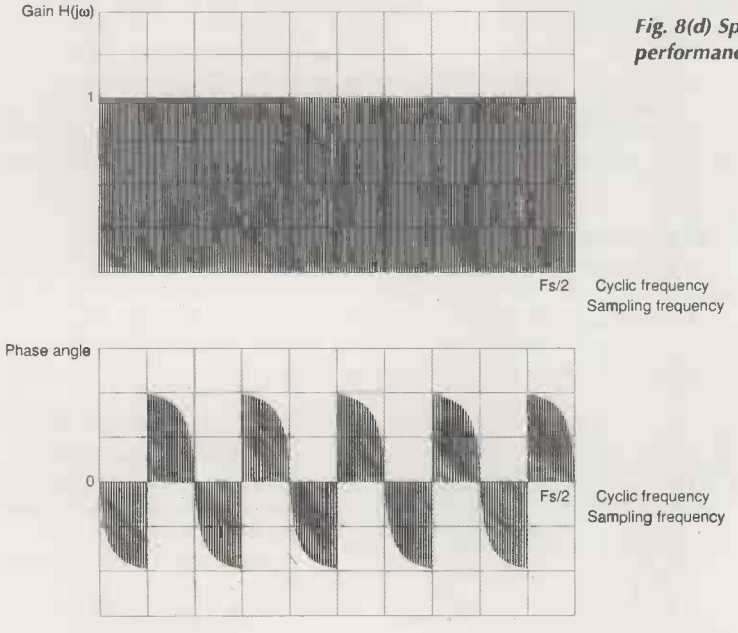


Fig. 8(d) Spectral performance.

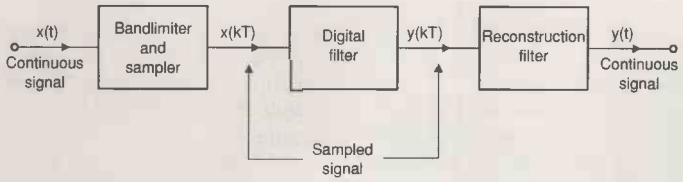


Fig. 9. For audio applications, the signal to be filtered is first band limited to prevent errors due to aliasing. The processed output is low-pass filtered to remove the effects of sampling. A MAX275 configured as a 4th order Butterworth gave excellent results, together with a low component count.

$m=0,1,...k-1$

Expressed as a power series in z^{-k}, the alternating characteristic of the impulse response follows quite naturally,

$$H(z)=1-gz^{-k}+g^2z^{-2k}-g^3z^{-3k}+...$$

The comb-filter is so named because its steady-state amplitude response, is considered to resemble the teeth of a comb. The spacing between the maxima of the teeth of the comb is equal to the natural frequency

$$F_0=1/kT$$

Referring to the amplitude ratio, **Fig. 6**, the depth of the minima, $1/(1+g)$, and the height of the maxima, $1/(1-g)$ are determined by the choice of g, values close to unity yield more extreme maxima and minima.

My routine for this (listing 4 on disk) functions as a real-time reverberation system, with parameters: g=0.5, T=25µs, T_r=1.02s.

Phase flanging

Typically, digital filters are designed to obtain a specific steady-state amplitude response; although developments in this section will detail the performance of discrete LTI systems designed to realise a particular impulse response. The effects described will include echo and phase flanging.

Flanging is an effect created by adding together two identical signals separated by a very short time interval. If the time delay is typically less than 25ms, the ear is usually unable to resolve the direct and delayed signals into two separate and distinct sounds. Instead, a complex sound is heard, described by Bartlett (1970) as "A hollow swishing, an ethereal effect, something like a jet plane without all the roars and rumblings". O'Haver (1977) identifies the signal processing effect as 'resonant' or 'twangy'.

Claiming that with speech or solo singing it gives a voice doubling effect, as if two people were speaking in synchronism. Six-string guitars are reminiscent of twelve-string instruments and concert pianos sound like 'honky-tonks'. Bartlett claims it can be applied most effectively to drums and identifies a number of contemporary recordings in which flanging is most pronounced and audible. Itchycoo Park by the Small Faces is my favourite.

It has already been discussed how the consequence of having the same time delay for all frequencies creates a different phase delay for different frequencies. Such an idea can be extended, and the audio effect made more pronounced, by causing the 'delay' to change continuously in real-time. This will cause the frequency response of the comb filter to sweep through the audio spectrum in real-time.

The linear signal processing operation is called 'phase flanging', and can be imposed on sounds by using a delay line whose delay-time can be varied on a sample-to-sample basis. (This processing operation is visualised with the assistance of listing 5 on disk). The performance of the system is characterised by the transfer function,

$$H(z)=1+z^{-k}$$

Initially, the program requests the delay parameter k. This will be an integer. The program then computes and plots the associated: pole-zero diagram, impulse response and Fourier transform. Use the monitor graphics to follow the dynamic behaviour as the parameter k is decremented, hit any key to view the next time-frame. **Fig. 7** shows the effect of making k=8.

On the disk, listing 6 contains the details of a pc-based real-time system designed to produce phase flanging; the delay loop is made up of a circular buffer composed of 4096 cells, the sampling interval T=20µs.

All-pass filter, chorus effect

Reverberation is a repetitive echo, made more pronounced if the time delay between reflections is shorter than a single echo. By adding the present non-delayed input to weighted previous inputs, the signal will re-cycle through the processor until it becomes inaudible. Typically, the intention is to obtain smooth sounding reverberation, free of repetitive echoes.

To avoid the reverberant ringing, colouration of the sound, associated with the equi-spaced peaks of certain comb filters, it will be useful to consider the theoretical performance and practical behaviour of the following all-pass filter, designed to generate a chorus effect. A suitable system function $H(z)$, designed to achieve a more natural sounding reverberation is formed by subtracting from the output a portion of the input,

$$H(z)=z^{-k}(1-gz^{+k}/1-gz^{-k})$$

Investigated as the convolution of two sequences, which separately characterise the non-recursive comb filter and recursive low-pass filter, this interesting all-pass system embodies the computational advantages of recursion. For example, once a signal has been applied to the input of the system, the reverberation time (T_r), can be several orders of magnitude greater than the delay time (t), giving effectively much longer memory, than the number of taps.

It is rewarding to reconsider the performance of this particular filter from a steady-state perspective and pole-zero model, **Fig. 8**. The pole-zero model geometrically illustrates the all-pass nature of the filter. The effect of each pole (resonance) being offset by a radially displaced zero (notch) relative to the circumference of the unit circle. The steady-state amplitudes of the spectral components of the sound will not be altered. Of course this does not mean the filter is transparent to signals as inspection of the phase and impulse response reveals.

The details of the real-time software are in contained in **Listing 1** (listing 8 on the disk). The analytical choice of the amount of delay and amplitude scaling are likely to remain subjective. Listing 1 utilises a delay of 4096 samples together with a scaling factor g equal to 0.7, to produce a reverberation time of 1.89s at a sampling frequency of 40kHz. As previously outlined, by generating two channels of audio output, the 'direct' and 'processed' it will be possible to control the amount of reverberation developed across the loudspeaker using the balance potentiometer.

Hardware requirements

To run the real-time programs detailed in this article, will require a pc-based *8088* processor, or generic equivalent; with from 1K to 6Kbytes of programmable memory, an 8-bit input port connected to a fast 8-bit analogue to

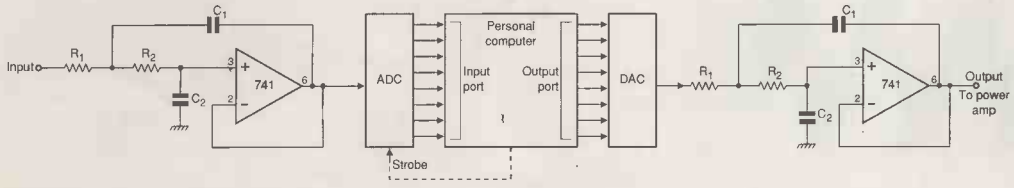


Fig. 10. The DSP system must include an anti-aliasing filter before the a-to-d and a low-pass signal reconstruction filter after the d-to-a. Ensure the –3dB frequency is identical in both filters.

digital converter, and a latched 8-bit output port connected to a digital to analogue converter. An additional output port and digital to analogue converter will be required for stereo applications.

Several different manufacturers now market general-purpose data acquisition boards, these usually include a multi-channel a-to-d converter, one or more d-to-a converters, and a programmable clock as standard features, with options such as programmable gain, variable sampling rate and direct memory access.

The general scheme for a pc-based audio signal processing system is shown in **Fig. 9**. To avoid errors due to aliasing, the continuous signal to be numerically processed is first band-limited by an analogue low-pass filter, before being converted into digital form by an a-to-d converter. The input data is manipulated mathematically by a numerical signal processor, and then converted back into analogue form by a d-to-a. Finally the sampled analogue output is further low-pass filtered to remove unwanted high-frequency components.

Although computer programming takes a good deal of time and effort, nowadays it is usually taken for granted and the code kept separate from the main document. Many of the listed programs are lengthy on account of the accompanying text and graphics and for this reason are available on disk.

However, the central message of this article has been how to utilise relatively long delays in real-time audio processing. For that reason it was considered appropriate to detail a short listing showing how the delay is organised within a circular buffer. On each pass of the loop, the program refreshes two buffer registers: input_data[x] and output_data[x] with the current sampled input, x(n), and current processed output, y(n), respectively.

Blue Chip *ADC-42* data transfer

The principal component of the digital signal processing system described here is a commercially available data acquisition board that fits in one of the expansion slots of the pc. Additional hardware will be required to process the bipolar audio signal, normally, an anti-aliasing filter and dc offset circuit prior to unipolar a-to-d conversion. Following manipulation in the pc, the sampled output of the d-to-a will be reconstructed through a low-pass filter before being amplified and developed across a loudspeaker.

A typical analogue input-output board designed for bus compatibility might occupy a total of fourteen bytes of memory. These memory locations will usually include a multi-channel a-to-d converter, one or more d-to-a converters, and a number of i/o ports as standard features.

Several manufacturers now supply general-purpose data acquisition boards. I have used the PC-Lab Card *PCL-818*, the Blue Chip *ACM-44* detailed in the text, and the Blue Chip Technology general-purpose i/o card *ADC-42*, used here for the purpose of description.

Software on disk

The following listings are available on pc-compatible disk for £14.99 – fully inclusive. Please send cheque or postal order to Audio C, EW Editorial, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.

Listing 1. Graphical amplitude-ratio and phase response of rational z-function.
Listing 2. Real-time echo generator written in Turbo C++, generates a delay of 0.2048s.
Listing 3. Computer-managed instruction designed to visualise the performance of a reverberator. The software written in Microsoft C plots the pole-zero diagram, impulse response and amplitude-ratio on the monitor. The user can vary the parameter *k* to observe the effect on spectral performance.
Listing 4. Real-time reverberator written in Turbo C++, designed to generate a reverberation time of 1.02s.
Listing 5. Computer-managed instruction designed to visualise the dynamic performance of a phase flanging processor. The software written in Microsoft C will plot the position of the characteristic zeros, the impulse response and the amplitude-ratio for a selected value of delay (*k*). Thereupon the effects of decrementing the delay are visualised using the monitor graphics.
Listing 6. Real-time flanging system written in Turbo C++, the sampling interval is 20μs and the circular buffer is composed of 4096 cells.
Listing 7. Real-time all-pass filter written in Turbo C++, designed to generate a chorus effect. Parameters g=0.7, number of cells in circular buffer = 4096, T=25μs and

The Blue Chip board can be given instructions, and have information read from it, using for example, the port-mapped Turbo C functions: outportb and inportb respectively. The comparable Microsoft C functions are outp and inp. The Base i/o address set at the factory is $0\times300_{16}$. To avoid bus contention this is selectable throughout the prototyping region, (0×300–$0\times31F_{16}$). If this port-mapped space is already monopolised, a reasonably safe range of addresses is (0×200–$0\times21F_{16}$).

The analogue input section features a software-controlled multiplexer giving access to 16 single-ended, or 8 differential channels. Full-scale input voltage is link selectable in the unipolar ranges 0–5V or 0–10V. Bipolar provision is provided with link selectable: ±2.5V, ±5.0V and ±10.0V ranges. Analogue to digital conversion is achieved using the Analog Devices *AD7572A*, a 12-bit successive-approximation converter. Software control is straightforward; strobing the a-to-d by simply reading the data input port starts conversion, which is completed in 10μs. This may be followed by an optional 'flag-test' to determine end-of-conversion.

Provision for digital to analogue conversion is provided by two 12-bit converters, (*AD7537*) the output voltage being in the range 0–10V full-scale. In addition, the board provides a single *8255* PPI offering 24 input-output lines through three ports.

Before data transfer can begin, the interface must be set up. In this particular case the initialisation procedure includes selecting the required channel number (0–15) by writing to Base + 12. Next, the start conversion is initiated by reading Base + 2. The resultant data is of no consequence and may be discarded. Conversion is complete when bit 7 of Base + 0 goes high. It is good practice to set up a polled-loop to monitor for e.o.c. To recover the 12-bit word accessed in two parts it will be necessary to, firstly, read Base + 1 for the four msbs of the high byte result. The card automatically puts zeros into the four lsbs. The low byte result is obtained by reading Base + 2. Finally, it will be necessary to reconstruct the 4 and 8 bits into a 12-bit data word prior to digital filtering in the pc.

The relationship between low-pass filtering, sample rate and aliasing will now be briefly reviewed.

A sampling frequency of 20kHz restricts the bandwidth of the digital signal processor to 10kHz if aliasing is to be avoided. An idealised band-limited filter with a well-defined pass-band of rectangular shape is needed when operating close to the Nyquist frequency. However, such a filter is not practically realizable and a compromise is required. Elementary first-order low-pass filters are inadequate for this application because the high frequency rate of 'roll-off' is not steep enough.

A satisfactory approximation to the 'brick-wall' characteristic is provided by a second-order Butterworth filter. This gives a maximally flat response in the pass-band, a sharp corner frequency and a reasonably rapid transition to the attenuation band. A unity-gain Sallen and Key design was satisfactorily used as an anti-aliasing and signal reconstruction filter, **Fig. 10**. Following the practical discussions of (O'Haver, 1978), the –3dB frequency of both filters was designed to be 1/4 of the sampling frequency. ■

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Precise current regulator

Contributions by Brotton and Bradbury[1] and Baert[2] gave rise to this circuit, which is a combined voltage regulator and voltage-to-current converter using the Zetex *ZR78LO5* voltage regulator and an *LM324*.

Normal output of the regulator is 5V, this being increased in this application by the inclusion of $R_{1,2}$ to lift the ground input by 2.2V, giving 7.2V output – often used for battery charging. Positive and negative regulators are separated by the transformer windings and can therefore be referred to ground to give ±5-20V output.

For current conversion, two op-amps are used as followers to provide the input impedance, current mirrors formed by the other two op-amps and transistors being driven by the supply currents I^+ and I^- of the op-amps to give $I_{out} = I^+ - I^- = (V_1 - V_2)/R$. Choose R to give the required current. Resistors in the current mirrors should be 1% types.

Kamil Kraus
Rokycany
Czech Republic

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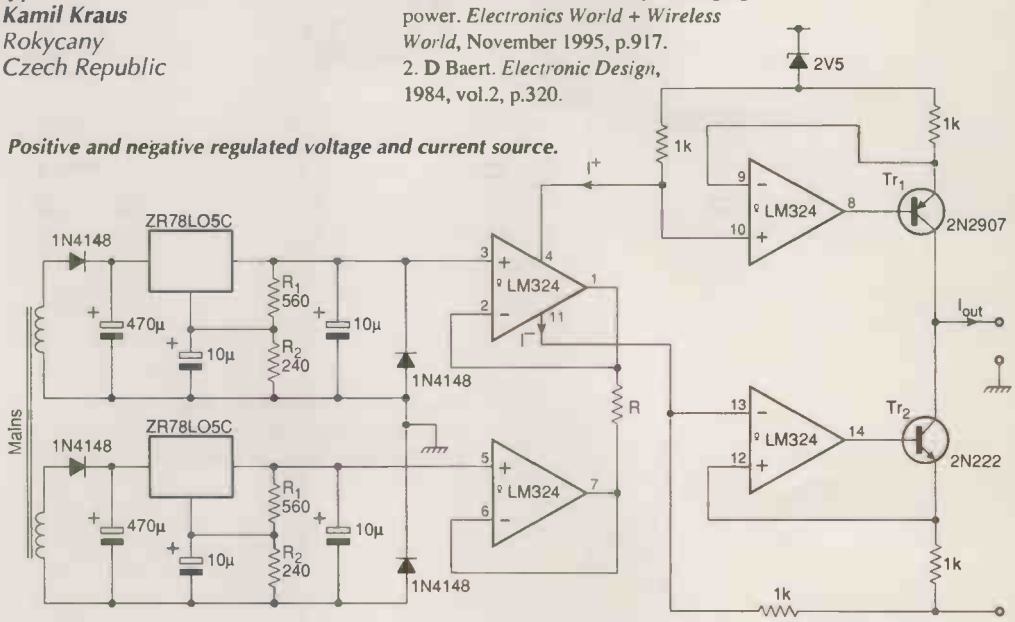
1. D Brotton and D Bradbury. Managing power. *Electronics World + Wireless World*, November 1995, p.917.
2. D Baert. *Electronic Design*, 1984, vol.2, p.320.

Positive and negative regulated voltage and current source.



Electric whisker

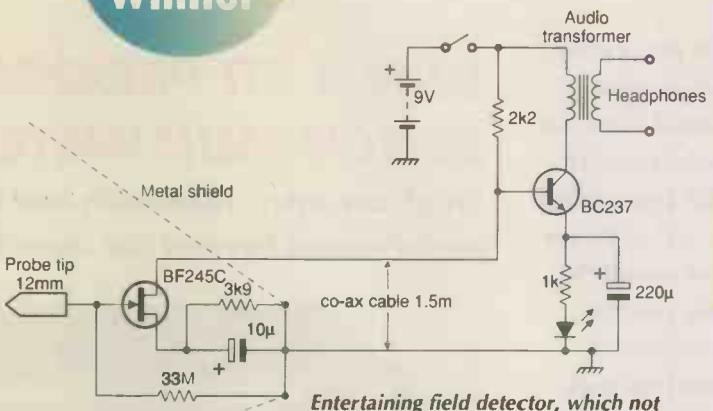
£100 winner

With this electric field detector, you can probe for buried cables or pipes to a depth of 3-5cm and detect the field of an appliance.

The 'probe' is the gate lead of a *BF245C* fet, a 33MΩ resistor being soldered to it where it enters the transistor with as short a lead as possible; cover the probe with spongy plastic to avoid contact with a surface. A 1.5m length of coaxial cable connects the high-impedance section to the amplifier, in which the output transformer was, in the original, salvaged from an old transistor radio. Volume is adequate and a 33Ω pot could be inserted in the transformer output, if required. To improve directionality, connect a 5cm metal screen as shown.

Aside from detecting fields, you can use this device as a telephone monitor: just bring the probe near a cable and hear the conversation – and the noise. It also works as a microphone by putting a bit of kitchen plastic film, used as a membrane, between your mouth and the probe. Tape the output and hear it later; quality is surprisingly good.

D Di Mario
Milan
Italy



Entertaining field detector, which not only detects fields, but works as a gramophone stylus. Shape the tip to a point and lean it against a vinyl disc. Quality is unremarkable, but you can hear it.

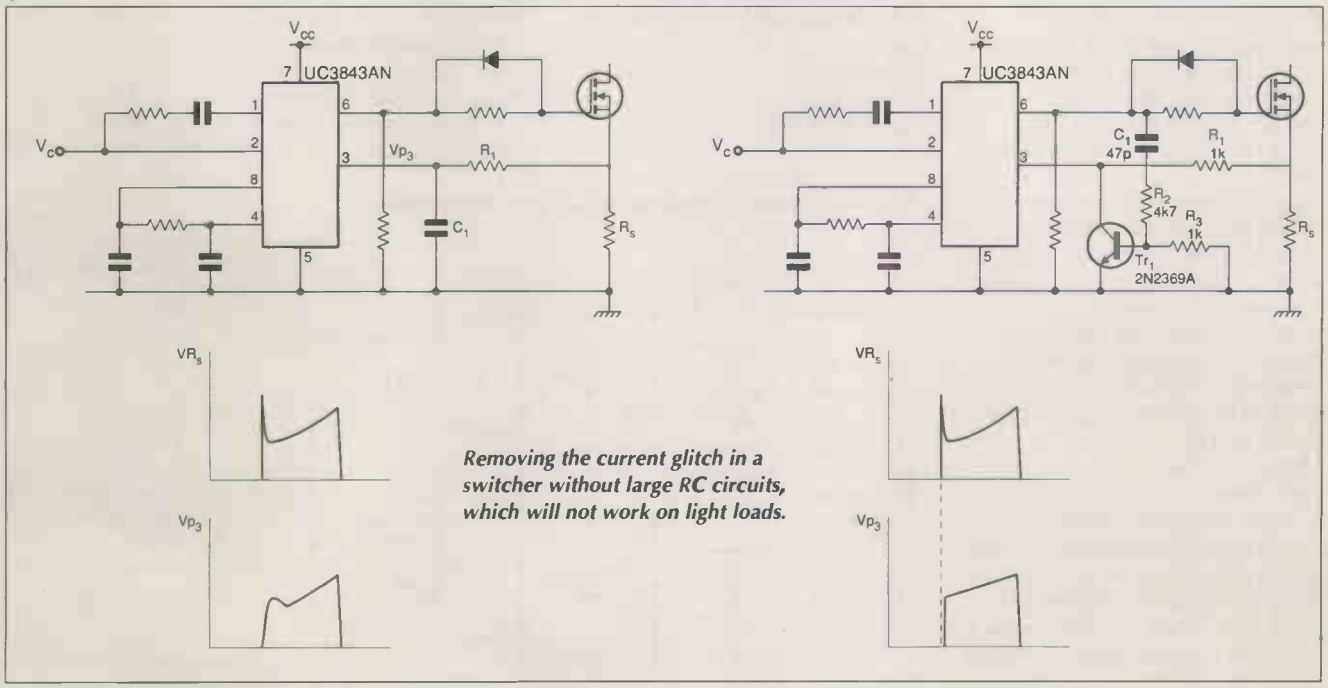
Deglitcher for more stable switching power supplies

Many switching power supplies use an *RC* circuit to suppress the current glitch, which is the result of diode reverse charge and parasitic capacitances in the circuit. The remedy is not particularly successful, however, because the *RC* widens the glitch, as well as decreasing its height, as shown in the left-hand circuit.

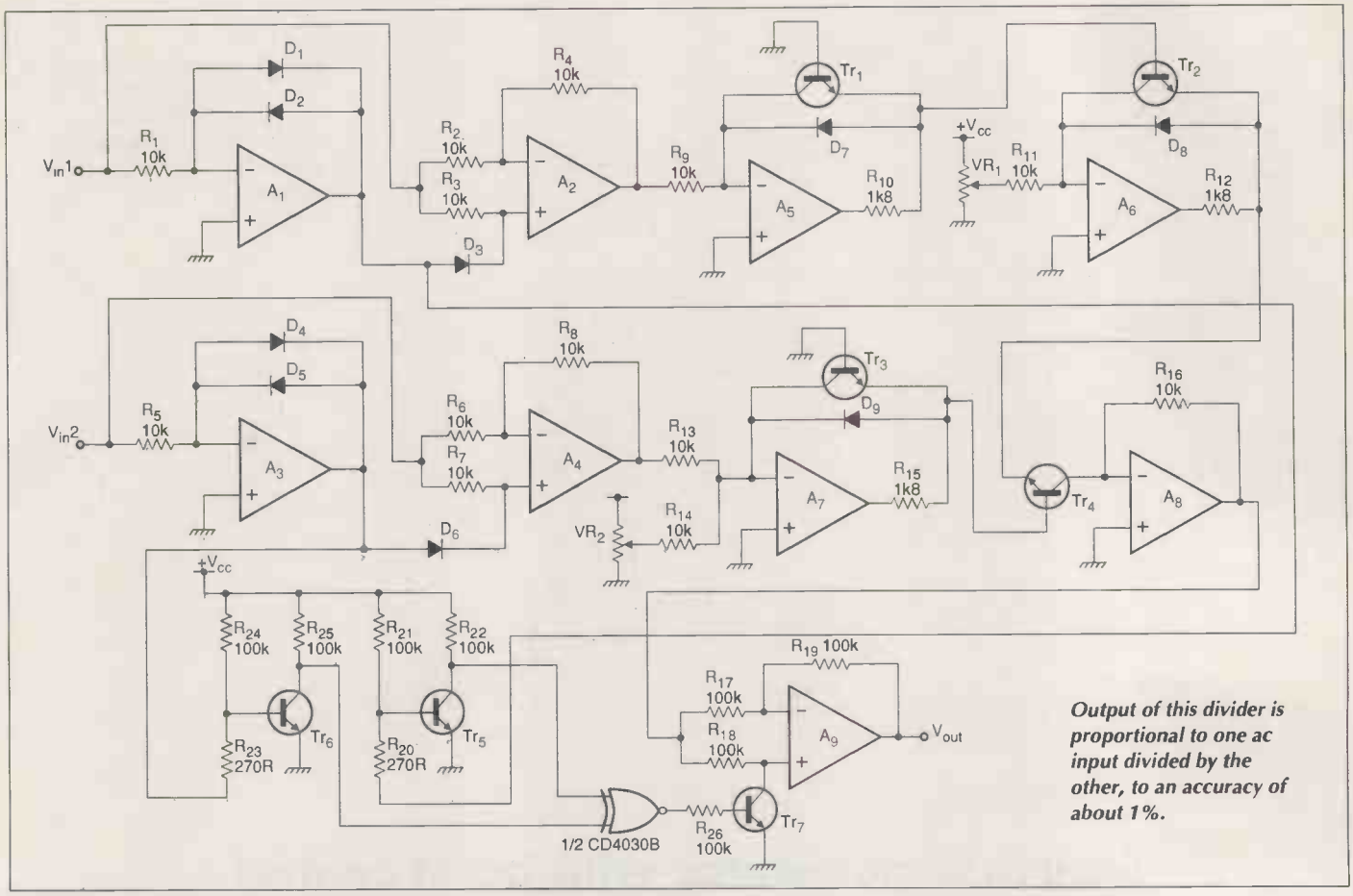
On light loads, the switching waveform is so narrow that its width is little more than that of the glitch, widened by the *RC*. The internal comparator of the ic may now try to work during the negative slope of the pulse and cause erratic behaviour. Increasing the *RC* time constant in an effort to avoid this can throw the baby out with the bath-water, the switching pulse disappearing with the glitch.

The new circuit solves this problem. The positive slope of the gate pulse turns on Tr_1 through C_1, reducing the current during the glitch. This greatly improves stability at low loads. The values shown work for most converters from 20W to 400W.

Francesc Casanellas
Barcelona
Spain



Removing the current glitch in a switcher without large RC circuits, which will not work on light loads.



Output of this divider is proportional to one ac input divided by the other, to an accuracy of about 1%.

Alternating voltage division

Two alternating voltages produce an ac output proportional to one divided by the other, the proportion being adjustable. Accuracy can be as high as ±1% with well chosen components. Amplifiers $A_{1,2}$ and $A_{3,4}$ form two rectifiers for inputs $V_{in1,2}$, the second amplifier of each pair being alternately inverting and non-inverting as dictated by the diodes $D_{6,3}$ and the ±0.7V output of the first stage. The outputs are, therefore always positive-going.

Amplifiers $A_{5,6,7,8}$ form a direct-voltage divider, giving an output $k|V_{in1}|/|V_{in2}|$, where k is adjusted by VR_1. A voltage of 1-30mV is inserted by VR_2 to avoid the embarrassment of a zero voltage from V_{in1}, but this does introduce a bit error when V_{in2} is small.

The rest of the circuit is a dc-to-ac converter, which gives an output of kV_{in1}/V_{in2} with inputs of the same or opposite polarities.

Rather better performance can be obtained when the *LM324* op-amps are replaced by *LM308*s and the transistors by *2N2920*s.

Jihai Zhang
Hangzhou, China

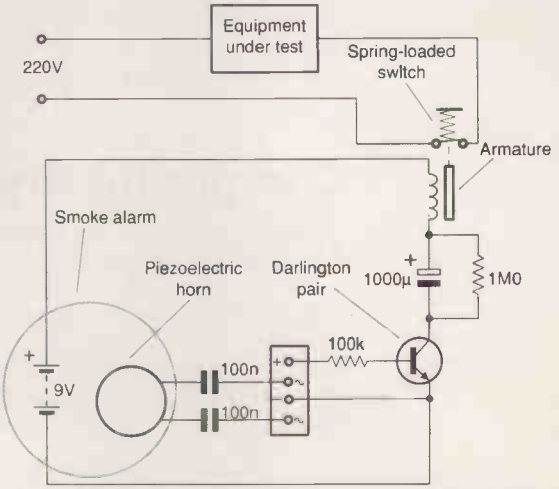
Smoke alarm removes power

There are occasions in which smoke alarms, while working to order, are ineffective because the sound is drowned by other noise or because people are unable to react for a variety of reasons. Again, relaying the alarm to the fire station imposes a delay during which the fire can take a hold. This circuit sounds an alarm in the normal way, but also disconnects power to electrical equipment, which is often responsible for the fire.

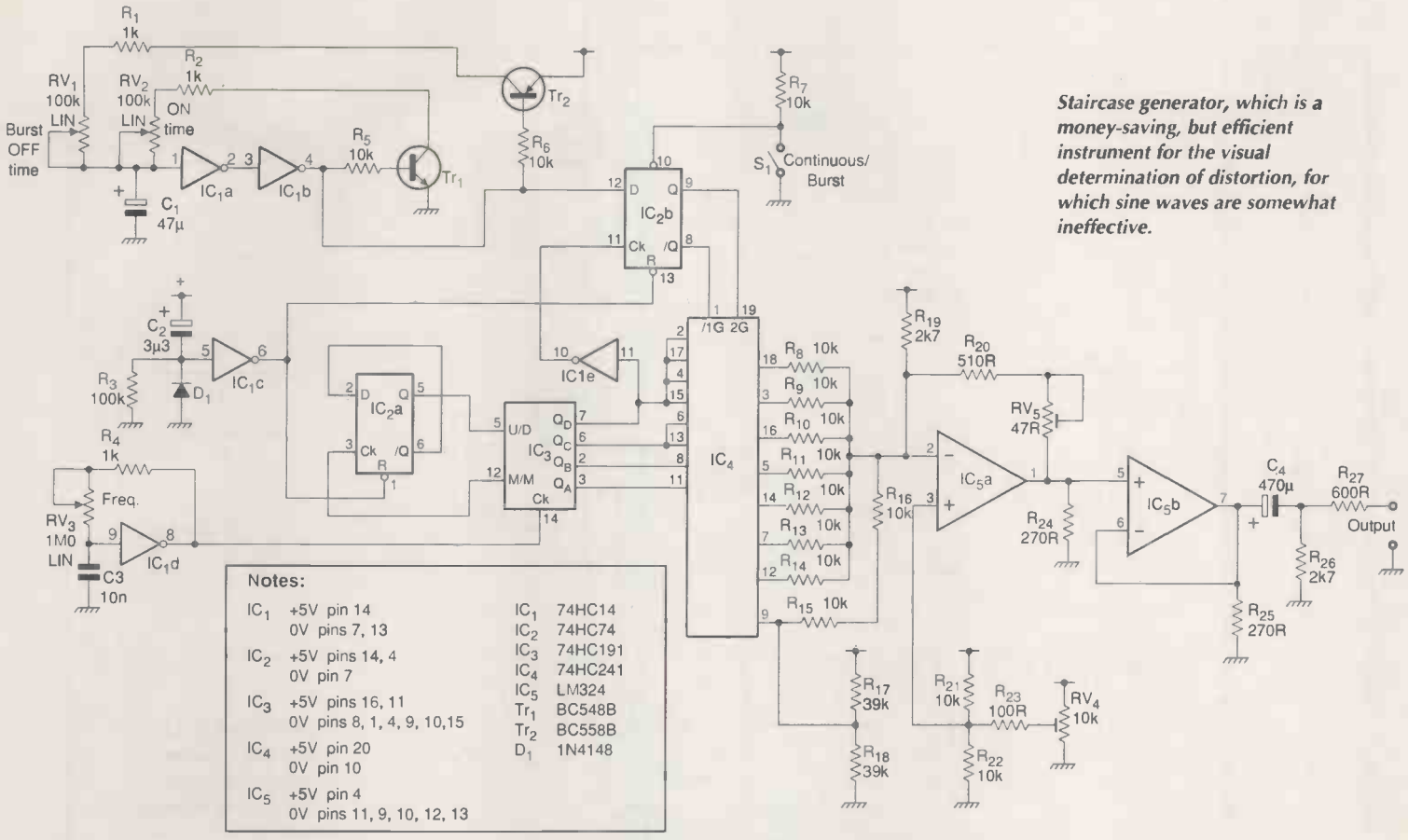
For example, in a laboratory where equipment undergoes extended, 24-hour testing, smoke on the alarm activates the piezoelectric transducer, the voltage across that being used to control a solenoid to break the supply circuit and, possibly, take more positive action such as turning on sprinklers.

A variation would be to allow an scr to pass a 30mA earth current so that an earth-fault switch would disconnect the supply.

Scott Arnesen
Oslo
Norway



Use this smoke alarm arrangement to isolate electrical equipment in case of fire.



Notes:

IC_1	+5V pin 14 0V pins 7, 13	IC_1	74HC14
IC_2	+5V pins 14, 4 0V pin 7	IC_2	74HC74
IC_3	+5V pins 16, 11 0V pins 8, 1, 4, 9, 10,15	IC_3	74HC191
IC_4	+5V pin 20 0V pin 10	IC_4	74HC241
IC_5	+5V pin 4 0V pins 11, 9, 10, 12, 13	IC_5	LM324
		Tr_1	BC548B
		Tr_2	BC558B
		D_1	1N4148

Staircase generator, which is a money-saving, but efficient instrument for the visual determination of distortion, for which sine waves are somewhat ineffective.

Staircase generator with burst control

As a visual aid to distortion estimation on an oscilloscope, a sinusoid is virtually useless, unless the distortion is gross. On the other hand, a staircase waveform readily shows quite small amounts of distortion and its steps may be used as markers to indicate an approximate level of the distortion.

This is a fairly standard circuit, but with some unusual features. A variable-frequency oscillator drives an up/down binary counter, the outputs being used in a digital-to-analogue converter followed by an op-amp output to give the staircase.

Weighted resistors to supply the currents from the octal buffer used as the converter are arranged so that only one value is needed, being combined in series or parallel to give an $R/4$, $R/2$, R and $2R$ sequence, on the assumption that the cmos outputs approach the supply rail closely enough for accuracy. Potentiometer RV_4 ensures that the summing node of the summing op-amp is precisely centred between the rails for best linearity of staircase.

In the tone-burst circuit, it is essential that burst switching occurs exactly at zero dc level to avoid dc shift on switching. In this case, the most significant bit of the counter is used as the trigger, which is as good as one can get. Tone switching is obtained by switching the whole converter on and off with the tri-state control, rather than trying to switch a small analogue signal. To get the on and off times, a low-frequency variable oscillator is used, the transistors avoiding the diode drop commonly found in the charge/discharge path of the timing capacitor.

A regulated 5V supply is essential for amplitude stability: a *7805* works well and an *LM317* even better.

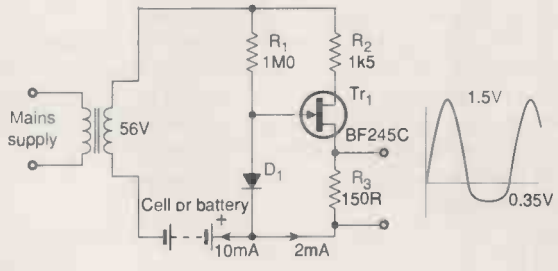
Sujit Liddle
New Delhi
India

Charger for dry cells or NiCds and batteries

Alternate charge and 20% discharge cycles enable this circuit to cope with most cell types or batteries, regardless of the number of cells in series or the battery voltage, so long as transformer and fet limits are not exceeded.

For a higher charge rate, use a fet giving a greater saturation current, adjusting the values of $R_{2,3}$ to give the new fet bias voltages.

Bob Philp
Luxembourg



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32K768 1MHZ 1M8432 2M000 2M1432 2M304 2M4576 3M000 3M2768 3M400 3M579545 3M58564 3M600 3M6864 3M93216 4M000 4M190 4M194304 4M2056 4M33614 4M608 4M9152 5M000 5M0688 6M000 6M041952 6M200 6M400 7M37280 8M000 8M06400 8M448 8M863256 8M8670 9M3750 9M8304 10M240 10M245 10M368 10M70000 11M000 11M052 11M98135 12M000 12M5 13M000 13M270 13M875000 14M000 14M318 14M7450 14M7456 15M0000 16M000 17M6250 18M432 20M000 21M300 21M400M15A 24M000 25M000 26M995 BN 27M045 RD 27M095 OR 27M145 BL 27M145 YW 27M195 GN 28M4696 30M4696 31M4696 31M4696 34M368 36M75625 36M76875 36M78125 36M79375 36M80625 36M81875 36M83125 36M84375 38M900 48M000 51M05833 54M1916 55M500 57M7416 57M7583 69M545 69M550 96M000 111M800 114M8 ..... £1 ea

TRANSISTORS

MPSA42 ..... 10/£1
MPSA92 ..... 10/£1
2N2907A ..... 10/£1
BC477, BC488 ..... 10/£1
BC107 BCY70 PREFORMED LEADS
full spec £1 £4/100 £30/1000
BC557, BC238C, BC308B £1/30 £3.50/100
2N2907 PLASTIC CROPPED £1/15 £4/100
BC548B SHORT LEADS £3/100 £20/1000

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OC29 ..... £2 ea
2SC1520 sim BF259 ..... 3/£1 100/£22
TIP 141/2 £1 ea TIP 112/42B ..... 2/£1
IRF620 TO-220 12A 200v ..... 2/£1
SE9301 100V 1DA DARL SIM TIP121 ..... 2/£1
BD680 ..... 4/£1
PLASTIC 3055 OR 2955 equiv 50p ..... 100/£35

TEXTOOL ZIF SOCKETS

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CNY65 OPTO ISOL 3000 available 50p
OPTO ICS also available TLP550 TLP666GF
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100 wa PLCC SKT 100 available ..... £1.50 each
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TL081 OP AMP ..... 4 for £1
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DIN 41612 96-WAY A/B/C/ SOCKET WIRE WRAP PINS £1.30
DIN 41612 64-WAY A/C SOCKET WIRE WRAP PINS £1
DIN 41612 64-WAY A/C PLUG PCB RIGHT ANGLE £1
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BANDOLIERED COMPONENTS ASSORTED Rs, Cs, ZENERS £5/1000
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1N4004 SD4 1A 300V ..... 100/£3
1N5401 3A 100V ..... 10/£1
1N5819RL 20K Ex stock ..... 1000 +10p
BA158 1A 400V fast recovery ..... 100/£3
BY254 800V 3A ..... 8/£1
BY255 1300V 3A ..... 6/£1
6A 100V SIMILAR MR751 ..... 4/£1
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6A 100V BRIDGE ..... 2/£1
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25A 200V BRIDGE £2 ..... 10/£18
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BY297 ..... 10/£1
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TIL81 PHOTO TRANSISTOR £1
TIL38 INFRA RED LED ..... 5/£1
4N25, OP12252 OPTO ISOLATOR 50p
PHOTO DIODE 50P ..... 6/£2
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FS22BW NTC BEAD INSIDE END OF 1 Inch GLASS PROBE RES 20°C 200R ..... £1 ea
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100n, 220n 63V 5mm ..... 20/£1 100/£3
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100n 250V radial 10mm ..... 100/£3
100n 600V Sprague axial ..... 5 for £1
2µ2 160V rad 22mm, 2µ2 100V rad 15mm ..... 100/£10
10n/33n/47n 250V AC x rated 15mm ..... 10/£1
1µ 600V MIXED DIELECTRIC ..... 50p ea
1µ0 100V rad 15mm, 1µ0 22mm rad ..... 100/£6
0.22µ 250V AC X2 RATING ..... 4/£1
0.22µ 900V ..... 4/£1

RF BITS

SAW FILTERS SW662/SW661 PLESSEY SIGNAL TECHNOLOGY 379.5 MHZ ..... £1.50 ea
FX3286 FERRITE RING ID 5mm OD 10mm ..... 10 for £1
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XTAL FILTERS 21M4 55M0 ..... £2 ea
ALL TRIMMERS ..... 3 for 50p
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P2N2222A PLASTIC ..... 10/£1
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10n 50V 2.5mm ..... 100/£4.50
100n 50V 2.5mm or 5mm ..... 100/£6
100n ax short leads ..... 100/£3
100n ax long leads ..... 100/£5
100n 50V dil package 0.3 Inch rad ..... 100/£8
1µF 50V 5mm ..... 8 for £1 £10/100

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12V 50watt LAMP TYPE M312 ..... £1 ea HOLDERS 60p ea
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MOONSHINE BIBLE 270 page book covering the production of alchohol from potatoes, rice, grains etc Drawings of simple home made stills right through to commercial systems. £15 ref MS1

NEW HIGH POWER MINI BUG With a range of 800 metres or more and up to 100 hours use from a PP3 this will be popular! Bug measures less than 1" square! £28 Ref LOT102.

SINCLAIR C5 MOTORS We have a new ones available without gearboxes at £50 ref LOT25

BUILD YOU OWN WINDFARM FROM SCRAP New publication gives step by step guide to building wind generators. Armed with this publication and a good local scrap yard could make you self sufficient in electricity! £12 ref LOT81

PC KEYBOARDS PS2 connector, top quality suitable for all 286/386/486 etc £10 ref PCKB. 10 for £65.

TRACKING TRANSMITTER range 1.5-5 miles, 5,000 hours on AA batteries, also transmits info on car direction and motion!Works with any FM radio. 1.5" square. £65 ref LOT101

ELECTRIC DOOR LOCKS Complete lock with both Yale lock and 12v operated deadlock (keys included) £10 ref LOT99

GALLIUM ARSENIDE FISHEYE PHOTO DIODES Complete with suggested circuits for long range communications\switching £12 complete.

SURVEILLANCE TELESCOPE Superb Russian zoom telescope adjustable from 15x to 60x! complete with metal tripod (imposible to use without this on the higher settings) 66mm lense, leather carrying case £149 ref BAR69

WIRELESS VIDEO BUG KIT Transmits video and audio signals from a minature CCTV camera (included) to any standard television! All the components including a PP3 battery will fit into a cigarette packet with the lens requiring a hole about 3mm diameter. Supplied with telescopic aerial but a piece of wire about 4" long will still give a range of up to 100 metres. A single PP3 will probably give less than 1 hours use. £99 REF EP79. (probably not licensable!)

CCTV CAMERA MODULES 46X70X29mm, 30 grams, 12v 100mA. auto electronic shutter, 3.6mm F2 lens, CCIR, 512x492 pixels, video output is 1v p-p (75 ohm). Works directly into a scart or video input on a tv or video. IR sensitive. £79.95 ref EF137.

IR LAMP KIT Suitable for the above camera enables the camera to be used in total darkness! £5.99 ref EF138.

REMOTE CONTROL TANDATA TD1400 MODEM/VIEWDATA Complete system comprising 1200/75 modem, auto dialler, infra red remote keyboard, (could be adapted for PC use?) psu, UHF and RGB output, phone lead, RS232 output, composite output. Absolute bargain for parts alone!!£9.95 ref BAR33

9 WATT CHIEFTAN TANK LASERS

Double beam units designed to fit in the gun barrel of a tank, each unit has two semi conductor lasers and motor drive units for alignement. 7 mile range, full circuit diagrams, new price £50,000? us? £349. Each unit has two gallium Arsenide injection lasers, 1 x 9 watt, 1 x 3 watt, 900nm wavelength, 28vdc, 600hz pulse frequency. The units also contain an electronic receiver to detect reflected signals from targets. five or more units £299 ea. £349 for one. Ref LOT4.

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NEW LOW PRICED COMPUTER/WORKSHOP/HI-FI RCB UNITS Complete protection from faulty equipment for everybody! Inline unit fits in standard IEC lead (extends it by 750mm), fitted in less than 10 seconds, reset/test button, 10A rating. £6.99 each ref LOT5. Or a pack of 10 at £49 90 ref LOT6. If you want a box of 100 you can have one for £250!

RADIO CONTROLLED CARS FROM £6 EACH!!!! All returns from famous manufacturer, 3 types available, single channel (left,right,forwards,backwards) £6 ref LOT1. Two channel with more features £12 ref LOT2.

THOUSANDS AVAILABLE RING/FAX FOR DETAILS!

MAGNETIC CARD READERS (Swipes) £9.95 Cased with flyleads, designed to read standard credit cards! they have 3 wires coming out of the head so they may write as well? complete with control elctronics PCB. just £9.95 ref BAR31

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ZENITH 900 X MAGNIFICATION MICROSCOPE Zoom, metal construction, built in light, shrimp farm, group viewing screen, lots of accessories. £29 ref ANAYLT.

AA NICAD PACK Pack of 4 tagged AA nicads £2.99 ref BAR34

PLASMA SCREENS 222x310mm, no data hence £4.99 ref BAR67

NIGHTSIGHTS Model TZS4 with infra red illuminator, views up to 75 metres in full darkness in infrared mode, 150m range, 45mm lens, 13 deg angle of view, focussing range 1.5m to infinity. 2 AA batteries required. 950g weight. £199 ref BAR61. 1 years warranty

LIQUID CRYSTAL DISPLAYS Bargain prices,
16 character 2 line, 99x24mm £2.99 ref SM1623A
20 character 2 line, 83x19mm £3.99 ref SM2020A
16 character 4 line, 62x25mm £5.99 ref SMC1640A

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GOT AN EXPENSIVE BIKE? You need one of our bottle alarms, they look like a standard water bottle, but open the top, insert a key to activate a motion sensor alarm built inside. Fits all standard bottle carriers, supplied with two keys. SALE PRICE £7.99 REF SA32.

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DAMAGED ANSWER PHONES These are probably beyond repair so just £4.99 each. BT response 200 machines. REF SA30.

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IBM PS2 MODEL 150Z CASE AND POWER SUPPLY Complete with fan etc and 200 watt power supply. £9.95 ref EP67

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1.44 DISC DRIVES Standard PC 3.5" drives but returns so they will need attention SALE PRICE £4.99 ref EP68

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DELL PC POWER SUPPLIES (Customer returns) Standard PC psu's complete with fly leads, case and fan. +12v,-12v,+5v,-5v SALE PRICE £1.99 EACH worth it for the bits alone! ref DL1. TRADE PACK OF 20 £29.95 Ref DL2.

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PASTEL ACCOUNTS SOFTWARE, does everything for all sizes of businesses, includes wordprocessor, report writer, window ing, networkable up to 10 stations, multiple cash books etc. 200 page comprehensive manual. 90 days free technical support (01342-326009 try before you buy!) Current retail price is £129, SALE PRICE £9.95 ref SA12. SAVE £120!!

COMPLETE PC 200 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. LAST FEW TO CLEAR AT £49 SAVE £30 ref LOT61

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

4.5mw LASER POINTER. BRAND NEW MODEL NOW IN STOCK!, supplied in fully built form (looks like a nice pen) complete with handy pocket clip (which also acts as the on/off switch.) About 50 metres range! Runs on 2 AAA batteries. Produces thin red beam ideal for levels, gun sights, experiments etc. just £39.95 ref DEC49 TRADE PRICE £28 MIN 10 PIECES

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VIEWDATA SYSTEMS made by Phillips, complete with internal 1200/75 modem, keyboard, psu etc RGB and composite outputs, menu driven, autodialler etc. SALE PRICE £12.99 REF SA18

AIR RIFLES .22 As used by the Chinese army for training puposes, so there is a lot about! £39.95 Ref EF78. 500 pellets £4.50 ref EF80.

PLUG IN POWER SUPPLY SALE FROM £1.50 Plugs in to 13A socket with output lead, three types available, 9vdc 150mA £1.50 ref SA19, 9vdc 200mA £2.00 ref SA20, 6.5vdc 500mA £2 ref SA21.

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***FM BUG BUILT AND TESTED** supenor design to kit. Supplied to detective agencies. 9v battery req'd. £14 REF: MAG14

TALKING COINBOX STRIPPER COMPLETE WITH COINSLOT MECHANISMS originally made to retail at £79 each, these units are designed to convert an ordinary phone into a payphone. The units have the locks missing and sometimes broken hinges. However they can be adapted for their original use or used for something else?? SALE PRICE JUST £2.50 REF SA23

GAT AIR PISTOL PACK Complete with pistol, darts and pellets £12.95 Ref EF82B extra pellets (500) £4.50 ref EF80.

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FIBRE OPTIC CABLE BUMPER PACK 10 metres for £4.99 ref MAG5P13 Ideal for experimenters! 30 m for £12.99 ref MAG13P1

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4X28 TELESCOPIC SIGHTS Suitable for all air rifles, ground lenses, good light gathering properties. £19.95 ref R/7.

GYROSCOPES Remember these? well we have found a company that still manufactures these popular scientific toys, perfect gift or for educational use etc. £6 ref EP70

HYPOTHERMIA SPACE BLANKET 215x150cm aluminised foil blanket, reflects more than 90% of body heat. Also suitable for the construction of two way mirrors! £3.99 each ref O/L041.

LENSTATIC RANGER COMPASS Oil filled capsule, strong metal case, large luminous points. Sight line with magnifying viewer. 50mm dia, 86gm. £10.99 ref O/K604.

RECHARGE ORDINARY BATTERIES UP TO 10 TIMES! With the Battery Wizard! Uses the latest pulse wave charge system to charge all popular brands of ordinary batteries AAA, AA, C, D, four at a time! Led system shows when batteries are charged, automatically rejects unsuitable cells, complete with mains adaptor. BS approved. Price is £21.95 ref EP31.

TALKING WATCH Yes, it actually tells you the time at the press of a button. Also features a voice alarm that wakes you up and tells you what the time is! Lithium cell included. £7.99 ref EP26.

PHOTOGRAPHIC RADAR TRAPS CAN COST YOU YOUR LICENCE! The new multiband 2000 radar detector can prevent even the most responsible of drivers from losing their licence! Adjustable audible alarm with 8 flashing leds gives instant warning of radar zones. Detects X, K, and Ka bands, 3 mile range, 'over the hill' 'around bends' and 'rear trap facilities. micro size just 4.25"x2.5"x.75". Can pay for itself in just one day! £79.95 ref EP3.

3" DISCS As used on older Amstrad machines, Spectrum plus3's etc £3 each ref BAR400.

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ACTIVE

A-to-D and D-to-A converters

24-bit delta-sigma a-to-d. From Burr-Brown, the *ADS1210/1211* wide-range, delta-sigma a-to-d converters with 24-bit resolution, using a single 5V supply. *1210* is a single-channel type and *1211* a 4-channel multiplexed device. Both have an SPI-compatible synchronous serial interface and two-wire control mode for low-cost isolation. Both also offer the 'turbo' mode to allow the selection of oversampling ratio to suit the application, from sampling at 20kHz to give 21bit at 10Hz to sampling at 320kHz to achieve 20bit at 1kHz. Burr-Brown International. Tel., 01923 233837; fax, 01923 233979.

A-to-d for video. AKM's *AK5482* is a 10-bit, 20MHz pipeline analogue-to-digital converter for use in the digitising of still colour video images such as photographs or scans. The 3V device offers differential linearity of ±0.5 lsb and integral linearity of ±1.5 lsb. Power consumption is 50mW at 20MHz. DIP International Ltd. Tel., 01223 462244; fax, 01223 467316.

Linear integrated circuits

Bias stabiliser. To stabilise the bias current of n-p-n bjts or n-channel fets, Motorola's *MDC5000T1* in an SOT-143 package allows the controlled device to have its emitter or source grounded while still working with a stable collector or drain current. It is mainly meant for use with rf stages on a low supply voltage, but is suitable for use with any linear stage to avoid the need for emitter/source by-passing while providing better control of bias over temperature and device parameter variations. Motorola Semiconductors. Tel., 01355 565000; fax, 01355 234582.

Rf cascodes. Motorola's *MRF1C0916* 900MHz general-purpose cascode SOT-143 amplifier is designed with internal chip-bias circuitry and off-chip matching for greater adaptability. Frequency range is 100-2500MHz with an output power of 2.3dBm at 1dB gain compression and a 2.7-5V supply. Small-signal gain is 18.5dB typical at 850MHz and reverse isolation 44dB typical. Tel., 01354 688040; fax, 01354 688248.

Microprocessors and controllers

Process controller. Athena's *XT32* Series of panel-mounted process controllers is a 1/32 DIN microprocessor-based indicating type providing closed-loop control of temperature or other quantities characterised by a linear input. It offers on-demand auto tuning and takes input from K, J and T thermocouples, rtds or linear inputs from other devices. There is a large display, dual output, selectable input, alarms and 'bumpless' auto/manual transfer. Hysteresis is adjustable and the quantity displayed is selectable. Athena Controls Ltd. Tel., 0161 4853536; fax, 0161 4853537.

Microprocessor reset. Maxim has the *MAX6315* microprocessor-reset chip, which emits a reset signal when the supply voltage falls below a preset internal threshold, maintaining the reset for a programmed, fixed time after the supply is restored. Thresholds are available in 100mV increments between 2.5V and 5V and there are four reset times from 1ms to 1120ms. The device ignores short transients and it includes a debounced manual-reset input. Maxim Integrated Products UK Ltd. Tel., 01734 303388; fax, 01734 305511.

Motors and drivers

Pwm motor controller. *UC3638*, by Unitrode, provides control of torque, velocity or position in dc motors, and drives Class-D amplifiers for audio and uninterruptible power supplies. It contains all necessary circuitry to generate an error signal and to modulate two bi-directional pulse trains in proportion to the error signal magnitude and polarity. Its features include a programmable, high-speed triangle oscillator, a differential current sensing amplifier with a gain of five, an error amplifier, pwm comparators, open-collector and ±500mA totem-pole outputs. Unitrode (UK) Ltd. Tel., 0181-318 1431; fax, 0181-318 2549.

Oscillators

Clock oscillators. Hy-Q's new range of oscillators is said to address the shortcomings in quality and delivery found in other makes. Three standard temperature ranges of 0 to 70°C, –30 to 75°C and –40 to 85°C are available in stabilities from ±100ppm to ±15ppm at frequencies in the 1-70MHz range. Output is compatible with HCMOS and ttl and there is a tri-state option at no extra cost. Hy-Q International (UK) Ltd. Tel., 01223 834444; fax, 01223 834589.

Lowest-cost PICs. One-time-programmable, 8-bit microcontrollers, *PIC16C923/924* are said to be the lowest-cost devices of their type. They combine an 8MHz clock speed and 500ns cycle time with a 4K-by-14 on-chip eprom program memory and 176 by 8 general-purpose registers. They also have 60 special function hardware registers, an 8-level deep hardware stack, interrupt, 25 i/o pins, pwm output and an SPI/I^2C synchronous serial port, in addition to a 5-channel, 8-bit a-to-d converter and a programmable lcd timing module. This new *PIC16C9XX* family is supported by the *PICMASTER* development system. Arizona Microchip Technology Ltd. Tel., 01628 851077; fax, 01628 850259.

Power semiconductors

Mosfet driver. A half-bridge, n-channel power mosfet driver, the *LT1336* from Linear Technology, has an on-board boost switching regulator to maintain high-side gate drive voltage at high duty cycles, including 93%-100%. The top-side driver is a floating-gate drive with no direct ground path, using rails up to 60V, and the internal boost switching regulator generates the floating high-side driver output voltage at 10.6V above the high-voltage rail to ensure enhancement of standard threshold mosfets. The device will drive into 10,000pF. Micro Call Ltd. Tel., 01844 261939; fax, 01844 261678.

PASSIVE

Passive components

Dual varicaps. Zetex's *ZDC833A* is a dual variable-capacitance diode in one SOT-23 package. The diodes exhibit a hyperabrupt *C/V* characteristic and show a large capacitance change for a small voltage input. Typical capacitance is 33pF and capacitance ratio 5 minimum for a 2-20V voltage. Q factor is a minimum of 200 at 50MHz and 3V reverse bias, which represents a series resistance of 0.5Ω. Matching of diodes in one package is within 0.25%. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5467.

Surface-mounted resistors. Ralec surface-mounted chip resistors come in a range of sizes from 0603 to 2512, with values down to 0.01Ω and in tolerances from 5% to 0.5%. The resistors are on a high-alumina substrate, the resistance element being of epoxy-coated ruthenium oxide, with nickel and solder-plated terminals. Legacy Distribution Ltd. Tel., 01243 533041; fax, 01243 536772.

Audio products

Stereo a-to-d. *AK4320*, an AKM 1-bit stereo digital-to-analogue converter, operates at three sampling frequencies: 32, 44.1 and 48kHz, and two master clock frequencies of 256 and 384f_s. It has a 20-bit oversampling filter and switched-capacitor filtering for the output. Dynamic range is

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100dB; s:n ratio 110dB. DIP International Ltd. Tel., 01223 462244; fax, 01223 467316.

Connectors and cabling

Surface-mount connectors. *PAK-5* board-to-board, surface-mounted connectors by Robinson Nugent have 'floating' contacts on the receptacle to take up torsional and up to 0.3mm of lateral position errors. The connectors come in sizes from 20 ways to 100 ways in 0.5mm pitch. Stacking heights are 3-8mm and a positive click is provided by the locking mechanism to achieve resistance to vibration and shock. Insulation resistance is 1000MΩ, dielectric withstand 150V ac, current rating 0.2A per contact and rated voltage 60V ac or dc. Robinson Nugent (Europe) Ltd. Tel., 01256 842626; fax, 01256 842673.

Optical/electrical connector. Radiall makes the *BOC 1/2* series of connectors for installing single-mode optical cable or mixed optical and electrical links in harsh conditions; the watertight system has been used for retransmitting hdtv programmes. Two units are made: the *BOC1*, with a screw locking method, has four channels to be fitted with all optical or mixed optical/electrical channels in any arrangement; *BOC2* takes two of each type of channel and uses a push-pull locking system. Cables from 7mm to 11mm diameter may be used. Optical insertion loss for a plug/adaptor/plug connection is 1.2dB and for a plug/receptacle 0.6dB. For electrical channels, rating is 16A and 8mΩ contact resistance. Transradio Ltd. Tel., 0181-997 8880; fax, 0181-997 0116.

Displays

Tough monitors and terminals. Production samples of the Regisbrook ruggedised display monitors and terminals are now available. They can be fitted with any lcd or electroluminescent display, an integral analogue touch switch and support electronics and power supplies. These monitors have line drives to allow connection to remote analogue and digital sources at a distance of over 50 metres – soon to be 100 metres. Packaging is stainless steel or powder-coated metal, a plastic seal rendering the equipment waterproof, even when partly submerged. Regisbrook Group Ltd. Tel., 01235 554433; fax, 01235 528971.

Graphic lcd. Epson's *SEK 1018 BOA* graphic lcd module has a viewing area of 97.12-by-74.08mm and is not much bigger overall, taking 2mA without the backlight and 20mA with it on. A *SEK 1330* lcd controller is in the package and is compatible with 80-series and 68-series microprocessors. Hero Electronics Ltd. Tel., 01525 405015; fax, 01525 402383.

10.3in colour lcd. Densitron Perdix announces its new display screen, which costs less than any of its others. *LMG8343E-DF2* is a 10.3in, 640 by 480 type, the whole package measuring 264 by 183 by 10.5mm and has a ccfl backlight to give a surface brightness of 75cd/m^2 and contrast ratio of 30:1. Response time is 270ms. *PCX535* is a matching vga controller. Densitron Perdix. Tel., 01959 700100; fax, 01959 700300.

Filters

Video filters. Faraday announces a range of single-in-line active video filters. Having a cut-off rate of 1.45, they are low-pass, phase-equalised designs intended as antialiasing and reconstruction filters in video and data conversion. Pass-band widths are 2MHz to 20MHz and the devices have high input, low output impedances or as specified by the customer. Gain is selectable at 0dB or 6dB. Faraday Technology Ltd. Tel., 01782 661501; fax, 01782 630101.

Hardware

Chip coolers. Chip coolers with fans and heatsinks by Sanyo Denki in the *San Ace MC* range are designed for use with cpus such as the Pentium family, but are equally at home with other types of semiconductor such as dsp circuitry and power devices. They come in four sizes between 45mm square and 66-by-62mm, each being available for 5V or 12V supplies. Coolers are quiet at around 28dB(A), locked-rotor protection is present and an alarm output is provided. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Emc-compliant chassis. Elma's *Series D* pc chassis is a complete enclosure, ready to use, designed to meet 89-336 EWG requirements without compromising appearance and cost. It is 4Y high, 84T wide and 448mm deep and is available in versions with four or eight slots. Aperture size is smaller and a CE-marked power supply is fitted. Overpressure cooling is used, outlets being designed to ensure cooling of all slots and, in particular, the hard disk drive; a temperature sensor controls the fan speed and an alarm is fitted. Radiatron Components Ltd. Tel., 01784 439393; fax, 01784 477333.

Membrane illumination. Wasp is now able to provide various type of illumination for front-panel membranes. Leds can be incorporated into the flexible membrane or, for overall lighting, leds or filament lamps can be spread about the membrane and the light conveyed by light paths in the membrane to give a uniform output. For low-level light, an electroluminescent layer is sandwiched in the membrane. Wessex Advanced Switching Products Ltd. Tel., 01705 453711; fax, 01705 473918.

Cabinet cooling. The Meech-ARTX *Control Cooler* prevents cabinets full of electronics from overheating, being a low-cost air-conditioning system powered by factory compressed air and having no moving parts. It will provide cooling capacity of 2500Btu/h, which represents a cabinet of 1.8 by 1.8 by 0.6m and is an alternative to a blower method of cooling with its attendant risk of air-blown dust; it is mounted in a standard knockout to keep IP65 rating. Operation is by the conversion of compressed air into two streams, hot and cold, the hot stream exhausting to atmosphere and the cold, which is 34°C colder than the supply, goes into the cabinet to be distributed by a manifold. Meech-ARTX Ltd. Tel., 01993 706700; fax, 01993 776977.

Computer safe. If you consider your computer to be in peril, you will welcome this safe. It is designed to take mini-towers up to 350mm high, 220mm wide and 500mm deep, is made from 2mm thick steel and has a seven-lever lock on the inset door. Its hinges are concealed and the safe is of welded construction, a dog-bolting arrangement preventing the removal of the door unless it is first unlocked. In the event that an unusually dogged thief decides to walk off with the safe itself, it has holes in the floor to take bolts. Ventilation is provided. Intek Electronics Ltd. Tel., 01352 810603; fax, 01352 810403.

Test and measurement

Digital/analogue audio analyser. Rohde & Schwarz has the *UPL* audio analyser for analogue and digital or combined audio analysis, having generators and analysers for dual-channel measurement and an integrated pc which therefore needs no keyboard or monitor. Functions include FFT analysis, jitter analysis, interface testing, programmable filtering, automatic test sequences and drivers for all commercial printers. Results are processed by the internal pc, which stores them for later use. Rohde & Schwarz UK Ltd. Tel., 01252 811377; fax, 01252 811447.

Thermal imager. *ThermaCAM SC1000* from Inframetrics gives full-screen temperature measurement to within ±2% or 2°C. Improvements mean that the camera uses less power, a standard camcorder battery lasting two hours with a battery belt for twelve hours as an option, and a new colour viewfinder provides better resolution; there is also a 4-in colour lcd viewfinder option. The 12-bit video output interfaces with company's *ThermaGRAM PRO 95* Windows 95-based software to allow a number of storage and analysis functions. Lens and filter options are many, one of them being a 15µm-resolution microscope. Inframetrics Infrared Systems Ltd. Tel., 01256 50533; fax, 01256 50534.

Clip-on millammeter. *mA-2000* from F W Bell provides non-contact current measurement of ac and dc.

Measuring ranges of this 3.5-digit, hand-held instrument are 0-200mA and 0-2000mA ac or dc up to 100kHz and an analogue output is included for oscilloscope or recorder. Resolution is 0.1mA and accuracy 1% of reading on dc, 2%-4% for ac, depending on frequency. Magnetics Consultants. Tel., 0191-528 4408; fax, 0191-515 2837.

Emc testing. Seaward has a range of instruments for emc testing, the latest member of the family being the *Orb* harmonics and flicker meter, which is for conformance test of single-phase equipment at up to 16A, carrying out Class A, B and C Fourier harmonics test up to the 40th. Also in evidence: the *Thor* surge generator, testing for immunity to emi to IEC 1000-4-5, again for conformance testing. Its range of output voltages simulates surges of the type caused by lightning and other sources, the software enabling its use by technicians. Sceptre is a pc-controlled spectrum analyser for the 150kHz-450MHz range of emissions, equipped with a line stabiliser. Finally, the *Mace* mains interference simulator, microprocessor-controlled to give three test routines in the one instrument. Seaward Electronic Ltd. Tel, 0191-586 3511; fax, 0191-586 0227.

Literature

ITT on CD-rom. Integrated circuits, discrete semiconductors and Hall sensors from ITT are all described on a new cd-rom catalogue for pcs and Macs, which also shows data sheets. Graphics and text may be printed out and may be copied to other, compatible software packages for inclusion in users' own documents. For screen display, the CD contains a copy of Adobe *Acrobat Reader 2*. The cd has not ousted paper, which is to continue. ITT Semiconductors. Tel., 01932 336116; fax, 01932 33148.

Servoamplifiers. Copley's new catalogue contains information on 60 brush and brushless amplifiers for motion control, in powers from a few watts to 20kW, and accessories including transformers, power supplies and mounting hardware. A range of techniques is employed: tachometer, encoder, resolver and Hall feedback, some of them multi-axis types, and there are low-cost types. The company's pulse-width modulated power amplifiers are also described. Copley Controls. Tel., 001 617 329 8200; fax, 001 617 329 4055.

Clean rooms. Cleaning the clean room is not, apparently, simply a matter of skipping round with a duster and a can of spray polish. So esoteric is it, in fact, that MVI has produced a video on the subject: *Preparing to clean the clean room* not only shows how it should be done and how to prepare the materials, but also how to prepare the people who are going to do it. Micron Video International Ltd. Tel., 01705 670550; fax, 01705 670543.

Snap-action switches. Matsushita has a new brochure to describe a range of snap-action switches which have their mechanism sealed in rubber and the terminals in epoxy resin to IP67/IP50. Ratings are 3A at 250V ac to 1mA at 24V dc, life span being over 500,000 operations. Pin-plunger, hinge lever or roller lever actuators are available. Matsushita Automation Controls Ltd. Tel., 01908 231555; fax, 01908 231599.

Materials

Flexible ferrite. Flexible film of ferrite polymer composite, made by Siemens, makes it possible to produce cores of exotic shapes, previously not realisable. For, example, reticulated flange cores for car immobilisers or non-welding pot cores for inductive proximity switches can all be made from the material. Now there is a film developed by Siemens and Matsushita which uses the ferrite shielding effect for emc applications. Other advantages of the film include magnetic stability, lightness and mechanical strength. At 25°C, relative initial permeability is 9; rel. dissipation factor <0.005 at 10MHz and <0.4 at 1GHz; resistivity 500Ω and specific dielectric constant 700, both at 1kHz. Free samples may be obtained from Siemens Response Centre on 0345 000 444. Siemens plc. Tel., 01344 396313; fax, 01344 396721.

Production equipment

Braid cutter. If, when cutting polyester braided sleeving, you find that it instantly turns itself into a mass of fibres, here is the answer. *Sealsnip* looks like a hacksaw, but the 'blade' is a hot wire that cuts all sizes of sleeving up to 30mm and gives the cut a neat, welded edge. There is a separate transformer for mains power and spare hot wires are provided. Systems and Electrical Supplies Ltd. Tel., 01734 873461; fax, 01734 752124.

Power supplies

Ups management. All Fiskars's uninterruptible power supplies are now complete with software to manage its affairs during a long power cut. *LanSafe III* or *FailSafe III* packages saves all data and performs a graceful shutdown of the system even when work has not been saved or if the computer is unattended. On detecting a power cut, the systems can handle fax and e-mail and will monitor, test and re-boot automatically. The package consists of the ups and a cd-rom with the software, installation data and on-line help. Fiskars Electronics Ltd. Tel., 01734 306600; fax, 01734 305868.

Fibre termination. *JTK-4000 Universal Fibre Termination Kit*, from Jensen, combines a basic kit of tools and supplies, which may be extended with tools specific to a given type of installation. Any techniques or connectors may be used. There is also the *Benchtop Fibre Tool Kit*, which is intended for the termination of AMP Light Crimp fibre connectors. Jensen Tools. Tel., 0800 833246 (free); fax, 01604 785573.

Pentium power. *MP55C* by Semtech is a voltage regulator module for the Intel *P55C* processor, which fits existing socket headers and integrates regulator, heatsink, capacitors, resistors and 30-pin connector in a form specified by Intel. The *P55C* uses split-voltage supplies and the use of the regulator avoids the need to redesign power supplies for up-dating a motherboard. Semtech Ltd. Tel., 01592 773520; fax, 01592 774781.

Hybrid regulators. Allegro announces a family of switched-mode dc-to-dc converters using hybrid ic techniques. *STR-7000* and *STR-7100* series and the *SI-8020* controllers are available for outputs of 5, 12, 15 and 24V at 6A and 12A. Input range is 11-40V for the 5V units and 30-50V for 24V types. Separate chopper excitation is used and there is provision for adjustable constant-current protection and externally-set foldback overcurrent handling. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.

Thrifty regulators. Toko intends its new range of regulators, the *TK112/113AM* series, for use in equipment spending much of its life on standby, an on/off control reducing consumption to 0.1µA. Voltage drop is 0.16V at 60mA. Output voltages are available in the 1.5-5.5V range, in steps of 0.5V. Cirkit Distribution Ltd. Tel., 01992 444111; fax, 01992 464457.

Bus regulator. *UC382* by Unitrode is a 2A, low dropout (450mV at 3A) linear regulator having a very fast transient response that, with a 3A/µs output current transient, passes only 12mV output voltage change. Separate bias and V_{in} pins are provided, the latter supplying the output transistor only. The 5-pin package allows Kelvin sensing, eliminating the effects of lead and trace resistance. Output voltage is 1.2-5V. Unitrode (UK) Ltd. Tel., 0181-318 1431; fax, 0181-318 2549.

250W, quad output. From Astec, the *LPQ250* series of BABT-approved, quad-output, 250W supplies, all with power-factor correction and contained in a U-channel extrusion. Inputs can be ac or dc at 85-264 and 120-370V respectively. Two models provide three low-voltage rails each with an adjustable floating output of ±5-25V. There is an emi filter and power-fail and remote inhibit facilities, as well as full protection. Chloride Powerline. Tel., 01734 868567; fax, 01734 755172.

375W, pfc supplies. Power-One's *PFC 375* series of power supplies are now CE-marked, taking in the Low-voltage Directive and the 89/336/EEC directive. All have power-factor

correction and optional fans and are contained in a package measuring 266.7 by 127 by 63.5mm. They are fully regulated with remote sensing on the main output, up to four outputs being of 5-48V and, unusually, 24V 10A. Power-One Europe. Tel., 01769 540744; fax, 01769 540756.

Radio communications products

1.8-2.4GHz power amplifiers. Three amplifiers from Anglia cover the 1800-2400MHz PCN/PCS communications bands. *ACAM 7690* is a mains-powered, rack-mounted type having a minimum power output of 10 with 1dB compression and flatness of ±1dB, power gain 40dB and third-order intercept point at 52dB. *ACAM 7915* is a cased version, using 15V at 5.5A, and covers 1700-2000MHz with minimum power gain of 30dB, ±1dB flatness and third-order intercept at 50dBm. *ACAM7963*, for 1800-2000MHz, is cased, using 15V, and having a minimum output power of 5W (1dB compression) and minimum power gain of 20dB ±1dB. Anglia Microwaves Ltd. Tel., 01277 630000; fax, 01277 631111.

Steel circuits. A new service from Corintech – the design and manufacture of thick-film circuits on stainless steel bases. They are mainly for sensors, high-power circuitry or circuits needing a very rigid fixing, such as those on an engine. CorinTech Ltd. Tel., 01425 655655; fax, 01425 652756.

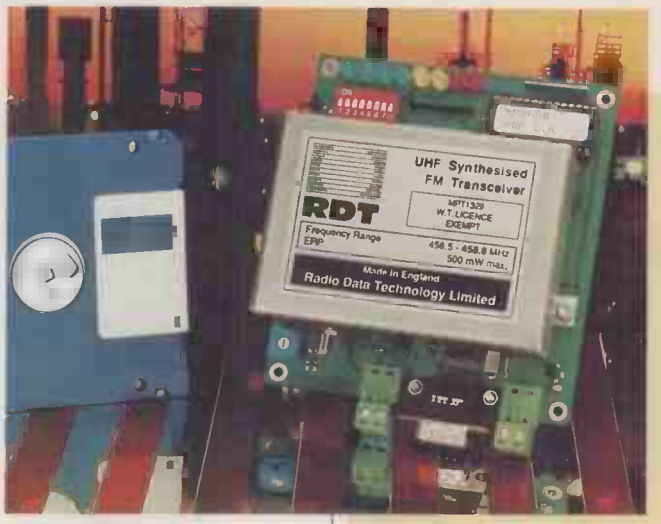
Switches and relays

Monitor relay. In those processes using multi-element heating, it is often necessary to ensure that all elements are carrying current. For this task, Crydom offers the *SMR System Monitoring Relay*, a standard 25A, 50A or 90A relay modified to take intelligent monitor circuitry to check current flow, line voltage, relay control voltage and other quantities. If a fault is present, an alarm output is activated and a led indicator shows. Crydom Europe. Tel., 0181-763 0550; fax, 0181-763 0499.

Double relay for cars. Siemens' *Double Mini Relay* is meant specifically for use in cars for immobilisers, sun roofs and seat adjusters. It has two separate 12V coils, the changeover contacts carrying 20A at 12V dc. The pcb-mounted case is 17 by 16 by 13mm. B&R Controls. Tel., 01279 443351; fax, 01279 415481.

Television components

Channel 5 retuner. In what seems to be an obvious answer to the problem of retuning many millions of video recorders and other equipment to avoid interference from Channel 5 television, Pace has introduced a retuner module that shifts the television signal to an unused part of the spectrum (channel 69) also providing a bonus by way of gain. The device plugs into a mains wall socket and connects by standard coaxial plugs and sockets. Pace Micro Technology. Tel., 01274 537082; e-mail, andrew.bone@pace.co.uk.



Transducers and sensors

Diff. pressure transducer. HBM's *Digibar* range of digital pressure transducers now includes a differential type, the *PDE300*, which has both digital readout and a form of analogue display, including min/max storage and trend. Ten ranges cover 100mbar to 2bar and the transducer is suitable for either battery power or two and three-wire (4-20mA) techniques, which give an analogue output for transmission to other locations and also limit relays for equipment control. HBM United Kingdom Ltd. Tel., 0181-420 7170; fax, 0181-420 7336.

Pressure transducer. Endevco offers the *8544-300M11*, which is a piezoresistive pressure transducer that will work at temperatures up to 177°C, being designed to operate inside engine transmissions; its Teflon cable is impervious to automatic transmission fluid. Temperature compensation is internal. Range is 0-300lb/in^2 and the device copes with burst pressure to 1000lb/in^2; output is 100mV full scale. Endevco UK Ltd. Tel., 01763 261311; fax, 01763 261120.

COMPUTER

Software

ChipLab for Windows. Data I/O has introduced a Windows interface for the *ChipLab* project programmer that also works with the company's *2700* programming system and will be made available for use with other Data I/O programmers shortly. The interface removes any need to consult handbooks and re-learn the system at each session, since it is completely intuitive and prompts are available at each step. Requirements are a 386 or better, 2Mbyte of extended memory, a 3.5in floppy drive, a parallel port, vga and at least 5Mbyte on the hard disk. Data I/O Ltd. Tel., 01734 440011; fax, 01734 448700.

Emc guidance. *Expert Consultant* from Seaward is upgraded to keep up with the latest European Directive on electromagnetic compatibility. The package is Windows-based and provides knowledge about emc and the implications of the directive for design of electronic equipment, test standards and routes to conformance. In addition, the program has been modified to ease its use and understanding. Minimum requirements: 386SX 25MHz; Dos 5.0; Windows 3.1. It also needs an 800 by 600 graphics card and 10Mbyte of free hard disk. Seaward Electronic Ltd. Tel., 0191-586 3511; fax, 0191-586 0227.

EN61000 testing. Voltech announces EN61000 Windows-based software to test equipment for EN61000-3 (EN60555) conformance. The standard is to do with current distortion and voltage fluctuation in ac power lines that may be caused by electrical equipment and this software allows all relevant tests to be performed quickly. It is meant for use with Voltech's *PM3000A* NPL-certified power analyser, software controlling both analyser and ac test source. Tests include steady-state and fluctuating harmonics, voltage change and voltage flicker, and there is an automatic Class D waveform check. Voltech Instruments Ltd. tel., 01235 861173; fax, 01235 861174. ■

Data communications

Radio modem. Radio Data Technology's *RM 9600* transceiver is said to be the world's fastest medium range, low-power radio modem, meant for use in wireless data and control links. It is a 500mW unit, working at 9600baud with forward error correction to allow programming and down-loading at the normal operating speed of a pc. The transceiver has both RS 232 and RS485 serial ports, so that the unit may be used for logging or for full-function, IEEE-compliant control. Operation is single-frequency, half-duplex in bands of up to 32 channels between 406MHz and 470MHz. Output power is adjustable in the 50-500mW range to minimise interference. Radio Data Technology Ltd. Tel., 01376 501255; fax, 01376 501312.

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5¼" CDC 94205-51 40mb HH MFM I/F RFE tested	£69.95(C)
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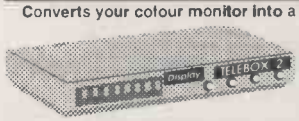

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TELEBOX ST for composite video input type monitors	£36.95
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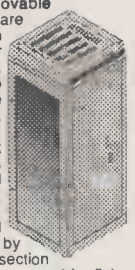

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CIRCLE NO. 139 ON REPLY CARD

Precise *frequency generation*

Nick Wheeler shows how you can derive almost any desired frequency up to 5GHz with quartz precision.

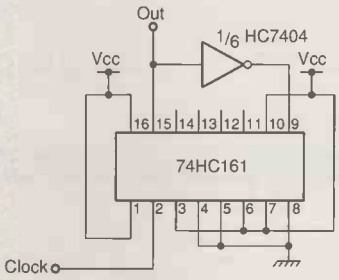

Fig. 1. The '161, shown here dividing by 7, has four program inputs for division to 16.

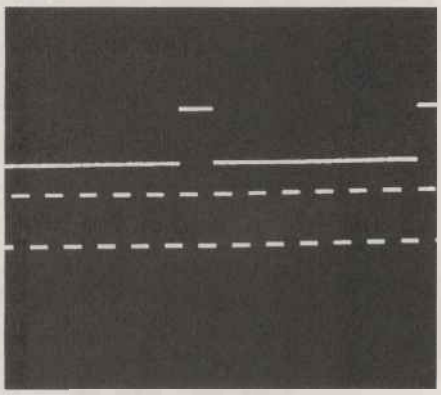

Fig. 2. Output versus clock input for Fig. 1.

With the exception of a few national standard broadcasts and satellite navigation systems, virtually all radio transmitters and many receivers derive their frequency control from quartz crystals. The frequencies are those of the crystal, or its harmonics, or are produced by voltage controlled oscillators, or vcos. With the vco, frequencies are compared, after digital division, with low frequency crystal oscillators. This is now the preferred method for most applications.

Cheap and accurate crystals are mass-produced for clocks and television applications and can be recognised in catalogues by the fact that the frequency is specified to four or even six decimal places. In this article I shall show how such crystals can be used to control almost any frequency.

General considerations

The phase locked loop, or pll, can take many forms. Possibly the most convenient form is based on the *HC4046* ic. The *4046* actually contains a useful vco, operable to about 15MHz, but this is not further considered below. Basically all systems work by comparing the phase of two pulse trains of the same nominal frequency to produce the voltage which is applied to the vco. If the vco drifts, this voltage changes and is arranged, in the manner of a servo loop, to return the vco to the correct frequency. There are two generally recognised approaches to phase comparison.

Phase comparison

Phase comparison between two pulse trains requires that the pulses be evenly spaced. This rules out the use of division schemes that, while the number of pulses output over a period may average out at the required number, and can therefore be counted, are unusable for pulse-by-pulse comparison. Comparators are usually described as type 1 or type 2.

Type 1 comparators require that the compared pulse trains require to be of close to unity mark-space ratio. In this case the comparator can be based on an exclusive-or gate.This is no problem where every divider chain has a divide-by-two last stage. Type 1 comparators are simple and have good noise rejection.

Type 2 comparators operate on pulse edge comparisons and are therefore insensitive to duty ratio. Many useful divider chains have duty ratios which have duty ratios equal to the division ratio. A type 2 comparator will work on a pulse train suitable for a type 1, but not *vice versa*.

In this article I shall only discuss how these two pulse trains, one derived from the crystal and the other from the vco, can be produced. There is a full coverage of pll design in ref. 1.

The question of accuracy needs to be considered. Ordinary, affordable, crystals are commonly specified to an accuracy of 50 parts per million, or sometimes at 20ppm. This is inconsistent with specifying the actual frequency to six decimal places. What this means in practice is that provided the division ratios chosen are accurate to within 20ppm then the resultant system is as accurate as can be expected.

Methods of implementation

Given that the end result desired is a particular frequency then by far the easiest method is to obtain a crystal which has a fundamental frequency which is a binary sub-multiple of the desired figure. Then a chain of a suitable number of divide-by-two stages is all that is required. If the desired frequency is below 100MHz this can usually be done with an AC part followed by an HC part. When the AHC family becomes readily available, these parts

are very nearly as fast as the AC types.

However, this convenient solution will almost always call for a non-standard crystal. These are readily available but have to be specially cut which takes time and, for small numbers, is relatively expensive. But always look in your supplier's catalogue to see if you are in luck, not necessarily exactly but within 20ppm. Also, a specially cut crystal will be subject to the faster ageing which affects new crystals whereas manufacturers presumably schedule the production of standard frequencies so that the worst of the ageing occurs before release.

For desired frequencies above 100MHz prescalers, which currently use emitter-coupled logic, ecl, will be required.

Table 1 lists some available types.

Table 1. ICs for dividing oscillator frequencies between 0Hz and 5GHz.

Type	Maker	Max freq	Min freq	Div Ratio
IFD-53010	H-P[1]	5.5GHz	0.15GHz	4
IFD-53110	H-P	3.5GHz	0.15GHz	4
SA 703N	Philips[2]	1.1GHz	DC[3]	128/129/144
SA 702N	Philips	1.1GHz	DC	64/65/72
SA 701N	Philips	1.1GHz	DC	64/65 128/129
SP 680B[4]	GEC Plessey	575MHz	10MHz	10/11

Notes
(1) Formerly Avantek.
(2) Other manufacturers offer equivalent parts.
(3) Minimum slew rate 32V/µs.
(4) This part has a ttl-compatible output.
All the others only have ecl output levels.

Achieving the required scaling factor

Cheap, accurate crystals are commonly available in the 2-10MHz range whereas phase comparison is more painlessly done in the hundreds of kilohertz region. So we arrive at the requirement that $X/n_1=O/n_2=F$, where X is the crystal frequency, O is the oscillator frequency and F is the operating frequency of the comparator.

To make what follows general I shall assume that O is greater than 100MHz necessitating the use of one of the prescalers listed above. This means that $n_2=P\times n_3$, where P is one of the available prescaler factors.

If the phase comparator is to operate at,say, 250kHz then crystals in the 2-10MHz region will require n_1 to lie in the range 8-40. Counters in this range can easily be made to operate at any integer value, in many cases using only one ic. They can also be made to work at many non-integer values. This is discussed below.

The high cost of the two divide-by-four prescalers means that they can only seriously be considered for use in the gigahertz region.

Frequencies and division ratios

It is not reasonable to contemplate the following method without the use of a computer. There are four variables in this problem, given the required value of output frequency. They are, in the terminology used above, X, the crystal frequency, n_1, the crystal frequency division ratio, P, the prescaler ratio and n_3, the post-prescaler division ratio.

A usable program takes the form of four nested FOR/NEXT loops which try all possible values of these variables against the criterion that X/n_1 lies within the range of $O/P/n_3$ ± the tolerance in ppm. A pc-compatible with a 75MHz Pentium executes a typical program in 35 seconds. If the desired frequency can be achieved exactly this will generally be possible with a large number of combinations of X, P and n_3. It is necessary to detect this in order to prevent the printer outputting reams of paper.

'All possible values' take the form of look-up tables of readily available crystal frequencies, entered directly from a suppliers catalogue, possible values of P as noted above and sensible integer values for n_1 and n_3.

If the program yields an odd number value for n_1 or n_3 then it may be necessary to add a divide-by-2 at the end of both chains, if a Type 1 phase comparator is to be used.

I ran an extremely lengthy program – taking 24 hours – which established that well over 95% of all frequencies between 100MHz and 1200MHz can be synthesised to better than 20ppm accuracy from at least one of the available crystal frequencies listed below. If the required frequency is an integer number of megahertz then some 85% of cases will be a 'direct hit' ie there will be zero tolerance – other than that of the crystal, of course) The remainder – non-integer targets – all fall within the 20ppm range.

Table 2. Megahertz values of the more readily available crystals.

2.2476	2.5	3	3.2768
3.567	3.577	3.579545	3.582
3.6864	3.7	4	4.096
4.194304	4.433619	4.608	4.9152
5.12	6	6.144	6.5536
7.3728	8	9	10
10.24	10.245	10.5	10.6985
10.7	10.7015		

Frequencies shown in Table 2 will be recognised as having widespread application in television, communications and clocks. As a result they are cheap and, since they come from long production runs, can be expected to be accurate and stable.

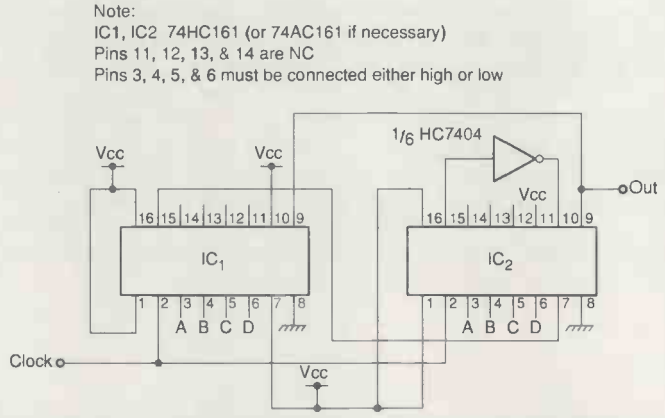

Fig. 3. Two xx161 parts cascaded allow programming of all division ratios up to 256 or, using this arrangement, division of all integers ratio to 241. Programming inputs are A, B, C and D.

Non-integer division ratios

If a solution with a good enough tolerance does not emerge, then a non-integer division ratio can be considered. This ratio will be the result of dividing one integer by another.

It should be said right away that one is here venturing into a potential minefield. It is possible to set up many non-integer dividing circuits, but many of the simpler solutions involve interpolation by analogue means such as delay lines or monostables.

Other solutions result in output pulse trains which, while the pulse count is indeed a non-integer function of the input frequency, have the property that the pulses are unevenly spaced and of differing lengths. Phase comparators cannot work on such inputs unless both are the same, which is unlikely.

It is easy to double the frequency of a pulse train, simply by using an inverter to produce positive (or negative) going signals twice per cycle. At this point an analogue element is necessary to generate pulses of half the duration of those in the original train. Again, this is easy and in the case of Type 2 comparators is undemanding. What this does mean, however, is that a limitation is placed on the range of frequencies which will work properly with a given arrangement of analogue timing parts.

The double-frequency, or if necessary a fur-

ther multiplied train, can now be divided by one of many easily achieveable integer ratios to produce a non-integer sub-multiple of the original. The result will almost always be suitable for Type 2 comparison only. Suitable doubler circuits can be found in the Circuit Ideas Pocket Books, available via *EW*.

Using the *xx161*

The literature contains many references to counters, but the approach which I have found almost universally usable is that based on the synchronous *xx161*. As the CLEAR function is not used in these circuits *xx163* parts may be used interchangeably. For applications up to 25MHz clock rate the HC type should be used. The AC type is good up to over 100MHz but should not be used unless necessary since the very rapid switching can cause emc problems. Both circuits work with HC and AC parts down to a few kilohertz clock rate. Though I have never tried this, the old *CD40161*, with its maximum clock rate of 2MHz with $5V_{DD}$ might even be better in this respect. These parts are still readily available.

The *161* part, described in Texas Instruments' terminology, is a programmable four-stage binary counter. It can be,

- Cascaded, without glitch problems, up to 18MHz in the HC version.
- Be hard-wired to yield any division ratio from 1 to 16 per chip.

The four programming pins, A,B,C and D have weights of 1, 2, 4 and 8 respectively. If the sum of the weights of those pins wired High is *N*, then the division ratio is (16–*N*).

Figure 1 shows the a circuit with A and D wired high (B and C low). Uniformly spaced output pulses at $F_{CLOCK}/7$ can be seen in the oscillogram of **Fig. 2**.

In the *161* part all transitions occur on the positive-going edge of the clock pulses, hence the duration of each of the output pulses, at the ripple-carry output, pin 15, is the clock period. A pulse train of this kind is suitable for Type 2 phase comparators.

Two *xx161* parts cascaded will yield all the binary division ratios up to 256 at the Q outputs, or, using the circuit of **Fig. 3**, every integer ratio up to 241. This circuit is taken from reference 2 but it will be found in practice that the division ratio is given by: $D=(256-N-15)$.

Where *N* is the sum of the weights of those programming pins which are high. The weights of the pins of IC_2 are 16, 32, 64 and 128. The circuits of Figs 1 and 3 have been thoroughly tested and the division ratios are as noted above. The output pulse will have a duration of one clock period, regardless of the value of *D*.

Conclusions

I have shown that unless you are very unlucky, almost any desired frequency can be synthesised accurately from commonly available as opposed to specially cut crystals. The higher the frequency the more likely this is to be so, since there is scope for dividing the vco frequency by larger integers while still maintaining a reasonable frequency of operation for the phase comparator.

The approach chosen is only made possible by the fact that fast pc compatibles are now readily and cheaply available. ■

References

1. The Art of Electronics, Horowitz and Hill
2. Don Lancaster, TTL Cookbook, Sams

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My hybrid power amplifier, which I call Vacusolid, consists of a solid-state phase splitter driving the push-pull valve output stage. It features an op-amp based integrator to avoid dc offset problems in the toroidal output transformer.

Push-pull valve output stage

A push-pull output stage is more expensive than a single-ended stage, but it offers a number of important advantages. Firstly, it provides much more power. According to Philips' 1965 data books, a class AB push-pull output amplifier with two *EL34*s operated as pentodes can deliver an output power of up to 100W – about eight times as much as that from a single-ended stage. This is due to class AB operation in conjunction with the high heat capacity of the electrodes, which allows the maximum nominal dissipation to be exceeded during signal peaks in this class of amplifiers. A *proviso* is that the bias current must be carefully chosen to avoid cross-over distortion.

A second advantage is that – if due attention is paid to accurate dc balance – there is no dc bias in the output transformer. This is due to the opposition of the magnetic fields in the primary windings. Furthermore, even harmonics cancel out if the output valves are carefully matched, leading to low open-loop distortion. This benefit is found especially with triodes but also to an extent in 'ultra-linear' designs. Finally, push-pull valve output stages have a high supply-voltage ripple rejection. In the ideal case – identical output resistance in both output valves, primary windings of output transformer fully balanced – the ripple on the supply voltage is completely suppressed. This is another reason for using well matched output valves.

Triode, pentode or ultra-linear design?

Figure 1 shows the distortion characteristics of a triode and a pentode. You will see that d_2, the second harmonic, is dominant in the case of a triode.

Since the second harmonics cancel out in push-pull operation, triodes give very low global distortion here. An additional advantage is that resonance of the output transformer, due to the leakage inductance and the winding capacitance, is effectively damped by the low output impedance of the triode. On the other hand, triode-based designs have the disadvantage of a low efficiency.

Push-pull output stages with pentodes give higher efficiency but also higher distortion, because d_3 is dominant in these valves. Stability with ac is poor too, because the high output impedance of a pentode means low damping of transformer resonance.

'Ultra-linear' designs give a compromise between triode and pentode operation by connecting the screen grids of the pentodes to an output transformer tap – preferably the 40% point. This yields about 65% of the maximum output power of a pentode output stage, while keeping the distortion and output impedance just about as good as with triodes.

Conventional phase splitter

The phase splitter delivers the two equal and antiphase input signals needed to drive the push-pull output stage. You could of course use a conventional valve design for this purpose, like that shown in **Fig. 2**. A high R_k in the common cathode circuit gives a good approximation to current-source operation. If

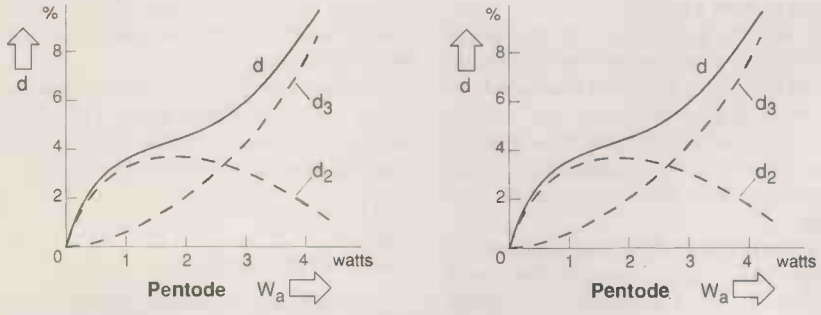


Fig. 1. Harmonic distortion – triode and pentode compared.

Wim de Jager is at University of Twente, Enschede, The Netherlands

the two anode resistances are matched, a balanced output is obtained.

The left-hand control grid is driven by a pentode preamplifier with dc coupling, while the right-hand control grid is grounded for ac via C_1. Voltage amplification is fairly low, because of the low μ of the triodes.

A typical valve amplifier incorporating such a valve-based phase splitter is described in reference 1. The low gain of the phase splitter means that a three-stage design is needed here, especially if part of the open-loop gain is used for global feedback.

Phase splitter with pnp transistors

The amplification factor μ of a bipolar transistor is typically about 30 times that of a triode. Maximum voltage gain can thus in principle be bigger too, allowing us to reduce the overall size of the amplifier from the conventional three stages to two.

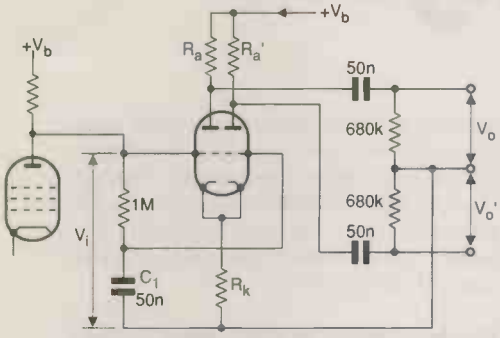


Fig. 2. Conventional phase splitter, R_k=68k, R_a=100k, A=V_o/V_i=25, thd=1.8%, V_o=25V_{rms}.

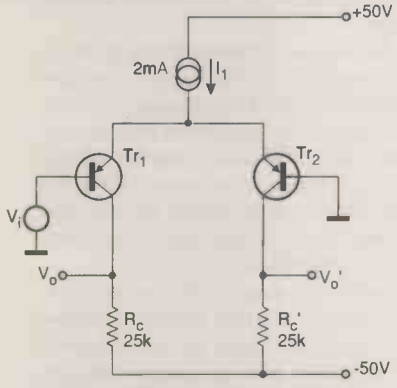


Fig. 3a. Solid-state alternative to Fig. 2.

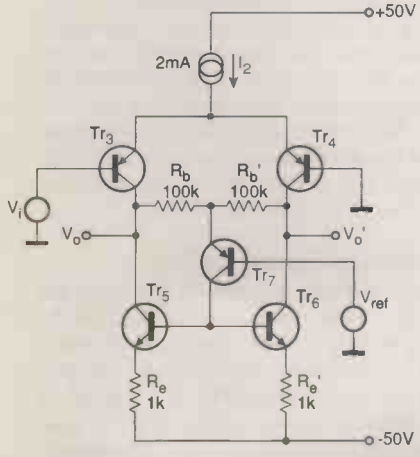


Fig. 3b. Improved solid-state phase splitter reduces distiortion and reduced influence on outpout gain.

Figure 3a shows a solid-state alternative for the circuit of Fig. 2. By opting for pnp transistors here, we make it possible to use the collector potentials of $Tr_{1,2}$ as the negative control-grid voltages for the output valves.

Using dc coupling in this circuit renders the inconvenient coupling capacitors found in a conventional valve-based design obsolete. The basic version of this circuit has two drawbacks, however. Firstly, distortion is about 25% at 90% full drive – much more than that of the output stage, and thus unacceptable. Secondly, the gain is coupled with the dc output level: if you adjust I_t or R_c to change the setting of the output valves, the gain will change too.

Both these problems can be solved by modifying the circuit as in **Fig. 3b**. This works as follows. Collector currents of $Tr_{3,4}$ activate $Tr_{5,6}$ via R_b, R_b' and Tr_7, which is connected in common-base configuration. As regards the dc setting, you then have $I_C(Tr_3)=I_C(Tr_5)$ and $I_C(Tr_4)=I_C(Tr_6)$. When Tr_3 and Tr_4 are at full drive, $I_C(Tr_5)$ and $IC(Tr_6)$ remain roughly constant, ie this configuration operates as a current source. The dc level of V_o and V_o' can be adjusted with the aid of V_{ref}. The maximum value of R_b is limited by the maximum permissible voltage drop across R_b, due to the base currents of Tr_5 and Tr_6.

We chose R_b=100kΩ in this design. As a result of this choice – and other factors – V_o is about 1.5V positive with respect to V_{ref}. This voltage can be used to adjust the desired negative grid voltage of the output valve without affecting the gain appreciably.

Gain, $A=V_o/V_i\cong800$, is mainly limited by the μ of Tr_3 and Tr_4, and is slightly reduced by R_b and R_b'. Emitter degeneration from R_e and R_e' is used on Tr_5 and Tr_6 to raise the output impedance and hence gain. Thanks to the low current modulation in Tr_3 and Tr_4, this circuit has a distortion of about 0.5% at 90% full drive – about a factor 50 better than the circuit of Fig. 3a.

Summarising, the high gain, low distortion and possibilities of dc coupling and dc adjustment offered by this circuit make it very attractive as the driver for a push-pull output stage. It also avoids the disturbances due to filament hum and microphony which can be troublesome in valve preamplifiers.

The integrator circuit

Use of toroidal transformers in the output stage[2] can give a large power bandwidth and improved stability of the global-feedback loop. Much of th is is due to their low leakage inductance. However, a toroidal transformer is more sensitive to core saturation due to dc bias than a transformer with conventional E/I laminations.

Direct coupling of a high-gain preamplifier aggravates the dc offset problem, and makes negative dc feedback necessary. In order to sense the cathode current of the output valves, cathode resistances are included in the circuit. These resistances are dimensioned to give a voltage drop of 400mV at 40mA (10W). This choice gives a voltage which is large enough for accurate processing, while limiting the mutual conductance of the output valve due to the current feedback to about 10%.

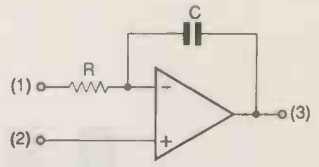


Fig. 4a. Op-amp based integrator used to provide dc negative feedback.

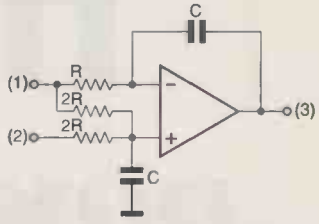


Fig. 4b. Integrator with a differential input.

I decided to use an active integrator with an op-amp to provide the negative dc feedback, in view of the high dc gain and low dc offset required. The principle of this integrator is shown in **Fig. 4a**.

The ac transfer from (1) to (3) in this circuit, with (2) grounded, is given by $A(1)\rightarrow(3)=-1/j\omega RC$ while that from (2) to (3), with (1) grounded, is $A(2)\rightarrow(3)$, which is $1+1/j\omega RC$. Presence of the additive term 1 in the second equation is a disadvantage for the intended application. This is because the output signals to the integrator are distorted due to the class AB operation of the output stage: if the amplifier is near full drive, these signals have more or less a single-sided rectified waveform. At low frequencies, as well as dc, negative feedback is produced.

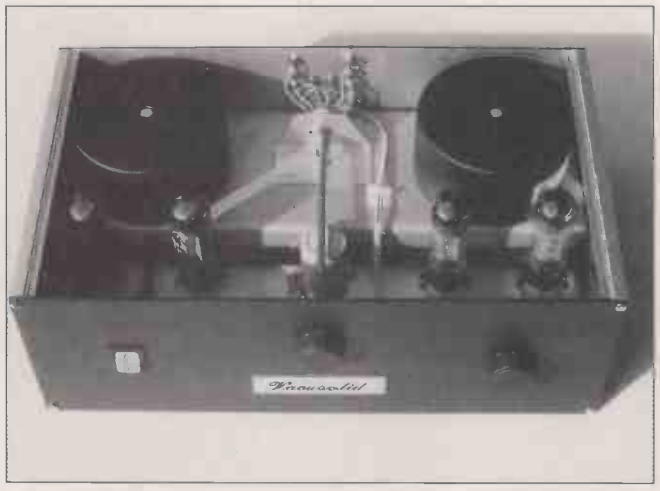
If signal (1) has a transfer function different from signal (2), this leads to distortion. The choice of very large values of R and C – representing a large time constant – can reduce the transfer at low frequencies. But the results are still unsatisfactory because of the presence of the additive term 1.

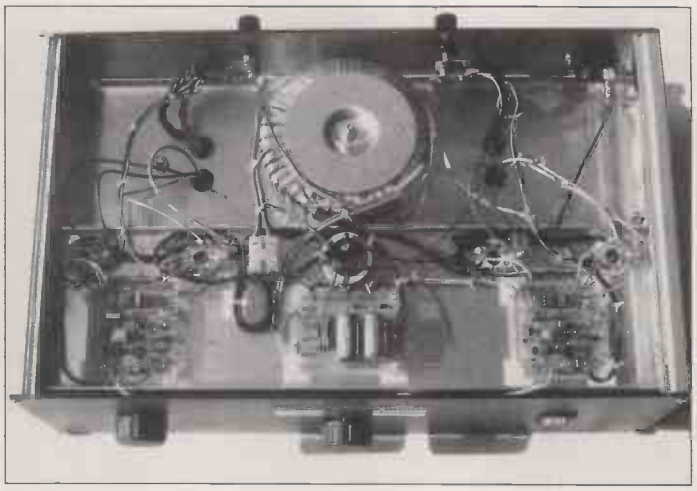
Using large time constant has another important disadvantage: the dc negative feedback is also active if the amplifier is overloaded. This gives a correction signal for dc balance which is much larger than under normal drive conditions. The result is a long recovery time: it takes the amplifier a long time to return to a normal setting after it has been overloaded.

An alternative here is to use an integrator circuit with differential input, **Fig. 4b**. By adding two resistors of value $2R$ and a capacitor of C oproduces the transfer functions,

$A(1)\rightarrow(3)=-1/2j\omega RC$
and,
$A(2)\rightarrow(3)=1/2j\omega RC$.

This integrator circuit suppresses low-frequency ac signals effectively, so that good





Photos 1, 2. Top and bottom views of the prototype power amplifier. Mounting the valves directly on the chassis offers benefits.

results are obtained with moderate values for R of 50kΩ and C of 100nF.

The complete amplifier

Figure 5 shows the complete amplifier. It is built round an ultra-linear push-pull output stage with two *EL34*s, an Amplimo type *VDV3070PP* toroidal transformer and a 470V power supply.

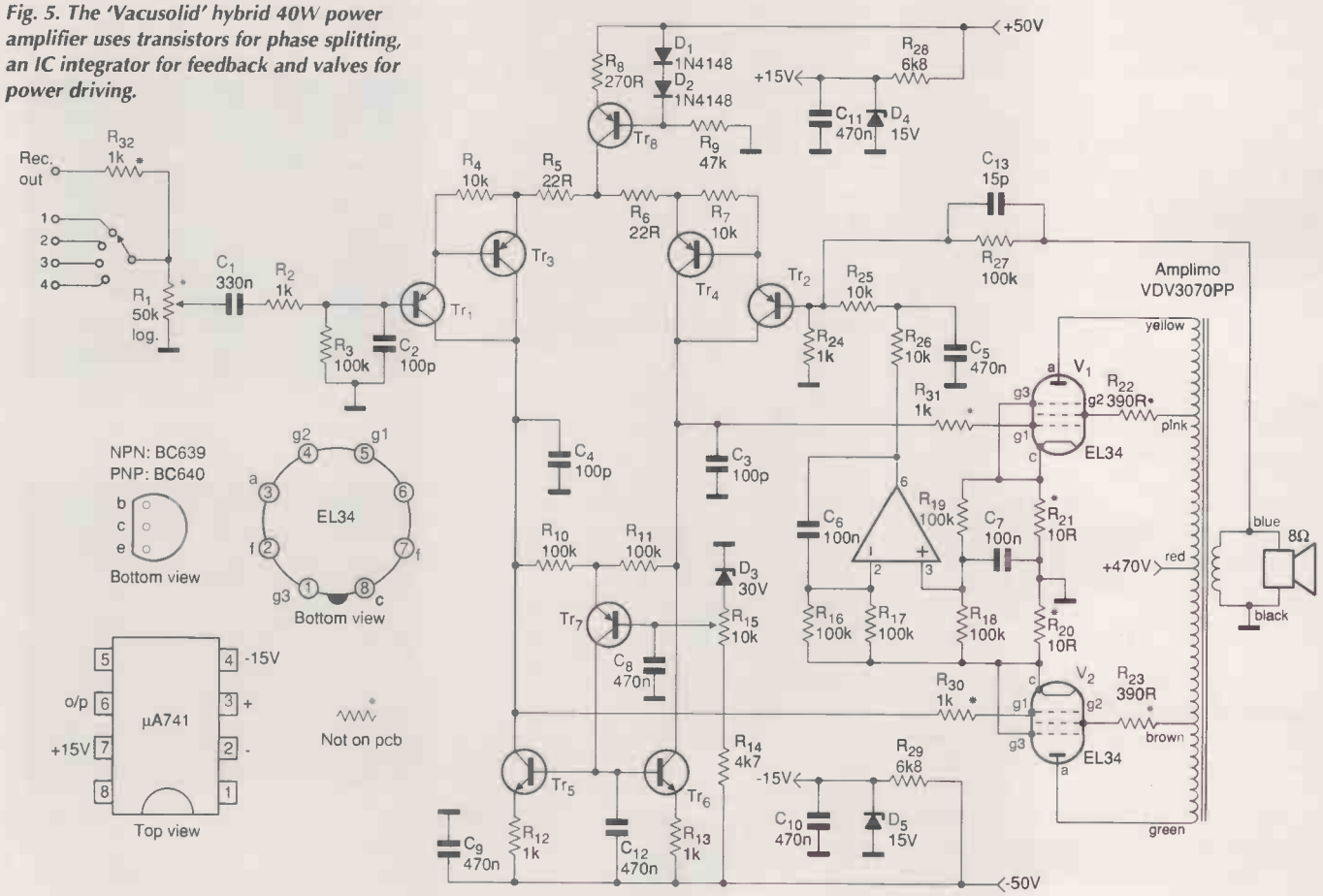
With an 8Ω load, maximum output power is more than 40W. Closed-loop gain is practically equal to R_{27}/R_{24}, i.e. 100. It follows that input sensitivity is typically 150mV rms. The output stage is driven by the solid-state phase splitter described above, to which two extra emitter followers, Tr_1 and Tr_2, have been added to increase input impedance and reduce the voltage drop across R_3 as a result of the base current of Tr_1.

Resistors R_4 and R_7 make sure that the bias currents of Tr_1 and Tr_2 are not too low. This would impair the hf performance. Emitter degeneration, involving R_5 and R_6, improves the input-stage dynamic range. Zener diode D_3 included in the voltage divider is used to determine the negative grid voltage, to make the set value more independent of variations in the negative supply. Resistors $R_{30,31}$ limit the risk of parasitic oscillations in the output stage, but are only effective if mounted close to the control-grid terminals.

The differential-input integrator circuit described above delivers a signal to the base of Tr_2 – the inverting input – via the negative feedback network. The op-amp does not work as well at higher frequencies, when signal distortion and noise may be produced at the output. To stop these signals from influencing the inverting input, an extra low-pass filter comprising R_{26} and C_5 is included in the β network.

The dc value of β, $R_{24}/(R_{24}+R_{25}+R_{26})$, gives a control range of about ±650mV at the base

Fig. 5. The 'Vacusolid' hybrid 40W power amplifier uses transistors for phase splitting, an IC integrator for feedback and valves for power driving.



of Tr_2. This is more than enough to compensate for the offset of the preamplifier and the output stage, and the voltage drop across R_1 due to the base current of Tr_1. The power supply for the first part of the circuit (±50V) is stepped down to the value required for the op-amp (±15V) with the aid of the parallel stabilisation (R_{28}–D_4 and R_{29}–D_5).

Supplying power

The circuit diagram of the power supply is shown in **Fig. 6**. The Amplimo type *7N607* toroidal transformer used here has a 340Vac ht winding. After rectification with 4×*1N4007* in a bridge circuit, this gives a dc voltage of 470V.

It is important to realise that this voltage, together with the current delivered by the transformer, is potentially lethal: users used to the low voltages of solid-state circuitry tend to forget the dangers of touching the live parts of valve-based equipment. Another hazard arises within the first minute or so after the amplifier is switched on. Until the cathodes of the output valves have fully warmed up, not enough current flows through the valves to discharge the HT capacitor if the circuit should be switched off at this early stage. This hazard can be avoided by including a 3W metal film 100kΩ shunt resistor in the circuit.

Supply voltages for the preamplifier are obtained by single-sided rectification from the 40Vac winding. Fairly large capacitors are used in connection with the single-sided rectification.

In order to protect the cathode filament insulation against puncture due to electrostatic charges, the filament current winding is grounded on one side.

Photos 1 and **2** show the prototype's housing. It is important to mount the valves vertically to prevent cathode sagging. I mounted the valve bases directly on the chassis rather than on the printed-circuit board, due to the high filament currents, high voltage, heat problems and the possibility of parasitic oscillation. ■

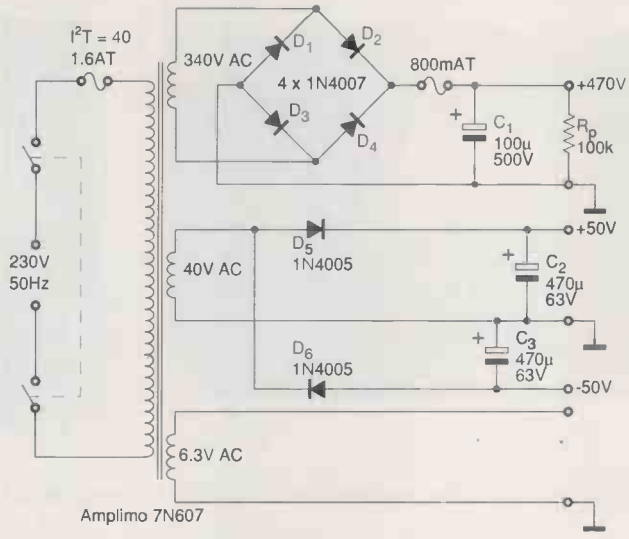


Fig. 6. Power supply circuit comprising ht at the top for the valves – lethal don't forget – ±50V for the phase splitting circuit and 6.3V for the valve heaters.

Performance of the hybrid power amplifier

Control range of cathode current is 10–90mA. The operating point was chosen at 40mA, at which setting the negative grid voltage is about 35V. At full drive, the output signal of the preamplifier is clipped at –50V. However, no current flows through the valves at –50V, so this does not affect performance adversely.

Screen-grid dissipation is 470V×5mA=2.35W (max. permissible 8W), while anode dissipation is 470V×35mA=16.45W (max. permissible 25W). Class A power in this setting is max. 8W for 8Ω.

- DC offset at cathode resistances <2.5mV (<0.625%)
- Negative feedback factor=5.6 (15dB)
- Loudspeaker damping factor=10 (1kHz)
- Input voltage at 40W output power=170mVrms.
- THD=0.5% at 40W, 8Ω, 1kHz, 1% at 40W, 8Ω, 10kHz
- Max. output power at 1kHz, thd=1%=44.6W at 8Ω, 37.8W at 4Ω
- Signal-noise ratio =95dB (104dB 'A' weight).
- Small-signal lf (–3dB) cutoff <10Hz, indicating that the CMRR of the integrator circuit (22dB at 16Hz) and chosen values of R at 50kΩ and C at 100nF give good results.
- LF cut-off point at 40W is 30Hz.
- HF –3dB cut-off =35kHz.

Frequency compensation used, C_3=C_4=100pF and C_{13}=15pF, limits bandwidth to the value indicated above. The excellent stability obtained is illustrated in the measured square-wave response with open output, **Photo 3**, at 8Ω, **Photo 4** and at 1µF, **Photo 5**.

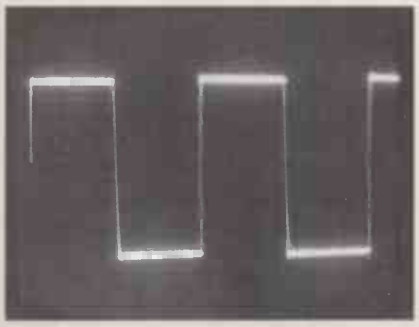


Photo 3. Square-wave response (2kHz) with open output.

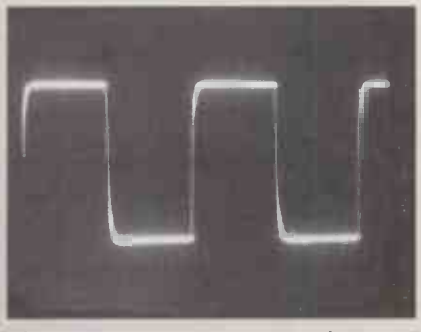


Photo 4. Square-wave response (2kHz) at R_L=8Ω.

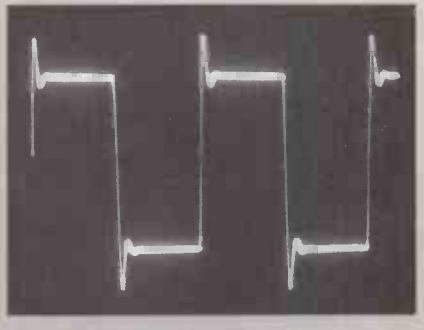


Photo 5. Square-wave response (2kHz) at C_L=1µF.

References

1. Jones, M., Classic valve power, *Electronics World + Wireless World*, Dec. 1995, pp. 1034–1038.
2. van der Veen, M., Theory and Practice of Wide Bandwidth Toroidal Output Transformers, preprint 97th AES Convention, Nov. 1994, San Francisco.

Technical support

The output and power-supply transformers, special fuses, ht capacitor (2×50µF, 500V), matched *EL34* valves, valve bases and printed-circuit boards are all obtainable from Amplimo, Vossenbrinkweg 1, Delden, The Netherlands, Fax. No. +31 74 3763132).

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ARTICLES FOR SALE

INDEX TO ADVERTISERS

	PAGE
Anchor Surplus	822
Bull Electrical	886, 901
Chelmer Valve	871
CMS	867
Conford Electronics	867
Crossware Products	871
Dataman	OBC
Devantech	870
Display Electronics	891
Electromail	880
Equinox Technologies	IBC
Field Electric	846
Halcyon Electronics	846
Hart Electronic Kits	896
ICE Technology	829
Jenving	837
Johns Radio	824
JPG Electronics	869
Kenwood	895
Keytronics	885
Labcenter Electronics	832
M & B Radio (Leeds)	866
Milford Instruments	IFC
Niche Software	895
Number One Systems	852
PICO Technology	869
Quickroute Systems	818
Radio-Tech	895
Ralfe Electronics	904
Robinson Marshall	837, 839, 841
Seetrax CAE	846
Stewart of Reading	870
Surrey Electronics	869
Technology Sources	839
Telford Electronics	867
Telnet	854
Those Engineers	841
Tie Pie Engineering	835
Tsien	894
Ultimate Technology	826
Wood & Douglas	871

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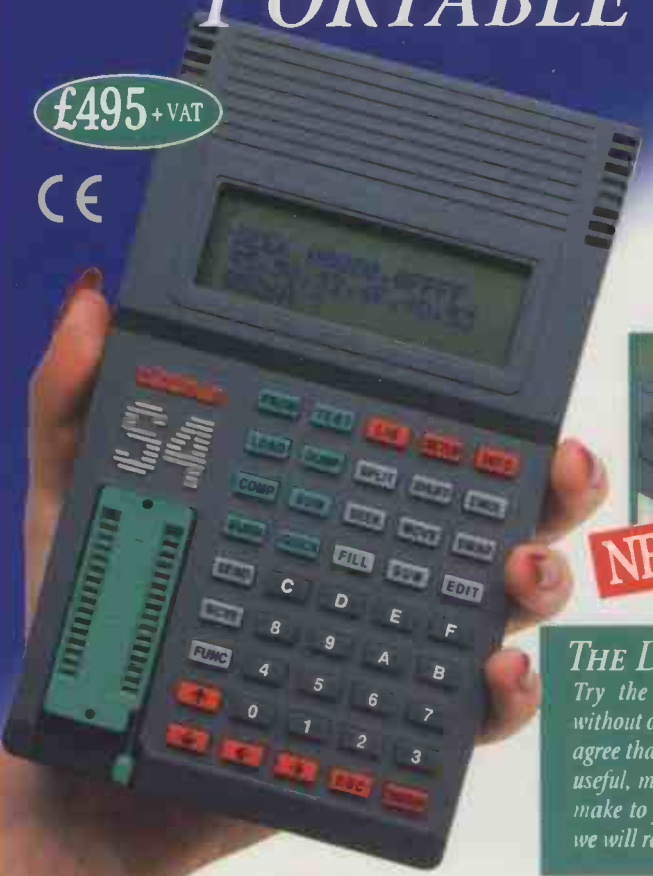

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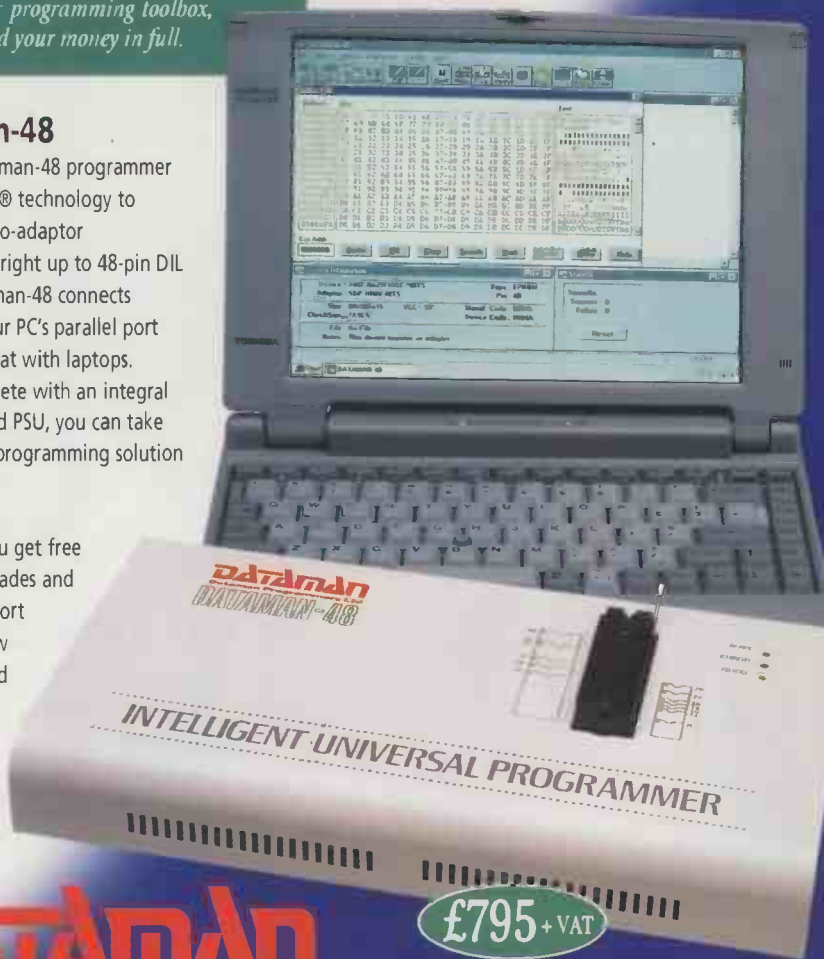

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