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RF DESIGN
Direct conversion SSB receiver

REVIEW
Easy acquisition under Windows?

PC ENGINEERING
Parallel port for data exchange

AUDIO
The ultimate slew rate?

APPLICATIONS
Power solutions to supply problems

Circuit design for RMS measurement

CONTENTS

FEATURES

READING RMS.................719
Metering sine-wave amplitudes is easy but estimating the true value of a complex waveform is fraught with difficulties – particularly at higher frequencies. We look at devices aimed at simplifying the task.

THE OPTICAL TECHNOLOGY DRIVE..............735
Both worm and rewritable optical disks can now hold up to 1.5Gbyte, and the technology is becoming much more accessible as prices fall. Martin Eccles looks at three frequently used media types.

NEW-WAVE MICROWAVES.............................736
Mike Hosking examines the fascinating world of microwave tuning, which involves ferrites and ceramics that are of little use at lower frequencies.

DIRECT-CONVERSION SSB RECEIVER.............743
Direct conversion reception is producing alternatives to expensive and bulky components. Frank Dorey shares his SSB development board – the principles of which can be adopted for other applications.

LIGHTING SWITCHES..752
Performance of medium-power switching transistors – such as those presented free on this month's cover – is now so high that it is possible to make tiny fluorescent lamp drivers with around 90% efficiency. Martin Eccles looks at how.

REAL-WORLD CONTROL VIA LPT...................755
Implementing i/o via PC expansion slots produces versatile but complex and relatively expensive results. For many simple tasks, the Centronics port can be more than adequate, as John Davies describes.

HIGH-SPEED AUDIO POWER..........................760
Although not wholly convinced of the merits, Doug Self embarked on the task of designing an audio power amplifier with a high slew rate. At first, the job seemed an easy one.

RF TRANSISTORS 772
Compensation terms and networks: Norm Dye and Helge Granberg show how frequency affects the way that impedance compensation networks are designed and why negative feedback is so effective.

RIPPLES IN THE ETHER.....................................778
In 1894, Marconi, embarked on research that enabled the world to take the first steps towards modern communications. John Powell Riley pinpoints the key moments in the life of this genius.

REGULARS

COMMENT 707
The sound of indifference.

UPDATE...708

RESEARCH NOTES...713
Thunderstorms that rain gamma rays, Do memory systems forget civil liberties, Babbling helps make sense of cerebral palsy, Joint approach finds new solder approach, Hard rain, Filters tune with fuzzy logic, Neural nets put the squeeze on moving pictures.

PC ENGINEERING 725
Although not a real-time operating system in itself, Windows provides an ideal graphical user interface to process-control type platforms. Allen-brown looks at *Visual Designer* – a PC data acquisition package.

DESIGN BRIEF 730
The golden ratio, e, pi, (√5+1)/2, tossing a coin? Ian Hickman investigates the relevance of magic numbers to modern electronics.

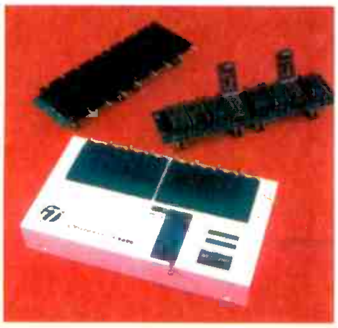
APPLICATIONS..748
Versatile switching regulator IC, Low-cost evaluation for PowerPC, Using Doppler in car alarm applications.

NEW PRODUCTS 767
Comprehensive round-up of the industry's new products, presented in the industry's most readable format.

Your free transistors
The two high-performance Zetex transistors, presented free to UK readers on this month's front cover, are detailed on page 752.

CIRCUIT IDEAS 784
Capacitive fluid-level detector, VFO uses a single current-mode IC, Power isolator, Micropower logic coupler, Switched-mode constant-current charger, Mosfet stabilises Wien amplitude, Simple servo, Gate-voltage generator.

LETTERS 787
Objective assessment, Quantified listening pleasure, Simpler circuit, Amp designers top ten, No military/civil distinction, Where is non-magnetic power? Hall's well that ends well, Old radio club, c change? Dying light, Bad references, Clock mechanism, Private progress – public property, Two wrongs...?



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ISSN 0959-8332

REED BUSINESS PUBLISHING

The sound of indifference

It is now five years since the nicam television stereo sound system first entered public service. Currently only about 20 per cent of all sets sold are equipped to receive stereo transmissions.

Those of us who have a nicam set will testify to the stunning quality available on ITV and Channel 4 sound tracks. They invariably demonstrate a crystal clear spatial sound image – particularly in drama productions – which adds immeasurably to programme enjoyment.

I would go so far as to say that the attention which production staff put into sound quality exceeds that of most radio programmes, more noticeably so when played through a good pair of speakers separated physically from the television.

Most people would never know this The television companies have done virtually no promotion on the nicam system. For those with long memories, compare situation to the publicity which surrounded the launch of the colour service nearly 30 years ago.

I personally have never heard a BBC nicam sound transmission but I am sure that it would be equally good: the transmitter which serves our area will not be delivering a nicam service for at least another ten years.

If you talk to broadcasters – particularly the BBC – they will tell you that they simply don't have the money to adapt transmitters in service to radiate the extra subcarrier signal. They can only afford to equip for stereo sound when station equipment naturally comes up for replacement. In the meantime, the below-the-line policy is not to mention this service so that people who are being forced to wait for decades "won't get jealous" in the words of one Corporation person.

The reticence of the of the independent television companies is harder to understand. One would have thought that one company or region would have made whatever competitive capital it could out of the enhanced sound service. Perhaps they feel that their viewers are too stupid to appreciate sound quality and that they don't want the advertisers to know.

Either way, nicam offers benefits to viewers and licence payers which most will never appreciate. This is ironic since the BBC played a major role in the design of this most excellent system.

One understands the pressures from the political agenda which broadcasters face; the process leading up to publication of the recent White Paper on the future of the BBC must have put all forms of capital expenditure on hold. Its publication should have cleared the air but reports coming from inside the Corporation suggest that financial easement will benefit programmes rather than engineering development.

Even though the Corporation has produced world class technical developments, continued financial constraint is likely to prevent these entering service. Thus when it makes pronouncements on such things as digital audio broadcasting or DVB, no one should take it seriously, least of all the setmakers who would otherwise invest heavily in new design and production. This is a great shame because it stunts development of a much wider electronics industry infrastructure.

When the Government allowed the renewal of the Corporation's charter on largely unchanged terms, it missed an opportunity to enable the BBC to participate in the future development of broadcasting. This is far more important than it might appear.

Frank Ogden.

Electronics World + Wireless World is published monthly. By post, current issue £2.25, back issues (if available) £2.50. Orders, payments and general correspondence to **L333, Electronics World + Wireless 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 0AU 071 261-5555.
Subscriptions: Quadrant Subscription Services, Oakfield House, Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 0444 445566. Please notify change of address. Subscription rates 1 year (normal rate) £30 UK and £43 outside UK.
USA: $52.00 airmail Reed Business Publishing (USA), Subscriptions office, 205 E. 42nd Street, NY 10117.

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, 205 E. 42nd Street, NY 10117.Telephone (212) 867-2080. Tlx 23827.
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
Typeset by Marlin Graphics 2-4 Powerscroft Road, Sidcup, Kent DA14 5DT

Widescreen television for UK

After fifty years of watching box-shaped tv screens, British tv viewers will now get the chance to receive wide-screen broadcasts. On new tv sets these will give clearer pictures of shape similar to a cinema screen. But a deep-rooted policy difference between the BBC and commercial tv stations means that viewers will have to chose between two different transmission technologies, or buy two different wide screen reception systems.

The commercial tv stations will use an analogue system called PALplus which builds on the existing PAL tv system used in Europe and Australia and will be ready this year. The BBC will wait three years and use all-digital technology which is not yet ready.

The BBC's digital wide-screen programmes will be completely separate from its analogue channels BBC1 and BBC2, use spare frequencies which are slotted between them, and be wholly incompatible with all existing sets. The Department of National Heritage recently cleared the way for this in its White Paper, The Future of the BBC. Viewers who

Pressure on for better engine management

In an internal combustion engine, nitrogen-oxide and carbon-dioxide pollutants can be minimised by increasing the air to fuel ratio for the engine.

Running at a much higher air to fuel ratio dramatically reduces these emissions but at the same time will increase the chances of misfiring which in itself increases emissions of unburnt fuel.

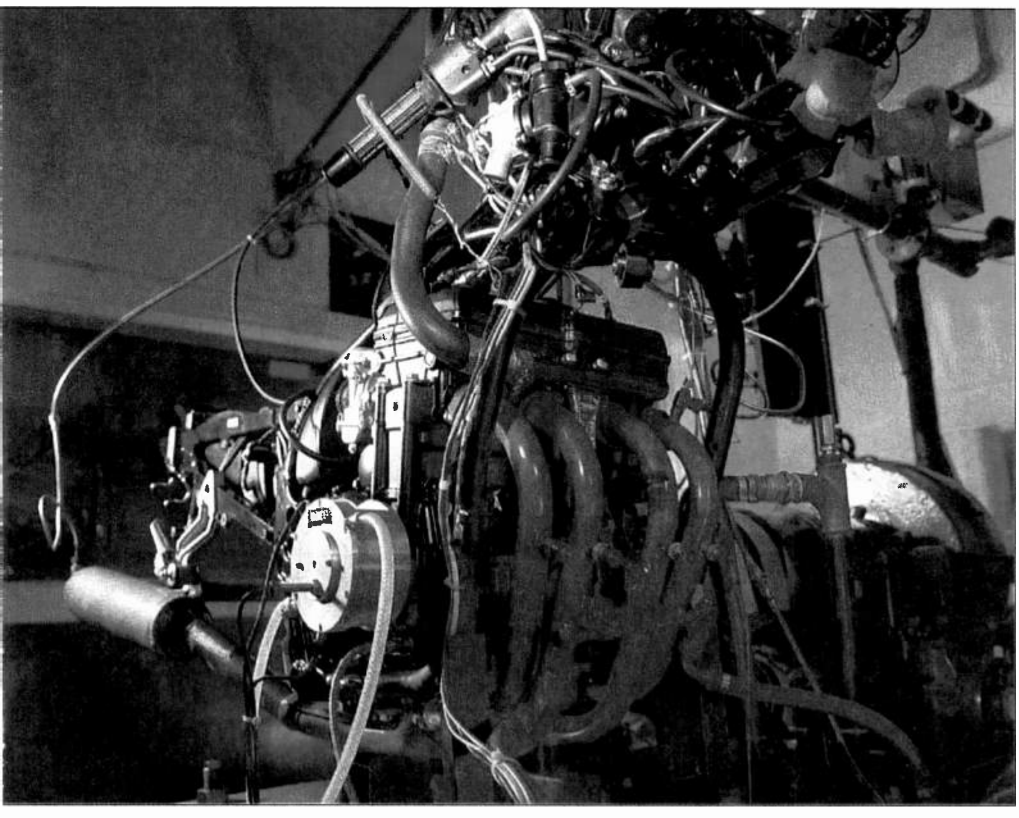
To minimise this effect, Toyota has developed a pressure sensor that fits in a vehicle's combustion chamber enabling the engine management system to detect misfiring and adjust the vehicle's air to fuel ratio accordingly, thus controlling the air-polluting nitrogen-oxide emissions.

"I was there in Japan at Toyota and saw the sensor. We even tested it for research purposes. It is more or less used for misfiring detection and it works close to the lean burn engine limit," said Rolf Kuratle, a product manager for combustion engine measurements at the Swiss firm Kistler Instruments which develops and manufactures automotive electronics.

Toyota claims that this is the first combustion chamber pressure sensor in the world to be mass-produced as an automotive part. It has been fitted in the Toyota Carina E, at the moment only available in Japan.

This sensor cannot be retrofitted in a car but it looks set to become an integral part of future car engines.

"In theory this is fine," said Peter Lanscott, a representative of Kistler Instruments UK. "But it is very expensive.



A Kawasaki *GPX750R* motorcycle engine undergoing high-speed combustion analysis – 10,500rev/min – at Ricardo, the Shoreham based research, development and consulting organisation. Readings for the Toyota pressure sensor are averaged over several engine cycles.

want to receive the new programmes will have to buy a new wide-screen tv set, with digital decoder. Everyone else will continue to watch BBC1 and BBC2, as normal, on their existing PAL sets.

Commercial station Channel Four wants to get in earlier and will this October start wide screen broadcasting with analogue PALplus. Granada has confirmed it will start PALplus transmission before the end of the year. Viewers with PALplus sets see pictures which are not just wider than normal, but clearer too, with more fine detail and fewer artefacts like spurious colour patterns on check jackets. Viewers with existing PAL tv sets receive the same programmes, but they will see them in letterbox format, with a black border at the top and bottom of the screen.

PALplus was developed by a consortium of European electronics companies and broadcasters. Work is a year ahead of schedule and electronics company Nokia of Finland will have wide-screen tv sets with built-in PALplus decoders ready for sale for 1300 pounds this October.

A conventional PAL tv picture, of 4:3 boxy aspect ratio, is built from 625 horizontal scanning lines. Of these, 576 are 'active' lines which are used to trace the picture on the screen. The other 49 lines define the black borders which are largely lost off the top and bottom of the screen.

With PALplus, the programme is broadcast in wide-screen 16:9 aspect ratio, using 432 active lines. The other 144 lines appear as thick black bars at the top and bottom of a 4:3 screen. A PALplus tv set with wide-screen picture tube expands this letterbox image to fill the full screen. So the expanded image is built from only 432 lines instead of 576, and would, without extra tricks, look very coarse.

A PALplus tv set rebuilds the lost resolution by using an analogue "helper" signal which conveys high frequency, fine picture detail. This helper is buried in the 144 black border lines for transmission.

The helper is generated before transmission, by equipment which separates the active picture lines into groups of four, and then converts each of these groups into a group of three picture lines plus one helper line. PALplus wide-screen sets combine the helper lines with their corresponding three-line groups, to reconstitute the groups of four lines. The receiver needs 6M-bytes of computer memory to do this.

Channel 4 says it will transmit 500 hours of wide-screen material before the end of 1995.
Barry Fox

Damages for multipath viewers

Viewers whose tv pictures have been spoiled by the erection of a tall building in the path of the transmitter have been anxiously awaiting the result of a recent High Court action. Seven hundred people in East London put their names on a writ which claims damages from Canary Wharf Limited, owners of the Tower in Docklands which blocked their tv signals.

These viewers now have their pictures back, thanks to a secondary transmitter built by the BBC, and are suing for at least two lost years. **BF**

We developed transducers for the Mercedes-Benz engine that won the championship in 1990 at Le Mans and it was expensive."

Although Toyota sensor's prime objective is to keep the nitrogen-oxide emissions low, it also improves fuel consumption without affecting vehicle performance, and therefore lengthens engine life.

To reduce the nitrogen-oxide levels, Toyota engineers introduced feedback into the control loop between the combustion chamber and the air intake manifold. The pressure sensor and the in-vehicle computer form an advanced engine management system.

It Is in the air intake manifold that the air to fuel ratio is increased close to the ideal ratio of an advanced lean-burn combustion engine which is 24:1 (compared to 14:1 ratio in an average petrol engine).

This in itself has proved difficult for the Toyota engineers to achieve, as increasing the air to fuel ratio means that optimal functioning and stability of the engine is difficult to maintain.

The pressure fluctuations in the combustion chamber, usually a sign of an unstable engine, are sensed and several consecutive readings are taken .by the in-vehicle computer which enables adjustments to the air to fuel ratio to be made.

"It is very dangerous to keep adjusting this ratio after every reading taken in the combustion chamber. That's why a few readings are taken and the amount of adjustment needed statistically calculated," said Katsuhito Hirose, the assistant manager at the homologation and research department at the Toyota Motor Company Europe in Brussels.

Two devices were suitable for use as combustion chamber pressure sensors appropriate for the vehicle environment. One was a sensor made of a piezoelectric material, PZT ceramic

chamber pressure sensors appropriate for the vehicle environment. One was a sensor made of a piezoelectric material, P~T ceramic (lead-zirconium-titanate), and the other made of a single silicon crystal.

The single silicon crystal was chosen due to its superior mechanical properties, such as thermostability and durability, and its immunity to electrical noise. The silicon crystal can achieve toughness against large stresses, and accuracy at high temperatures.

Silicon also has a high piezoresistive coefficient (the ratio of electric resistivity to applied stress).

The combustion pressure sensor is installed directly into the vehicle's engine and consists of a pressure detecting part, which converts stress into force, a force detector, which converts the force into electric signals and a built-in amplifier which suppresses noise.

A key part is the force detector which works on the basis of the piezoresistive effect. In this case Toyota has used a piezoresistive material whose directions of force, driving current and detecting electric field are perpendicular to one another.

The force detector, of a size 1.7 x 1.7 x 1.9mm, consists of a metal hemisphere, a transmission block, a silicon chip, and a base block. The metal hemisphere rests over the force transmission block, which is placed over the silicon chip and the base block. A hemisphere design was chosen due to the fact that force is equally distributed in all directions over the silicon chip.

The silicon chip is p-type with diffused boron atoms. On the chip's surface two pairs of electrodes are connected to each other in a perpendicular manner. The electrodes are there to supply the driving current (input electrodes) and to detect the voltage (output electrodes).

The input and output electrodes' lengths, positions and impedances are important for the detector as they influence its sensitivity.

By experimenting, Toyota's optimal figures for the output electrodes' length are in the range of 50 to 100μm and the length of the input electrodes should be equal to the width of the force impressed area.

The force transmission block and the base block are made of devitrified glass, which is a material doped with impurities to gain a multi crystal structure. This makes it suitable for electrostatic bonding with a thermal expansion coefficient close to that of silicon.

The detector is driven by a dc voltage. The driving current flows throughout the silicon chip and the electric potential is distributed on its surface.

The output voltages, proportional to the applied force, are detected between the output electrodes as a differential potential.

The force is applied to the top of the detector. When no force is applied, the output voltage is nearly equal to zero. A metal diaphragm converts the combustion pressure into the force which is transmitted to the force detector by a transmission rod.

The output amplifier consists of an op-amp integrated into the silicon wafer in order to minimise the effects of electrical noise picked up by the cabling between the sensor and the in-vehicle computer. This noise is of the order of several millivolts.
Svetlana Josefana, *Electronics Weekly*

Next generation of wafers set at 12in

The next generation of silicon wafer size has been fixed at 12in. at a closed door meeting between major non-Japanese chip users and equipment suppliers at the Semicon West exhibition held in San Francisco last week.

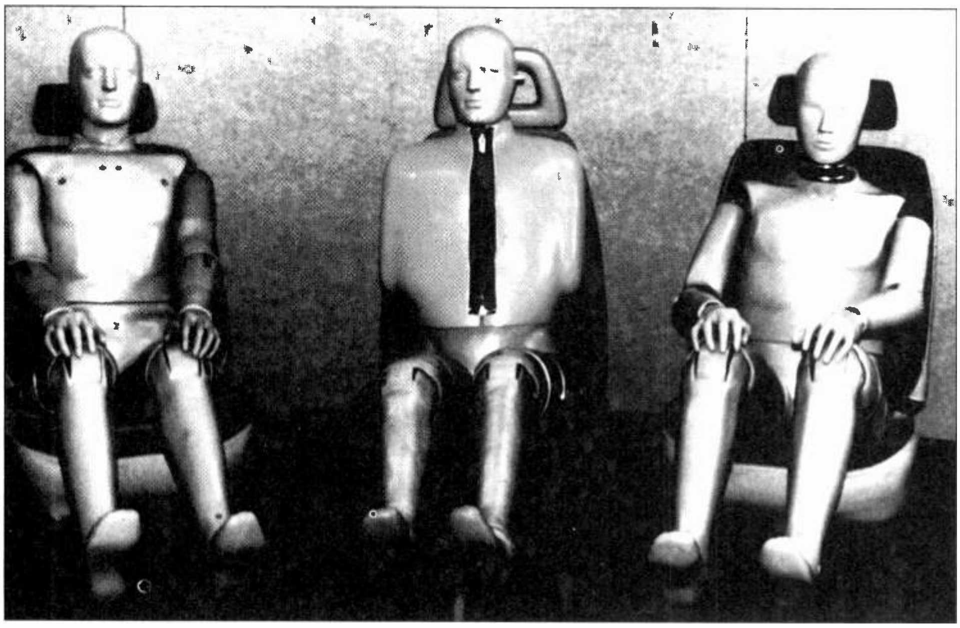
But the Japanese Ministry of Trade and Industry (MITI) has proposed that an international consortium should be formed involving German, US and Japanese chip makers to work together on the next-but-one wafer size – 16in. To make sure that Japanese companies have a major say in it, MITI is offering to put up 70% of the $176m estimated cost.

The 12in decision, which has the support of the US government-backed semiconductor consortium Sematech, is also supported by Europe's Joint European Submicron Silicon Initiative, whose representatives attended the meeting.

Ten-inch wafers were rejected because they would only give a 56% increase in area over today's 8in wafers, whereas 12in wafers give a 125% increase.

But the decision has not met with universal approval. Most Japanese companies are still trying to fully utilise their 8in wafer fabs. The world's largest supplier of wafer fab equipment, Applied Materials, would have liked to have seen the industry agree on 14in wafers – such a move could have squeezed out smaller equipment manufacturers.

Paul Gregg, *Electronics Weekly*



Dummies with intelligent heads. An element of a new range of crash-test dummies from Vector Research is a new magnetohydrodynamic sensor from Endevco. It measures angular rate to help in the assessment of head injury criteria.

Pentium prices will plummet

Intel is planning to slash the price of its Pentium microprocessors by up to 50% as it fights According to documents leaked from Intel, it will cut the price of Pentium chips by between 35 per cent and 50 per cent over the next nine months.

As predicted in Electronics Weekly last month, the Pentium price cuts are much sharper than Intel's standard price cutting strategy for previous microprocessor lines such as the 386 and 486, where prices fell only 25 per cent per year.

Intel's plans for the fourth quarter of this year call for the 60MHz Pentium microprocessor to fall in price to below $400 compared to its recently cut price of $575. The price of the 90MHz Pentium will fall to around $600.

By the second quarter of 1995, 100MHz Pentium prices will be reduced and the 486DX4 will replace the 486DX2. The 75MHz Pentium will then replace the 60MHz Pentium at the same price point.

The steeper price cuts represent a potentially dangerous strategy for Intel since Pentium microprocessors are more expensive to make than rival high end 486 or Risc microprocessors and Intel risks losing profits needed for future investment.

Play time... US chip firm LSI Logic has unveiled pictures of the central processing unit it has designed jointly with Sony for the Japanese firm's Playstation video game machine. Based around a 32bit Mips Risc microprocessor, the CPU also has a 3D graphics engine and a full-motion video decoder based on the JPEG standard. Sony plans to launch the Playstation, which will run software delivered on CD-ROMs, in Japan later this year, followed by a US launch in 1995.

Sony MiniDisc in computer data storage challenge

Sony is launching its MiniDisc audio technology as a data storage format which it hopes will replace the 3.5in. floppy disc.

The move, announced at last month's PC Expo show in New York, is part of a drive to establish Sony as a major branded computer peripherals supplier, spearheaded by products based on two of its most famous technologies: MiniDisc and its Trinitron colour tv tube.

Sony's MiniDisc re-recordable disc technology was developed as a replacement for the compact cassette in the audio market. Each magneto-optical disc, measuring just 2.5in. in diameter, can store 140Mbytes.

Sony said several major computer makers, including IBM, are interested in integrating MD drives into their portable machines. But initial versions are too high and so will be sold as standalone peripherals.

Sony also launched Trinitron 15in and 20in computer monitors.

Sony already makes unbranded computer peripherals for other computer companies. It hopes to cash in on this expertise and on its strong brand recognition in the consumer market.

At the press conference to announce the peripherals move, Sony also gave the first public glimpse of the personal digital assistant it plans to launch next year. The prototype pen-based device is based on a Motorola processor. It has an internal modem but no wireless communications capabilities. Sony said it will make further announcements about the device in September. ■

CIRCLE NO. 108 ON REPLY CARD

RESEARCH NOTES

Jonathan Campbell

Thunderstorms that rain gamma rays

Here on Earth, thunderstorms can mean spectacular light shows caused by dramatic discharges of electricity. But scientists are now coming to realise that there may be more to see above the clouds too – bursts of gamma ray radiation that originate in Earth's the upper atmosphere.

Up to now, such activity has been hard to detect and 'evidence' has tended to be regarded as spurious noise. But a US team making observations with multiple detectors as part of the burst and transient source experiment (Batse) running on the Compton Gamma Ray Observatory now say they have positive proof of these gamma ray flashes (*Science*, Vol 264, pp.1313-1316).

The Compton Observatory has been monitoring gamma activity since its launch in 1991. But the researchers say that the rare gamma-ray flashes have not been reported before because it was unclear that they were real.

More to lightning than meets the eye?

Two features of the burst are their extremely hard spectra and their short duration. In addition, they differ from other known gamma-ray behaviour and originate from the Earth's atmosphere around 30km up.

Scientific speculation is that the phenomenon is the result of a rare-type of high-altitude electrical discharge above thunderstorm regions. The researchers have plotted the approximate locations over the earth of the gamma ray bursts, occurring at a less than one per two months and have obtained a clear correlation with average annual thunderstorm activity. In addition, specific concurrent weather information has showed thunderstorm conditions coincident with the events.

The possibility of strong electric fields producing ionisation at altitudes high above thunderstorms was discussed over 70 years ago. The researchers say that fields intense enough, over a large enough area, could not only cause ionisation but also 'runaway' electrons and subsequent bremsstrahlung x-rays. They point to the fact that the electric field due to lightning falls off less rapidly with height above the clouds than does the atmospheric density which determines the break-down potential of air. Calculations suggest that an electric field strength at 60km exceeding 500V/m could have the effect. The field would need to accelerate electrons over several km to achieve the megaelectron volt electrons necessary to produce the gamma bursts. But the scientists say that glow-like discharges observed from planes and the ground over heights of 40-80 km – extending well over 10km vertically and 10 to 5km horizontally – could be capable of producing the large field changes.

Upward-going lightening events have been recorded by the space shuttle and by pilots. But this is believed to be the first investigation of gamma radiation from atmospheric electrical discharges. That, combined with findings still in the initial stages, means that the cause can still not be explained for certain. But at least now the scientists are convinced that there is something to explain.

Do memory systems forget civil liberties?

Civil liberties groups worried about current data protection legislation could one day look back on the 1990s as nostalgic days of untrammelled privacy. In a brave new world where we no longer have to rely on our memories to recall where we put that file, or what we said to someone, we could all simply become entries in someone else's electronic diary.

A glimpse of this forget-me-not future is given by Mike Lamming and his team at the Rank Xerox Cambridge EuroParc (*The Computer Journal*, Vol 37, No 3, 1994, pp.153-163).

In a project to design a human memory 'prosthesis', the Xerox group has looked at how various different research projects could be linked together to create a system that records every place we go, everyone we speak to, what we say and what we do. The work springs out of the need to improve office systems, easing the problems of finding files, papers and notes, recalling names of people and places, procedures and lists, and remembering to perform tasks.

Relevant hardware and software is already around, from Amstrad's *PenPad* to the more advanced Casio/Tandy *Zoomer* and Apple/Sharp *Newton MessagePad* devices. Xerox also has been developing its bleeper-sized *Parc Tab*, with a touch-sensitive screen and a beacon to indicate its location. Infra-red links each Tab the user's home computer.

"I come here to forget"

EuroParc has already been test gathering data on people's movements in its Pepys project, using Olivetti infra-red badges to log the location of people. Walking from an office to the common room to meet a visitor generates events at a whole series of sensors along the way.

Pepys location data could be augmented, say the researchers, by a video diary and snapshots. A video network around the building would be directed by the badges to switch to the camera nearest a particular person as they move around the building. Attraction of video is that it seems to help people recall considerably more activities than the Pepys diary alone.

Pepys is also limited in that though it will recall meetings, it does not record what happened. But two systems undergoing testing that could tackle this shortcoming are NoTime and Marcel. NoTime electronically captures the hand-written notes made during a meeting. Each writing stroke is date-stamped and the notes are linked to the appropriate part of an audio and video recording of the meeting.

The Marcel system recognises activities involving paperwork by using a video camera mounted over the user's desk. Images of documents are digitised, and then compared with a database of known documents. The researchers report that Marcel could be a plausible way of logging document movement.

Keeping track of activities at a workstation is in some ways easier, in that all file movements and commands can be logged. But the researchers note that the data records, though extensive, still are not easy to interpret in terms of what someone was doing.

After gathering data, any memory system must be able to prompt the user when a particular location is entered or when a certain person is encountered. Parc Tabs are already being used to generate audio reminders in this way, with messages being set at any time and anywhere in range of a Tab sensor.

The aim of a memory prosthesis, say the researchers, is to sense aspects of a user's environment and make records which can be later used to help recall events which a user did not even realise they needed to remember. So any system must automatically capture as much data as possible.

Implications for privacy are considerable. The researchers say that users should be clear about what is being recorded, and systems should be configurable to what individuals consider as acceptable intrusion. But for non-prosthesis holding individuals there is no such choice. Their movements, conversations, hand-written notes and even facial expressions could all be electronically recorded as part of the diary of people they meet. Such a prospect makes current concern over identity cards and the proliferation of video cameras quite tame in comparison

Babbling helps make sense of cerebral palsy

Northeastern University's baby babble blanket could help disabled babies interact with the world around them.

Computer game technology, so often blamed for robbing young people of their social and communication skills, is being put to work at Northeastern University, to help improve communication capabilities.

A software development group led by Harriet Fell and a field testing group led by Linda Ferrier, have transformed a Ninentendo Powerpad from a games peripheral into a 'baby babble blanket' that helps severely disabled babies to experiment with non-sensical babbling. Early babbling is thought to be vital to development of later talking.

Their blanket, linked up to a Mac computer running specially developed software, enables babies to trigger sounds and an audio track of other baby babbles by rolling around on the large pressure sensitive switches. The babble software also allows the child to turn on electrical toys or household gadgets such as fans.

Fell and Ferrier hope that this interaction will help fight the 'learned helplessness' of disabled infants. Unfortunately, by the time children with physical disabilities reach school age, they may already be poor communicators because they have never learnt to interact with the world around them.

So far the blanket has been tested with a wide variety of children, including non-disabled babies as young as four-and-a-half months, and children with multiple disabilities up to twelve years of age. It is currently being tested in homes and classrooms.

A spin-off project has been development of an early vocalisation analyser. The analyser takes a digitised waveform and uses a base-line noise threshold to count the number of infant vocalisations in specified time. Vocalisations are then sorted by their characteristics.

Currently a prototype is being use to answer the question: 'How frequent and how long are vocalisations of normal infants compared to infants with cerebral palsy?'

Joint approach finds new solder approach

Worries over possible health and environmental drawbacks to current soldering practice, upon which so much of the electronics industry depends, have been forcing researchers to seek alternatives to lead-based solders and the volatile organic compounds used in fluxes. Now a collaborative project led by GEC-Marconi (GEC Journal of Research, Vol 11, No 2, pp.76-89) suggests that solders based on Sn-(2 or 3.5)Ag or Sn-0.7Cu could prove workable options in both wave and reflow soldering.

One other clear conclusion of the study is that all components of the solder joint such as the metallisation on both boards and components and the fluxing system – in addition to the solder alloy – must be taken into account.

The engineering performance of various lead-free solders was investigated as well as their economic and environmental aspects. The tin/silver and tin/copper compounds performed well and the team also identified the importance of an inert atmosphere as the key to developing highly reliable processes.

Cracking in through-hole joints following the formation of low melting point compositions caused by badly matched solder alloy and component metallisation was one of the factors emphasising the importance of looking at a system as a whole.

The authors say that further trials will be necessary before all-embracing recommendations can be made.

Hard rain

The evocative sound of rain blowing against a window brings back 'sweet memories' according to a 1970s classic soul number.

Now J McLoughlin, DJ Saunders and RD Ford have reprised that theme for the 1990s, interpreting the mournful sound of rain gently falling on a roof (*Applied Acoustics*, Vol 42, No 3, 1994, pp.239-255):

"The sound intensity level radiated from a single-skin corrugated roof of trapezoidal section when excited by the impact of water has been shown to depend on the sixth power of the impact velocity, third power of the drop diameter and be directly dependent upon the water impact rate".

Need some work on the scansion there I think guys.

Bending battery technology. A prototype plastic high-energy battery that is rechargeable and can be bent into any shape, has been developed by telecomms research company Bellcore in the US. The battery looks like a solid – no liquid leaks out if it is punctured – but acts as a lithium-ion battery. Bellcore says it will 'revolutionise' the consumer electronics and telecommunications industries. Its performance relies on a polymer matrix: the elements are permanently bonded together then covered by a moisture-proof barrier coating. At 3.8V its energy density clearly makes it competitive with normal nickel-cadmium and lead-acid batteries, without the environmental worries – it contains no toxic metals.

Filters in tune with fuzzy logic

Electrical Engineers at A & M University College in Texas have developed an expert system, exploiting fuzzy logic, that they claim is a simple way to bring out-of-spec filters back into line.

Butterworth and Chebychev approximation techniques are often used to fit a frequency response of an analogue filter into a specified window constraint. But when approximations are implemented in hardware, component variations can mean the filter may not meet its specification. Inclusion of a tuning system can adjust some of the components. But adjustable components usually produce non-linear changes in filter frequency response; variations in one component can modify several characteristics of the filter; and the implemented circuit will contain parasitic components and have other non-ideal effects.

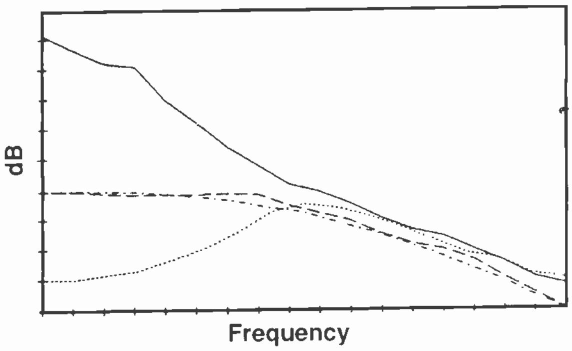
The Texas approach (*Electronics Letters*, Vol 30, No 11, pp.846-847) takes advantage of the fact that a filter window specification can allow any curve – as long as it is in the window. Once achieved, the system can optimise the filter to approximate the desired function.

The fuzzy logic involved was designed to approximate a Butterworth filter with maximum attenuation in the stopband.

By measuring the output at certain frequencies, the system modifies the filter parameters accordingly, applies the test signals and repeats the same process until the frequency response is within the window.

Texas's system has been successfully tested on a low pass filter implemented with transconductance op-amps and the researchers say that fuzzy logic has now been proved as a useful technique for tuning filters and should be a useable method for other electronic circuits or systems.

Correcting out-of-spec filters using fuzzy logic. Untuned output (solid line) was tuned (dashed line) after 14 iterations. In the second case the untuned output (dotted line) was tuned (dot/dash line) after nine iterations.



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CIRCLE NO. 110 ON REPLY CARD

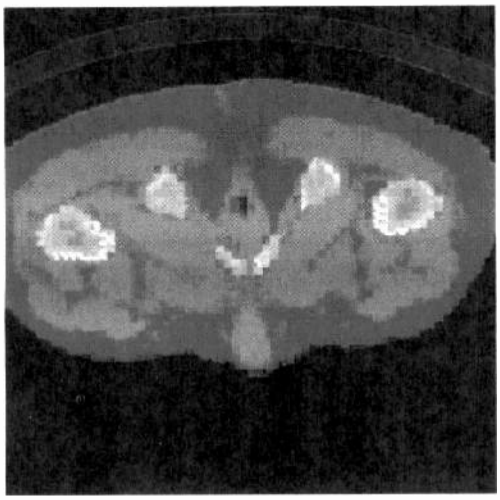
Neural nets put the squeeze on moving pictures

Neural networks, adapting on-line to a changing image input, could be the way forward for engineers looking to achieve that frustrating balance between picture quality and effective compression of moving images. Early work by researchers at GEC-Marconi has shown that the technique can work – though much research still needs to be done.

NP Walker, SJ Eglen and BA Lawrence (*GEC Journal of Research*, Vol 11, No.2, 1994, pp.66-75) took as their starting point, the compression of single images using a Kohonen network.

To start, a Kohonen network – a network of 16 input nodes joined to 1024 output nodes by random weights – was 'trained' by presenting it with 4×4 pixel blocks from an image. The output node whose weight most closely resembles the input pixel intensity has its weight adjusted so that the match is even closer. As training progresses, the weights associated with different output nodes come to represent various patterns and textures from the image. Characteristics of the network also mean that neighbouring nodes will come to represent similar patterns within the image.

At the end of training, the weights associated with the output nodes form a codebook. Now when a picture is compressed, the image is broken down into numbers that point to the relevant codebook blocks that will reconstruct it.



Reconstructed hip image from the single image-block codebook.

Next step was to try neural network compression of moving images. Each separate frame image could be separately compressed as before. But this would ignore inter-frame redundancy – where there is no change in an area of the image. So the researchers took 4×4 blocks of pixels from four consecutive images, in effect creating a 4×4×4 cube of 64 pixels to be analysed. For the complete four images, the 64-dimensional input vectors are presented to the network to produce a single list of appropriate codebook pointers.

Success of the approach was tested by compressing a sequence of 50 medical cat scans through a patient's hip joint.

Initially, the researchers limited themselves to training a network (64 input and 1024 output nodes) using only the first four images from the sequence. Once trained, reconstruction of the original images of the first four frames was good, with an acceptable error difference compared to the originals. But for images that the network had not seen before, the error was much larger. Plainly a new codebook was needed that was more representative of the unseen patterns.

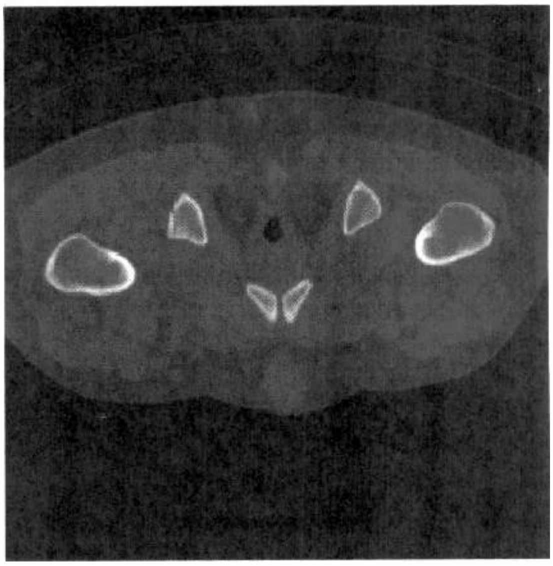
The answer, say the researchers, is to develop an adaptive network that could reflect the changes in the scene it must represent. A codebook could be updated every 10 frames or so, or when it no longer represented the image.

Whether or not a codebook is adequate can be ascertained by measuring the error for each windowed image block. Blocks that have errors above a given level could then be used to help update the codebook. The neighbourhood concept allows one input vector to alter the weights of many nodes in the codebook – a considerable advantage of the Kohonen network over other techniques.

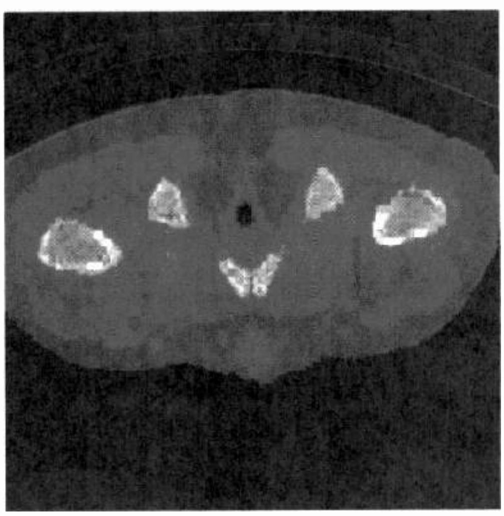
The team says that as well as providing good reconstructions of the image in question, an adaptive network would help produce a codebook that was better able to represent a wider range of images. To prove their point, they trained the network using four groups of four-frame images, but still leaving frames 29-50 unseen.

This time the reconstruction of the unseen images was much better (see figures) and suggests that a more generalised codebook formed by an adaptive network would give a better performance

Currently the team is investigating how storage or timing problems caused by the blocking process can be avoided. One path is to exploit redundancy by transmitting an initial image and then coding the subsequent inter-frame differences. ■

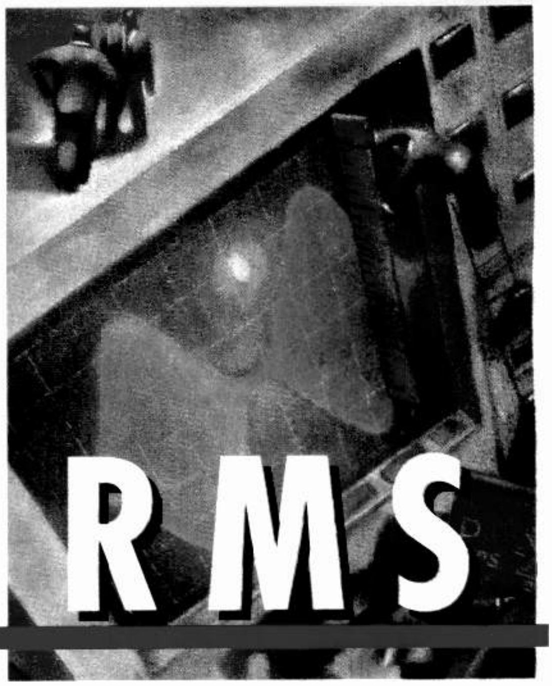


Original hip image. (Courtesy St Thomas's Hospital Medical Physics Dept)



Reconstructed hip image from the multiple image-block codebook, showing a less pronounced blocking effect, better contrast and an improved reconstruction of the dark area in the centre of the image.

True rms measurement has become straightforward thanks to dedicated ICs. The information in this article has been compiled by Dan Ayers and *EW+WW* staff.



READING RMS

Since the advent of digital multimeters, engineers have been able to make quick, simple and relatively accurate voltage measurements for both dc and ac. A common and unhappy side effect of this however is an over-reliance on the seemingly definitive number on the readout.

Reference to the meter's specifications often shows that the last digit displayed may be far from the real value. A more fundamental question is whether even the range is appropriate. Although a low to middle priced multimeter is adequate for dc and certain ac measurements, a crucial range is usually missing - true rms.

Why rms ?

Virtually all electronic systems call for some means of monitoring ac voltage. It is easy to obtain the peak, or peak-to-peak, value of a signal by pumping a capacitor with a rectifier, and subsequent op-amp buffering is straightforward. This is useful to indicate when an amplifier or similar system is approaching its clipping limits.

A strategy used in many ac voltmeters is to show the mean average deviation, or MAD, of a signal from a predetermined reference, usually the mean. This so-called ac average can be useful, but a more versatile measure is the rms voltage of a signal. This fundamental quantity provides information about the energy available or used over time.

When applied to a resistive load for a given period of time, any signal of the same rms voltage would cause the same amount of heat dissipation. Sometimes described as effective voltage, rms corresponds to the dc voltage that would produce the same heating effect.

Often, the mean average deviation is displayed on a scale calibrated in rms volts. But this setup only shows a correct reading when the waveform applied is of the same shape as the waveform used to calibrate the meter. Many digital multimeters only give a valid ac reading for fairly low frequency, sinusoidal waveforms below around 400Hz.

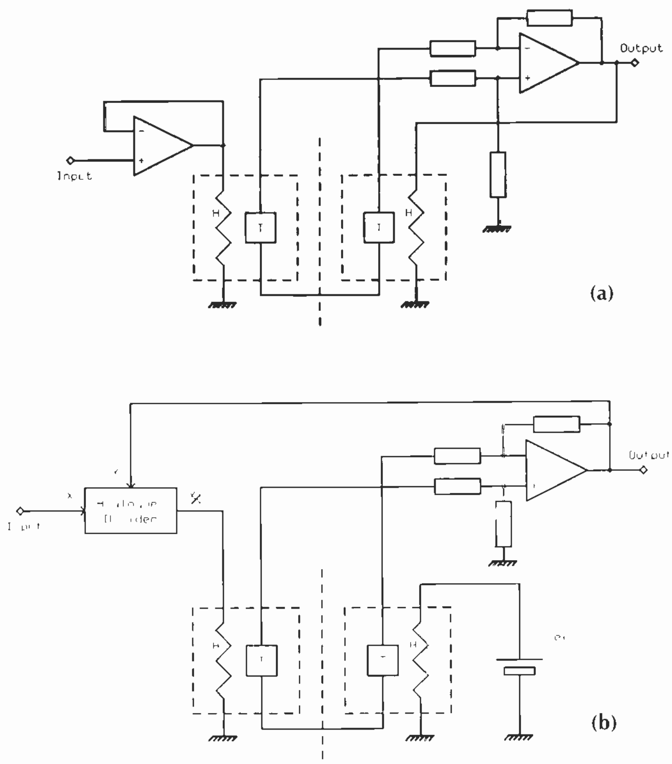


Fig. 1a). Deriving the rms value of a signal using two heater-sensor combinations. A DC voltage applied to a resistor produces exactly the same amount of heat as its equivalent in rms.
b). Dynamic range of the thermal converter is improved by adding analogue divider.

As long as the waveform is known, the true rms value of a signal can be calculated from the MAD. With many real-world signals such as noise and those associated with distortion however, this can cause problems. Comparing the MAD values with the true rms values for differing waveshapes clearly demonstrates the limitations, **Table 1**.

It is helpful that if unrelated signals are summed, then the rms of their sum is equal to the square root of the sum of the squares of their individual rms values. The rms value is also convenient for assessing signals with random characteristics. It represents the statistical standard deviation of a stationary zero-mean random process[1].

Circuit methods for true rms

For high accuracy, thermal methods of deriving the rms level of a signal are the most appropriate. This is because the heating effect of an ac voltage corresponds directly with the rms value, ie. that of the dc voltage required to produce the same heating in the same load. There are many drawbacks here, mainly due to the time taken for the temperature of different parts of the system to stabilise.

In the simplified thermal converter of **Fig. 1b)**, two units, each comprising a heater *H*, thermally coupled to a temperature sensor *T*, are thermally insulated from each other. The first is heated by the applied signal, the second is forced by the difference amplifier to the same temperature. If both units have identical thermal paths to the environment, then the output voltage is proportional to the rms value of the input.

A practical system might have thermocouple sensors and a chopper-stabilised device for the difference amplifier. This configuration suffers from limited dynamic range. Power through the heaters is proportional to the *square* of the rms voltage, and heater overload is a distinct possibility.

This problem is overcome in **Fig. 1b**). Here, the output amplifier still strives to maintain the temperature difference at zero, but now the power in the second heater is fixed. An analogue divider maintains equilibrium as its control voltage Y is proportional to the rms of input voltage X. As a result, the rms function is provided without the heaters having to function over an unmanageable range[1].

Convenience is much enhanced by using computational elements to obtain the rms value. Analogue-to-digital converters and digital processing are relatively expensive however. Fortunately, old-fashioned analogue techniques with modern manufacturing methods have resulted in accurate and easy to use integrated circuits.

The complete function required is:

$$E_{rms} = \sqrt{\left(\frac{1}{T}\int_0^T V_{in}^2 dt\right)}$$

Computation is simplified by considering the integration and division by T as a running average. In practice, this is valid for most types of signal encountered, so:

$$V_{rms} = \sqrt{\overline{V_{in}^2}}$$

There are two basic approaches to obtaining the true rms value of a signal – explicit and implicit[1]. The explicit or direct approach is shown in **Fig. 2 (a)**. Two inputs of a four-quadrant multiplier are fed with the input signal, producing a squaring function. Positive-going voltage created is averaged over time, and the square root of this dc value is taken. This can be done by inserting a squarer into the negative feedback loop of an amplifier.

Although good accuracy is possible, this approach is more complex and more expensive. In addition, dynamic range is at least an order of magnitude narrower than with a comparable implicit arrangement.

Dynamic range is particularly significant when measuring signals with a high crest factor, or cf. This is the ratio of peak to rms voltage. Obtaining a valid measure of a signal with a large crest factor needs a proportionately greater headroom.

The implicit approach follows from a little manipulation of the rms equation to:

$$V_{rms} = \frac{\overline{V_{in}^2}}{V_{rms}}$$

producing the more elegant configuration Fig. **2b**). Assuming an adequate *CR* time constant, the rms voltage output is held constant over the period of the signal being averaged and division by this value can be carried out before the average is taken.

Table 1. Comparison between mean absolute deviation and rms voltages for common waveforms. Mean absolute deviation is also known as ac average.

Waveform		RMS	MAD	CF
V_p	DC	V_p	V_p	1
V_p	Sine	$\frac{V_p}{2}$	$\frac{2V_p}{\pi}$	$\sqrt{2}$
V_p	Triangle	$\frac{V_p}{\sqrt{3}}$	$\frac{V_p}{2}$	$\sqrt{3}$
V_p, t, T	Pulse	$V_p\sqrt{\frac{t}{T}}$	$\frac{V_p t}{T}$	$\frac{1}{\sqrt{\frac{t}{T}}}$
$2V_p$	Gaussian (white) noise	—	RMS $\times\sqrt{\frac{2}{\pi}}$	∞ typically 1 - 6

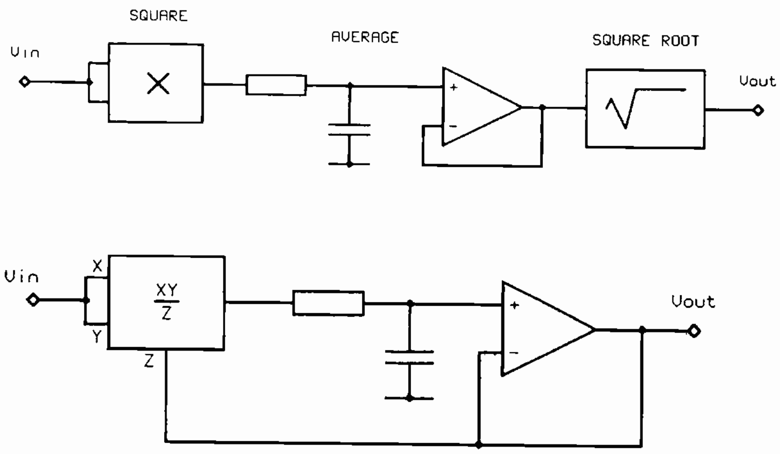


Fig. 2. Computation of rms voltage can be explicit, but implicit computation, (b lower), provides greater dynamic range.

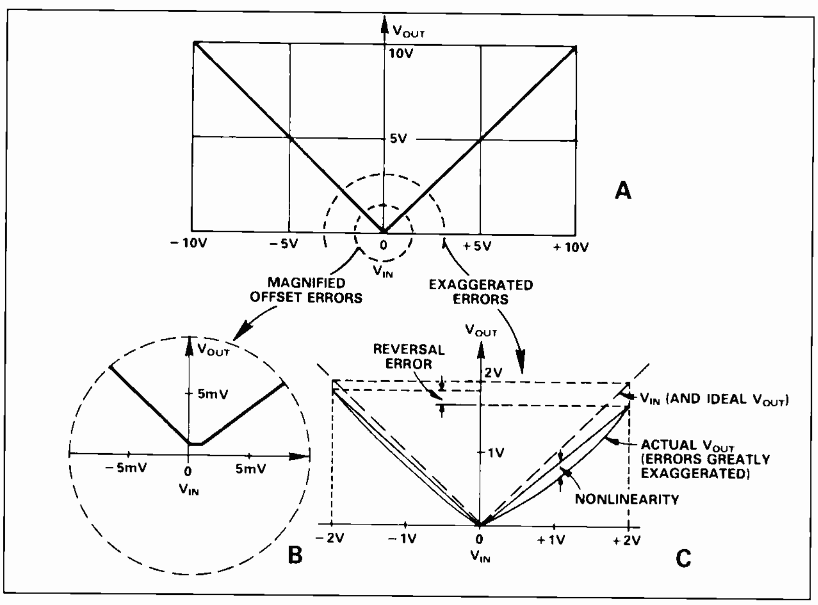


Fig. 3** (bottom). **Static errors in rms-to-dc converters. These errors are combined and expressed as a percentage of reading plus a constant.

Error sources in rms conversion

An ideal rms converter provides a dc output voltage exactly equal to the rms value of its input voltage, regardless of the amplitude, frequency, or shape of the input waveform. Of course a practical rms converter has errors.

Static errors are offsets and scale factor errors that apply to dc and low-frequency sinewave to about 1kHz. Under these conditions, the finite bandwidth of the converter – and the effective averaging time – can be made negligible compared to the input and output offset, and scale factor errors. Here, rms can be interpreted as the square root of the low pass filtered, or averaged, square of the input voltage.

An rms to dc converter's overall 'static' error is specified in percent of reading plus a constant. As shown in **Table 2**, the *AD637J* is specified at 1mV +0.5% of reading. This should be interpreted to mean that at any point within the *AD637J*'s 0V to 7V rms input dynamic range, converter output voltage will differ from the precise value of the rms input by at most 1mV plus 0.5% of the correct rms level. Note that this is less absolute error than the *AD536AJ* converter.

To illustrate this point, consider a sinewave input of 1V rms at 1kHz applied to the input of an *AD637J*. Actual *AD637* output voltage will be within: ±(1mV+0.5%×1V)=±(1mV+5mV). This is 6mV from the ideal output of 1.0V, or between 0.994 and 1.006V dc. These static errors can be classified into the standard categories of offset voltage, scale factor (gain) error, and nonlinearity errors.

Every practical rms converter has an input/output transfer characteristic that deviates from the ideal. The detailed error explanation given by **Figures 3a,b)** illustrate the major classes of errors commonly encountered.

At low levels, the rms converter's input offset voltages can flatten the point of the ideal absolute value transfer and take it more positive relative to the zero output voltage level with zero input voltage applied. Practical effects of these offset errors determine both the resolution and accuracy of the converter for low-level input signals.

For the ICs discussed here, the combined total of offset errors is typically less than 1mV. At higher input levels, of the order of few hundred millivolts, scale factor and linearity errors may dominate offset errors. A scale factor error is defined as the difference between the average slope of the actual input/output transfer and the ideal *1* to *1* transfer. If a 100mV rms input change produces a 99mV change in output, then the scale factor error is 1%.

In addition to the single polarity example just given, there can be a different scale factor for both negative and positive input voltages. The difference in these scale factors, termed the 'dc reversal error', is shown in Fig. **3c)**. When testing this parameter, a dc voltage is applied to the converter's input, say +2V, and then the polarity of the input voltage is reversed to –2V. Difference between the two readings will equal the dc reversal error.

Nonlinearity, as its name implies, is the curved portion of the input/output transfer characteristic. This is shown in an exaggerated form in Fig. 3c. This error is due to non-ideal behaviour in the rms computing section and cannot be reduced by trimming offset or scale factor.

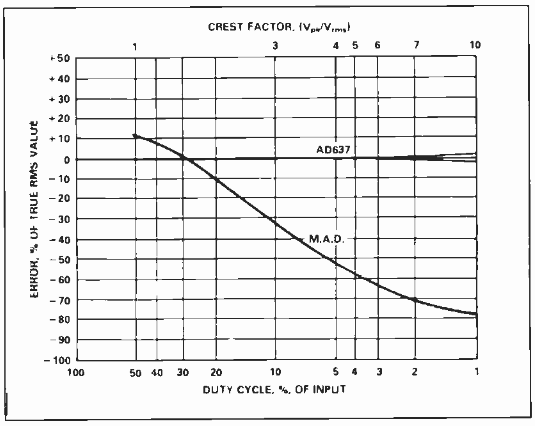


Fig. 4. Error versus duty cycle for an MAD ac detector and AD637-based rms converter.

Therefore, nonlinearity sets a limit on the ultimate best case accuracy of the rms converter.

For the *AD637*, nonlinearity is typically better than 1mV (0.05%) over a 2V full-scale rms range; for the *AD536A* the nonlinearity equals 5mV or less. Typically the *AD636* has less than 1mV nonlinearity over its 0 to 200mV specified input range.

As shown by **Fig. 4**, the errors of true rms to dc converters, although varied, are considerably lower than those errors found in precision mean-absolute deviation rectifiers when the duty cycle of the input waveform is varied.

TABLE 2: Typical rms-to-dc converter specifications.

	AD536AJ	*AD637J*	*AD636J*
Input dynamic range	7Vrms	7Vrms	1V rms
Nominal fsd rms	2V rms	2V rms	200mV
Peak trans. Input	±20V	±15V	±2.8V
Max total error			
No external trim			
mV/% reading	5mV ±0.5	1mV ±0.5	0.5mV ±1
Bandwidth, (-3dB)			
Full Scale	2MHz	8MHz	1.3MHz
0.1 V rms	300kHz	600kHz	800kHz
Error at Crest Factor			
of 5, rms	-0.3%@1V	±0.15%@1V	-
0.5%@200mV			
Power supply			
Volts min	±3	±3	+2/-2.5
max	±18	±18	±12
Current typ.	1mA	2mA	800μA
max	2mA	3mA	1mA

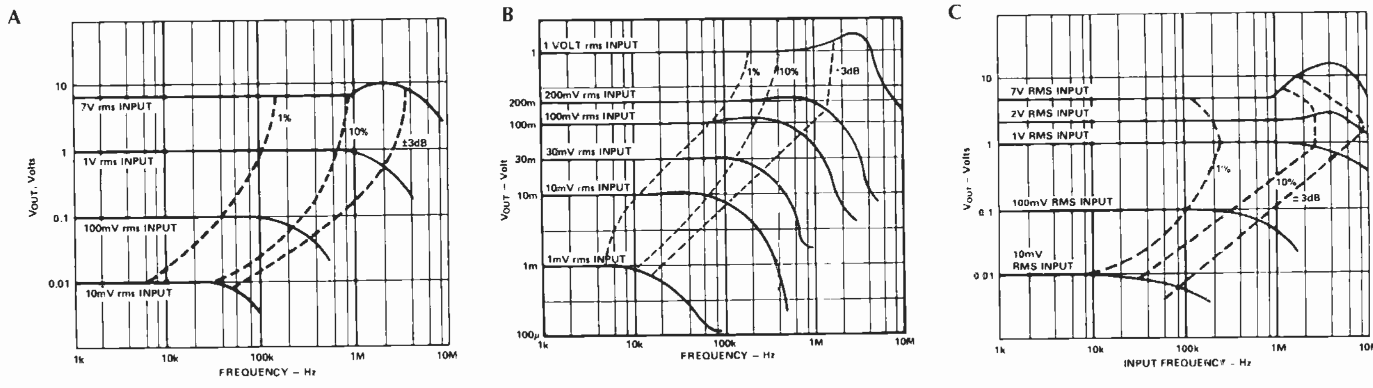


Fig. 5. High frequency response for the three converters – AD536A *at a),* AD636 *at b) and* AD637 *at c).*

Bandwidth considerations

In practice, ac inputs are of the most interest to users of rms converters. For 1kHz sinewave inputs, there is negligible difference between readings at this frequency and performance at dc. As a result, dc measurements provide a convenient way of determining errors at around 1kHz.

At higher input frequencies, bandwidth characteristics of the rms converter become most important. As shown by **Figs 5a,b,c)**, ac bandwidth drops off as the input level is reduced; this is primarily due to gain-bandwidth limitations in the absolute value circuits.

Caution should always be used when designing rms measuring systems which must deal with complex waveform amplitudes above 1V rms. Trimming is recommended for applications needing the lowest possible offset and scale factor errors, **Figs 6a,b,c)**. Ground the signal input point, V, and adjust trimmer R_4 for an output of zero volts. Alternatively, R can be adjusted to give the correct output with the lowest expected value of V_{IN} applied. This second method allows the lowest possible error over the expected input range, but results in higher errors below this range.

Connect a 1kHz calibrated full scale input to V_{IN}. Adjust trimmer R_1 to give the same output voltage. This adjustment provides specified accuracy with a 1kHz sinewave input and slightly less accuracy with other input waveforms.

With correct trimming, the remaining errors in an rms converter will be due to nonlinearity effects of the device; unfortunately, nonlinearity errors cannot be reduced by external trimming.

Practical circuits

Although it is possible to produce close approximations to squaring and square root functions relatively directly, log/antilog blocks can give greater accuracy and simplify initial setting up[2]. These blocks are often based on the exponential response of transistors.

Figure 7 uses two standard chips to produce a log/antilog implicit rms converter which is adequate for many applications; the separate computing elements are clear to see.

Several companies produce dedicated rms chips, and the circuit of **Fig. 8** shows how straightforward such devices are to apply. The *SSM2110* is a particularly versatile device. With a minimum of external components it can provide rms, absolute value and peak conversion, or alternatively the log of any of these[3]. **Figure 9** would be suitable for a meter calibrated in decibels – very useful for audio work.

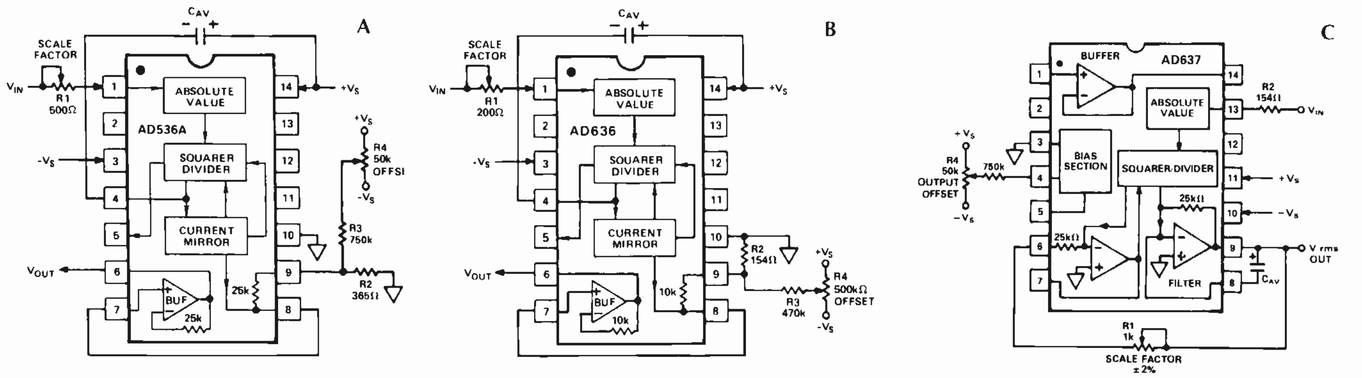


Fig. 6. Internal functions together with external offset and scale factor trimming circuits for AD536A *at a),* AD636 *at b) and* AD637 *at c).*

High impedance input rms dpm and dB meter

Only two integrated circuits and a liquid crystal display are needed to produce high quality, dpm/dB meter.

Voltage input to the meter feeds through a 10MΩ input attenuator to pin 7 of the *AD636*. Buffer output, pin 6, is ac coupled to the rms converter's input, pin 1. Resistor R_6 provides a 'bootstrapped' circuit to keep the input impedance high.

Output from the rms converter is selected by the linear/dB switch; selecting pin 8 for linear, pin 5 for dB. The selected output travels from the linear/dB switch through low pass filter R_{15}, C_6 to the input of the meter chip, which is a *7106* type a-to-d converter. The *AD589*

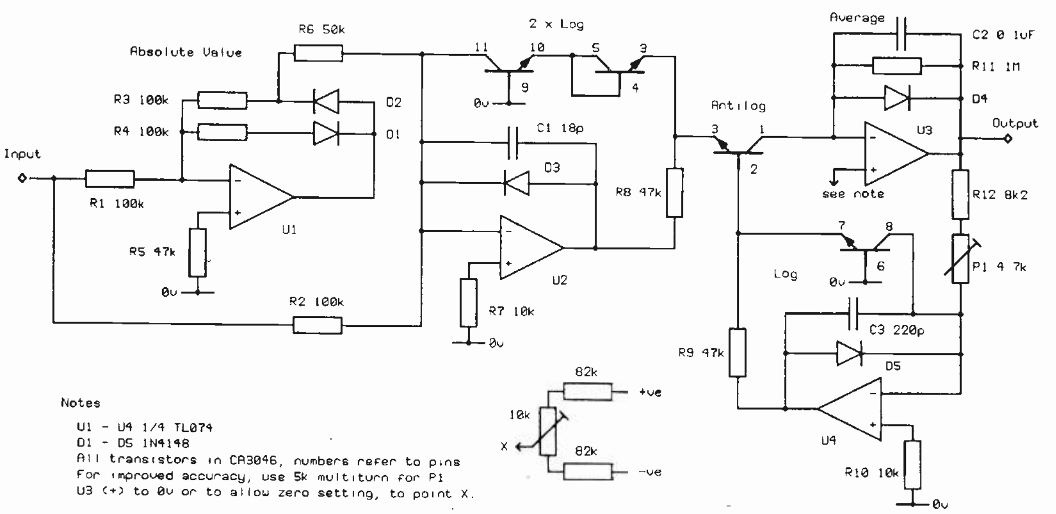


Fig. 7. Converter for rms measurement using standard chips shows log-antilog calculation of square/square-root functions.

provides a stable 1.2V reference voltage for supplying the calibration circuitry.

To calibrate, first adjust trimmer R_9 for the 0dB reference point. Next, set R_{14} for the decibel scale factor, and finally, adjust R_{13} to set the linear scale factor. Total current consumption is typically 2.9mA from a standard 9V transistor radio battery.

This circuit uses the *AD636* low power rms converter to extend battery life and provide a 200mV full scale sensitivity. It provides better accuracy and bandwidth at 200mV rms input than the *AD536A*, which would need preamplifier to achieve similar results.

Programmable-gain rms measurement

Measurement of the rms of complex waveforms of varying magnitude normally requires a high quality, compensated input attenuator. In contrast, the programmable gain rms preamplifier circuit of **Figure 10** features an *AD544* bifet operational amplifier as an inverting input buffer with four remotely switchable gain ranges: 200mV, 2V, 20V, and 200V full scale.

Switching gain resistors in the buffer feedback loop allows the use of a low voltage cmos multiplexer to remotely control the gain of potentially high voltage input signals. The preamplifier's input is well protected on all ranges for input voltages up to 500V peak.

Input connects to J_1, with R_1 and diodes $D_{1,2}$ forming the amplifier's input protection. Capacitor C_1 prevents high frequency roll-off, which would occur due to the *R/C* time constant of the 1MΩ input resistor and the stray capacitance at the *AD544* summing junction. The *AD7503* cmos multiplexer switches the appropriate feedback resistor for each gain connecting the resistor between the operational amplifier output, pin 6, and its summing junction, pin 2.

Capacitors C_{4-7} are compensation capacitors which are adjusted for flat response at each gain setting. Address lines A_{0-2} select the desired input range of the preamplifier. Resistors $R_{4,6,10,12}$ are gain calibration controls for each selected gain. Output of the *AD611* operational amplifier is converted to its rms equivalent voltage by the *AD536A* rms-dc converter.

Input ranges are 200mV, 2V, 20V and 200V rms. For the respective ranges, –3dB bandwidth points are >4kHz, 600kHz, 1.5MHz and 600kHz. For the lowest range, bandwidth will vary with the degree of stray capacitance at pin 9 of the *AD7503*.

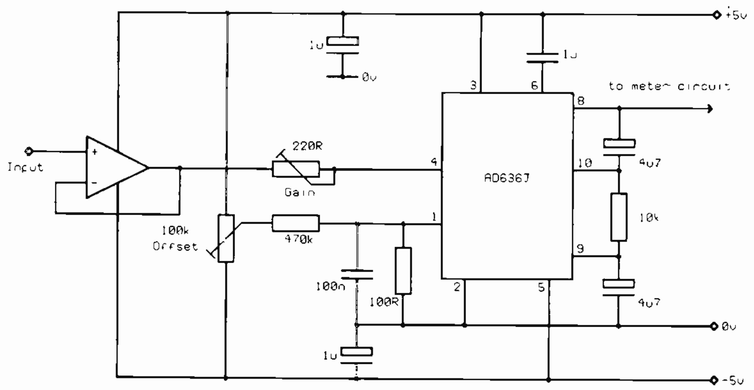


Fig. 8. Dedicated rms converter chips can reduce component count and improve accuracy.

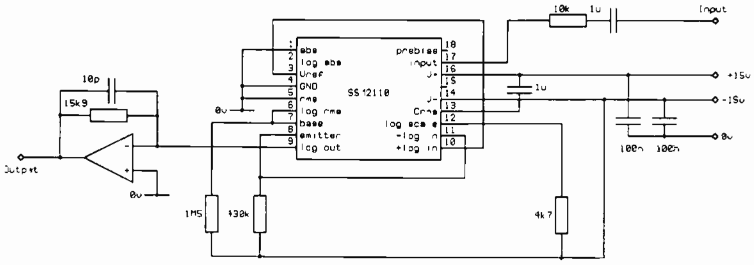


Fig. 9. Versatile converter chip configured for log of rms conversion. This configuration is useful for audio decibel metering.

Testing converters

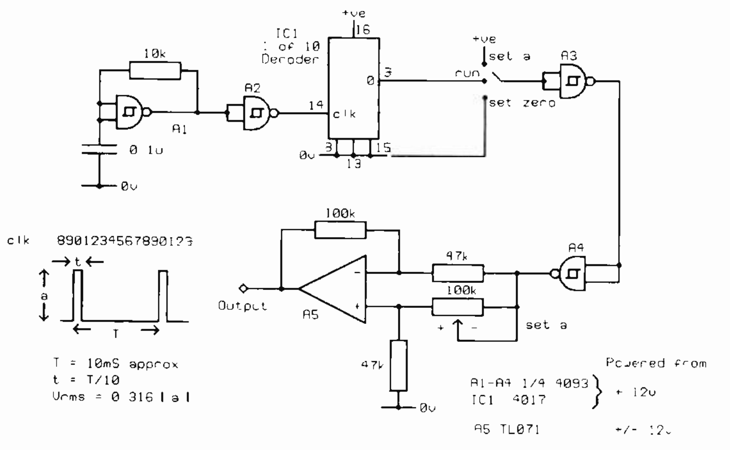
To calibrate and assess the accuracy of an rms converter, many factors need to be considered – particularly the dc response (offset), frequency response (gain) and dynamic range. Laboratory equipment is desirable, but a good overall picture can be gained by feeding a pulse waveform of known amplitude and mark/space ratio into the converter. This is because the pulse contains frequency components extending to infinity – in theory at least – and calculation of the crest factor and true rms value is straightforward.

The circuit shown generates reasonable pulses with variable amplitude and a mark/space ratio fixed at 1:10. A simple clock with a frequency around 1kHz is built around a Schmitt trigger. This frequency may be varied over a wide range by altering the resistor and/or capacitor values. Clock output is sent to a 1 of 10 decoder to fix the mark/space ratio and the pulses are cleaned up by the remaining Schmitt triggers.

For controlling amplitude, an op-amp is configured to provide variable gain giving an output pulses from –10V to +10V referred to ground. The circuit can suffer from ringing on the pulse edges. This affects the rms level, especially at higher pulse rates. Should this be a problem, it is advisable to strap a variable resistance of around 47kΩ between the op-amp input pins and trim for best shape.

Before testing an rms circuit, the zero should first be checked and any offset noted. A suitable dc reference voltage should then be set at, say, 5V. It is important that the circuit under test is connected before setting the reference to avoid loading errors. This will also confirm that the converter is responding correctly to dc.

Switching the pulse generator switch to run should produce a dc voltage at the output of the converter of around 1.6V. Its true rms value is 0.316x5=1.58V.



Simple pulse generator with fixed mark/space ratio and variable amplitude allows easy assessment of rms converter accuracy.

Noise referred to the amplifier input is 360μV on the 2V range while the signal-to-noise ratio is 75dB. Output settling time is 397ms to reach 1% of input.

Address lines $A_{0\text{-}2}$ should be set for each gain. Calibration trim potentiometers $R_{4,6,10,12}$ should be individually adjusted for the correct gain on each range.

Compensation capacitors $C_{5,6,7}$ should be adjusted for flat response on each range. For this, use a variable frequency sinewave input signal and an oscilloscope to monitor the *AD544* output, pin 6. Alternatively use a digital voltmeter on its dc scale connected to the converter's output.

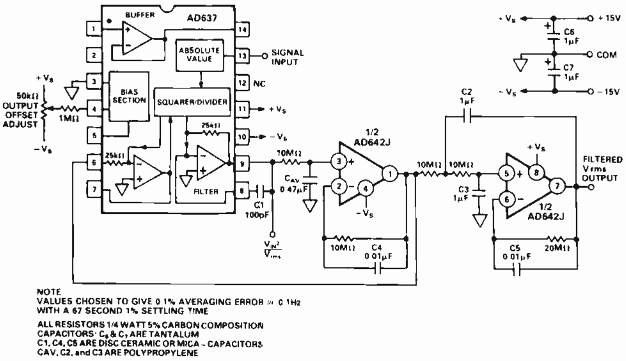


Fig. 10. With signals as low as 0.1Hz, this circuit exhibits less than 0.1% averaging error.

Reading ultra-low frequencies

Reducing input frequency requires lengthening the averaging and filtering time constants to maintain the same levels of dc error. Consequently, successively larger values of C_{AV} are needed. With very large values of averaging capacitor, needed for frequencies below 10Hz, C_{AV} can become physically too large and also prohibit the use of low-leakage devices.

Figure 10 uses two very low input bias current amplifiers, permitting large values of averaging resistance – in this case 10MΩ. This circuit has been optimized to exhibit less than 0.1% averaging error for input signals as low as 0.1Hz. The V_{IN}^2/V function appears at pin 9 of the *AD637*.

As a result of transient noise spikes, the circuit may overload because the filter stage averaging capacitor has been drastically reduced. Normally, the averaging capacitor is called C_{AV} but in this case it has been renamed C_1. Reducing the capacitor allows output at pin 9 of the *AD637* to respond to the square of the input signal rather than to the average of the input square

For applications where high crest factor-low frequency signals are to be measured, C_1 should be increased to 3.3μF. In conjunction with the internal 25kΩ filtering resistor, this capacitor forms a low-pass filter with a 2Hz corner frequency. This attenuates higher frequency signals – transients – by the ratio of the transient frequency to that of 2Hz. This means that in the case of 60Hz transients, they will be reduced by 60Hz/2Hz or 30 times. Practically speaking, there will be effective transient protection.

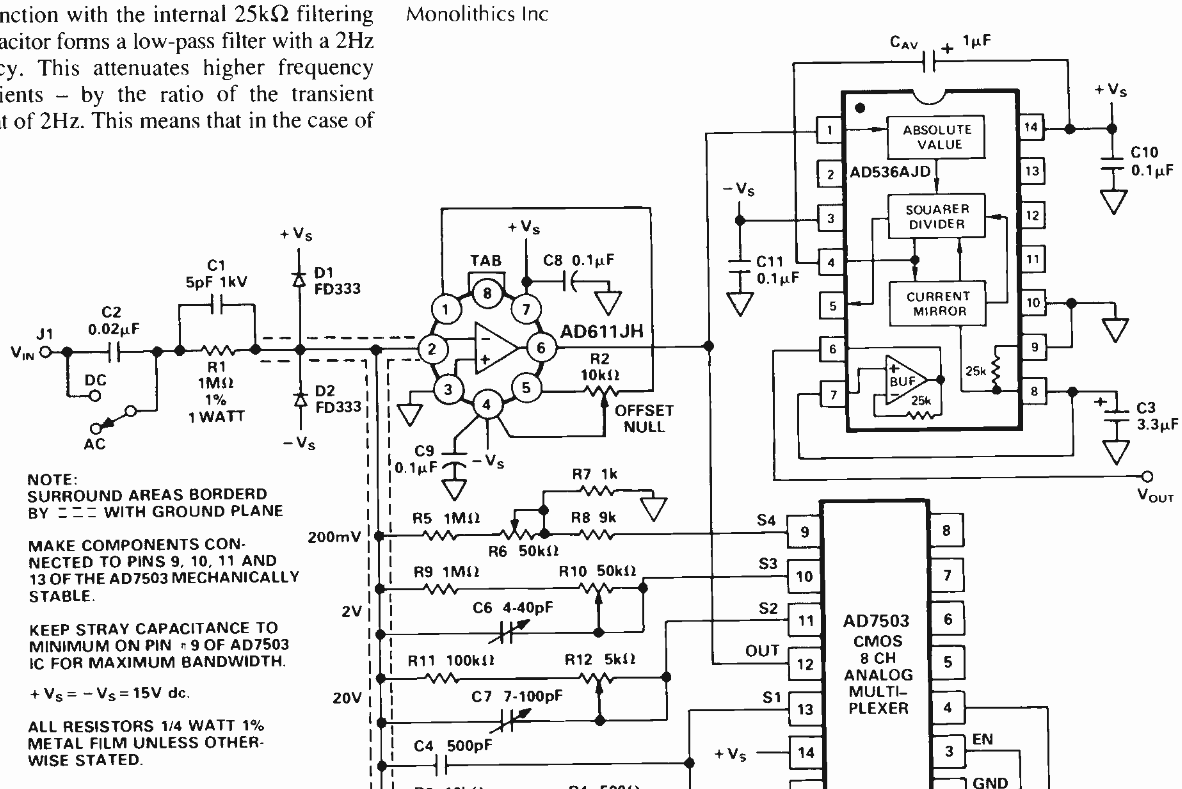
In addition, larger or smaller values of C_1 may be used as required by the specific application. If a low-pass filter is used ahead of the *AD637*, out-of-band signals are less likely to cause an overload. This allows smaller values of C_1 to be used in these circuits.

Since raising C_1 causes increased averaging of higher frequency signals, the V_{IN}^2/V_{rms} function will be linearly converted to the average of V_{IN}^2/V_{rms} as the input frequency goes up. This prevents the instantaneous square of the input signal from appearing at pin 9 of the *AD637*. ■

References

1. D H Sheingold (Ed.), *Nonlinear Circuits Handbook*, Analog Devices Inc. ISBN 0-916550-01-X.
2. D.Ayers, *The Twisted World of Non-Linear Electronics, Electronics World & Wireless World*, Feb 1993.
3. *SSM Audio Products Audio Handbook*, Vol. 1, Precision Monolithics Inc

Fig. 11. Measuring rms of complex signals normally involves an expensive attenuator. This programmable-gain circuit, with 200mV, 2V, 20V and 200V ranges, does the same job.



GAIN	A0	A1	A2	MAX V_{IN}
10	H	H	L	200mV
1	L	L	L	2 Volts
1/10	H	L	L	20 Volts
1/100	L	H	L	200 Volts

EW+WW gratefully acknowledges the help of Analog Devices in the preparation of ths article

Visual Designer: Easy to acquire?

Although doubts remain about the capability of Windows to run real time tasks such as data acquisition and process control, its well known user interface makes it a logical choice for first time system builders. Allen Brown looks for compromise in this new data acquisition package

The growth in PC data acquisition products continues and although the theme remains basically the same, there are variations in style. Much of today's software has the instant appeal of being readily accessible to the new user. This has come about by the almost universal acceptance of Microsoft Windows with its Graphics User Interface front end. Gone are the days when a new commercial software package demanded a entirely fresh learning curve.

For those who want to perform data acquisition without a steep learning curve, Intelligent Instrumentation have recently released *Visual Designer 2* which is a combined software and hardware product. The software is essentially an icon driven package and, as its name suggests, has a highly visual aspect to it. It forms a friendly interface between the user and the proprietary data acquisition cards in the PC.

Operational systems are created by pulling functions – represented by icons or blocks – out of the libraries and positioning them on the PC screen. Once in place they are linked as required to generate what are known as FlowGrams. On issuing the RUN command, the design is compiled into usable code, referred to as FlowCode and executed. These latter compilation stages are quite transparent to the user.

As the newly designed system is operating, the appropriate output devices (panel meters and 'scope displays) are updated to provide real-time operation.

To realise the full potential of the software it is necessary to have appropriate i/o expansion card(s) in the PC. As yet, software library support is only offered on cards manufactured by Intelligent Instrumentation, the *PCI-2000* Series for example. They also manufacture Visual Designer supported i/o cards with the EISA interface.

The software is installed with ease, provided that older versions are flushed out. During installation, Visual Designer replaced the Microsoft Windows Direct Memory Access (DMA) Manager with its own (pcivdmad.386 is added to the Windows system.ini file). This appears to be a common practice with a number of data acquisition software packages. Is this due to a lack of confidence in Microsoft's version one asks?

The software comes in two sections: the first is equivalent to an editor, where instead of using words, the user enters schematic blocks or icons from the libraries. The second section, called RUN is where the FlowGrams are compiled and executed. The two sections are quite separate in their execution. DIAGRAM basically acts as an editor where the design is constructed whereas RUN is the environment where the design is realised. Calling RUN from DIAGRAM is quite seamless but can be rather slow on a 386-PC, especially for complex tasks.

Blocks

A system design constructed in *Visual Designer* will consist of interconnected blocks which are accessed from the blocks options in menu bar. These blocks (for example add, subtract analogue i/o, plot and chart) will have either input or output channels (or both) depending on their on function. The actual blocks are stored as Windows dynamic link libraries (DLLs) and are accessed when the design is compiled (RUN). The concept behind this construction lies in the probable need to have the DLL library easily updated as new blocks become available either from the manufacture or through the user's own industry.

The design procedure for creating a system consists of three parts; selecting the required operational blocks; forming the interconnections between them and lastly, configuring the blocks to your specifications – for example, setting sample rates on an i/o card. Selection is achieved by simply accessing the drop-down menu from the blocks option and clicking the mouse or the required function. The

block position defaults to the centre of the screen, then by using the mouse it can be positioned where needed. This is particularly useful especially for constructing complex structures with many tens of blocks.

Initially is it difficult to give any style consideration to the design; it must be sufficiently fluid to allows changes as the structure build up. The blocks can be cut, copied and pasted like any other Windows object – all the elements of a good text editor.

Once the blocks are in place, they can be linked together by using the wiring tool. The whole action of wiring is achieved by using the mouse's click and drag facility. Examples are shown in the figure. When moving a block the wire(s) connecting it to other blocks automatically follows the block. Each wire has a direction arrow attached to it to indicate the direction of the data flow. As the link is accepted on a target block, a block connection dialogue box appears which gives details of what's connected to the two ends of the wire – the inputs and the outputs. The dialogue box gives a brief description of the connecting blocks.

Although each block has a default set of parameters, the user has the option of configuring each block according to their specification. This is third part of the design: double-clicking with the mouse will generate the Configuration dialogue box for the block. The user is then able to make adjustments to the operation of the block. These include the optional data types which can have various integer or floating point sizes. When the completed structure is ready to compile, the user has the option of generating a list of all the blocks (DLL list) which act as a sort of parts list.

I/O features

Since *Visual Designer* is specifically tailored for the i/o cards from Intelligent Instrumentation, its designers have included many blocks which exploit the full range i/o features. On selecting the i/o option from the blocks menu as expected an i/o icon is generated on the screen (**Fig. 1**). The software then offers a variety of modes of configuring the i/o features. Great emphasis is placed on the DMA features found on a number of expansion cards. The data transfer can occur via a DMA channel which does not involve the PC's CPU direct operation or through software which is executed by the CPU.

When deciding on the *Visual Designer* product, the potential user must be able to specify their data acquisition requirements, this will determine the choice of cards. For example, for following i/o cards all have DMA features for rapid data transfer:

PCI-20501C Series:	High performance EISA cards
PCI-20098C Series:	Multifunction cards
PCI-20091W-1:	High speed analogue input card,
PCI20377W-1:	Low power multifunction card.

Whatever expansion cards you use, the software will have a support library for it. At this time it's not possible to integrate the software to other manufacturers i/o cards – you have to choose the cards available from Intelligent Instrumentation.

For very high speed data transfer (input or output) there is 10MHz Input buffer block which is selected from the drop-down menu. This uses the Intelligent Instrumentation *ZPB6064* Buffer Board. It has a circular fifo buffer which can be set up to a value of 65,535 (unsigned integers). To achieve the 10MHz data rate the *ZPB6064* requires the external *ZPD1002*, 12-bit analogue box.

Alternatively for 16-bit resolution, you must sacrifice capture rate and settle for the *ZP1003* analogue input box which samples at 500kHz. Other i/o features include the BurstGenerator block for generating multiple TTL pulses, RateGenerator block for generating periodic TTL pulses and the Timer block for producing low frequency TTL pulses – all quite useful for driving stepper motors.

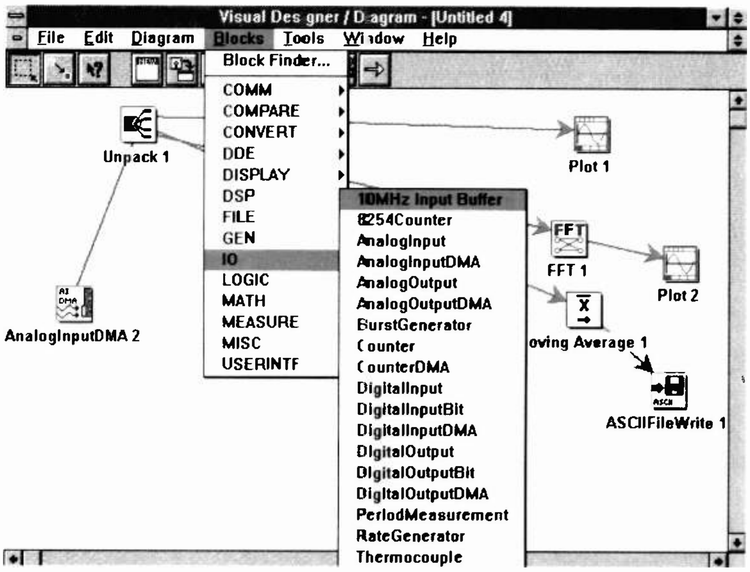


Fig. 1. The blocks are accessed from the drop-down menus, here the i/o options are shown.

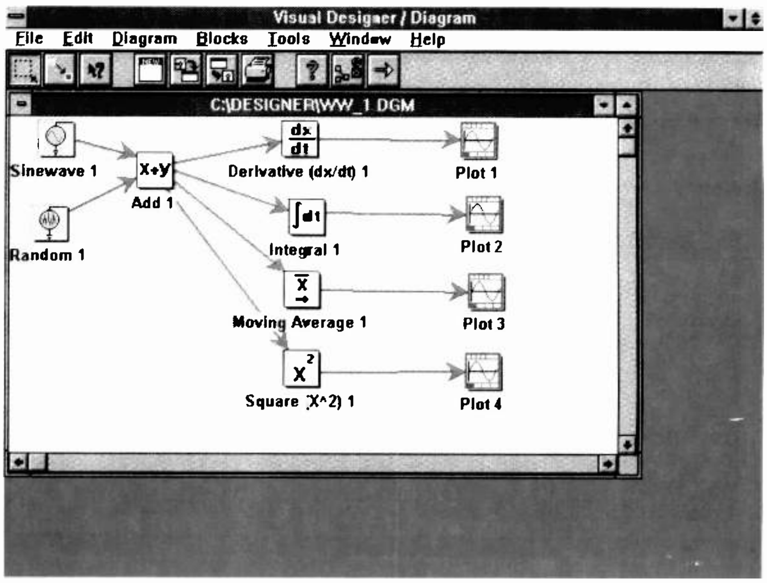
Functional blocks

Once the blocks feature from the menu bar is evoked, *Visual Designer* is not dissimilar to writing a pictorial computer program without the obvious flow control constructs. However it does possess functions for performing mathematical operations, general data processing, data display and i/o.

An example of some of the maths function blocks is shown in **Fig. 2**, the resulting RUN is shown in **Fig. 3**. In these figures the signal sources are a random noise and a sine wave generator. Note that one of the blocks performs a moving average which has the effect of cleaning up the signal. Another function from the maths menu of note is the $mX+b$ block. This has the effect of scaling the input X and adding an offset. It could be used to convert the units of X (say a voltage derived from a pressure transducer) into units of pressure (Newtons per m^2) – quite useful provided the transducer behaves in a linear manner.

Visual Designer also has a facility for enabling data transfer up to 115,200bps via a serial port on the PC (RS232 interface). All the options for setting the port are accessed through the comm block. A lot of effort has gone into the design of this feature which allows the user to

Fig. 2. An example of the mathematical operations on offer from Visual Designer.



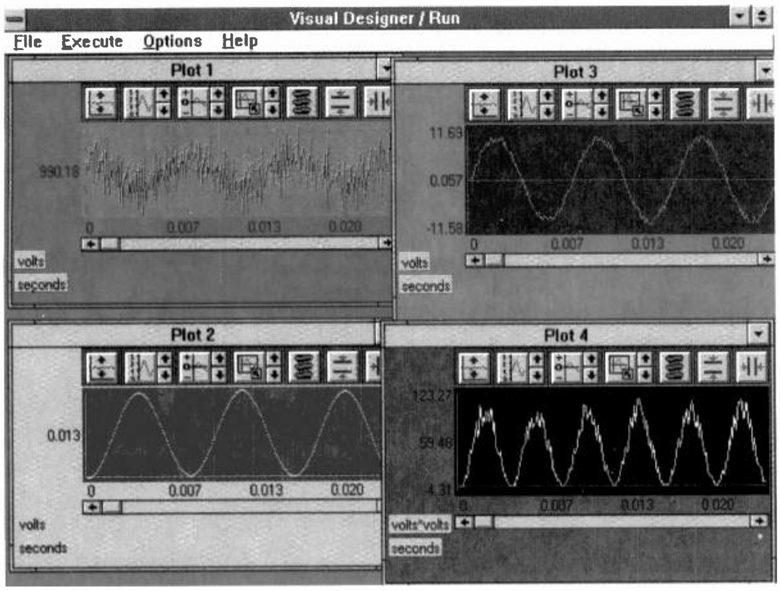


Fig. 3. The graphical outputs of the operations from Fig 2.

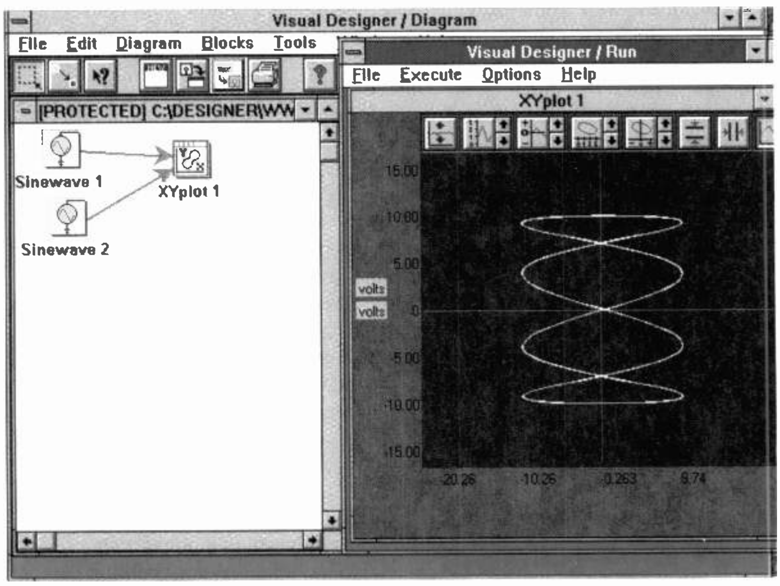
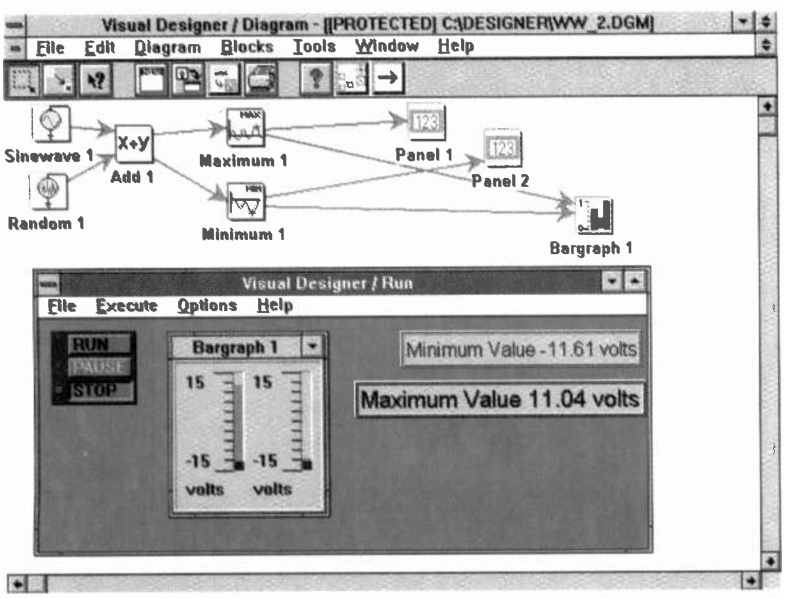


Fig. 4. Illustration of the XYPlot block and display.

Fig. 5. Panel and Bar graph displays generated from visual Designer.



exercise considerable control over fixed or dynamic data together with buffer sizing and data formats. However there is no mention of the possible problems that can arise from the universal asynchronous receiver transmitter (UART) in the PC. On most models, even 486-PC, the UART is the 8-bit *8250* which can only accommodate rates up to 19,200 when running Windows. Ideally if using the PC for high transfer rate then the 16-bit *16550* UART should be used which has a large buffer. This device can accommodate the constant interrupts from Windows without loss of data.

Dynamic Data Exchange

One of the main principles in the design of Windows, or for that matter any other multitasking operating system, is the ability to allow concurrent tasks to exchange data. In Microsoft Windows this is done by means of the Dynamic Data Exchange (DDE). Visual Designer has included blocks which effect DDE, they are the DDEServer and the DDEClient. The DDEServer will accept data from any source within a Visual Designer structure (a block) and pass it onto other concurrent Windows tasks such as *Excel* or *Word for Windows*.

Three attributes are attached to a DDEServer: the server topic name which identifies the DDEServer block to use (each must have a unique name), the name of the server application and the name of the data to be transferred. The parameters of the data buffer connecting the client and the server (size, units, weighting and offset for example) can be adjusted by the user. The user also has the option of controlling the data flow, whether it is transferred as soon as it becomes available or when *Visual Designer* has stopped. The DDEClient block allows the software to receive data from other concurrent Windows tasks.

The optional output of the DDEClient block is a data buffer (its size specified by the user) for holding double precision floating point numbers. As with DDEServer it requires a server name, a topic name and an item name.

It can also take advantage of the local area networking facilities of Windows 3.11 For example, a PC on the LAN could be at one location running a spreadsheet, whilst the PC with *Visual Designer* could be at another location. As the data becomes available from the acquisition card, by using the DDEServer it can be immediately transported via the LAN to the spreadsheet running on the remote PC.

Screen display blocks

There are a number of screen display blocks available in Visual Designer. The XYPlot can be used to display data from two separate buffers to create Lissajous type figures (**Fig. 4**). The plot block produces a displays which are not too different from that of an oscilloscope. There are number of interactive control icons on the top of the plot to change the display parameters. However one of the drawbacks of the plot feature is it only allows one signal to be displayed.

The panel block allows the user to display slowly varying information such as peak or minimum values – as seen in **Fig. 5**. Again the user has a lot of freedom in designing the appearance of the panel. The chart block gives the opportunity to display a continuous stream of data – as in a chart recorder proper except masses of paper is not generated. Up to eight input channels can be fed into each chart block which gives it the feel of a real chart recorder – the data streams can also be saved.

Alternatively the input data may be represented using the bargraph block.

A curious variant on the display is the analogue meter block. The sales brochure shows a design for an engine test system which uses analogue meters quite effectively for representing fuel flow, fuel quantity and rpm.

User's Manual

The User's Manual comes with a Reference Manual and a Guide on the expansion card(s). On the whole these are well written with good presentation. Bearing in mind that Visual Designer should appeal to the user who has no prior experience of data acquisition, I feel it would be have been appropriate to include more examples in the manuals. An enlargement of the *Getting Started* could go some way of addressing the needs of the new user.

And finally...

It can be argued that having a proprietary package of software and hardware eliminates many of the difficult problems that arise when trying to integrate manufacturer *A*'s software to manufacturer *B*'s hardware. With *Visual Designer* this problem does not arise since the software is designed specifically for the hardware expansion cards. The design of the package is very appealing and instantly accessible. The new user will gain confidence very quickly having taken the plunge. Although the block library is quite extensive there are certain areas which are sparsely serviced. The DSP sub-menu is thin with only FFT and power spectrum options. Generally the signal conditioning operations are not as abundant as one would like (filters for example). However, the modular design will allow libraries to be augmented as the product grows. With these reservations, *Visual Designer* is well worth considering for the user who wants a data acquisition system up and running within half hour. ■

SYSTEM REQUIREMENTS

486DX-PC
8MB of system ram
10MB of Hard disc space
Mouse
Graphics accelerator card for SVGA
Intelligent Instrumentation data i/o acquisition card Microsoft Windows 3.11

SUPPLIER DETAILS

Intelligent Instrumentation Ltd, Suite 5, 2 Penn Place, Northway, Rickmansworth, Hertfordshire WD3 1RE. Phone: 0923-896989

PRICE

Visual Designer 2 Software:	£595
PCI-20098C Multifunction Card:	£879

Magic numbers IN ELECTRONICS

The Greeks knew about the golden ratio, π and (√5+1)/2. But although Pythagoras and Euclid never had the opportunity to apply their work to filter design, it still remains a valid proposition. Ian Hickman brings civilisation up to date.

Ever since man started to count, numbers have fascinated him. Starting with the positive whole numbers (up to ten, perhaps, initially) man at some point realised that there is no largest number, and eventually came realise that there were other 'numbers' in between the whole numbers he was used to.

For instance, while rulers of lengths 3, 4 and 5 cubits would let him build nice tidy right angles at the corners of a palace or house, the circumference of a barrel obstinately refused to equal a whole number times the diameter, although twenty-two sevenths seemed to be near enough for most practical purposes.

Nowadays pi crops up in technical contexts, in electronics as elsewhere, for instance in $\mu_0=4\pi 10^{-7}$, the permeability of free space.

Most readers of this journal will be conversant not only with pi but also with e, the base of exponential or Naperian logarithms. Like pi, e is a truly magic number, popping up all over the place. It is also a number to be wary of – the exponential function has a dangerous tendency to explode.

For instance, suppose that in 1066 near Hastings, one of William the Conqueror's soldiers wantonly did £1 worth of damage to the property of a local landowner, and that the landowner's descendants today obtained a court order for the payment of this sum with interest at a modest rate of, say, 2.5% per annum compound. Then the successors in title of William the Conqueror face a bill of eight thousand nine hundred and forty eight million, four hundred and thirty four thousand eight hundred and ninety eight pounds.

If instead, 1.25% interest had been added six-monthly, the figure would have been slightly higher. If interest had been added not six-monthly, monthly or even daily, but one millionth of the annual rate added every 31.5 seconds (one millionth of a year), the figure would have been slightly higher still, the effective annual rate becoming 2.531%. The total, instead of increasing in yearly steps, would have mounted up following a smooth curve described virtually exactly by an exponential function, in this case:

$$Total=Principal \times e^{0.025t},$$

where t is in years. You can imagine how rapidly the exponential function explodes if t is in seconds or even microseconds, especially if 'a' in e^{at} were unity or larger, rather than 0.025.

That an exponential cannot go on growing for ever is well known to engineers, but completely unknown to politicians (with their talk of continuous sustainable growth), over-optimistic business men, or the poor unfortunates taken in by cleverly

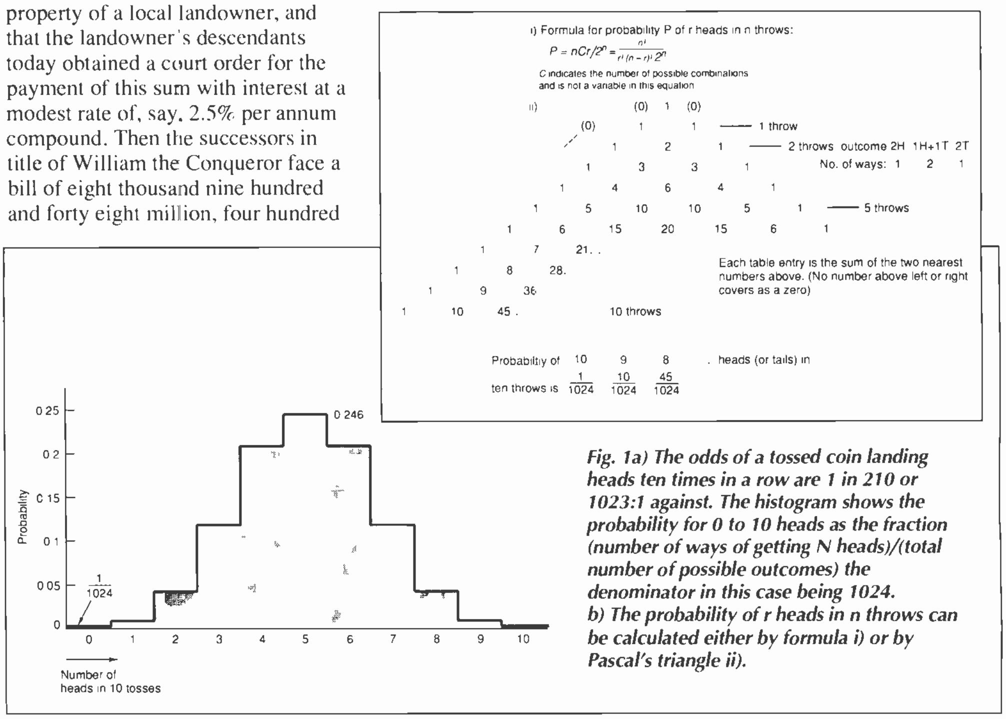


Fig. 1a) The odds of a tossed coin landing heads ten times in a row are 1 in 210 or 1023:1 against. The histogram shows the probability for 0 to 10 heads as the fraction (number of ways of getting N heads)/(total number of possible outcomes) the denominator in this case being 1024.
b) The probability of r heads in n throws can be calculated either by formula i) or by Pascal's triangle ii).

disguised chain-selling schemes. Exponential decay is much more well behaved, the voltage across a parallel resistor/capacitor combination dying away, like the world, with a whimper according to the equation

$$V = V_0 e^{\frac{-t}{CR}} \qquad \ldots 1$$

This tells you what voltage is left across the capacitor t seconds after some arbitrary time of observation t_0 at which the voltage across the capacitor was V_0. Assuming that originally the capacitor was charged up to some enormous voltage, you can find out what the voltage was at any time *before* t_0 by letting time run backwards. Just substitute $-t$ for t in equation 1, converting it into a positive exponent and a growing exponential.

In equation 1, the variable is time, but e appears in other equations where the variable is squared. I can't think off hand of any equations where the variable is time squared, but other equations with e to the power (a variable squared) often occur. Naturally, to avoid explosions, the squared variable has an associated negative sign, just as time does in equation 1.

Because there is no difference between x^2 and $(-x)^2$, a function defined by such an equation dies away as the variable increases in either a positive or a negative direction away from the norm or mean, as the following equation shows. It describes Gaussian noise – noise having a Gaussian or 'normal' distribution:

$$\text{Probability density of voltage } V = K_1 e^{-K_2 V^2}$$

There is no maximum value to this function: in theory you could get a voltage spike of near infinite magnitude, but as the probability of this is near zero, you would have to wait for ever for the chance to observe it.

Incidentally, the same equation governs something as mundane as the tossing of a coin – at least if you do it often enough. **Figure 1** shows the possible outcomes of tossing a coin ten times in a row. Interestingly, five heads and five tails is not the most likely outcome. Six of one and four of the other (not specifying which) is much more likely, though five of each is marginally more likely than six heads and four tails (or four heads and six tails). If you make a histogram of the number of possible ways of getting 0, 1, 2 heads, the result closely resembles the distribution of Gaussian noise shown in **Fig. 2**. As the number of tosses approaches infinity, the histogram converges ever more exactly on the normal curve.

We count in tens because we have five fingers on each hand. But 5 – or rather its square root – is the basis of another magic number which seems to be not at all well known, and which I will call K. This number is

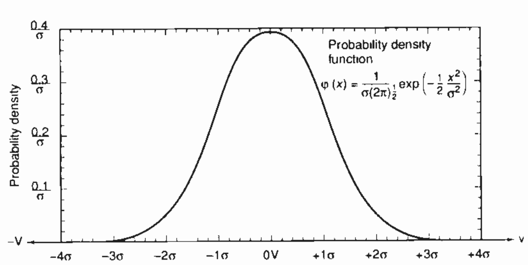
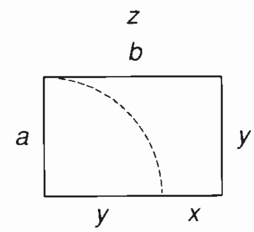


Fig. 2. The probability of the instantaneous value of random noise falling at any particular value are described by the normal curve, also known as the Gaussian distribution.



$\frac{x}{y} = \frac{y}{z}$ gives the same result

Let $\frac{a}{b} = \frac{b}{a+b}$

then $b^2 = a^2 + ab$

$a^2 + ab - b^2 = 0$

by formula for roots of a quadratic:

$$a = \frac{-b \pm \sqrt{b^2 + 4b^2}}{2}$$

taking the positive root

$$a = \frac{(\sqrt{5} - 1)\,b}{2}$$

$a = 0.618 \,..\, b$ whence $b/a = 1.618 \,.. = K$

Fig. 3. Derivation of the magic number K.

$K=(\sqrt{5}+1)/2=1.618$. Clearly if a number is greater than one, its reciprocal is less than one, and vice versa. As it happens, K is the (only) number which differs from its reciprocal by exactly unity, so $K-1=1/K=0.618$ – a number I shall call K'.

Like e, K crops up all over the place. As **Fig. 3** shows, it describes the relative dimensions of a sheet of paper where the ratio of the short side to the long side is the same as the ratio of the long side to the sum of the long and short sides. I have a sneaking feeling this is called the *golden ratio*, but can't find it mentioned in any of my maths textbooks or encyclopaedias. It is said to be the most aesthetically pleasing ratio for a sheet of paper, being in fact only slightly squarer than the long and lanky-looking foolscap. A4 differs in the other direction, being squarer than K:1. In fact it is $\sqrt{2}$:1, so that on halving it to A5, the ratio is still the same.

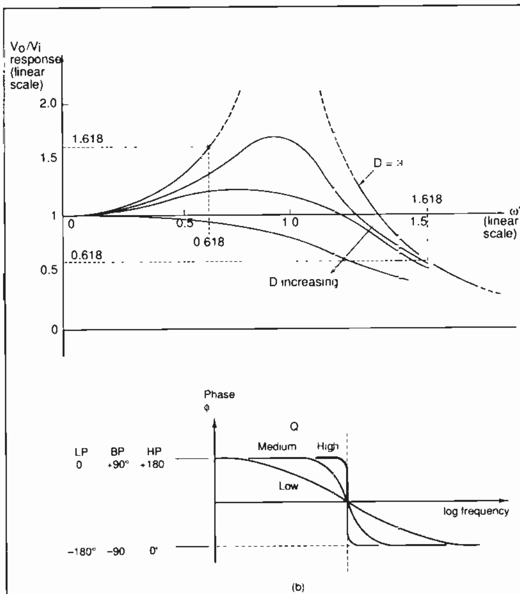


Fig. 4a) Amplitude response of a second order lowpass filter b) Phase response of a second order lowpass filter.

What I hadn't realised until a few years ago is that K and K' crop up in electronics – particularly in connection with filters. The equation defining the response of a second order low pass filter is:

$$\frac{V_o}{V_i}=\frac{1}{s^2+Ds+1}\quad\text{generally, or}$$

$$=\frac{1}{(j\omega)^2+j\omega D+1}\quad\text{in the steady state} \quad \ldots 3$$

where ω equals $2\pi f$, and f is the frequency in Hz. For convenience, make ω' the normalised frequency, i.e. the actual frequency divided by the filter's cut-off frequency. Thus at half the cut-off frequency, ω' = 0.5 and etc, keeping the sums simple.

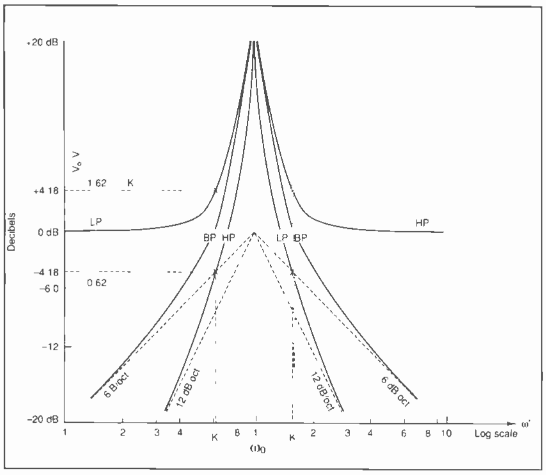


Fig. 5. Amplitude response (on a logarithmic scale) versus frequency (on a logarithmic scale) for low-band- and highpass filters, showing how the gain is K', 1 or K at frequency K' or K, according to type. Since K x K'=1, these three points on the logarithmic frequency axis are equally spaced. K and K' on the amplitude scale correspond to ±4.179dB. Note that the bandpass curve always lies between the other two.

D represents the damping term, which determines how high the peak at the upper end of the passband is, relative to the response at 0Hz, before the response falls away into the stop band. If D=0, corresponding to a Q of infinity since Q=1/D, then the peak reaches infinite proportions. **Fig. 4a** shows the response of a second order lowpass filter for various values of Q up to infinity. With the response at ω'=1 being infinite, one might expect that it would still be very large, even an octave above or below this frequency.

In fact this is not the case, even at ω'=K or K', distinctly less than an octave away. As Fig. 4a shows, the response at a frequency K is K' and at K' is K, as you may verify for yourself (or see from the accompanying text panel) by substituting the appropriate values of ω in equation 3, with D equal to zero. In fact, you will find that the response at K' is $+K$ and that at K is $-K'$, indicating no phase shift in the former case and 180° in the latter. For as **Fig. 4b** shows, when Q=∞ there is no phase shift anywhere in the passband and the stopband phaseshift is 180° at all points

K and its stablemate K' show numerous relationships, which can be simply verified by algebra by substituting $(\sqrt{5}\pm1)/2$ as appropriate. Note the following relationships where $K=(\sqrt{5}+1)/2$ and $K'=(\sqrt{5}-1)/2$:

$1/K=K'$; $1/K'=K$; $K'+1=K$; $(K+1)/K=K$; $(1-K')/K'=K'$

beyond cutoff frequency. Amplitudes K and K' are sort of 'point asymptotes' or pegs in the ground. As the Q rises towards infinity the response at ω'=K' and K approaches these points, but can never exceed them.

The response of a second order bandpass or highpass filter is the same as in equation 3, but with s or s^2 respectively replacing 1 in the numerator of the righthand side. You might expect K and K' to be involved here too,

For a second order lowpass filter

$$\frac{V_o}{V_i}=\frac{1}{(j\omega')^2+j\omega' D+1} \text{ and if } D=0\text{, then at } \omega'=\frac{\sqrt{5}-1}{2}=K',$$

$$\frac{V_o}{V_i}=\frac{1}{\left(j\frac{\sqrt{5}-1}{2}\right)^2+1}=\frac{4}{-\left(5+1-2\sqrt{5}\right)+4}=\frac{2}{\sqrt{5}-1}=\left(\frac{\sqrt{5}-1}{2}\right)^{-1}$$

$$=\frac{1}{K'}=K$$

For the second order bandpass case,

$$\frac{V_o}{V_i}=\frac{j\omega'}{(j\omega')^2+j\omega' D+1} \text{ and if } D=0\text{, then at } \omega'=\frac{\sqrt{5}+1}{2}=K,$$

$$\frac{V_o}{V_i}=\frac{\frac{j\left(\sqrt{5}+1\right)}{2}}{-\left(\frac{\sqrt{5}+1}{2}\right)^2+1}=\frac{\frac{j\left(\sqrt{5}+1\right)}{2}}{\frac{-\left(5+1+2\sqrt{5}\right)}{4}+1}$$

$$=\frac{j\left(\sqrt{5}+1\right)}{-\left(1+\sqrt{5}\right)}= \quad -j= \quad 1\angle-90°$$

For the second highpass case,

$$\frac{V_o}{V_i}=\frac{(j\omega')^2}{(j\omega')^2+j\omega' D+1} \text{ and if } D=0\text{, then at } \omega'=\frac{\sqrt{5}-1}{2}=K',$$

$$\frac{V_o}{V_i}=\frac{-\left(\frac{\sqrt{5}-1}{2}\right)^2}{-\left(\frac{\sqrt{5}-1}{2}\right)^2+1}=\frac{\frac{6-2\sqrt{5}}{4}}{\frac{6-2\sqrt{5}}{4}-\frac{4}{4}}=\frac{\frac{3-\sqrt{5}}{2}}{\frac{1-\sqrt{5}}{2}}=\frac{1+\frac{1-\sqrt{5}}{2}}{\frac{1-\sqrt{5}}{2}}$$

$$=\frac{1-K'}{-K'}=\frac{1-\frac{1}{K}}{-\frac{1}{K}}=\frac{K-1}{-1}=-K'=K'\angle+180°$$

and you wouldn't be wrong. In the case of the infinite Q bandpass filter, K' and K are the frequencies where the response is j and –j respectively, i.e. the amplitude response is unity, the phase being 90° leading at K' and lagging at K. In the case of the infinite Q highpass filter, the response at K' is $-K'$ (–4.18dB and leading by 180°), and at K is K. This is shown in **Figure 5**, along with both the lowpass and bandpass results. The curves are not exactly to scale but are the right general shape. They are displayed on logarithmic axes, which permits the display of the 6 and 12dB/octave asymptotes as well. ■

Optical storage is now at the stage where a single $5^{1}/_{4}$in disk can hold up to 1.5Gbyte – whether write-once or rewritable. Martin Eccles outlines three technologies currently in widespread use.

the optical TECHNOLOGY DRIVE

Development work on optical storage started in the early sixties. But due to continual advances in hard disk drives, light-technology drives were slow to appear on the market.

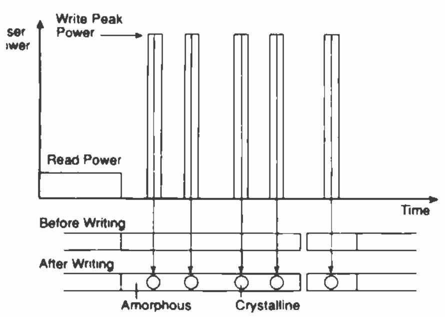
Over the past few years, write-once optical disks – worms – have become well established for mass-storage archiving. More recently, rewritable optical disk drives have evolved into serious alternatives to hard drives, mainly due to advances in their data throughput. In burst mode, some rewritable drives can achieve figures of 5M-byte/s.

Write-once essentials*

Worm drives involve a disk containing an active layer embedded between two transparent substrates – typically made of polycarbonate or glass. This active layer can have its optical characteristics permanently changed by applying a high powered laser beam.

As its active layer, a typical disk has a material which is amorphous in its unrecorded state, i.e. automatically unstructured. By locally heating this material with a high-powered laser beam, it undergoes rapid crystallisation, forming a crystalline, automatically ordered, spot in the amorphous material – a phenomenon known as phase change.

When reading the disk, because the crystalline spots have a higher reflectivity than the amorphous areas of the active layer, data is easily detected by the variation in intensity of a reflected low-power laser beam.



Unrecorded, the active layer in worm media is amorphous and absorbs light. Recording involves changing the phase of a data area by crystallising it using heat from a laser. This results in light absorbing and reflecting areas: each represents a logic level.

Phase-change technology

Phase change erasable media are similar to phase change worm types. The major difference is in the active layer material. It has the property of reversible phase change, and is originally formed with a crystalline structure, resulting in high reflectivity.

During writing to the disk, a very high powered laser beam is used to locally heat the active layer wherever a data bit is to be recorded. This momentarily melts the crystalline structure, which rapidly cools to form an amorphous, lower reflectivity, spot.

As with worm technology, reading is accomplished by detecting the difference in reflectivity between the crystalline and amorphous spots on the disk. For this job, a low powered laser beam is used.

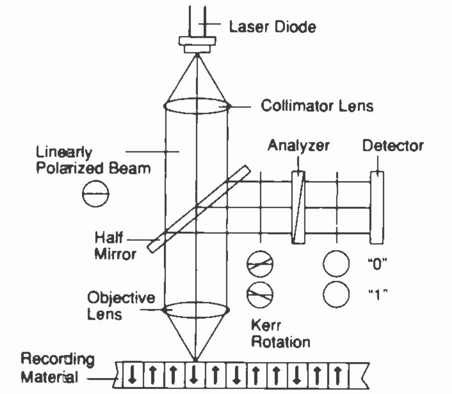
With phase-change technology, rewriting the disk involves directly overwriting data. By reheating an amorphous spot on the disk with a high powered laser beam, below the melting point of the active layer, recrystallisation occurs. The spot reverts back to its original crystalline structure with high reflectivity. This means that laser beam temperature alone can be used to change the active layer to either crystalline or amorphous state, according to the data to be recorded, in a single pass.

Magneto-optical disks

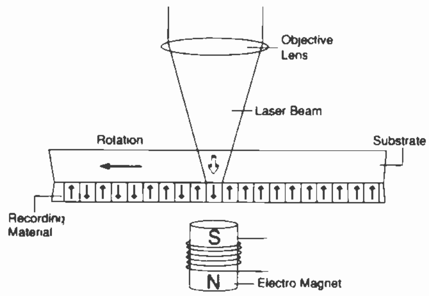
As the name suggests, magneto-optical technology is based on a combination of magnetic and optical effects. This technology has the benefit that there are standards based on it, but it also has the drawback that erasing is inherently slower than for phase-change systems.

Construction of the disk is similar to other optical disks, but the active layer is formed of a magnetic material. In its initial state this material is uniformly magnetised. Each magnetic domain is aligned perpendicular to the plane of the disk and with the same polarity.

Unlike conventional magnetic materials, the magnetisation of this material is not easily altered at room temperature. Beyond the Curie temperature however, polarity of a magnetic domain can be easily changed by an external magnetic field. Wherever a data bit is to be recorded, a high powered laser beam is used to heat the material above this temperature, allowing this reversal to take place.



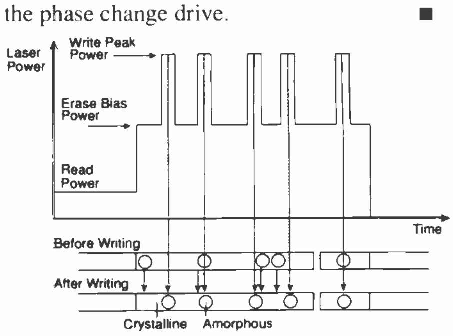
In a magneto-optical rewritable disk, polarity of the magnetic bit region can only be changed while the area is heated by a laser.



To read the disk, the drive needs to determine which domains have been 'flipped' using a plane polarised low-power laser beam. The reflected beam has its plane of polarisation rotated clockwise or anticlockwise, depending on the domain polarity – a phenomenon known as the Kerr effect. This rotation is detected and interpreted by the drive.

Before data can be overwritten, the relevant area of the active layer must be returned to its initial state of magnetisation, i.e. erased. This needs a separate rotation of the disk, during which all magnetic domains are heated and the magnetic field is applied in the reverse direction.

After erasing the old data, new data can be written in the normal way. This means that overwriting data is a two pass process, which results in a slower write throughput than with the phase change drive. ■



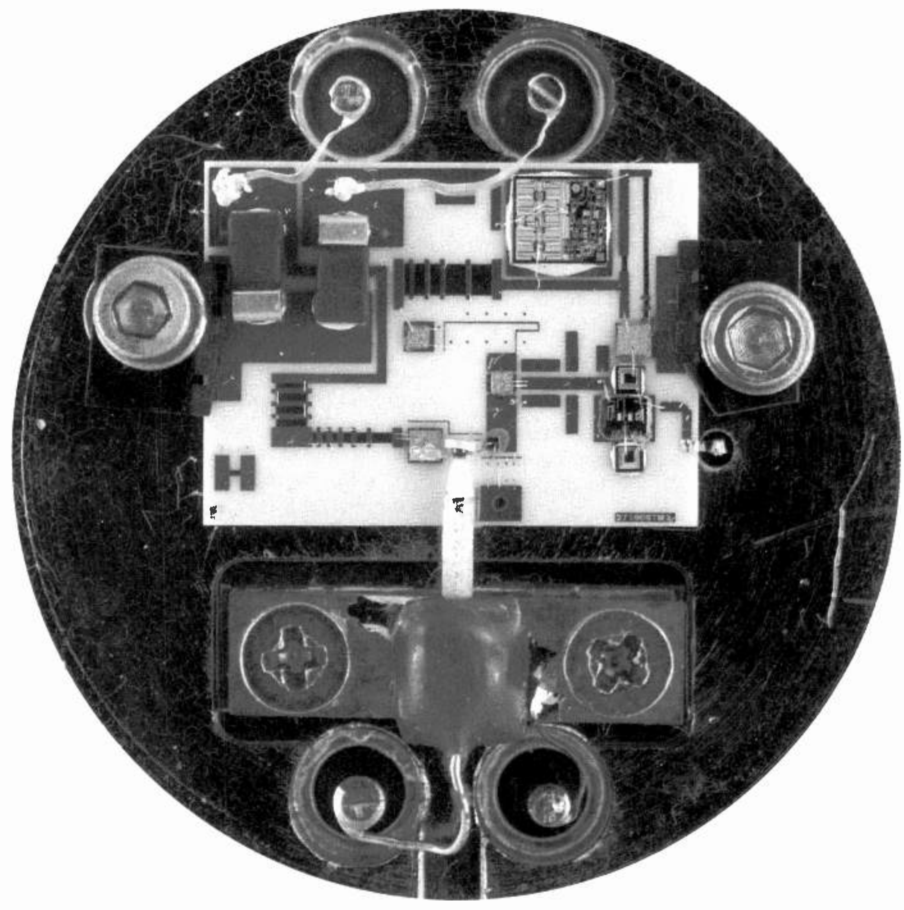
Similar to worm disks, phase-change rewritable types have amorphous and crystalline areas but the crystalline areas formed by melting can be reversed by heating to just below their melting point.

*Artwork used in this article was derived from drawings supplied by Matsushita/Panasonic

NEW WAVE MICROWAVES

6: oscillator frequency control

Tuning systems for microwaves frequently rely on ferrites and ceramics for frequency control, techniques not available on lower frequencies. Mike Hosking examines the fascinating world of solid state microwave tuning.



Any device capable of amplifying at the frequency of interest may be used as a fundamental oscillator. It will require control of frequency and spectral output.

There are, of course, other solid state methods of generating microwave signals, but these tend to be indirect methods and are only mentioned in passing. For instance, direct digital synthesis with subsequent frequency multiplication. This technique integrates digitally generated increments of phase, each of which are given an amplitude weighting corresponding to a sine (or cosine) function. After D to A conversion and low pass filtering, the final output is an analogue sine wave. Integration is approximated by an accumulator and frequencies of 100's of MHz are possible with ECL or GaAs logic elements. An advantage of this technique is the small frequency resolution obtainable ($<10^{-3}$Hz) together with fast frequency switching. However, spurious outputs generated by sampling errors can be high. Phase noise mirrors the sampling clock.

A second, indirect technique is that of frequency synthesis using high-speed dividers and a phase-lock loop. **Fig. 1** shows a typical

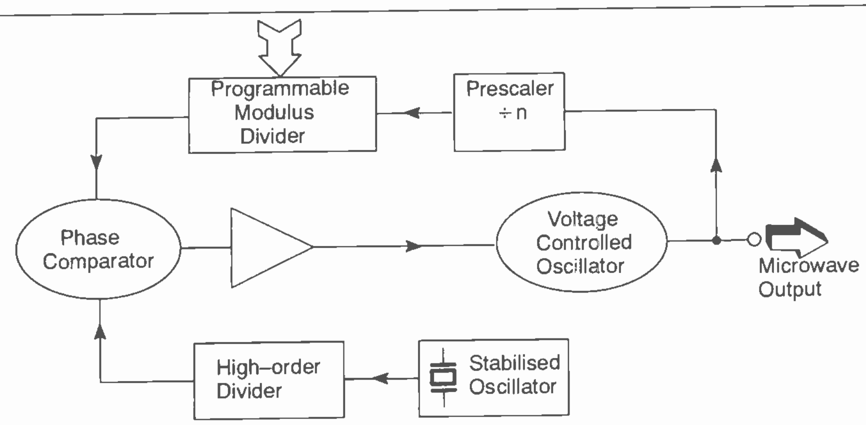


Fig. 1. A practical form of microwave frequency synthesiser uses a fundamental VCO in a phase locked loop. The loop incorporates high frequency dividers, one having a programmable modulus.

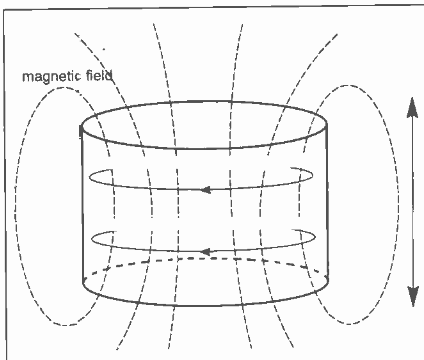


Fig. 3. Electric and magnetic fields external to the disc allow coupling to the microwave circuit.

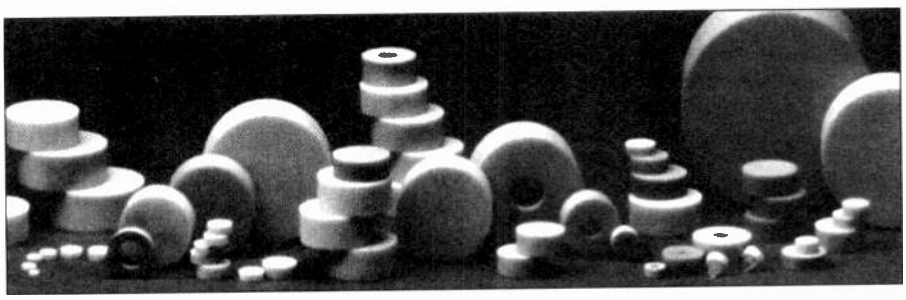


Fig. 2. Assorted sizes and shapes of ceramic discs suitable for high Q dielectric resonators. Diameter ranges from a few millimetres to several centimetres. An optimum diameter to height ratio would be 2:1.

synthesizer circuit. A VCO of several GHz output frequency is possible and is controlled by the error voltage generated from the phase comparator. Noise output is dependent upon the VCO design and the PLL noise, with the output frequency being quantised by the divided reference oscillator, i.e. 8 kHz steps for a 2^{10} divider and 8.192 MHz reference.

Dielectric resonator

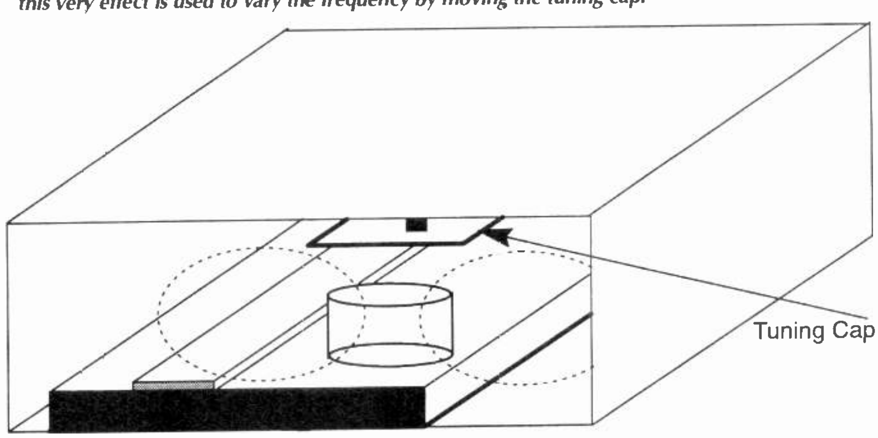
A dielectric resonator is a high-Q circuit element. The resonator itself consists of a small disc of ceramic material, having a high dielectric constant, typically lying between 30 and 40GHz although operation to 100GHz is possible. By bringing this element physically close to the electromagnetic fields surrounding a microwave oscillator circuit, coupling of the fields to the resonator will occur; resulting in frequency locking of the oscillator. A range of such discs is shown in **Fig. 2**. Barium titanate and zirconium titanate are most commonly used as the basic materials with various additives to control the frequency/temperature coefficient.

A dielectric resonator system works like this. A length of microwave transmission line, short circuited or open circuited at both ends, appears as a resonant circuit having a fundamental and well-defined resonant frequency. Furthermore, the resonant circuit also supports higher-order modes, providing they conform to the particular boundary conditions of the circuit. In this respect, the disc (cylinder) of the dielectric resonator behaves in similar fashion to a short-circuited length of air-filled, circular, metal waveguide. In such a resonator, the fields are internally reflected by the metallic walls and the physically realisable modes must conform to the boundary conditions where there are no tangential component of E-fields and no normal component of H-fields at the conductor surfaces. In an analogous fashion, the dielectric resonator can be modelled as a length of cylindrical waveguide having magnetic walls which behave as magnetic conductors, i.e. magnetic short circuits. However, an accurate analysis, leading to prediction of the resonant frequency, is more complex because the dielectric boundaries are not perfect conductors. The E and H fields radiate beyond the boundaries of the disc and must be accounted for, but it is this characteristic which allows coupling of the resonator to an external circuit. **Figure 3** shows the electromagnetic field configuration for the transverse electric fundamental mode, normally used in circuit design.

Due to the low dielectric loss tangent of the modern ceramics, the Q-factor of the disc resonator can be exceptionally high; Q's of 10,000 at 25GHz and 24,000 at 10GHz are routinely available. In terms of frequency coverage, disc resonators become too large below about 1GHz and, above about 40GHz, are prone to the generation of unwanted modes. In addition, the resonator Q-factor decreases with frequency. Thus, with present materials, the general range of application is from about 1.5GHz to 40GHz for a simple disc. Higher frequency performance can be achieved with changes to the simple geometry, which generate a circumferential "whispering gallery" mode and is currently pushing performance through the millimetre wavelengths.

The dielectric resonator is particularly well suited to the microstrip type of circuit where

Fig. 4. A disc resonator coupled to a microstrip line within an enclosed box. The walls of the box, together with the proximity of the circuit affect the resonator field pattern and must be allowed for. However, this very effect is used to vary the frequency by moving the tuning cap.



Chips! Wrapped or unwrapped?

With the application of reverse bias, the depletion region widens and the junction capacitance of the varactor will decrease from its maximum value at zero bias $C_{j(0)}$ to a value set by the tuning voltage $C_{j(V)}$. The maximum value of V is determined by the diode breakdown voltage. If C_j were to constitute the capacitive element of a tuned circuit, then the corresponding change in resonant frequency would be given by

$$\frac{f_{max}}{f_{min}} = \left(\frac{C_{j(0)}}{C_{j(V)}}\right)^{\frac{1}{2}}$$

Hence, the ratio

$$\frac{C_{j(0)}}{C_{j(V)}}$$

becomes a specification parameter of the varactor as it indicates the available tuning range. Other factors, such as the actual frequency of operation and required Q-factor, determine the absolute value of $C_{j(V)}$ to be selected. The actual way in which C_j itself changes with bias voltage is given by

$$C_{j(V)} = \frac{C_{j(0)}}{\left(1+\frac{V}{\phi}\right)^{\gamma}}$$

where ϕ is the "built-in", or contact potential, of the diode typically lying between 0.8 and 1.3 volts depending upon construction and material. The exponent γ depends upon the doping profile of the p-n junction. Linear tuning occurs when γ=2. In practice, C_j is not the only capacitance present: circuit strays also add to the total, so that the circuit itself contributes to the value of γ.

Packages add significant parasitic reactances which affect and sometimes limit device performance. This is particularly true for the varactor and can be illustrated by the following example.

A diode suitable for, say, 10GHz operation might have $C_{j(0)}$ of 0.6 pF for the chip alone and 0.1 pF at a reverse bias of –30 V. The capacitance ratio is thus 6:1 producing a theoretical maximum frequency change is about 2.4:1. A chip package would normally add 0.18 pF. In addition, the packaged varactor must still be embedded within the microwave circuit and thus there will be fringing capacitances to the mounting structure, perhaps 0.05 pF. The total capacitance values now becomes 0.83 pF max and 0.33 pF min, a ratio of 2.5:1. This results in a tuning range of 1.6:1. So, even though these parasitic elements are small, their effect at microwave frequencies can be drastic.

tight coupling can readily be achieved and the disc may be bonded or bolted directly to the substrate. As shown in **Fig. 4**, though, the more practical design situation must include the effects of the microstrip substrate and surrounding packaging on the resonant frequency.

In fact, these effects may be used to advantage to allow mechanical frequency tuning over a small range. Adjusting the metal plate closer to the disc will increase its resonant frequency and raising the disc above the substrate surface with a plastic spacer will lower the frequency.

When designing, say, a fet oscillator in microstrip, various configurations are possible and involve positioning the dielectric resonator so that it acts as a feedback element. Essentially, at resonance, the disc appears as a band-stop filter having a high reflection coefficient. A typical series feedback circuit is shown in **Fig. 5a**. With the resonator coupled to the matched transmission line (say 50Ω) of the gate, a maximum amount of the oscillator output power will be reflected back at the resonant frequency of the disc. The phase of the reflected power with respect to the output can be optimised by adjusting the distance between resonator and fet (approximately a half-wavelength) in order to achieve injection locking at the disc frequency.

A push-pull oscillator circuit using two fets and a single disc is shown in **Fig. 5b**. This arrangement has the advantage that the fet noise sources can be added in anti-phase, thereby improving the FM noise performance.

Frequency stability with temperature proved an early problem for DROs. Not only do the physical dimensions of the disc change with temperature but so, too, does the actual dielectric constant. However, these problems have been overcome by the use of various additives to the disc material. It is now possible to select material with a specified temperature coefficient, typically lying between –4 and +10ppm/°C and to use this to compensate for temperature effects elsewhere in the circuit for near-perfect overall stability.

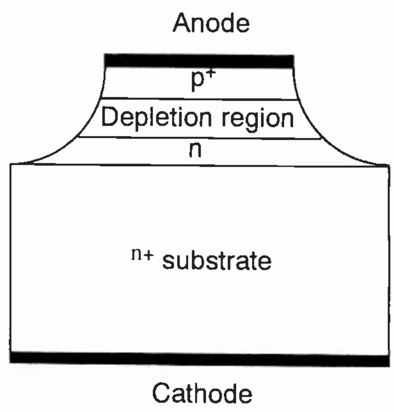
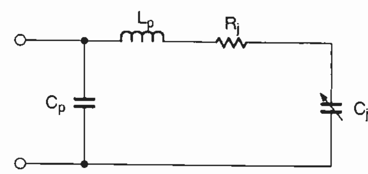


Fig. 6. Varactor mesa doping profile and the approximate equivalent circuit of the packaged chip. Cp and Lp are parasitics while Rj and Cj are the junction resistance and capacitance.



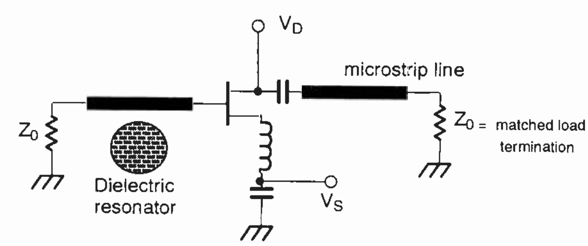
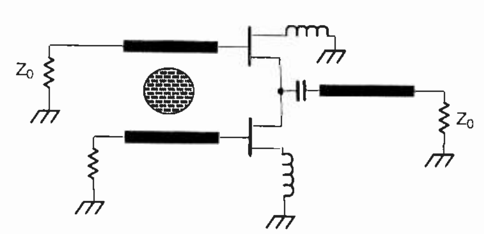


Fig. 5. Schematics of two possible microwave fet oscillator circuits.

a) the dielectric resonator in a series feedback configuration
b) push-pull arrangement which reduces fet noise.



Frequency tuning

Microwave VCOs lie at the heart of receiver systems and RF instrumentation, often over a wide frequency range. Two devices typically perform this function: the variable capacitance diode (usually referred to as the varicap at low frequencies and the varactor at microwave) and the YIG sphere which is made from ferrite material.

The varactor diode is a p-n junction, made from either GaAs or Si, of the form shown in **Fig. 6a**. The doping profile of the junction results in an equivalent circuit **6b**, which also includes the parasitic elements of a typical package. The varactor operates under conditions of reverse bias to appear as a small series resistance (about 0.5Ω) and a variable junction capacitance C_j. This capacitance is,

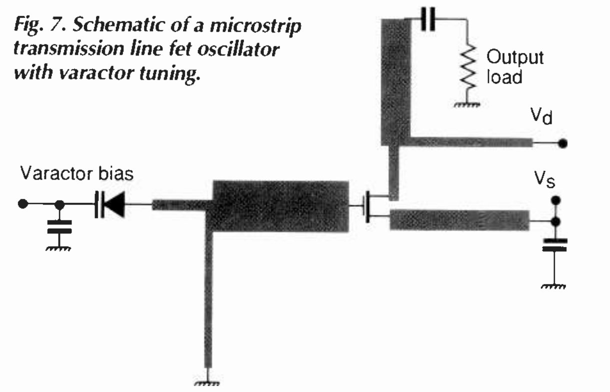


Fig. 7. Schematic of a microstrip transmission line fet oscillator with varactor tuning.

in fact, shunted by the junction resistance but, as this is very large (10MΩ) under reverse bias, it has little effect. When designing an oscillator, if C_j is made to be all or part of the tuned circuit, then changing the reverse bias voltage will change the resonant frequency of operation.

Most oscillator circuits may be frequency tuned by varactor, although the physical realisation of the resonant circuits will vary considerably. In waveguide Gunn or impatt oscillators, the varactor would be port mounted in similar fashion to the active device. **Figure 7** shows a schematic of a microstrip mesfet oscillator which uses capacitive feedback from the source to the gate, with the varactor included in series with the gate. If the oscillator was centred at, say, 10GHz, then it would be possible to tune over the full X-band frequency range (8-12GHz).

By changing the doping profile of the p-n junction, it is possible to vary the value of the exponent, γ, in the tuning equation to create two categories of microwave varactor. When γ=0.5 it results in the abrupt junction device and a value of γ=1.2 to 1.5 produces the hyperabrupt varactor. The latter device offers a wider tuning capability and, with a combination of circuit design, can give a highly linear characteristic between tuning voltage and frequency.

Varactor tuned oscillators allow fast frequency tuning; slew rates of 10GHz per μs are possible. This allows high data rate FM modulation or for fast, signal intercept receivers for frequency lock loops. Voltage breakdown limits the amount of change possible in C_j, reducing maximum frequency change to one octave. **Figure 8** shows typical tuning curves for both abrupt and hyperabrupt diodes.

Other important considerations include settling time and post-tuning drift of an oscillator. Drift is usually of more concern in free-running CW oscillators where thermal effects at the varactor p-n junction may cause problems.

YIG tuning

Ferrite materials are used to make a number of non-reciprocal microwave components such as isolators, circulators, gyrators. In addition, it is also possible to design very small resonators having a high Q-factor with this material and these have widespread applications as the tuning element in wideband oscillators and filters.

The material itself is a polycrystalline ceramic made from sintered oxides and having ferrimagnetic properties. Most commonly, yttrium and iron oxides are used to form a garnet material (hence YIG), sometimes with a doping of GaAs.

In a ferrite, not all of the electrons are paired with opposite spins (as in a non-magnetic material) but instead, there is a surplus of un-paired electrons, resulting in a net magnetic moment and a small magnetisation. The properties of a ferrite (particularly its permeability) can be influenced by the strength and direction of an applied dc magnetic field. Simultaneous interaction with an alternating microwave field allows a variety of components possessing non-reciprocal properties to be designed. The application here uses interaction between the microwave field, a dc field and the spinning electrons.

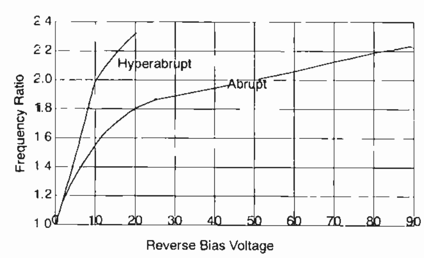


Fig. 8. Tuning curves of theoretical frequency ratio for a hyperabrupt and abrupt junction varactor

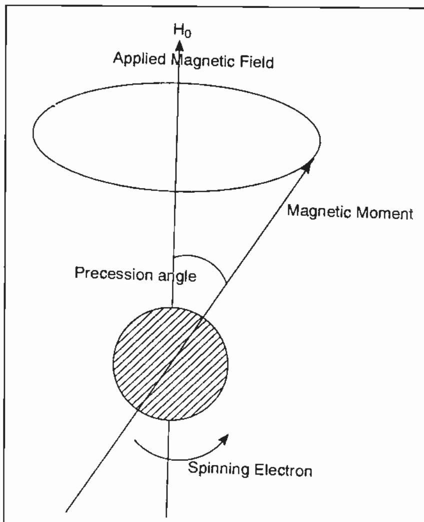


Fig. 9. The spinning YIG electrons create a magnetic moment which, under the influence of a steady external magnetic field, precess about the field axis.

Figure 9 shows the effect of applying a steady, external magnetic field to the YIG material. Initially, the un-paired and spinning electrons are lined up to produce an overall magnetic moment.

If a dc magnetic field, H_0 is applied to the ferrite, the magnetic moment and will start to precess around this axis at a precession angle proportional to H_0. The effect is analogous to the precession of a spinning gyroscope when the spin axis is not aligned with the gravitational force.

Two important physical relationships arise from this situation: firstly, the ratio of the magnetic moment of the ferrite electron to its angular momentum is called the gyromagnetic ratio, γ, and has the value 0.035MHz A^{-1} m (or, more colloquially, 2.8MHz per Oersted).

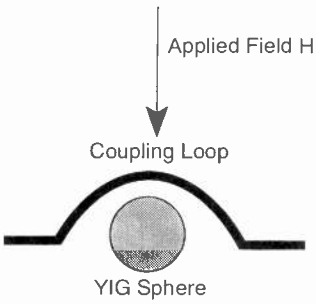
Secondly, the precession frequency (also called the Larmor frequency) is related to the magnetic field by $f=\gamma H$.

To produce a resonator, the YIG material is formed into a small, highly polished and accurately dimensioned sphere; diameters range typically from 0.2 to 2mm. Such a sphere is naturally resonant at the precession frequency and, most importantly, possesses a very high unloaded Q-factor (2000 or more). If a microwave field is coupled into the YIG sphere, then there will be a strong interaction when resonant frequency and microwave frequency are equal. Such coupling normally takes the form of an inductive loop in close proximity to the sphere, as shown in **Fig. 10a** and has the overall equivalent circuit of **Fig. 10b**. If a YIG plus coupling loop is made part of a microwave oscillator circuit, the frequency may be tuned by varying the magnetic field applied to the sphere. An electromagnet, with the sphere accurately located between pole pieces, is used to produce the field. Thus frequency becomes proportional to coil current.

YIG tuned oscillators (YTO's) can provide tuning ranges of several octaves; 2-18GHz is possible. The technology finds use in electronic warfare, as well as sweep generators and spectrum analysers. Coil inductance limits speed of tuning or slew rate. It typically takes 1ms for full band coverage. However, some oscillators incorporate a small, subsidiary coil for faster (but limited) tuning in FM or FLL circuits.

The oscillator itself may use any of the devices previously discussed, although the physical realisation and capabilities of the circuit will obviously vary. Bipolar or fet devices are common arranged to provide a negative resistance applied to the YIG coupling loop. This produces oscillation at the natural frequency of the system. A complete microwave YTO circuit in microstrip form is shown in **Fig. 11**.

The YIG sphere, about 0.5mm diameter, can be seen, partially hidden by the coupling loop and is mounted at the end of a beryllium oxide rod. The rod is heated to a temperature above the normally-expected ambient level and accurately stabilised. This reduces the frequency drift with temperature, as well as allowing ini-



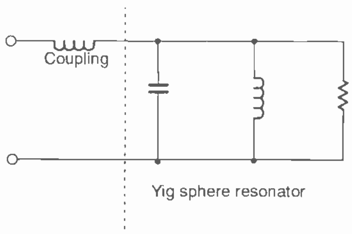


Fig. 10. Coupling to the YIG sphere resonator is effected by a wire loop resulting in the simplified equivalent circuit.

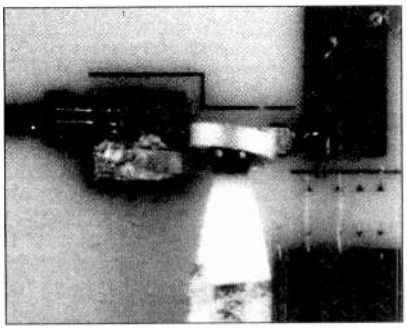


Fig. 11. An actual YIG circuit with the resonator ball, a sort of spinning electron gyroscope, mounted on a beryllia arm underneath a coupling loop connected to an oscillator circuit.

tial alignment of the sphere for optimum coupling. Beryllium oxide is used for the rod as it possesses excellent thermal conductivity while being an electrical insulator. The white substrate is alumina, about 23 x 17mm and contains a thin-film bipolar oscillator in the top right-hand corner and an output buffer amplifier in the lower RH corner. Output power is 10mW and the oscillator tunes over 2-8GHz. A sealed lid, incorporating the tuning coil completes the circuit.

Historically, the YIG oscillator has been used for wideband applications but, recently, much effort has been put into miniaturising the circuit by using permanent magnets and limited tuning The incentive has come from the mobile communications market, which make use of the high Q-factor available from YTOs (instead of the wide frequency coverage) to produce low, phase-noise sources.

Finally, an example of the oscillator devices in use is provided by the Hewlett Packard sweep generator. The *HP83592C* plug-in for the generator covers the 10MHz to 20GHz band in one unit, a span of nearly 11 octaves. **Figure 12** shows the main microwave components.

The basic signal is generated by a low phase noise, bipolar transistor oscillator YIG tuned over 2.3 to 7.0GHz. To produce the lower frequencies, the output of this YTO covers the range 3.81 to 6.2GHz and is routed to a mixer driven from a stabilised 3.8GHz local oscillator. Thus, when the YTO is at 3.81GHz, an IF of 10 MHz is generated, increasing to 2.4GHz when the input is 6.2GHz. This band is switched to the output via amplification and filtering. Higher frequencies are generated from the output of the YTO via a step recovery diode multiplier, from which either the fundamental, 2nd harmonic or 3rd harmonic can be selected. Selection is done by a tunable filter, again using a YIG element as the high-Q circuit. ■

Dielectric Disc Resonator

Because there is a very large difference between the dielectric constant of the ceramic disc (ε_{rd}) and that of the surrounding medium (ε_{ra}), the internal fields are reflected at the dielectric interfaces. For example: the ratio of reflected electric field E_r to incident field E_i at normal incidence to the interface is the voltage reflection coefficient, given by

$$\frac{E_r}{E_i} = \frac{\sqrt{\varepsilon_{rd}} - \sqrt{\varepsilon_{ra}}}{\sqrt{\varepsilon_{rd}} + \sqrt{\varepsilon_{ra}}}$$

For the case of a zirconium titanate disc having $\varepsilon_{rd} = 38$ and $\varepsilon_{ra} = 1$ for air, the reflection coefficient is 0.72. Thus, a resonator can be formed without electronically conducting boundaries; the interface appearing as an approximate open circuit to the E-field and a short circuit to the H-field.

The reason for using a dielectric disc rather than an equivalent waveguide structure is that of size. As the fields are propagating largely within the dielectric medium, their velocity and, hence, wavelength will decrease by a factor of approximately $\sqrt{\varepsilon_{rd}}$, which ≈6 in the above example. The resonant frequency of the disc itself can be calculated to within an

$$f = \frac{6.8 \times 10^2}{D\sqrt{\varepsilon_{rd}}}\left(\frac{D}{2H} + 3.45\right) Hz$$

with D in metres. A typical ratio of D/H is about 2:1. This, however, must be substantially modified in practice, due to the fact that: (a) the disc itself must be positioned on and coupled to the microwave circuit; (b) the total circuit is almost always packaged within a metal enclosure. Both affect the radiated fields from the disc.

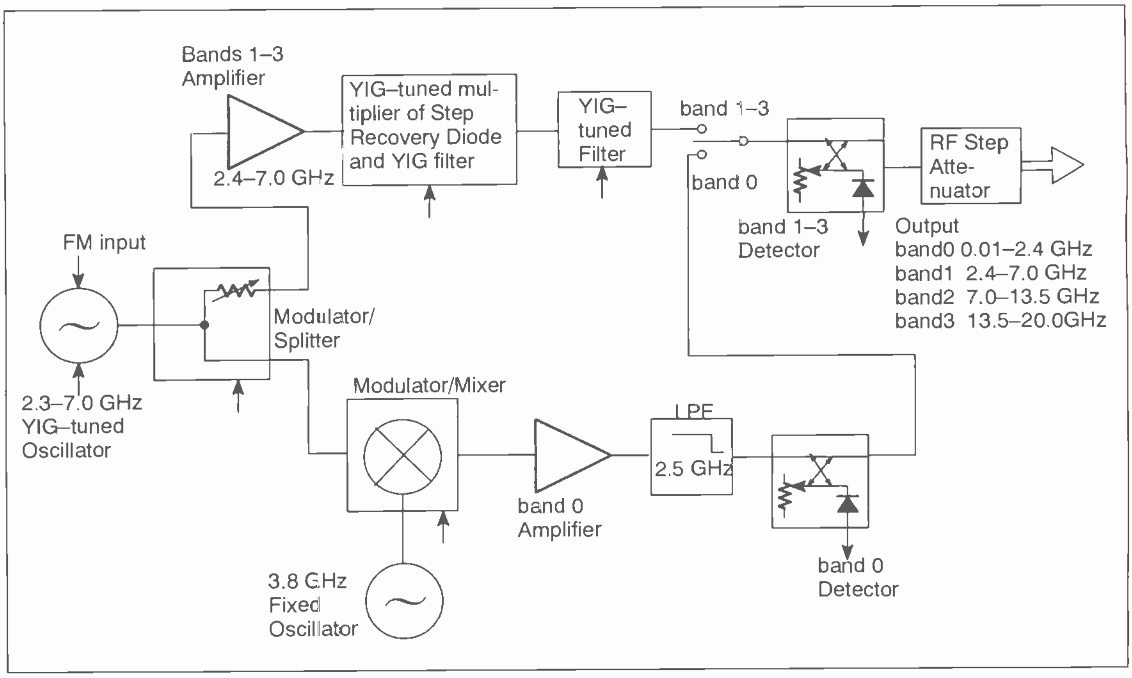
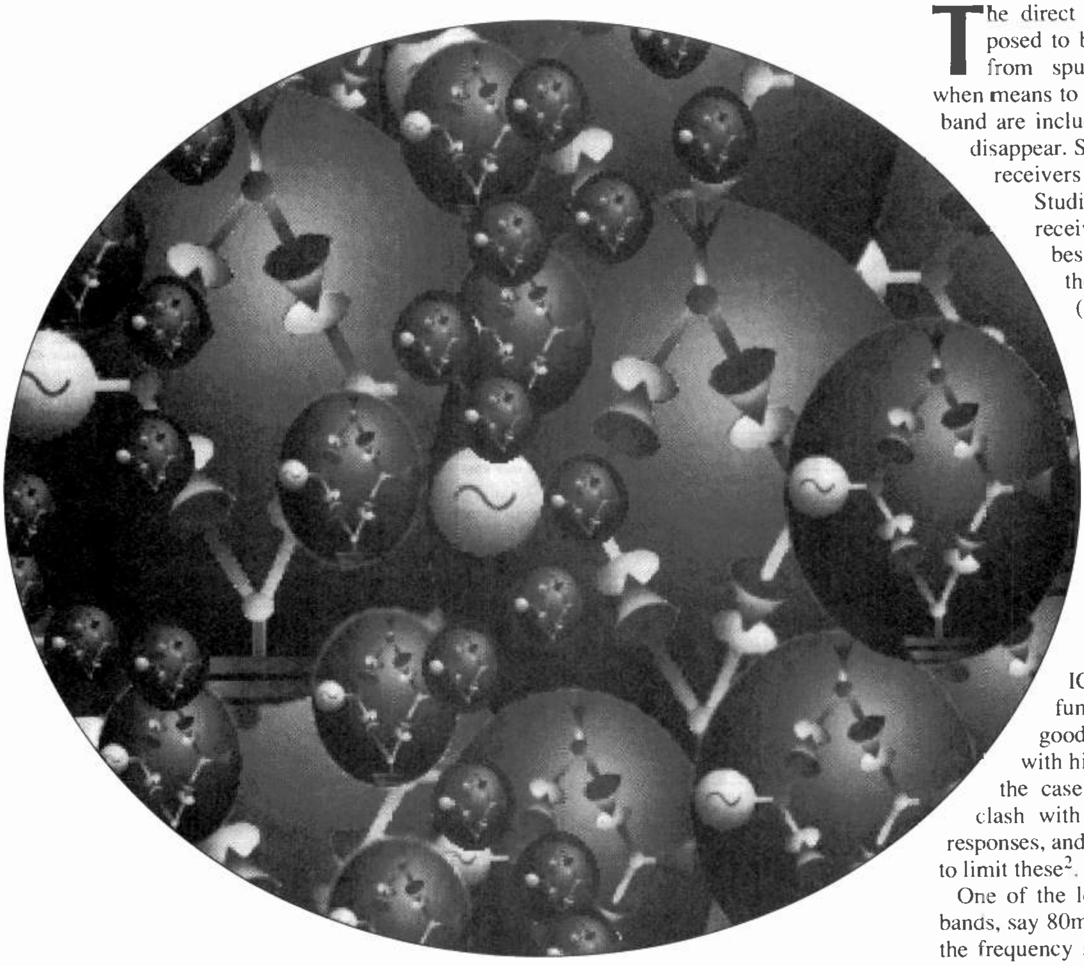


Fig. 12. This plug-in for a H-P signal generator provides an excellent example of using several technologies to produce signals over an 11-octave range.

Direct conversion ssb receiver

Direct conversion for rf reception is now widely used in integrated form because it delivers good performance without the use of mechanical filters and other expensive, bulky items. Frank Dorey has put together a development breadboard for direct conversion SSB use, the principles of which can be adopted for many other applications.



The direct conversion receiver is supposed to be simple and relatively free from spurious responses. However, when means to eliminate the unwanted sideband are included, the simplicity seems to disappear. So practical designs for dc ssb receivers are still rare.

Studies on direct conversion receivers have concluded that the best option for ssb reception is the "Weaver type" receiver (known to radio amateurs as the *third method*). But it has to be ac coupled to get over the problem of a constant tone out of the second oscillator when dc drift occurs in the first balanced modulator. While there is a "hole" in the audio response, this, if not too wide, does not seriously detract from speech intelligibility[1].

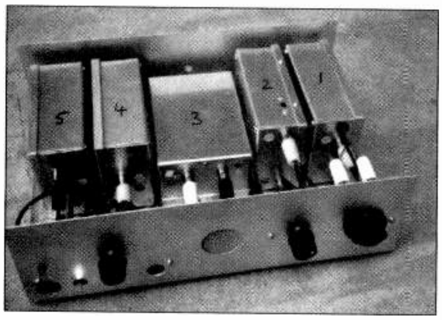
Although the use of digital ICs to realise some of the circuit functions required may seem a good idea, the use of square waves with high harmonic content tends, in the case of the Weaver receiver, to clash with the aim of no out-of-band responses, and extra work might be required to limit these[2].

One of the lower frequency HF Amateur bands, say 80m, seems to be a good place in the frequency spectrum to try out a design, using only analogue techniques so as to ensure freedom from spurious responses. As this is a design study, most of the circuitry is divided between five separate modules: rf amplifier and dual first mixers; 2-phase rf oscillator; dual 7-pole low-pass filters; dual second mixers, audio amplifier and agc circuit; two-phase second oscillator and power supply regulator.

Circuit operation

Referring to the block diagram, suppose that a transmission on a nominal carrier frequency of 3600kHz is in fact lower sideband with an audio range which would just fit this particular receiver. Then it would contain frequencies from 3599.8kHz down to 3596.4kHz. The first oscillator in the receiver should be tuned to 3598.1kHz to receive this transmission.

Sidebands of the local oscillation are produced extending for 1.7kHz on each side of it, a range of frequencies 3.4kHz wide. But what comes out of the balanced modulator consists of two sets of audio frequencies as superimposed signals, each in the range zero to 1.7kHz being the two halves of the original audio spectrum fed into the transmitter modulator. The ac coupling at the output of the first balanced modulator deliberately curtails the zero end of each band of frequencies, removing a dc component but putting a gap in the middle of any subsequent reconstitution of the original modulation frequency range. Also, one of the signals has a reversed frequency spectrum compared to the original modulation.

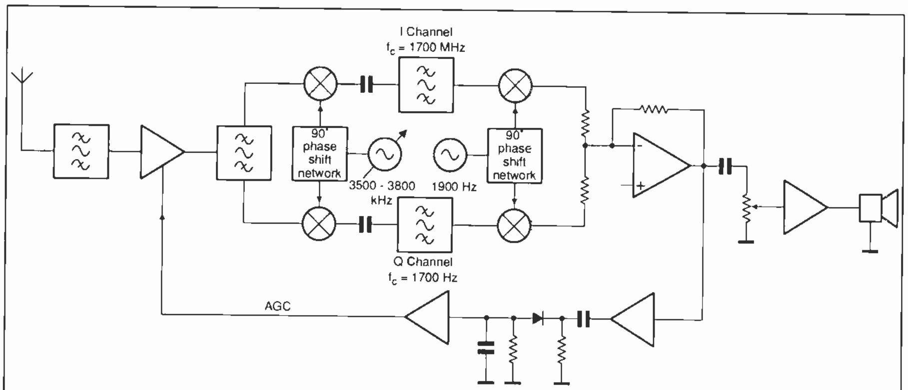


The prototype unit was built up as a series of modules wired together. This achieves a high level of screening, important even to basic direct conversion equipment.

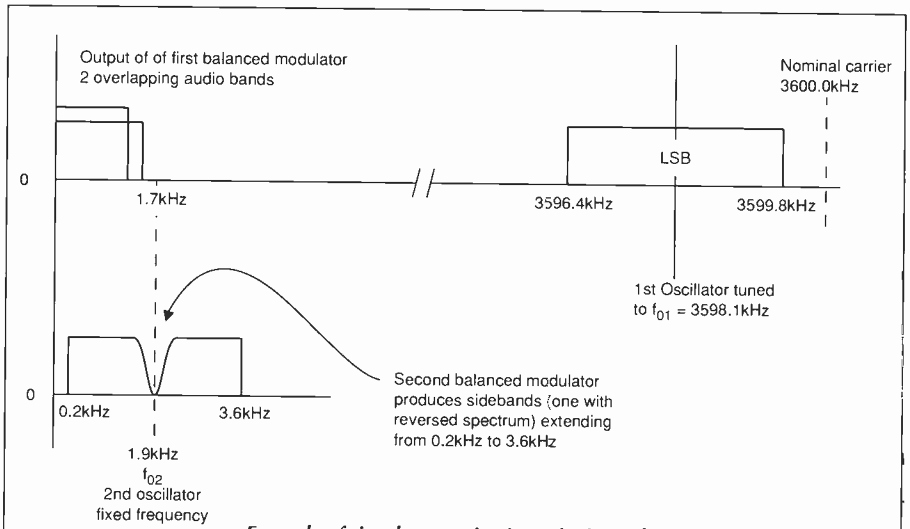
What has just been described is happening, of course, in two channels, the I (in-phase) and the Q (quadrature) channels. This permits signal processing by using another pair of balanced modulators fed from a 2-phase second oscillator. This is followed by a summing op-amp to re-invert one half of the audio spectrum and put together again the full range of modulation frequencies (except for a small gap in the middle).

The sideband range of the second oscillator, running at 1.9kHz, contains frequencies from 0.2kHz (ie. 1.9–1.7) to 3.6kHz (ie. 1.9+1.7). Correct phasing of the second oscillator inputs to the second balanced modulator channels will ensure that the output summation will reinforce wanted signal components and cancel the unwanted ones.

Note that incorrect phase choice will result in USB instead of LSB reception; also that success of the process depends on matched gains in the two channels up to the summing op-amp. For the maths to bear this all out, see the appendix to ref.1).



Receiver block diagram. Most of the circuitry is divided between five separate modules: rf amplifier and dual first mixers; 2-phase rf oscillator; dual 7-pole low-pass filters; dual second mixers, audio amplifier and agc circuit; two-phase second oscillator.



Example of signal processing in each channel prior to summation. Note the 'hole' caused by the fact that frequencies near zero from the overlapping audio bands, cannot pass through the ac coupling inherent in the circuit.

Prototype details

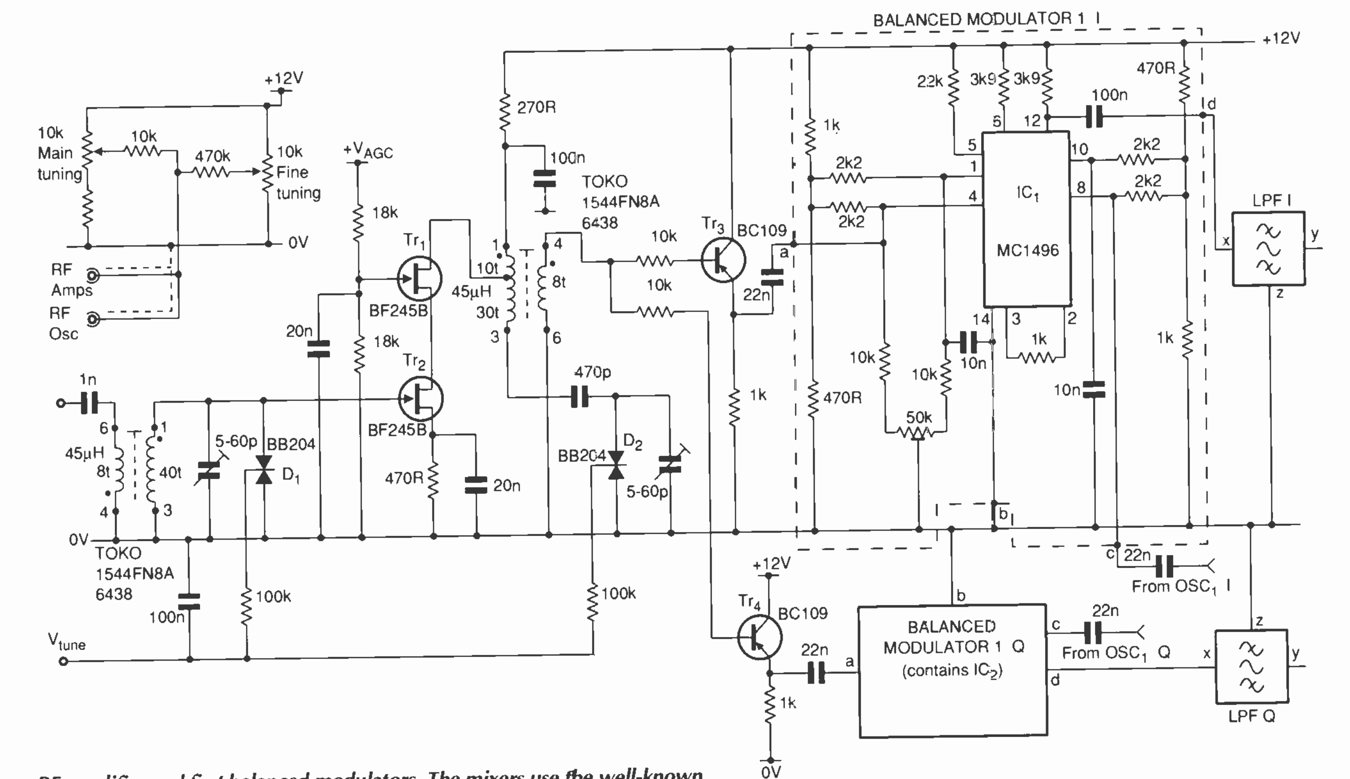
The signal interconnections between modules were made with phono connectors, colour-coded to identify channels. Each module was built on a printed circuit board contained in a small aluminium box. The five modules were housed in a case with panel mounted controls. No display of frequency is included. Since the channel to which the receiver is tuned is always (fo1+1.9)kHz, it is convenient to connect a monitoring frequency meter to an auxiliary output from the first oscillator.

No attempt was made to include a power supply inside the case, the regulator input coming from an external unstabilised supply of at least 14.5 volts dc placed some way from the receiver. The current taken from this supply was found to be 190mA.

The cascoded junction fet rf stage uses Toko coils for 3.5 to 3.8MHz coverage, and incorporates diode tuning. Source follower buffers couple it to the following balanced mixers. The mixers use the well-known *MC1496*. Substitution for the later *NE602* should reduce the component count and simplify the layout.

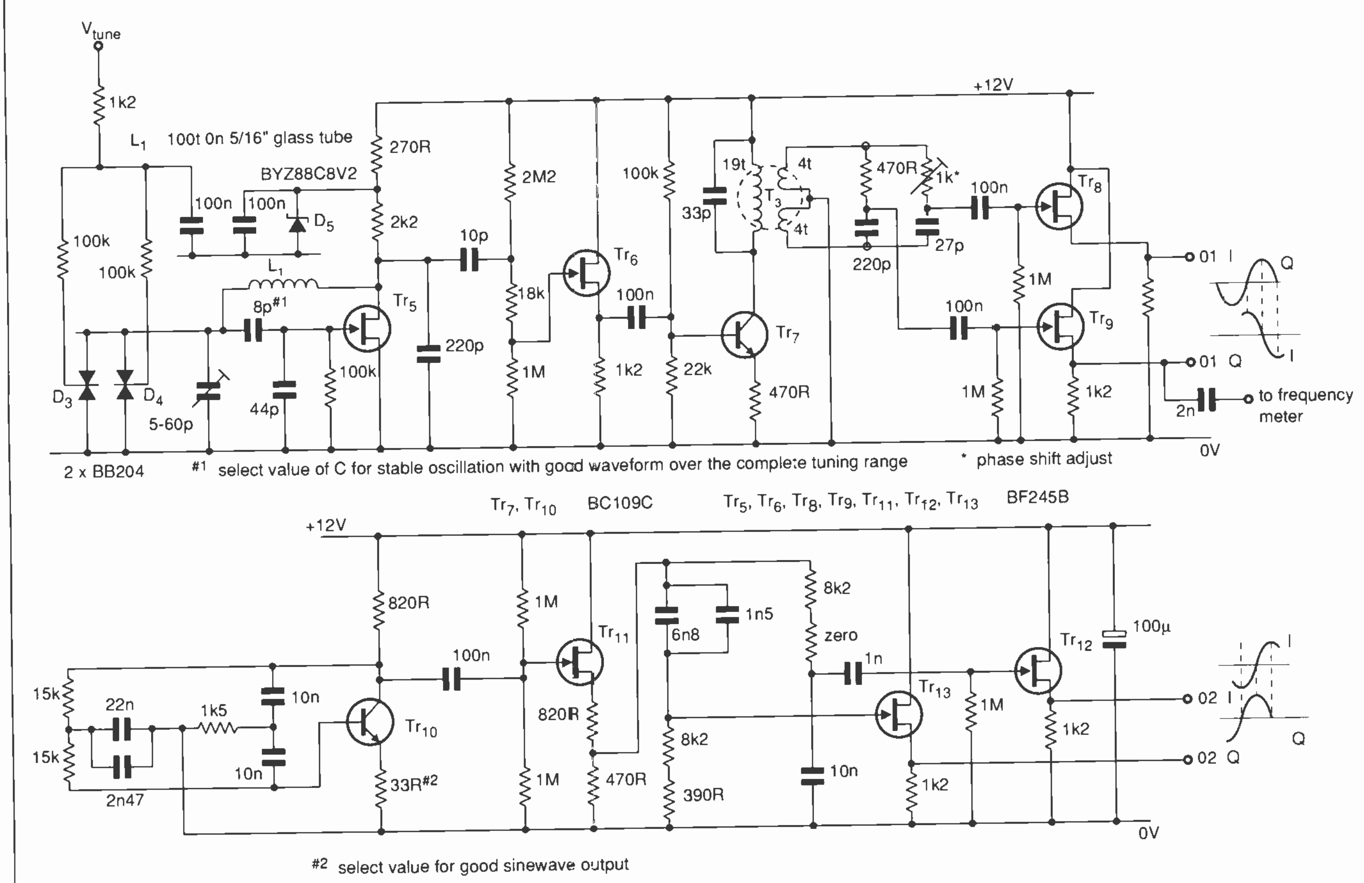
The fet Vackar first oscillator circuit is built around a coil wound on a 5/16 inch glass former with 100 turns of 0.15mm enamelled wire. The gate drive is adjusted so that oscillation is just stable over the complete frequency band, with a good clean waveform.

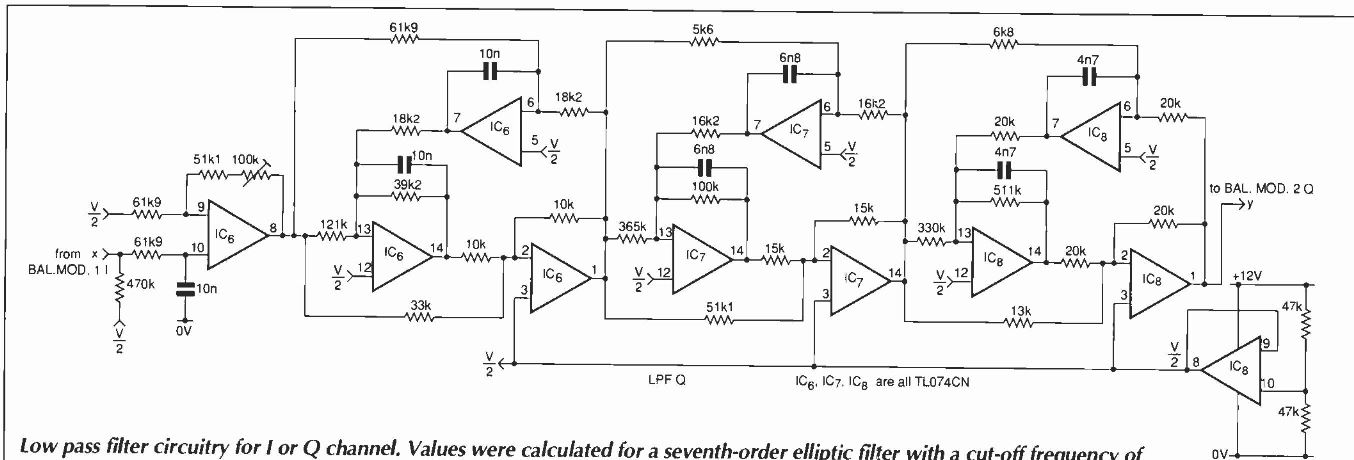
The rf phase-shift network was suggested by Ref 3. Operation was checked using an oscilloscope, having first ensured that no phase difference was indicated when the same signal was applied to both Y-channels. This frequency is too high for the bandwidth of the average X-channel so that a Lissajous figure method cannot be used unless the scope X-input can match the phase response of the Y-channel.



RF amplifier and first balanced modulators. The mixers use the well-known MC1496. Substitution for the later NE602 should reduce the component count and simplify the layout. High level parts such as the SL6440 type will improve strong signal performance but require care in matching.

First RF and second AF oscillators.





Low pass filter circuitry for I or Q channel. Values were calculated for a seventh-order elliptic filter with a cut-off frequency of 1700Hz, pass band ripple width 0.3dB, and minimum stop band loss 80dB. This last value puts it in a category where little or no trimming should be required if capacitors with a 5% tolerance are used and resistors are within 1% of the calculated value. Other channel is identical.

The low pass filter design is detailed in Ref.4. Values were calculated for a seventh-order elliptic filter with a cut-off frequency of 1700Hz, pass band ripple width 0.3dB, and minimum stop band loss 80dB. This last value puts it in a category where little or no trimming should be required if capacitors with a 5% tolerance are used and resistors are within 1% of the calculated value.

Using K=3 in each stage, including the first-order stage, gives each filter a pass band voltage gain of about 80. Adjustable gain in each first-order stage permits balancing the overall gains of the I and Q channels.

The second oscillator uses a twin-T circuit followed by a simple phasing network. The circuits were laid out so as to allow the late addition of series or parallel trimming additions to get the frequency exact and the phase difference exactly 90°, with equal amplitudes of the I and Q outputs. With a frequency in the audio range it was possible to switch the oscilloscope to its X/Y mode and trim for a near-perfectly circular Lissajous figure.

The audio amplifier is a simple unity-gain output stage following the summing amplifier and the agc circuit is fed from a gain-of-5 stage, using spare amplifiers in the *LM324* which already contained the summing amplifier at the end of the I and Q channels. The volume control and the audio output stage are fed direct from the summer.

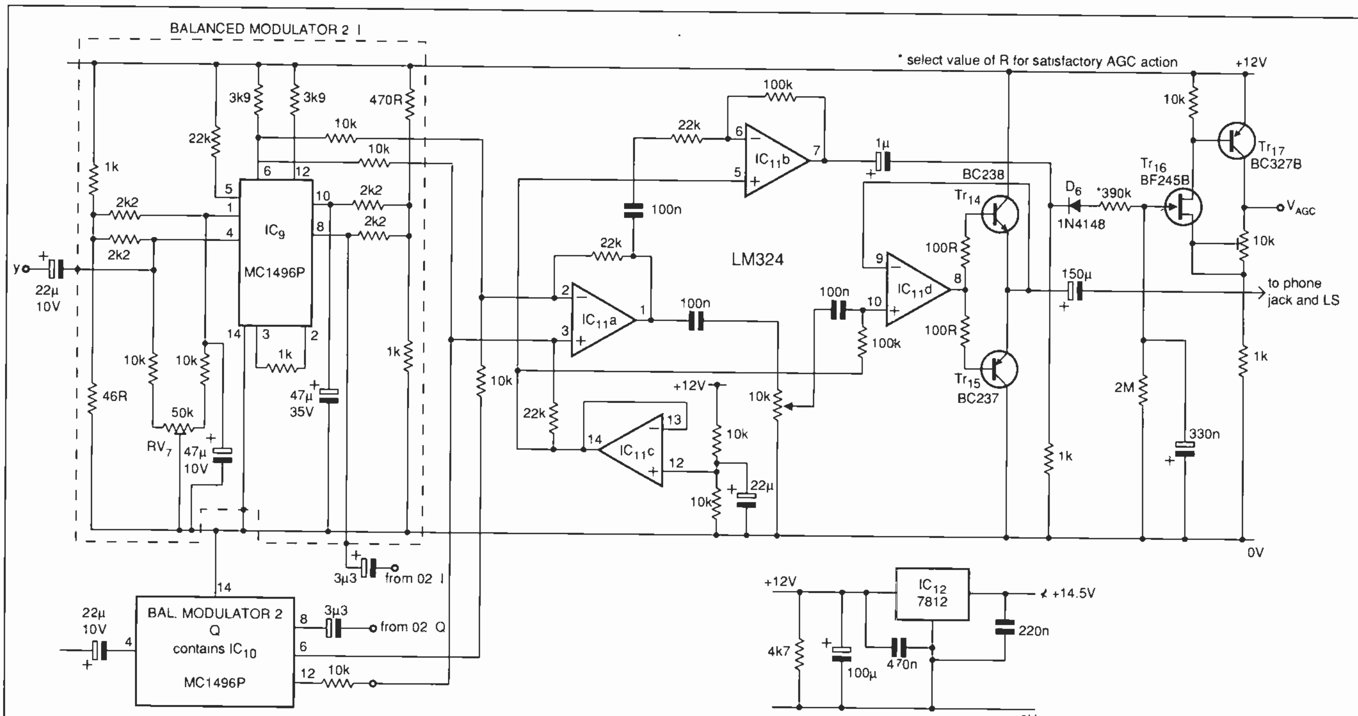
The agc circuit was suggested in Ref.5, and as stated there, it is necessary to choose carefully, how much control to use. Excessive values lead to a particularly annoying "pinging" sound as each word is spoken in this type of receiver.

Levels of carrier injection are 300mV rms for each of the rf carrier components and 50mV rms for each AF carrier component.

Alignment

Check or adjust the drive levels of both the rf and af oscillators to ensure good sinewaves. Tune the first oscillator to the correct range of 3500 to 3800kHz by adjusting (alternately) the capacitive trimmer and the value of resistor in series with the main tuning control.

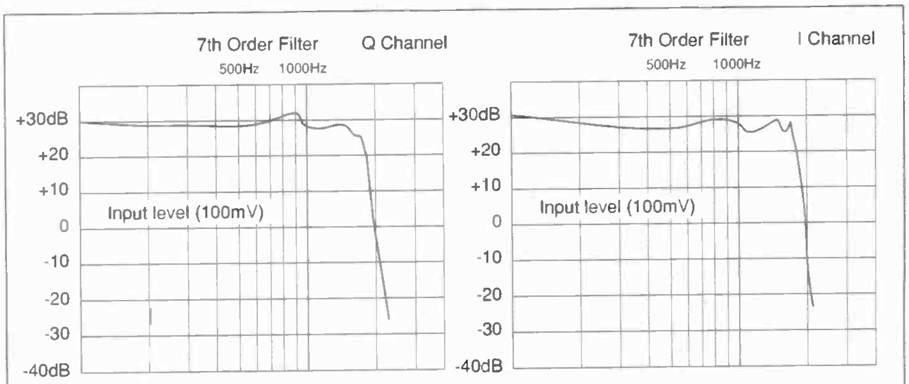
Set a phase difference of 90° between the rf carrier I and Q outputs by adjustment of the variable resistor in the phase shift network. The amplitudes should be reasonably equal. Trim the second oscillator frequency to 1900Hz by adjustment of the value of the auxiliary capacitor in the centre leg of the T-network with a resistive top.



2nd balanced modulators for I and Q channels and audio amplifier/power supply. IC11c provides a half-supply rail bias point for other op-amps in the system. The pre-amp sections would benefit from a lower noise device in place of the LM324 used in the prototype.

Trim the phase of the af carrier outputs, by adjustment of the values in the four arms of the phase shift network for a good Lissajous circle. Balance the first and second balanced modulators, observing their waveforms at the respective output pins. Apply an input CW signal, at a level of say 50μV, and observe that LSB reception is being achieved, ie. that the output tone increases in frequency as the signal generator frequency is decreased. If the opposite occurs the second oscillator carrier components should be swapped over at the balanced modulator inputs in the I and Q channels. Tune the rf amplifier circuits in the usual way, ie. coil cores at the low frequency end of the band and capacitive trimmers at the high end.

Observe levels due to a received cw signal at the outputs of the I and Q channel low-pass filters and adjust the gains of the first order stages until the channel amplitudes are equal. Then tune through the sideband range slowly. Probably, as the correct tone increases, a spurious output tone will be heard to correspondingly decrease. A careful attempt should be made to adjust the gain of one channel to maximise the loudness of the correct, ie. wanted, tone as compared to the spurious one. The only adjustment which remains is to choose the optimum value for the resistor in series with the agc diode to give satisfactory action on a strong received speech signal.



Low pass filter response curves for I and Q channels.

Performance

In use the receiver needs to be carefully earthed to minimise hum and feedback. Feeding inputs from a signal generator via a standard dummy aerial showed usable outputs from signals down to a few microvolts. On a short outdoor aerial the receiver compares well with a conventional superhet incorporating a mechanical filter, while it is pleasantly free from spurious whistles. ■

References

1. *Direct Conversion ssb receivers*, Dr.S.R.Al-Araji and Professor W. Gosling, *The Radio and Electronic Engineer*, Vol.43, No.3, March 1973, p.209.

2. *ICs simplify design of single-sideband receivers*, I.Hickman, *EW+WW*, November 1991, p.939.

3. *PA0KSB's 2-phase receiver*, a summary of the work of K.Spaargaren included by Pat Hawker,G3VA, in Technical Topics, *Radio Communication*, November 1970, p.761.

4. *Active Elliptic Audio Filter Design using Op-Amps*, D.H.G.Fritsch, G0CKZ, *Radio Communication*, February, p.98 and March, p.179 1986.

5. *ARRL Handbook 1988*, Chapter 28, *Audio and Video Equipment* page 28-11, Fig.14 and associated text.

APPLICATIONS

Please mention *Electronics World + Wireless World* when seeking further information.

Versatile switching regulator IC

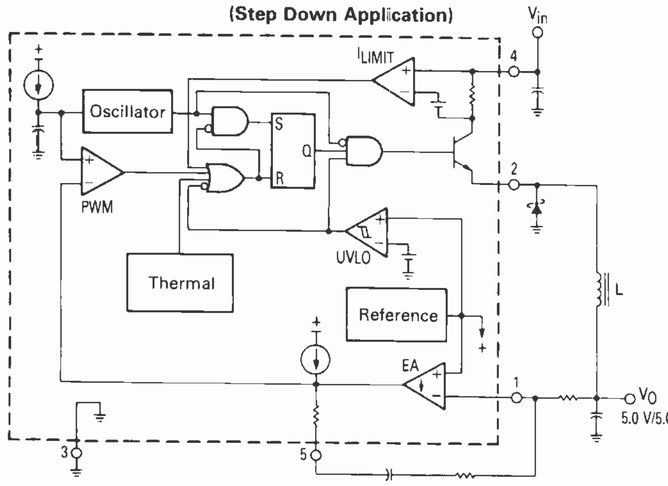
Two high performance power switching regulators containing the primary functions required for dc-to-dc converters are described in Motorola note *MC34167/D*.

These fixed frequency devices were specifically designed to be incorporated in step-down and voltage-inverting configurations with a minimum number of external components. They can also be used cost effectively in step-up applications.

Called the *MC34167* and *MC33167*, the ICs comprise an internal temperature compensated reference, fixed frequency oscillator with on-chip timing components, latching pulse width modulator for single pulse metering, high gain error amplifier, and a high current output switch.

Protective features consist of cycle-by-cycle

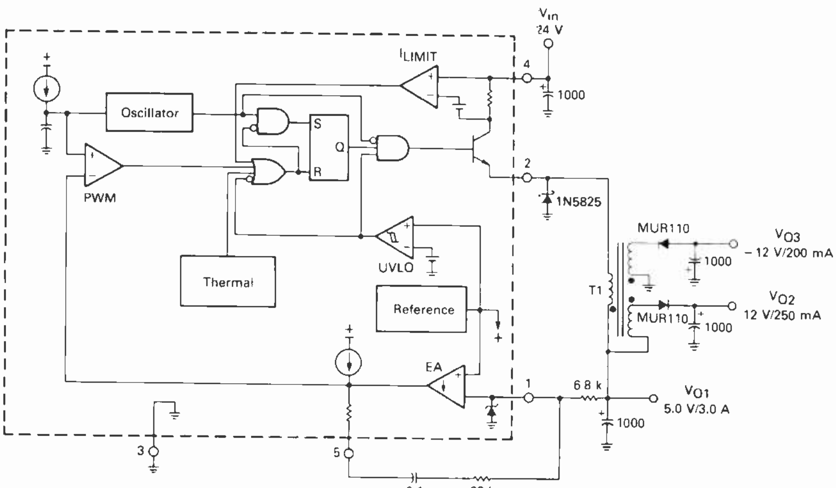
SIMPLIFIED BLOCK DIAGRAM
(Step Down Application)



Housed in a five-pinned flat pack, the MC34167 switching regulator needs few external components yet can be configured for a wide range of DC converter topologies. It delivers more than 5A switch current and operates from inputs to 40V.

Test		Condition	Results
Line Regulation	5.0 V	V_{in} = 15 V to 30 V, I_{O1} = 3.0 A, I_{O2} = 250 mA, I_{O3} = 200 mA	3.0 mV = ±0.029%
	12 V		572 mV = ±2.4%
	−12 V		711 mV = ±2.9%
Load Regulation	5.0 V	V_{in} = 24 V, I_{O1} = 30 mA to 3.0 A, I_{O2} = 250 mA, I_{O3} = 200 mA	1.0 mV = ±0.009%
	12 V	V_{in} = 24 V, I_{O1} = 3.0 A, I_{O2} = 100 mA to 250 mA, I_{O3} = 200 mA	409 mV = ±1.5%
	−12 V	V_{in} = 24 V, I_{O1} = 3.0 A, I_{O2} = 250 mA, I_{O3} = 75 mA to 200 mA	528 mV = ±2.0%
Output Ripple	5.0 V	V_{in} = 24 V, I_{O1} = 3.0 A, I_{O2} = 250 mA, I_{O3} = 200 mA	75 mV_{p-p}
	12 V		20 mV_{p-p}
	−12 V		20 mV_{p-p}
Short Circuit Current	5.0 V	V_{in} = 24 V, R_L = 0.1 Ω	6.5 A
	12 V		2.7 A
	−12 V		2.2 A
Efficiency	TOTAL	V_{in} = 24 V, I_{O1} = 3.0 A, I_{O2} = 250 mA, I_{O3} = 200 mA	84.2%

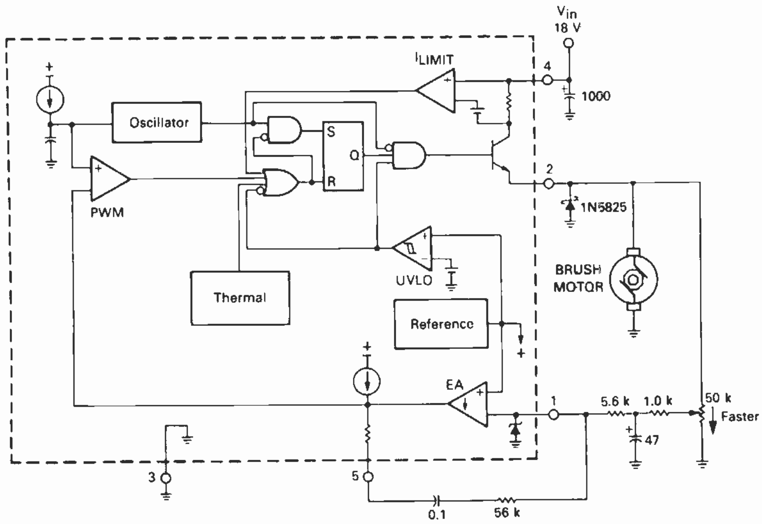
T1 = Primary — Coilcraft M1496-A or ELMACO CHK1050, 42 turns of #16 AWG on Magnetics Inc. 58350-A2 core.
Secondary — V_{O2} — 69 turns of #26 AWG
V_{O3} — 104 turns of #28 AWG
Heatsink = AAVID Engineering Inc. 5903B, or 5930B.



Triple-output converter. Multiple auxiliary outputs are easily derived by winding secondaries on the main output inductor. These must be connected so that the energy is delivered to the auxiliary outputs when the switch output turns off. During the off time, voltage across the primary is regulated by the feedback loop, yielding a constant volts/turn ratio. The number of turns for any given secondary voltage can be calculated by:

$$\text{Secondary turns} = \frac{V_{O(sec)} + V_{F(sec)}}{\left(\dfrac{V_{O(pri)} + V_{F(pri)}}{\#\,turns_{(pri)}}\right)}$$

Note that the 12V winding is stacked on top of the 5V output. This reduces the number of secondary turns and improves load regulation. For best auxiliary regulation, the auxiliary outputs should represent less than 33% of the total output power.



This variable motor speed controller incorporates emf feedback sensing. For any supply from 12 to 24V, line regulation is 1% measured at 1760rev/min, and 6% at 3260rev/min.

The output transistor is designed to switch a maximum of 40V with a corresponding peak collector current of 5.5A. This sets the power rating for both motor control and power conversion applications. Internal thermal shutdown circuitry is provided to protect the chip when junction temperature exceeds 170°C.

current limiting, undervoltage lockout, and thermal shutdown. Also included is a low power standby mode that reduces power supply current to 36μA.

Output switch current exceeds 5A, under control of a 72kHz fixed-frequency oscillator with on-chip timing. For an output of 5.05V, no external resistor divider is needed. The operating range is 7.5 V to 40 V.

The note incorporates seven design examples, including those shown, plus full design notes.

Motorola European Literature Centre, *88 Tanners Drive, Blakelands, Milton Keynes MK14 5BP. Tel. 0908 614614, fax 0908 618650.*

Low-cost evaluation for PowerPC

A module designed to aid evaluation of the first implementation of the *PowerPC* chip is described in application note *AN486* from Motorola. Broken down into four modules – cpu, ram, rom and i/o – the evaluation board is designed to combine ease of implementation and straightforward manufacture.

The risc-technology *PowerPC* chip is the only master on the board, but there are facilities for an external arbiter/master. A single bank of dynamic ram is accommodated. At 55MHz with 60ns rams, the board reads memory with 5-3-3-3 clock cycles per burst and writes with a 4-2-2-2 pattern.

For ease of programming, dual 256K-byte eprom space is eight-bit wide. Maximum ram capacity is 8Mbyte.Two standard chips from the *68000* family are used for the i/o, namely the *68681* and *68230*. Running at the system bus speed, these provide two RS232 channels, a 16bit event counter and Centronics printer interface.

Within the comprehensive note are details of *PCB* design, signal timings and software drivers.

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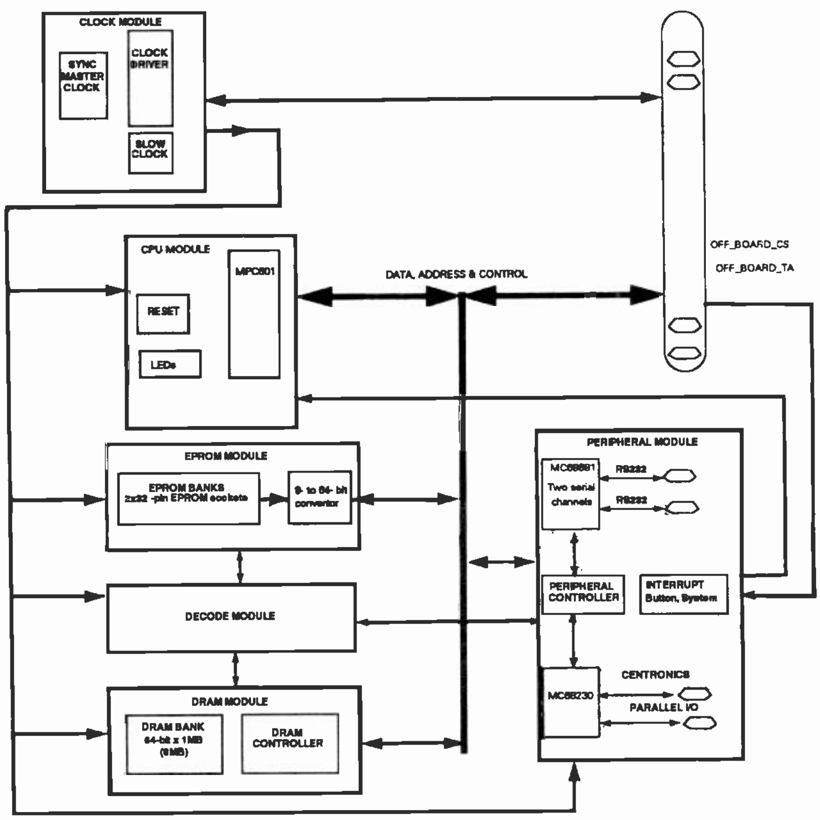


Elements of an evaluation system for Motorola's new PowerPC chip. Within the note describing this system are full details including software drivers and pcb layouts.

Using Doppler movement detection in car-alarm applications

A microwave Doppler radar module designed specifically for car alarm applications is the subject of GEC Plessey application note *AN3818-1.2*. Called the *DA5813*, the module is designed for compatibility with most existing alarm systems currently used, provided they have a volumetric sensor input. Such an input is generally used for an ultrasonic sensor. Installing and setting up the device is said to be simple.

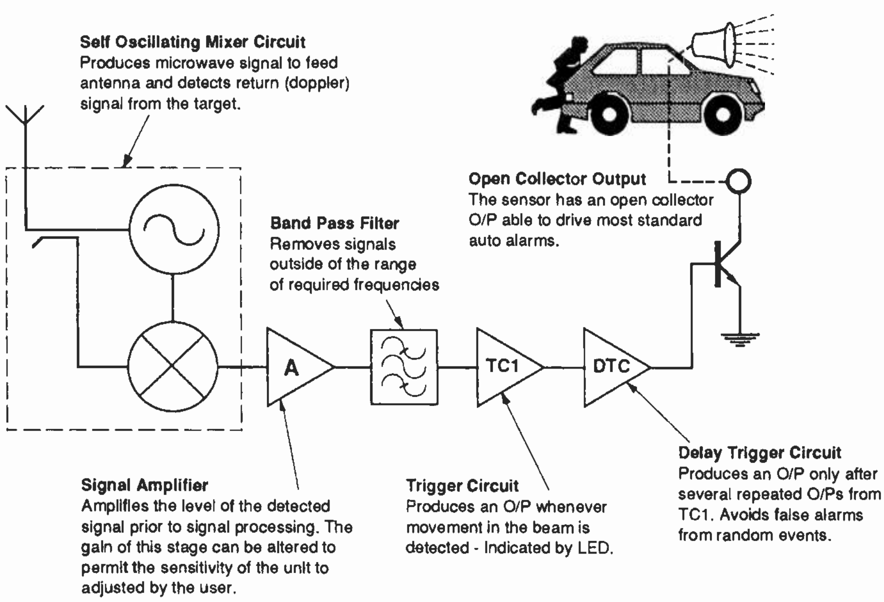
Part of the alarm sensor is a very stable transistor oscillator for generating the law power microwave signal. Operating frequency is 2.45GHz. This signal is transmitted via a simple printed dipole antenna to give an approximately omni-directional beam pattern.

Stationary objects, i.e. the vehicle interior, reflect the microwave signal with no Doppler frequency shift, hence there is no output. However, if any object within the range of the unit is moving, then a Doppler frequency shift is generated. This frequency shift is proportional to the velocity of the target.

The Doppler shifted frequency is received via the same antenna and the non-linear action of the oscillator transistor mixes the transmitted and received signals together. Output from the oscillator is a small amplitude low frequency difference signal.

Amplitude of the difference frequency is proportional to the distance and size of the target. This low frequency signal is amplified and actively filtered to reject frequencies outside the known area of interest. Window comparators decide if the signal is large enough – i.e. a real target – to trigger the output. A simple *RC* charge pump on the output comparators smoothes out spurious alarm events.

A reasonably omni-directional beam pattern is provided by the dipole antenna. If the unit is placed above a large metal surface, for example the car floor pan, the polar diagram, or area of coverage, is similar to that shown.



This Doppler radar module, built into a bolt-on box, provides an output compatible with existing alarm systems and is easily retrofitted.

The sensor uses the floor pan as a reflector. Ideally the unit should be mounted approximately 20mm above the floor pan, central in the vehicle, for example behind the hand brake. Distance of the sensor above the floor pan determines the efficiency of the unit. If this distance is significantly reduced, the sensor's detection range will be affected.

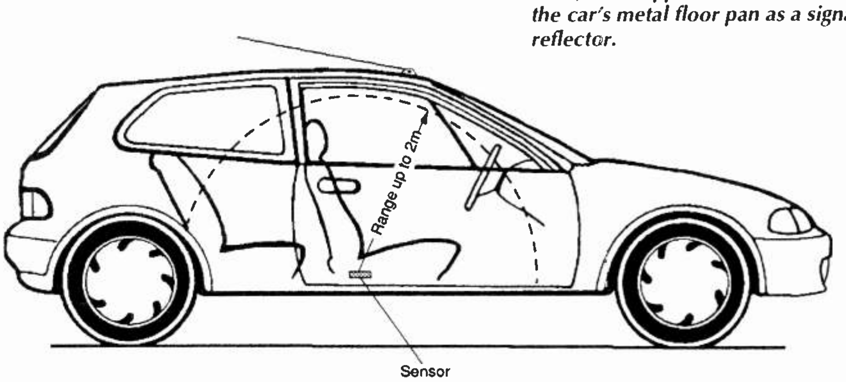
It is possible to mount the sensor under or behind plastic but not in a metal enclosure. As indicated in the polar diagram, the system exhibits a slight asymmetry. This is due to the antenna being close to one side of the box. In some applications, this effect may be useful.

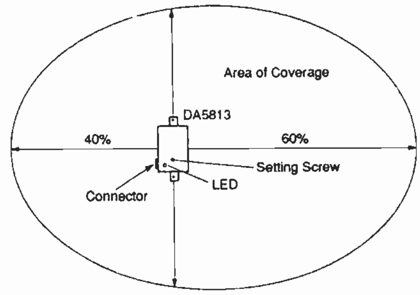
Adequate gain is available by adjusting a sensitivity control potentiometer to give a minimum of two metres coverage range when the sensor is mounted as described. Metal objects in close proximity to the sensor may distort the beam pattern and lead to areas of poor coverage. Moving the sensor slightly can overcome this.

Further information in the note discussed how the alarm is mounted and adjusted. ■

***GEC Plessey Semiconductors**, Cheney Manor, Swindon, Wiltshire SN2 2QW. Tel. 0793 518000, fax 518411.*

Mounting, say, behind the hand brake lever, the Doppler alarm module uses the car's metal floor pan as a signal reflector.





Because it is unlikely that the Doppler radar module will be mounted exactly in the middle of the car, its effective area is asymmetrical. Although proportions of the area covered are fixed, the size of the area is adjustable by simply turning a screw.

Switches for lighting

Modern high-performance chip designs make it possible to drive a fluorescent lamp using a transistor with a footprint of just 2 by 4mm. Martin Eccles looks at a number of efficient switching designs for applications ranging from emergency beacons to lcd backlighting.

Since backlit lcd appeared, designers have striven to produce ever more efficient fluorescent lamp drivers to maximise battery life. This article provides an insight into some of the key loss mechanisms in driver circuits and looks at some efficient and cost-effective designs.

Efficient fluorescent lamp drive

Figure 1's circuit meets fluorescent lamp drive requirements. Efficiency is 88% with an input voltage range of 4.5V to 20V. This efficiency figure can be degraded by about 3% if the *LT1172* V_{IN} pin is powered from the same supply as the main circuit V_{IN} terminal.

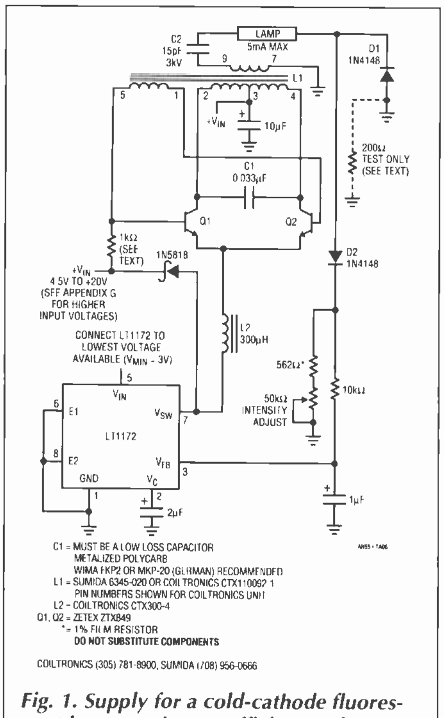


Fig. 1. Supply for a cold-cathode fluorescent lamp reaches an efficiency of 88%, provided efficient transistors are used.

Lamp intensity is continuously and smoothly variable from zero to full intensity. When power is applied, the *LT1172* switching regulator's feedback pin is below the device's internal 1.2V reference. This causes full duty cycle modulation at the V_{SW} pin, **Fig. 2**, trace A,. Inductor L_2 conducts current, trace B, which flows from L_1's centre tap, through the transistors, into L_2. This inductor's current is deposited in switched fashion to ground by the regulator's action.

Inductor L_1 and the transistors comprise a current driven Royer class converter which oscillates at a frequency primarily set by L_1's characteristics – including its load – and the 0.033μF capacitor. Driven by the *LT1172*, inductor L_2 sets the magnitude of the Q_{1-2} tail current, hence L_1's drive level.

The *IN5818* diode maintains L_2's current flow when the *LT1172* is off. The chip's 100kHz clock rate is asynchronous relative to the push-pull converter's (60kHz) rate, accounting for trace B's waveform thickening.

The 0.033μF capacitor combines with L_1 to produce sinewave voltage drive at $Q_{1,2}$ collectors, traces C and D respectively. Inductor L_1 furnishes voltage step-up, and about 1400Vpk-pk appears at its secondary, trace E.

Current flows through the 15pF capacitor into the lamp. On negative cycles lamp current is steered to ground via D_1. Positive waveform cycles are directed, via D_2, to the ground referred 562Ω-50kΩ potentiometer chain.

The positive half-sine appearing across the resistors, trace F, represents $^1/_2$ the lamp current. This signal is filtered by the 10kΩ/1μF pair and presented to the *LT1172*'s feedback pin. A control loop, which regulates the lamp, is closed by this connection.

At the IC's VC pin, a 2μF capacitor provides stable loop compensation. The loop forces the *LT1172* to switch-mode modulate L_2's average current to whatever value is required to maintain a constant current in the lamp.

Value of the constant-current, and hence

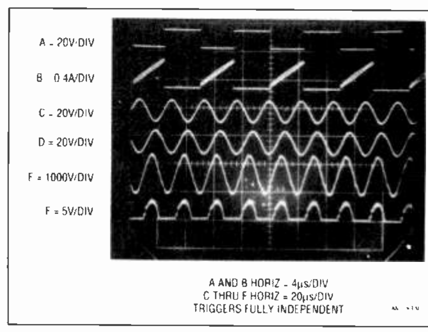


Fig. 2. Waveforms for the cold-cathode fluorescent lamp supply. Triggering is independent on traces A and B, and C through F.

lamp intensity, is varied via the potentiometer. Constant-current drive allows full 0 to 100% intensity control with no lamp dead zones or 'pop-on' at low levels. Lamp life is enhanced because current cannot rise as the lamp ages.

The circuit's 0.1% line regulation is notably better than some other approaches. This tight regulation prevents lamp intensity variation when abrupt line changes occur. This typically happens when battery powered apparatus is connected to an ac powered charger.

High line regulation performance derives

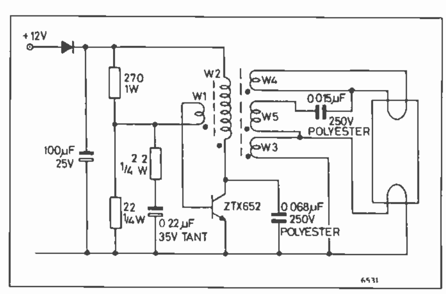


Fig. 3. Inexpensive 8W fluorescent lamp inverter operates efficiently from supplies between 10 and 16.5V.

Transformer coil details.
Core type FX3439 with 0.125mm (0.005in) spacer. Former DT2523 with enamelled copper wire, W_1=4t by 34SWG, W_2=17t by 26SWG, W_3=7t by 28SWG, W_4=7t by 28SWG, W_5=130t by 36SWG.

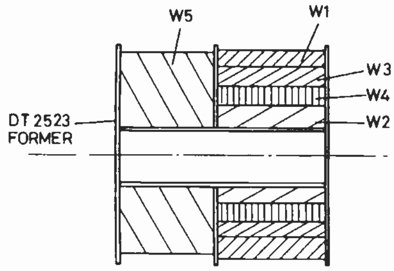
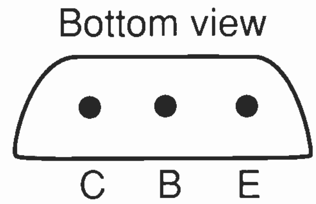


Fig. 4. Winding format the for 8W fluorescent lamp inverter.

Two free *ZTX689B* 3A transistors

Produced by UK semiconductor manufacturer Zetex, the two *ZTX689B*'s medium-power transistors given free with UK issues of this month's *EW+WW* are 20V n-p-n devices capable of up to 3A continuously – 8A peak.



Featuring high gain, combined with low saturation voltage, this medium-power device has a minimum h_{FE} of 400 and a $V_{CE(sat)}$ of 0.5V maximum at 1A collector current. High gain at high current makes the *ZTX689B* useful for driving lamps or relays directly from logic outputs. With turn-on and turn-off times of 50 and 1000ns respectively, the device is also ideal for high-efficiency power converters such as fluorescent lamp drivers – in some cases increasing efficiency by up to 20%.

Designed using Zetex matrix chip technology, the ZTX689B is TO-92 style, dissipates up to 1.5W, and operates over a wide temperature range of –55 to 200°C.

from the fact that L_1's drive waveform never changes shape as the input voltage varies. This characteristic permits the simple 10kΩ-1μF combination to produce a consistent response.

Compared to true rms conversion, *RC* averaging produces a serious error, but the error is constant and 'disappears' in the 562Ω shunt.

Efficiency of the circuit is 88%. Value of the base drive resistor – nominally 1kΩ – should be chosen to provide full saturation without inducing base overdrive or beta starvation.

Driving an 8W lamp

The circuit shown in **Fig. 3** is designed to drive an 8W fluorescent lamp from a 12V source using an inexpensive inverter based on the *ZTX653* bipolar transistor.

The inverter will operate from supplies in the range of 10 to 16.5V, thus making it suitable for use in on-charge systems such as caravanettes as well as periodically charged systems such as camping lights or outhouse lights etc. Other features of the inverter are that it oscillates at an inaudible 20kHz and that it includes reverse polarity protection.

The 270Ω and 22Ω resistors bias a *ZTX653* transistor into conduction, where the positive feedback given to the transistor by W_1 drives it into saturation, thus applying the supply voltage across W_2, **Fig. 4**. This causes a magnetising current to build up in W_2 until the transformer's ferrite core saturates. When this happens, the base drive given to the transistor by W_1 decays, causing rapid turn off.

Until the fluorescent tube strikes, the transformer is only loaded by the tube heater filaments which present only a minimal load. When the transistor turns off the transformer 'rings' for half a cycle at a frequency governed by the windings inductance and the 0.068μF capacitor, reversing the magnetising current and turning the transistor on again. This 'ring' induces a high voltage pulse across the fluorescent tube which will cause it to strike once the heaters have warmed up.

Once the tube has struck, it loads the transformer heavily, swamping this ringing action and so greatly reducing the peak voltage induced across W_2 and the transistor. This extends the non-conducting period of the inverter cycle and during this period, energy stored in the transformer in the form of magnetising current is dumped into the fluorescent tube. When all this energy has been dumped, the voltage on the transistor collector falls and it switches on once more.

Emergency fluorescent lighting

Battery powered fluorescent lighting is an important application area that significantly benefits from the very low saturation voltage *ZTX689* and *ZTX869* transistors. Housed in the E-line (TO92 style) package, these transistors replace the TO126 and TO220 types commonly used in this application, giving savings in cost and size while providing improvements in efficiency.

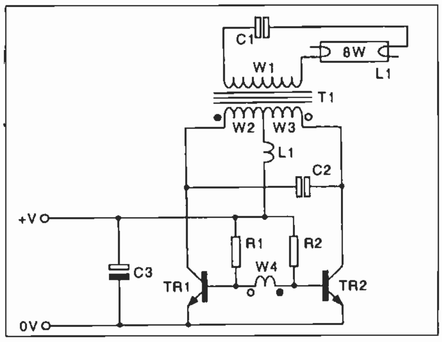


Fig. 5. Emergency fluorescent lamp controller runs from 2.4 or 4V. Low operating voltage results in fewer power cells, reduced volume and lower cost.

The circuit shown in **Fig. 5** can be used with either 2.4V or 4V supplies by selecting the appropriate component values from **Table 1**. Both designs operate from just two series connected NiCd/NiMH or lead-acid cells. The fewer cells used in a battery pack, the cheaper and more volume efficient the system will be. In addition, replacing the TO220 type transistors normally used with E-line *ZTX689* or *ZTX869* types further reduces component costs and board size.

These transistors give by far the lowest saturation voltage of devices in their class. This translates directly into improved circuit efficiency and extended battery life. With most of the remaining losses occurring in the wound components, efficiency of the 4V design is around 87% and the 2.4V design a creditable 82%.

TABLE 1. Emergency fluorescent lamp – components for 2.4 and 4V operation.

+V	2.4V	4V
$R_{1,2}$	120Ω, 0.5W	120Ω, 0.5W
Tr_1	ZTX869	ZTX689B or ZTX869
Tr_2	ZTX869	ZTX689B or ZTX869
C_1	2.2nF, 1000V polypropylene	2.2nF 1000V polypropylene
C_2	0.47μF 100V polyester	0.15μF 100V polyester
C_3	100μF 6.3V electrolytic	100μF 6.3V electrolytic
L_l	25μH (25t, 1mm copper wire on 9mm dia 25mm long ferrite rod.)	60μH 35t, 0.71 mm copper wire on 9mm dia 25mm long ferrite rod.)
T_1	FX3440 cores with 0.55 spacer or FX3670 cores with 0.65mm spacer. DT2484 coil former	FX3440 cores with 0.34 spacer or FX3670 cores with 0.45mm spacer or LA1630 (3C85 ferrite) pre-gapped core. DT2484 coil former
W_1	500T 0.18mm neatly wound (first winding)	400t 0.18mm neatly wound (first winding)
$W_{2,3}$	3t each, 0.5mm bifilar wound second and third windings)	4t each, 0.5mm bifilar wound second and third windings)
W_4	3T 031mm (fourth winding)	3t 0.31mm (fourth winding)

Note: Use insulating tape between W_1 and other windings. Core spacer must be made from non-conducting

The 4V design works for battery voltages in the range of 1.5V up to 8V while the 2.4V design operates from 0.95V to 6V. These wide operating ranges mean that the circuits will withstand the high supply voltage that can occur with rapid charging, yet they are capable of ringing the last ounce of charge from failing battery packs.

These designs give enhanced reliability in several areas. The low power losses of the *ZTX689B* and *ZTX869* transistors minimise temperature rises in the converter – important in reliability terms. Eliminating the bulk of TO126 or TO220 type transistors removes potential susceptibility to vibration. Also the circuits will withstand reverse battery connection and indefinite operation without a fluorescent tube, which is important in an unattended application.

The designs operate at a frequency of around 100kHz during striking, falling to 25kHz once struck. The circuits give an instant-start characteristic as no heater warm up time is required.

If necessary, the circuit can be adapted to operate from a single 2V cell. Other possible variants will drive higher wattage tubes at reduced power levels, giving emergency back-up for normally mains powered tubes. ■

This information is based on information provided by Zetex Semiconductors and Linear Technology.

Cold-cathode fluorescent lamps as circuit loads

Fluorescent lamps are complex transducers, with many variables affecting their ability to convert electrical current to light. Factors influencing conversion efficiency include the lamp's current, temperature, drive waveform characteristics, length, width, gas constituents and the proximity to nearby conductors.

These and other factors are interdependent, resulting in a complex overall response. **Figures 6-9** shows some typical characteristics. A review of these curves hints at the difficulty in predicting lamp behaviour as operating conditions vary.Lamp current and temperature are clearly critical to emission, although electrical efficiency may not necessarily correspond to the best optical efficiency point. Because of this, both electrical and photometric evaluation of a circuit is often required.

It is possible, for example, to construct a ccf lamp circuit with 94% electrical efficiency which produces less light output than an approach with 80% electrical efficiency.

Similarly, the performance of a very well matched lamp-circuit combination can be severely degraded by a lossy display enclosure or excessive high voltage wire lengths. Display enclosures with too much conducting material near the lamp have huge losses due to capacitive coupling. A poorly designed display enclosure can easily degrade efficiency by 20%. High voltage wire runs typically cause 1% loss per inch of wire.

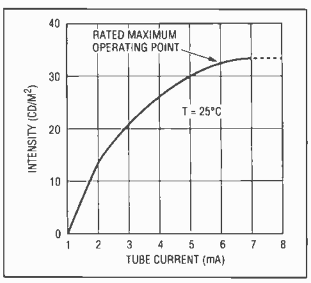


Fig. 6. Emissivity for a typical 6mA fluorescent lamp. Curve clearly flattens badly above about 6mA.

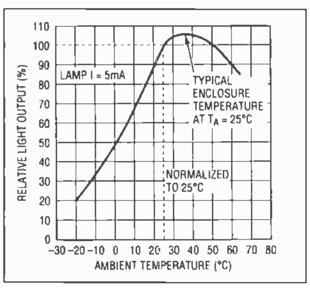


Fig. 7. Ambient temperature effects on emissivity of a typical 5mA lamp. Lamp and lamp enclosure must come to thermal steady state before measurements are made.

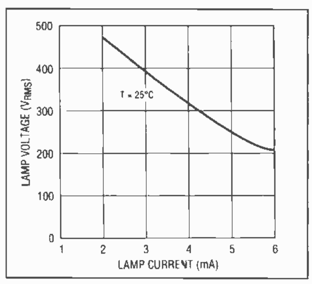


Fig. 8. Current versus voltage for a lamp in its operating region.

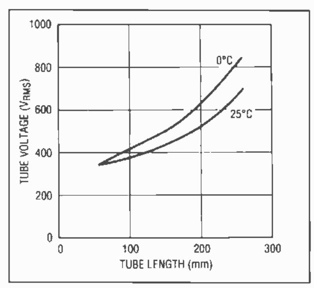


Fig. 9. Running voltage against lamp length at two temperatures. Start-up voltages are usually 50 to 200% higher over temperature.

Load characteristics of ccf lamps

Fluorescent lamps are a difficult load to drive, particularly for a switching regulator. They have a 'negative resistance' characteristic; the starting voltage is significantly higher than the operating voltage.

Typically, the start voltage is about lkV, although higher and lower voltage lamps are common. Operating voltage is usually 300V to 400V, although other lamps may require different potentials.

Fluorescent lamps will operate from dc, but migration effects within the lamp will quickly damage it. As a result, the drive waveform should be ac only

Figure 10a shows an ac driven lamp's characteristics on a curve tracer. Negative resistance induced 'snap-back' is apparent. In Fig. **10b,** another lamp, acting against the curve tracer's drive, produces oscillation.

These tendencies, combined with the frequency compensation problems associated with switching regulators, can cause severe loop instabilities, particularly on start-up. Once the lamp is in its operating region it assumes a linear load characteristic, easing stability criteria.

Lamp operating frequencies are typically 20 to 100kHz, and a sine-like waveform is preferred. The sine drive's low harmonic content minimises rf emissions, which could cause interference and efficiency degradation*.

A further benefit to the continuous sine drive is its low crest factor and controlled rise times. These are easily handled by the ccf lamp. Efficiency of the rms current-to-light output of a fluorescent lamp is degraded by fast rise high crest factor drive waveforms.

* Many of the characteristics of cold-cathode fluorescent lamps are shared by hot-cathode fluorescent types.

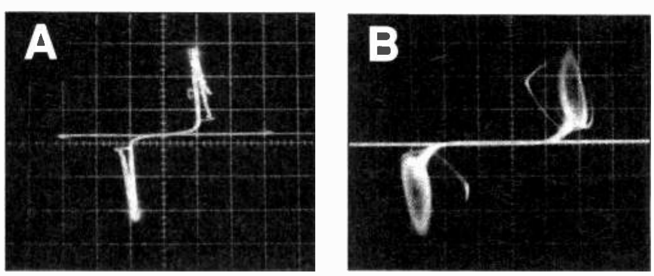


Fig. 10. Negative resistance characteristic for dual cold-cathode fluorescent lamps. The 'snap-back' effects show clearly as oscillation in b). These characteristics complicate power supply design.

Real-world control via LPT

For simple, byte-wide or analogue i/o, PC expansion slots offer a far from cost-effective solution. But there is a very low-cost and easy-to-implement alternative path to simple i/o in the guise of the standard Centronics printer port, as John Davies explains.

Rapid growth of PCs – originally designed as a stopgap to compete with early Apple computers – has resulted in some strange developments. The 640K-byte memory limit is one. Much effort has gone into designing memory drivers to overcome this limitation.

Another quirk is the PC's rather strange i/o extension capability. A limited extension bus is usually available, but it relies on rather oddly shaped and expensive pcbs. Some portable computers have no expansion capability at all. However there is a way of interfacing cheaply to any PC, portable or not, and that is via the parallel printer port. As a bonus, this method removes the need for expensive pcbs.

Printer port structure

On PC compatibles, the printer port is complies with the Centronics standard, which involves transfer of eight parallel bits of data at a time. To control data flow, a number of additional handshaking bits are needed, bringing the total number of lines to 17. These comprise 12 outputs and 5 inputs. All the lines are at ttl compatible, simplifying interfacing.

Connection is via a standard 25-way D-type at the PC, and via a standard 36-way Centronics connector at the printer. **Table 1** summarises connections at the PC end.

For controlling printer data flow, there are three bytes in the PC i/o map. Addresses of these bytes depend on which of the four LPT printer ports is being accessed. Generally, LPT1 is address 378_{16} and LPT2 is at 278_{16} but these addresses may vary, **Table 2**.

Following sections describe two simple applications – one involving reading from the printer port and the other writing. Both examples use the interrupt capability of the port. In each case LPT1 at address 378_{16} and interrupt line 7 are assumed.

Using PC interrupts

Installing and using interrupt drivers on the PC is rather cumbersome but nevertheless relatively straightforward. There are two interrupt sources, namely software or hardware.

Software interrupt sources are generated either by an 'INT xx' instruction or when a particular type of error occurs in the software.

Table 1. Pin-out of the almost universal Centronics printer port. These control signals can be adapted via simple interfacing to produce both digital and analogue i/o.

Pin	I/O	Centronics signal	Pin	I/O	Centronics signal
1	O	Data strobe	14	O	Auto line feed
2	O	Data bit 0	15	I	Error status
3	O	Data bit 1	16	O	Initialise
4	O	Data bit 2	17	O	Select
5	O	Data bit 3	18	-	0V line
6	O	Data bit 4	19	-	0V line
7	O	Data bit 5	20	-	0V line
8	O	Data bit 6	21	-	0V line
9	O	Data bit 7	22	-	0V line
10	I*	Data acknowledge	23	-	0V line
11	I	Busy	24	-	0V line
12	I	Out of paper	25	-	0V line
13	I	Selected			

Note: The asterisk indicates that the input is inverted.

Table 2. Within the PC i/o map, these three bytes are available for contrtolling the printer port. Their positions in the map depend on whether they relate to LPT1, 2, 3 or 4.

Bit	Base	Base+1	Base+2
0	Data bit 0	Unused	Data strobe
1	Data bit 1	Unused	Auto linefeed
2	Data bit 2	Unused	Initialise
3	Data bit 3	Error status	Select
4	Data bit 4	Selected	Interrupt enable
5	Data bit 5	Out of paper	Not used
6	Data bit 6	Data acknowledge	Not used
7	Data bit 7	Busy	Not used

Notes: Bit 0 is the least significant bit. Base is 378_{16} for LPT1, 278_{16} for LPT2. Base and base+2 are write only, Base+1 is read only. Bit 4 of base+2 is not available on the port but is an internal line used to enable printer port interrupts.

An attempt to divide by zero for example will cause a software interrupt.

On a PC, the most common use for software interrupts is as an entry point for the bios and dos calls. Dos calls execute primarily via interrupt number 20_{16}. Hardware interrupts on the other hand are generated by hardware, examples of which are disk drives and system timers.

Within the PC microprocessor, interrupts are handled via a jump table comprising interrupt vectors. Each vector is four bytes and points to the interrupt routine for its associated interrupt. These four bytes comprise two segment bytes and two offset bytes for the address of the interrupt routine. When an interrupt occurs, the processor calls the routine pointed to by the relevant vector. There are 256 individual vectors stored in memory starting at location 0000:0000.

Whereas software interrupts are easily handled, hardware interrupts are not so straightforward. Interrupting hardware must have some way of informing the processor of which interrupt vector it wants to trigger. This is achieved by a separate chip called a programmable interrupt controller, or pic.

In most PCs, the pic used is an Intel *8259A*, or equivalent. This device has eight separate interrupt inputs. It is programmed by the processor to generate a specific interrupt vector when each of the interrupt inputs is triggered.

Initial setting up of the vectors and other information such as whether the inputs are edge or level triggered is done by the PC bios but there is one register in the pic that has to be modified by the programmer. This is the eight-bit wide interrupt mask register, which individually enables or disables the interrupt inputs. If a bit is set to logical one, then the associated input is disabled, and vice versa if set to zero.

When modifying a particular bit, take care not to affect any of the other bits. This is achieved by first reading the interrupt mask, modifying the required bit and writing it back the interrupt mask. Code in **Listing 1** illustrates how bit 4 is enabled.

One further noteworthy point is that after an pic-generated interrupt has been processed, it is necessary to write an 'end of interrupt' instruction to the pic. This indicates that further interrupts may now be generated. Signalling an end-of-interrupt is achieved by writing 20_{16} to pic register address 20_{16}.

Before interrupt processing can begin, the relevant processor interrupt vector needs to be pointed at the by the interrupt software routine. The correct way of doing this to use dos call 35_{16} to read the current interrupt vector and call 2516 to write the new interrupt vector. It is good practice to save the current vector and to restore it after you have finished with the interrupt. Code fragment **Listing 2** illustrates the process.

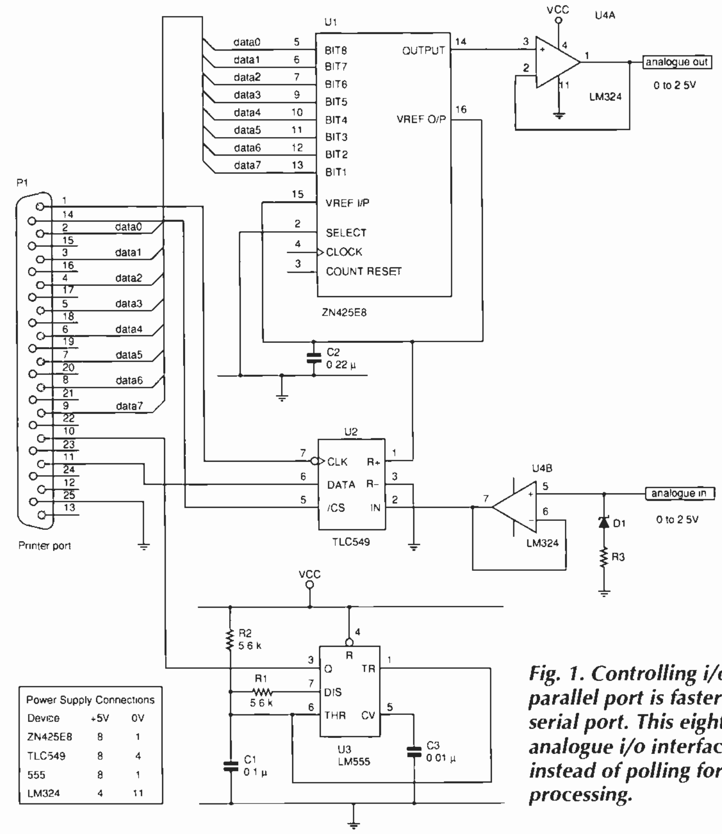


Fig. 1. Controlling i/o via the PC parallel port is faster than using the serial port. This eight-bit resolution analogue i/o interface uses interrupts instead of polling for more efficient processing.

LISTING 1. This example for enabling bit 4 in the PC interrupt controller's mask register demonstrates how to make changes to individual bits without affecting the other bits.

```
MOV AH,11101111B   ;Bit to be enabled
MOV DX,21H         ;Address of interrupt mask
register
IN AL,DX           ;Read current mask
AND AL,AH          ;Clear bit 4
OUT DX,AL          ;Write new mask
```

LISTING 2. Before interrupt processing can begin, the relevant processor interrupt vector needs to be pointed at by the interrupting software. This routine shows how to carry out the task properly.

```
        |
  MOV AH,35H
  MOV AL,int_no              ;Interrupt vector number
  INT 21H                    ;DOS call
  MOV old_vector,BX          ;Offset of old vector
  MOV old_vector+2,ES        ;Segment of old vector
  MOV AH,25H MOV AL,int_no ;Interrupt vector number
  MOV DX,OFFSET int_serv.  ;Offset of serv. routine
  MOV DS,SEG int_serv.       ;Segment of serv. rou-
tine
  INT 21H
        |
int_serv.:                   ;Interrupt serv. routine
        |
        |
```

Analogue i/o via LPT

Via the printer port, analogue i/o with eight-bit resolution and 1kHz sampling rate is straightforward. This bandwidth is obviously not sufficient for audio processing, but is more than enough for measuring room temperature, light level, battery voltage, etc. Cost of the interface hardware is also low, with a total chip cost of under £10.

The main difficulties lie in matching the connection requirements of analogue i/o devices with the pins available on the parallel port. In summary the pins available are:

Address	Port	Lines
Base	0	8 outputs
Base+1	1	5 inputs (1 interrupt)
Base+2	2	5 outputs

Analogue output is achieved using an eight-bit GEC-Plessey d-to-a converter called the *ZN4258*. There are three main reasons for choosing this chip. Firstly it is cheap, at around £3. Secondly it has a voltage output whereas most alternatives have a current output and require additional components for current-to-voltage conversion. Finally, the device has a built in voltage reference brought out on a pin. This can be used not only for the d-to-a converter but also for an a-to-d device.

To drive it, the *4258* needs only the eight data bits. Since no associated control bits are required, the device fits neatly on port zero. An output settling time of 1µs is ample for a 1kHz sample rate.

Analogue input is generally a more difficult to implement than analogue output. In addition to the data bits for reading the converted input value, two control lines are usually needed, one to start the analogue-to-digital conversion and one to indicate the end of the conversion.

Since the PC parallel port has few input pins, analogue input can be difficult. There is however an 8-bit converter – the *TLC549* – designed to circumvent this type of interfacing problem.

Rather than presenting its digital output in parallel form, the device produces serial words. These are controlled by a clock line from the reading device, in this case the PC. In this way, only device select and data clock outputs together with a data input line are needed. This requirement is met by an input pin from port 1 and two outputs from port 2.

An application note on the *TLC549*, available from TI, details operation of the device. The chip needs a voltage reference, which is supplied from the d-to-a converter as described above. Time for an analogue-to-digital conversion is 17μs, which easily meets the 1kHz sampling requirement.

Both of the above devices also operate conveniently from a 5V supply.

Circuit details

Figure 1 shows the complete i/o interface. Port zero's eight output bits are taken directly to the digital inputs of the d-to-a converter. Output of the d-to-a converter's built-in voltage reference, filtered by a capacitor to ground, is taken to its voltage reference input. This voltage reference is also taken to the positive reference input of the a-to-d converter.

Output of the d-to-a converter is buffered via U_{4a} to provide rudimentary protection against accidental short circuits, etc. Analogue input voltage is also buffered, via U_{4b}, for protection of the a-to-d converter. Zener diode D_1

LISTING 3. Routine for outputting and inputting analogue signals via the PC printer port using the circuit of Fig. 1. Elements of this routine are also useful for controlling parallel i/o.

```
NAME PARALLEL_PORT ASSUME CS:_TEXT,DS:_TEXT,SS:STACK
_TEXT                segment byte public 'CODE'

;Equates
   base_address   EQU     378H            ;LPT1 base address
   data_out_0     EQU     base_address    ;8 bit output port
   data_in        EQU     base_address+1  ;4 bit input port
   data_out_1     EQU     base_address+2  ;4 bit output port
   par_int        EQU     0FH             ;Printer interrupt no.
   pic             EQU     20H            ;Base address of PIC
   pic_stat        EQU     pic+1          ;PIC status register
   pic_eoi         EQU     20H            ;End of interrupt command
   int_mask        EQU     01111111B      ;Mask to enable ints in PIC
   clk_bit_clr    EQU     00000001B       ;Clear ADC clock bit
   clk_bit_set    EQU     11111110B       ;Set ADC clock bit
   cs_bit_clr     EQU     00000010B       ;Clear ADC CS bit
   cs_bit_set     EQU     11111101B       ;Set ADC CS bit
   int_bit_set    EQU     00010000B       ;Set interrupt bit
   int_bit_clr    EQU     11101111B       ;Clear interrupt bit
```

LISTING 3. (contd)

```
;Start of program

;Load DS with the relevant value
  PUSH CS
  POP DS
;Set up default values for ports
  MOV AL,0
  MOV DX,data_out_0
  OUT DX,AL
  MOV AL,00000001B
  MOV DX,data_out_1
  OUT DX,AL
  MOV p2_status,AL
  MOV data_avail,0
  MOV analogue_out,0
  MOV analogue_in,0
;Start interrupts
  CALL vectors
  CALL intson OR p2_status,int_bit_set
  MOV AL,p2_status
  MOV DX,data_out_1
  OUT DX,AL
;Loop reading input until
;a key is pressed read_input:
  CMP data_avail,0FFH
  JNE read_input
;Read/write data
  MOV data_avail,0
;Data processing
  MOV AL,analogue_in
  MOV analogue_out,AL
;Check for key pressed
  MOV AH,1
  INT 16H
  JZ read_input
;Terminate program
  AND p2_status,int_bit_clr
  MOV AL,p2_status
  MOV DX,data_out_1
  OUT DX,AL CALL stop_ints
  ;Remove the interrupt routine
  MOV AX,4C00H
  ;and terminate normally
  INT 21H
;Write to analogue output
  ;port (8-bit data in AL)
write_port PROC NEAR
  MOV DX,data_out_0
  OUT DX,AL
  RET
write_port ENDP
;Read from analogue input port
  ;(8 bit data returned in AL)
read_port PROC NEAR
;Put CS low to select chip
  OR p2_status,cs_bit_clr
  MOV AL,p2_status
  MOV DX,data_out_1
  OUT DX,AL
;Read 8 data bits
  MOV CX,8
  MOV BL,0
read_bit:
;Read bit
  MOV DX,data_in
  IN AL,DX
  RCL AL,1
  CMC
  RCL BL,1
;Clock in next bit
  MOV DX,data_out_1
  AND p2_status, clk_bit_set
  MOV AL,p2_status
  OUT DX,AL
  OR p2_status,clk_bit_clr  MOV AL,p2_sta-
tus
  OUT DX,AL
;Next bit
  LOOP read_bit
;Put CS high
  AND p2_status,cs_bit_set
  MOV AL,p2_status
  MOV DX,data_out_1
  OUT DX,AL
;Copy value
  MOV AL,BL
  RET
read_port ENDP
;Setting new interrupts vectors PROC NEAR
  MOV AH,35H ;Get old irq vector
  MOV AL,par_int
  INT 21H
  MOV oldint,BX
  MOV oldint+2,ES
  MOV DX,OFFSET service ;Get new irq vec-
tor
  MOV AH,25H ;Set vector
  MOV al,par_int
  INT 21H
  RET
vectors ENDP
;Activating interrupts
intson PROC NEAR
  IN AL,pic_stat ;Read PIC mask
  AND AL,int_mask
  OUT pic_stat,AL
  STI ;Enable interrupts
  RET
intson ENDP
;Stopping interrupts
stop_ints PROC NEAR
  MOV DX,oldint
  MOV DS,oldint+2
  MOV AH,25H
  MOV AL,par_int
  INT 21H ;Restore old vector
  RET
stop_ints ENDP
;Interrupt service routine
service PROC NEAR
  PUSH DS
  PUSH AX
  PUSH BX
  PUSH DX
  PUSH CS
  POP DS ;Data is in code space
  CALL read_port
  MOV analogue_in,AL
  MOV AL,analogue_out
  CALL write_port
  MOV data_avail,0FFH
  MOV AL,pic_eoi
  ;Send end of interrupt
  OUT 20H,AL
  POP DX
  POP BX
  POP AX
  POP DS
  IRET
service ENDP

;Data area
  oldint  dw 2 dup (?)
  ;Old interrupt vector
  p2_status db 1 dup (?)
  ;Copy of clock / CS output
  data_avail db 1 dup (?)
  ;data available flag
  analogue_in db 1 dup (?)
  ;current analogue data in
  analogue_out db 1 dup (?)
  ;current analogue data out
  _TEXT ENDS
  STACK SEGMENT STACK 'STACK'
  DB 256 DUP (?) STACK ENDS  END
```

and resistor R_3 may be fitted depending on the danger of an overvoltage on the input.

The data sheet for the *TLC549* recommends that its analogue input voltage does not exceed 5V. As a result, D_1 should be no more than 4.7V and R_3 selected depending on what the largest overvoltage is likely to be.

Negative reference voltage for the a-to-d converter is connected to signal ground. Two digital output lines connect from port 2, bit 0 for the clock and bit 1 for chip select. One input line is taken to bit 7 of port 1.

Clocking at 1kHz is provided by the 555 timer, which is wired for a 50% mark-space ratio square wave output. This signal feeds the parallel port interrupt line on port 1. Resistors $R_{1,2}$ may need to be trimmed if an exact 1kHz signal is needed or, for greater accuracy, a crystal oscillator with an appropriate divider chain could be used.

As with all mixed analogue/digital systems, careful circuit layout is needed to avoid interference. Analogue and digital sections should be kept as separate as possible. Individual ground and power busses should be connected at a single point as near as possible to the supply. Any additional digital devices should be adequately decoupled – preferably via a 100nF capacitor per chip – and any spare digital input lines should be tied directly to digital signal ground.

Software details

Software demonstrating how to drive the analogue i/o port comprises three main parts. One part writes data to the analogue output port, one reads from the analogue input port and one, an interrupt routine, coordinates the reading and writing, **Listing 3**.

Software for writing to the analogue output port is trivial. Eight-bit data to be written is stored in the AL register and the routine called which simply writes the data to port 0.

The procedure for reading the analogue input is more involved. To further complicate matters the two outputs and one input used, for chip-select, clock and data, are inverted in the printer port hardware. As a result, a logic one written to the port appears as a logic zero on the output.

Initially, the analogue-to-digital converter is selected by setting the chip-select line low. Converted analogue data is then read in bit by bit by toggling the clock line.

Assembly of the input byte is carried out by reading one data bit at a time into bit 7 of AL. Each bit is rotated it into the carry bit, and the carry bit is inverted to reverse the inversion caused by the hardware. Finally, each bit is rotated back from the carry bit into the BL register.

The LOOP instruction uses the CX register as a count to repeat the reading and rotating procedure eight times. Finally the chip-select line is brought high and the converted data returned in the AL register.

Distribution of processing between the interrupt routine and the normal background processing is as follows. The interrupt routine reads the analogue input and copies it to the byte variable 'analogue_in'. Flag byte 'data_avail' is then set to FF_{16} to indicate that a new value is available. In addition the byte variable 'analogue_out' is copied to the analogue output.

Linking of the input and output is done in the background routine, which loops round, reading the flag byte until data is available. Analogue input data is then copied to the analogue output data and the flag byte cleared. Any digital filtering or processing should be put in the background routine rather than in the interrupt routine.

My code was developed on a PC XT running at 12MHz. Based on measurements of how much time the *TLC549* chip-select is set low, the analogue i/o processing took around 20% of the available processing time.

Enhancements

The circuit and software in this article are fairly basic and the minimum needed for a useful system. Several straightforward enhancements are possible.

On faster PCs it should be possible to run at a higher conversion rate than 1kHz. It is not possible to say exactly how much faster as the software uses many i/o instructions. These may not run much faster than on a basic XT. The best way to check is by using an oscilloscope to look at how much time the chip-select line spends at logic zero.

This interface has only one input and one output. Expanding the inputs should be possible by adding a multiplexer on the input and controlling it using some of the spare port 2 lines. Note that switching of the multiplexer must be synchronised with reading the data because of the sampling method used by the *TLC549* as explained in the panel.

Expanding the outputs is rather more tricky as there are no spare output ports available. Output of the d-to-a converter must therefore be switched through an output multiplexer with some form of sample and hold circuit. This expansion is not likely to be as useful as input expansion.

As mentioned earlier any serious digital processing will need a more stable frequency reference than the 555 timer. A crystal oscillator and divider circuit is probably the easiest way to achieve this. ■

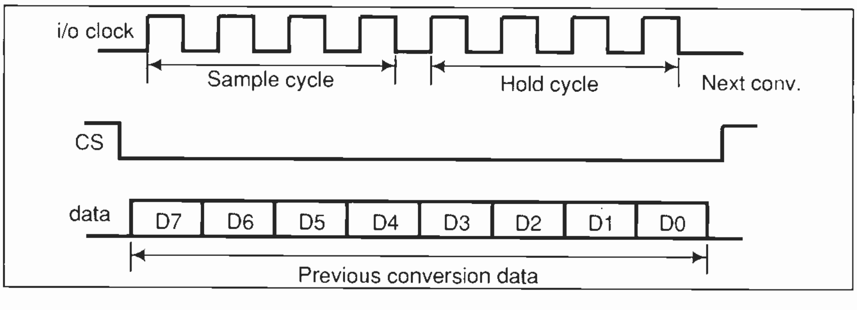
Integrated a-to-d converter with serial output

Within the *TLC549* are a timing clock, a sample and hold circuit, an eight-bit a-to-d converter and control logic. There are two control inputs, a clock and a chip select line, and one output for the data string. The internal system clock and i/o clock are used independently and need no special speed or phase synchronisation. This means that the controlling processor needs only to be concerned with the reading of the data. Relationship between the various data lines is shown below.

To start a normal control sequence, chip-select is lowered to select the device. The most significant bit of the previous conversion appears on the data pin. Falling edges of the first four i/o clock cycles shift out the second, third, fourth and fifth bits of the previous conversion result. The on-chip sample-and-hold begins to sample the analogue input after the fourth high to low transition of the i/o clock.

Three more i/o clock cycles are then applied which shift out the sixth, seventh and eighth conversion bits. The eighth i/o clock cycle is applied, setting the sample and hold circuit into the hold state. This hold state continues for the next four system clock cycles, after which the conversion cycle starts for 32 system clock cycles.

After the eighth i/o clock cycle the chip-select line must go high for at least 36 system clock cycles, i.e. 17µs for the conversion to complete correctly. If the analogue input is multiplexed then the multiplexer should be switched at this point, but note that the next value read will be the previous conversion. The control sequence can then be started again.



Further reading

Data sheets on the GEC-Plessey *ZN425E8* and Texas Instruments *TLC549* should prove useful. There is also an application note explaining how the *TLC 549* is accessed.

HIGH speed audio power

Douglas Self argues that increasing slew rate of domestic power amplifiers above a certain easily achievable limit offers only marketing benefits. Nevertheless he took on the challenge of designing for speed, and at first glance, his task seemed an easy one.

It seems self-evident that a fast amplifier is a better than a slow one. But what is a fast amplifier? Closed-loop bandwidth is not a promising yardstick. Almost certainly, any power amplifier with negative feedback will have a basic closed-loop frequency response well in excess of any aural requirements – even if the overall system bandwidth is defined at a lower value by earlier filtering.

There is constant debate about the importance of an amplifier's open-loop bandwidth, much of it depressingly ill-informed. Frequency of dominant pole P_1 defines open-loop bandwidth. I have demonstrated that this pole is variable and a rather shifty quantity that depends on transistor beta and other undependable parameters[1]. I also showed how it can be subtly increased by reducing open-loop gain below P_1.

While P_1 may vary, the actual gain at hf, say 20kHz, is thankfully a much more reliable quantity that is set only by frequency, input stage transconductance, and the value of the C_{dom} capacitor[2]. This is probably the only meaningful way to describe the amount of nfb that an amplifier enjoys.

Maximum slew rate is the most meaningful definition of amplifier 'speed'. The minimum slew rate for a 100W/8Ω amplifier to cleanly reproduce a 20kHz sinewave is easily calculated as 5V/μs. Consequently, 10V/μs is adequate for 400W/8Ω – a power level outside the realm of normal domestic audio.

A safety margin is desirable. If a factor of two is chosen, then it can be argued that 20V/μs is enough for any hifi application. There is a less obvious but substantial safety margin already built in. Maximum-level signals at 20kHz are mercifully rare in music; the amplitude distribution falls off rapidly at higher frequencies.

Wants and needs

Firm recommendations on slew rate are hard to find. Peter Baxandall made measurements of the slew rate of vinyl disc signals, concluding that they could be reproduced by an amplifier with a slew limit corresponding to maximum output at 2.2kHz. For the 100W amplifier this corresponds to 0.55V/μs slew[3].

Nelson Pass made similar tests. With a moving-magnet cartridge, he quoted a not dissimilar maximum of 1V/μs at 100W. A moving-coil cartridge doubled this to 2V/μs, and Pass reported[4] that the absolute maximum possible with a combination of direct-cut discs and moving-coil cartridges was 5V/μs at 100W. This is comfortably below the 20V/μs figure arrived at above theoretically; Pass concluded that even with generous 10:1 safety factor, 50V/μs would be the highest speed ever required from a 100W amplifier.

In the real world, the 'numbers game' also has to be considered. Everything else equal, the faster an amplifier is the better it sells. For example, it has been recently reported in the hifi press that a particular 50W/8Ω amplifier has been upgraded from 20V/μs to 40V/μs slew rate[5]. This is clearly expected to elicit a positive response from intending purchasers.

Such reports are the exception since equipment reviews in the hifi press do not usually include slew rate figures. This makes it difficult to determine the state-of-the-art. My archives reveal that top-end equipment is usually specified at around 50V/μs; slew rates are always quoted in suspiciously round numbers. There was an isolated claim of 200V/μs, but I doubt this figure.

The Class-B amplifier of **Fig. 1** has been published previously[6]; original component numbers have been preserved. This generic circuit has many advantages, though an inherently good slew performance is not necessarily one of them. However, since the topology remains the basis for the overwhelming majority of amplifiers, it seems the obvious place to start.

In a 1993 *EW+WW* article[6], I glibly stated that the amplifier's slew rate calculated at

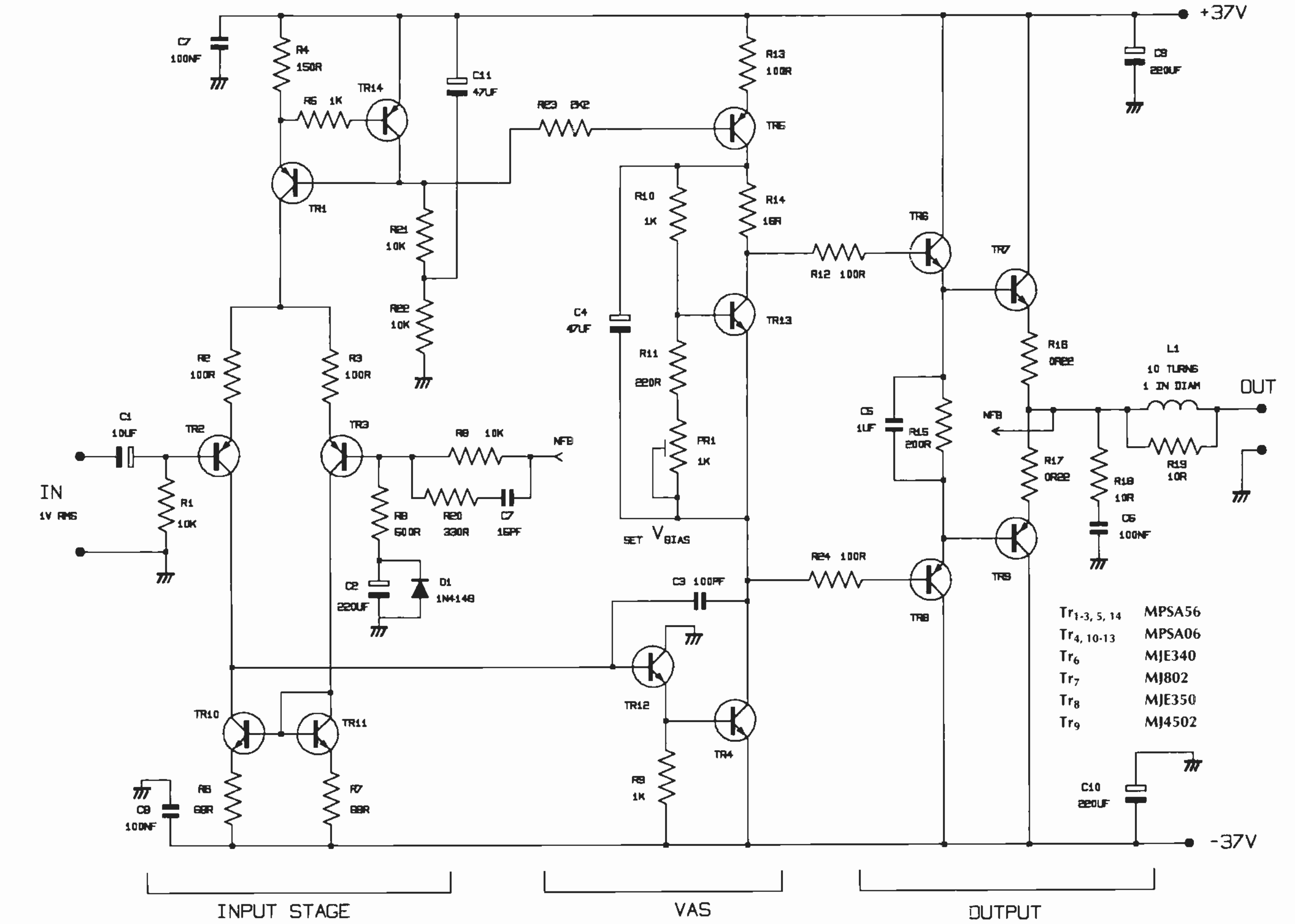


Fig. 1. The Class-B amplifier from Part 7 of Distortion in Power Amplifiers. At the simplest level the maximum slew rate is defined by current source Tr1 and the value of C_{dom}.

40V/µs, which by the above arguments is more than adequate. However, consider how improvements might be made in this figure to make the amplifier sell better.

Gaining slew

At the simplest level, slew rate in a conventional amplifier configuration like Fig. 1 depends on getting current in and out of C_{dom}.

In this case C_{dom} is C_3, and slew rate is given by the convenient relationship I/C_{dom}, in V/µs, where I is in microamps, and C_{dom} in picofarads. For Fig. 1, the relationship yields 4000/100, or about 40V/µs. This is as quoted above, if we assume – as all textbooks do – that the only current limitation is the tail source of the input pair.

If the differential pair has a current-mirror collector load – and there are good reasons why it should – then almost all tail current is available to service C_{dom}. Increasing slew rate by raising tail current seems very simple. But tail current is not the only limit on the slew current in C_{dom}. This point is touched on in an earlier article of mine[7].

Figure 2 shows the current paths for positive and negative slew limit. It is immediately clear that positive current can only be supplied by the current-source load in the voltage amplifier stage. This reduces the maximum positive rate, causing slew asymmetry if the voltage amplifier current source cannot supply as much current as the tail source. In contrast, for negative slewing Tr_4 can turn on as much as required to sink the C_{dom} current, and voltage amplifier collector load is not involved.

In most designs the voltage amplifier stage current-source value does not appear to be an issue. This is because the voltage amplifier is run at a higher current than the input stage to ensure enough pull-up current for the top half of the output stage. However it will transpire that the voltage amplifier source can still cause problems.

Measurement

Directly measuring the edge slopes of fast square waves from an oscilloscope screen is not easy. Without a delayed timebase it is virtually impossible.

A much easier, and far more accurate, method is to pass the amplifier output through a suitably scaled differentiator circuit; slew rate then becomes simple amplitude, which is much easier to determine from a graticule[2]. **Figure 3** provides a handy 100mV output for each V/µs of slew; note that the RC time constant must be short for reasonable accuracy.

Drive for the differentiator was provided directly by the amplifier, and *not* via an output inductor. Be aware that this circuit needs to be coupled to the oscilloscope by a proper ×10 probe, using the local grounding clip. Capacitance of plain screened cable produces serious under readings. Sub-microsecond pulse techniques are involved, so bear in mind that waveform artefacts such as ringing are as likely to be due to test cabling as to the amplifier, and care is essential.

Applying a fast-edged square-wave to an amplifier does not guarantee that it will show its slew-rate limits. If the error voltage so generated is not enough to saturate the input stage then the output will be an exponential response, void of non-linear effects.

For most tests described here, the amplifier had to be driven almost to clip to ensure that the true slew limits were revealed; this is due to the heavy degeneration that reduces the transconductance of the input pair. Degeneration increases the error voltage required for saturation, but does not directly alter slew limits.

Running a slew test on the circuit of Fig. 1, with an 8Ω load, sharply highlights the inadequacies of simple theory. The differentiator revealed asymmetrical slew rates of +21V/µs up and –48V/µs down, which is both a letdown and a puzzle considering that the simple theory promises 40V/µs. To obtain results worse than theory predicts is merely the common lot of the engineer; to simultaneously get results that are *better* is grounds for the gravest suspicions.

Faster, faster

Looking again at Fig. 1, the value of the voltage amplifier current-source is apparently already bigger than required to source the current C_{dom} requires when the input stage is sinking hard. As a result, R_4 can confidently be decreased to 100Ω, to match R_{13}, in an attempt to accelerate slewing.

Disappointingly, it appears that the slew rate only changes to +21V/μs, –62V/μs; the negative rate still exceeds the new theoretical value of 60V/μs. So what is wrong?

At first it seems unlikely that the voltage amplifier stage current source is the culprit. With equal values for R_4 and R_{13}, the source should be able to supply all the input stage can sink. This belief can be tested by increasing the voltage amplifier source current while leaving tail current at its original value, revealing that R_4=150Ω, R_{13}=68Ω gives +23V/μs, –48V/μs. The small but definite increase in positive rate shows clearly there is something non obvious going on in the voltage amplifier source.

This straightforward method of slew acceleration by increasing standing currents means a significant increase in dissipation for voltage amplifier and its current source. There is a danger of exceeding the capabilities of the TO92 package, leading to a cost increase. At the input stage, the problem is less acute since dissipation is split between at least three devices.

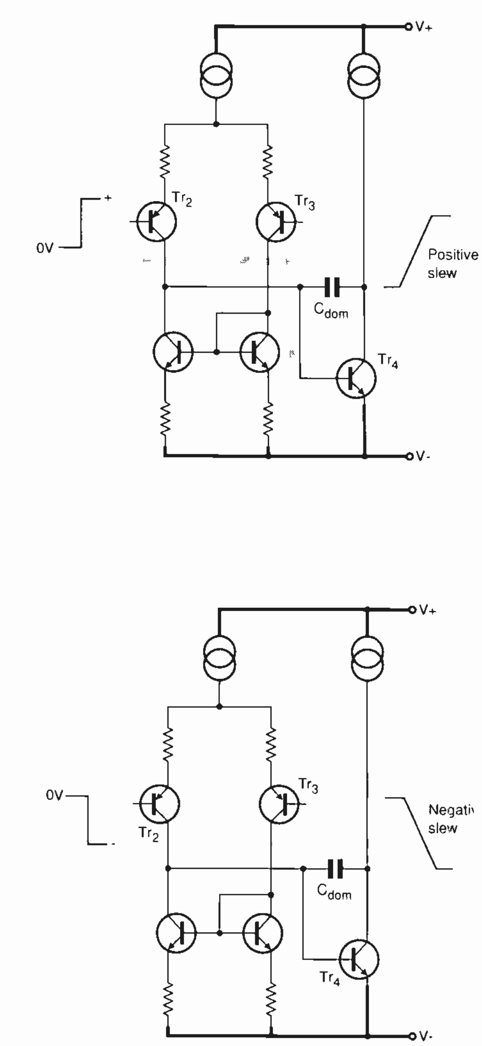


Fig. 2.a) Current path for positive slewing. At the limit, all the slewing current has to pass through the current mirror, Tr_2 being cut off. 2b) Current path at negative slew limit. Transistor Tr_1 is saturated and the current mirror cut off.

Simulating slew

Replacing the Class-B output stage with a small-signal Class-A emitter follower, the circuit was reduced to a 'model'. This model was then subjected to various Pspice simulations.

Figure 4 shows the positive-going slew of this model amplifier. Both actual output voltage and its differential are evident, the latter scaled by dividing by 10^6 so it can be read directly in V/μs from the same plot. **Figure 5** shows the same for negative-going slew. Plotting is repeated for a series of changes to resistors $R_{4,23}$ that set the standing currents.

Several points need to be made about these plots. Firstly slew rates shown for the lower $R_{4,23}$ values are not obtainable in the real amplifier with output stage – for reasons that will emerge. Note that almost imperceptible wobbles in the output voltage put large spikes on the plot of the slew rate. It is unlikely that these are being simulated accurately, if only because circuit strays are neglected. To obtain valid slew rates, read the flat portions of the differential plots.

Using this method, I obtained the first insight into slew-rate asymmetry. At audio frequencies, a constant-current source provides a fairly constant current, making it the usual choice for the voltage amplifier stage collector load. As a result the collector is exposed to the full output swing and full slew rate.

When an amplifier slews rapidly, there is a transient feedthrough from the collector to the base, **Fig. 6**, via collector-base capacitance. If the base voltage is not tightly fixed, then fast positive slewing drives the base voltage upwards. This reduces voltage at the emitter and hence the output current. Conversely, for negative slew the current-source output briefly increases[8]. In other words, fast positive slewing itself reduces the current available to implement it.

Having discovered this hidden constraint, the role of isolation resistor R_{23} immediately looks suspect. Simulation confirms that its presence worsens the feedthrough effect by increasing the impedance of the reference voltage fed to the base of Tr_5. As is usual, input-stage tail-source Tr_1 is biased from the same voltage as Tr_5; this minor economy complicates things significantly, as the tail current also varies during fast transients, reducing for positive slew, and increasing for negative.

Real life

Bias isolation resistors as used in Fig. 1 are very common. My own purpose in adding R_{23} was not to isolate the two current sources from each other at ac – something it fails to do entirely – but to aid fault finding.

Without this resistor, if the current in either source drops to zero, for example if Tr_1 fails open circuit, then the reference voltage collapses. In turn, this switches off both sources, and determining which device has failed can be a nuisance.

Accepting this, we return to the original Fig. 1 values and replace R_{23} with a link; the measured slew rates at once improve from +21, –48 to +24, –48V/μs. This is already slightly faster than our first attempt at acceleration, without the thermal penalties of increasing the voltage amplifier standing current.

The original amplifier used an active tail source, with feedback control by Tr_{14}; this was a mere whim, and a pair of diodes gave identical thd figures. It seems likely that reconfiguring the two current sources so that the voltage amplifier source is the active one would make it more resistant to feedthrough. In this case, the current-control loop is now around Tr_5 rather than Tr_1, with feedback applied directly to the quantity showing unwanted variations, **Fig. 7**. There is indeed some improvement, from +24, –48 to +28, –48V/μs.

This change seems to work best when the voltage amplifier stage current is increased, and R_4=100Ω, R_{13}=68Ω now produces +37, –52V/μs – a definite improvement in positive slewing. Note that the negative rate has also slightly increased, indicating that the tail current is still being increased by the feedthrough effect.

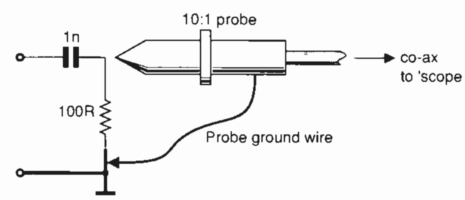


Fig. 3. A simple – but useful – differentiator. A local probe ground is essential for accuracy to exceed ±10%.

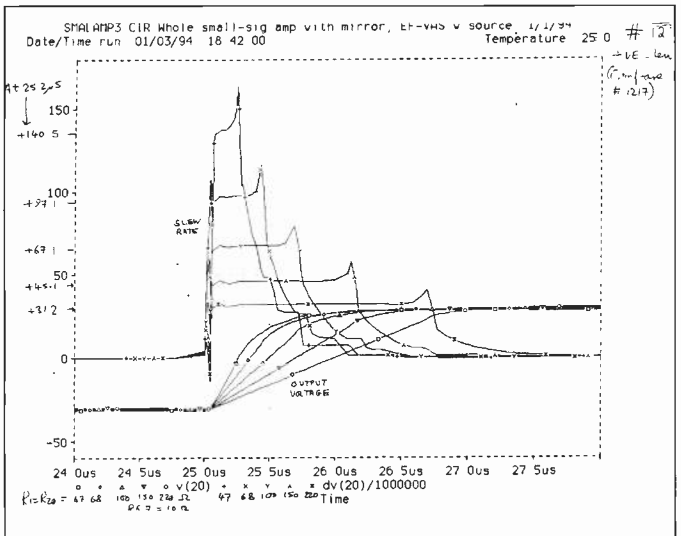


Fig. 4. Positive slewing of simulated model amplifier. The lower traces show the amplifier output slewing from –30 to +30V while the upper traces are the scaled differentiation.

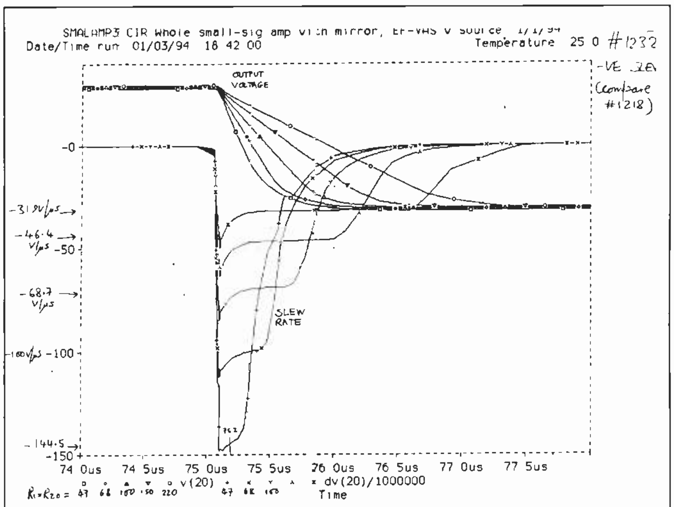


Fig. 5. Negative slewing of simulated model amplifier. Increasing the slew rate limit causes a larger part of the output transient to become exponential, as the input pair spends less time saturated. Thus the differential trace has a shorter flat period.

Minimising this transient feedthrough seems desirable, as it works against us just at the wrong time. One possibility would be a cascode transistor to shield Tr_5 collector from rapid voltage changes; this would require more biasing components and would reduce positive output swing, albeit only slightly.

Since it is feedthrough capacitance of the voltage amplifier current-source that causes the main problem, can it be turned against itself? Can the circuit be altered so that an abrupt voltage transition increases the current available to sustain it, rather than reducing it? Yes it can.

A small capacitance C_s added between Tr_5 collector carries the full voltage swing, sensing the feedback point of the active tail source. As the voltage amplifier collector swings upward, the base of Tr_{14} is also driven positive. This tends to turn it off and hence increases bias applied to voltage amplifier source Tr_5 via R_{21}.

This technique is highly effective, but it smacks of positive feedback and should be used with caution; C_s must be kept small. I found 7.5pF to be the highest value usable without degrading the amplifier's rf stability.

With R_4=100Ω and R_{13}=68Ω, adding C_s=6pF takes us from +37, –52 to +42, –43V/µs. The slew asymmetry that has dogged this circuit from the start has been corrected. Fine adjustment of this capacitance is needed if good slew symmetry is important.

Further complications

Other unexpected effects were uncovered in the pursuit of speed. It is not widely known that slew rate is affected both by output loading and the output stage operating class. For example, above I have stated that R_4=100Ω and R_{13}=68Ω yields +37, –52V/µs for Class-B and an 8Ω load. With 4Ω loading this changes to +34, –58V/µs, and again the loss in positive speed is the most significant.

If the output stage is biased into Class-A and loaded with 8Ω load then +35, –50V/µs is measured. This is explained by the fact that the output stage draws significant current from the voltage amplifier stage, despite the cascading of drivers and output devices. In the 4Ω case, the drivers draw enough base current to divert extra current from C_{dom}, and current is in shortest supply during positive slew.

With Class-A, the effect is more severe because the output device currents are always high. Even when quiescent, the drivers require more base current, and again this will be drained off from the voltage amplifier stage collector.

Speeding up this amplifier would be easier if the Miller capacitor C_{dom} was smaller. Does it really need to be so big? Well yes, because if the nfb factor is to be kept reasonably low, for dependable hf stability, the hf loop gain must be limited. Open-loop gain above the dominant pole frequency P_1 is the product of input stage g_m with the value of C_{dom}, and the g_m is already as low as it can reasonably be made by emitter degeneration.

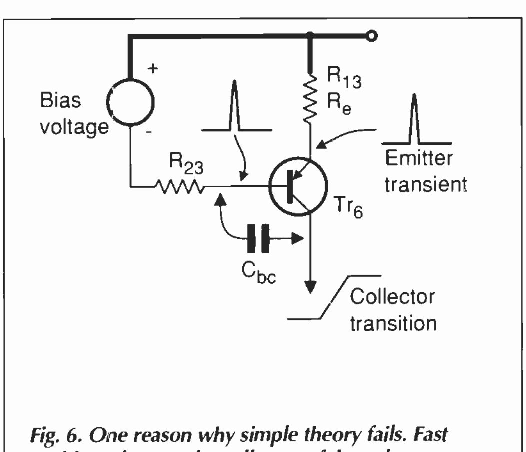


Fig. 6. One reason why simple theory fails. Fast positive edges on the collector of the voltage amplifier source Tr_6 couple through the internal C_{bc} to momentarily reduce standing current.

At 100Ω, emitter resistors $R_{2,3}$ are large enough to mildly compromise the input offset voltage. This is because the tail current splits in two through a pair of resistors that are unlikely to be matched to better than 1%. Noise performance is also impaired by this extra resistance in the input pair emitters. Thus for a given nfb factor at 20kHz, C_{dom} is fixed.

Despite these objections, the approach was tested by changing the distribution of open-loop gain between the input and the voltage amplifier stages. Resistors $R_{2,3}$ were increased from 100Ω to 220Ω, and C_{dom} reduced to 66pF. This does not give exactly the same nfb factor, but in essence the transconductance of the input stage is halved, while gain of the voltage amplifier is doubled. This gain doubling allows C_{dom} to be reduced to 66pF without affecting stability margins.

With R_4=100 and R_{13}=68 as before, slew rate is increased to +50, –50V/µs with C_s=6pF to maintain slewing symmetry. This is a 25% increase in speed rather than the 50% that might be expected from simple theory, and indicates that other restrictions on speed still exist – in fact Pspice showed there are several.

One of these restrictions is as follows: when slewing positively, Tr_4 and Tr_{12} must be turned off as fast as possible, by pulling current out of C_{dom}. The input pair therefore causes Tr_{10} to be turned on by an increasing voltage across Tr_{11} and R_7. As Tr_{10} turns on, its emitter voltage rises due to R_6. At the same time the collector voltage must be pulled down to near the –ve rail to turn off Q_4.

At the limit, Tr_{10} runs out of V_{ce}, and is unable to pull current out of C_{dom} fast enough. The simplest way to reduce this problem is to reduce the resistors $R_{6,7}$ that degenerate the current mirror. This risks hf distortion variations due to input pair I_c imbalance, but values down to 12Ω have given acceptable results. Once more it is the positive rate that suffers.

Another way to reduce the value needed for C_{dom} is to lower the loop gain by increasing the feedback network attenuation. In other words, run the amplifier at a higher closed-lop gain. This might be no bad thing. The current 'standard' of 1V for full output is, I suspect, due to a desire for lower closed-loop gain in order to maximise the nfb factor, in turn, reducing distortion. I recall JL Hood advocating this strategy back in 1974.

I have, however, left closed-loop gain alone. Of course the input signal could be attenuated, allowing more amplification, but I have an uneasy feeling about this sort of thing; amplifying in a preamp then attenuation in the power amplifier implies a headroom bottleneck – if such a metaphor is permissible. It might be worth exploring this approach; this amplifier has very good open-loop linearity and I do not think excessive thd would be a problem.

Having spent some effort on minimising distortion, I do not want to compromise the thd of a Blameless amplifier. Mercifully, none of the modifications set out here have any significant effect on the overall thd, although there my be minor variations, around 10-20kHz.

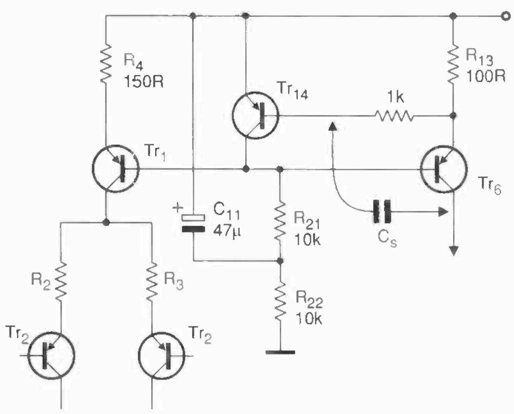


Fig. 7. A modified biasing system that makes Tr_6 current the controlled variable, and reduces the feedthrough effect.

Conclusion

Results I have obtained do not seem stunning at first sight. They do however have the merit of being as realistic as I can make them. I set out in the belief that enhancing the slew rate would be fairly simple, but the reverse has proved to be the case. It may well be that other voltage amplifier configurations, such as the push-pull stage examined in reference 1, will prove more amenable to design for rapid slew rates; however, such topologies have other disadvantages to overcome. ■

References

1. Self. D. *Distortion in Power Amplifiers, Part 3*. EW+WW. Oct 1993. p. 824.
2. Self. D, *ibid* Part 1. EW+WW. August 1993. p. 631.
3. Baxandall, P. *Audio Power Amplifier Design*. *Wireless World*, Jan 1978, p. 56.
4. Pass, N. *Linearity, Slew Rates, Damping, Stasis &...*, *Hi-Fi News & RR*, Sept. 1983, p. 36.
5. Hughes, J, *Arcam Alpha5/Alpha6 Amplifier Review*. *Audiophile*, Jan. 1994, p. 37.
6. Self. D, *Distortion in Power Amplifiers*. Part 7. EW+WW. Feb 1993.
7. Self. D, *ibid*, Part 2 EW+WW. Sept. 1993. p.736.
8. Erdi, GA, *300V/μs Monolithic Voltage Follower*, *IEEE J. of Solid-State Circuits*, *Dec. 1979, p. 1062*.

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Power Supply with Extras. Mains input is fused and filtered and the 12V DC output is voltage regulated. Intended for high class equipment, this is mounted on a PCB and, also mounted on the board but easily removed, are two 12V relays and Piezo sounder,**£3**, Order Ref: 3P80B.
Insulation Tester with Multimeter. Internally generates voltages which enable you to read insulation directly in megohms. The multimeter has four ranges, AC/DC volts, 3 ranges DC milliamps, 3 ranges resistance and 5 amp range. These instruments are ex-British Telecom but in very good condition, tested and guaranteed OK, probably cost at least £50, yours for only **£7.50** with leads, carrying case **£2** extra, Order Ref: 7.5P/4.
Mains Isolation Transformer. Stops you getting "to earth" shocks. 230V in and 230V out. 150 watt, **£7.50**, Order Ref: 7.5P/5 and a 250W version is **£10**, Order Ref: 10P97.
Mains 230V Fan. Best make "PAPST", 4½" square, metal blades, **£8**, Order Ref: 8P8.
2MW Laser. Helium neon by Philips, full spec. **£30**, Order Ref: 30P1. Power supply for this in kit form with case is **£15**, Order Ref: 15P16, or in larger case to house tube as well **£18**, Order Ref: 18P2. The larger unit, made up, tested and ready to use, complete with laser tube **£69**, Order Ref: 69P1.
12v 8ohm speaker, only £1.50 and waterproof.
Solar Charger. Holds 4AA nicads and recharges these in 8 hours, in very neat plastic case **£6**, Order Ref: 6P3.
Ferrite Aerial Rod. 8" long x ⅜" diameter, made by Mullard. Complete with two coils, 2 for **£1**, Order Ref: 832P.
Air Spaced Trimmer Caps. 2-20pf, ideal for precision tuning UHF circuits, 4 for **£1**, Order Ref: 818B.
Modem Amstrad FM240 As new condition but customer return, so you may need to fault find, **£6**, Order Ref. 6P34.
Amstrad Power Unit. 13.5V at 1.9A or 12V at 2A encased and with leads and output plug, normal mains input **£6**, Order Ref: 6P23.
80W Mains Transformer. Two available, good quality, both with normal primaries and upright mounting, one is 20V 4A, Order Ref: 3P106, the other 40V 2A, Order Ref: 3P107, only **£3** each.
Project Box. Size approx. 8"x4"x4½" metal, sprayed grey, louvred ends for ventilation otherwise undrilled. Made for GPO so best quality, only **£3** each, Order Ref: 3P74.
Sentinel Component Board Amongst hundred of other parts, this has 15 ICs, all plug in so do not need soldering. Cost well over **£100**, yours for **£4**, Order Ref. 4P67.
Sinclair 9V 2.1A Power Supply Made to operate the 138K Spectrum Plus 2, cased with input and output leads. Originally listed at around **£15**, are brand new, our price is only **£3**, Order Ref. 3P151.
Experimenting with Valves. Don't spend a fortune on a mains transformer, we can supply one with standard mains input and secs. of 250-0-250V at 75mA and 6.3V at 3A, **£5**, Order Ref: 5P167.
15W 8 Ohm 8" Speaker & 3" Tweeter. Made for a discontinued high quality music centre, gives real hi-fi and only **£4** per pair, Order Ref: 4P57.
Water Pump. Very powerful, mains operated, **£10**, Order Ref: 10P74.
0-1mA Full Vision Panel Meter. 2¾" square, scaled 0-100 but scale easily removed for re-writing, **£1** each, Order Ref: 756.
VU Meter. Illuminate this from behind becomes on/off indicator as well, 1½" square, **75** each, Order Ref: 366.
Amstrad Keyboard Model KB5 This is a most comprehensive keyboard, having over 100 keys including, of course, full numerical and qwerty. Brand new, still in maker's packing, **£5**, Order Ref. 5P202.
1 RPM Motor. This is only 2W so will not cost much to run. Speed is ideal for revolving mirrors or lights. **£2**, Order Ref. 2P328.
Unusual Solenoid. Solenoids normally have to be energised to pull in and hold the core, this is a disadvantage where the appliance is left on for most of the time. We now have magnetic solenoids which hold the core until a voltage is applied to release it. **£2**, Order Ref. 2P327.
Mains Filter. Resin impregnated, nicely cased, pcb mounting. **£2**, Order Ref. 2P315.
200VA Mains Transformer. Secondary voltages 8v-0-8v. So you could have 16v at 12A or 8v at 25A. Could be ideal for car starter charger, soil heating, spot welding, carbon rod welding or driving high powered amplifiers etc. **£15**, Order Ref. 15P51.
Prices include VAT. Send cheque/postal order or ring and quote credit card number. Add £3 post and packing. Orders over £25 post free.

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ACTIVE

Asics

Large, fast FPGA. Claimed by AT&T to be the largest and fastest field-programmable gate array on the market at 15,000 gates and 48MHz, the *ORCA ATT2C15* is now in production. Software on offer is *ORCA"x"press+* and *Verilog* logic synthesis tool from Exemplar Logic and the *ORCA Development System* (ODS 2.0), which provide a complete design system for all ORCA FPGAs. AT&T Microelectronics. Tel., 0732 742999; fax, 0732 741221.

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

Colour imaging. The 10-bit *MP8830* colour imaging subsystem by Micro Power reduces analogue front-end costs and dsp requirements and improves resolution by correcting the image pixel-by-pixel, a method that allows users to scale data from the a-to-d converter to improve accuracy and speed up image capture. It contains three a-to-ds, each having simultaneous sampling and independent, digitally controlled gain and offset adjustment. Micro Power Systems UK Ltd. Tel., 0932 857315; fax, 0932 858761.

'Fastest' d-to-a. Rockwell has the *R161008*, a new 1.2Gsample/s 10-bit digital-to-analogue converter, which it claims to be the world's fastest, operating at clock and data speeds of over 1.2GHz, while using less than 800mW. It is intended in the main for optical-fibre communication, in particular for domestic digital communications – television, voice, data and multimedia, converting fibre digital data to analogue information for domestic television receivers. A suggested alternative is to use it in conjunction with a sine-rom accumulator to perform direct digital synthesis, replacing plls in spread-spectrum transceivers such as wireless telephones. The device operates on a single –5.2V supply, interfaces to ecl and, being based on heterojunction bipolar technology, has a settling time of less than 1ns to 0.5 lsb, with a glitch impulse of under 1ps. Rockwell International Corp. Tel., 081 577 2800; fax, 081 577 2257.

Discrete active devices

Dual sm mosfet. The SO-8 package is an 8-lead soic type with a 1.8mm height off the board and IR use it to house two *Hexfet* mosfet dice, which benefit in cooler operation and higher efficiency. There are three dual n-channel devices, one dual p channel type and one dual n/p-channel module, varying from 20V and 100mΩ to 50V and 300mΩ. International Rectifier. Tel., 0883 713215. fax, 0883 714234.

1100V, 1Ω fet. Tersely named *IXTH13N110*, IXYS's new mosfet supports the claim to be the highest voltage (1100V BV_{DSS}) mosfet in the TO-247 package, also offering an $R_{DS(ON)}$ of 0.92Ω and being rated at 13A continuous. Its companion *IXTH14N100* is a 1000V, 14A device with an on resistance of 0.82Ω. IXYS Corporation. Tel., 0101 408 982 0700; fax, 0101 408 496 0670.

Linear integrated circuits

Triple 125MHz op-amp. Harris has the *HA5013*, a 14-pin package of three op-amps with rgb/composite video specification and a –3dB bandwidth of 125MHz, 0.07dB gain flatness to 20MHz for rgb and 0.03°, 0.03% differential phase and gain for composite video – said to be the best available in a ±5V/±15V triple device. Output into 150Ω is 20mA. Harris Semiconductor UK. Tel., 0276 686886; fax, 0276 682323.

Memory chips

100MHz drams. 100MHz synchronous drams in Fujitsu's *MB81116420/6820* series are available in 16M capacity and have kept, as far as possible, the architecture used in earlier dram design, so that only minor modification to equipment is needed. The devices are in a two-bank form, the pair acting alternately to allow continuous data transfer. The two organisations currently available are 2M by 4 by 2 banks and 1M by 8 by 2 banks. Hawke Components Ltd. Tel., 0256 880800; fax, 0256 880325.

Microprocessors and controllers

2.5V PIC. *PIC16C54* from Microchip is a 0.9μm 8-bit processor operating at up to 4MHz on a 2.5V supply, such as a single lithium-iron battery. On 6V, the PIC processor runs at 20MHz and provides faster instruction execution than any other 8-bit microcontroller in the price range, says the company. Features include an on-chip eprom fuse configurator to select on-chip *RC* timing and clock options. There are 512 word of eprom and 25byte static ram. Polar Electronics. Tel., 0525 377093; fax, 0525 378367.

"Green" microprocessors. 486 microprocessors from AMD are now more environment-friendly, in that they now use much less power. Power-managed *Am486* processors include a 66MHz and 50MHz clock-doubled *Am486DXL* devices and a 40MHz *Am486DXL*. Power reduction is to 30W from 100W for a typical clock-doubled 66MHz system; reducing clock frequency and turning off inactive peripherals takes it to less than 30W which, it appears, is the PC industry's definition of a green PC. Advanced Micro Devices (UK) Ltd. Tel., 0483 740440; fax, 0483 756196.

Mixed-signal ICs

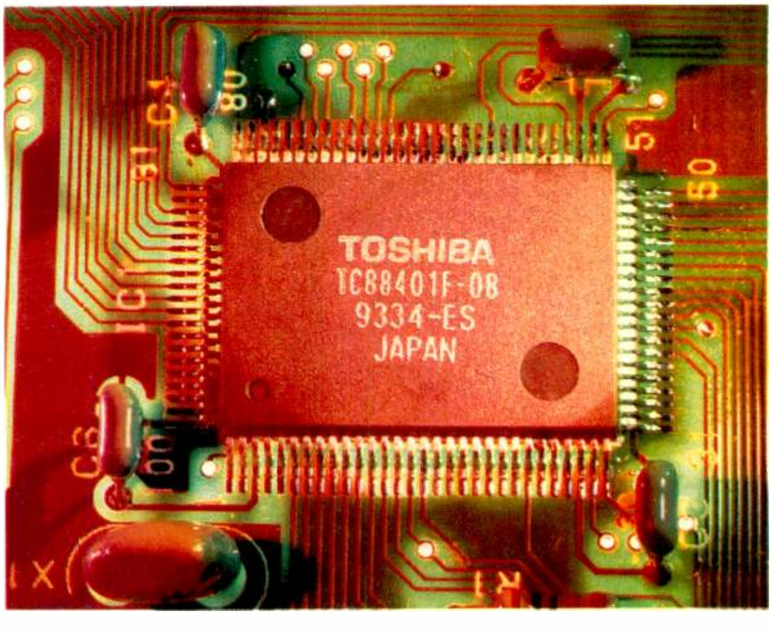
Video/audio decoders for multimedia. Hitachi and GC Technologies have a pair of decoders for MPEG1 systems that need only a 4Mb dram for a complete system. *HD814103* is a video decoder which, with the dram, supports all video functions including image magnification and reduction, while *HD814102* is an audio decoder needing no external memory. *HD184103* decodes 352 by 240 pixel images at 30frames/s or 352 by 288 pixels at 25 frames/s, allowing display size and position, window size, position and border colour to be selected. Interfaces are continuous serial, CPU command and DMA. Hitachi Europe Ltd. Tel., 0628 585000; fax, 0628 585200.

Optical devices

Red laser diode. *HL6312G* is an upgraded version of Hitachi's earlier *HL6411G* true red (633nm) visible index-guided laser diode, in that output power is now 5mW and temperature –10°C to 50°C. Luminosity is six times greater than 670nm diodes and the device operates at 2.7V maximum. Hitachi Europe Ltd. Tel., 0628 585000; fax, 0628 585200.

Digital light sensor. TI's *TSL230* is a

Record/playback chip. Toshiba's *TC88401F-08* is a cmos voice record and playback device for message systems and answering machines, using adaptive predictive coding with maximum likelihood quantisation (APC-MLQ) to provide high sampling rates and compact storage. A complete system requires the Toshiba device, a codec, memory, an audio circuit and a microcontroller, speech being recorded and played back directly or as addressable phrases in response to command or voice trigger. Up to one hour of speech in up to 256 messages are possible. Toshiba Electronics (UK) Ltd. Tel., 0276 694600; fax, 0276 691583.



programmable light sensor producing a digital output directly. Working in the 300-1100nm part of the spectrum, the device consists of a photodiode and converter which provides a pulse train whose frequency corresponds to light intensity, sensitivity being pin programmable to give a 160dB range of intensities. Output can go directly to a microcontroller, DSP or logic circuitry, the pulse-counting technique giving 16-bit accuracy and 1µs response to intensity changes. The output-enable pin allows the use of multiple sensors on one line. Non-linearity is 0.02% from zero to 100kHz. Texas Instruments. Tel., 0234 270111; fax, 0234 223459.

Oscillators

MC68040LV clock driver. *ACT22040LV* is a clock driver developed for Motorola's 32-bit *MC68040* 3.3V microprocessor and is available in standard frequencies of 25MHz and 33MHz, with alternatives between 20MHz and 70MHz as options. Rise and fall times are 2ns maximum. A 5V version is also offered for the *MC68030* and *68040*(5V). Frequency tolerance is ±1000ppm at 25°C. Advanced Crystal Technology. Tel., 0635 528520; fax, 0635 528443.

Power semiconductors

3W UHF power mosfet. Motorola's *MRF5003* 520MHz sm power mosfet puts out up to 3W at 7.5V and has a minimum gain of 9.5dB at 512MHz. Feedback capacitance is typically 4.4pF and the device will withstand 20:1 load vswr at any phase angle. Motorola Inc. Tel., 0908 614614; fax, 0908 618650.

Power mosfets. At a cell density of 6×10^6/in^2, Motorola's *HDTMOS* p-channel power mosfets exhibit on resistances of 30-150mΩ with logic-level inputs. This means that they are a viable alternative to n-channel devices in load management and high-side switching, in which they need no charge pump or power supply to boost the input. Blocking voltage is 30 or 60V and drain currents 15-50A. Motorola Inc. Tel., 0908 614614; fax, 0908 618650.

10V/5V low dropout regulator. Two outputs from the Cherry *CS-8147* low dropout regulator are the 10V ±2.5%, supplying 500mA and a secondary at 5V ±5% giving 70mA, the latter being inherently stable without an external capacitor. Both outputs are controlled by an enable pin, sleep mode needing only 70µA. Both outputs are protected against overvoltage, short and thermal runaway conditions. Clere Electronics Ltd. Tel., 0635 298574; fax, 0635 297717.

3V regulators. Allegro Microsystems has a new family of low dropout 3V regulator chips delivering 3V at up to 75mA or 150mA transient. The *8182/83/84* devices have a pmos pass element giving a 100mV dropout voltage at 60mA. Quiescent current is constant at 50µA as dropout is reached, so that data is not compromised. The *8184* is the basic version; *8183* has an enable input for control of power-up, standby and power-down; while the *8182* has the enable and a reset input. Allegro MicroSystems Inc. Tel., 0932 253355; fax, 0932 246622.

PASSIVE

Passive components

NTC thermistors. Intended to protect nickel-cadmium and nickel-metal-hydride batteries while recharging, Philips's *640 5* and *640 6* negative temperature-coefficient thermistors have R_{25} values within a 1% tolerance. They are used in chargers designed to switch off if temperature increases above a prescribed limit – around 40°C, switching on again when cooled down. Philips Components. Tel., 01031 40 722790; fax, 01031 40 724547.

Stable metal film. Resistors in Neohm's high-stability *RI* resistors hold their characteristics in a range of loads and hostile conditions. Four temperature coefficients are the *CRI*, 4.02Ω-1MΩ at 50ppm; *ERI*, 29.4Ω-301kΩ at 25ppm; *YRI*, 49.9Ω-240kΩ at 15ppm; and *ZRI*, 100Ω-100kΩ at 10ppm. All are offered in ±0.1% tolerance, with ±0.05% as an option. Voltage coefficient is 5ppm/V maximum. Surtech Interconnection Ltd. Tel., 0256 51221; fax, 0256 471180.

Chip capacitor/resistors. Murata's *CR Chip* range combines a multilayer ceramic capacitor and a resistor in one surface-mounted package. It is meant for high-speed bus termination and eliminates inductive connecting pcb traces. Values range from 10pF to 220pF and 10Ω to 1kΩ. Murata Electronics (UK) Ltd. Tel., 0252 811666; fax, 0252 811777.

RF smt interconnection. MMS from 3M under agreement with Radiall is a dual-sourced microminiature rf coaxial interconnection system including coaxial assemblies, surface-mounting receptacles and adaptors and a range of assembly and test accessories. System vswr is 1.07 at 2GHz and losses are low up to 6GHz. Straight or right-angled plugs are used

Chip capacitors. Cal-Chip has a new series of high-voltage, multilayer, ceramic chips, the *CHV series*, in three dielectrics including the very stable Class 1 COG type with a C change over –55°C to 125°C of ±20ppm/°C. Voltage ratings are 500V dc or 1000V dc, capacitance values from 3.9pF to 1nF and a best tolerance of ±1%. Terminal styles are silver palladium, nickel barrier or high leach resistant silver palladium. Cal-Chip Electronics Inc. Tel., 0101 215 672 5500; fax, 0101 215 672 5501.

Dielectric filter. AVX announces the *PDFC* series of dielectric filters meant for use in telecomms, particularly in the DECT sector. Frequency range is 1.8-2GHz, insertion loss 3dB and, for compatibility with the newest equipment, size is 6.5 by 5.5 by 3mm. Filters to provide lower insertion loss and improved stop-band attenuation are available to order. AVX Ltd. Tel., 0252 336868; fax, 0252 346643.

Displays

Electroluminescent display. Planar has a 130mm by 110mm EL display using the company's Integral Contrast Enhancement technique, which has 100% greater contrast than conventional displays, in many applications being able to replace a 5in crt. It uses less than 3W and gives a luminance better than 25cd/m^2 Planar International Ltd. Tel., (Finland) 010 358 0 420 01; fax, 010 358 0 422 143.

Packaged leds. Two new series of led packages by Dialight for use as circuit board indicators have been developed for high-density pcbs. Series *547* and *555* use 2mm leds and are available in single or quad packs in either right-angle or top-viewing mounts. Colours are red, green or yellow or combinations thereof in luminous intensities of 0.5-3mcd at 1.6-2.4V. Dialight. Tel., 0638 665161; fax, 0638 660718.

Compact lcd. Displays in Hitachi's *LMG74XX* series are liquid-crystal modules giving a black-and-white display of 240 by 128 pixels or 40 characters by 16 lines. The display controller and sram are built in, as is the fluorescent back light.The devices use the supertwist nematic technique for contrast of 20:1 and response time of 270ms. Controller may be the Hitachi *61830B* for characters or bit-mapped systems or the Toshiba *T6963* for graphics and text. Hitachi Europe Ltd. Tel., 0628 585000; fax, 0628 585200.

Filters

Cable-clamp EMI filters. The *ZCAT* series of emi filters by TDK eliminate common-mode radiation from power and interface cables without insertion, being clamped onto the cable. They come in five versions for cable diameters from 2.5mm to 13mm. Impedance at 10-100MHz is 25-80Ω and 50-150Ω from 100MHz to 500MHz. The 13mm type handles large surges without saturating. TDK UK Ltd. Tel., 0737 772323; fax, 0737 773810.

Saw resonators. *RO2101* and *RO2103* low-loss surface-acoustic wave resonators from rf Monolithics are meant for low-power rf paging and telemetry at 418MHz and 433.92MHz respectively. Insertion loss is a typical 1.6dB and the devices are frequency stable to within ±70kHz. RO2101 complies with the European ETS-300-220 standard and the 2103 with the DTI MPT1340 standard. Quantelec Ltd. Tel., 0993 776488; fax, 0993 705415.

Hardware

LCD membrane. Enco Industries has overcome the problems associated with combining a membrane control panel with an lcd and its control circuitry; previously, the need for accurate positioning of the lcd and emi/rfi protection were drawbacks. Using the Enco system, the lcd can be mounted anywhere and joined by a 'flexible tail' using a heat seal connection – a recent technique, so that the lcd is positioned exactly in the aperture in the membrane graphics panel, a screen layer being included as part of the membrane also protecting the lcd. Its control and drive circuitry can also be included within the membrane panel, so that the pcb design is simplified. A single flexible lead controls all lcd functions. Enco Industries Ltd. Tel., 05057 5151; fax, 05057 5165.

Fuse holder covers. Transparent plastic covers for 5mm by 20mm base-mounted and pcb-mounted open fuse holders by AF Bulgin shroud the live parts of the *FX0321/0267/0267/PC* models,but there are two holes in the covers to allow entry to a test probe. The cover retains the fuse when removed. Gothic Crellon Ltd. Tel., 0734 788878; fax, 0734 776095.

Touch screens. Elmwood touch screens use resistive overlay techniques for rapid response and good resolution. They come in flat or curved form with either matrix or analogue action. and can be integrated with crts, gas plasma, electroluminescent and liquid crystal. Radiatron Components Ltd. Tel., 081 891 1221; fax, 081 891 6839.

Dual-height headers. A F Bulgin has a new set of dual-height headers in the range of rising-cage pluggable terminal blocks. The headers allow double the density of circuit connections in the same board space as single-height types. 90° headers are open-ended (stackable) or closed-ended types, which mate with standard scalloped-edge units. Open types come in four and six-circuit modules, while the closed version are in standard lengths of 4-24 circuits. A F Bulgin & Co. Ltd. Tel., 081 594 5588; fax, 081 507 2691.

Instrumentation

Digital panel meters. More 'intelligent' than is normally the case, *DPMXX/4000XX* digital panel meters by Amplicon Liveline carry out data logging, non-linear scaling and linearisation of ten types of thermocouple and filtering, as well as the usual voltage and current measurement. All three in the range are controllable from the front panel or by RS-232 from a computer or comms network one having open-collector output to allow in-built alarms to operate externals, another having two alarm relays and the third four relays and the alarm signals. Quantities measured are *V* dc, *I* dc, *V* ac, *I* ac, and outputs from PT 100 sensors, in addition to temperature and 0 or 4-20mA output. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 0273 570215.

CATV test system. Using the cable itself to transmit data on head-end carrier levels to test receivers, Wavetek's *Stealth System Sweep* performs real-time, non-intrusive, full-spectrum testing of cable networks, carrier-level drift thereby being eliminated from the equation. Hand-held receivers provide a display of frequency response to 1GHz and signal analysis including video and audio carrier level and frequency and hum and carrier-to-noise levels. It is compatible with normal,scrambled and digitally encoded transmission and is frequency-agile to avoid contention with occupied channels. Wavetek Ltd. Tel., 0603 404824; fax, 0603 483670.

Cost-effective dsos. Cost of the three digital storage oscilloscopes in the *Gould 600 Series* has been kept down by eliminating those features often used by a minority of engineers, and there is a move away from extensive screen menus in favour of front-panel controls and leds. All instruments have at least a 100Msample/s rate, automated measurement and RS-423 and IEEE-488. *610* is a general-purpose type; *620* samples at 400Msample/s and *630* has a 50Kword store. All three have 100MHz bandwidth in all circumstances and there is glitch detection. Gould Instrument Systems Ltd. Tel., 081 500 1000; fax, 081 501 0116.

Audio oscillator. Kenwood offers the *AG203A* low-distortion oscillator covering the 10Hz-1MHz frequency range in five bands to an accuracy of ±3% +1Hz, with the facility of synchronisation to an external signal. Sine distortion is less than 0.1% from 400Hz to 20kHz at 7V rms, with an output-voltage flatness from 10Hz to 1MHz of ±5dB. There is a square-wave output of 10V into an open circuit with a 200ns rise time and duty cycle of 45:55 or better at 1kHz. A 0-50dB attenuator operates in 10dB steps and there is a fine adjustment control. Thurlby Thandar Instruments Ltd. Tel., 0480 412451; fax, 0480 450409.

Pressure sensor. Pressure sensors in Control Transducers's *XPRO* range are moisture-proof, water-resistant, in a stainless-steel body with invulnerability to shock and vibration. Built-in amplifiers give a conditioned output of 5V or 4-20mA, driving controls or indicators without extra amplification. Accuracy is better than 1% due to all causes and the range includes 1bar-13.8bar in absolute scaling and in sealed gauge up to 490bar. Control Transducers. Tel., 0234 217704; fax, 0234 217083.

RF/microwave hazard measurement. Holaday's battery-powered *HI-4000* consists of a number of electric and magnetic field probes to cover the 10kHz-40GHz frequency range, an lcd readout module and accessories, all in a fitted case. The probes are optically coupled to the display by low-loss cables up to 300m long to avoid perturbing the field, measurements being possible between 0.1V/m and 3000V/m. A data logger in the display provides analogue output for a plotter. Holaday Industries. Tel., 0628 478155; fax, 0628 476871.

Literature

Rack and cabling catalogue. *Vero Electronics's* 96-page catalogue describes *IMRAK 400* enclosures and *IMRAK 1400*, *2400* and *3400* floor-standing racks and cabinets. There are also details of patch panels for different connector formats, cable management products and enclosures for both copper and fibre cabling. Vero Electronics Ltd. Tel., 0703 266300; fax, 0703 265126.

PSU guide. *XP*'s 1994 power supply guide covers linear and switching types, dc-to-dc converters from 250mW 200W, Eurocassettes, 19-in rack and DIN-rail units and 'lump-in-a-cord' devices. XP plc. Tel., 0734 845515; fax, 0734 843423.

Amplicon catalogue. Sixteen new product series in the 1994 *Amplicon Liveline* catalogue include automatic data switches, optical-fibre links/repeaters, optical modems, intelligent data acquisition boards and multifunction panel meters. There is also a glossary of terms. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 0273 570215.

Power supplies

Mains converter. Accepting any voltage and frequency input and converting them to any other standard, the *Behlman Power Passport* is controlled by either a front-panel keyboard, with a *V, I, f* display, or by the RS-232 interface, which also provides communication. It has full circuit protection and the output voltage is adjustable by ±20%. Kingshill Electronic Products Ltd. Tel., 0474 327833; fax, 0474 564796.

DC power. Two more supplies in Farnell's P range are available. Model *PDS1101A* puts out 0-110V dc at up to 1A in constant current or constant voltage mode and has an lcd to show *I, V* and power simultaneously. *PSA3505A* provides 0-35V dc at 5A and uses two analogue meters. Farnell Instruments Ltd. Tel., 0937 581961; fax, 0937 586907.

24V-12V regulator. From an unstable input varying by up to ±25%, Avel-Lindberg's *XR 100DC* switching regulator provides an output regulated to better than ±1%. Rated at 100W, the 24V-12V device meets BS6527, VDE 0879, FCC Class B and other current European standards. It is input and output protected and measures 140mm by 76mm by 36mm high. Avel Lindberg Ltd. Tel., 0708 853444; fax, 0708 851040.

External equipment psu. Having a universal 90-264V/47-440Hz input and a range of single and dual outputs, the *CL25* 25W external power supply by Computer Products is intended to ease the size and shape constraints on the design of desktop and portable equipment. Single outputs are 5.1V, 9.5V or 12V, with others as options, and the dual versions have a fixed 5.1V output with floating 12V. 15V or 24V rails. All outputs are protected against short-circuits and have automatic restart. An internal battery charger is an option. The units meet major safety standards. Computer Products, Power Conversion Ltd. Tel., 0494 883113; fax 0494 883419.

Radio communications products

DECT bandpass filters. Murata's *DP* series of *Gigafil* bandpass filters represent a 60% volume reduction over the conventional type at 0.2 cubic centimetre. *DFC21R* is meant for DECT cordless telephone application in the 1880-1900MHz band. Insertion loss is 1.8dB and attenuation is 36dB at 1690MHz, 24dB at 1790MHz and 30dB at $2f_0$. Murata Electronics (UK) Ltd. Tel., 0252 811666; fax, 0252 811777.

Integrated telemetry transmitter. *RFM HX* AM transmitters are in surface-mounting packages measuring 8.6 by 10.2mm, 3mm high and need only the antenna for use in low-power telemetry. *HX1003* is for the UK 418MHz band, meeting the DTI MPT1340 standard and *HX1000* for 433.92MHz in Europe to ETS-300-220 test spec. Power is from a 3V lithium supply at 7-7.5mA. Quantelec Ltd. Tel., 0993 776488; fax, 0993 705415.

Switches and relays

SM push-button. Fujisoku smt push-button switches are made in high-temperature thermoplastic and withstand reflow soldering and temperatures up to 270°C for five seconds or 350°C for three seconds. Actuators are detachable to allow soldering, immersion washing in fluorine, alcohol or water-based fluids, and assembly. Devlin Electronics Ltd. Tel., 0256 467367; fax, 0256 840048.

Transducers and sensors

Linear displacement. Gemco *Series 951* are magnetostrictive linear displacement transducers from MagneTek. They contain all necessary signal conditioning, are field-programmable and produce outputs of 10V analogue voltage or 4-20mA current, ttl digital level, RS422 start/stop pulse or RS422 pulse-width modulation. Eurosensor. Tel., 071 405 6060; fax, 071 405 2040.

Capacitive accelerometer. Endevco's *Model 729A* is a low-g, variable-capacitance accelerometer with its own internal electronics. Near-critical gas damping produces stable frequency response with temperature and the unit is able to withstand shocks of 10,000g and still measure

Narrowband radio modules. A matched transmitter/receiver pair by Circuit Design for the pan-European ETS-300-220 frequency of 434MHz is a now available. CDP-TX-01 is a narrowband transmitter using a crystal oscillator with a ±4ppm stability and direct fsk modulation to allow 12.5kHz channel spacing.The CDP-RX-01 receiver is a crystal-controlled double superhet with 120dBm sensitivity and ±5kHz selectivity. Low Power Radio Solutions Ltd. Tel., 0993 709418; fax 0993 708575.

mg accelerations immediately afterwards. Features include dc response, ±2V output and temperature –54°C to 121°C. Full-scale ranges are 2-100g and power regulation, amplification and filtering are built in. Endevco UK Ltd. Tel., 0763 261311; fax, 0763 261120.

Two-wire absolute encoder. Control Transducers's *AD series* of absolute digital encoders are single-turn, non-contacting, optical sensors that report shaft position within a 360° range without reset or a homing cycle, since the output codes are unique to a given position. The units include an RS232/485 interface, 2-65535 codes/rev, 38.4, 57.6 or 115.2kbaud at 9 or 12-bit accuracy and up to 15 encoders on one SEI bus. Maximum shaft speed is 10,000rev/min. Control Transducers. Tel., 0234 217704; fax, 0234 217083.

COMPUTER

Computer board-level products

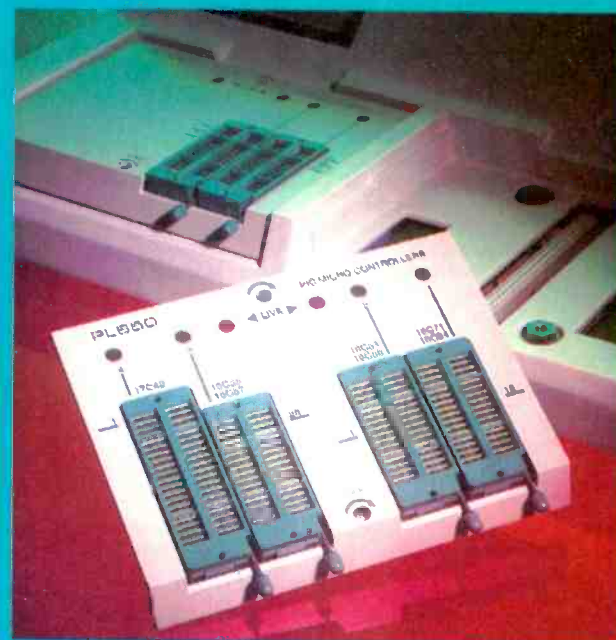
100MHz 486 board computer. New to Ampro's Little Board family of 486 single-board computers is the *Little Board/486 DX4* 100MHz embedded controller, which has a 1.5Mbyte on-board bootable solid-state disk, allowing eprom, flash eprom or static ram to substitute for magnetic disks. The unit is effectively a PC AT and several expansion cards in the space of a half-height 5.25in disk drive. A single PC/104 module fits within the board volume. Diamond Point International Ltd. Tel., 0634 722390; fax, 0634 722398.

VME-PCI adaptor. Bit 3's *Model 617 Adaptor* connects a PCI Local Bus computer to a VMEbus system, sharing memory and special-purpose boards and providing data transfer between systems at up to a sustained 26Mbyte/s. Either system can be a bus master on the other. It consists of two cards; a short form factor PCI bus card and a 6U VMEbus card, interconnected by up to 25ft of cable or optical-fibre using a Bit 3 interface. Bit 3 Computer Corporation. Tel., 0101 612 881 6955; fax, 0101 612 881 9674.

Data communications

Parallel-port data acquisition. Computer Instrumentation's *mini-POD* series of data acquisition modules are connected to a PC or notebook parallel port, taking power from the port and needing no other connection than the signal. The range includes an 8-channel, 12-bit programmable a-to-d converter, a 4-channel 18-bit a-to-d, 2-channel thermocouple converters for types K, T and R and a PT100 resistance thermometer with a resolution of 0.01°C. Cost includes software drivers for Basic, Pascal, C and Visual Basic, and CI's SoftScope and data logging software. Computer Instrumentation Ltd. Tel., 0903 700755; fax, 0903 700788.

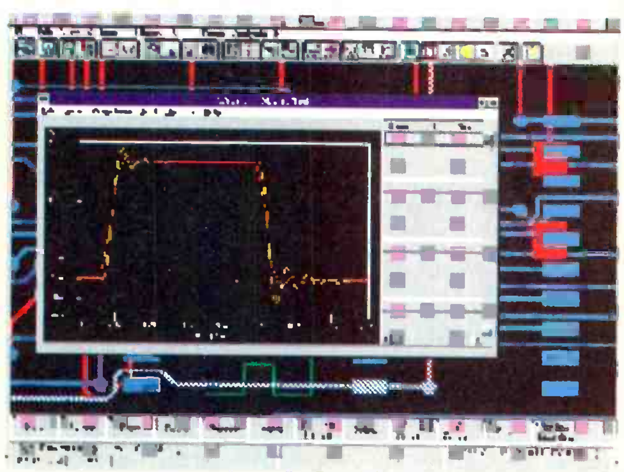
Transducer interrogation. Katon Ingram offers the *TDP2511 Transducer Data Pod*, which interrogates up to 128 strain gauges or pressure transducers. It links the transducers to a host computer using an RS-232 interface, each transducer having an address in a sequence. Three modes are used: multidrop, in which data is requested from any address from the computer; each address sending information at a predetermined, adjustable rate; or one address in use, data being continuous. The interface consists of a strain-gauge amplifier with a continuously adjustable sensitivity from ±0.68mV/V to ±8mV/V, the ±10,000 count output being calibrated in the relevant engineering units. Having been set, the instrument retains all digital tare load and calibration information in non-volatile memory. *TDP2511* is for use with bridge transducers with full or half bridge resistances of 120Ω-10kΩ. Calibration is by on-board analogue adjustment or digitally by the computer. Katon Ingram Ltd. Tel., 0983 822180; fax, 0983 822181.



ST62xx gang programming. Two new modules for the Lloyd Research L9000 modular programmer handle 20-pin and 28-pin devices such as the ST62E20/25 and ST62T20/25. Unusually, the L9000 will take a second module, so that capacity can be increased as a new model's production volume expands. The PL620 module has four dil sockets and the PL621 four SOIC sockets and, since the L9000 is a general-purpose programmer, a second module can be fitted to program devices such as eproms in dil, PLCC or LCC packages, or other microcontrollers. Lloyd Research Ltd. Tel., 0489 574040; fax, 0489 885853.

Computer systems

CRT workstation. A crt version of the Blue Chip lcd workstation is announced. An 8U high enclosure contains motherboard and passive processor options from 386SX to P24 and bus options are PCI, VL and ISA with up to eight available slots. Sealing is to IP65 and rack, panel and desk versions are made. crt resolution is 1024 by 768 and there is a 57-key pad. PCMCIA devices or floppies are in a lockable bay and the minimum hard disk is 170Mbyte. Blue Chip Technology. Tel., 0244 520222; fax, 0244 531043.



Windows NT PCB design. Two new pcb design tools from Intergraph Electronics are the *Veribest PCB Designer* and *VeriBest Signal Integrity*, both running under Windows NT. New features in the former include shape-based auto routeing and editing, improved database access, automatic test-point generation and user-configured reports. Signal Integrity is an analysis tool providing crosstalk estimation, etch delay figures, resistance, capacitance and characteristic impedance from the routeing and editing environment. Waveforms resulting from transmission-line simulations are displayed on screen, where 'what if?' analysis can be performed. Dynamic "push-and-shove" allows interactive routeing in which features blocking a route can be moved manually. Routeing can be selected to allow for timing or length constraints. Intergraph (UK) Ltd. Tel., 0793 492733; fax, 0793 492940.

Development and evaluation

8051 emulator. Metalink's *iceMaster-PE 8351FX* supports the 8351FX range of microcontrollers – a second-source 8051. The device connects to the host computer via RS-232 and directly to the target system cpu socket with no probe or host card necessary. There is transparent trace memory to allow viewing during emulation, a performance analyser and symbolic and source-level debugging. Reflex Technology Ltd. Tel., 0494 465907; fax, 0494 465418.

Universal gang programmer. Using universal pin drivers, Concentrated Programming's *Sprint Multisyte* is a gang device programmer that copes with virtually anything in sight – arrays, memory and microcontrollers. Three versions exist: dual, quad and octal provide two, four or eight programming sites, each being configured with a variety of TOP modules to take almost any type of package and each being programmable separately for running test vectors. Concentrated Programming Ltd. Tel., 0279 600313; fax, 0279 600322.

PIC processor programmer. Lloyd Research's *L9000* programmer will now handle PIC processors from Arizona Microchip, such as *16C5x*, *16C71/84* and *17C42* with the *PL650* socket module. *L9000* operates alone or may be operated from a PC or a PC batch file. Most devices are programmed in around three seconds, those normally programmed serially on development programmers being handled in parallel. The device has positions for two modules from different families. Lloyd Research Ltd. Tel., 0489 574040; fax, 0489 885853.

PLD training. *PAL Trainer* from Flight Electronics is a PC-based learning aid, specifically for third-year students but also useful for engineers, to help people learn about programmable logic devices and Palasm, and to function as an array programmer. The training course progresses through initial logic design, through PC simulation, programming and test, using examples provided. Students then program devices themselves using Palasm V.4. Hardware includes a cased PCB combining a GAL (Lattice Semiconductor's gate-array logic) programmer and the test unit, an interface card, cable, a disk, sample chips, Palasm and a manual. Flight Electronics International Ltd. Tel., 0703 227721; fax, 0703 330039.

Socket modem kit. *SocketModem Designers' Kit* from Rockwell is an evaluation platform for single-board voice/data/fax modems, the kit containing a dip modem and a BABT-approved line interface. The mother board has 128K of rom, serial DTE connector, led indicators, telecom connectors and a power connection for the supplied 5V supply. All the user needs is a PC XT or higher with a text editor and a prom programmer, all the other software being supplied. The unit copes with a large range of modem standards, depending on the version used. RCS Microsystems Ltd. Tel., 081 979 2204; fax, 081 979 6910.

Computer peripherals

486DX upgrade. The Aries *Upgrade Socket* and *Upgrade Adapter* contain the necessary modifications to allow a DX4 to be used in 486DX/DX2 computers. For pin-grid arrays, the DX4 goes into the *Socket*, while the Adapter takes the DX4 in SQFP form, in which case soldering is needed, Aries supplying the solder package if needed. Aries Electronics (Europe). Tel., 0908 260007; fax 0908 260008.

Software

State-diagrams-to-ANSI C. Using *Abel2C* from Visual Software, graphical state diagrams can be converted to ANSI C, the compiled C code being executable. It works in conjunction with *StateCAD*, which allows state diagrams to be drawn and compiled into Abel HDL, which now feeds *Abel2C*. Applications include hardware emulation, failure analysis and test vector simulation. A simulation procedure library, created automatically when a file is converted, allows inputs to be set, results to be compared, errors isolated and state machine flow to be analysed. ARS Microsystems Ltd. Tel., 0256 381400; fax, 0256 381685.

Spice library. Analog Devices's *ADSpice* Revision 1 disk contains 392 advanced Spice models, 40 of them new, including video amplifiers, voltage references and the BUF04 high-speed buffer. The simulation techniques used allow emulation of ac and dc performance and thermal, noise and other characteristics. The free Spice library comes on a 3.5in PC-compatible disk. Analog Devices Ltd. Tel, 0932 253320; fax, 0932 247401. ■

USING RF TRANSISTORS

Compensation terms and networks

Norm Dye and Helge Granberg show how frequency affects design of impedance compensation networks and explain why negative feedback is so simple – and effective.

From the book RF Transistors: principles and practical applications.

Amplifier frequency compensation equalises the input impedance of a transistor so that the matching element can look into a relatively constant *R* and *Z* over a given bandwidth. Narrow band designs, using *L-C* matching elements, do not usually require it since bandwidth is limited to 5% or 10% by the matching element.

Wide-band designs with bandwidths greater than 10% are generally combinations of *L-C* or microstrip, and wide-band transformers; or wide-band transformers alone. With complex *L-C* or microstrip designs, transistor impedance matching over bandwidths of half an octave or more is possible. But it is not really good practice. Input impedance of a transistor (bjt or fet) varies with frequency much more than does the output impedance, so only the input usually needs to be compensated.

At power levels higher than a few watts, where output impedance is low, losses in the compensation networks make output compensation impracticable. But it is sometimes carried out using just a series inductance, for example, with a capacitive output, or with shunt capacitance and an inductive output. *Ls* and *Cs* can not both be used with wide-band transformers because shunt capacitance is used to compensate for leakage inductance.

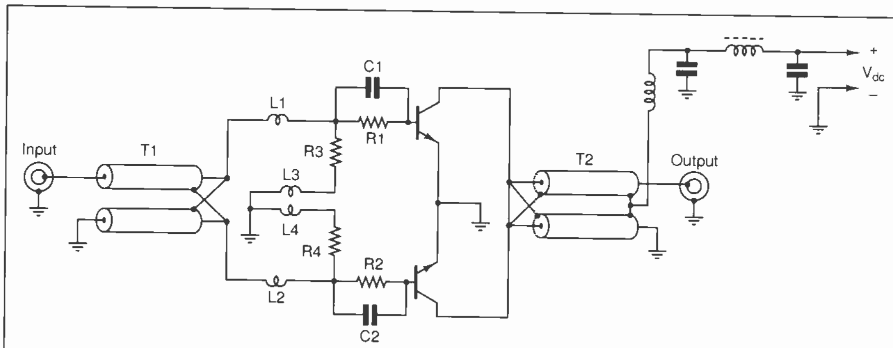


Fig. 1. Push-pull amplifier showing networks for input impedance compensation as well as for levelling of the power gain in wide-band applications.

Network effects

In certain inter-stage matching arrangements, losses must be tolerated. If a power amplifier operates at a power level of 150-200W and has a power gain of 6-7dB, then the driver power output would be 30-50W and would have (for a 12V design) an output impedance of around 1.5Ω. Assuming the power amp input has a frequency compensation network, part of the drive power will be dissipated in it as well as in the matching network itself. In the above case, the result would be considerable power loss, lowering overall efficiency of the system and possibly making necessary an additional amplifying stage.

Wide-band amplifiers tend to use push-pull designs because they make it easy to achieve low emitter-emitter or source-source inductances – much easier than low emitter/source-to-ground inductances (important in a single-ended design).

The input/output impedances are also higher, simplifying design of wide-band impedance matching networks.

Transistor input impedance is high at low frequencies, and low – and more reactive – at high frequencies. The change is around 40-80% per octave depending on frequency spectrum and device type.

Such behaviour is true for both bjts and fets, although input impedance of a fet for a given electrical size is higher, particularly at lower frequencies. If the device input crosses over from capacitive to inductive within the desired frequency band, compensation-network design

Table 1. Typical component values for the networks of Fig. 1 applied to a 2-30MHz 200W amplifier design.

L_1, L_2	27-33nH
L_3, L_4	35-40nH
C_1, C_2	2000-2800pF
R_1, R_2	10-15Ω
R_3, R_4	8.2-12Ω.

becomes even more difficult.

For a capacitive input, a shunt LR combination (**Fig. 1**) 12-1) R_3 /L_3 and R_4/L_4 will serve as an initial compensating network.

Reactance, inductance and frequency

Ideally, reactances of inductances L_3 and L_4 will be very large at the high-frequency end of the band, and the shunt circuit will have negligible effect.

At the low-frequency end, the reactances of L_3 and L_4 become low, effectively leaving only R_3 and R_4. Since the reactances of the series inductors L_1 and L_2 are also at their minimum values, the series combination of R_3/R_4 will, in fact, be in parallel with the output of T_1, presenting an artificial load to it. At high frequencies the reactances of L_1 and L_2 are adjusted to a value, which in series with C_1/R_1, and C_2/R_2, results in a load to T_1 comparable with the low input impedances of Tr_1 and Tr_2.

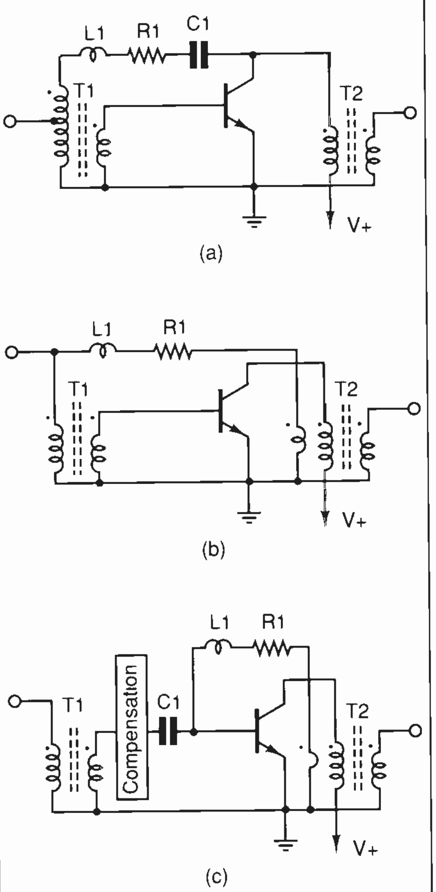
C_1/R_1 and C_2/R_2 are actually used for gain-levelling rather than for frequency-compensation. The idea is that the reactances of C_1 and C_2 are low at high frequencies, where power gain is lowest. At low frequencies, the higher device power gain is lowered by the increased reactances of C_1 and C_2 to produce a more even response, leaving R_1 and R_2 as the main power carriers to the bases. Thus R_1/R_3 and R_2/R_4 form π attenuators, with the input impedances of Tr_1 and Tr_2 serving as the second shunt leg. The values of all Rs can then be calculated when the transistor input impedance and the desired power gain slope are known.

Typical component values for the networks of Fig. 1 applied to a 2-30MHz 200W amplifier design are shown in Table 1.

Designing a network to match a "load" that changes from capacitive to inductive as a function of frequency is a difficult task. But with internally matched transistors, the situation is different. Power gains and input impedances for their specified frequency ranges are much more constant than those of non-internally matched transistors. So frequency compensation is often not required at all since the maximum bandwidths that can generally be obtained with internal matching are less than three octaves.

Very low Q, broad-band circuits used to "match" the input or output of a transistor can be realised at low frequencies (below 100-200MHz) using low Q matching networks – for example, broad band transformers. When designing such circuits, the input/output impedance of the transistor can normally be expressed as the magnitude of Z_{in}/Z_{out} without worrying about phase angle. The larger the value of R_{in}/R_{out} relative to X_{in}/X_{out}, the more accurate will be this approximation. But for narrow-band circuits – all circuits above 200MHz – matching networks using both the real and imaginary parts of the load impedance should be used (in this instance, the input/output of the amplifying transistor). In rf device data sheets, input/output impedances are usually given in complex form

Fig. 2. In circuit a), negative feedback is derived directly from the collector. Adjustment of the feedback voltage source with respect to the base is achieved in T_1 by providing a high impedance input point. In circuit b), a lower impedance point than the collector is created by adding a third winding in T_2. This allows the feedback voltage to be fed directly to the input of T_1. In c), the feedback voltage is fed to a low impedance point (the base), necessitating a low impedance voltage source. This is also achieved by adding a third winding in T_2.



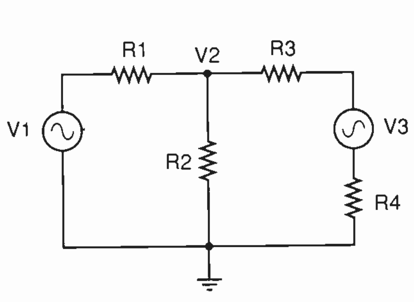


Fig. 3. Simplified model of a negative feedback network which can be used to determine the loop parameters with sufficient accuracy. Design of this model is based on a series RLC loop.

sheets. In many cases computer software can generate and optimise the elements of the matching network.

Negative feedback

One other impedance-compensation and gain-levelling method, with advantages and disadvantages compared to LCR networks, is negative feedback.

The technique involves feeding part of the output power, out-of-phase, back to the input. Part of the input voltage and the fed-back voltage then cancel. Advantages include simplicity and a stabilising effect on the amplifier, and the only disadvantage is that power is dissipated in the feedback network, lowering overall efficiency of the system.

Power lost depends on the amount of gain reduction desired at low frequencies, ie, the amount of feedback.

The out-of-phase feedback voltage is set to a certain amplitude, with respect to the input voltage, which holds at any power level – providing the input impedance remains constant. To produce power, the input voltage must exceed in amplitude the voltage fed back. As well as gain reduction, negative feedback also lowers the effective input impedance of the device(s). The device input impedance itself remains unchanged, but the out-of-phase voltage fed back to the input lowers the load impedance to the input matching element.

In a wide band amplifier, feedback voltage should be inversely proportional to the frequency and big enough so that the gain is reduced to the correct amount at all frequencies below the high-end of the band. With simple networks consisting only of R and L, where the feedback voltage source is the collector (or drain) of the output transistor and the voltage is fed back to the base (or gate) directly, this is not possible. An exception is low-power design, where impedance levels are relatively high.

A collector-to-base feedback circuit is illustrated in **Fig. 2a**. Note that the input impedance for the feedback voltage is set by T_1. The same kind of feedback with a lower impedance source results from adding a third winding in T_2 (**Fig. 2b**). Again feedback voltage is fed to the input through the primary of T_1, having a higher impedance level than the base. In **Fig. 2c**, a third winding is again added in T_2. It has a very low impedance since the feedback voltage goes directly to the base, itself having a low impedance.

Negative feedback to the base is the most commonly used arrangement with bjts since the base impedance is well defined, leaving only one variable – the third winding in T_2. Instead of T_2, the third winding for deriving the feedback voltage can be located in the collector/drain dc feed choke, sometimes giving a more convenient option because of its proximity to the input.

The voltage swing across the choke is equal to that across the output transformer. But its use as the source for feedback voltage increases flexibility of circuit design since its impedance ratio to the feedback winding is easily adjustable without affecting output matching of the transistor.

BJT or fet

Negative feedback loops for fets are easier to determine because the fet is a voltage-controlled device. With bjts, voltages must be converted to currents since base voltage variations are small and it would be difficult to achieve sufficient accuracy with calculations.

But models for negative feedback loops (**Fig. 3**) meant primarily for fet amplifiers, can also be used with bjts in modified form.

Figure 3 refers to the push-pull amplifier design given in **Fig. 4.** At 10MHz, the low-frequency end of this example, magnetic cores needed in the input and output transformers are not shown in the schematic for simplicity. In the model, the feedback voltage is derived directly from the fet drains, limiting optimisation of the system. A peak in power gain of about 2.5dB will remain around the middle of the 10-175MHz spectrum.

(If flatter gain response is required, methods shown in Figs. 2a and 2c are recommended).

There is considerable phase deviation from 180° at 175MHz, resulting from the series L. So there will also be about 1dB gain reduction at this frequency due to the finite reactances of the inductances.

At low frequencies, where feedback is at its maximum, the phase error is negligible and the model shown in Fig. 3 produces fairly accurate values. In the model, the series inductance used to shape the gain slope, has been omitted. This L can be treated as an additional variable – its value for the spectrum in question would probably be lower than the minimum achievable with the physical size of the circuitry.

Ideally reactance of the series L should be infinite at the high end of the spectrum and zero at the low end.

C_1 and C_2 in Fig. 4 are dc blocking capacitors. Their values are not critical, but they must be large enough to present a low reactance at the lowest frequency of operation.

Tr_1 and Tr_2 are assumed to be *MRF151* devices. But at 10-175MHz, these could be replaced with a single push-pull *MRF151G*. (The *MRF151G* is equivalent to two *MRF151*s in a single package, but tested to 175MHz specifications. The *MRF151* is tested at 30MHz, although is usable up to at least 175MHz).

Calculating feedback resistors

From the data sheet and by simple calculations, the nearest full integer impedance ratios of 9:1 (input transformer) and 1:4 (output) can be found to be the closest practical at 175MHz, with a 50Ω interface. For Fig. 3, from the data sheet, we can also deduct the following parameters:

G_{PS} at 10MHz is approx 26dB;
G_{PS} at 175MHz is approx 16dB (lowered to 15dB with feedback);
$P_{in}1$ (f = 10MHz, P_{out} = 300W) = 0.75W, V_{in}rms = 2.03V (V_2);
$P_{in}2$ (f = 175MHz, P_{out} = 300W) = 9.5W, V_{in}rms = 7.23V (V_1);
V_3 = V_{out}rms (drain to drain) = 61.25V;
R_1, R_2 (transformer source and gate-to-gate impedances) = 5.5Ω; R_3 = feedback resistor; R_4 = (output load) = 12.5Ω.

The value of the feedback resistor is given by:

$$R_3 = \frac{V_2+V_3}{\left(\frac{V_1-V_2}{R_1}\right)-\frac{V_2}{R_2}} - R_4$$

Substituting

$$\frac{2.03+61.25}{\left(\frac{7.23-2.03}{5.5}\right)-\frac{2.03}{5.5}}-12.5$$

$$=\frac{96.6}{2} \text{ or } 48.3\Omega \text{ each resistor.}$$

= 96.6Ω/2 or 48.3Ω each resistor.

Total power dissipated in the feedback resistors at the low-frequency end of the spectrum of operation, the worst case, is

$$(V_2+V_3)\times\left[\left(\frac{V_1-V_2}{R_1}\right)-\frac{V_2}{R_2}\right]$$

or 63.28 x 0.58 = 36.7W, 18.35W per resistor, though this assumes that the series L has zero reactance. (No simple formulas are available to calculate the values of R and series L versus frequency response, though some computer programs can plot the amplifier's response characteristics for given values of these elements.)

Any series reactance would be treated as added series R at a given frequency, and deducted from its original value. Since there will be a voltage drop across the reactance, the voltage across R will be lower, resulting in reduced dissipation.

At the low-frequency end, the series L is customarily selected with its reactance approximately equal to the input impedance of the device. In this case that value is 80nH (5Ω) at 10MHz and phase delay is negligible. At 175MHz, the same inductance represents a reactance of 90W resulting in a phase delay of about 15°. This is normal and only becomes dangerous if 180° is approached, causing the feedback to turn positive in phase and create instabilities. Such a phase shift could occur only if the initial value of L is unnecessarily high or if the amplifier bandwidth is 7-8 octaves or more – possible in certain low power designs. The Q value of the series L, already reduced by the series R, X/R, can be further controlled with a parallel R (R/X).

In practice typical Q values for the series L are less than ten in most cases.

Power loss

These examples show that power loss at low frequencies is considerable, in this case amounting to 6-7% of overall efficiency. The feedback resistor values can be rounded to 50Ω, and the reactance of the series L is 5Ω, but the dissipation factor of R is reduced only by 10%.

Recalculating, using the formula above, gives dissipation figures of: (63.28–6.33) x 0.58 = 32.5W or 16.75W for R_1 and R_2 of the push-pull amplifier in Fig. 4.

Notice that at the low-frequency end of the amplifier's frequency band, adding the series L causes minimal change in efficiency. But at the high end (175MHz), the effective value of the feedback resistor is increased from 50W to 140W. Including the phase delay, this results only in an approximate 1dB gain loss. If the loss of efficiency with negative feedback is not acceptable in an application, a combination of RLC compensation and negative feedback (Figs. 1 and 2) usually yields excellent results. ■

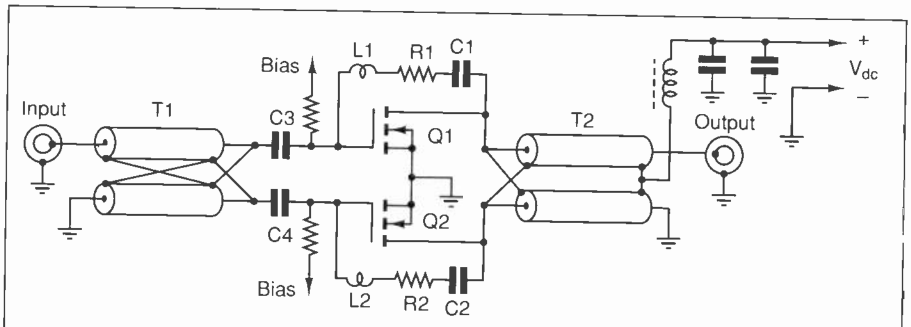
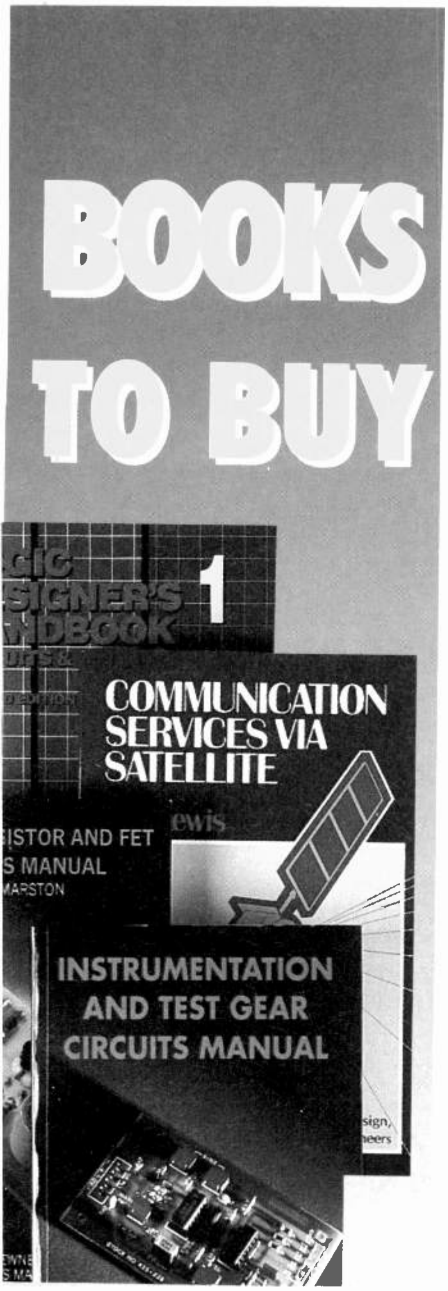


Fig. 4. A fet rf push-pull amplifier with negative feedback. The component values for RLC feedback networks can be established with the model of Fig. 3.



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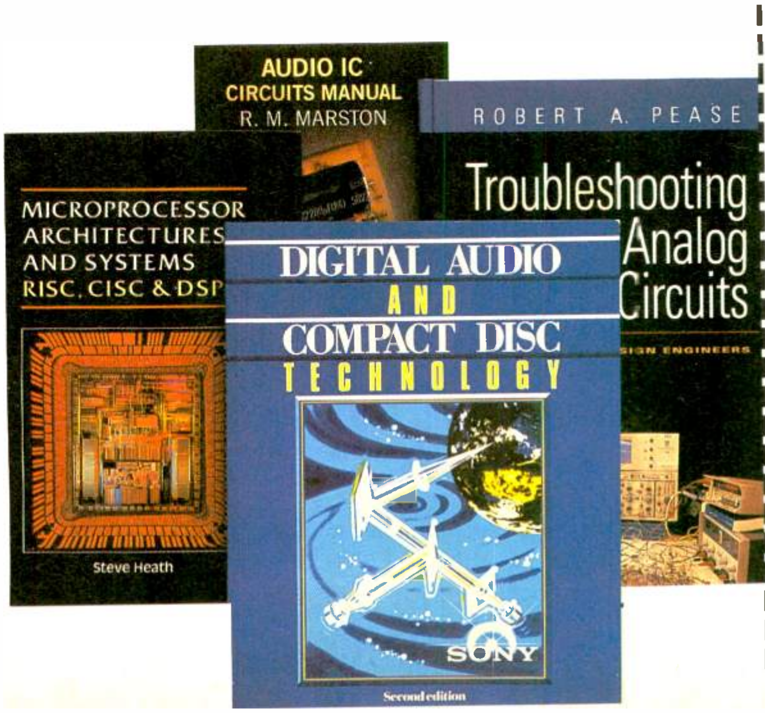

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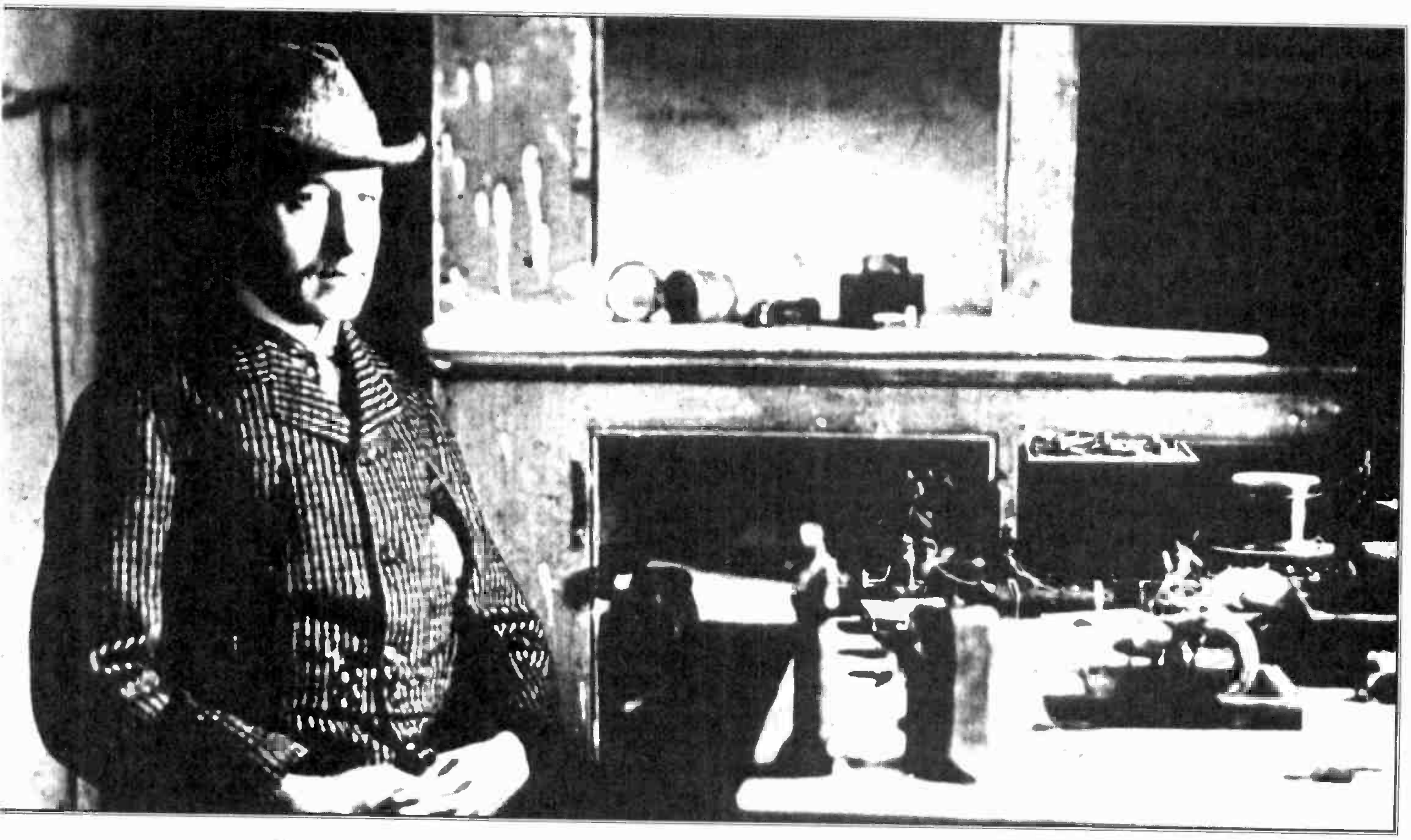
the MAN

Signal Hill, St Johns, Newfoundland, with apparatus used in the first transatlantic wireless tests. Marconi knew he was playing for big stakes, both commercially and in terms of his scientific reputation. (Courtesy GEC-Marconi Ltd).



who started ripples in the ether

In 1894 Marconi began the research that enabled the world to take the first steps towards modern communications. 100 years on, John Powell Riley pinpoints the key moments in a genius's life and looks at exactly what Marconi risked in his greatest experiment.

No single person invented the radio. But the pioneer towering above all those brilliant people who made essential contributions is, without question, Guglielmo Marconi.

Marconi began his experiments at a time when no one knew how radio waves propagated. Range was generally thought to be limited to about the distance a man could throw a stone. Today we understand that the theoretical basis of wave propagation is vastly complicated. But in those early days what was needed was an experimenter with a unique intuitive and inventive gift. Marconi was that man.

He was born on the 25th of April, 1874 in Bologna, Italy. His Irish mother, Annie Jameson, was a member of the Jameson whiskey family and his father, Giuseppe Marconi, was a comfortably well-off, land-owning Italian.

At the age of five, Marconi came to England for his first two years of schooling, going back to Italy to complete his education in Livorno, where he later studied physics. One of his earliest influences was the eminent Professor Rigi whose work on electromagnetic radiation first began to capture his imagination. At that time, his aim was to become an officer in the Italian navy. But his ambition changed when he failed the entrance examination.

Marconi's awakening came when, vacationing in the Italian Alps in the summer of 1894, he read a paper describing Heinrich Rudolph Hertz's laboratory experiments with "Hertzian waves". Immediately he returned home, and set up a lab in the attic of Villa Grifone, the family home in Pontecchio near Bologna.

From that moment, single-minded fascination with the subject shut out all else and wave propagation became his lifelong obsession.

Encouraged by his mother – and strongly opposed by his father who thought he should devote his energies to improving the grape harvest – Marconi nevertheless steadily increased the distance over which messages could be sent "through the ether".

Within a year he was able to transmit morse signals over distances of about a kilometre, to positions hidden beyond the crest of a mountain.

England beckons

Starting with nothing more than a laboratory experiment, Marconi had proved that practical wireless communication was feasible, an achievement that marked the advent of radio and profoundly changed our lives.

But with astounding lack of vision the Italian government, like Marconi's father, showed no interest in his accomplishments. England, then the greatest maritime nation in the world, appeared to be a better place to further his ambition and at the urging, both of his mother and the Irish branch of the family, the reserved, self-confident young man of 21 moved there with his mother.

Helped by his cousin Jameson Davis, a man of influence who lived in London, he applied for his first patent on June 2, 1896 at the age of 22. On March 2, 1897 he filed an improved specification and on July 2, 1897 the London Patent Office granted the young man his patent. It was the first in any country for a "system of telegraphy without wires by means of electromagnetic waves".

By July 1897, again with the help of his cousin, Marconi had formed the Marconi Company (originally the Wireless Telegraph and Signal Company, Limited; later Marconi's Wireless Telegraph Company; now GEC-Marconi Limited). He had also increased reception range to 7-9 miles over water.

Competitive tension

The young Marconi deeply impressed William Preece, chief engineer of the Post Office, the licensing authority for communications. Preece had been experimenting with electromagnetic waves but had failed where Marconi succeeded. In a moment of candour he praised Marconi and referred to him as "the boy wonder." Though intended as a sincere compliment, the remark was promptly seized upon by mature physicists and engineers who were embarrassed at being unable to explain Marconi's accomplishments in terms of known science.

Over the years Preece alternately praised Marconi and joined the ranks of his critics. He strongly supported the cable companies that were operating under license agreements granting them exclusive rights to certain important point-to-point links. Cable companies quickly recognised the threat posed by "wireless" and emerged as bitter and often unprincipled rivals. Cable couldn't compete in maritime applications and that afforded Marconi a foot-hold in a slowly growing but conservative captive market – together with another arch rival, Telefunken, in Germany. Professor Adolphus Slaby, of the Technical High School at Charlottenburg, Berlin (also associated with Siemens & Halske and Telefunken) saw himself as a colleague but Marconi regarded him as his *bête noire*. The Marconi Company continued to make technical strides at an almost unbelievable pace. German rivals, under Slaby's technical guidance, struggled to keep up, but whenever Marconi got too far ahead, the German government would contact the British government through diplomatic channels and request that Professor Slaby be permitted to attend a Marconi demonstration as an interested observer. In this way the obsequious Slaby conducted what Marconi considered to be commercial espionage and on return to Germany promptly upgraded the Telefunken equipment. In fairness to Slaby, one of his technical tips to Marconi, concerning matching the impedance of the coherer to that of the antenna, proved invaluable.

The British government was also slow to see the significance of radio in Naval communication. The application of wireless to the safety of life at sea seemed to Marconi to be obvious and it strongly motivated him from the start.

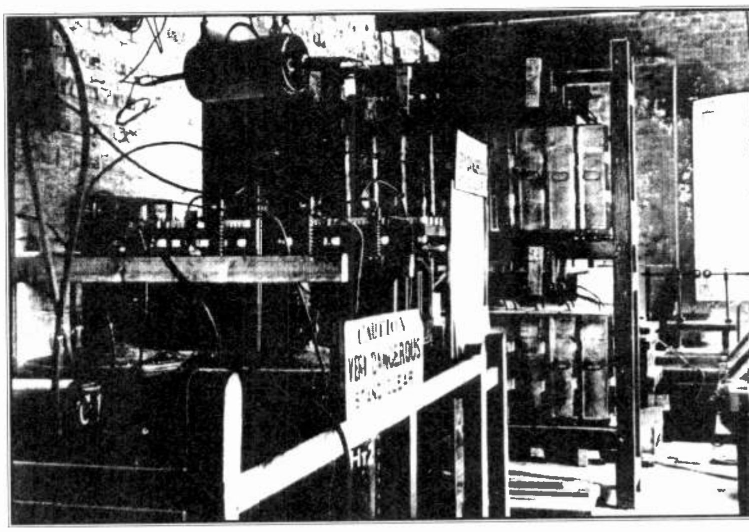
Today we take radio for granted. When Nasa put men on the moon, few stopped to reflect on the other scientific miracle that allowed us to talk to them.

Calling Newfoundland

Marconi's work inevitably brought criticism from the establishment (see box). To tackle this – and for a much more rational commercial reason – Marconi knew he must attempt a transatlantic test as soon as possible. This was the best way, ultimately, to break the monopoly on international communications, held by the cable companies.

To attempt it, Marconi needed a much more powerful transmitter, a more elaborate antenna, and a better station site. But at that time, the Marconi Company was hard pressed to pay the ever-mounting cost of research and development. Business was not good and a transatlantic test called for elaborate preparations which would be an almost unacceptable financial burden.

> *"I was about to test the truth of my theories, to prove that the patents that the Marconi Companies and myself had taken, and the tens of thousands of pounds which had been spent in experimenting and in the construction of the great station at Poldhu, had not been in vain."*
>
> *G Marconi*



Apparatus at Poldhu in Cornwall that transmitted the "s" received by Marconi in Newfoundland. The spark gap can be seen, right, in the background. (Courtesy GEC-Marconi Ltd).

Marconi met with his board of directors and, using all his powers of persuasion, eventually obtained their less than enthusiastic support for a risky attempt to bridge the Atlantic.

Moving fast, he selected an ideal new station site called Poldhu, in Cornwall near Mullion – about six miles north of Lizard Point. The inventor described it with poetic imagery, as "hard and bleak" and said it "possessed an inexpressible charm because of the soft airs, pungent with salt, that blew over it".

Building construction started in October of 1900. Marconi was determined that Poldhu would have the most powerful transmitter yet constructed and its antenna would be equally formidable. Critics now began to refer to his new more powerful transmitting station as "the thunder of Poldhu." They predicted it was going to drown out the smaller local stations.

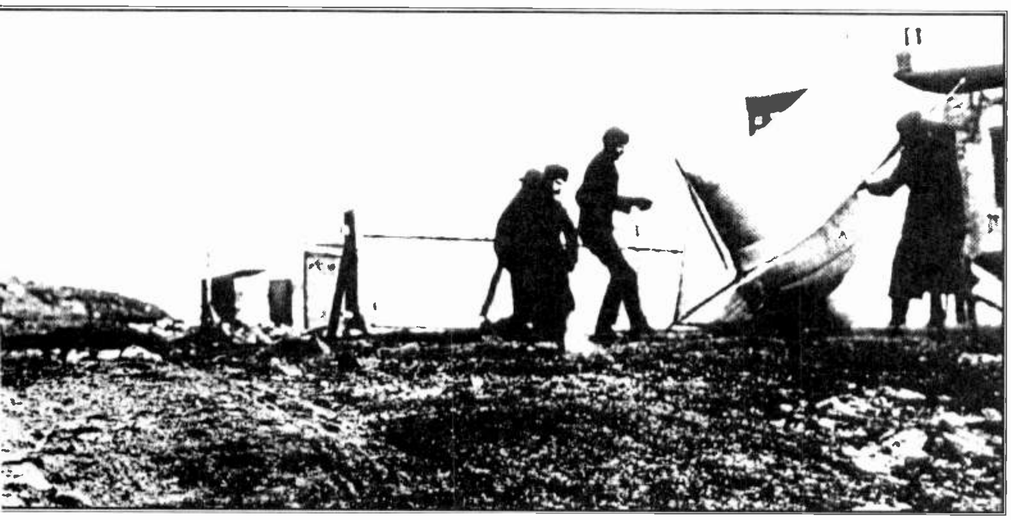
Certainly, a major problem had to be overcome. Spark-gap transmitters emitted a tremendously broad power spectrum so that a receiver placed near one would pick up signals no matter to what it was tuned. The crude but remarkably sensitive "coherer" detectors used as receivers also lacked selectivity.

In December 1990, Marconi moved a station from Dovercourt, near Harwich, to Poldhu to start transmission tests. On January 23, 1901 the station at Poldhu easily established communication with one at St Catherine's Point, Isle of Wight, 196 miles distant. This was well beyond line of sight and Marconi saw it as proof that earth curvature would not be an obstacle on the longer transatlantic path.

Media scepticism

Some elements of the technical press in England, especially at the start of his career, were devastatingly cruel to Marconi. It nettled him all his life. When transmission range was only a few miles, they said "the usefulness of radio was obviously very limited." Later, when greater distances were spanned, they said that the five word per minute speed of transmission was too slow. It was in fact 22 words per minute.

When Marconi again extended the range so that ships at sea could receive messages and lives were saved by radio, the critics still hounded him. One, writing for a technical trade paper, claimed that to transmit at a speed of 50 words per minute, twenty gigantic stations – ten at each end of the path – would need to be constructed.



Balloons failed, but Marconi was able to use a kite to raise his aerial high enough in Newfoundland for him to receive the transmission from England. (Courtesy GEC-Marconi Ltd).

Dr J A Fleming, a consultant to Marconi and a pioneer in his own right, made many improvements at the transmitter site, greatly increasing power during the next four months, while Marconi turned his attention primarily to his newly-developed tuned circuits (see box), adding them both to the transmitter and to the receiver to avoid interference with other services. The result, according to his notes, "was a transmitter more powerful than anything of the kind, until then, constructed". (10 to 12kW).

Transatlantic wireless telegraphy was at last feasible in Marconi's opinion. Furthermore it was attractive as a commercial proposition, and tests proved that nearby stations now had nothing to fear from the "thunder" of Poldhu.

Line of vision

As preparations raced ahead, there was no let up from those in the scientific community who lacked Marconi's vision. A favourite theme now was that very long range communication was not possible because of the curvature of the earth. There was, they pointed out, a "hump of water more than 100 miles high" that radio waves could neither bend around nor penetrate to reach a distant point below the horizon on the other side of the Atlantic. One facetious argument was that transatlantic wireless transmission would only be possible if antennas at both ends of the path were supported on masts two hundred miles high.

Marconi had convinced himself they were wrong by successfully communicating over lesser distances that were nevertheless well beyond line of sight. In doing this he had become aware of what now is called the ground wave, following the surface of the earth and propagating particularly well over salt water.

By trial and error Marconi had found that vertically polarised antennas working against ground, what we today call Marconi antennas, accentuated this propagation mode. As wavelength increases, ground wave attenuation decreases.

In fact this preoccupation with ground wave propagation misled experimenters, Marconi included, into believing that long distance transmission required long wavelengths of the order of 360 to 2000m – and consequently gigantic antennas.

Eventually they would learn that there are also sky wave modes of propagation (see box) even better suited to long distance communication.

Disaster

By the end of August, 1901, an antenna at Poldhu consisting of twenty poles, 220 feet tall, was erected on a circle of 200 feet diameter, with their tops connected by triatic stays to which the many semi-vertical elements were fastened. Work was nearly completed when, on September 17th, the worst storm in memory swept the English coast. A triatic stay snapped and the poles fell to the ground.

Directors of the Marconi Company were appalled at the disaster and at the funds

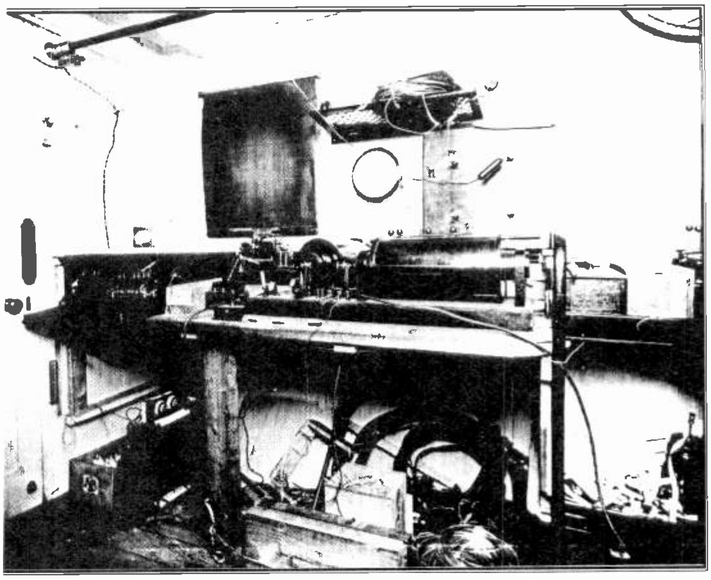
Genius in the Edison mould

Marconi was a wizard in the sense that Thomas Edison was. Both were single-minded men whose intense determination and intuitiveness set them apart, and fitted them ideally for their empirical approaches to science. Both lived, most of the time, in the intensely fascinating dream worlds known only to those whose immense creativity requires total concentration. As a result they were often oblivious to all else.

Edison left his bride on their honeymoon, got off the train and went back to his lab. Similarly Marconi, conducting radio experiments on his yacht, the Elettra, took his youngest daughter Gioia to the nearest beach, left her alone to play – and forgot her. She tried to dog-paddle back to the yacht, and it was only a miracle that allowed her to be found before she drowned.

expended with nothing to show for it, but Marconi only redoubled his efforts. He did wonder though if the site was a mistake – the coast at Mullion was highly vulnerable to violent storms.

At first the delay looked like months, but Marconi was at his best when challenged by adversity and he had erected a simpler but nevertheless effective antenna within eight days. This one used two 160ft poles 200ft apart to support a triatic stay, stretched between them, from which he suspended sixty copper wires. At the top they were spaced at 3ft intervals and at the bottom they came together for connection to the transmitter.



Critics claimed Marconi could have staged his transatlantic triumph, and questioned whether he could be believed. To confound them Marconi repeated the feat using a wireless installation aboard the SS Philadelphia, inviting in many witnesses to observe reception. (Courtesy GEC-Marconi Ltd).

Using this "aerial" in mid-November the Poldhu station sent such strong signals to Marconi's most distant station at Crookhaven in South Ireland that he felt confident they could be detected at a ten times greater distance. The experiment again reinforced his conviction that the curvature of the earth was not going to be a problem.

Covert activity

At last Marconi felt ready to attempt the transatlantic test. He decided to site the receiving station in Newfoundland – that being the nearest point of land in North America. As the test was such a dramatic extension of range he thought it best to avoid publicising his plans: if he succeeded, the effect would be all the more dramatic.

The reason he gave for visiting Newfoundland was that he wanted to investigate the possibility of signalling to Cunard liners on Atlantic passage to beyond the Grand Banks, 400 miles off the Newfoundland coast. Before departing England, Marconi told the chief operator at Poldhu to expect a cable. On the day following receipt of that cable, Poldhu was to send the letter "*s*" (three dots) at regular intervals daily between 3pm and 6pm Greenwich time until told to stop. The corresponding time at St John's, Newfoundland, would be 11:30 am to 2:30pm.

We now realise that the schedule he chose, based merely on convenience, was almost disastrous. During daylight hours solar absorption, then unknown, greatly diminishes the strength of long wave sky waves. Transmissions on the wavelength used should actually have been scheduled during hours of darkness when sky waves, unattenuated by solar absorption would have been far stronger than ground wave. But having made these arrangements, Marconi sailed on November 26th from Liverpool on the Allan liner SS Sardinian with assistants, George Stevens Kemp and P W Paget.

Not mistaken

After an uneventful crossing, the ship docked at St John's, Newfoundland on Friday, December 6, 1901. The next day, Marconi visited the Governor, Sir Cavendish Boyle, the Premier, Sir Robert Bond and other members of the Ministry. Everyone promised wholehearted cooperation and went all-out to help him find radio station sites at Cape Race, Mistaken Point, and other likely places. All sites were rejected by Marconi – perhaps he worried how a site near "Mistaken Point" would be exploited by a hostile and sceptical home technical press.

He eventually settled on Signal Hill near the city of St John's, in his notes described as: "A lofty eminence overlooking the port and forming the natural bulwark which protects it from the fury of the Atlantic gales".

Because of inclement winter weather and time considerations, Marconi decided not to erect high poles to support the aerial. Instead he tried to use a captive balloon. He had brought two for that purpose as well as six kites as back-ups.

The top of Signal Hill has a small plateau of about two acres which he thought suitable for handling either the balloons or the kites. From a crag on this plateau rose the new Cabot Memorial Tower, designed as a signal station commemorating the discovery of Newfoundland. Close to it there was an old military barracks which was then used as a hospital. He set up his equipment in this building and prepared for the great experiment.

After wrestling with the balloon then turning to the kites to elevate the aerial to 400ft he at last heard the "three sharp little clicks corresponding to three dots" that indicated the passage of the *s* transmission, across the Atlantic

Marconi's "sternest test"

Marconi describes preparations – and what was at stake – for his test transmission from Cornwall to Newfoundland:

"On Monday, December 9th, barely three days after my arrival, I began work on Signal Hill together with my assistants. I had decided to try one of the balloons first as a means of elevating the aerial and by Wednesday we had inflated it, and it made its first ascent during the morning. Its diameter was about fourteen feet and it contained some 1000 cubic feet of hydrogen gas, quite sufficient to hold the aerial which consisted of [500 feet of] wire weighing about ten pounds. Owing, however, to the heavy wind that was blowing at the time, after a short while the balloon broke away and disappeared to parts unknown. I came to the conclusion that perhaps the kites would answer better and on Thursday morning, in spite of the furious gale that was blowing, we managed to elevate one... to a height of about four hundred feet".

Marconi went on to describe why it was so vital that his experiment was a success:

"I was about to test the truth of my theories, to prove that the patents that the Marconi Companies and myself had taken, and the tens of thousands of pounds which had been spent in experimenting and in the construction of the great station at Poldhu, had not been in vain.

"In view of the importance of all that was at stake, I had decided not to trust to the usual arrangement of having the coherer signals recorded automatically through a relay and a morse instrument on paper tape, but to use instead a telephone connected to a self-acting coherer, the human ear being more sensitive than the above-mentioned recorder. Suddenly, about half past twelve there sounded the sharp click of the "tapper" as it struck the coherer, showing me that something was coming and I listened intently.

"Unmistakably, the three sharp little clicks corresponding to three dots, sounded several times in my ear but I would not be satisfied without corroboration.

"Can you hear anything, Mr. Kemp? I said, handing the telephone to my assistant. Kemp heard the same thing as I and I knew then that I had been absolutely right in my calculations. The electric waves which were being sent out from Poldhu had traversed the Atlantic, serenely ignoring the curvature of the earth which so many doubters considered would be a fatal obstacle, and they were now affecting my receiver in Newfoundland. I knew that the day on which I should be able to send full messages without wires or cables across the Atlantic was not very far distant. The distance had been overcome and further development of the sending and receiving apparatus was all that was required."

George Stevens Kemp (left) and P W Paget (right) sailed with Marconi (centre) to Newfoundland to help Marconi set up his most ambitious experiment.
(Courtesy GEC-Marconi Ltd)

from Podhu (see box: Marconi's own words). Marconi wireless had conquered the Atlantic and silenced, for the moment, some of his critics.

With a mixture of exultation, and trepidation Marconi sent a cable after three days of testing, having barely managed to hear the letter *s* twenty eight times. The news created a sensation around the world with some elements of the media jubilant and others outraged.

Messages of congratulation poured in from many nations, but the response from England was mixed. The consensus of the technical press there was one total scepticism. Almost immediately though, Marconi received a cable from a lawyer representing the Anglo American Telegraph Company. It accused him of violating the exclusive right – that the cable company had under its contract – to send communications from England to Newfoundland. The company threatened a lawsuit if Marconi did not desist immediately. The move, Marconi noted, was a tacit admittance that they believed the signals had been received.

In marked contrast, the reaction in America was warmly congratulatory. The prestigious American IEE (Institute of Electrical Engineers, now the IEEE) gave a dinner at the Astoria Hotel in New York in his honour attended by 300 members including many of our most distinguished men of science. Its President was the famous Steinmetz, mathematician and electrical genius of General Electric who held 200 patents. Alexander Graham Bell, inventor of the telephone was there. He conceivably might have seen wireless as a competitor, but instead he lavishly praised Marconi, and offered him the use of his estate at Cape Breton, Nova Scotia for the erection of a wireless station (Marconi declined). Thomas Edison and Nicola Tesla who could not be there both sent messages of sincere congratulations.

It was a heady experience for a young man of 27.

Some critics in England, however, were still unconvinced. They claimed Marconi was either perpetrating a hoax or was confusing normal atmospherics with test signals, citing the "immutable and well understood laws of physics" as their authority. Some were quick to point out that the world had only Marconi's word for what really happened. There were no ink recordings to document his claims.

Results confirmed

Stung by insinuations, Marconi sailed for New York aboard the SS Philadelphia, with C S Franklin as his wireless operator. (Franklin later distinguished himself as a prolific inventor. One of his many designs, the Franklin array, a directional antenna, came into wide use by point-to-point services.) The ship departed Cherbourg on February 22, 1902 and was fitted with equipment of the same type Marconi used in Newfoundland.

During this voyage he received complete messages to a range of 1551miles and they were automatically recorded by a morse inker in the presence of the ship's master, Captain Mills, other officers, and many interested passengers.

Each ink recording was signed and dated by one of these witnesses, usually by the captain. When range would no longer permit the reception of complete texts, Marconi continued to record parts of messages or the letter *s* until 2099 miles distant from Poldhu.

During his passage on the SS Philadelphia, Marconi made another discovery of fundamental importance. It was that long wave radio signals traverse greater distances over areas of the earth that are in darkness (due to the solar absorption phenomenon mentioned earlier). Without understanding it at the time, he observed the differences between ground wave transmission giving a steady signal day and night but attenuating rapidly with distance; and sky wave transmission, traversing vast distances with less attenuation if the time of day is right. Most critics were confounded and silenced by the shipboard demonstrations, but a few diehards thought that Marconi's latest feat was accomplished by relaying signals from ship to ship in a chain spanning the Atlantic ocean. Even today some radio engineers question the authenticity of the first transatlantic transmission, Poldhu to St Johns, (because of high ground wave attenuation at that range and also because of high solar absorption of sky waves at the 366m wavelength used).

But no one can deny his triumphant demonstrations conducted on the SS Philadelphia in February, 1902, at a time when his competitors were able to span only about one tenth that range. He had finally documented long range wireless transmission and silenced most shortsighted critics.

Of course, successful transmission of sentence fragments and the letter *s* across the Atlantic fell far short of what was required for a marketable service. But after many setbacks, and five years of further development, regular commercial service finally began between a Marconi station in Clifden, Ireland and one in Glace Bay, Newfoundland, on 15 October, 1907.

The monopoly of the submarine cable services was broken and the genie was out of the bottle. ■

In tune with coherer technology

The coherer was invented by the Frenchman E Branly, based on the discoveries of the Anglo-American D E Hughes, then further updated by the Englishman Sir Oliver Lodge as well as the Russian A S Popoff. The device consisted of a tube filled with fine metal filings – Marconi used 95% nickel and 5% silver. When rf energy passed through them, the particles cohered and resistance dropped. To respond to changes of state (signal or no signal), some form of "decoherer" such as a solenoid tapper or a mechanical shaker was necessary.

The clumsy approach limited the speed of morse transmission until the advent of the vacuum tube. More importantly, their lack of selectivity and the fact that spark gap transmitters were "broad as a barn door" had begun to turn the spectrum into a bedlam.

The further development of radio would not have been possible without a solution to the problem. Marconi found it by using tuned circuits to add selectivity both to transmitters and to receivers.

Sir Oliver Lodge had experimented earlier with tuned circuits but failed to reduce his ideas to practice as required by patent law. The improvement permitted more than one pair of stations to operate in the same area at the same time. The innovation became famous as "the Marconi four sevens patent" (No 7777).

Acknowledgment

Thanks to Roy Rodwell of GEC-Marconi Ltd for kind assistance in verifying certain details.

CIRCUIT IDEAS

SEND YOUR CIRCUIT IDEAS TO THE EDITOR, ELECTRONICS WORLD, QUADRANT HOUSE, THE QUADRANT, SUTTON, SURREY SM2 5AS

Do you have an original circuit idea for publication? We are giving £100 cash for the month's top design. Other authors will receive £25 cash for each circuit idea published. We are looking for ingenuity in the use of modern components.

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Capacitive fluid-level detector

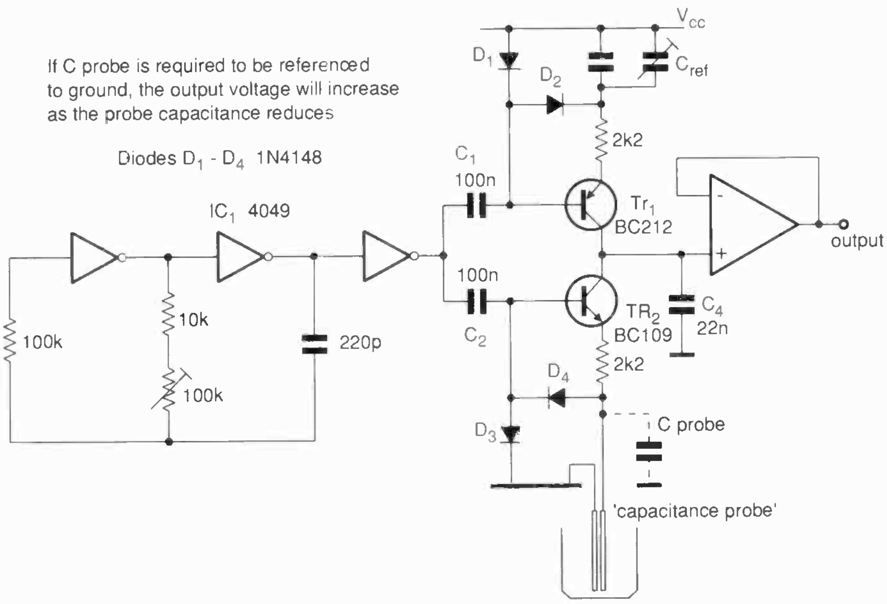
An output voltage proportional to a capacitance indicates the level of a fluid – in the original, oil.

The three buffers form a square-wave generator and driver. When the square wave goes high, C_{ref} charges positively with respect to V_{cc} via D_2 and C_1. When it is low, C_{probe} charges negatively via C_2 and D_4.

Transistors $Tr_{1,2}$ conduct alternately, so that the voltage on C_4 and therefore the output depends on the values of C_{ref} and C_{probe}, which is simply two parallel or concentric conductors with a small air gap. Each of the capacitors charges or discharges C_4.

Oscillator frequency may be adjusted to give the required output excursion for a given change in C_{probe} and the trimmer in C_{probe} sets the output voltage. The two capacitors may be transposed to reverse the output voltage.

Glyn Roberts
Walsall
West Midlands



Capacitive level detector provides analogue indication.

Micropower logic coupler

It is common to electrically isolate peripherals from logic circuitry by means of optical couplers. They work well, but suffer from the drawback that one coupler needs about 50mW of power. This inductive coupler avoids the problem.

The cmos square-wave generator in Fig. 1 drives tank circuit L_2C_1 and loss resistance R_p. In the absence of an input, Tr_1 is open and the square wave developed across the tank is rectified by the diode to produce a logic 0 at the Schmitt buffer output. A logic 1 at the input turns Tr_1 on, shorting the tank, and the output is logic 1.

For a frequency of 500kHz, C_1 330pF and Q 120, R_p is given by

$$R_p = \frac{Q}{2\pi f C_1} = 116\text{k}\Omega.$$

Resistor R_1 in parallel with R_p gives about 100kΩ, which results in a required power of 22.5μW. Increasing the frequency calls for several hundred microwatts and the use of 10.7MHz filter coils.

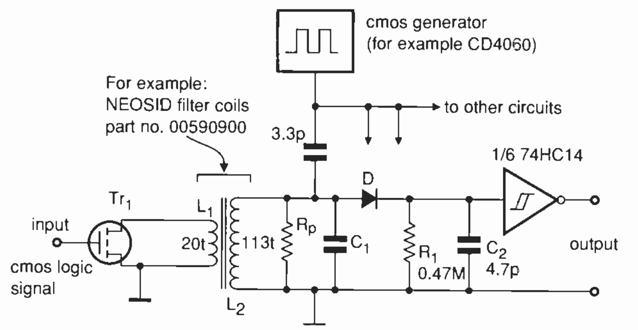


Fig. 1. While providing effective logic isolation, this inductive circuit consumes about 2000× less power than an opto-isolator.

Even simpler, in the circuit of Fig. 2, a diode replaces the fet, passing or blocking the input from the cmos gate. A logic 0 at the input to the diode produces 1 at the output of the buffer.

Franz Braunschmid
Vienna
Austria

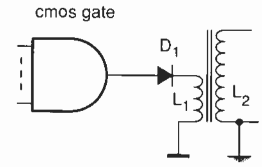


Fig. 2. An alternative to the mosfet switch in Fig. 1.

Switched-mode, constant-current charger

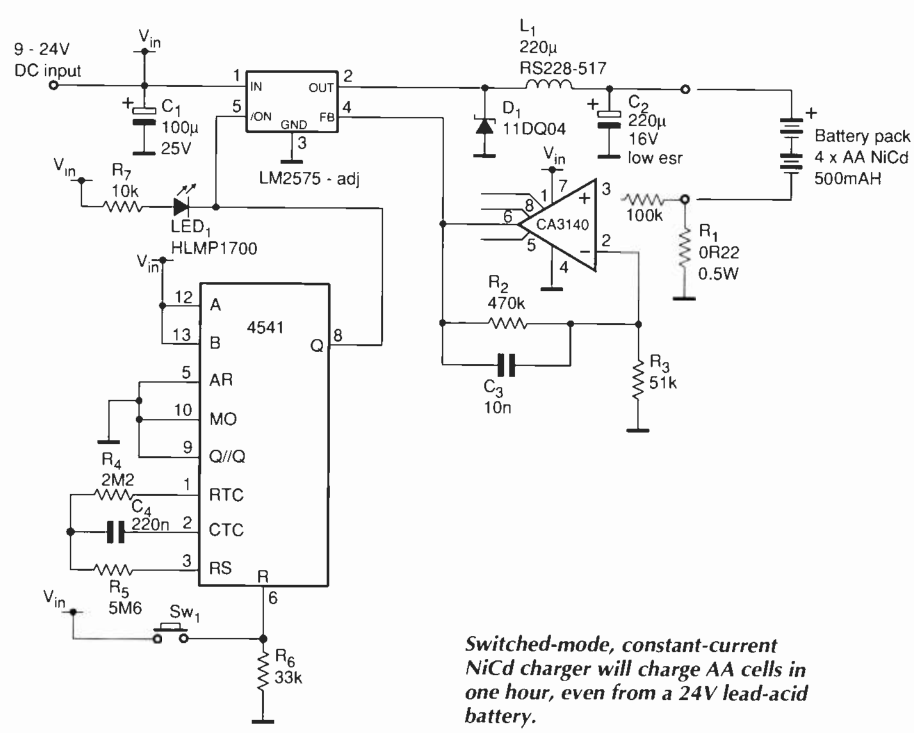
A common method of fast charging NiCd batteries is to inject a constant current equal to the battery capacity – one hour at 500mA for a 500mAh battery. Monitoring tests for a reduction in terminal voltage to indicate overcharging and an increase in the rate of temperature rise. This type of charger is often built into portable equipment, in which it is a simple matter to measure temperature and where the battery consists of a fixed pack. When used to charge batteries from other equipment, results can be erratic. Additionally, there is the power dissipation in the series-pass element usually used, which has to cater for the nearly 2:1 rise in terminal voltage as charging approaches completion.

This circuit uses a switched-mode series element to increase efficiency; in this case it is a National Semiconductor 1A *LM2575* voltage regulator with current-derived feedback. Constant current flowing through R_1 produces a 0.12V drop, which is amplified by a factor of 10 and applied as feedback to the regulator, so that a varying output from the regulator appears across C_2 to maintain the current through R_1 and the battery. The integrating capacitor C_3 slows down the regulator's internal control loop.

With values shown the circuit generates a constant current of 520mA, which can be altered by adjusting the value of R_3, perhaps by switching various resistors to obtain a programmed charger. For higher charge rates, a 3A version of the regulator, the LM2576-ADJ, can be used, with up-rated D_1 and L_1.

No heatsink is needed for charge currents in excess of 1A and AA cells charge within an hour. The 4541 timer disables the regulator after one hour, this interval being set by C_4, R_4 and R_5 and Sw_1 providing a reset and another hour's charging for NiMH or larger NiCd cells. The led indicates charging taking place.

Huw Jones
Gyrus Medical Ltd.
St Mellons
Cardiff



Switched-mode, constant-current NiCd charger will charge AA cells in one hour, even from a 24V lead-acid battery.

VFO uses a single current-mode IC

Using only one IC, two capacitors and four resistors, this variable-frequency oscillator uses one element to provide independent control of frequency and amplitude. The IC in question is a Phototronics *PA630* current-mode amplifier consisting of a second-generation current conveyor and two buffers.

Figure 1 shows the arrangement, in which the frequency is:

$$\omega_0 = \omega_x \sqrt{1-v} \quad (0 \le v \le 1),$$

and the condition for oscillation is:

$$k_0 = 1 + \frac{C_2}{C_1}\left(1 + \frac{R_2}{R_1}\right)$$

where,

$$\omega_x = \left(R_1 R_2 C_1 C_2\right)^{-1/2},$$

and,

$$k_o = 1 + R_F / R_G$$

Therefore, frequency control is totally independent of the condition for oscillation, which may also be set by R_F or R_G. Since the current conveyor and output buffer form a current-feedback amplifier, amplitude stability is good.

Taking into account amplifier characteristics, the non-inverting amplifier's transfer characteristic is

$$k_{(s)} = \frac{k_0}{1 + sR_F C_T}$$

where $R_F C_T$ is τ, the amplifier time constant and C_T is the internal transcapacitance. There is no gain/bandwidth trade-off.

The earlier equations now become

$$\omega = \omega_0\left(1 + \tau\omega_x k_0\right)^{-1/2}$$

and

$$k = k_0 + \tau\omega_x\left(1 - \frac{\omega^2}{\omega_x^2}\right)$$

The theoretical absence of slew-rate limiting in the current-feedback amplifier makes for higher operating frequencies, larger amplitudes and less distortion; frequency being only limited by parasitic capacitance. In an op-amp circuit, the same configuration would be limited to an upper frequency of about 16% of the gain/bandwidth of the op-amp. Again, large component values are unnecessary for LF oscillators.

Figure 2 shows the result of performance measurements, using a *PA630* with ±10V, a symmetrical passive network and R_F of 470Ω. The amplifier time constant was found to be 14.5ns. Distortion is less than –35dB. The saturation of the output buffer affords amplitude limiting, but a temperature-sensitive R_G would be desirable.

Santiago Celma
Pedro Martinez
University of Zaragoza
Spain

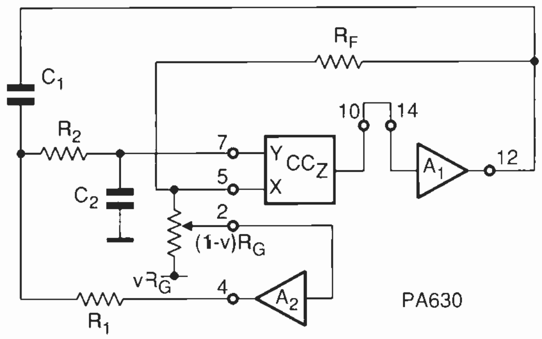


Fig. 1. Current-feedback amplifier circuit allows independent control of amplitude and frequency.

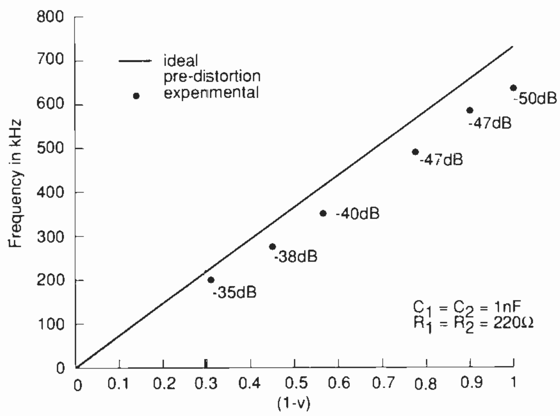


Fig. 2. Results of testing agree closely with predicted characteristics. Output frequency is shown plotted against

Power isolator

In the event of a power supply attempting to impose overvoltage on the load, or the load trying to draw too much current, this circuit arrangement isolates the load and indicates the fault condition by led.

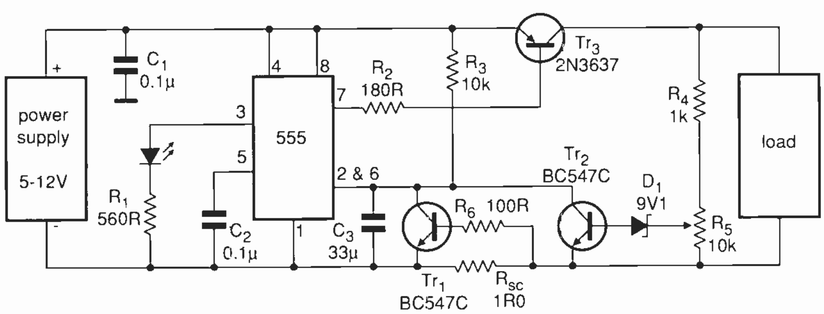
In normal operation, Tr_1 and Tr_2 are off, the *555* is reset and the *555* discharge transistor is on and draws base current to saturate Tr_3, passing current to the load.

If the load demands excess current, the voltage drop across R_{sc} turns Tr_1 on and triggers the monostable formed by the *555*, turning off its discharge transistor and therefore Tr_3, which isolates the load. The monostable times out and retriggers continually so long as the fault remains.

To protect the load against overvoltage, D_1, R_4 and R_5 present a fraction of the load voltage, determined by R_5, to Tr_2. If this voltage exceeds the set limit V_T, Tr_2 conducts, triggers the monostable and again switches off Tr_3. In either case, the led lights.

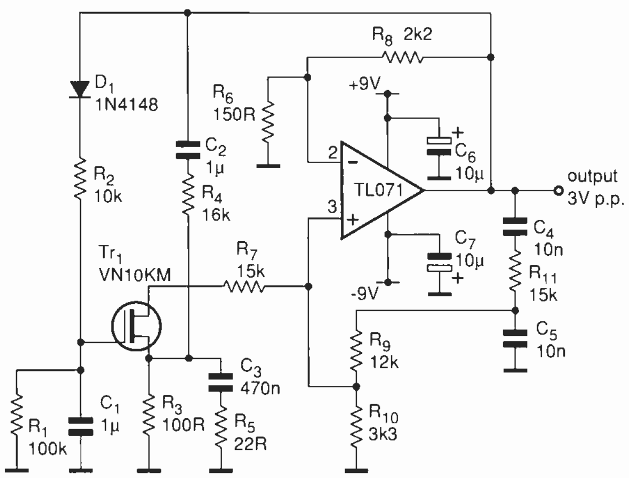
The series pass transistor Tr_3 is either on or off and therefore need only be a medium-power type. Voltage of the zener D_1 should be $(V_z+0.7) \approx 0.8V_T$, V_T being in the 5.5V-15V range. The sense resistor R_{sc} is $0.7/I_T$, I_T being the trip current.

M S Nagaraj
ISRO Satellite Centre
Bangalore
India



If either the power supply or load attempt to impose unreasonabl e demands on the other, this circuit breaks the connection.

Mosfet stabilises Wien amplitude



Spot-frequency Wien-bridge oscillator uses a mosfet instead of a thermistor for amplitude control, while retaining a THD of about 0.02%.

A common method of stabilising a Wien-bridge oscillator is to use a thermistor, which is expensive. This 1kHz spot-frequency oscillator uses a mosfet to do the same job, with low distortion.

Positive feedback via the frequency-determining network of the bridge is attenuated by $R_{9,10}$, negative feedback through $R_{6,8}$ being reduced slightly more. Diode D_1 rectifies the output and charges C_1 to bias the fet partially on which, since the fet is in series with R_7 across R_{10}, reduces positive feedback until it equals negative feedback and the output amplitude stabilises at about 3Vpk-pk. Signal voltage across the fet is small, keeping distortion to a reasonable level, being almost all second harmonic at 0.4% without the further network around in the fet source circuit.

Output signal through C_2 and $R_{4,3}$ modulate the fet source voltage in the correct phase to reduce the residual distortion, C_3R_5 adjusting the phase for minimum distortion, which now consists of third and higher harmonics.

Tests show THD at 0.02%-0.025%, for a number of fets. Output amplitude is completely stable for supply voltages from ±6V to ±15V, with minimum distortion at ±9V; varying R_4 sets minimum distortion in the whole supply voltage range.

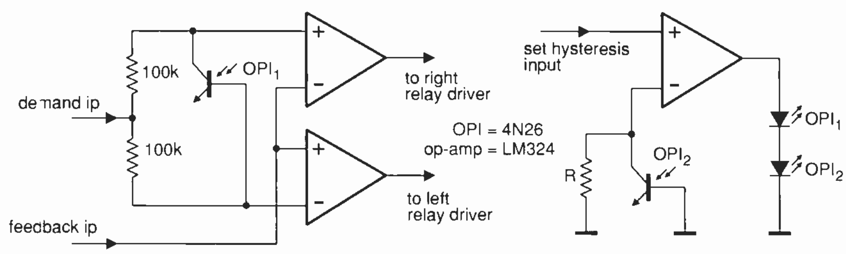
Output frequency of the circuit shown is 1009Hz. Resistors should be 1% or 2% types and capacitors 5% or better.

Ian March
Waterlooville
Hampshire

Simple servo

A very simple servo loop controls two relays that switch a reversible DC motor having a slider potentiometer to measure its position.

To prevent the servo loop hunting, the amplifier is provided with hysteresis; when the feedback is within the hysteresis band, both relays open and brake the motor. Unusually, an opto-isolator generates a "floating" voltage to define the hysteresis, with another to compensate for temperature variations.



Servo, originally used to control steering on a small vehicle, uses opto-isolators to provide a floating hysteresis voltage to define dead zone.

Resistor R should be the correct value for equal opto-isolator voltages. Adjust the "set hysteresis" input until the servo does not hunt; in the original, the potentiometer provided 600mV from full right to full left and 40mV of hysteresis was needed, which is equivalent to 3% worst-case error. Solid-state switching would improve this performance.

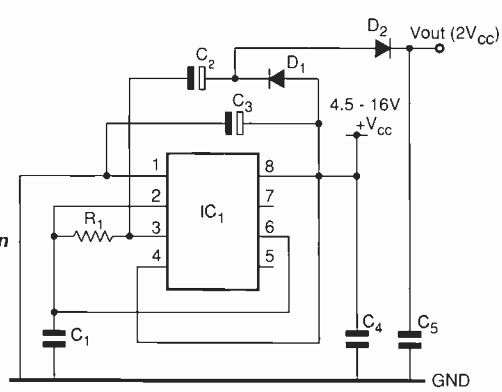
W Gray
Farnborough
Hampshire

Gate-voltage generator

The advantages of mosfets when used to control small power devices are unquestioned, but do not include the problem of gate voltage provision, which can be troublesome since it must be up to 9V greater than the source voltage – not ideal for battery-powered circuitry.

The 555 oscillator drives a voltage multiplier supplying about twice its own supply voltage. Current output is small, but mosfets take negligible gate current; this circuit has driven ten mosfets with no voltage drop.

George Smith
Fencehouses
Tyne & Wear



Voltage multiplier to supply gate voltage to power mosfets, particularly in battery-powered designs.

LETTERS

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

Objective assessment

The sole purpose of audio equipment is reproduction of music. From this, it follows that there is only one standard against which an audio design may be assessed: its effect on music reproduction – the quality criterion.

If the objective methodology used for evaluation relies on standards other than the quality criterion, for example objective measurements of some kind, then objective evidence must be provided that proves those standards accurately reflect the quality criterion. Without such proof, the results of the evaluation have no validity.

Similarly, if evaluation test conditions differ from designed-for conditions, such as amplifier load not a loudspeaker, or the test signal not music, then evidence must be provided that *proves* the changed conditions do not affect the results.

As the quality criterion is essentially subjective in nature, I would be surprised if the evidence to support an objective approach actually exists. In that case, this 'objective' method is just another form of subjectivity.

I would welcome alternative views adding to our collective understanding of this fascinating subject. But, to paraphrase Douglas Self, if you don't have anything worthwhile to add, please don't add anything!

Stephen J Merrick
Stockport

Quantified listening pleasure

Jerry Mead's confidence (Letters, *EW+WW*, July) provokes a wry grin. If, as he asserts, he is a trained listener and able to choose the listening conditions, then I wouldn't be in the least surprised he is able to detect an audible change when one element – say, the amplifier – is simply substituted in a reproduction chain. But he and many others are missing the whole essence of the objectivist argument.

First of all, if a change is genuinely audible, then a change such as a difference of 0.5dB in mid-band gain is measurable, quantifiable and correctable – something the ear/brain interface can never do. Moreover, if the change is a perceptible deterioration in fidelity, then it can be identified by well-understood procedures.

There is no magic, whatever anyone likes to think – the well-known Quad tests in which I took part, taught me that, if nothing else.

To compound the growing impression that I am a dreadful old reactionary, I also happen to be quite confident that a very good engineer would have to try hard to design a bad amplifier these days – such is the state of the science as now fully understood. The essential topology of a competent design, so lucidly outlined by Doug Self, adds a great deal to what we knew already. However, even Doug will agree that if Ben Duncan favours us with an amplifier of his own, that it will probably sound no different from his. But if it costs more, Ben, then the exercise is a failure.

What I and many other experienced design engineers take vigorous exception to is any assertion that ears alone can determine whether any item of audio hardware sounds 'good' or 'bad'. This is what the self-proclaimed experts in the audio comics continuously try to persuade their largely technically-ignorant readers. There are far too many variables involved: psychological, physiological, emotional – you name it. It also perpetuates the myth that domestic sound equipment is getting better. It is not to any significant degree.

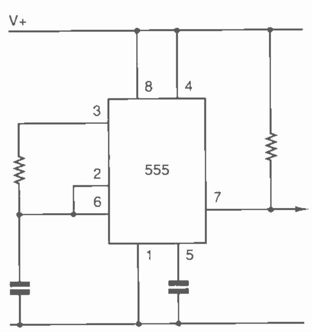
For example, the only tangible difference with each successive model of a compact disc player is an increase in the amount of air inside.

A decade ago, when rather more affluent, I would have happily put up a four figure sum as Peter Walker did, fully confident that given mutually agreed test procedures, the Moir/Quad test could be repeated with identical results. Incidentally, the magazine guru who provoked it chickened out of participation at the eleventh hour and hasn't been heard of since.

Reg Williamson
Staffs

Simpler circuit

The objective of I C Rohsler's circuit (*Circuit Ideas, Square waves from a 555*, July) can be more simply realised by reversing the roles of pins three and seven. Pin three is normally the output and has active pull-down and active pull-up. Pin seven is normally the discharge route, and has only active pull-down (being open-collector).



The timing resistor is now connected to pin three and the output taken from pin seven, which may or may not require a pull-up resistor to the supply depending on the application. An automatic 50% duty cycle results and there is no need to mess around with pin five. See also *Circuit Ideas*, December 1990.

D M Bridgen
Surrey

Amp designers top ten

The Self/Duncan controversy is fascinating! But, there's one glaring flaw: Where do they get their sources of distortionless music. And what sort is it – highbrow, middle, pop, Motown or jazz?

I have a good collection of several hundreds of recordings on assorted media including cassette tape, reel-to-reel, vinyl, CDs plus a digital four-band radio. But how much of it is true hifi? With my hand on heart I have to admit a much smaller percentage than I'd like.

Could Mr Self and Mr Duncan be induced to tell us their top ten, with full catalogue numbers? After all, if it turns out that one likes brass bands and the other South American, much would be explained.

Ronald G Young
Sussex

No military/civil distinction

I was most interested to read R M Burfort's view that military spending in the electronics industry letter has been a bad thing. Looking back over my career in electronics I recall an early Pye television receiver that incorporated the if strip out of an airborne radar, type *H2S*.

As a keen radio ham listener in those days I had an *R1155* bomber receiver. I also built a superegenerative set for 10m using a CV2 valve, a double-horned triode, and heard the New York taxis on it.

When I joined the Ministry of Transport & Civil Aviation in 1951 most of our equipment was ex-Air Force. The vhf receivers were *R1132* manual tuning and *R1392* crystal tuning, the transmitters were *T1131* for vhf and *T1190* for hf; the hf receivers being *CR100*.

Marconi Instruments when I went there in 1958 had one third military business, one third civil and one third export. At MEL, in 1969, I experienced mostly military contracts but did work on Madge which was a helicopter landing aid. We also did some tests for North Sea oil rigs.

My last company, Rediffusion Flight Simulation had both military and civil contracts.

My point is that, unfortunately, a balance of military and civil is required – though it would be better for mankind if so much effort were not put into weapons of mass destruction.

P D Somerville
Sussex

Where is non-magnetic power?

Two years ago, in July the 1992 issue, you printed an article entitled *Electricity without magnetism*, pp. 540-542.

Unless I missed any follow-up or letters from readers there has since been a conspicuous absence of comment. This surprises me because in a power hungry world, the hypothesis put forward would seem to be of monumental importance.

The inventors, Aspen and Strachen, would surely not still be endeavouring to overcome the non-availability of the essential PFDV (?) material referred to. Could it be perhaps that the inventors fooled the US Patent Office into granting them a patent. Or perhaps one of the many powerful industrial bodies has successfully persuaded the inventors to forget it.

On p542 Editor Frank Ogden and consultant Derek Rowe seemed suitably impressed. But I suspect that neither of them really understood just how the device worked. The interpretational drawing and editorial comments included does not satisfy all the claims made by the inventors: where does that leave us?

Either the claim is a complete sham and US Patent Office has been conned, or perhaps the article was a rather late April fool.

R L Tufft
Thirsk

Neither Derek nor I would claim total understanding of the phenomenon reported in the article and I don't think the authors would either. Every so often an experimental observation is made which defies explanation by either you, me or the US Patent Office. It must be right to report that observation in the hope that a proper understanding follows in the fullness of time. Ed.

Hall's well that ends well

It was indeed with misty eyes that I read the article about Hall effect devices (*Applying Hall to good effect*, *EW+WW*, July, pp.576-580). I first had to work with these devices some years ago, when employed as a maintenance engineer at a local semiconductor production factory.

Night shift technicians kept leaving notes for the day shift engineering group to investigate a malfunction in a simple machine used to crop component leads. It appeared that the machine was not indexing accurately. Stopping and starting positions were set by a Hall effect switch mounted on the fixed frame of the machine, and little cylindrical permanent magnets on a rotating wheel, close to the Hall switch.

Investigation with a sophisticated field strength tester – the end of my screwdriver – revealed a noticeable difference in the 'pull' from each magnet, causing the switch to trigger at the wrong place at each station on the index wheel.

Being a battle-hardened maintenance man, I took the only sensible option available: rip the magnets out of the wheel and fit a roller action micro switch on a bracket beneath the Hall switch. The holes in the wheel provided excellent detents for the switch, and indexing was perfect.

If any purists want to discuss this inelegant solution, I can only refer to the shift technicians who thought it a doddle to set up, and the production people, who wanted to know why it hadn't been done correctly in the first place – a simple answer that clearly demonstrates that we should all take in to account the secondary effects (and beyond) of our designs.

Peter Thornton
Oldham

Old radio club

I am researching the history of the Southend and District Radio Society. I have proof of its existence in November 1923 as I have committee meeting minutes back to that time. I also have some evidence, in the form of leaflets and handwritten notes, that indicates the Society was formed in 1920. But there is no confirmation of this because these documents are not dated or signed.

If the Society was founded in 1920, this will be quite important to us because it means next year will be our 75th anniversary, and of course we shall stage a specials event to mark the occasion.

I would be very grateful for any information anyone may have on the Society as it will be important to future members to have a complete history of one of the oldest amateur radio clubs in the UK. In the early days it was known as the Wireless Society of Southend.

Can any *EW+WW* reader help m out with any more background?

LJ Burchell
Southend and District Radio Society
PO Box 88
Rayleigh
Essex SS6 8NZ

c change?

May I firmly rebut Michael Williams' contention (Letters, *EW+WW*, July) that the decrease in speed of light is due only to the inaccuracy of early equipment. Surely he could give Nobel prize-winners such as Michelson credit for knowing the difference between results that vary around a true value and those that show a consistent decrease. This is what Michelson found with equipment whose accuracy and sensitivity was more than adequate to detect changes.

If Mr Williams, or any of your readers (UK only for the moment), cares to send me a loose £1 stamp and an A4 addressed-envelope, I will gladly loan them the 90-page 1987 Stanford Research Invited Report with, 377 references, written by Trevor Norman and Barry Setterfield.

I think recipients will find it blows away Mr Williams' belief that the decrease in c need not be taken seriously.

As the author of *Science vs Evolution*, referred to by Mr Goldberg (Letters, July) I can assure readers that a 'scientific and media mafia' most certainly does exist. They operate to ensure that what reaches the public are only orthodox views on the many controversial subjects that include evolution, relativity and heliocentricity.

For that reason, your publication is to be congratulated as being one of the very few to have the courage and freedom to publish alternative evidence on sensitive subjects where open debate is discouraged.

Malcolm Bowden
Kent

Dying light

Statistical analysis of over 600 observations of 12 different atomic quantities measured by 25 disparate methods over 300 years has confirmed the decrease in the speed of light. For example, R, G and e are truly fixed constants, whereas h, e/mc, h/e, and $2e/h$ are non-constant, and have varied exactly as would be expected with a falling c (*c-decay* or *cDK* as it is called by Australian researchers).

Yet Michael Williams (Letters, July 1994) attributes the relentlessly diminishing values of c to observational errors or instrument limitations in the past. But if this were the case, there would be a spread of results on either side of a true constant value, not a steady, monotonous decrease.

Since the 1960s, we have been basing our standard of time not on the astronomical clock, but on the atomic clock, which uses the frequency of vibrations of electrons in caesium. The frequency is bound up with c, so that if c is not constant, our time standards will be faulty.

In 1984, van Flandern of the US National Bureau of Standards showed that the atomic clock is slowing down relative to astronomical time. All c measurements since the 1960s have been calibrated using atomic clocks. The clocks are themselves moving lock-step with any change in c, so no variation can be noted with that method.

cDK is subtler, better-developed, and having wider ramifications than Michael Williams can imagine. It has radically reduced the age of the universe, recalibrated radiometric dating downwards, forced a reinterpretation of the Red Shift, eradicated the Big Bang Theory, sunk the Nebular Hypothesis, undermined uniformitarian geology, and destroyed Darwinism – all at the same time.

Amnon Goldberg
London

Bad references

This *Letters* column is becoming increasingly disturbing. It should be insisted that people who quote experiments purporting to show that all modern science is erroneous must at least give their sources.

Your correspondent Amnon Goldberg is particularly bad at this as he quotes all kinds of experiments without ever giving usable references. On the very few occasions when I have found his sources, they have either not said what he thought they did or were highly contentious and difficult to accept.

But the nadir came with Michael Williams' letter attacking Mr Goldberg (May). This was in execrable bad taste and gratuitously offensive. He accused Mr Goldberg of being related to a modern mass murderer. Even if this were a joke it is not the type that should appear in a respectable journal. He called Mr Goldberg a 'loony chabadnik'. The first term is highly offensive. The second can only be known to a very small number of your readers and I can assure you is not an accurate description whatsoever of Mr Goldberg.

To cap it all, even Mr Williams' science is wrong. The point about the hollow space in the centre of a sphere of uniform mass is that for an inverse-square law field, the interactions due to the outer layer all cancel out. As a result, the measurement of G at the Earth's surface is correct. This is first year undergraduate problem given in many text books.

Perhaps in future you will exercise your editorial prerogative to edit out offensive remarks.

Michael Slifkin
Address unreadable or witheld.

I don't necessarily endorse the opinions expressed in the Letters column. Occasionally they make me wince, they frequently make me laugh. Every so often, I encounter something which makes me rethink accepted truths. I hope that other readers react in a similar fashion. It would be a very thin Letters *column if we restricted ourselves to the eminently worthy. And boring too. -Ed*

Private progress – public property?

I believe there is no such thing as 'intellectual property' (Patently unclear, May, pp.433-436) – except in the sense of memory of private experiences. Otherwise why should we oblige every citizen to share at least nine years of compulsory education, asserting that knowledge is as universal as it is everyone's property?

The path to modern technology commenced with Newton's physics, was refined by the French mathematicians who shared their discoveries freely, and every new law is exploited as a newly discovered aspect of nature.

Michael Williams
Beth Shemesh
Israel

Two wrongs...?

We recently saw , as part of the *Heretics* series on *BBC1*, the molecular imprint of a heart drug supposedly transferred via a magnetic circuit to a container of distilled water. I am prepared to believe that water may have some memory feature, though to transfer if by electromagnetic means using simple components is rubbish.

We have also seen, mostly within the pages of *EW+WW*, much debate on how ac transmission lines could concentrate gamma rays, and so lead to cancer.

Though both theories seem unlikely, I have an observation to make based on these two apparently unrelated effects.

Given that EM fields can affect the 'memory' of water (supposition); that the human body is more than 90% water (true); that ac magnetic fields erase magnetic tape by randomising the magnetic domain element (true); and that the human immune system is helped by the transfer of 'templates' by water (big supposition), then prolonged exposure to overhead ac power lines could erase the information carried by the body's water.

If that were so, the immune system may not be as good as it could be and people with a tendency towards a terminal illness would be more likely to acquire it while living under a power line than those who do not.

It could be a case of two rubbishy theories being used to create a third one. I neither believe them nor disbelieve them but I am prepared to be convinced.

Peter Barnes
Bristol

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INDEX TO ADVERTISERS

	PAGE
Amdat	732
Anchor Surplus Ltd	711
BK Electronics	718
Bull Electrical	739
Chelmer Valve Company	718
Dataman Designs	BC
Display Electronics Ltd	742
Field Electric Ltd	759
Halcyon Electronics Ltd	729
ICE Technology Ltd	706
John Morrison	729
John's Radio	712/734
JPG Electronics	732
Kestral Electronics	775
Keytronics	765

	PAGE
Lab Center	751
Langrex Supplies Ltd	759
M&B Electrical	766
M&B Radio (Leeds)	716
MQP Electronics	726
Number One Systems	726
Powerware	718
Ralfe Electronics	792
Seetrax Ltd	771
Smart Communications	IFC
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