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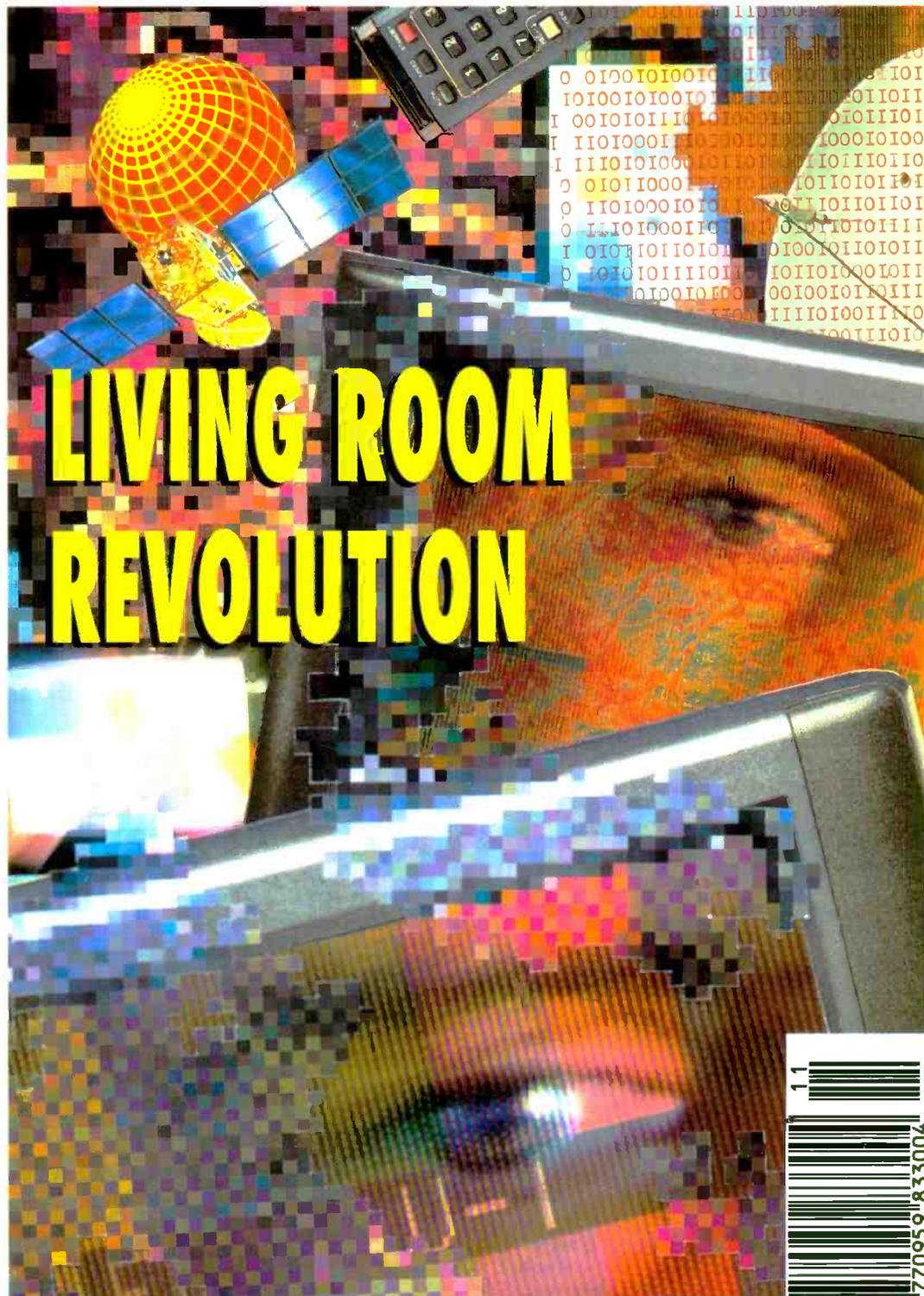
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Cover illustration
Kevin Clarkson

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Programmable logic devices are essential to compact logic design and have been so for a long time. They save component count and board area and provide adaptability of circuit function. Despite this utility their operation is not always well understood. In the first part of a new series on logic design Geoff Bostock explains the workings of programmable logic.

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Do it yourself spectrum analysis. There is nothing like a spectrum analyser to help things along if you work or play with RF. There is also nothing like the price tag. Ian Hickman presents a good performance, penny pinching DIY design.
Also next month: a single chip PC.

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386 AT - PC386

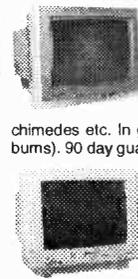


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EDITOR

Frank Ogden
081-652 3128

DEPUTY EDITOR

Jonathan Campbell
081-652 8638

CONSULTANT

Derek Rowe

DESIGN & PRODUCTION

Alan Kerr

EDITORIAL ADMINISTRATION

Lorraine Spindler
081-652 3614

ADVERTISEMENT MANAGER

Carol Nobbs
081-652 8327

SALES EXECUTIVE

Pat Bunce
081-652 8339

ADVERTISING PRODUCTION

Shirley Lawrence
081-652 8659

PUBLISHER

Susan Downey

EDITORIAL FACSIMILE

081-652 8956

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Picture of opportunity

The European electronics industry hasn't had much cause for celebration in recent years. The consumer interests suffered badly at the hands of the Japanese while the ending of the Cold War took the steam out of the defence business. Just six months ago, its great hope for a new generation of broadcasting based on the MAC analogue TV standard fell apart as digital broadcast strategies began to emerge.

But it has become clear that the opportunity presented by digital television will be massive... Enough to replace the ailing defence interests with market driven civilian business while at the same time contributing to technology at the first stage of its development.

There is good evidence that the initiative will not be lost on this occasion. Some 85 European broadcasters, equipment makers, satellite operators, network carriers, research organisations and Government departments have signed a memorandum of understanding to establish world class digital TV standards and infrastructure in Europe by the end of the decade. Collected interests cover all aspects of digital video: origination, production, networking to satellite and terrestrial broadcasting.

The initiative will be called the digital video broadcasting project (DVB) and the first services to be run under its unified standards will be on cable and satellite by the end of 1995. This will be followed by digital terrestrial services some two to three years later.

The mix of interests contained in the Memorandum will ensure that it either disintegrates spectacularly or forms the basis of a world digital broadcasting standard. For instance Japanese interests are represented by Matsushita, Sony, JVC

and Toshiba. Given that the compression element of the package relies heavily on the US driven MPEG-2 technology, it is likely to succeed all over the world.

It is almost impossible to overstate the significance of this announcement to European electronics. The start point for all aspects of this venture are at least equal and possibly favourable to local interests. For instance, it seems likely that terrestrial broadcast modulation standards will use coded orthogonal frequency division multiplexing, a close relative of spread spectrum modulation, and pioneered by NTL in this country. The same organisation has also made an impressive contribution to MPEG and its work features prominently in Rupert Murdoch's satellite broadcasting plans. Tom Woodford has written a comprehensive article on the technology in this issue.

Real technical advances in digital TV technology have been made by European electronics companies such as Philips, Thomson, Nokia and Grundig, all parties to the Memorandum. They will make new equipment which, unlike MAC based sets, we will want to buy along with the rest of the world.

Unified digital TV technology will expand vastly the number of channels and types of service which broadcasters provide. Information on demand will create a new market for business and domestic terminals, essentially TV sets with digital storage facilities and two-way communications channels built-in. It will offer new business opportunities from new technology leading to the development of a new communications infrastructure.

It would be hard to find a more optimistic picture on any other channel.

Frank Ogden

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UPDATE

Digital chips to change the face of radio

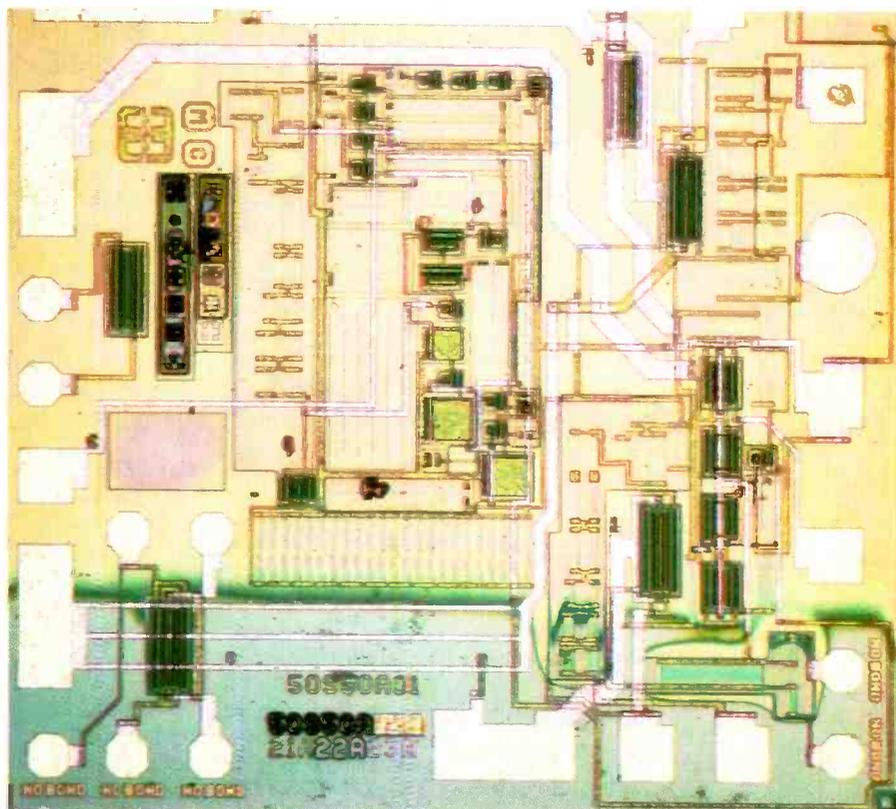
A new class of dsp based processing chips can replace traditional analogue radio architecture with direct digital processing. Developed by the US company Harris Semiconductor for the portable telecommunications market, the devices perform filtering, second local oscillator and quadrature demodulation functions without reducing the signal to baseband as in conventional receivers.

The 16bit 52Msample/s processing rate allows the construction of filter equivalent functions with more than 100dB of attenuation at a knee frequency of 13MHz.

Being a true digital architecture, the radio dsp family is programmable allowing design of system functions under software control. This flexibility includes the construction of separate I/Q channels for data recovery.

The company has also produced a front end chip with a 2.2dB noise figure rf amplifier (at 900MHz) integrated with a +10dBm intercept double balanced mixer on the same piece of silicon. A similar device for 1.8GHz is being developed.

Although hand-held radio is the intended application, the multiple function dsp chips will find use elsewhere. For instance the HSP50016 comprises a direct digital synthesiser with internal sine lookup table, mixer and two lowpass filters tunable with a resolution of 0.0091Hz. The local oscillator function is claimed to be spurious free down to -102dB below carrier level. And the filter function is claimed to have less than 0.04dB of ripple in the passband. It provides a complete downconversion subsystem on a chip.



2.2db noise factor: HFA 3500 silicon front end

Other dsp based devices released initially include a 52Msample/s half-band filter and a serial i/o filter with more than 140dB of stopband attenuation.

Placing specialist radio functions in dsp

chips will eliminate much of the analogue signal chain and its associated set up adjustments. It also reduces the need for block filtering; complete data recovery can be made with very few high frequency conversions stages.

Frank Ogden



Glint in Du Pont's eye: The first merchant accelerator chip in the world for the OpenGL 3D graphics standard may be launched next year by Du Pont Pixel. Nicknamed *Glint*, it will be able to manipulate images like the one above at up to 300,000 Z-buffered Gourand-shaded triangles a second. High level 3D manipulation instructions are taken by the chip and performed on objects stored in a set of off-chip buffers. It will run in Unix and Windows NT host systems. The first version will not handle geometry transformations; this will be integrated into later versions of *Glint*, hopefully by 1996. A colour-map chip, due for launch in 1995, will be integrated into *Glint* by 1997.

P6 is out of order

More details have been released about the P6, Intel's planned successor to its Pentium (i586) chip. The P6, or i686, is in development and is scheduled for volume production in 1995.

It will be built with 0.6µm features in an attempt to pack more than five million transistors onto a single chip. Initial versions though may be in a multi-chip set.

Speed will be 133MHz and about 200Mips. There will be four integer and two floating point processors.

Out of order instructions will be able to be executed with the processors trying to predict which instructions it will receive. A DX version will be capable of running in multiprocessor systems and the LX version will be for single processor desk-top systems.

Engineers must have green fingers

The Engineering Council is putting UK engineers and technicians in the front line for safeguarding and improving the environment.

With the launch of the Code of Professional Practice on Engineers and the Environment, the 290,000 engineers on the council's register were called upon to make the environment a key consideration in every project they undertake. The code comes into effect in March.

The announcement coincided with the news that Anthony Convery of County Tyrone had won the council's 1993 Environment Award for Engineers for developing a machine to recycle waste concrete from ready-mix lorries and on-site concrete mixers.

A commendation was given to an inexpensive radar for insect pest forecast and control developed by Joseph Riley, Alan Smith and Douglas Gregory of the Natural Resources Institute in Kent.

* For the second year running a girl has won the Engineering Council's Young Engineer for Britain 1993 award. Lucy Porter, 16, of Bath invented a swing exerciser for children with special needs who are unable to use their legs. The Mountbatten prize for the best project using electronics went to Samantha Haines, 18, from Guildford for a physiotherapy aid. And the best user of information technology in a project award was won by Stuart Cowley, 18, for a lubrication advice service.

Laser linewidths cut to 3.6kHz

A semiconductor laser with a spectral linewidth of 3.6kHz, 3000 times narrower than conventional semiconductor lasers, has been developed by Hitachi.

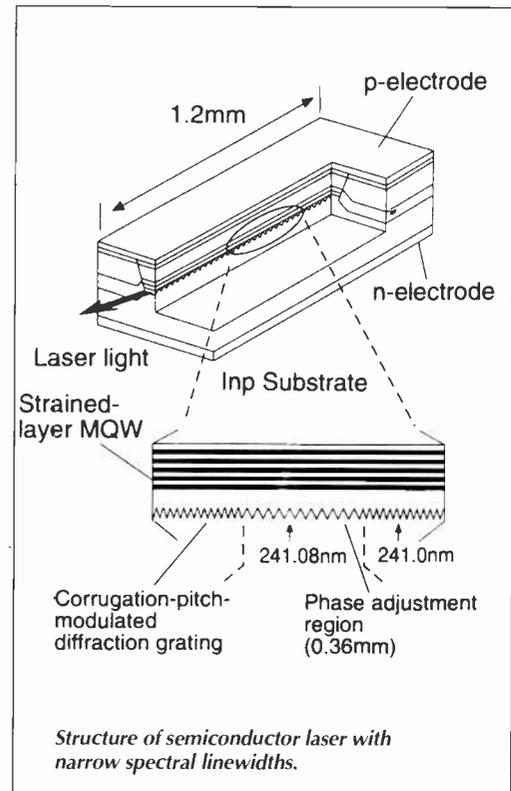
This narrow bandwidth may make it suitable for use in ultra-high speed coherent fibre optic communications systems operating at 200Gbit/s or more.

Inside the laser is a stack of multiple InGaAsP layers on an InP substrate. Infrared radiation is emitted at a wavelength of 1.55µm and output of 55mW.

Narrow wavelength distributed feedback diode lasers have a diffraction grating etched onto the device. This sets up a pattern of multiple interfering wave trains that cancel out except in a narrow frequency range. But the resulting bandwidth of about 10MHz is too wide for coherent fibre optic communications.

To get round this, Hitachi has adjusted the pitch of the diffraction grating on the device so that the grating spacing is slightly longer in the centre of the resonator. The interference pattern produced results in a much narrower bandwidth.

The company also uses a strained layer multiple quantum well structure with alternate layers having different lattice constants.



Level playing field wanted for pcbs

Europe's pcb makers are lobbying the European parliament in a bid to give pcbs the same protection in Europe as semiconductors. Brian Haken, executive director of the UK's Printed Circuit Interconnection Federation, said European pcb makers have a disadvantage compared with the Japanese when it comes to getting money from the banks.

He said that "Japanese board makers can borrow at low preferential interest rates and

for longer than their European counterparts. All we want is a level playing field."

He pointed out that Sony had received financial assistance from the EC to establish a pcb plant in France.

Miti, Japan's industry body, has targeted pcbs and semiconductors as two crucial areas for future growth and it plans to double Japan's income from pcbs in the next five years.

Transistor breaks noise barrier

Production has started of a redesigned high electron mobility transistor, which maker Toshiba claims has a lower noise level than any other on the market. Noise output is 0.45dB at 12GHz.

The biggest source of noise in this type of transistor is resistance in the channel layer, normally made of GaAs. Toshiba has used InGaAs, adjusting the level of indium

to make a lattice with minimal resistance.

The crystal structure has also been improved in the AlGaAs layer, which supplies electrons to the channel layer. This makes the electron flow 40% higher than in plain GaAs devices.

Mass production is scheduled to start November reaching 100,000 pieces by April.

Work ethic for wireless communicator

Motorola is to use Microsoft's At Work software in a new wireless personal communicator.

The company has not announced when these communicators will be out or what their price will be.

But Bob Growney, executive vice

president, said that "developing wireless personal communicators based on Microsoft's hand-held system allows us to leverage their strength in desk-top software to provide a pragmatic solution for today's mobile professional".

GaAs moves down commercial pipeline

The move from military to commercial applications for GaAs and the design of antenna arrays rather than simple aperture or wire types were the two most notable trends at September's European Microwave Conference in Spain.

GaAs technology has largely come of age in the form of MMICs. Until recently microwave applications for GaAs have been mainly as discrete devices, particularly the mesfet and low noise high electron mobility transistor. These devices have been integrated into planar technology circuits to form hybrid modules.

But GaAs is also a suitable substrate material for microwave transmission lines and so the complete MIC can be formed monolithically.

With the prospects appearing for consumer and commercial applications, the rewards in terms of numbers sold is very high. In military contracts a "large number" was still several orders of magnitude below that of present prospects.

A paper from E Pettenpaul of Siemens claimed an estimated market of 20 million direct broadcast satellite receivers by 1995 for Europe alone. And large scale use of personal and mobile communications in the 900MHz and 1.8GHz bands is estimated at 100 million units in Europe by 2000.

For office use, including wireless lans, there is a GaAs MMIC effort in all the allocated bands to 60GHz and above.

A paper by N Diehl of Daimler-Benz pointed out that another vast application for microwave communications is mobile computing. For example, test or calibration engineers would be able to perform complex analysis and checks on machinery, equipment and installations, and have full real time access to all data on such equipment.

There seems little doubt that in a few years there will be a steady growth in intelligent vehicle communication systems – again a potential market for 100s of millions of sensors. Some aspects of such links in the 5.8GHz band were presented in a paper by Bucks and Grabow. On higher frequencies to 60GHz there are also applications for automatic collision avoidance and high-speed train links.

In all these systems the market volumes are so high that the pressure on microwave GaAs MMIC technology to achieve low cost, reliable performance and small size is very high.

On antenna design, some present and certainly future satellites will have narrow spot beams that are intelligently switchable in direction. Thus arrays are needed that have the advantage of being able to combine a total transmit power from one amplifier per array element, or perform spatial signal processing using a separate receiver in each element.

Incorporating attenuators and phase shifters with these functions lets the beams be steered and shaped.

Applications also include satellite to satellite links. But such arrays have 100s to 10,000s of individual elements and, to be viable for consumer or commercial use, the costs of the microwave circuits in each element must be dramatically small. Again, the emphasis is on ingenious circuit design, accurate performance modelling, and a low-cost manufacturing technology.

The conference also saw a selection of papers looking at possible applications for superconducting materials with a critical temperature above 77K. These included microwave components as high-Q circuit elements, non-dispersive signal processing circuits, or low-loss small antennas.

Another developing field with wide applications is that of the interaction of microwaves and optical systems. For example, a paper from D Wake of British Telecom talked of a microwave transmitter needing no electrical power and a fibre optic input. It used video pattern data to frequency modulate a 4.1GHz microwave oscillator, which in turn drove a laser. The fm output of the laser then passed into a dispersive delay line to generate harmonics of the microwave frequency plus modulation.

The optical signal is converted to microwaves using a zero bias edge coupled photodiode with the seventh harmonic of 28.7GHz selected and transmitted to a superhet receiver. All power requirements in the transmitter section were derived from the initial optical signal. Possible applications are links in lans for broadband access.

At the associated exhibition, Thomson-CSF announced that it is acquiring the travelling wave tube and coaxial tube business from Siemens. Also announced was that Eesof is merging with Hewlett-Packard subject to government approval. Eesof simulation software is used for microwave cad.

Mike Hosking

Vodafone wins cellular survey

Vodafone provides a more reliable service than Cellnet according to a survey of cellular networks by industry watchdog Oftel.

Some 29,000 attempts were made to make two min phone calls from mobile to fixed networks and vice versa.

On mobile to fixed network calls 97.1% were connected and completed on Vodafone compared with only 95% on Cellnet.

And on fixed network to mobile calls Vodafone again came out on top with 95.1% set up and completed compared with only 94.3% on Cellnet.

Beeb dabbles with digital

The BBC has started engineering tests of digital audio broadcasting (dab) using high power transmitters in the London area.

A 10kW erp transmitter at Crystal Palace is being joined by 1kW erp transmitters at Alexandra Palace, Reigate and Wrotham. They will operate at 226MHz.

Survey vehicles will measure the field strengths of each transmitter and the way they work together to form a single frequency network.

Following the tests the BBC expects to produce proposals for introducing dab services in the UK. The plan is for national and local dab services to broadcast terrestrially using vhf frequencies.

Joint project for commercial gps

GEC Plessey Semiconductors and Canadian Marconi are jointly developing receiver technology for commercial gps applications.

GEC will be responsible for making and distributing the MicroGPS family of global positioning engines. And Marconi will offer a customisation service for those who want gps functions in other equipment.

Ray Gleason, GEC's marketing director, said that among the technical advantages of MicroGPS is the "very fast 15GHz bipolar process that enables low-cost low-power gps rf technology to become available for the first time to the commercial market place".

The family will be available as standard oem modules for direct plug-in applications.

Audio compression for comms ICs

Enfield firm Silicon Systems has been licenced to use California-based DSP Group's Truespeech audio compression algorithm in its communication ICs for lan, modem, wireless and multimedia products.

The algorithm can compress a 1min voice file down to 60Kbyte with no noticeable degradation. This compares with about 960Kbyte for standard audio compression algorithms.

The firm has also started to integrate DSP Group's Pine dsp technology in the ICs to make them suitable for mobile computers, personal digital assistants and other portable computing applications.

Silicon Systems can be contacted on 081-443 7061. ■

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CIRCLE NO. 102 ON REPLY CARD

RESEARCH NOTES

Sulphur chemistry recharges battery technology

Scientists have developed an aqueous cathode cell based on solid sulphur with a theoretical storage capacity several times that of conventional aqueous system cathodes based on nickel, manganese, mercury or silver.

The room-temperature battery could be a significant step forward in the search for new, more efficient, energy storage, fuelled by applications as diverse as environmentally friendly vehicles and the rapid growth of cordless communications. In each case the requirements are broadly similar: high energy storage per unit weight, good safety, room-temperature working, long life and low cost.

So far, no battery has met all these requirements.

The highest energy systems use either molten electrolytes, or expensive or environmentally-unfriendly metals for the electrodes. Many other promising room-temperature aqueous systems are beset with technical problems. Established technologies have weight penalties, environmental disadvantages, modest capacities or high cost. But that could change with the announcement by Dharmasena Peramunage and Stuart Licht of Clark University in

Worcester, Massachusetts, of what looks like a promising new room-temperature battery using an aqueous cathode based on solid sulphur.

Their research follows on from earlier studies into the use of aqueous sulphide solutions – unusual in that they will dissolve elemental sulphur. Sulphur is normally insoluble and non-conducting at room temperature. The resulting polysulphides have been used successfully in experimental rechargeable polysulphide cathode/tin anode cell.

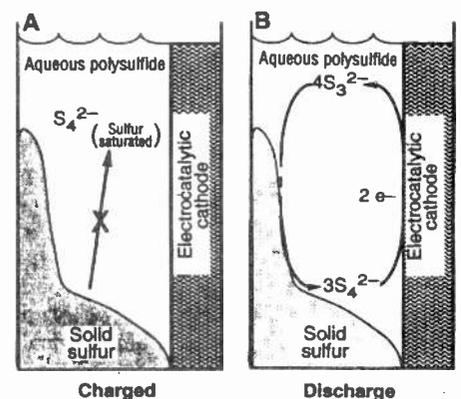
But the latest development takes this approach a stage further with the development of a cathode capable of direct reduction of elemental sulphur at room temperature. It makes use of solid sulphur in contact with a fully saturated solution of polysulphide. The theoretical storage capacity of this composite sulphur cathode is several times larger than that of conventional aqueous system cathodes based on nickel, manganese, mercury or silver.

Experimental batteries using an aluminium anode with the new sulphur cathode have a 40% greater capacity than the earlier experimental aluminium/polysulphide batteries. Open

circuit terminal voltage is also a healthy 1.28-1.30V.

As for an overall comparison with existing commercial aqueous batteries, the sulphur cathode system is said to develop 220W.Hr/kg – even at relatively high discharge rates – compared to 110 for zinc/air systems and up to 95W.Hr/kg for alkaline manganese batteries.

Peramunage and Licht say they are optimistic of further improvements.



Solid sulphur cathode in charged state (left) and discharging state (right).

Sound treatment for pollution

Acoustic agglomeration is a useful technique for separating smoke, ash and even fog. But no-one knew how it worked. Until, that is, researchers, led by Dr Gary H Koopmann of the Center for

Acoustics and Vibration at Pennsylvania State University, announced they had found the answer.

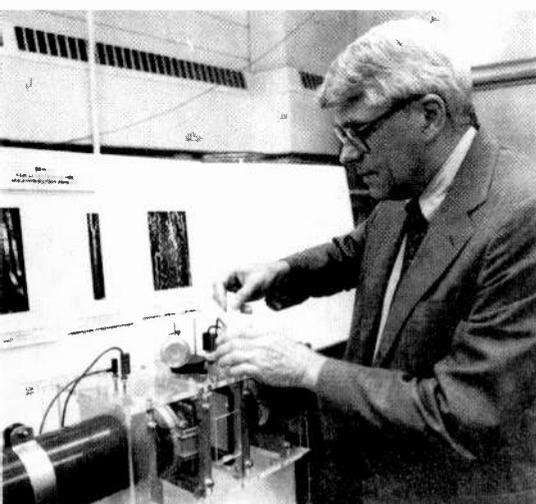
Koopmann and his team constructed a small, controlled experimental set-up with two face-to-face loudspeakers operating in phase, so that the air between them moved back and forth, rather than being compressed by the sound waves. Using commercially available 8µm beads (smaller beads agglomerate naturally), they dispersed small amounts in the form of a cloud falling through the chamber. A variety of sounds from 600 to 1000Hz were tested, but 800Hz was found to provide the best agglomeration.

Key feature of the experiment was a video camera that recorded the scene at 30 frames/s. Frame by frame examination revealed that, as the particles fell through the sound field, the larger ones appeared to be pursued by smaller ones travelling slightly faster. Eventually the smaller particles caught up with the larger ones and coalesced.

The Pennsylvania researchers believe that this phenomenon is a scaled-down version of what happens when a car gets into the slipstream of a lorry: the car tends to get sucked along. With suspended particles, the acoustic wind causes the smaller ones to "slipstream" larger particles and to be pulled closer before eventually colliding and clumping together.

"Not only did we see the particles coming closer together and joining, but there is also a cascade effect. Once two particles join, a third smaller one will draught the agglomerated larger particle and eventually join," says-Koopmann.

Acoustic agglomeration plants normally generate sounds at the 160dB level, horrendously loud, but fortunately confined to chambers that prevent too much escape. The aim of such techniques is to increase particle size to above 20µm, where they can easily be filtered out using conventional techniques. One of the main applications is the removal of fly ash from the waste stacks of coal-burning plants.



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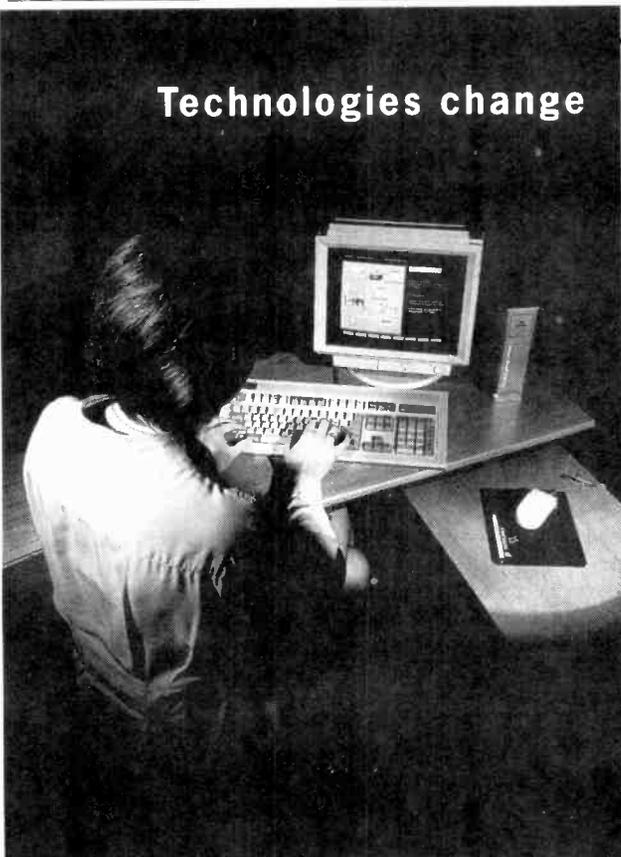
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Robots glimpse improved vision processing

If robots could flick their cameras around in the same way as humans switch their gaze back and forth, then they could make a much better job of image processing – with a lot less memory. That is the conclusion of research currently in progress at the University of Rochester in New York.

Mary Hayhoe, associate professor of psychology and a member of the University's Center for Visual Science, has been trying to understand why human vision is so markedly superior to that of robots, even though we mortals can only see clearly in the very centre of our visual field. Why is it that robots are fooled by something as simple as a change in lighting, when they are equipped with edge-to-edge razor-sharp cameras and megabytes of digital image-processing capability?

Hayhoe and her colleagues discovered that

the key to human image processing is eye movement. Unlike the robot eye, which stares continuously at a scene, the human eye constantly flicks around. These short jerky eye movements, known as saccades, are how we circumvent the need for massive amounts of image storage and processing power.

"Eye movements provide a really efficient way to do things," says Hayhoe. Instead of keeping track of every detail in the environment, you use the world as your memory. You don't get information until you need it."

Other workers at Rochester have built on Mary Hayhoe's work and produced a number of versatile robots that mimic the way humans study their visual field. Instead of boosting processing power to enable their robots to see every detail in their

environment – the usual approach to robotic vision – the Rochester group adopt an approach which they call "active vision." With this technique the robot uses sensory input from its environment to make its decisions, including the ability to move its eyes. By using artificial saccades, the Rochester robots simplify the image processing task enormously and considerably enhance their ability to make sense of the visual field. With no remote control, and in real time, these robots have been able to dodge tennis balls, manipulate toys and search for identified objects such as cereal boxes.

Saccades – humans make over 100,000 every day – are vital in our ability to make sense of the world. Our eyes move so fast that we have the illusion of being able to see the whole of our visual field in sharp focus. Tests at Rochester have shown that when people are asked to copy a pattern of building blocks, they don't rely on an image stored in the brain. Instead they make hundreds of saccades, through which they compare the original with the duplicate they are making. As Hayhoe puts it, "you don't just look once at a scene and then replicate it. Instead, your eyes are constantly going back and forth, picking up little bits of information again and again."

Traditionally artificial intelligence has assumed that huge amounts of memorised information are necessary for even the simplest pattern recognition tasks. But if the Rochester robots can successfully duplicate the full range of human eye movements, then it may, one day, be possible to perform even complex tasks without resorting to the usual elaborate models of the world



Why lightning is forked

Why does lightning zig-zag its way across the sky rather than travelling in a more-or-less straight line? As with many apparently simple questions, the answer is still being investigated. But plasma physicists at Moscow's Lebedev Institute of Physics and the Los Alamos Laboratory in New Mexico think they have the solution.

Finding a single theory to explain why lightning forks, why it generates x-rays and even why it takes place at all has been extraordinarily difficult. Numerous measurements of electric fields in thunderclouds show them to be too weak to accelerate electrons to the speed necessary to create atmospheric breakdown. To create an ionised trail for lightning to follow, an electron has to gain enough acceleration to overcome the loss of energy it experiences every time it hits an air molecule.

But Robert Roussel-Dupre and his associates in the USA and Russia have now formulated a theory that provides a coherent

explanation for all the above phenomena. According to the Roussel-Dupre's team, a lightning strike is initiated when a cosmic ray entering the Earth's atmosphere from outer space collides with an air molecule, ejecting a fast-moving electron.

If the cosmic ray imparts enough energy to that electron, then the electric field strength in a typical thundercloud is quite enough to accelerate the electron further, keeping it moving as it hits and breaks apart air molecules. Every collision creates an avalanche of high-energy electrons, each in turn accelerated by the electric field and leaving behind a trail of ionised air molecules.

According to Roussel-Dupre, this avalanche rumbles downwards for about 100m before it peters out, leaving a pool of electrically charged particles with its own associated field. The process might well stop there except that there are large numbers of cosmic rays that can re-trigger it.

After a pause of only a few microseconds, a new cosmic ray collision sends a fast-moving electron through the pool of charged particles, creating another avalanche and another ionised trail. This new trail is linked to the first one, but heads off in a different direction.

This step by step conducting stairway heads rapidly ground-ward until it is close enough for a streamer to head up and close the circuit.

It is a fascinating theory that explains both the forked nature of lightning and also the large scale – often 300m – over which it operates.

The researchers' mechanism further explains the emission of x-rays, which are thought to result from the high energy electrons generated by each successive avalanche. ■

Research Notes is written by John Wilson of the BBC World Service.

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 HP Signal Generator type 8656A - 0.1-990Mc/s. AM/FM - £2000.
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 Marconi microwave 6600A sweep osc., mainframe with 6650 PI - 18-26.5GHz or 6651 PI - 26.5-40GHz - £1000 or PI only £600.
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 Microwave Systems MOS/3600 Microwave frequency stabilizer - 1GHz to 40GHz £1k.
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 Alltech Stoddart receiver type 17/27A - 01-32Mc/s - £2500.
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 HP 3580A LF-spectrum analyser - 5kHz to 50kHz - LED readout - digital storage - £1600 with instruction manual - internal rechargeable battery.
 Tektronix 7D20 plug-in 2-channel programmable digitizer - 70 Mc/s - for 7000 mainframes - £500 - manual - £50.
 Datron 1065 Auto Cal digital multimeter with instruction manual - £500.
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 Schlumberger 2741 Programmable Microwave Counter - 10Hz to 7.1GHz - £750.
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 HP 2225CR Thinkjet Printer - £100.
 TEK 576 Calibration Fixture - 067-0597-99 - £250.
 HP 8006A Word Generator - £150.
 HP 1645A Data Error Analyser - £150.
 Texscan Rotary Attenuators - BNC/SMA 0.10-100-100DBS - £50-£150.
 HP 809C Slotted Line Carriages - various frequencies to 18GHz - £100 to £300.
 HP 532-536-537 Frequency Meters - various frequencies - £150-£250.
 Barr & Stroud variable filter EF3 0.1Hz-100kc/s + high pass + low pass - £150.
 S. E. Lab SM215 Mk11 transfer standard voltmeter - 1000 volts.
 Alltech Stoddart P7 programmer - £200.
 H.P. 6941B multiprogrammer extender. £100.
 Fluke Y2000 RTD selector + Fluke 1120A IEEE-488-translator + Fluke 2180 RTD digital thermometer + 9 probes. £350 all three items.
 H.P. 6181 DC current source. £150.
 H.P. 59501A - HP-IB isolated D/A/power supply programmer.
 H.P. 3438A digital multimeter.
 H.P. 6177C DC current source. £150.
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 H.P. 741B AC/DC differential voltmeter standard (old colour) £100.
 H.P. 6209B DC power unit.
 Fluke 80 high voltage divider.
 Fluke 431C high voltage DC supply.
 Tektronix M2 gated delay calibration fixture. 067-0712-00.
 Tektronix precision DC divider calibration fixture. 067-0503-00.
 Tektronix overdrive recovery calibration fixture. 067-0608-00.
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 Marconi TF2163S attenuator - 1GHz £200.
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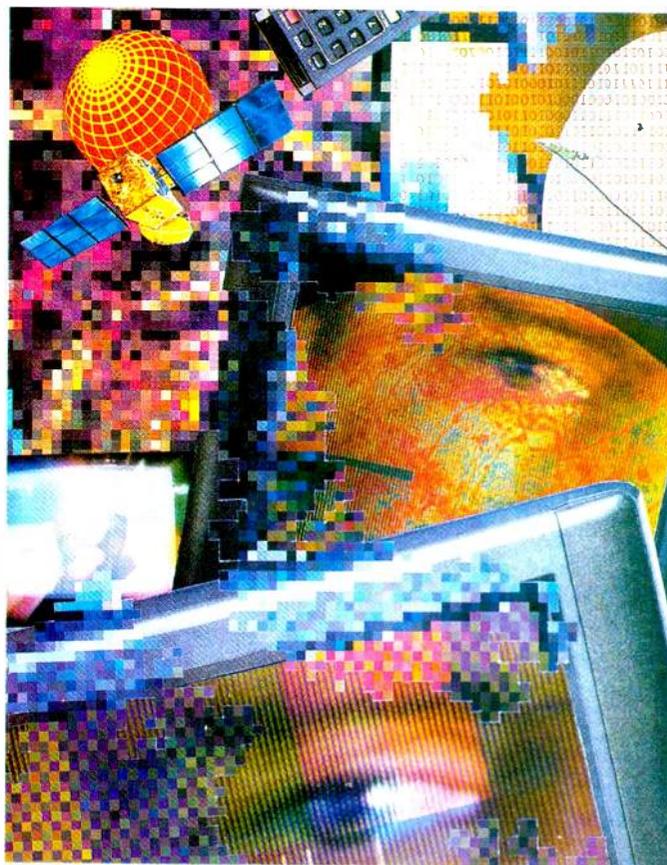
B&K 4812 calibrator head.
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 Racal/Dana signal generator 9082 - 1.5-520Mc/s - £800.
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 Racal 4D recorder - £350 - £450 in carrying bag as new.
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 Adret FX synthesizer 2230A - 1Mc/s. £250.
 Tektronix - 7S12-7S14-7T11-7S11-S1-S52-S53.
 Rotek 610 AC/DC calibrator. £2k + boot.
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 Marconi automatic distortion meter type TF2337A - £150.
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 HP4815A RF vector impedance meter c/w probe. £500-£600.
 Marconi TF2092 noise receiver. A, B or C plus filters.
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 Tektronix oscilloscope 485 - 350Mc/s - £500.
 HP180TR, HP182T mainframes £300-£500.
 Bell & Howell CSM2000B recorders.
 HP5345A automatic frequency converter - 015-4GHz.
 Fluke 8506A thermal RMS digital multimeter.
 HP3581A wave analyser.
 Philips panoramic receiver type PM7800 - 1 to 20GHz.
 Marconi 6700A sweep oscillator + 6730A - 1 to 2GHz.
 Wiltron scaler network analyser 560 + 3 heads £1k.
 R&S signal generator SMS - 0.4-10+0Mc/s - £1500.
 HP8558B spectrum ANZ PI - 1-1500Mc/s - o/c - £1000. N/C - £1500 - To fit HP180 series mainframe available - £100 to £500.
 HP8505A network ANZ + 8503A S parameter test set + 8501A normalizer - £4k.
 HP8505A network ANZ + 8502A test set - £3k.
 Racal/Dana 9087 signal generator - 1300Mc/s - £2k.
 Racal/Dana VLF frequency standard equipment. Tracor receiver type 900A + difference meter type 527E + rubidium standard type 9475 - £2750.
 Marconi 6960-6960A power meters with 6910 heads - 10Mc/s - 20GHz or 6912 - 30kHz - 4.2GHz - £800-£1000.
 HP8444A-HP8444A opt 59 tracking generator £1k-£2k.
 B&K dual recorder type 2308.
 HP8755A scaler ANZ with heads £1k.
 Tektronix 475 - 200Mc/s oscilloscopes - £350 less attachments to £500 c/w manual, probes etc.
 HP signal generators type 626 - 628 - frequency 10GHz-21GHz.
 HP 432A-435A or B-436A - power meters + powerheads - 10Mc/s-40GHz - £200-£280.
 HP3730B down converter - £200.
 Bradley oscilloscope calibrator type 192 - £600.
 Spectrascope SD330A LF realTime ANZ - 20Hz-50kHz - LED readout - tested - £500.
 HP8620A or 8620C sweep generators - £250 to £1k with IEEE.
 Barr & Stroud variable filter EF3 0.1Hz-100kc/s-high pass+low pass - £150.
 Tektronix 7L12 analyser - 1Mc/s-1.8GHz - £1500 - 7L14 ANZ - £2k.
 Marconi TF2370 spectrum ANZ - 110Mc/s - £1200-£2k.
 Marconi TF2370 spectrum ANZ + Tk2373 FX extender 1250Mc/s + trik gen - £2.5k-£3k.
 Racal receivers - RA17L-RA1217-RA1218-RA1772-RA1792 - P O R.
 Systron Donner microwave counter 6057 - 18GHz - nixey tube - £600.
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Clear highway for digital television



Although there are still no signs of it in the shops, a new TV technology will shortly make your video recorder and TV set obsolescent. Not only will picture quality visibly improve, but the availability of more programme channels will provide a new imperative to the broadcasters. Tom Woodford reports on a digital TV revolution.*

*Tom Woodford has spent much of his adult life in the semiconductor industry and now runs a consultancy specialising in communications projects, radio and data broadcasting. Someday soon he plans to move to France, to sail, write and grow tomatoes.

Current TV broadcast technology has its origins before the Second World War. It predates magnetic tape recording, radar, stereo, and digital watches. With a few minor improvements – like better resolution, some sort of colour, and less fuzzy sound – you watch TV which is little different from Baird's and Blumlein's.

In thirty years of research the only identifiable achievement was MAC, which Fortress Europe believed would protect it from the Far Eastern invasion when the drawbridge was raised in 1993. Proving again that democracy is a poor method of choosing winners, the real threat was not a land attack from the Orient, but a Sky-borne raid from the Antipodes.

While Europe discovered that DMAC was not the consumer dream of the 1990s, the Americans and Japanese were having the odd skirmish about high-definition standards. All was pretty quiet until a small US start-up, SkyPix, hit the market with a vapourware launch of a fully digital DTH TV broadcast system in mid-1991. SkyPix would deliver some 64 TV programmes direct to the home by packing five "or more" digital channels onto each of sixteen high-power satellite transponders. Cable operators and terrestrial broadcasters would be out of business overnight...

Suddenly hundreds of developments appeared in a rash of similar announcements. Presumably this was what SkyPix had hoped since, the day after its own launch to the press, it was rumoured to be phoning around the

industry trying to buy into any working system. Like schoolchildren in their first communal shower, the industry had unwillingly been exposed to itself, only to find that everyone had been developing the same things, none of which yet actually worked.

With the secrecy and embarrassment lifted, the world was thrust into a new television era. Digital compression became public knowledge, although still unattainable. Digital TV became a reality, while yet still undefined.

Digital video compression is unlike the data compression used (for example) in file exchange by modem or in PC hard-disk utilities like *Stacker* or *SuperStor*. PC Data compression uses coding techniques to eliminate redundant data. On recovery, the data is re-assembled *exactly* in its original form. However, in any system of video or audio compression some of the original information is deliberately thrown away in the coding process; it is lost forever and is never recoverable.

The advent of digital video had a reverse effect on programme makers, who returned to silver halide technology. The "Inspector Morse" TV series was filmed in 35mm; Super-16 format is more common. Producers of high-value programmes want to ensure that their work is stored for the future with minimum loss of quality. They feel that current electronic storage is inadequate.

Given that TV pictures are already a feeble representation of a real image, what is the enthusiasm for digital compression, and why has it suddenly become the technological hot spot?

Squeezing a bunch of lemons

In a universe with limited RF spectrum,

Squeezing a bunch of lemons

In a universe with limited RF spectrum, straight digital transmission is a relative non-starter as an alternative to analogue techniques. A 625-line video picture with a reasonable resolution of 400 vertical lines (S-VHS) requires 250,000 samples per frame. At 25 frames per second, and eight bits per sample, this will require at least 50 megabits per second to transmit. Add some colour and some sound, and 60Mb/s is not unreasonable, but this still only gives a high-end-consumer picture quality.

Full studio broadcast standards would require over 140Mb/s, or at least 150MHz of bandwidth. This compares with the 6MHz or so that a broadcast PAL signal needs in conventional vestigial-sideband AM. Nevertheless, the perceived benefits of digital transmission – picture accuracy, dynamic range and noise immunity – have made it a very desirable objective, and this has spawned a vast amount of research into data compression algorithms.

Much of a video signal is the same from frame-to-frame; in real life very few large changes occur in a twenty-fifth of a second. Obviously it should be possible to transmit only the differences between frames, and

avoid the redundant burden of repeated information. There are some potential logistic problems with this approach, particularly “appearances” (see box) but the major difficulty is the actual transmission medium itself. (For the time being, “transmission” will be taken to include storage for subsequent replay)

Simple compression techniques are successful where high transmission powers are available to guarantee adequate signal-to-noise ratios at the receiver. With any sequential, differential system, a burst of noise can mask a significant change in one area of a picture. This error will remain until the damaged area is subsequently updated, which could potentially be many minutes.

In the world of terrestrial TV it is commonplace to bung a few megawatts up a big stick to broadcast distances of a few kilometres; for satellites only fifty watts are available to transmit tens of thousands of kilometres. Conversely, the satellite engineer has over 36MHz of bandwidth to play with compared to his terrestrial colleagues’ eight or so, and multipath is not a problem with less received signal power than the sound of a burning candle.

In practice it is terrestrial and cable broadcasting, with severe multipath, adjacent-chan-

nel and local interference potential, where the greatest difficulties are found.

Broadcasters see it as essential that compatibility among all delivery methods is ensured, or the commercial burden of standards conversion will invalidate any likely benefit. This argument has been the justification for the many new, rival TV standards worldwide. *HDTV*, *HD Divine*, *Spectre*, *Sterne*, *System 2000*, *Race 203* and *Diamond* are only a very few of the more prominent. However, the recent pre-emptive strikes by *DirectTV* in the US and by Rupert Murdoch in Europe have forced an uncharacteristic unity, and the de-facto future standard has become MPEG by default.

It would appear that the American system has won, but in reality the world’s commercial grown-ups have just noticed that their technical children are still playing long after their bed-time, and have imposed some discipline.

MPEG plus, plus plus, plus minus

The Motion Picture Experts Group of ISO, the International Standards Organisation, grew out of JPEG, the Joint Photographic Experts Group. JPEG was formed to define standards for digital storage, processing and display of still pictures on the (then) growth technology

The mpeg data stream structure

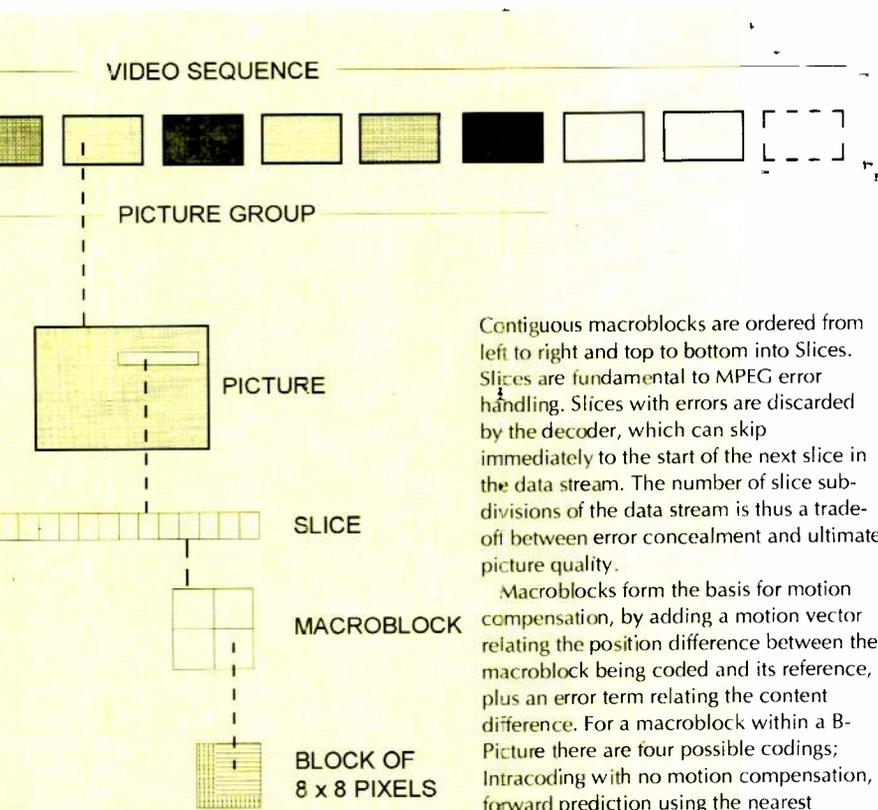
The general form of the MPEG stream comprises two layers. The System Layer contains timing and other information used to demultiplex the audio from the video and subsequently to synchronise the two.

The Compression Layer contains the compressed video and audio data streams. Audio coding will be covered in a later article.

Within the video stream there is an hierarchical data structure. The video sequence contains a sequence header code, one or more groups of pictures, and an end-of-sequence footer.

Picture Groups are strings of one or more pictures which allow access into a Sequence.

A Picture consists of three rectangular matrices representing the Luminance (Y) and the Colour Difference (Cb and Cr) values. The luminance matrix has an even number of both rows and columns. The colour-difference matrices are each half the size of the luminance matrix in both axes, so there are one each Cb and Cr for every four Y values. The Y and C values



are arranged as 8 x 8 sets known as Blocks, so a chrominance block describes a picture area four times bigger than a luminance block.

The Y and C Blocks are arranged into macroblocks, each of which contains one Cb block, one Cr block and four Y blocks.

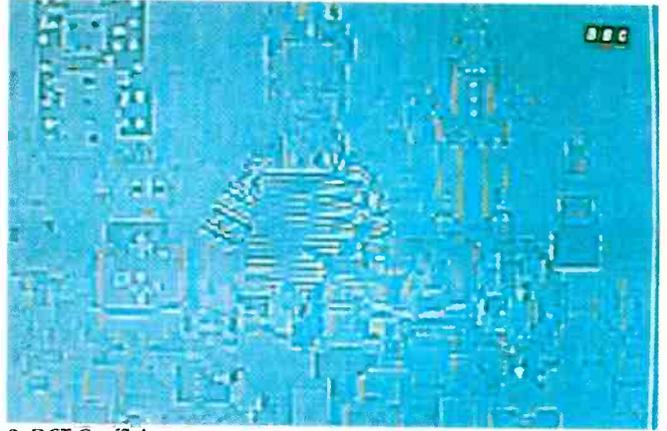
Contiguous macroblocks are ordered from left to right and top to bottom into Slices. Slices are fundamental to MPEG error handling. Slices with errors are discarded by the decoder, which can skip immediately to the start of the next slice in the data stream. The number of slice subdivisions of the data stream is thus a trade-off between error concealment and ultimate picture quality.

Macroblocks form the basis for motion compensation, by adding a motion vector relating the position difference between the macroblock being coded and its reference, plus an error term relating the content difference. For a macroblock within a B-Picture there are four possible codings; Intracoding with no motion compensation, forward prediction using the nearest previous I- or P- picture as reference, backward prediction referencing the nearest future I- or P- picture, and bi-directional prediction using the nearest past and future I- or P- pictures. Backward prediction can be used to overcome appearances by providing the uncovered, new information in advance.

that



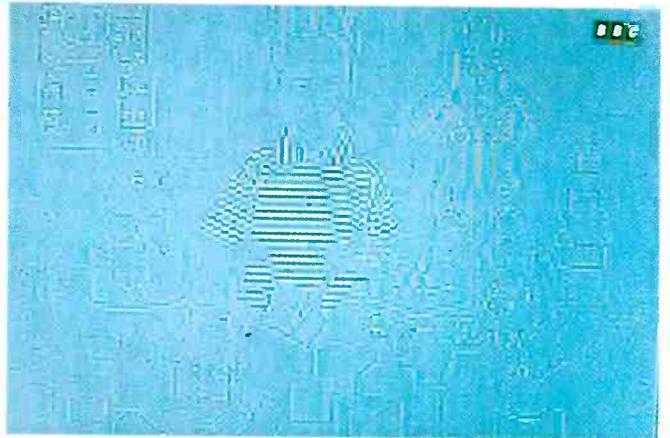
1. DCT Coefficient 8 x 8 DC Term



2. DCT Coefficient 8 x 8 Layer 1 contribution



3. DCT Coefficient 8 x 8 4 Coefficients



4. DCT Coefficient 8 x 8 Layer 2 contribution



5. DCT Coefficient 8 x 8 9 Coefficients



6. DCT Coefficient 8 x 8 16 Coefficients



7 (left). DCT Coefficient 8 x 8 64 Coefficients

DCT demonstration: This still image has been stored as a 720 by 576 image and cut into 90 by 72 blocks. Each block is an 8 by 8, frame based, pixel area. These blocks were discrete cosine transformed to produce 64 coefficients to 12-bit accuracy.

The series of images show the effect of individual contributions of selected coefficients to full accuracy in reconstructing the final picture.

The images are numbered from 1 to 7. The first image shows the result of using just the DC term. The second image does not display a difference image but the picture information carried in three coefficients of the first layer from the DC term. The third image shows the combined image of the previous two, a total of four coefficients. The images then progress through the layers until all 64 coefficients are used eventually to obtain a complete picture.

Paul Burfield, BBC Research Department.

of Personal Computers, and MPEG followed to continue the work in the area of pictures that moved.

Originally MPEG was an occasional, typical gathering of like-minded engineers, evaluating compression methods and discussing the various possible applications of digital video. Participants would attend one of the routine meetings bringing a tape of their latest video, which had defeated the current state-of-the-art compression algorithms. Everyone would have a jolly good laugh, and then go up the pub to chat about it. By this process of intellectual attrition the MPEG algorithms were steadily refined, until a level of performance was reached which the consensus agreed was adequate.

It is important to understand that these early developments originated in the PC environment, and were limited by the (late 1980s) standard PC bus and clock speeds with a maximum data rate of 1.5Mb/s. Most of the participants had their backgrounds in information technology, and their objective was to achieve a reasonable performance in reasonable timescales, aimed at the multi-media markets.

Hardware evolution progressed in parallel with the specification and standardisation process, so that demonstrable systems existed and kept pace with the theoretical work, in particular the essential operational silicon building-bricks. Like the DMAC development, the MPEG specification was an evolving summary of what had already been achieved, or was about to be.

By early 1991 the MPEG standard had reached a performance level which the majority of contributors considered adequate, and MPEG 1 was effectively defined, although still only in its "travelling draft" form. Multimedia consumer products were released to the market, in particular Philips CD-Interactive or CDi, although PC expansion cards offering MPEG-1 capability had already been available for some time.

Meanwhile, the professional TV broadcasters had been pursuing their own digital pathway. Starting from the minimum data rate of around 140Mb/s, their approach was to see how much this could be reduced while still retaining acceptability. By the late 1980s it was felt that 25 - 35Mb/s was looking promising, and working systems were built to prove the point to the doubters.

Casting around for available hardware, some developers "discovered" the MPEG-1 sub-systems and were delighted to find that much of the silicon would work at much higher data rates than the 1.5Mb/s for which it was specified. These engineers were looking to upgrade existing links for broadcast distribution networks, and had only a passing interest in consumer markets.

Simultaneously, the programme makers were evaluating the new digital technology, to establish its viability as a lower-cost replacement for VHS tape and to expand cable networks. They were stunned by the picture quality which could be achieved in only 1.5Mb/s, and started to wonder what might be possible

Discrete cosine transforms, run length encoding and quantization

The MPEG algorithms use numerous stages in the source coding process. The obvious first stage for a digital system is to digitise the analogue original. This implies inherent quantization into a finite number of pixels, each with a finite number of grey levels for the luminance information, and similarly limited chroma steps. The resulting array of pixels is then subdivided into blocks of 64 pixels, which are then translated to the frequency domain by Discrete Cosine Transformation.

DCT is usually explained in a heavily jargonised, esoteric form. This may be better understood by considering it as delta-modulation and coding in pairs; the image block is coded as a series of transitions from one grey level to the next. If the change in luminance between any pixel pair is large, the transformed equivalent pixel is white. If there is no change, the transform is black. There are obviously shades of grey in between these two extremes.

As luck would have it, the average video picture has few hard contrast transition edges in proportion to the total picture area, so the DCT version tends to have a large number of very black pixels, and the few, very white pixels tend to occur in specific groups. The statistics necessary to demonstrate this are beyond the scope of this box, but anyone who wants to prove the point and has a couple of years to spare can try pausing their home VCR on a typical tape, and then inspecting the resulting frozen video frame to perform a quick DCT in their head.

The resulting DCT form of a frame has large groups of all-black or all-white pixels, or long sequences of small change representations, so the frame can be further compressed by Run-Length coding in pairs.

In simple terms this means that, if there are 1000 identical pixels in a string, then one could transmit a coded message saying "here follows 1000 identical pixels, all

black", which is obviously a shorter message than all the 1000 original pixels, and a lot less boring.

For a smooth transition from black to white over a series of pixels, a similar coded message saying "the next 500 pixels increase by only one bit from one to the next..." is an equally efficient simplification.

The MPEG system interposes an additional step before run-length coding, by first zig-zag scanning the array of DCT pixels. This tends to produce long runs of similar or identical transformed pixels, and increases the RALC efficiency. Again, this is something to try for yourself at home on a typical TV frame or two.

The run-amplitude-length coding is also of variable word length, so that the more commonly occurring sequences have shorter codes, which further increases the coding efficiency. Conversely, some blocks need to be coded much more accurately than others. A smooth transition needs to be transferred as such, because the eye is more critical of harsh edges where they shouldn't occur than of blurred edges which ought to be sharp.

Since the original stage of the entire coding process is to divide the picture into blocks, it is vital that the edges of the blocks should join seamlessly in the final, re-assembled picture. The MPEG algorithm modifies the amount of quantisation for each 16 x 16 "macro-block" of pixels to overcome this, and this additional process is also used to allow smooth adaptation to changes in transmission bit rate.

Since this only describes the coding process used for one typical pair of video frames in very simple terms, you will not now be surprised that real-time MPEG encoding of raw video has only just become a reality. Indeed, one could be forgiven for wondering if there is enough raw silicon in the world to construct all the necessary MPEG encoders.

at only slightly higher data rates which would fit into their existing five or six megahertz bandwidth.

In the worldwide climate of de-regulation, telecomms operators were also evaluating digital TV, particularly for the embryonic pay-per-view and video-on-demand domestic movie services, but also with an eye on the very lucrative news gathering market. PTTs have transmission standards at 1.2 and 8 Mb/s, and thus they were also following the twin performance paths. Would 1.2Mb/s give acceptable quality for consumers, and could 8Mb/s be good enough for broadcast feeds?

TV setmakers, still smarting from the European MAC disaster and equally wary of the continuing unresolved acrimony over widescreen and high-definition formats, were

only too pleased to join in with anything that looked half sensible.

Apparently MPEG was already exploring all these areas, and it was natural for the various TV-oriented interest-groups to want to join in.

The original MPEG meetings typically attracted around fifty attendees. During 1992 the number of Representatives grew, and the gathering in Sydney earlier this year drew 270 Delegates. MPEG is now unofficially called MP/G, where the "I" stands for "Interested".

With this level of world-wide participation the speed of MPEG progress has accelerated, and now exceeds the ability of the politicians to keep up. With MPEG-1 effectively an established standard, although still officially a draft, the various Interested bodies have moved on ward to the next generation, MPEG-2.

Picture coding

The MPEG standard defines three type of pictures within the coded stream: Intra, Predicted and Bi-directional.

Intra Pictures, or "I-Pictures" are coded using only information present in the picture itself and therefore provide possible random-access points into the compressed video stream. I-Pictures use only moderate levels of compression coding, typically two bits per coded pixel.

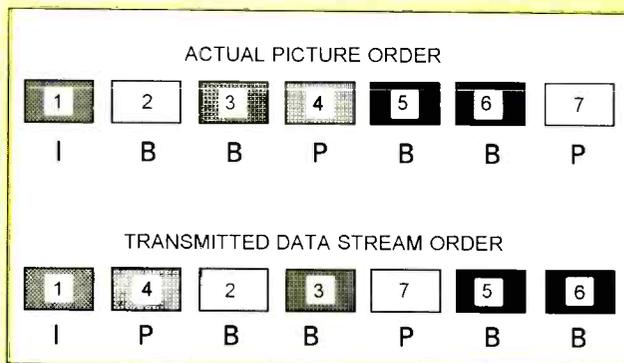
Predicted Pictures, or "P-Pictures" are coded with reference to the nearest, previous I-Picture or P-Picture. This is referred to as "Forward Prediction", and allows higher levels of compression than I-Pictures by using motion compensation. P-Pictures are used by subsequent P-Pictures and B-Pictures for prediction, and thus can propagate coding errors along the data stream.

Bi-directional Pictures, or "B-Pictures" use both a past and a future picture as a reference. By using bi-directional prediction some noise reduction is also achieved by the resultant averaging.

Although B-Pictures use the highest possible compression, they cannot propagate errors since they are never themselves used as a reference.

The MPEG algorithm allows the encoder to choose the number and positioning of I-Pictures within the data stream, depending on the original video composition (number of cuts and scene changes) and on the specific need for random accessibility.

Where random access is paramount I-Pictures occur typically twice per second.



The algorithm also determines the number of B-Pictures between any pair of P-Pictures or I-Pictures, depending on the video material content and also the encoder's memory capacity. Typically, two B-Pictures will occur between successive I- or P- pictures.

The MPEG encoder re-arranges pictures into a different order for transmission. Most importantly, the reference pictures needed to reconstruct any B-Pictures must be sent before the B-Pictures which actually use them as a reference. This re-arrangement or re-ordering of the video sequences requires complex re-synchronising with the sound signal in the receiver..

With the broadcasters coming down from 140Mb/s and the PTTs going up from 1.2Mb/s the inevitable clash occurred at around 15Mb. If the momentum of MPEG were not to be lost it was essential that some new Peg was put in the ground to mark the direction for onward development.

MPEG-1 silicon was already operating quite happily at 15Mb/s and faster, and was growing offshoots. The leader in MPEG silicon technology, Californian C-Cube Microsystems developed its *CL450* decoder with much of the actual algorithm soft-loaded as microcode. This has allowed amendments and upgrades to

basic MPEG-1 to be demonstrable almost as soon as it evolves.

This flexibility, and the higher data rates, could be used either to improve resolution or to reduce compression artifacts, or a bit of both. The speed upgrades had already become known as MPEG-Plus, and with multiple enhancements there is also MPEG Plus-Plus.

The confusion increased with the requirement that MPEG-2 should be backwards-compatible, and hence the parallel existence of MPEG-2-Minus...

At these various higher data rates, MPEG-1 has demonstrated performance equivalent to

existing systems. At 2Mb/s, picture quality is similar to domestic VHS, and approaches broadcast PAL standard at about 12Mb/s. In between is Betacam SP used in SNG, at around 8Mb/s. Since MPEG-1 already appears to have reached the various targets of the interested participants, what then is MPEG-2?

Until early September this was not a simple question since participants are party to non-disclosure agreements. The main improvements in MPEG-2 appear to be the possibility for flexible, seamless, dynamically variable data rates rather than the nominally fixed 1.5Mb/s of MPEG-1. In addition MPEG-2 makes better use of the temporal redundancy in interlaced TV pictures, which was not exploited in the non-interlaced IT origins of MPEG-1, and MPEG-2 is defined to be fully compatible with telecommunications networks data rates. Draft specifications of MPEG-2 are notionally available from standards bodies such as BSI, but these all date from earlier this year and are already obsolescent.

MPEG has about five meetings a year, with some ad-hoc gatherings in between as required. The Spring meeting in Sydney set out MPEG-2 in outline, and these standards were chilled at the July meeting in New York, in preparation for being frozen in the Korea meeting in early November.

At the International Television Symposium at Montreux in June the Europeans apparently capitulated, finally buried MAC and adopted MPEG as the future standard. A European Launching Group meeting was planned for September to lay out a co-ordinated plan for the introduction of MPEG-2 by mid-1997. It appeared that America had indeed won the digital race.

Timing and synchronisation

The MPEG system performs large amounts of signal processing before transmission and after reception but before display. Although the audio and video components are multiplexed together in the data stream, they use totally different coding techniques. In particular, the video frames must be transmitted re-ordered, so the audio and video data components at any point in the stream may be totally unrelated. The MPEG standard provides suitable timing references to allow the audio and video to be re-synchronised by the decoder.

The System Clock Reference is a sample of the 90kHz MPEG system clock. This provides 7.8×10^9 clocks in a 24-hour day. An SCR is transmitted as a 33-bit value, so can therefore identify any one particular clock cycle in a 24-hour period. SCRs are inserted into the data stream by the encoder at least once every 0.7 seconds. The system decoder in the receiver extracts the SCRs

and passes them on to the video and audio decoders to update their own internal clocks.

A potential time difference of nearly a second between the video and audio would be totally unacceptable, so the system also transmits Presentation Time Stamps. These are effectively 33-bit "clapper-board" markers which can identify related sound or vision Presentation Units to within one system clock tick. PUs are specific decoded video pictures or decoded audio sequences.

The encoder inserts a PTS into the data stream at least once every 0.7 seconds to identify specific video frames and audio sequences which must occur together. The video decoder checks incoming PTS markers against the current SCR before displaying a picture. If the PTS is "early" the previous picture is repeated; if it is "late" a picture is discarded.

Mapping out a digital highway

However, in a pre-emptive strike at the end of August, Rupert Murdoch announced that BSKyB would operate an MPEG-2 digital service by the end of next year. Sky will work jointly with the UK's National Transcommunications Ltd (the old IBA's technical wing), Comstream, Thompson and News Datacom, who provide the existing *VideoCrypt* conditional-access system. The Murdoch digital service will therefore be virtually identical to the Hughes *DirectTV* system in the US, scheduled for launch in mid-'94. Thompson and C-Cube are the technical and manufacturing resource behind *DirectTV*.

BSkyB will undoubtedly use the potential compatibility to good effect. In late August Murdoch purchased Delphi Internet Services, one of the top five US on-line data suppliers. He has announced deals for interactive systems with telecom operators, particularly BT, and with the Kirch Group in Germany for Pay-per-view services. Significantly, all the selected partners are commercially independent; there are no government influences or state monopolies in BSKyB's Digital future.

NTL's involvement follows from its *System 2000* digital transmission system, which they have been selling highly successfully as a broadcast feed since late last year. The development for Murdoch will probably be entirely new, as an MPEG-2-based consumer system. The contract expects hardware in the retail market no later than the end of 1994. With the final stages of MPEG-2 definition only expected after the meeting next November, these are tight timescales indeed for NTL, an organisation with limited experience in consumer product technology.

C-Cube is confident that it will have working MPEG-2 silicon available by Christmas, and volume product at consumer prices (under \$50) early in 1994. The apparent lead of C-Cube may be because R & D head Didier Le Gall is also chairman of the MPEG video committee.

NTL's significance is real-time MPEG encoding. At the end of 1991 the best available encoders took around 200 minutes to produce just one minute of programme output. By mid '92 this overhead time was down to about 70 minutes, and below 15 minutes by a year ago. While a dramatic improvement, this was still quite impracticable for any broadcast system.

Encoding was also a two-stage process, with a first pass to identify and edit out any hot spots which would defeat the MPEG algorithms. However, NTL was demonstrating real-time encoding before the end of last year, and shipping product to customers soon after. C-Cube did not deliver real-time encoders until last month, and at reputedly very much higher prices.

It is difficult to understand why BSKyB should want to move so quickly to digital, with a dominant installed base of PAL/*VideoCrypt* customers. Perhaps it is significant that none of Murdoch's chosen associates are manufacturers; they are developers

Movement prediction and appearances

Quite a lot of the information in any given TV frame is similar to a previous or subsequent frame. The MPEG system uses this sequential redundancy to transfer some pictures in a simplified form representing only their differences from some reference picture previously or yet to be transmitted.

This technique has excellent potential for problems, in particular with visual discontinuities known as "appearances".

The classic example is a static picture containing a black circle, which is actually the mouth of a tunnel from where, suddenly and unpredictably, bursts an express train. More subtle is the side view of an aircraft's wheels at touch-down when they (suddenly and unpredictably) start to rotate, accompanied by a puff of smoke. The opening of a door to reveal a hitherto undisclosed scene beyond is less sudden, but equally unpredictable.

Conversely, it is easy to detect consistent movement within a sequence of pictures by seeking groups of blocks which change only in their relative position to the overall picture from frame to frame. This can be

used to predict the future position of a moving group of blocks, and code a sequence of frames accordingly. This technique is known as "Movement Prediction" or "Motion Compensation", and it can improve compression ratios by a factor of three or higher. It can also create more serious problems...

I have actually seen a video of a horse race finish in which, in the digitally compressed version, a different horse wins. A wide-angle view of an applauding crowd can also create intolerable (but amusing) artefacts, since the overall movement of countless clapping hands is virtually random. Stormy seas or fields of wheat are equally incoherent.

A predicting algorithm needs to be able to decide when to give up trying and to wait around and see what happens next. The overall algorithm also needs to transmit whole pictures every so often, to allow the decoder to recover from transmission errors which would otherwise propagate endlessly from frame to frame.

and specifiers. While the litigation between the old BSB and its DMAC receiver suppliers still rumbles on, BSKyB is being careful not to incur any implied commitment to eventual sales of hardware. It is likely that Murdoch is protecting himself against the future, by trying to force it to look the way he wants. When, and how quickly, he follows his own lead will be intriguing.

Perhaps the biggest concern is the short-term replacement of VHS tape by Video CD. Nimbus is planning to have its add-on video decoder for audio CD machines on the market by the end of 1993. The Gang of Four (Philips, JVC, Sony and Matsushita) plan to enter the market with their rival "white book" CD-K machines early next year. Whichever system wins, either promises a drastic reduction in the cost of video rental by eliminating cheap, back-room piracy. With a common format on both sides of both the Atlantic and Pacific, the costs of programming must also fall. The quality improvement over tape will be dramatic and immediate for the ordinary consumer, with a price penalty which is low, or even zero.

The installed base of domestic TVs which can accept direct RGB inputs is small. Domestic RGB VCRs are non-existent. As with DMAC, the majority of viewers will watch their MPEG-broadcast TV as conventional PAL, SECAM or NTSC via composite video or even UHF. To time-shift a satellite movie they will certainly have to use an inferior analogue medium. Without the cost-

reduction benefits of digital, and extra features like video-on-demand and interactivity, it will become increasingly difficult for broadcasters to compete with the better quality and established convenience of renting a movie on CD from the local video store.

DirectTV and *Digital Sky* will be satellite-based, of course. With restrictions on coverage areas and due consideration for domestic consumers, the last European PAL transmitter will probably not be switched off until 2015 or even later. Much work has yet to be completed on alternative terrestrial transmission technology, although it looks as if NTL is again the leader with its OFDM *Spectre* developments. Whatever system is adopted to replace AM for terrestrial broadcast, it will be delivering digital programmes created using (by then) long-established technology. In twenty years' time every home could have a fibre-optic connection, and UHF broadcast may never be upgraded at all...

At Montreux, it was suggested that MPEG actually stands for Maximising Profit and Efficiency Gains. MPEG offers the ability for broadcasters to pack five, six or more TV channels into the space occupied now by only one. As far as the average TV viewer is concerned, this will mean only More, not Better, until he has saved up to buy his new MPEG television in some three years' time. One wonders how much extra he will have to pay for spiked feet? ■

CIRCUIT IDEAS

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Economical 27MHz phase modulator

This circuit phase-modulates a clock signal, a process finding application in PM and FM transmitters and in clock jitter testing. It exploits the properties of 74HC CMOS gates that (a) input logic threshold increases with increasing V_{cc} , (b) propagation delay increases with increasing V_{cc} and (c) V_{cc} is allowed to vary between 2V and 6V.

Triangular modulation of 1Vpk-pk is superimposed on the nominal 4V derived from the 4.5V battery via the diode, which catches the modulating input and results in a V_{cc} varying between 4V and 5V. There is sufficient noise immunity in 74HC and 74AC logic to allow correct drive to the following stage with these levels.

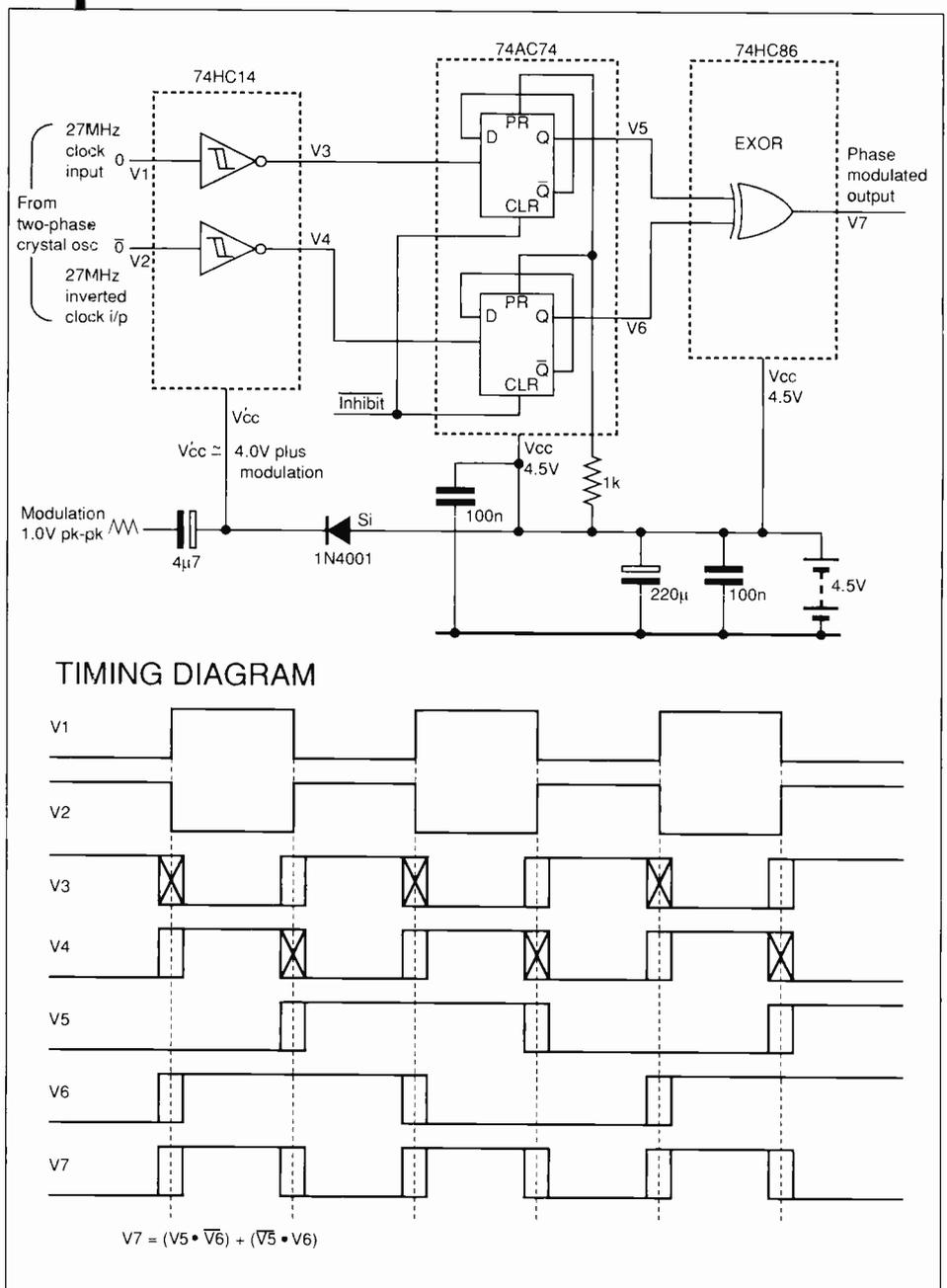
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Since the effects of varying V_{cc} on propagation delay and logic threshold are in the same sense only on falling edges of the clock input, both inverted and non-inverted clock signals are used, the timing from the rising edges being eliminated in the 74HC74, configured as a pair of divide-by-two flip-flops. A 74HC86 Ex-Or section combines the two outputs V_5 and V_6 to provide the circuit output at up to 27MHz.

Two of the spare gates in the 74HC86 will make the 27MHz crystal oscillator and the triangular-wave generator was made from spare Schmitt inverters, with a buffer, at audio frequency. A 4.5V supply is conveniently obtained from three alkaline manganese cells. I found it necessary to use a ground-plane board layout.

Laurence Richardson
Hersham
Surrey

At very low cost, an audio waveform phase-modulates a clock signal of up to 27MHz. Only negative-going input edges are used, since the properties of the CMOS logic exploited in the circuit have a tendency to cancel on rising edges.

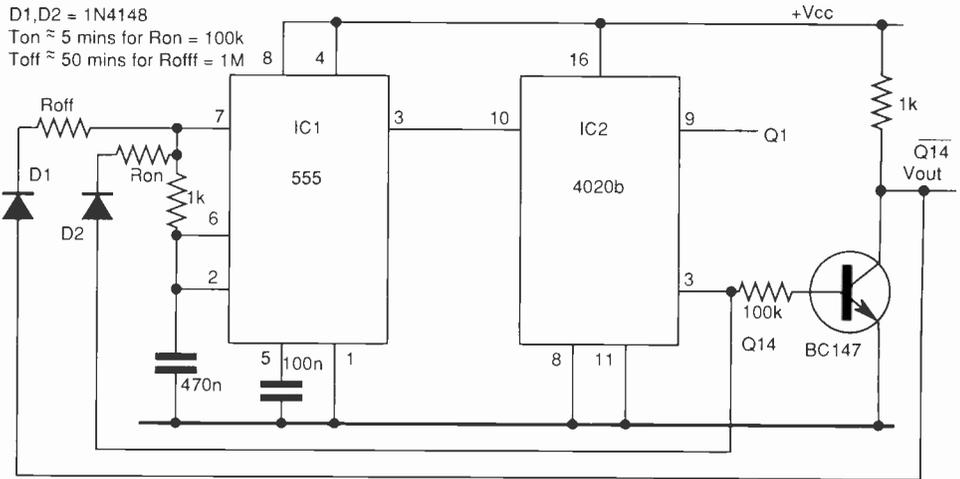


Independent on/off for long-period astable

Astable multivibrators using 555 timer ICs are flexible in that frequency and duty cycle are independently adjustable but, when long and independent on/off periods are needed, large and costly capacitors must be used. This circuit avoids the problem and gives periods up to several hours.

An 4020 14-stage binary counter divides the timer output, its Q₁₄ and Q₁₃ outputs connecting R_{off} and R_{on} at alternate transitions to give independent on and off periods. Output Q₁ gives an indication of the output duration.

Devadoss John
Hindustan Cables Ltd
Hyderabad
India



Low-loss lamp dimmer

Once having shut the car door in a dark car-park and thereby extinguished the internal light, you can't find the keyhole to lock it. But this circuit dims the light slowly until the door is locked.

Transistor Tr₂ drives the lamp and in turn is driven by the 3524 regulating pulse-width modulator. With the door switch closed (door open), C₃ is fully discharged, Tr₁ fully conducting and C₁ shorted. Pin 2 of the PWM, the non-inverting input, is high and the lamp fully on.

Closing the door and thereby opening the switch causes C₃ to charge through R_{5,6}, holding Tr₁ in saturation for about a minute. When Tr₁ cuts off, C₁ charges slowly through R₇, bringing pin 2 of the PWM to 0.6V, and slowly dimming the lamp.

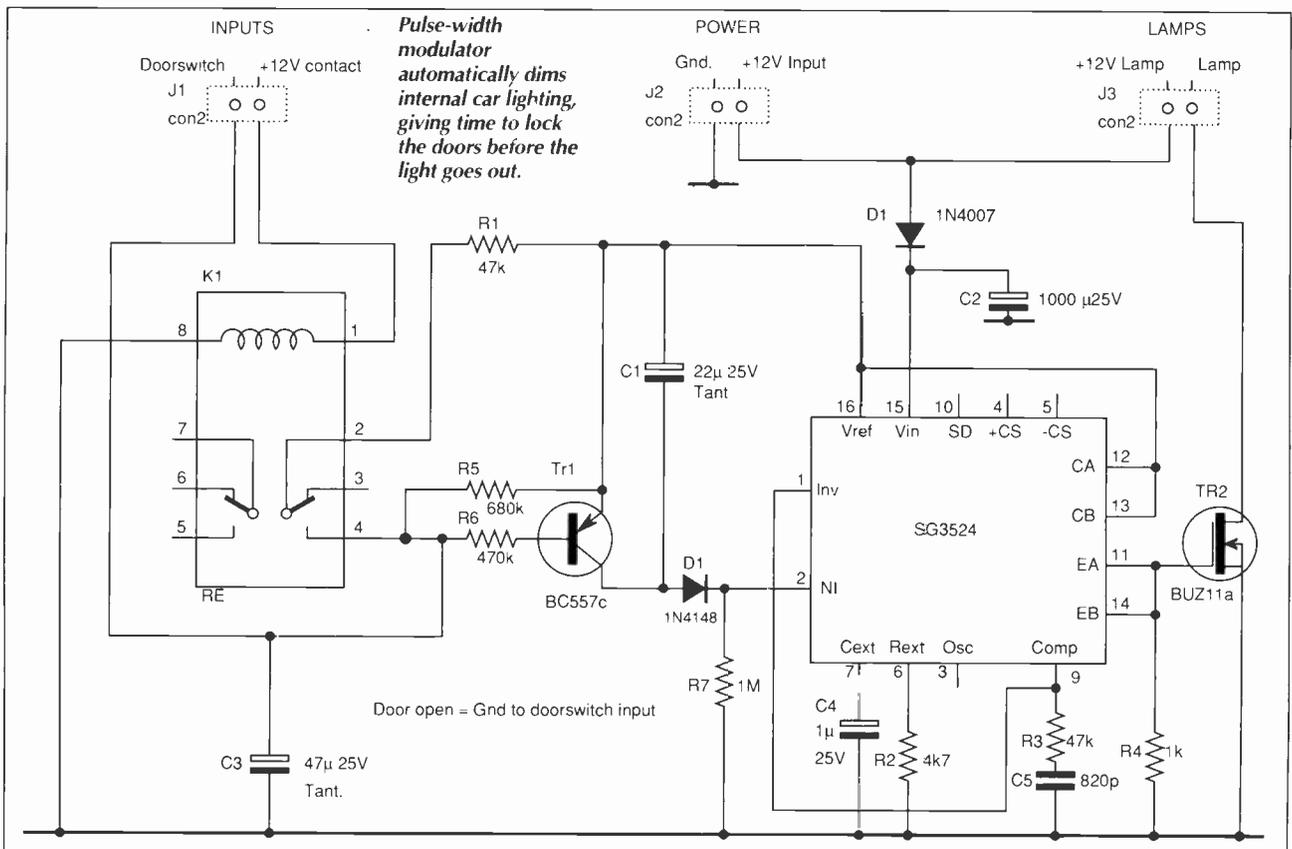
If matters were left there, one would have to wait for the lamp to go out completely, which would take some time, so the relay is actuated by the 12V contact on the lock to charge C₃ quickly through R₁. Diode D₁ and C₂ prevent interference when the engine is being

started.

The driver transistor needs no heat sink since it runs in hard switching.

Yves Delbrassine,
Hovenierslanden 3,
8200 Bruges,
Belgium

This principle is, of course, not limited to dimming car lighting; low-loss drive for high-power loads is common, although it is not often seen described for such simple projects. Among the possibilities that spring to mind is a miniature crystal oven using the PWM to drive a power transistor - Ed.



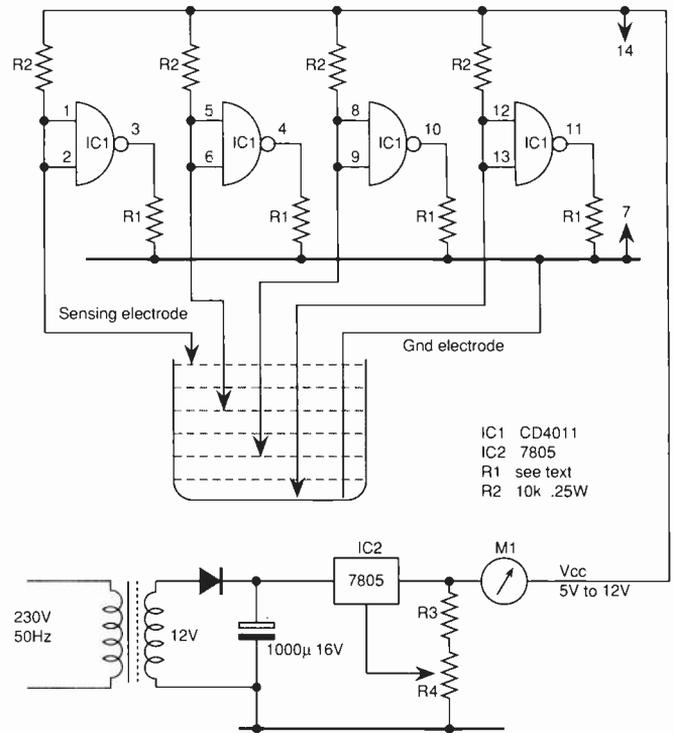
Two wire level indicator

Using one IC, this two-wire remote-reading instrument indicates that water in an earthed vessel is at or above one of four levels.

If all sensing electrodes are unwetted, all Nand outputs are low and the meter reads virtually zero. As an electrode touches water, its gate goes high and the meter reads a current V_{cc}/R_1 . A rising water level causes more gates to contribute to the current.

The optional regulator and voltage-setting resistor R_4 allow full-scale to be adjusted by shorting all the electrodes to ground.

K N N Narayanan and C V Raman Nagar
Bangalore
India.



Proportion indicator

To show the ratio of one of several inputs to the total of all inputs as a percentage, the obvious solution involving a microprocessor can be simplified by the use of three A-to-D converters, which perform all the functions of digitising, ratio determination and driving the percent display.

The non-inverting, summing op-amp pair adds the inputs — three, in this case, as an illustration — the sum being taken as the reference voltage for all A-to-Ds and the individual inputs being fed to $V_{in(hi)}$ of each converter. Display 1 now reads

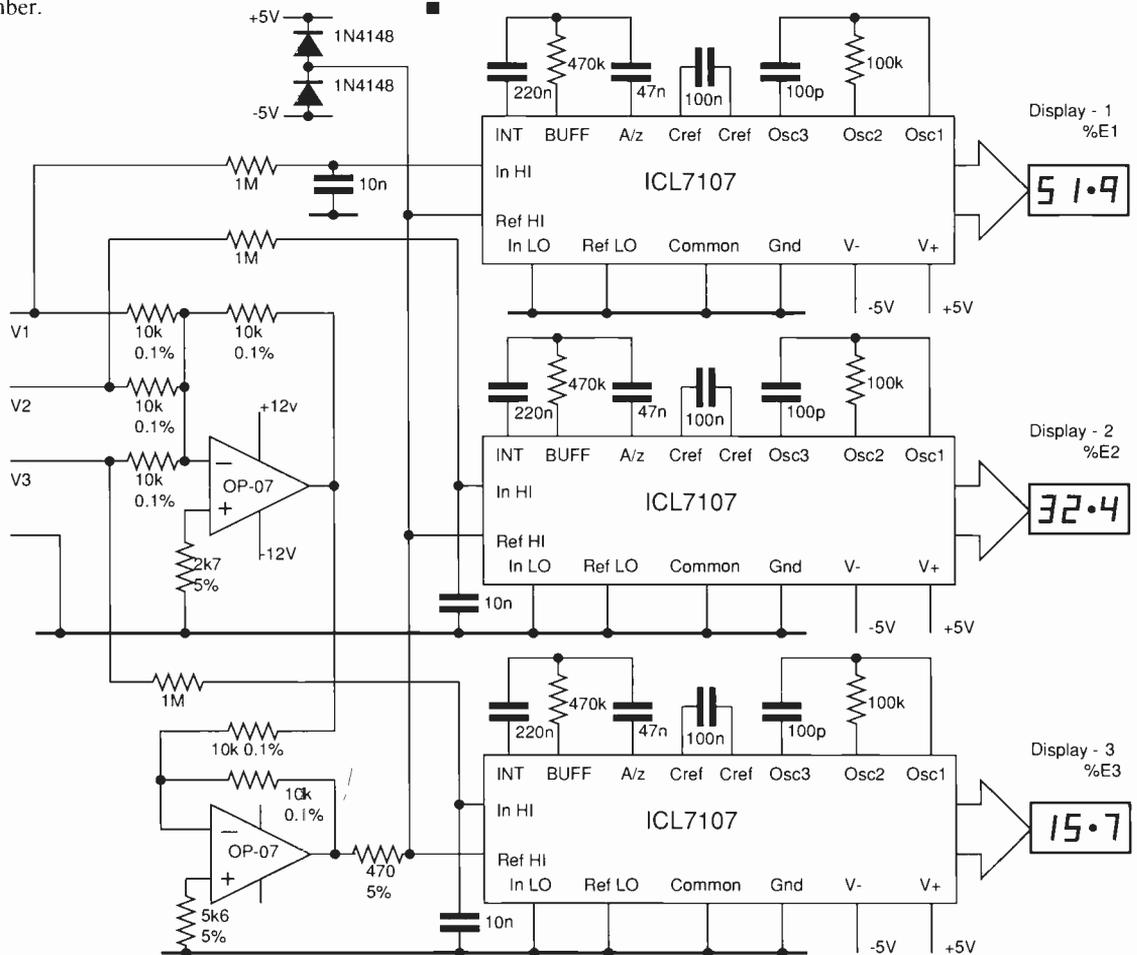
$$\frac{V_{in(hi)}/V_{ref(hi)}}{or}$$

$$\frac{V_1}{(V_1 + V_2 + V_3)},$$

and similarly for the other inputs. If the right-hand decimal point in the "tens" position is turned on, the display shows a percentage.

This circuit will handle a total input of 0.2-4.5V and give three readings per second; although three inputs were used in the prototype, it should work with any number.

M S Nagaraj
ISRO Satellite Centre
Bangalore



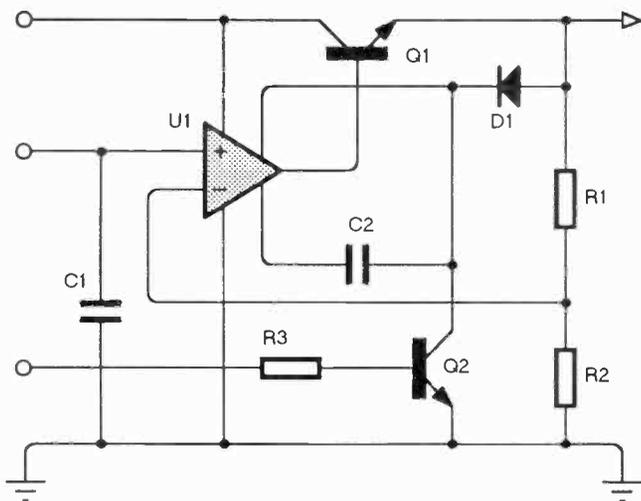
Indicator shows ratio of each input as a percentage of total input. Circuit replaces more exotic microprocessor-based arrangements.

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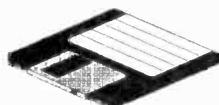
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CIRCLE NO. 107 ON REPLY CARD

Many op-amp applications process signals which swing about ground. This requires the designer to add a negative supply rail to the system. Frank Ogden reports on a new op-amp package which generates its own.

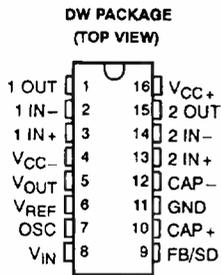
Charge pumped op-amp supplies the missing rail

Most portable communications and instrumentation designs must draw their power from a single battery or supply rail. The input signals which they process are usually referenced to ground. Logic level input signals pose little problem; they swing from ground towards the positive rail allowing the signal conditioning circuitry to operate from a single supply. Provided of course that there are enough volts on the supply rail to provide headroom for the signal being processed.

Most other signals swing around ground. Designers traditionally deal with these either by using CMOS op-amps such as the *TLC271* which include the negative rail in their common mode range or by putting in an extra power block to provide a negative supply rail referenced to ground.

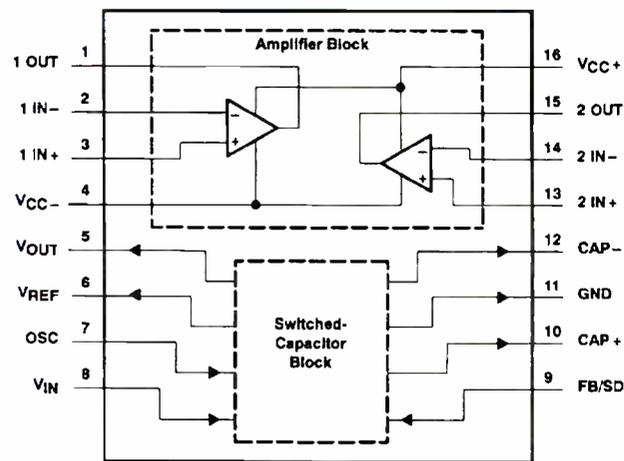
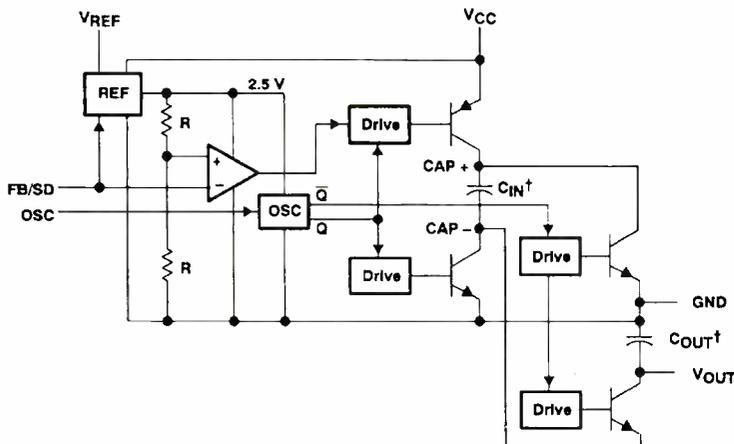
Using op-amps with extended common mode range is not always a good idea. They will tolerate a below-ground swing of just a few hundred millivolts. Exceeding this, for instance during a transient, incurs a real danger of device latch-up. However, the provision of a proper negative supply rail allows fully symmetrical signal handling and extended dynamic range at the expense of power supply complexity and board area.

Texas Instruments provides an answer to this dilemma in the *TLE2662*. Incorporating a pair of high output drive, jfet input op-amp blocks together with an adjustable charge pump negative rail generator, it allows the design of single 5V rail signal processing system with an input/output drive capability of 7Vp-p about ground within a single device package. The allowable supply voltage range of the converter system extends to 15V providing the amplifier blocks with $\pm 15V$ supply rails and 25Vp-p output swing about ground.



TLE2662 pinout

Charge pump converter internal block diagram



Functional block diagram

Suitable for modems, PCMCIA cards, portable phones, VCOs, level shifters, battery chargers and data acquisition systems, the device comes in a wide bodied SMD package.

The amplifiers

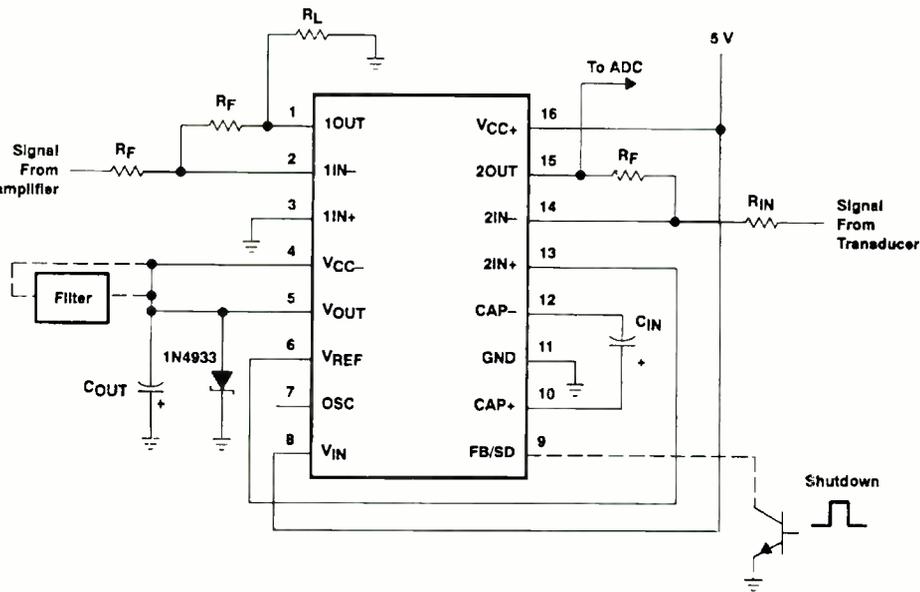
The identical amplifier blocks exhibit fairly typical jfet op-amp performance with 2V/ μ s slew rate, 25mA output drive capability and a millivolt of input offset voltage. However input bias current is just 3pA making the amplifiers highly suitable for electrometer type applications. The equivalent input noise current is also very low, about 1fA/ $\sqrt{\text{Hz}}$. Full output swing bandwidth extends to 140kHz at a supply voltage of $\pm 5V$. Quiescent no load current drain is about 600 μ A.

The voltage converter

The bipolar charge pump voltage converter doesn't share any pins with the amplifier blocks allowing total system separation if required. It will deliver up to 100mA provided that dissipation considerations are taken into account. Since the associated amplifiers would draw no more than about 6mA in small signal, high impedance load applications, the converter may be configured to supply a useful amount of current to other parts of the system. An extra pair of external diodes will allow its configuration as a boost converter to increase the available positive rail voltage to the amplifier section or other external systems.

The converter block also includes a 2.5V reference which may be used to regulate the output voltage through a resistive divider associated with a feedback/shutdown pin, or as an external reference for other circuitry.

Fig. 1. Basic application circuit showing a direct connection between the amplifiers and the internal negative rail power supply. Low level signal conditioning applications would normally require extra decoupling between converter and amplifier sections.



The power converter generally operates most efficiently at the internal oscillator frequency of 25kHz. However, this frequency may be raised, lowered or locked to an external clock with the appropriate placement of an additional external capacitor. The converter may also be strobed for interleaving with data acquisition operations. It can also be temporarily shut down. Converter supply current reduces to about 80µA under shutdown conditions.

Design considerations

Although the circuit blocks of the TLE2662 allow great versatility in system design, the high switching currents present in the converter section require careful handling. Many applications will be able to use a direct connection between the power converter and amplifier section: this basic arrangement can lead to 25kHz switching ripple appearing on the op-amp outputs amounting to a few tens of millivolts. A simple LC filter comprising a 50µH surface mount inductor feeding a 220µF capacitor in each amplifier supply leg will reduce output ripple by a couple of orders of magnitude.

Alternatively, the strobe facility on the power converter section may be used to stop the oscillator during sensitive data acquisition operations.

Attention should be given to decoupling adequately the main power supply to the converter at pin 8, VIN, particularly where long circuit tracks are involved. Also bear in mind that the converter ground return, pin 11, carries switching currents which could upset signal circuits if taken to a signal grounding point.

Typical application

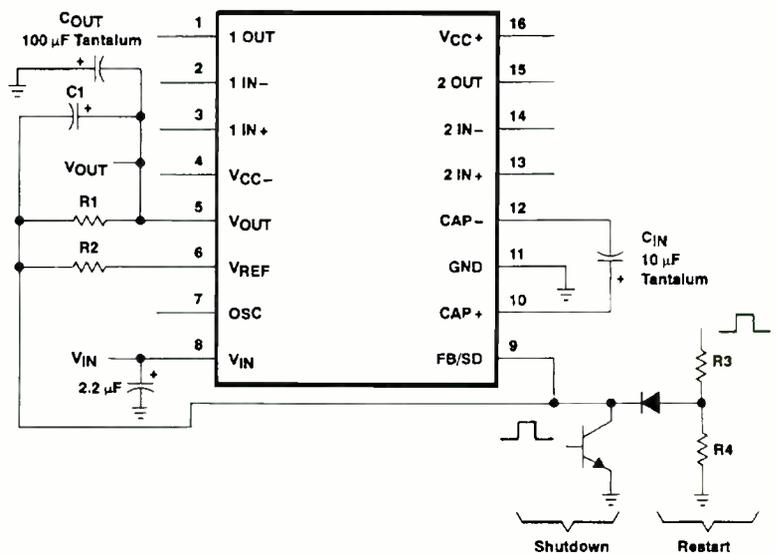
In its most basic configuration, the switched capacitor section of the TLE2662 provides the negative rail for the amplifiers in a single supply system. As shown in Fig. 1, the 5V positive supply is connected to VCC (pin 16) and VIN (pin 8). The negative output voltage, Vout, from the charge pump (pin 5) connects to the amplifier negative supply point, VCC- (pin 4). Only three external components are necessary not counting the components associated with the signal handling functions of the amplifiers: storage capacitors Cin, Cout and a small Schottky diode to prevent Vout rising above ground during startup. Once the negative rail is established, it provides no further function. As shown one amplifier is connected an inverter driving a resistive load while the other forms part of an ADC system using the TLE2662 internal reference.

The second circuit, Fig. 2, shows the converter section

set up to provide a regulated negative output voltage rather than the simple of dual of the supply input voltage. Its value is determined by the ratio of R1 to R2 and, assuming a Vref of 2.5V, is given by:

$$R_2 = R_1 \left(\frac{V_{OUT}}{\left(\frac{V_{REF}}{2} - 40mV \right)} + 1 \right)$$

R1 should be 20kΩ or greater since the reference current is limited to ±100µA. R2 should be in the range 100kΩ to 300kΩ. Frequency compensation is accomplished by adjusting the ratio of Cin to



Cout, a normal ratio of about 10:1. Capacitor C1, required for good load regulation, should be 2nF for all output voltages.

TLE2682

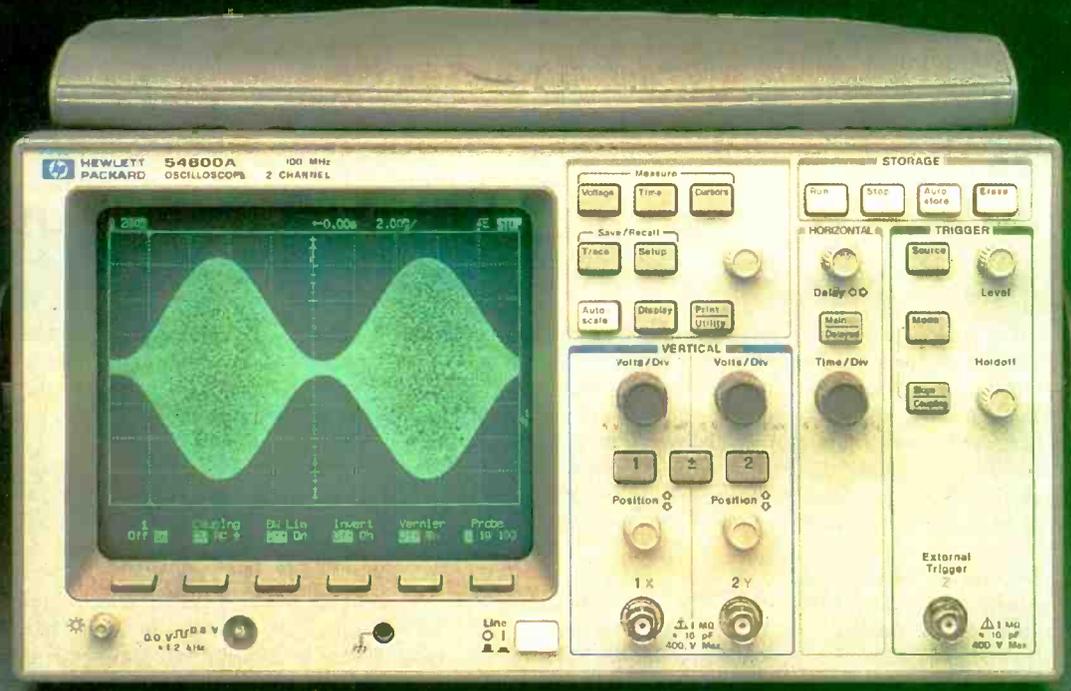
This is a higher speed derivative of the 2662 offering a typical slew rate of 40V/µs and a gain bandwidth product of 10MHz. For instance, settling time to 0.1% on a 10V step (1kΩ/100pF load) is about 400ns. In other respects it has similar characteristics to the TLE2662. ■

Fig. 2. The basic regulator function. Vout referenced to ground (pin 11) must be less than the total of the supply voltage less the voltage drop across the internal charge pump switches.

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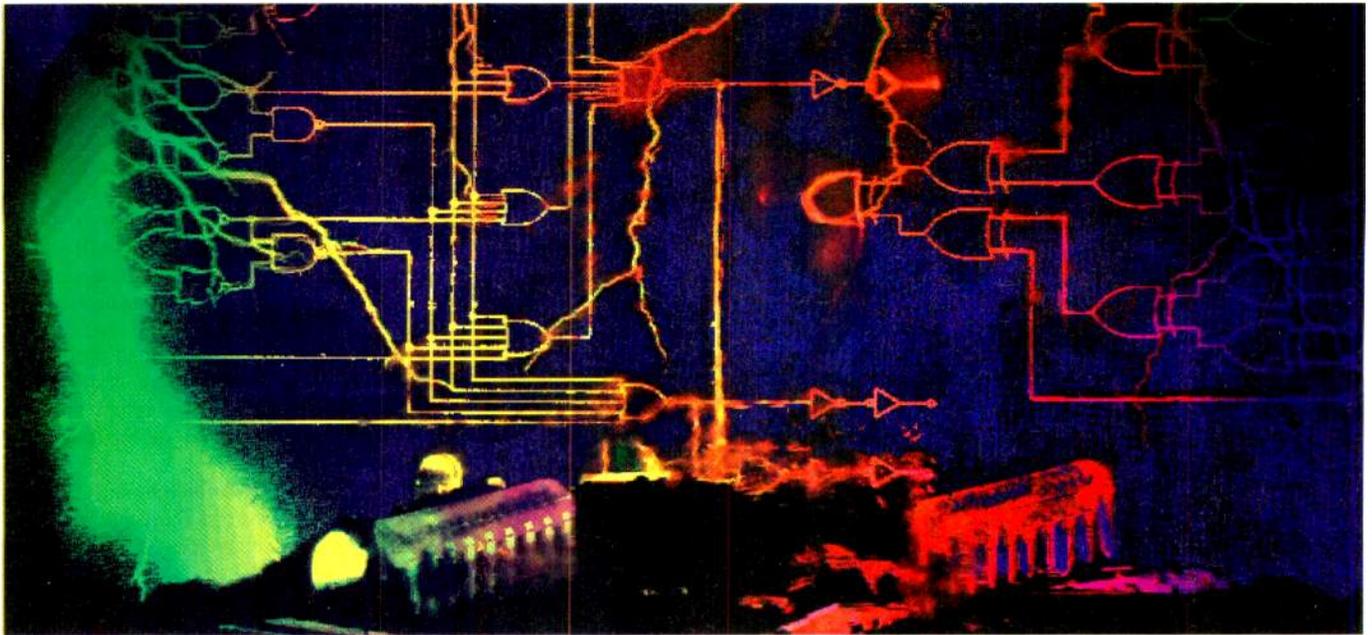
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TRACY MARTIN

WORKING

with programmable logic

*Programmable logic devices are essential to compact logic design and have been so for a long time. They save component count and board area and provide adaptability of circuit function. Despite this utility their operation is not always well understood. In the first part of a new series, logic designer **Geoff Bostock** explains the workings of programmable logic.*

Programmable Logic has come of age. It is at least 21 years since the first proms (programmable read only memories) came to the market.

It may seem perverse to start a series on programmable logic by mentioning memory devices, but one of the most common uses of proms is as logic devices. After all, every combination of input signals to a prom will result in a well defined set of output signals – exactly how a combinatorial logic device might be defined. Conversely, even a 2-input nand gate could be described in ‘memory notation’ as:

ADDRESS	DATA
0	1
1	1
2	1
3	0

One prom manufacturer, Monolithic Memories Inc. (now a subsidiary of AMD), even went as far as calling proms PLEs (programmable logic elements) when used in logic applications. If proms can be used as logic devices, why is there a need for pals, GALs, FPLAs and all the other PLDs (programmable logic devices) which have mushroomed onto the market? Indeed, why not just use TTL or CMOS logic? By the end of the series I will hope to have answered these questions.

Proms as logic devices

Examination of the prom structure, **Fig. 1a**, shows how they can operate as logic devices. The 5-input/8-output prom is a real device marketed by AMD as a 27S19, National as a 74S288, Philips as 82S123 and Texas as an 18S030, etc. The five input signals are fully decoded into 32 internal lines. Each line drives eight transistor bases; their emitters are connected via fuses to the input side of one of the eight output buffers.

Viewed as a memory, the five inputs define an input address; this sends one of the internal lines high. Any of the outputs connected via an intact fuse will be pulled high also; the inverting output buffer making a low appear on the corresponding output pin. Conversely, a blown fuse will let the output line stay low, sending the output high. Thus the pattern of blown fuses determines the output data corresponding to any particular input address.

The alternative way of describing the prom is to call the input decoder a fully decoded AND-array, and the output section a programmable OR-array. Line 0 in **Fig. 1a** is thus:

!A4 & !A3 & !A2 & !A1 & !A0

where ! represents an inversion and & the AND function. The lines driving the output buffers are wire-ORed, where the fuses are intact, so an output function formed by leaving the fuses

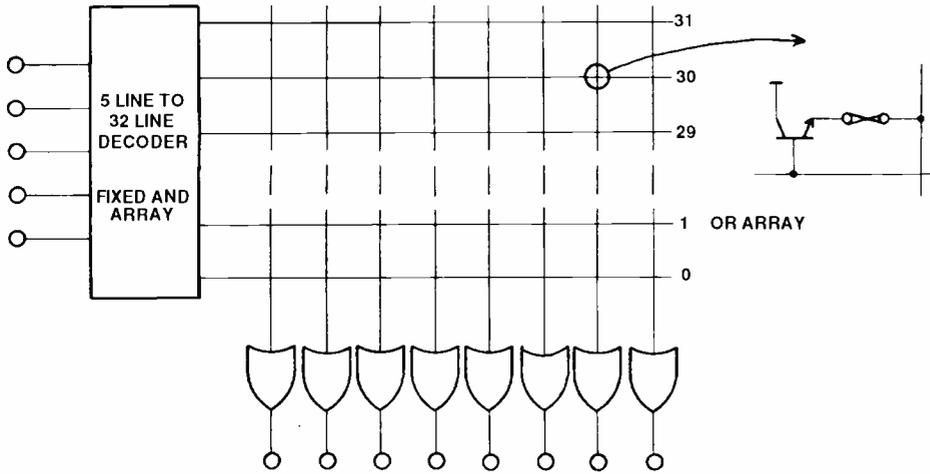


Fig. 1a. Prom architecture

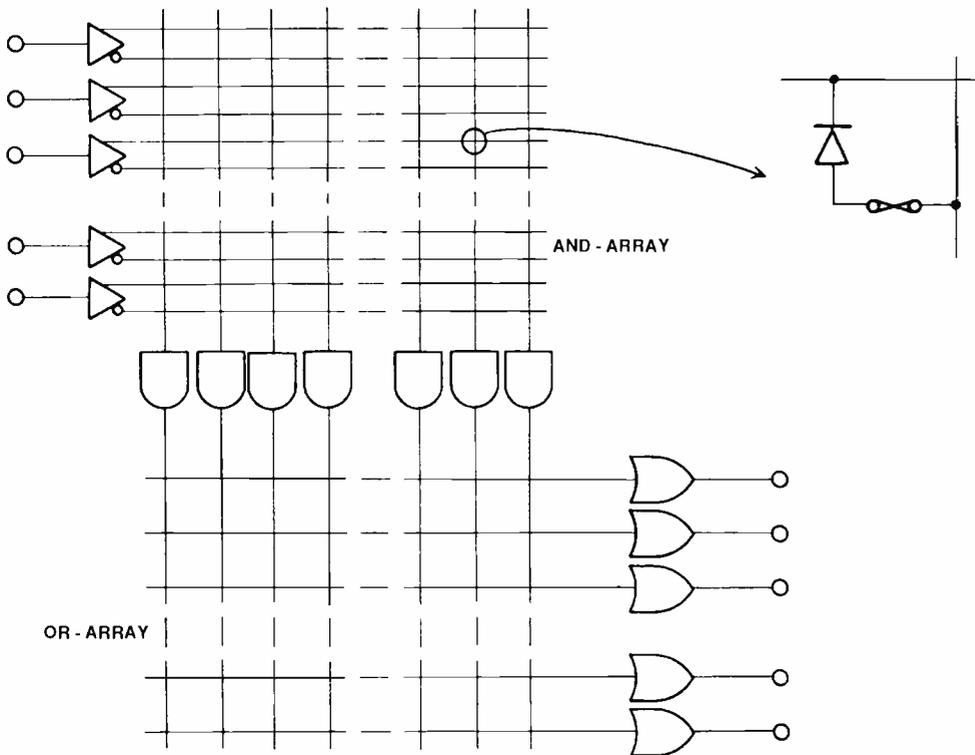


Fig. 1b. FPLA architecture

on lines 3 and 17 intact, but blowing out all the others could be written:

$$O0 = \overline{!A4 \& !A3 \& !A2 \& A1 \& A0 \# A4 \& !A3 \& !A2 \& !A1 \& A0}$$

where # represents the OR function. Note that O0 is inverted, because the output buffer is inverting.

We could equally well represent this logic function by a Karnaugh Map. Every cell in a five-bit Karnaugh Map corresponds exactly to one of the internal decoder lines so, if a logic function fits one it can be implemented in a prom. The Karnaugh Map for the above function is shown in Fig. 2. Note that we are only talking about combinatorial functions here.

It does not matter if the logic is defined as active-low, as here, or active-high. In the latter case the cells would be defined by 'H's instead of 'L's.

Proms are usually defined by a truth table, rather than equations or a Karnaugh Map. The truth table for the above equations would be:

	0	1	2	3	4	...
0000 -	01	01	01	00	01	...
0010 -	01	00	01	01	01	...

and so on, where addresses and data are in hexadecimal.

In practice, several software packages for PLD design include an assembler for proms. This allows logic to be defined as Boolean equations, and assembled into a format suitable for prom programmers. We will look at PLD assemblers and programmers in more detail later on.

One of the earliest applications for prom logic was in address decoding. Fig. 3a shows the memory map of a fairly simple, hypothetical microprocessor system, including some memory-mapped i/o, one port being read only, the other read/write. The enable signals for the various peripherals can be derived from the upper four address lines and the read/write output from the processor.

This logic may be converted directly to the truth table in Fig. 3b, more or less by inspection. For example, A15, A14 and A13 all low will address any location in the lowest 8K of the map. The only complication is the ram space, which needs to be split into three segments for addressing purposes.

To convert this truth table into a hex-ascii table, which is the format needed for loading into a prom programmer, we must first allocate pins to signals. Let us assume that R/W

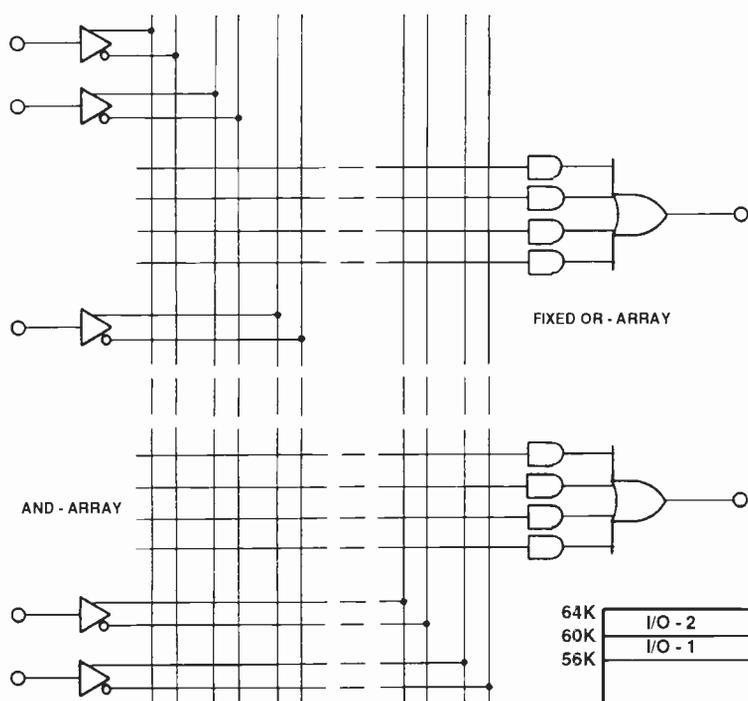
A question of notation

Many designers prefer to describe logic circuits by a circuit diagram, although most PLD assemblers require Boolean logic equations as their data source. Computer aided design systems with schematic capture usually have interfaces to the popular PLD assemblers, so direct conversion of circuit diagrams to logic equations is possible. If this is not available, the conversion must be done by hand. Figures 7a and 7b illustrate this process for a simple circuit.

The first step is to eliminate double inversions, then the circuit can be redrawn without the constraints of standard logic family packages. The internal signals can be labelled with their logic equations, starting from the inputs. By carrying this process

through to the outputs, the logic equations for the outputs can be deduced.

Internal signals which become too complicated can be redefined as a single symbol. Most PLD assemblers allow substitution of strings or internal signals by logic equations defined in terms of input signals. They will then perform Boolean algebra on the resulting complex logic equation. The most likely snag is that the chosen target device will not have sufficient capacity for the resulting equations. The way round this, assuming that the assembler has performed some minimisation, is to redefine the active levels of the outputs with most and terms, if the device allows this, or try a different device with higher and term capacity.



	0	0	0	0	1	1	1	1	A4
	0	0	1	1	1	1	0	0	A3
	0	1	1	0	0	1	1	0	A2
0	0								
0	1								L
1	1	L							
1	0								
A1	A0								

Fig. 2. Karnaugh Map for prom logic function

Fig. 1c. pal architecture

is the most significant address line and A12 the least significant; also assume that the outputs are assigned in order from prom1 as O0 to I/O2-WRITE as O7. The output data for O0 low and all other outputs high is 'FE', for O1 only low it is 'FD' and so on.

Prom1 is selected when A15, A14 and A13 are low irrespective of R/W and A12, which corresponds to addresses 0000, 0001, 0010 and 0011; similar reasoning will determine the active addresses for the other outputs and allows us to construct the prom table of Fig. 3c.

This is a fairly elegant one-chip solution for a circuit which would otherwise take five gate packages, or a 3 to 8 decoder and two gate packages.

However, prom logic does suffer from drawbacks. Firstly, each additional input requires the number of memory cells to be doubled. Further, the number of inputs and outputs in a particular prom is fixed with, usually, either four or eight outputs. As memory size increases, so do cost, power consumption and delay time. Thus, while proms may be suitable for simple logic circuits, they soon run out of steam in more complex situations.

The FPLA solution

The major prom drawback of doubling in size for each extra input comes about because the inputs are fully decoded. This is clearly a waste of resources in our example above because six of the outputs only need a single combination of inputs decoded while the other two need just three.

In an FPLA (field programmable logic array) the input decoder is programmable as well as the output or-array. This architecture is shown in Fig. 1b. Thus, in the PLS100, the first commercial FPLA, only 48 input combi-

64K	I/O - 2								
60K	I/O - 1								
56K	32K RAM								
24K	8K PROM								
16K	8K PROM								
8K	8K PROM								

Fig. 3a. Memory map for simple microprocessor system

R/W	A15	A14	A13	A12	I/O-2 READ	ACTIVE O/P
H	H	H	H	H	I/O-2 READ	07 (7F)
L	H	H	H	H	I/O-2 READ	06 (BF)
X	H	H	H	L	I/O-1	05 (DF)
X	H	H	L	X		03 (F7)
X	H	L	X	X	RAM-CE	
X	L	H	H	X		
L	H	H	L	X		04 (EF)
L	H	L	X	X	RAM-WE	
L	L	H	H	X		
X	L	H	L	X	PROM 3	02 (FB)
X	L	L	H	X	PROM 2	01 (FD)
X	L	L	L	X	PROM 1	00 (FE)

Fig. 3b. Address decoder truth table for simple microprocessor system

	0	1	2	3	4	5	6	7	8	9	A	B	C	D	E	F
000_	FE	FE	FD	FD	FB	FB	EF	DF	BF							
001	FE	FE	FD	FD	FB	FB	F7	DF	7F							

Fig. 3c. prom programming table for address decoder

nations are decoded from the 16 inputs compared with the 65536 decoded lines in a theoretical 256k prom, with 16 inputs.

On examining the structure of the and-array, we find that each and-gate (usually called an and term or product term) has two connections to each input line, one to the 'true' signal and one to its complement. By leaving the appropriate fuse intact, any signal or its complement can be present in any product term.

In our memory map example, leaving the three complement fuses of A15, A14 and A13 intact - and blowing all the other fuses - will provide the decode signal for prom1.

To connect the decode signal to an output, the appropriate fuse in the or-array between this product term and the 'prom1' output must be left intact and the fuses to the other output

lines blown. Left like this, the output will be high whenever these three address lines are low, and low at all other times; this is just the inverse of what we programmed into the prom.

There are two ways out of this problem. One is to use Boolean Algebra and rewrite the logic as:

$$prom1 = A15 \# A14 \# A13$$

The other way is to make use of the programmable inverter which is usually found in FPLAs and many other PLDs. It is an exclusive-or gate, shown in Figure 4, with one input from the or-array and the other to a pull-up which is grounded via a fuse. With the fuse blown the logic is inverted, if unblown it is not affected.

The format for entering this data into a pro-

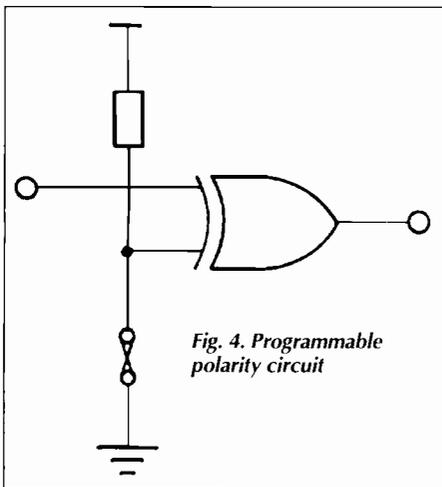


Fig. 4. Programmable polarity circuit

grammer, in order to blow the desired fuses is a dedicated truth table, although it is more usual to use a logic compiler with logic equations these days. Nevertheless, an FPLA truth table is a useful way of representing a logic function, and may still be used to by-pass logic compilation.

Fig. 5 shows the FPLA truth table for our memory map decoder. It has two halves; there is a section for defining the product terms, and a section for connecting the or-array and setting output polarity.

An 'H' in a signal column connects the 'true' fuse to that term, an 'L' the complement. A 'don't care', that is no connection at all, is shown by the '-' symbol. Putting '0' would cause both fuses to be left intact, so the and gate would see both polarities at once; it would, therefore, always be low which is an inactive state. The combination of Hs and Ls is the logic condition which will cause the product term to be active; that is to go high.

In the or-array, each column represents an output. The box at the head of each column defines the action of the programmable invert-

er. An 'H' means no inversion (active-high) and an 'L' causes inversion (active-low). To attach a product term to an output via the or-array, an 'A' is placed in the box where the product term row and output column intersect. A '.' in this box means that the fuse at that or-array intersection is to be blown.

The memory map example shows that the same logic which used 32 product terms in a fully decoded prom, needs only twelve terms in an FPLA. The PLS100, though, has sixteen inputs, eight outputs and 48 product terms, so its resources are very under-used. Even if there were other logic functions to be included, all the outputs are used up so the rest of the FPLA is redundant in this application. This FPLA shares one of the drawbacks of proms, namely, a fixed number of outputs leading to an inflexible allocation of logic resources.

The next generation of FPLA, the PLS153 and PLS173, overcame this by using the bi-directional i/o structure of Fig. 6. Each FPLA has ten i/o pins and, respectively, eight and twelve inputs, with 32 product terms. If a PLS173 were used for the memory map decoder, there would be seven inputs, two i/o pins and 20 product terms available for another logic function. However, the same FPLA could be used in an application requiring 20 inputs and just two outputs, for example.

An additional set of product terms must be programmed to define the direction of data flow through each i/o pin. An unprogrammed PLS153/173 has low levels on all the and gate outputs. The three state buffer on each i/o has an active-high control so, before programming, all the buffers are disabled and each i/o pin acts an input. If all the fuses in a product term are blown the and-gate will be unconditionally high; in this case the buffer is always enabled. If a logic function is programmed into the product term, the buffer will be enabled when the logic function is true.

PROM, PAL or FPLA?

There are three architectures commonly used for implementing combinatorial logic functions in programmable format: prom, FPLA and pal. Although proms are designed to be used in memory applications such as program storage and look-up tables, they are equally suited to use as logic devices. Every possible input combination can be programmed to give a specific output. Any logic function which can be written as a truth table can also, in principle, be programmed into a prom; the only restriction is the number of inputs and outputs which the prom supports.

An alternative approach is draw the Karnaugh Map for the logic function. Every cell in the Karnaugh Map can be programmed individually into a prom, because the inputs are fully decoded. All the decoded lines are or-ed together in the or-array, via fuses, which transmit a high to the output when the line is addressed, but leave it low if the fuse is blown.

Adding an input to a prom doubles the number of fuses needed in the or-array. In an FPLA the input decoder is also programmable, so only 'active' input combinations need to be programmed. Thus, while a ten-input prom has the equivalent of 1024 and-gates in its input decoder, FPLAs typically have only 32 or 48 programmable and-gates.

If a logic function has a truth table with more than 48 lines, there might be problems fitting it into an FPLA. The only way round this problem is by logic minimisation. For simple functions it may be possible to draw Karnaugh Maps for each output and reduce the number of logic terms by combining map cells. For more complex functions it may be necessary to use a minimisation program on the logic equations for the function.

FPLAs possess the property of being able to allocate any and-term to any one or more outputs. However, many circuits, particularly address decoders, need each and-term for only one output. Thus the programmable or-array can be replaced by fixed or-gates without adversely affecting the utility of the device. This is the principle of pals.

Apart from simplicity, there is a gain in performance from this approach. The or-array consumes power and adds capacitance to the decoded and-term lines; pals normally run cooler and faster than FPLAs.

The chief drawback is the limitation in and-terms per output. There are, at most, only seven or eight and-terms or-ed together for each output, in simple combinatorial pals, so functions needing more than this would require an FPLA.

INPUTS					OUTPUTS							
I ₄	I ₃	I ₂	I ₁	I ₀	O ₇	O ₆	O ₅	O ₄	O ₃	O ₂	O ₁	O ₀
ACTIVE LEVEL					L	L	L	L	L	L	L	L
H	H	H	H	H	A
L	H	H	H	H	.	A
-	H	H	H	L	.	.	A
-	H	H	L	-	A	.	.	.
-	H	L	-	-	A	.	.	.
-	L	H	H	-	A	.	.	.
L	H	H	L	-	.	.	.	A
L	H	L	-	-	.	.	.	A
L	L	H	H	-	.	.	.	A
-	L	H	L	-	A	.	.
-	L	L	H	-	A	.
-	L	L	L	-	A

Fig. 5. FPLA programming table for address decoder

It should be noted that, even when the output is enabled, the i/o pin is still connected to the and-array. Any output function can, therefore, be fed back and combined with input signals. The FPLA can implement multi-level logic in this way, or make latches and flip-flops as we shall see further on.

The code for entering control logic in truth table format is just the same as in the main logic array, with the addition of the symbol '0' to denote both fuses left intact. Thus, a control term for a dedicated input will usually consist of a row of 0s.

PALs – a simpler way

While FPLAs offer a more flexible solution than proms for programmable logic, re-examination of Fig. 5 shows that they can be unnecessarily complicated. Although each product term is available in the or-array to be gated into any output, in this application every product is used in only one output. In other words the programmability of the or-array is wasted. Each output could just as well use a fixed or-gate with the product terms shared between the outputs.

This is just the structure of a pal; pal stands for programmable array logic, and is illustrated in Fig. 1c. PALs were introduced by Monolithic Memories Inc. in the mid 1970s. The first pals occupied a 20-pin package and contained just sixteen product terms. These were shared among the outputs so that a two output pal had eight terms per output, a four output device just four terms, and so on. A similar 24-pin family was also made available.

Pals have a very logical numbering system. For example, a *PAL14H4* has fourteen inputs and four active-high outputs, while a *PAL16L6* has sixteen inputs and six active-low outputs.

Apart from being simpler to understand and to use, pals have performance advantages over FPLAs. A fuse array is a high capacitance structure which introduces propagation delay into the logic path. Removing the or-array in favour of a fixed or-gate saved 5ns in the first pal family, compared with the equivalent FPLA. Also, cutting down the number of product terms made a significant reduction in power consumption.

The first pal families suffered from the restrictions of fixed input and output numbers, and fixed output polarity. While adequate for simple circuits, even our simple memory map logic would not fit the pal with the appropriate i/o count, the *PAL10L8*. This is because two of the outputs need three product terms, but the *PAL10L8* has only two terms per output.

Introduction of more complex pals, such as the *PAL16L8*, overcame these drawbacks. This pal has eight product terms per output, but one of these controls the three-state output in the same way that bi-directional outputs are controlled in FPLAs. Six of the eight outputs are bi-directional, so i/o flexibility is available as well as increased logic power. There are still only ten direct inputs, but the fed back inputs make a total of sixteen into the whole array.

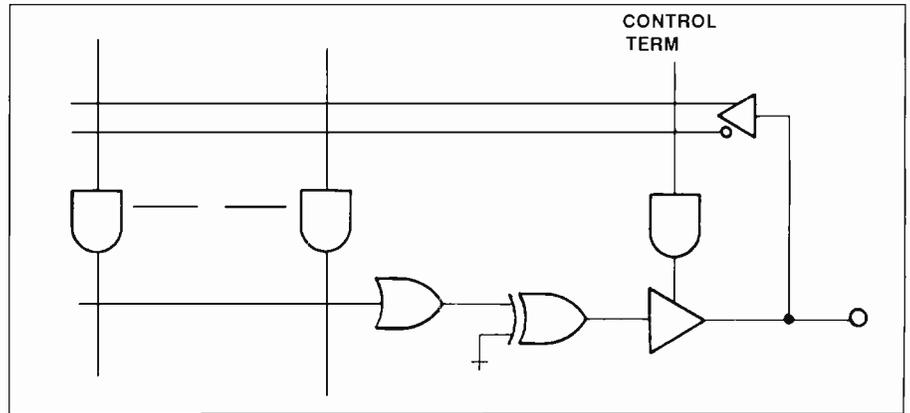


Fig. 6. Bidirectional i/o circuit

PAL16H8 and *PAL16P8* (programmable polarity) were also made but better solutions can be found now, as we shall see in a later article, so only the *PAL16L8* remains in current production.

Our address decoder will fit into a such a device, but truth table entry is not a preferred method of designing pals. Once personal computing became commonplace, MMI introduced a pal assembler called palasm. The designer uses logic equations to define the circuit and palasm assembles this into a file to the jedec specification. The jedec file can be loaded into a pal programmer, which blows the correct fuses in the pal to reproduce the logic in hardware.

Given that the tools exist we just have to define our circuit as logic equations. We can illustrate this by writing the logic equations for the memory map example. Each line of the truth table is a single and-term; where an output contains more than one line of truth table, the various and-terms are or-ed together.

```
!O7 = !RNV & A15 & A14 & A13 & A12
!O6 = RNV & A15 & A14 & A13 & A12
O5 = A15 & A14 & A13 & !A12
!O4 = !RNV & A15 & A14 & !A13
      # !RNV & A15 & !A14
      # !RNV & !A15 & A14 & A13
!O3 = A15 & A14 & !A13
      # A15 & !A14
      # !A15 & A14 & A13
!O2 = !A15 & A14 & !A13
!O1 = !A15 & !A14 & A13
!O0 = !A15 & !A14 & !A13
```

Note that the above equations would not assemble in palasm because that uses a different syntax: * for and, + for or and / for invert. These symbols are used for arithmetic functions in some other assemblers which are now available.

For completeness we should mention the other combinatorial pals which are still available. The *PAL20L8* is a 24-pin version of the *PAL16L8*; it has fourteen dedicated inputs but is identical in every other way. Another 24-pin device is the *PAL20L10*; this has only four product terms per output, one for control and three for logic. It has twelve direct inputs, two direct outputs and eight bi-directional terms.

While these devices are still being made, it is likely that most will be phased out in favour

of the cmos generic pals, which we will examine in the third article. The exception may be the 5ns and 7ns pals, which are still easier to make in bipolar than cmos technology.

Design entry methods

The standard way of defining logic for any combinatorial PLD is in terms of Boolean equations. Many assemblers are now available. Some are dedicated to a particular PLD manufacturer such as AMD (palasm), National (Opal) and Philips (snap), while others are marketed for universal use. Examples of universal software suppliers are Data I/O (abel), Isdata (LOG/iC), Logical Devices (cupl) and OrCAD (OrCAD-PLD). Many work stations also feature PLD assemblers.

Most software now includes features such as logic expansion, minimisation and simulation. The former allows many functions to be written as compact equations, with brackets and multi-level logic definition, while the latter gives the designer confidence that the finished product will do the job for which it is intended. Simulation will usually result in the production of test vectors. These instruct the programmer how to functionally test the PLD after programming. Occasionally, a device will be programmed correctly but will not function because of some small fault in the device, which cannot be tested before programming.

However comprehensive the software, the correct logic data must be entered in order to guarantee a working PLD. We have seen how a truth table can be converted into logic equations but, very often, the logic is not readily defined as a truth table.

The first step in any design is partitioning... deciding which input and output signals are to be included in the logic block. If the total number of signals is no more than twenty two, and there are ten or less outputs, it is likely that the logic will fit into one of the devices described so far if we allow only combinatorial logic.

The relationship between the input and output signals must eventually be defined by logic equations but some designers may not be comfortable taking that step directly. The other common way of visualising logic is with a logic diagram, or schematic. A logic

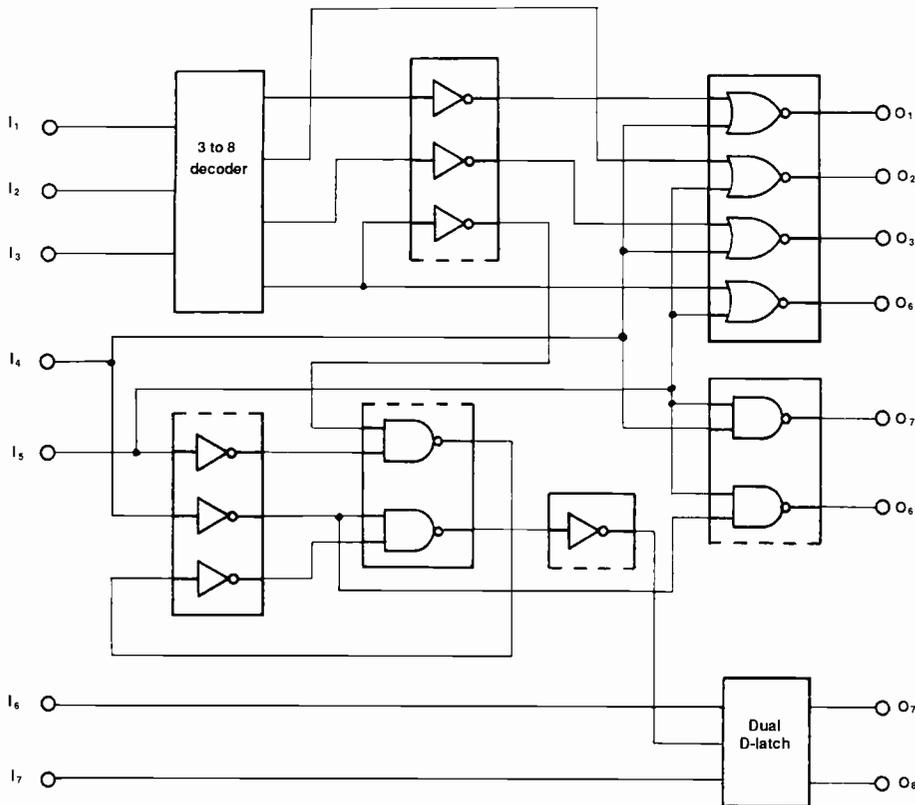


Fig. 7a. Typical logic circuit drawn for TTL or CMOS implementation

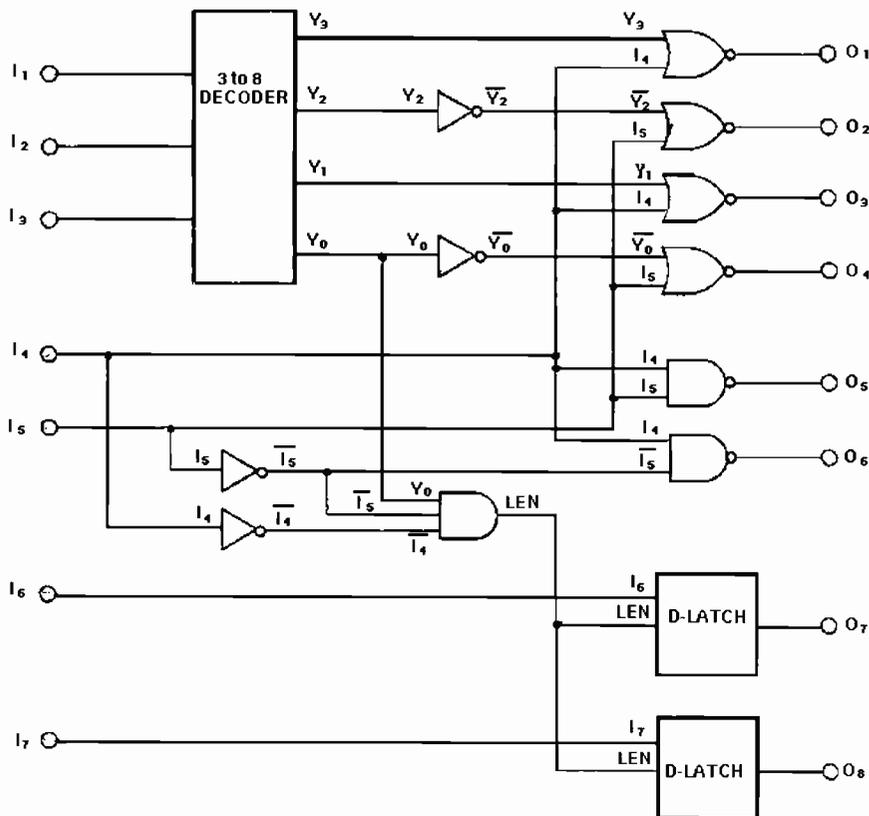


Fig. 7b. Typical logic circuit after compaction for Boolean logic equations

schematic is, perhaps, the traditional way of defining a logic system as the components are clearly laid down with their interconnections drawn in. Often, the components are functional blocks from a standard logic family with a scattering of single gates and inverters. A simple example is shown in Fig. 7a.

Schematic capture software will convert logic schematics directly to logic equations in a format suitable for most of the standard logic assemblers. If schematic capture software is not available, the designer is then faced with the task of converting the data manually. This is not as traumatic as it may appear. Fig. 7b shows the first stage, which is to label a simplified logic schematic with internal signal names.

The second stage of the conversion is to define the internal signals in terms of input signals and 'earlier' internal signals, thus:

$$\begin{aligned} Y_0 &= !I_1 \& !I_2 \& !I_3 \\ Y_1 &= !I_1 \& !I_2 \& I_3 \\ Y_2 &= !I_1 \& I_2 \& !I_3 \\ Y_3 &= !I_1 \& I_2 \& I_3 \\ LEN &= !I_4 \& !I_5 \& Y_0 \end{aligned}$$

If the PLD assembler can handle internal signals which are neither inputs nor outputs, LEN can be left unexpanded; otherwise it must be expanded to:

$$LEN = !I_4 \& !I_5 \& !I_1 \& !I_2 \& !I_3$$

The output signals may now be written in terms of input and internal signals. O1 to O6 should cause no problems, unless O2 and O4 need further expansion by hand,

$$\begin{aligned} !O_2 &= I_5 \# !Y_2 \text{ becomes:} \\ !O_2 &= I_5 \# I_1 \# !I_2 \# I_3 \text{ and } !O_4 = \\ I_5 \# !Y_0 \text{ becomes:} \end{aligned}$$

!O4 = I5 # I1 # I2 # I3 by the usual de Morgan conversions.

O7 and O8 are outputs from a dual D-latch, which is not a combinatorial circuit, but can be built from combinatorial components. Fig. 8 shows the schematic of a D-latch built from NAND gates, and the corresponding Karnaugh Map. The logic equation for the D-latch is:

$$Q = D \& LE \# Q \& !LE \# D \& Q$$

The third term (D & Q) looks superfluous but links the other two terms on the Karnaugh Map. Without it there would be a danger of glitches when the latch changed state; as a general rule, overlapping terms in a Karnaugh Map will prevent glitches. The only problem is that the design is now untestable without adding a control signal to allow the gates to be tested individually.

We can complete the derivation of the logic equations by substituting I6 and I7 for D, O7 and O8 for Q, LEN for LE; thus O7 becomes:

Fig. 7c. Logic equations derived from Fig. 7b

$$\begin{aligned} !O_1 &= Y_3 \# I_4 \\ !O_2 &= !Y_2 \# I_5 \\ !O_3 &= Y_1 \# I_4 \\ !O_4 &= !Y_0 \# I_5 \\ !O_5 &= I_4 \& I_5 \\ !O_6 &= I_4 \& !I_5 \\ O_7 &= I_6 \& LEN \# O_7 \& !LEN \# I_6 \& O_7 \\ O_8 &= I_7 \& LEN \# O_8 \& !LEN \# I_7 \& O_8 \end{aligned}$$

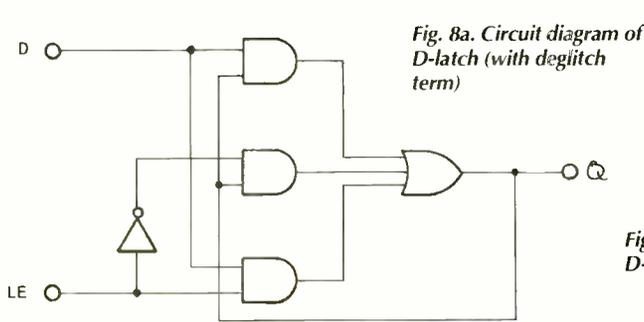


Fig. 8a. Circuit diagram of D-latch (with deglitch term)

	0	0	1	1	D
	0	1	1	0	LE
0			H		
1	H		H	H	
Q					

Fig. 8b. Karnaugh map of D-latch

programming data and most programmers can use this to functionally test the finished device. A small proportion of devices will not work correctly even when the correct fuses have been blown. This is because the device cannot be fully tested by the manufacturer before any fuses have been blown. Unless the device is made in an erasable technology, which many are now, the fuses cannot be blown before leaving the factory.

For similar reasons, a small proportion of devices will not program correctly either. Although programming failures can be due to manufacturing defects, they can also be caused by the programmer being out of calibration or by poor socket contact. For this reason, users performing their own programming should ensure that their programmer is kept well maintained. A fall in programming yield can indicate a need for corrective action.

Many users of programmable logic buy their PLDs ready programmed, or use a specialist sub-contractor. This has the advantage of removing yield problems, and the need for investing in equipment which has to be upgraded as new devices come onto the market. Once the design is proved, buying PLDs then becomes as transparent as buying standard logic and easier than masked semi-custom, because PLDs are mass-produced standard devices, unlike masked ASICs.

$$O7 = I6 \& LEN \# O7 \& !LEN \# I6 \& O7$$

If expansion is necessary, the following results:

$$O7 = I6 \& !I1 \& !I2 \& !I3 \& !I4 \& !I5$$

$$\# O7 \& I1 \# O7 \& I2 \# O7 \& I3$$

$$\# O7 \& I4 \# O7 \& I5 \# O7 \& I6$$

In most cases, expansion will not be necessary and the complete set of equations can be left as in Fig. 7c.

Design methods for sequential circuits are usually quite different from those illustrated above, and device architectures are more complex. These topics will be covered in next month's article.

Programming

The only issue left unresolved is program-

ming. Programmers come in all shapes and sizes, from large universal stand-alone machines to small dedicated units which are PC driven. Because the design will probably have been done on a PC, the data has to be transferred to the programmer. A standard format has been agreed between device manufacturers, design software houses and programmer makers.

This jedec standard numbers each fuse in the device and the design software will produce a file with a string of 1s and 0s indicating which fuses are to be blown and which left intact. This file is downloaded into the programmer which addresses the device to be programmed with the correct fuse locations, according to the data in the file.

Test vectors can be appended to the pro-

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WINDOWS

support for DSP design

DSPworks and QEDesign 1000 are available as a single package for digital filter design. But, apart from running under Windows, what else distinguishes them, asks Allen Brown?

Fig. 1. Windows DSPworks features are accessed via drop-down menus. Background window is that of a square wave with added Gaussian (normally) distributed noise (which has controllable mean and variance).

The number of digital signal processing (DSP) software packages running under Windows is still small. But Momentum Data Systems of Costa Mesa, California has added to that total with the release of two Windows packages for DSP design and analysis. *QEDesign 1000*, aimed at designing digital filters, and *DSPworks*, for signal analysis. *DSPworks* is a general purpose digital signal processing software package that allows many of the well known signal processing tasks to be carried out on data derived from an expansion card or previously stored in a disc file. Its easy-to-use graphics allow data to be displayed in a variety of formats. One immediate piece of good news is that *DSPworks* does not need a dongle: the bad news is that *QEDesign 1000* does. That said, once it is connected to the ever growing dongle string protruding from the back of your PC, you should be up and running in no time. (*QEDesign 1000* will actually work

without the dongle but it will not output the filter coefficients to disc).

In *DSPworks*, every waveform or function generated gets its own window, with the individual waveforms selected from a drop-down menu.

Reflecting the importance of DSP window functions, a large selection are available with *DSPworks* – including the Kaiser Bessel window, indispensable for designing finite impulse response filters (FIR) using the window technique.

Another broad choice is that of formats for the waveform generated, spanning 32-bit floating numbers, binary or ascii characters. Considerable control can be exerted over the data formats by using the utilities menu to, for example, round, truncate and convert formats. Conversion is useful when data are to be used with other applications packages.

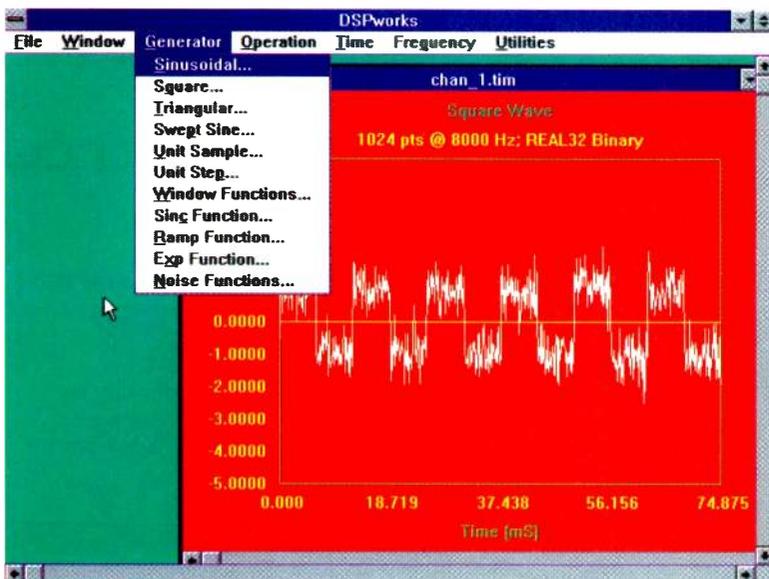
Waveforms that have been generated or imported can then be manipulated, with options including a smooth or moving average. QUANTISE FIXED POINT enables data to be quantised. This is useful when simulating the behaviour of digital filters implemented on fixed point processors (eg Motorola *DSP56000*), which can differ significantly from floating point simulations.

File names are allocated to the generated waveforms, and stored on disc as they are created. Whether this is a confidence measure or a method of reducing the amount of ram required, it is certainly a convenient safeguard.

Time domain signal processing operations include all the standard functions, with an oscilloscope option available to users with an appropriate expansion card resident in the PC. SIGNAL FILTERING is only useful if data from *QEDesign 1000* is present.

In this case, *QEDesign 1000* would have been used for the design of the digital filter, followed by code assembly then analysis of performance either using disc data or real-time data from an expansion card.

Of the frequency options, as with the time options, some can be performed with disc data while others are only offered when real-time data is available from a DSP card. In effect,



QEDesign allows a PC and DSP card to act as a real-time spectrum analyser with 2-d spectral displays and 3-d waterfall displays.

DSP expansion cards

To get the most out of *DSPworks*, a DSP card is a must – most engineers working in the field of DSP will probably have at least one DSP card at their disposal anyway.

Card manufacturers represented include Ariel, Loughborough Sound Images, Sonitech, DSP Research and Data Translation, a list that shows ample support for cards hosting Texas Instruments and Motorola chips. But for cards hosting Analog Devices *ADSP-2100* series chips, support is lacking, a deficiency that should be rectified since Analog Devices is capturing a sizeable part of the telecomms DSP market.

If the card has analogue I/O facilities, *DSPworks* also provides options relating to sample rate and channel selection. For example the Ariel *DSP32C* card has sampling rates adjustable from 2kHz to 100kHz.

A DSP card can be used in a number of ways. For example, where signal behaviour in given frequency regions is of interest, a digital filter could be placed between the incoming data and the display on the screen. Or the expansion card could act purely as a digital filter, digital data derived from its A-to-D converter being processed and the output digital data stream being passed to its D-to-A to generate a filtered analogue output.

QEDesign 1000

QEDesign 1000 is a specialised *Windows* software package, solely for designing digital filters. The user is asked, through dialogue boxes, to specify filter parameters, and the appropriate filter coefficients are produced. Various filter characteristics (eg transfer function and impulse response) are plotted in individual windows, giving instant visual access to filter performance.

In practice, digital signal processors are frequently used to perform filtering tasks and part of the course of designing a digital filter is to determine the appropriate coefficients.

One interesting feature of *QEDesign 1000* is the use of double precision (64-bit floating point) in design calculations. In design of very large filters, with many tens of taps, the numerical rounding effects can be significant. For example the Equi-ripple design method uses the Remetz exchange algorithm which is recursive and very sensitive to numerical precision. But it means that a maths coprocessor is indispensable (maths coprocessors for 33MHz 386-PCs can be found for under £60 and every engineering PC should have one anyway).

Choosing coefficients for a digital filter is somewhat similar to choosing the *R*s and *C*s in an analogue filter. *QEDesign 1000* presents seven plotting areas on the screen, and when a filter has been designed the results are displayed in the seven regions using the tile option from the window command.

IIR filter design

QEDesign 1000's infinite impulse response (IIR) filter design is based on the impulse invariant technique, using models of analogue equivalent filters. The models are Butterworth, Tschebychev, Elliptical and Bessel, all familiar to analogue filter designers.

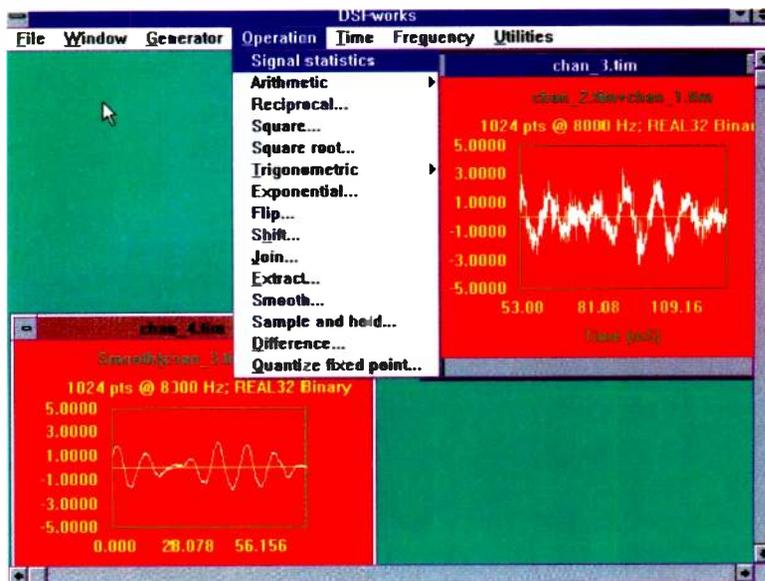


Fig. 2. *DSPworks* has several processing options. Smooth is shown in the background.

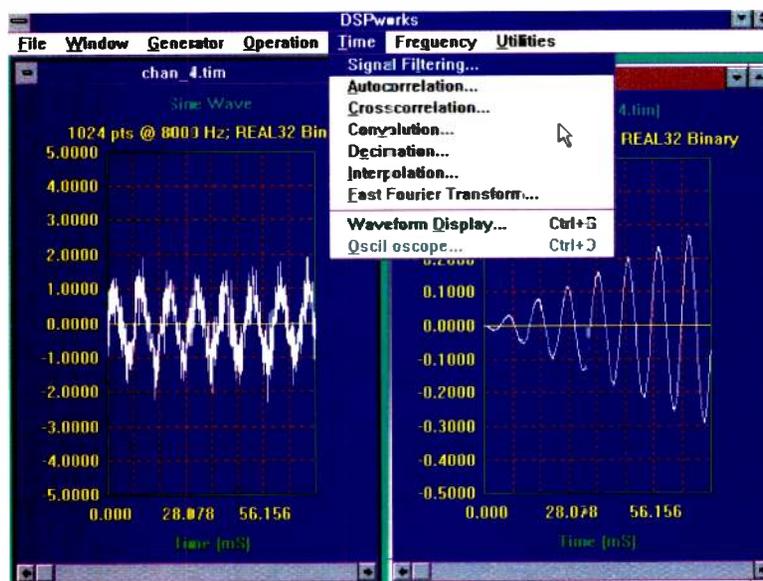


Fig. 3. Time domain processing options are instantly available. Screen shows the effect of applying the autocorrelation function to a noise contaminated sine wave.

EASIER CODE GENERATION

Part of *QEDesign 1000* is an optional code generator(s) which is not only tailored to a selected DSP processor, but also to a proprietary PC expansion card.

Once satisfied with the design of a digital filter, a user can avoid the task of coding in assembly language for a specific processor by using the optional code generator.

Code generators are not new – Atlanta Signal Processing, with their digital filter design package, provided a code generator for the Texas Instruments *TMS32010* in 1983. But they are convenient, especially when designed to work with specific PC expansion cards. Momentum Data Systems are currently producing code generators for several well established commercial expansion cards: eg the Ariel *DSP32C* board. But the software development tools (assemblers and linkers) will still be needed for the processor to produce executable code.

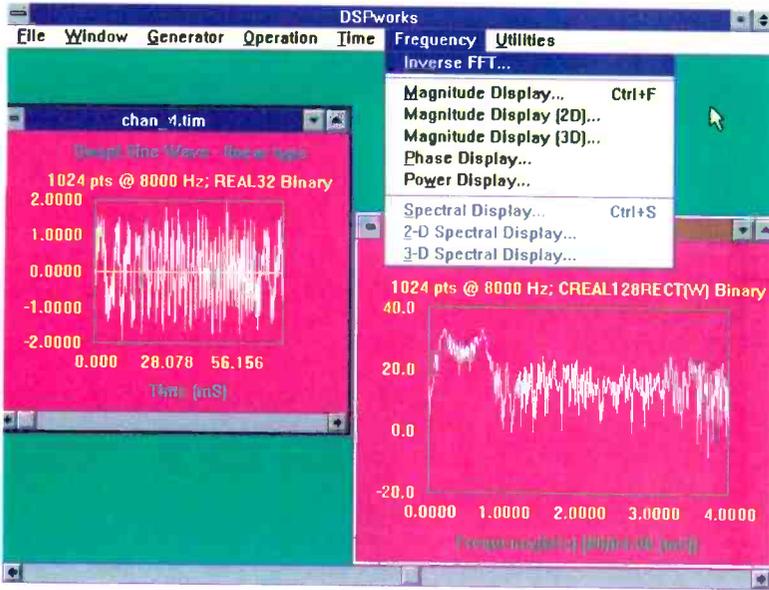
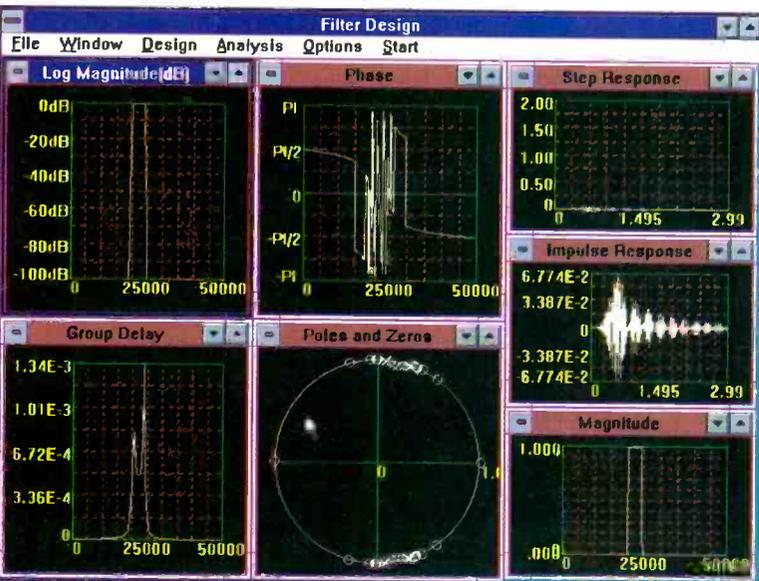
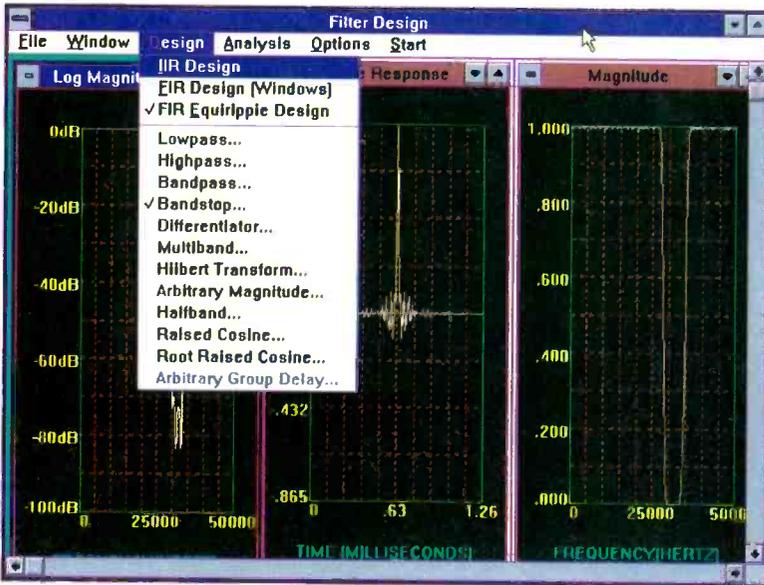


Fig. 4. A DSP card allows real-time frequency processing to be achieved

Fig. 5. QEDesign 1000 choice of filter types is large.



SYSTEM REQUIREMENTS

386 or 486 PC with VGA monitor
 Maths coprocessor (advisable)
 Expansion card hosting DSP chip (strongly advised)

DELIVER US FROM DONGLES

What happens if a user has several software packages, each requiring a dongle? The result is a ribbon cable and a string of cascaded dongles. There must be a better solution.

One alternative that would still prevent multiple copies of software from being made is a greater use of the cd-rom.

The cd-rom drive is continually falling in price (the Mitsumi LU005S is now available for under £120) and over the next year will become a common peripheral in PCs.

But it is not only cost that will act as a driving force, there are hard disc advantages too.

Historically, the problem with computers has been their insatiable appetite for memory. No matter how much memory a computer has, it will never be enough.

The dos memory limit managed to stem this problem for several years forcing software producers to work within the 640kbyte boundary. But the advent of Windows, memory managers and dos extenders has brought multi-Mbyte PC software packages.

Many commonly used software packages take up several megabyte of hard disc space, (for example Visual C++ from Microsoft takes up 40Mbyte of hard-disc and can take up to an hour to install). A 386-PC or 486-PC with only a single 100Mbyte disc will soon become inadequate.

By issuing software on cd-roms (either 12cm or 8cm), which can be executed directly, it will be possible to save on hard-disc space, save time by not having to install it, while giving a high degree of protection against illegal copying. The argument against using cd-roms usually relates to the data transfer rate from disc to ram which is much less than for hard-discs.

But imaginative use of cache techniques can minimise the problem – bearing in mind that cd-roms are read-only.

If monthly music magazines can generate giveaway cds then I'm sure, with a little imagination, software manufacturers can distribute their products on CDs.

AB

SUPPLIER DETAILS

DSPworks £500 + VAT, QEDesign 1000 £800 + VAT
 DSPworks Plus (Both packages) £1200 + VAT.
 Bore Signal Processing, 39 Hawkswell Close,
 Woking, Surrey GU21 3RS. Tel: 0483-740138 Fax: 0483 740136.

Fig. 6. Graphical results of an IIR filter design displayed in the seven plotting areas.

Filter profile parameters are entered by the user and, before the calculations begin proper, the package estimates the number of taps to realise the design. Results are shown in the seven plotting areas – quite an attractive feature really (Fig. 6) – showing frequency magnitude (linear), frequency magnitude (dB), phase response, zero-pole plot, group delay, step response and impulse response. The coefficients can not be accessed directly, though they can be printed from the saved file.

Most IIR filters are implemented on fixed point processors and so there are certain conventions regarding their realisation. An IIR filter can have many tens of taps. But it is normally configured as biquadratic structures cascaded in a series (see diagram), maximising stability of the filter. Filter coefficients are provided in groups of five for the five multiplications for each biquad.

Should the need arise to implement the filter on a floating point processor, *QEDesign 1000* also provides appropriate floating point coefficients.

In many applications where digital filters are required, lattice filter structures are the best choice. Unfortunately in this version of *QEDesign 1000* there is no provision for designing lattice filters: it would be nice to see this deficiency remedied in future versions.

FIR filter design

Finite impulse response filters can be designed in several ways, and *QEDesign 1000* supports window and equi-ripple techniques.

All linear phase FIR filters are derived from a sinc (FT) profile and truncating the function causes a problem. In the window design method, a window is applied to a function to minimise sharp truncation by softening its end values. Many windows can be used. In the frequency domain they have the effect of improving attenuation in the stop band regions – usually at the expense of broadening the pass band regions.

QEDesign 1000 offers no less than 17 window functions. In response to the filter profile parameters, the package gives a choice of window function with the number of taps required to realise the filter requirement.

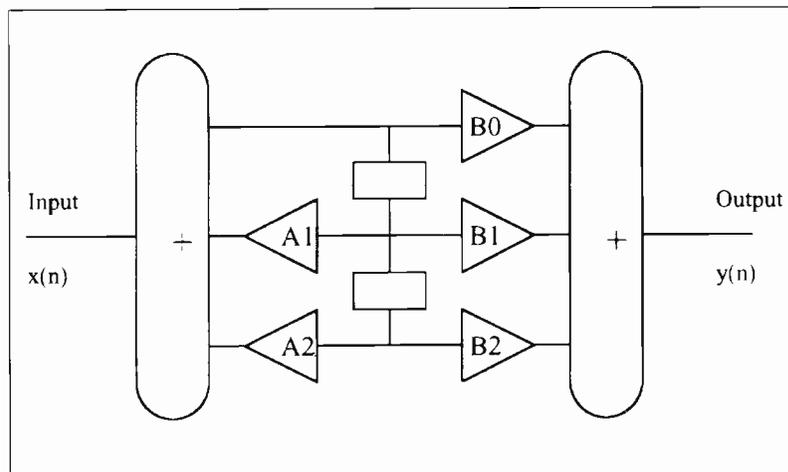
As in IIR filter design, results are displayed graphically, showing frequency magnitude (linear), frequency magnitude (dB) and impulse response. The other method of FIR filter design relies on the work done by Parks and McClellan in 1973. It is basically a curve fitting method, giving a minimum number of taps for a given filter profile. Computationally, it is quite intensive and does not always find a set of solutions (filter coefficients). But the technique does make design of multiband filters relatively easy.

The result is three plots showing the filter's characteristics.

Other options in the FIR design menu include differentiators and Hilbert Transformers (HT) – HTs being useful for generating quadrature components of waveforms. Given $\sin(FT)$, the HT will produce $\cos(FT)$ which can be used in quadrature amplitude modulation (qam) schemes.

QEDesign 1000 also has a facility to analyse the behaviour of transfer functions. Transfer function parameters are entered (either in the Z or S domain) and pole zero values are calculated. Quite complex transfer functions can be analysed, and the package will then design a biquadratic cascade ready for direct implementation of the transfer function.

On the whole, *QEDesign 1000*'s design path is well thought out and easy to learn. But it offers little more than the standard features generally found on digital filter design software. What would be really useful is a FIR multirate filter design feature for when sampling rate of the input device is different from processing rate. Cascading multirate FIR filters can save a lot of computational effort: maybe we shall see this in a future version.



Biquadratic structure with five multiplications, two delay elements and adders.

Competitive market

Looking at both packages, engineers designing FIR and IIR digital filters will certainly find something to attract them.

But there is also a lot of other software on the market performing very similar tasks with lower price tags – some is even free. I also feel the product is a little weak in not offering substantially more than its competitors.

Even so, as a suite, *DSPworks*, *QEDesign 1000* and a code generator, running under *Windows* makes an attractive package. The added facility of linking the product with a range of DSP expansion cards is particularly appealing.

The overall result is that the package is a well put together product that will surely prove a useful asset to the DSP engineer. ■

USER MANUALS

Curiously, the manuals, one for each package, are not actually bound as finished products. They come in A4 ring binders with the impression that as more options are purchased, new pages are added to the binder. The result is that the product looks to be still in a state of flux.

In reality, for most software products, no sooner has Version X been installed than Version X+1 is released, rendering the user manual out of date. At least with a ring binder updated chapters can be inserted as appropriate.

Layout of the manuals is good with lots of screen shots showing how to access the various options.

The *Windows* common user interface makes learning easier, but in general the detail is rather thin, with the feeling that the manuals are most readable by users who are very familiar with digital filter.

New users of these devices will need to have a good prior understanding of the nature of digital filters before launching into the product. What would have been useful here is a greater use of references to readable texts on the subject such as *Digital Signal Processing: A Practical Approach* by E.C. Ifeakor and Jervis (Addison-Wesley 1993).

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	£1.50 each

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Golden earring

I was interested to read Phil Deniss's response to Ben Duncan's article on op-amp distortion (Letters, *EW + WW*, October), but disappointed to note that he chose to write off the "claims of the golden eared or subjectivist club" with such facility and derision. The most valuable tool in any scientific research programme must surely be an open mind. And certainly the fact that we can't – yet – measure or quantify something that we can otherwise positively identify doesn't invalidate its existence. It simply means that we have to work harder to identify an appropriate measurement methodology.

Over the last ten or so years I have worked with Ben on the development of a number of commercial power amplifier designs. From that experience I would now rank "developmental listening" as being as valid in the design of an amplifier as is the quest for product safety, long-term reliability, applications suitability and an acceptable cost of production. Indeed I find it strange to find myself writing to defend a practice which I had always assumed to be a "given" in the development of any audio product – by definition a device which will always be employed subjectively.

However, at the risk of inviting derision from Phil Deniss and the many other members of his objectivist club, I'll describe the way in which I employ my own "golden ears". Where my description is incomplete it is because I don't currently understand the mechanisms involved, I hope that one day encephalography will perhaps tell us more.

My listening tests are carried out in Ben's sitting room, using a system which has remained essentially unchanged for the last eight years. My source material, both vinyl and CD, consists of tracks that I know well, chosen for instrumental or vocal content, performance quality or for some other aspect which I either like or have found useful.

At the start of each session (a session may last for several days) I spend as much time as I feel I need to in re-establishing a model of our reference system in my head. During this time I find I can either switch

from general "listening" to analytical "hearing" or I can't. Colds, coffee, cigarettes and fatigue can all make a difference, but not of degree. It's either yes or no. If the answer is yes, then the reference amplifier is swapped for the amplifier to be tested.

I set myself no time limits, remain relaxed, and play music, in any order which comes to mind that I know will highlight a particular aspect of amplifier performance. Both amplifier channels are always in the same mod state, and I listen in stereo at levels ranging from very quiet to very loud. Because these days we don't usually start listening until we've confirmed channel-to-channel amplitude matching of better than +0.05dB on the Audio Precision, and because our designs usually include the type and make of passive components which we know from experience can be expected to sound good, listening to the effects of each circuit or component change can often be quite prolonged. Long enough, in fact, for me to construct what I can only describe as a mental sonic map of what I am hearing. Longer if I am enjoying the experience.

"Hearing" is probably too limited a description of what seems to happen. Like others have reported, I find I can mentally chop up the source material to concentrate on a voice, or an instrument or a frequency band. But sense of vision also seems to play a part as well. As an example I should cite a test amplifier which incorporated a particularly low-quality preset potentiometer, positioned as a CMRR trimer across the inverting and non-inverting inputs of its differential front end stage. At any listening level I experienced quite involuntary, rapid and independent movement of each eye (behind closed lids), and eventually became quite nauseous as a result – I presume – of trying to resolve two sets of continuously-variable error-laden information into one stable and coherent image.

That the preset was the cause of the problem was later confirmed by Audio Precision testing, in which banging the pcb (and even shouting "hoi" at the preset) were shown to cause significant step changes in its value. Similar vision-related effects have been noted at other times, too.

Wire swapping

With regards to Roger Castle-Smith's (*EW + WW*, July) and RA Woolley's (August) letters concerning swapping live and neutral leads to an amplifier, Woolley's explanation of what happens in the transformer is very clear but the closing remarks make me put pen to paper.

First, most common components have tinned iron lead-outs, which will be affected by the magnetic field present in the amplifier. The effect of the trick is thus to reduce the various forms of distortion in these components. That is why high-end manufacturers all use expensive components with copper lead-outs and often outboard power supplies.

Secondly, the little hum actually created by the magnetic field may not be audible as hum, but different levels in two components of a hi-fi chain must be equalised by way of the only connection between the two – the signal connection. Consequently, the signal transfer will be given less than perfect conditions, which will affect the sound.

In most cases it is more important to connect all components in a hi-fi chain with the same orientation of live and neutral than it is to connect each and every component with the right orientation.

Hope this will cast some light on this piece of folklore, as Woolley prefers to call it.

Bjørn Jerle
Birkerød, Denmark

One commonly-used fet driver stage design is usually drawn to include a component we now refer to as the sea-sickness capacitor.

Lack of space precludes a description of several other recurring physical effects (all, of course, brain-centred) and my apparent ability to retain mental performance "maps" test-to-test. In any event by now Phil Deniss himself may be feeling quite nauseous. Nevertheless, and despite their subjective nature, our listening tests are as rigorous as we can make them. Only one change is ever made between tests, I note my impressions during each test, and each iteration of the amplifier under test is subjected to the same suite of Audio Precision analyses from which: hard copy output is compared. Interestingly this last will often produce no discernible differences under static testing, and it is largely this that prompted Ben's latest research into areas of harmonics and dynamic performance. We want to measure and explain the differences I hear – it makes good commercial sense to be able to do so – but we can't find them all. Yet.

I made the observation earlier that I don't understand the mechanisms involved in this, although naturally I am curious to find out more. In terms of their application, though, I

don't think it matters that much. I'm confident that I can – with reference to my experience as a studio and live sound balance engineer – use my odd physiology/training/ears to achieve the "best" sound obtainable from any well-designed audio product. The fact that our products, once finalised, tend to sound "good" on systems ranging from mid-range domestic to pro audio studio monitoring and touring reinforcement leads me to suspect that our procedures do have some validity.

The statement that "the vast majority" of people apparently can't "hear" the same things does not matter much either. I remain convinced that such a listener will benefit from our work, albeit unconsciously at an emotional level. (Sorry to introduce emotions, Phil. More difficult-to-measure mechanisms at work.) In our deliberately casual, unpressurised and completely random user tests, most subjects have reported an enhanced sense of listening comfort or satisfaction when auditioning our products.

I'm happy with that.

Jerry Mead
Mead & Company
Royston
Herts

Not on, 10-4

While I am concerned that the huge amount of money people have invested in equipment with the 88 to 108MHz broadcast band may have to be augmented by purchases of digital broadcast equipment with a possible lesser technical shelf life, I am concerned that another aspect of the use of the EM spectrum seems to be totally ignored.

I am referring to the use of communications equipment available to the general public for private purposes.

Here in these islands the only legal two-way radio communications equipment available to the public is the inefficient and interference-prone 27MHz fm citizens' band equipment. A more useless allocation could hardly have been chosen into which to dump hundreds of thousands of users throughout Europe.

In the US citizens can buy mobile and portable equipment with minimal paperwork and licence formalities that operate in the 150 or 460MHz regions. Australia, I think, still has a uhf citizens band with repeater facilities. Are governments afraid to let radio communications equipment become available to all citizens? Think of the safety, time and frustration saving aspects in having portable or mobile communications available. Use, identification and frequency control of such equipment could be easily controlled by credit-card type smart cards to minimise abuse of the airwaves.

With two blocks of frequencies a few megahertz apart in the upper vhf or uhf part of the spectrum, unit to unit or repeater use could make life

easier for many people.

Surely with encryption, scrambling, digital and spread spectrum techniques the military do not effectively use all the 225 to 400MHz spectrum? Time for them to hand most of it over and let the public have access to the spectrum as well.

Just who is using all the vhf and uhf frequencies. Surely even 0.5% of the spectrum could be used by (theoretically) 100% of the people? The 27MHz allocation is unusable. So why were we given it?

Des Walsh
Cork, Ireland

Genius in the genes

The article "Geniuses are made not created" (*EW + WW*, October) repeats a common major logical error. Evidence – however comprehensive – that top performance always implies high motivation and hard practice provides no proof for the converse.

Neither does the fact that many with high potential ability fail to maximise their ability prove that outstanding success may be achieved without high inherent potential.

Despite the continuing nature versus nurture argument there is much evidence for major genetic differences between individuals in animals and humans for physical, intellectual and personality characteristics.

For a wide range of physical skills even small differences in body conformation have major effects on extreme performance. The same body type does not produce top performance in all sports. Individual cases prove nothing. Top performers

at national or international level are not random samples but highly self selected for motivation and inherent ability. Small children normally persevere only with skills that provide satisfactory rewards.

First born children tend to produce a much higher proportion of those who achieve outstanding success in life. The normal diversion of concentrated parental attention following the birth of a second child suggests rejection and leads many first borns to make special efforts to regain a starring position.

Evidence from animals shows that similar training does not give similar results for all. Some breeds, strains and individuals do much better than others. Underlying differences in ability usually show early in life.

In domestic animals wide differences of behaviour occur between individuals treated uniformly and kept in a single group. Different breeds, especially in dogs, tend to have inherent behaviour characteristics.

Siblings, even in a similar environment, usually differ widely. This fact contrary to much popular belief supports rather than denies the argument for genetic causes.

The mathematician Norbert Wiener quoted the case of his own brother. Their father had believed that his special teaching system had produced Norbert's outstanding skills. He was highly disappointed to find that, for his younger son, the same system produced only normal adequate competence.

The success reported by Professor Ericsson refers to the somewhat special and narrow skill in remembering a sequence of numbers. However, a memory for 3000 digits suggests an unusual brain quality far beyond the average of 80 more widely achieved by hard practice.

Competence for any skill requires sufficient practice so that the skill becomes largely unconscious and automatic. Any need for conscious thought usually worsens performance. Small changes in target skills create major problems – Konrad Loren wrote at length on this point.

Extreme levels of performance require hours of regular practice. How does this concentration of interest affect the wider range of skills?

RG Silson
Tring, Herts

Over processed radio

Can you hear me, mother? There's a line for radio nostalgia fans. It was the catch-phrase of comedy ventriloquist Sandy Powell who used it to see if he could be heard at the back of the hall. If Sandy was working in radio today, he could be

sure we can hear him. Very loud – but not so clear.

My own sense of nostalgia only goes back a couple of decades, but at last it's official. The 1970s are back! This is the best news in ages as it means all my furniture has come back into fashion.

A Quad II, an ultralinear pair of KT66s allied to a distortion-cancelling pair of EF86s in a clever phase-splitter used to give me "the closest approach to the original sound" for years.

But not any more. To be honest, it sounds dreadful. Legal beagles down at Quad can save their sealing wax, there is nothing wrong with the tuner. Your only crime was to design it properly.

The problem lies in the audio processing that has slowly changed the sound balance since Mott the Hoople was in the charts. It started with wide-band compression. The BBC led the field with a limiter that gently reduced the dynamic range of all audio frequencies present by the same amount, giving an overall impression of loudness enough to counter reasonable domestic noise.

Then came the active systems. A bank of filters carve up the audio into anything up to six pass-bands. These are compressed at different rates preset by the broadcaster, the reconstituted audio then going for transmission.

In pop radio, some DJs can set their own processing at the desk leading to double compression effects which, as they have no musical analogy, can lead to listener fatigue simply due to the saturation of the sound.

Engineers say processing is here to stay – it has grown to be an industry in itself.

Radio marketing staff will tell you that those who shout loudest get the largest audience and so attract the advertising revenue. That's fine up to a point but with CD and digital audio mass storage setting new standards for source programming and radios chasing fashion by including user settings for equalisation – Megabass and the like – this must be the time for the broadcasters to reassess their use of processing to allow the final level of fidelity to align with the listeners level of investment in equipment. In other words, you'll get what you pay for.

As it is, audio processing has had the effect of putting the traditional hi-fi and the ghetto-blower on a middling common denominator.

With so much choice now in radio, isn't it time to move the technical goalposts?

Yes, we can hear you, Sandy. It's just that there seems to be rather a lot of you.

Robert Ellis
Derby

Pause for reflection

If, before reading Ian Hickman's article about rf reflections (*EW + WW*, October), I had known little about the theory and practice of transmission lines, I might have gained the impression by reading it that most transmission lines have phase or wave velocities of 200,000km/s and that somehow a coaxial configuration leads to this.

He states that "in practice any substantial increase [in velocity] proves to be impossible". What he does not point out is that this particular velocity applies only to transmission lines with solid polythene (or equivalent) dielectrics.

Transmission lines (coaxial and balanced) used by broadcasting engineers and others frequently have wave velocities in excess of 290,000km/s, which is due to the small amount of dielectric material used to separate the two conductors. In my opinion 290,000km/s is a substantial increase on 200,000km/s.

He also says that the "values of L and C per metre in free space are the lowest that can ever be achieved, giving $v = c$ and $Z_0 = 377\Omega$ ". Surely he means that the product of the values of L and C per metre is lowest when the conductors are separated by free space, giving $v = c$. L and C themselves may have any values, within reason. The ratio L/C determines the Z_0 of the line, which, in theory, may also have any value.

Dick Manton
Purley, Surrey

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The conventional way to learn about DSP is to grasp fully the maths first. Which is probably why so many otherwise competent design engineers fall down badly in applying digital signal processing. Although the basic algebra is really quite straightforward Jean-Jacques Dauchot maintains that the best way to understand the subject is to get hold of a DSP development system and experiment...



Experimenting with DSP

Processing signals in a numerical form using digital hardware offers significant improvements in reliability and stability over traditional analogue handling methods. With the introduction of microprocessors dedicated for digital signal processing, it is becoming economical to use. However, the use of digital signal processing techniques requires considerable mathematical skill.

Learning from books can be frustrating as they tend to lean on mathematics in a big way. The best way to find out about DSP is to

experiment. Using a low cost DSP kit and software developed by Tycho Designs and Hippo Solutions, this article introduces two typical uses of DSP: fir filtering and digital sinewave generation.

Experimenting with FIR

The main advantages in using digital filters are that filter characteristics are easily changed; they are stable against changes in temperature; they process low frequencies more effectively; frequency response characteristics can be made to approximate closely to the ideal; they

can be made to have no insertion loss; linear phase characteristics are possible; reliable and repeatable; no precision components or component matching; no timing required; superior performance.

There are two kinds of digital filters used: finite impulse response (FIR) and infinite impulse response (IIR) filters. Digital filters with a finite impulse response can exactly achieve linear phase, cannot become unstable, and can be easily realised on general purpose or customised hardware. IIR filters are more efficient than FIR filters, and can generally

give a sharper cut-off than an FIR filter of the same order. Because IIR filter algorithm uses feedback of a previous output signal, they are also called recursive filters. Consequently, they have a non-linear phase response and can become unstable.

Which filter is better for a particular application depends on the hardware used for the implementation of the filter. For example, the Texas Instruments *TMS320C series* family of signal processors include a special instruction to facilitate the implementation of an fir filter. The combined use of the instructions LTD and MPY are specially geared in the design of the fir filter.

The fir filter is also known as a non-recursive filter or convolution filter. From the time-domain point of view, it is also called a moving average filter. A fir filter works processing an output signal from the history of a number of previous samples each time a new sample signal is received. We can view the process of filtering through the changes in the differences in level between consecutive samples. The way the filter distinguishes between high and low frequencies is by the rate of change between samples. Large differences indicate high frequencies and small differences indicate low frequencies.

Consider the filter design in Fig. 1 and suppose a sequence of samples of values 1, 8, 6, 4, 1, 5, 3, 7 enter the system. The difference between consecutive samples ranges from 2 to 7. Using a coefficient value *g* of 0.25, the first three numbers enter the filter and are added and multiplied: $(1+8+6) \times (0.25) = 3.75$.

The next generation yields: $(8 + 6 + 4) \times (0.25) = 4.50$. After the entire sequence has passed through the system, the result is a new sequence which is 3.75, 4.50, 2.75, 2.50, 2.25, 3.75. The new inter-sample difference ranges from 0.25 to 2.25, and we have effectively achieved a low pass filter.

Filter by design

To begin with, coefficients for the filter must be calculated for a given set of parameters. The art of designing non-recursive filters is to achieve acceptable performance using as few coefficients as possible. Typically, a practical fir filter will employ between 10 and 150 coefficients. Theoretically, in order to achieve an ideal response, a fir filter will have an infinite number of taps or coefficients, but in reality a fir filter will have a finite number of taps due to the limitation of processing time and of the hardware.

This truncation effect introduces errors, the effect of which is ringing or ripples in the stop band region. Because an fir filter has a finite response, the actual amplitude response will be different to the theoretical value. This effect is known as the Gibb's phenomenon and has the frequency response of what is known as the rectangular window.

The spectrum of a rectangular window tends to the $\sin(x)/x$ or sinc, form. The method that can be used to reduce this effect is as follows. By multiplying the impulse response sequence with another window function smooths the

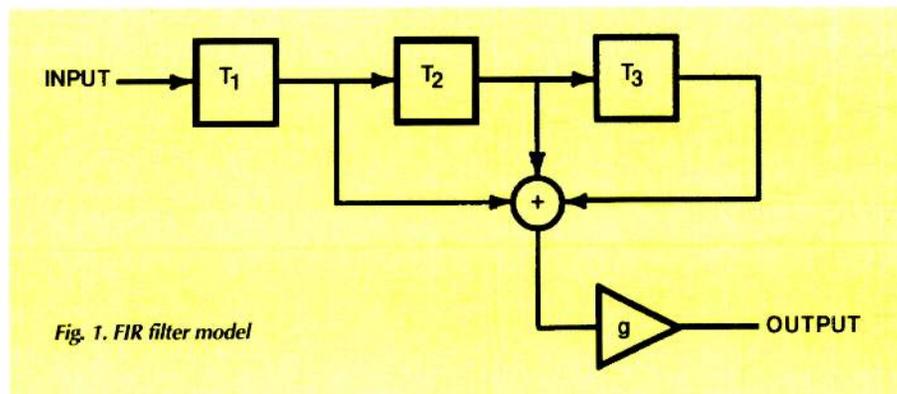


Fig. 1. FIR filter model

ripples in the stop band. There are several window functions that may be used and the two common ones are the Hamming and the Hanning windows.

$$\text{Hanning } 0.5(1 - \cos(2\pi n/N-1));$$

$$\text{Hamming } 0.54 - 0.46\cos(2\pi n/N-1)$$

The number of coefficients used in the filter will depend on the order of filtering required, the cut-off frequency and the sampling frequency used. The relationship of the cut-off frequency and the sampling frequency is an important factor to be considered when deciding how many coefficients should be used in the design. For the filter to be efficient, the length of the delay line (ie the number of taps) must hold at least one complete cycle of samples of the required cut-off frequency.

The calculations employed in the generation of filter coefficients are often complex and difficult to understand. These filter coefficients may early be calculated using computer methods, or may even be performed on a programmable scientific calculator.

The Fourier transform method with windowing is the simplest method of designing a linear phase fir filter. The basic formulae used to calculate the coefficients is

$$H(n) = \frac{\sin(2\pi W(M-n))}{\pi(m-n)} \quad 0 \leq n \leq (M-1)$$

Where *W* = cut-off frequency divided by the sampling frequency, and $M = 1/2(N-1)$ where *N* is the length of the filter.

To achieve exact linear phase, an fir filter has to have a unit-pulse response that is symmetrical around the point $(M-1)/2$. Fig. 2 shows how the coefficients for a 20 tap filter are organised to achieve linear phase. Coefficients 10 to 19 are a mirror copy of the coefficients 0 to 10.

- H(0) = H(19)
- H(1) = H(18)
- H(2) = H(17)
- H(3) = H(16)
- H(4) = H(15)
- H(5) = H(14)
- H(6) = H(13)
- H(7) = H(12)
- H(8) = H(11)
- H(9) = H(10)

Fig. 2. Linear phase filters require a symmetrical coefficient structure.

Once the coefficients have been calculated, the formulae shown below is used to realise the filter. $Y_n = X(N) \cdot H(N) + X(N-1) \cdot H(N-1) + X(N-2) \cdot H(N-2) + X(N-n) \cdot H(N-n)$ *Y_n* is the output of the filter, *X(N)* is the oldest sample, *X(N-n)* is the newest sample and *N* is the

length of the filter. As the values of the coefficients are normalised using real numbers ranging between $-1 \leq 0 \leq 1$, they must be scaled to the processor's data word size. The *TMS320C10* DSP has a 16-bit data word length, therefore all coefficient values must be scaled by multiplying them by 2^{15} to obtain values ranging between: $-32768 \leq 0 \leq 32768$. This process introduces rounding errors which degrade the performance of the filter further. A software package, *FDesign*, is to be used to design the filter. Not only will it calculate the coefficients, but it will generate the *320C10* code as well. The parameters necessary to generate the filter coefficients are: the sampling frequency, the cut-off frequency, the number of taps, the type of window damping.

Once the coefficients have been calculated, the frequency response of the filter is calculated and displayed in graphical form.

In this example, we will design a low pass filter for the developer. The specification of the filter will be as follows:

- Sampling Frequency: 10 KHz
- Cut off Frequency : 2.5 KHz
- Bandwidth of signal : 5.0 KHz
- Number of taps : 20
- Window type : Hamming

Filter coefficients calculated are shown below:

H[00]	=	H[10]	=	-1.39622e-03	ffd3
H[01]	=	H[11]	=	+6.36498e-03	00d0
H[02]	=	H[12]	=	+1.91480e-03	003e
H[03]	=	H[13]	=	-2.09039e-02	fd54
H[04]	=	H[14]	=	+3.14799e-03	0067
H[05]	=	H[15]	=	+5.00330e-02	0667
H[06]	=	H[16]	=	-2.62335e-02	fc5a
H[07]	=	H[17]	=	-1.07157e-01	f249
H[08]	=	H[18]	=	+1.18529e-01	0f2b
H[09]	=	H[19]	=	+4.74540e-01	3cbd

TMS320C10 program generated by *FDesign* which is ready to be assembled and linked.

```

idt 'fir filter'
mask equ 0
mask1 equ 1
xn equ 2
x01 equ 3
x02 equ 4
x03 equ 5
x04 equ 6
x05 equ 7
x06 equ 8
x07 equ 9x08 equ 10
x09 equ 11
x10 equ 12
x11 equ 13
x12 equ 14
x13 equ 15
x14 equ 16
    
```

```

x15 equ 17          data 0x003e          ltd x13
x16 equ 18          data 0xfd54          mpy h13
x17 equ 19          data 0x0067          ;
x18 equ 20          data 0x0667          ; ltd x12
x19 equ 21          data 0xfca5          mpy h12
h00 equ 22          data 0xf249          ;
h01 equ 23          data 0x0f2b          ; ltd x11
h02 equ 24          data 0x3cbd          mpy h11
h03 equ 25          data 0x3cbd          ;
h04 equ 26          data 0x0f2b          ; ltd x10 mpy h10
h05 equ 27          data 0xf249          ;
h06 equ 28          data 0xfca5          ; ltd x09
h07 equ 29          data 0x0667          mpy h09
h08 equ 30          data 0x0067          ;
h09 equ 31          data 0xfd54          ; ltd x08
h10 equ 32          data 0x003e          mpy h08
h11 equ 33          data 0x00d0          ;
h12 equ 34          data 0xffd3          ; ltd x07
h13 equ 35          ; End of coefficients mpy h07
h14 equ 36          data 0x0000          ;
h15 equ 37          tble data 0x0001          ; ltd x06
h16 equ 38          ; mpy h06
h17 equ 39          start lack tble          ;
h18 equ 40          larp 0          ; ltd x05
h19 equ 41          lark 0,one          mpy h05
yn equ 42          make tblr *          ;
one equ 43          sub one          ; ltd x04
aorg 0             banz make          mpy h04
b start          ;
data 0x7fff          loop bioz get          ; ltd x03
data 0x8000          b loop          mpy h03
data 0x0000          get in xn,pa0          ;
data 0x0000          zals mask          ; ltd x02
data 0x0000          xor xn          mpy h02
data 0x0000          sacl xn          ;
data 0x0000          zac          ; ltd x01
data 0x0000          ; mpy h01
data 0x0000          lt x19          ; ltd xn
data 0x0000          mpy h19          mpy h00
data 0x0000          ;
data 0x0000          ; ltd x18          ;
data 0x0000          mpy h18          ;
data 0x0000          ;
data 0x0000          ; ltd x17          ;
data 0x0000          mpy h17          ;
data 0x0000          ;
data 0x0000          ; ltd x16          ;
data 0x0000          mpy h16          ;
data 0x0000          ;
data 0x0000          ; ltd x15          ;
data 0x0000          mpy h15          ;
data 0x0000          ;
; Start of coefficients          ; ltd x14
data 0xffd3          mpy h14
data 0x00d0          ;

```

Loading a filter program

The filter program generated by *fdesign* must be assembled and linked using the Hippo solution's DSP development package. The output file generated by the linker is in Motorola S record format. The developer is connected to an 8-bit ADC+DAC circuit board via a 64-way backplane. The Developer's three channel timer is used to generate the sampling frequency and the input and output anti-aliasing filtering. The output of channel No. 2 is used to generate a signal that is connected to the BIO pin of the 320C10, the channel is programmed to generate a 10000Hz square wave.

The sampling frequency must be twice that of the highest frequency being sampled. Channel No1 and Channel No2 are used to clock switched capacitor filters which provides anti-aliasing filtering, they must be clocked 100 times that of the highest frequency being sampled. The *devcomms* software is used to communicate with the Developer via the PC's serial port. When invoked, it attempts to communicate with the developer. Once communication has been established, a success banner is displayed and the Developer is ready to receive commands and data.

The filter code is loaded with "Load Code", selected from the File menu or by hitting the F3. It takes about 5 seconds to transmit the code to the developer.

Before the DSP is allowed to run its program, the timer has to be programmed. By hitting CTRL T, a control panel is invoked which allows the timer to be programmed. The following frequencies are programmed into the timer:

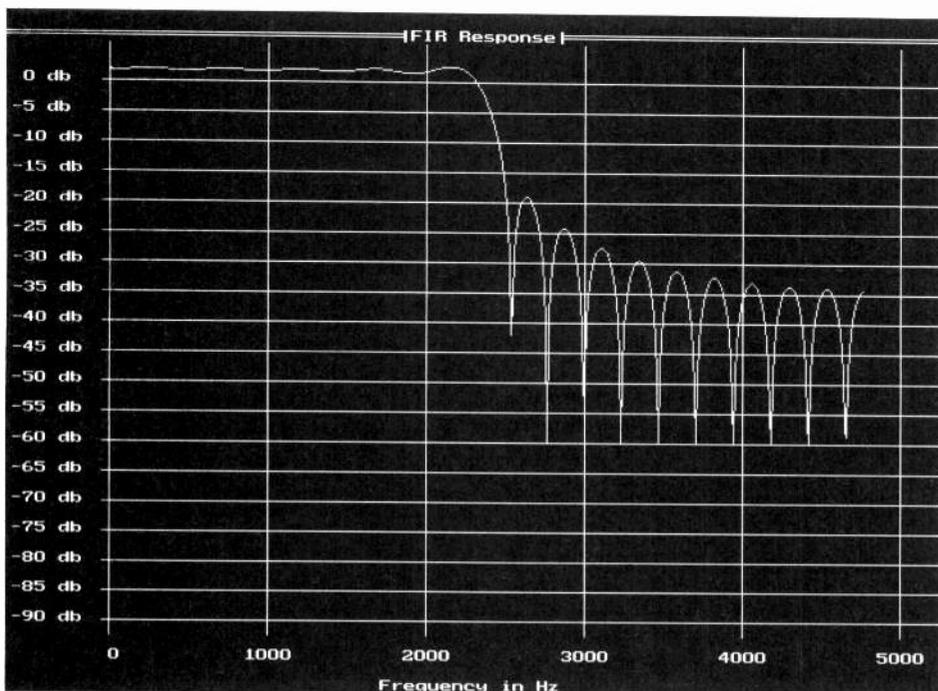
- Channel No. 1 500,000 Hz
- Channel No 2 10,000 Hz
- Channel No 3 500,000 Hz

With an audio signal generator and an oscilloscope connected to the analogue board, the loaded fir filter code can be tested. The filter design generated earlier achieved 60dB attenuation per octave. By varying the number of filter taps, the performance of the number of taps, cut-off frequency and sampling frequency can easily be demonstrated.

Generating sine waves

Sinewave generators are fundamental building blocks of signal processing systems in applications such as communication, instrumentation and control. In the past, engineers designed oscillators using analogue components with their inherent tendency to drift. Using DSP it is possible to produce stable, low distortion sine waves over a wide range of frequencies. Two methods of sinewave generations using the *TMS32010* will be described.

The first frequency response plot left shows the Gibb's effect using a rectangular window, the second plot right shows the effect reduced by using a hamming window. The plots are of a 21 taps FIR filter with a cut-off frequency of 2500Hz using a sampling frequency of 10,000Hz



The developer

The developer is based on the *TMS320C10* running at 20MHz. An *80C31* microcontroller is used as a master processor to the DSP and its serial port is used to connect to an IBM PC. A software package on the PC communicates to the DSP developer which allows DSP programs to be downloaded to the DSP program memory.

The DSP developer is constructed on an extended single euro-card with all the *320C10* data, control and input/output signals connected to the 64-way DIN connector.

The board includes a three channel timer chip controlled by the *80C31* cpu which can be used for anti-aliasing, sampling rate, and reconstruction filtering. The three timer channels outputs are connected to the 64-way DIN connector. A backplane is available to connect the DSP board to analogue/digital interface cards complete with 64 way DIN connectors.

The *80C31* has full access to the DSP program memory which can be loaded with a program or contents read. The output port of the *80C31* has control of the *TMS32010* reset line.

The developer comes with two software packages, the IBM PC Developer interface and real-time firmware in an EPROM plugged into the DSP developer board. The PC Developer interface software is a set of windows and pop down menus which allow the user to control the developer via the serial interface using simple commands.

The software package used to develop the programs for the *320C10* is the Hippo Solutions TMS320CXX development tools. Together with the Developer, it offers a complete and low cost development system for the DSP chip.

The software system consists of a compiler, assembler, linker and simulator. Broadly speaking, the compiler translates 'C' language subset source into assembly language. These sources are

combined with other assembly language sources, assembled and linked into a single executable image which can be either split into high and low byte images with the provided utility 'UIX' or loaded into the simulator or downloaded into the DSP Developer to be run.

On its own, the DSP developer board can do little. A suitable input/output board must be connected to the developer via the 64-way back plane. An eight bit adc/dac circuit is shown on Fig. xx. It uses the popular *DAC800* dac chip and the Analog Devices *ADC7575* analog ADC chip. Anti-aliasing filtering is accomplished with a 2nd order low pass filter and an 8th order clock tunable low pass filter using a *MAX291* chip from MAXIM. The clocking frequency for the filter is provided by the *82C53* timer channel 0 under the control of the *80C31* controller. The clocking frequency must be a hundred times that of the highest frequency being sampled.

Sampling is initiated by reading the *ADC7575* which loads the contents of the previous sample and starts the next sampling process. The ADC has a built in sample and hold circuit. The sampling period can be determined by the use of the *320C10*'s BIO pin connected to the output of the monostable. The *82C53* timer 3 is programmed to trigger the monostable.

The output of the dac is connected to another *MAX291* filter circuit which reconstructs the signal and filters out the sampling frequency. The conventional 2nd order filter circuit removes any residue clocking signal. The *82C53* timer 0 provides the clock for the switching capacitor filter.

The DSP developer and the TMS3201X Software development tools can be obtained from Tycho Designs 38, Playfields Drive, Branksome, Poole, Dorset BH12 2EQ Phone 0202-736791. Prices start from about £100.

The first is the fast direct table lookup method and the second is the use of IIR filters.

Direct table lookup sine generator

The first algorithm is a simple, fast table lookup process. The sine values of N angles which are uniformly spaced round the unit circle are stored in a table. The values have the

following format:

INDEX	SINE TABLE
0	SINE(0 * (360 /N))
1	SINE(1 * (360 /N))
2	SINE(2 * (360 /N))
3	SINE(3 * (360 /N))
N-2	SINE(N - 2 * (360 /N))

A sine wave is generated by stepping through the table at a constant rate, wrapping

round at the end of the table when 360° is exceeded. What determines the frequency are the step size in degrees between samples and the time interval between successive samples (ie the sampling interval f_s). The frequency is given by the equation:

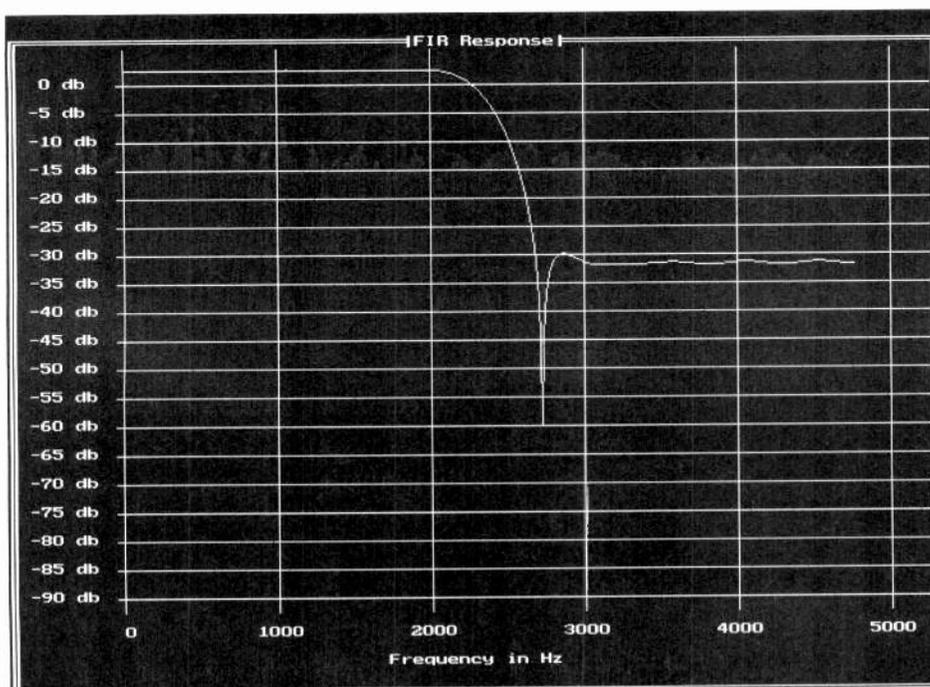
$$f = \text{step size} / (f_s \times \text{table size})$$

The accuracy of the sine wave depends on the size of the table, the greater the number of steps in the table the greater the accuracy is. To satisfy the Nyquist criterion there must be at least two samples generated for each sinusoidal period. The values in the table are all normalised using the two's complement hexadecimal Q14 notation. The decimal values between +1.0 and -1.0 are multiplied by 16384. Rounding is applied, rather than truncation to reduce further distortion. To run the sine wave generator into the DSP developer, use the assembler and the linker to build an executable .COD file from the following program.

```

B      START
DELTA EQU 0
ALPHA EQU 1           ;Mask to
confine addresses to table
SINA EQU 2           ;Start
address of sine table
TEMP EQU 3           ;Integer part
phase step
MASK EQU 4           ;Fractional
part phase step
CFSET EQU 5          ;Phase
accumulator integer part
CUTV EQU 7

```



Then load the program into the developer. The following frequencies are programmed into the timer:

- Channel No 2 10000 Hz
- Channel No 3 500000 Hz

Sinewave generation using IIR filters

One disadvantage using IIR filters is that they can become unstable under certain conditions. These conditions occur when the poles of the filter function approach and extend past the unit circle. Without going into too much detail, poles and zeros determines the behaviour and impulse response of a recursive filter. If an pulse of finite time and magnitude is injected into a second order IIR filter with its poles located within the unit circle, a sinewave will be generated with its amplitude decaying exponentially in time. However, if the same pulse is injected into the same type of filter with its poles located outside the unit circle, the same sinewave is generated. But this time, the amplitude of the wave will increase exponentially in time until the filter saturates. Now, if the poles of the filter are calculated so that they reside on the unit circle and a pulse is applied, the filter will oscillate forever. The formulae to generate the sinewave is as follows:

$$Y[n] = (y[n-1] \times (w)) - (C2 \times y[n-2]) + Xn$$

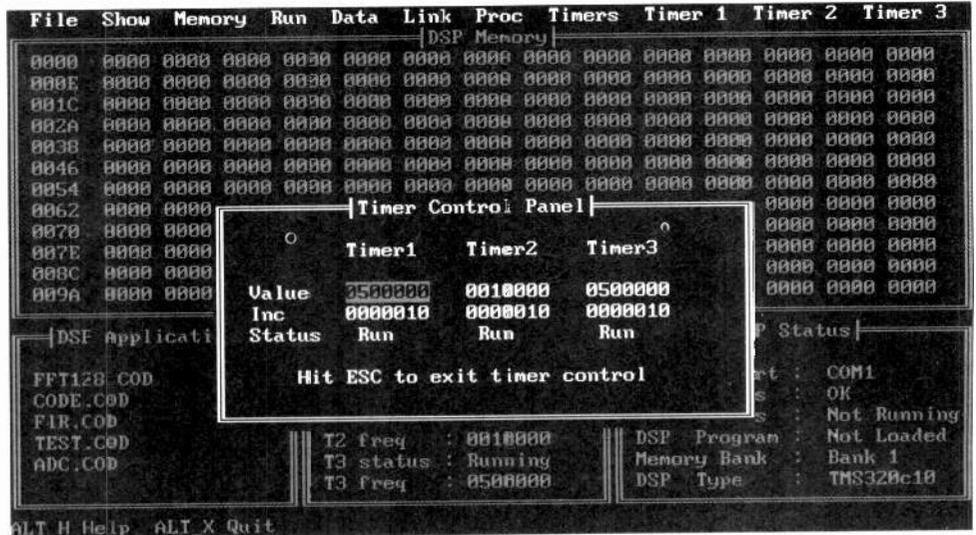
Where $Y[n]$ is the output of the filter $y[n-1]$ and $y[n-2]$ are the direct form II IIR filter delay elements.

C2 is the coefficient representing the radius of the pole of the filter.

The frequency is calculated using the following equation : $w = 2\pi(\text{Freq}/\text{Sampling Freq})$

Example : $F_{\text{sampling}} = 10\text{kHz}, F_{\text{sinewave}} = 500\text{Hz}$
 $w = 2 \times \cos(2\pi(500/10,000)) = 1.90211$.

w must be converted to Q14 notation, therefore $w = 1.90211 \times 16384 = 31164$. Below is



a program written in 'C' and compiled with the Hippo solution 'C' compiler and assembler. To demonstrate the filter will decay if the C2 is set to value of less than 1, change the value to 0.999999. You will see that the oscillation will decay. If the value of C2 is set to 1.0001, the filter will go unstable. Several sinewave can be generated simultaneously by duplicating the formulae variable and sine function subroutine. Care must be taken not to saturate the output of the DAC when adding the resulting output value of each of the tones.

The following frequencies are programmed into the timer:

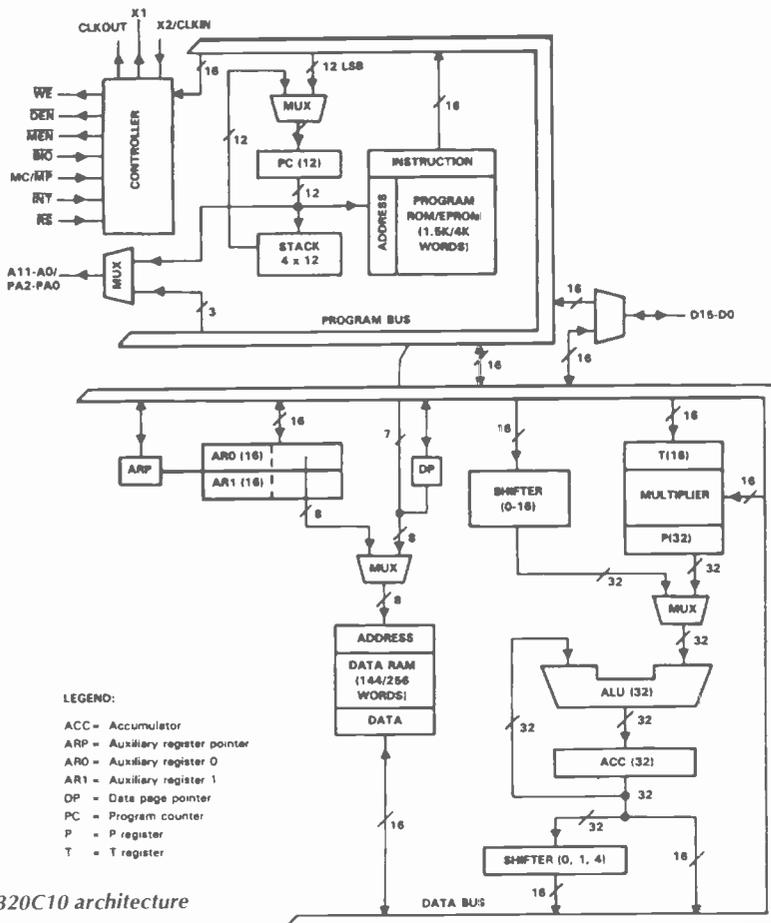
- Channel No 2 10,000Hz
- Channel No 3 500,000Hz

```
#define cut \
    #asm \
    out [00],01 \
```

The Developer has a three channel timer connected to the backplane. Channel No 1 is used to clock a switched capacitor filter circuit used for anti-aliasing filter, channel No 2 is used for generating the sampling frequency and channel No 3 is used to clock a filter circuit for the reconstruction of the signal.

```
#endasm
#define wait \
    #asm \
    copy 'wait.asm'\
    #endasm
int y1=0, y2=0; /* Direct Form II
IIF filter delay elements */
int OutVal; /* Output value
*/
int A = +1.90211 ; /* 500 Hz 2ccs(w)
*/
int C2 = 1.0 /* Radius of the
pole */

main()
{
    int impulse= 500; /* Initial
pulse amplitude */
    while (1)
    {
        wait ; /* wait for the BIO
pin */
        out(OutVal, 0); /* Output the
value */
        /* DAC8000 binary offset DAC */
        OutVal=(sine(impulse)- 2048) >> 4;
        impulse = 0 ; /* set pulse to
zero */
    }
    /* Sinewave generation subroutine */
    far sine(int in) /* y[n] = A x y[n-1]
- C2 x y[n-2] + x[n] */
    {
        int tmp, y;
        tmp= D1 <*> A -(C2 <*> D2) + in;
        D2=D1;
        D1=tmp;
        return tmp;
    }
    Assembler routine for the sampling of
the BIO pin which is copied
into the main 'C' program at compile
time.
loop bioz go
    b loop
go nop
```



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“Power tends to corrupt.” First Baron Acton.

Class B output stages differ in at least three different ways, presenting some intractable design problems. But class AB proves to be no answer, says Douglas Self.

Distortion in power amplifiers

4: the power amplifier stages

The almost universal choice in semiconductor power amplifiers is for a unity gain output stage, and specifically a voltage follower. Output stages with gain are not unknown¹, but they are not common. Most designers feel that controlling distortion while handling large currents is hard enough without trying to generate gain at the same time.

The first three parts of this series have dealt with one kind of distortion at a time, due to the monotonic transfer characteristics of small signal stages, which usually, but not invariably, work in class A². Economic and thermal realities mean that most output stages are class B, and so we must now consider crossover distortion, which remains the thorniest problem in power amplifier design, and HF switchoff effects.

It is now also necessary to consider what kind of active device is to be used; jfets offer few if any advantages in the small current stages, but power fets are a real possibility, providing that the extra cost brings with it tangible benefit.

The class war

The fundamental factor in determining output stage distortion is the class of operation. Apart from its inherent inefficiency, class A is the ideal operating mode, because there can be no crossover or switchoff distortion. However, of those designs which have been published or reviewed, it is notable that the large signal distortion produced is still significant. This looks like an opportunity lost, as of the distortion mechanisms discussed in the first part of this series, we now only have to deal with Distortion 1 (input stage), Distortion 2 (vas),

and Distortion 3 (output stage large signal non-linearity). Distortions 4,5,6 and 7, as mentioned earlier, are direct results of class B operation and therefore can be thankfully disregarded in a class A design. However, class B is overwhelmingly of the greater importance, and is therefore dealt with in detail.

Class B is subject to misunderstanding. The statement is often made that a pair of output transistors operated without any bias are working in “class B”, and therefore “generate severe crossover distortion”. In fact, with no bias each output device is operating for slightly less than half the time, and the question arises as to whether it would not be more accurate to call this class C and reserve class B for that condition of quiescent current which eliminates, or rather minimises, the crossover artifacts.

A further complication exists: it is not generally appreciated that moving into what is usually called class AB, by increasing the quiescent current, does not make things better. In fact, the THD reading will increase as the bias control is advanced, with what is usually known as “ g_m doubling” (ie a voltage gain increase caused by both devices conducting simultaneously in the centre of the output voltage range) putting edges into the distortion residual that generate high order harmonics in much the same way that underbiasing does. This important fact seems almost unknown, presumably because the g_m doubling distortion is at a relatively low level and is completely obscured in most amplifiers by other distortion mechanisms.

The phenomenon is demonstrated in **Figs. 1a,b,c** which shows spectrum analysis of the distortion residuals for under biasing, optimal,

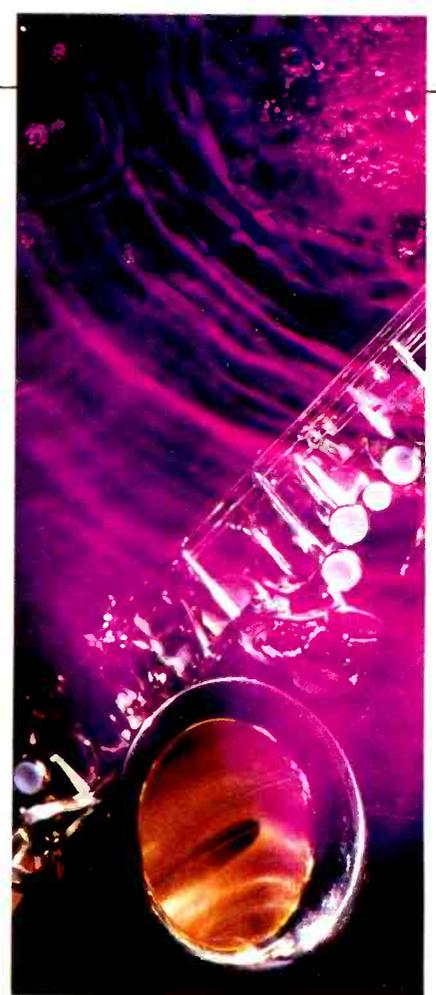
and over biasing of a 150W/8 Ω amplifier at 1kHz.

As before, all non-linearities except the unavoidable Distortion 3 (output stage) have been effectively eliminated. The over biased case had its quiescent current increased until the g_m doubling edges in the residual had an approximately 50:50 mark/space ratio, and so was in class A about half the time which represents a rather generous amount of quiescent for class AB. Nonetheless, the higher order odd harmonics in Fig.1c are at least 10dB greater in amplitude than those for the optimal class B case, and the third harmonic is actually higher than for the under-biased case as well. However the under biased amplifier, generating the familiar sharp spikes on the residual, has a generally greater level of high-order odd harmonics above the 5th; about 8dB higher than the AB case.

Bearing in mind that high order odd harmonics are generally considered to be the most unpleasant, there seems to be a clear case for avoiding Class AB altogether, as it will always be less efficient and generate more high order distortion than the equivalent class B circuit. class distinction therefore seems to resolve itself into a binary choice between A or B.

It must be emphasised that these effects can only be seen in an amplifier where the other forms of distortion have been properly minimised. The rms THD reading for case **1a** was 0.00151%, for case **1b** 0.00103%, and for case **1c** 0.00153%. The tests were repeated at the 40W power level with very similar results. The spike just below 16kHz is interference from the test gear VDU.

This may seem complicated enough, but there are other and deeper subtleties in class B.



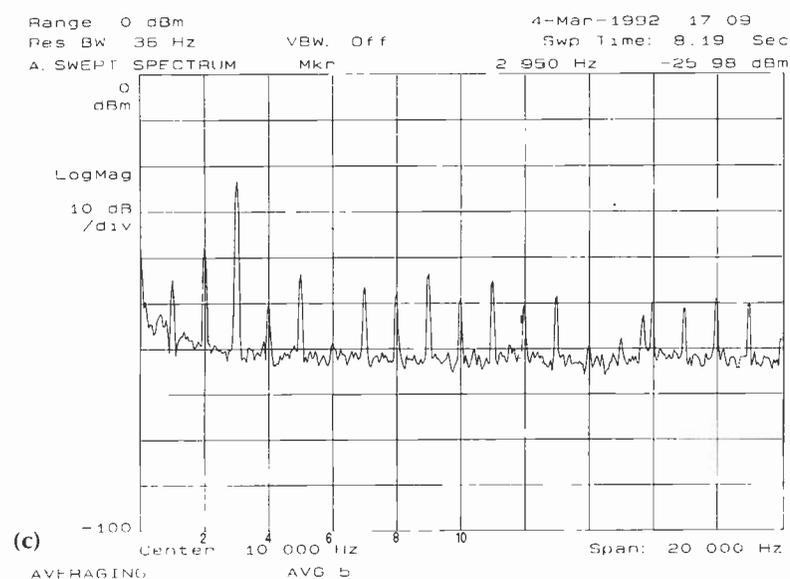
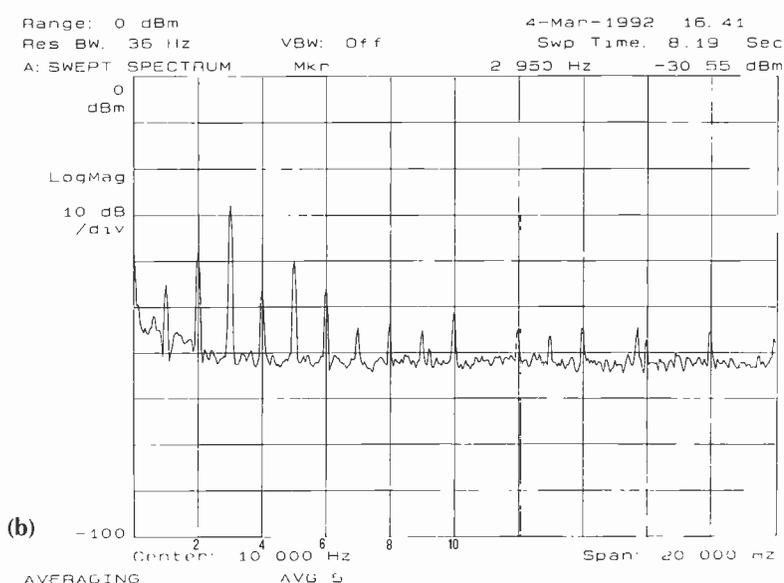
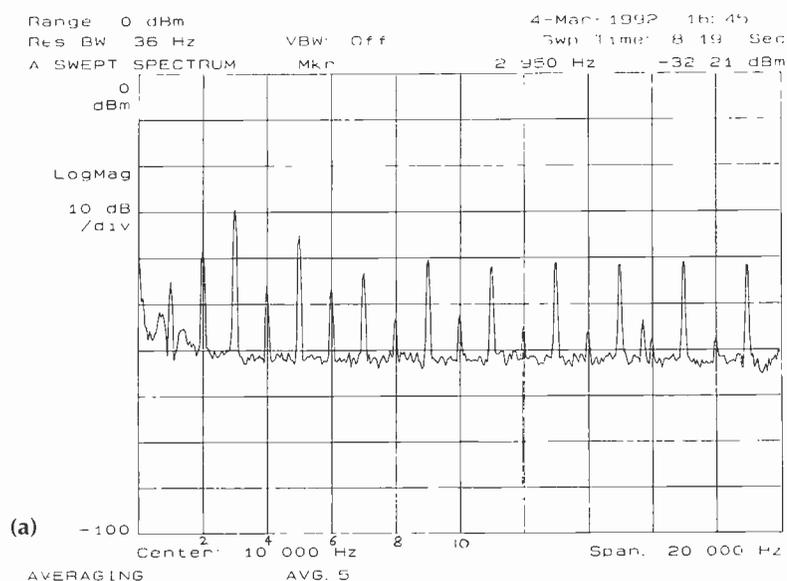


Fig. 1. Spectrum analysis of class B & AB distortion residual. 1a) Under-biased class B; 1b) Optimal class B; 1c) class AB.

Distortions of the output

I have designated the distortion produced directly by output stages as Distortion 3 (see Part 1); this subdivides into three categories. Mechanism 3a describes the large signal distortion produced by both class A and B, ultimately because of the large current swings in the active devices. In bipolars, but not fets, large collector currents reduce the beta leading to drooping gain at large output excursions. I shall use the term "LSN" for large signal non-linearity, as opposed to crossover and switchoff phenomena that cause trouble at all output levels.

The other two contributions to Distortion 3 are associated with class B only; Distortion 3b the classic crossover distortion resulting from the non-conjugate nature of the output characteristics, and is essentially frequency independent.

In contrast, Distortion 3c is switchoff distortion generated by the output devices failing to turn off quickly and cleanly at high frequencies. This mechanism is strongly frequency dependent. It is sometimes called switching distortion, but this allows room for confusion, as some writers use "switching distortion" to cover crossover distortion as well. I refer specifically to charge storage turn off troubles.

One of my aims for this series has been to show how to isolate individual distortion mechanisms. To examine output behaviour, it is perfectly practical to drive output stages open loop providing the driving source impedance is properly specified; this is difficult with a conventional amplifier, as it means the output must be driven from a frequency dependant impedance simulating that at the vas collector with some sort of feedback mechanism incorporated to keep the drive voltage constant.

However, if the vas is buffered from the output stage by some form of emitter follower, as described in the last part, it makes things much simpler, a straightforward low impedance source (eg 50Ω) providing a good approximation of a vas-buffered closed loop amplifier. The vas buffer makes the system more designable by eliminating two variables – the vas collector impedance at LF, and the frequency at which it starts to decrease due to local feedback through C_{dom} . This markedly simplifies the study of output stage behaviour.

The large signal linearity of various kinds of open loop output stage with typical values are shown in Figs. 6-15. These diagrams were all generated by spice simulation, and are plotted as incremental output gain against output voltage, with the load resistance stepped from 16Ω to 2Ω. The power devices are MJ802 and MJ4502, which are more complementary than many transistor pairs, and minimise distracting large signal asymmetry. The quiescent current

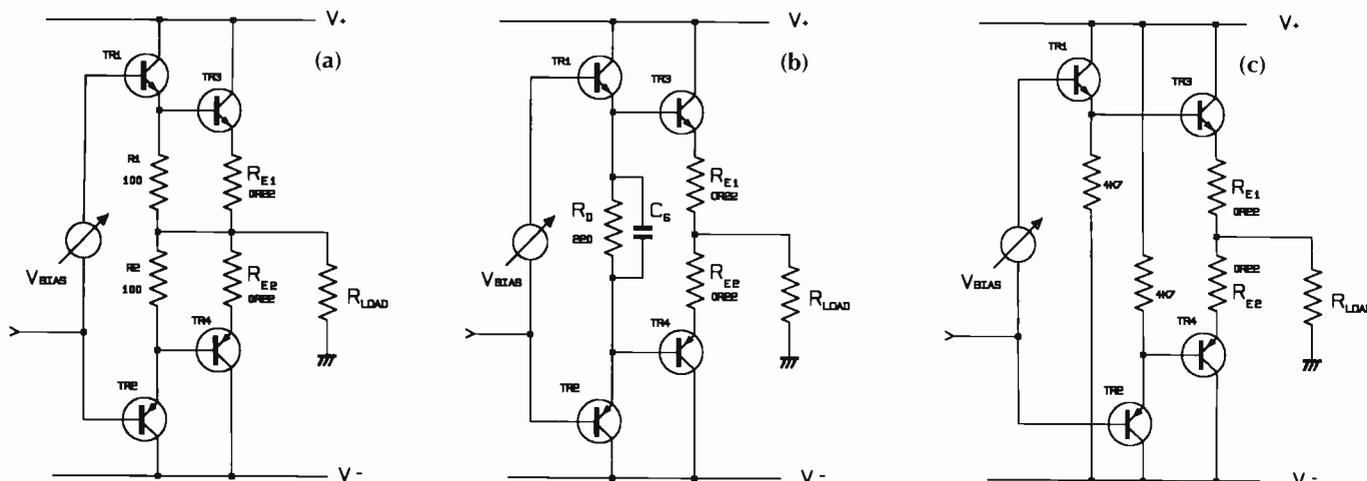
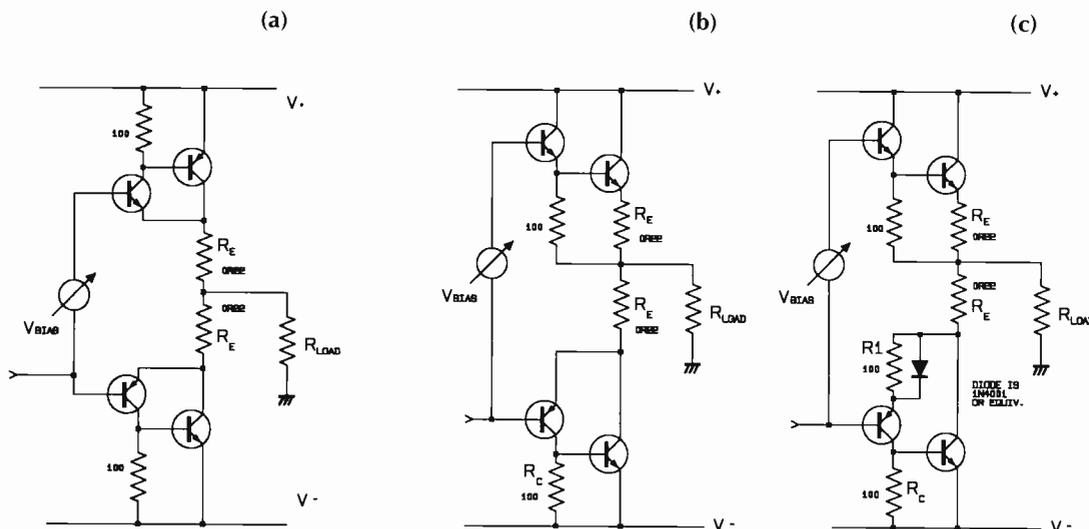


Fig. 2. Three types of emitter follower output stages.

Fig. 3. CFP circuit and quasi complementary stages.



is in each case set to minimise the peak deviations of gain around the crossover point for 8Ω loading; for the moment it is assumed that you can set this accurately and keep it where you want it. The difficulties in actually doing this will be examined later.

There are at least 16 distinct configurations in straightforward output stages not including error correcting³, current dumping⁴ or Blomley⁵ types. These are as follows:

Emitter Follower	3 types	Fig. 2
Complementary Feedback Pair	1 type	Fig. 3
Quasi Complementary	2 types	Fig. 3
Output Triples	At least 7 types.	Fig. 4
Power FET	3 types	Fig. 5

The emitter follower output

Figure 2 shows three versions of the most common type of output stage; the double-emitter follower where the first follower acts

as driver to the second (output) device. I have deliberately called this an emitter follower rather than a Darlington configuration, as this latter implies an integrated device with associated resistors. As for all the circuitry here, the component values are representative of real practice.

Two important attributes of this topology are: The input is transferred to the output via two base emitter junctions in series, with no local feedback around the stage (apart from the very local 100% voltage feedback that makes an emitter follower what it is);

There are two dissimilar base emitter junctions between the bias voltage and the emitter resistor R_e , carrying different currents and at different temperatures. The bias generator must attempt to compensate for both at once, though it can only be thermally coupled to one. The output devices have substantial thermal inertia and thus thermal compensation

represents a time average of the preceding conditions. Fig. 2a shows the most prevalent version (type I) which has its driver emitter resistors connected to the output rail.

The type II configuration in Fig. 2b is at first sight merely a pointless variation on type I, but in fact it has a valuable extra property. The shared driver emitter resistor R_d , with no output rail connection, allows the drivers to reverse bias the base emitter junction of the output device being turned off.

Assume that the output voltage is heading downwards through the crossover region; the current through R_{e1} has dropped to zero, but that through R_{e2} is increasing, giving a voltage drop across it, so Tr_4 base is caused to go more negative to get the output to the right voltage. This negative excursion is coupled to Tr_3 base through R_d , and with the values shown can reverse bias it by up to 0.5V, increasing to 1.6V with a 4Ω load. The speed

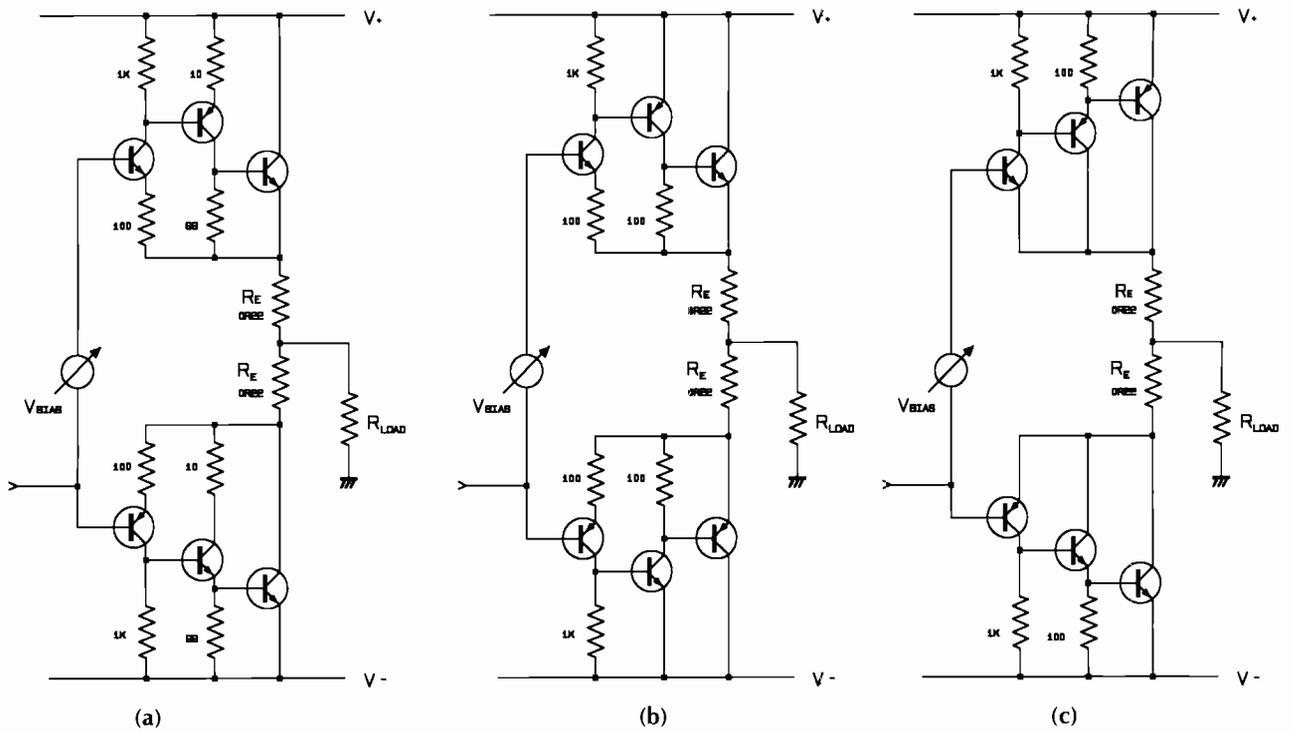


Fig. 4. Three of the possible output triple configurations.

up capacitor C_s markedly improves this action, preventing the charge suckout rate being limited by the resistance of R_d . While the type I circuit has a similar voltage drop across R_{e2} , the connection of the mid point of R_1, R_2 to the output rail prevents this from reaching Tr_3 base; instead Tr_1 base is reverse biased as the output moves negative, and since charge storage in the drivers is usually not a problem, this does little good. In the type II circuit the drivers are never reverse biased, though they do turn off. The important issue of output turn off and switching distortion is further examined in the next part of this series.

The type III topology shown in Fig. 2c maintains the drivers in class A by connecting the driver emitter resistors to the opposite supply rail rather than the output rail. It is a common misconception⁶ that class A drivers somehow maintain better low frequency control over the output devices, but I have yet to substantiate any advantage myself. The driver dissipation is of course substantially increased, and nothing seems to be gained at LF as far as the output transistors are concerned, for in both type I and type II the drivers are still conducting at the moment the outputs turn off, and are back in conduction before the outputs turn on, which would seem to be all that matters.

Type III is equally good as type II in reverse biasing the output bases, and may give even cleaner HF turn off as the carriers are being swept from the bases by a higher resistance terminated in a higher voltage approximating constant current drive; I have yet to try this.

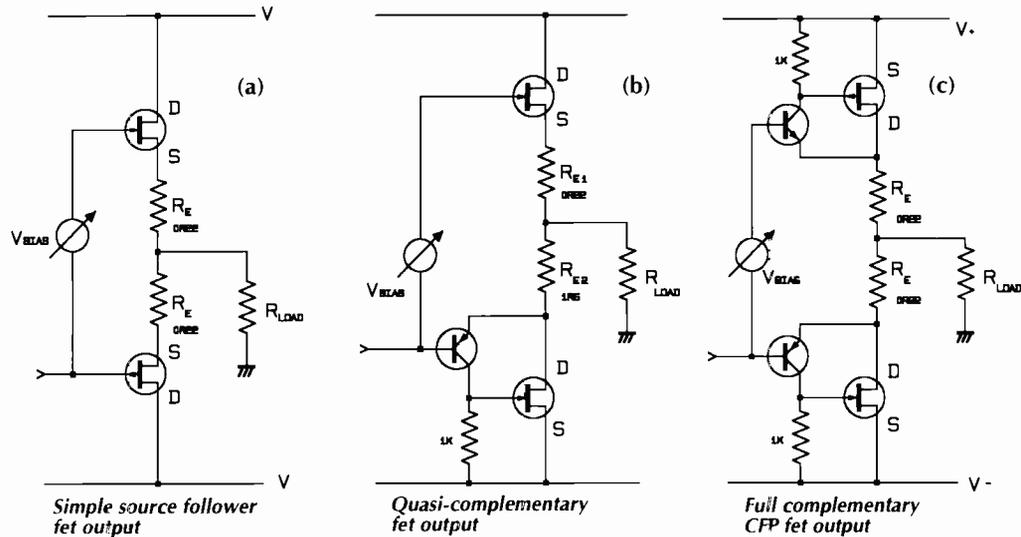


Fig. 5. Three mosfet output architectures.

The large signal linearity of the three versions is virtually identical – all have the same feature of two base emitter junctions in series between input and load.

The gain/output voltage plot is shown at Fig. 6; with BJTs the gain reduction with increasing loading is largely due to the emitter resistors. Note that the crossover region appears as a relatively smooth wobble rather than a jagged shape. Another major feature is the gain droop at high output voltages and low loads indicating that high collector currents are the fundamental cause of this.

A close up of the crossover region gain for 8Ω loading only is shown in Fig. 7; note that no V_{bias} setting can be found to give a constant or even monotonic gain; the double dip and central gain peak are characteristic of optimal adjustment. The region extends over about $\pm 5V$, independent of load resistance.

Complementary feedback output

The other major type of bipolar output is the complementary feedback pair (CFP) sometimes called the Sziklai Pair. Fig. 3a. There seems to be only one popular configuration, though versions with gain are possible. The drivers are now placed so that they compare output voltage with that at the input. Wrapping the outputs in a local negative feedback loop promises better linearity than emitter follower versions with 100% feedback applied separately to driver and output transistors.

This topology also has better thermal stability, because the V_{be} of the output devices is inside the local feedback loop, and only the driver V_{be} affects the quiescent current. It is usually simple to keep drivers cool, and thermal feedback from them to the V_{bias} generator transistor can be much faster and mechanically simpler.

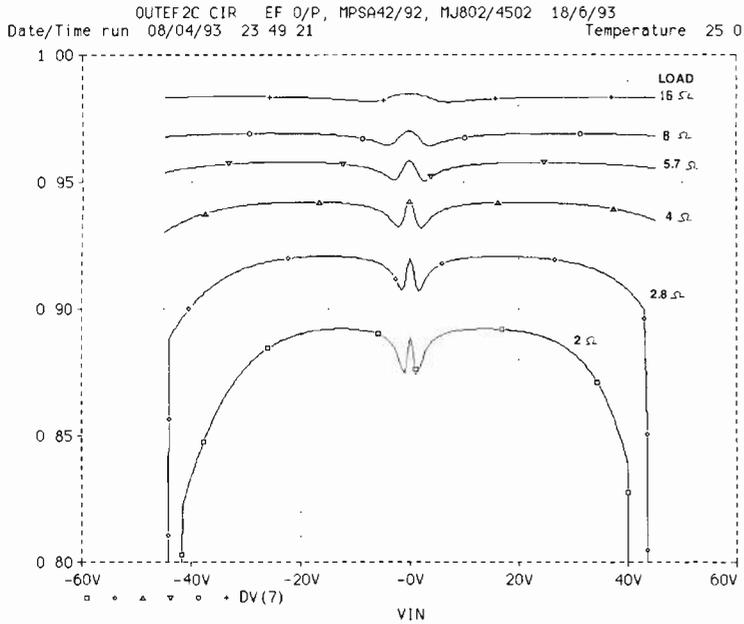


Fig. 6. Emitter follower large signal gain vs output.

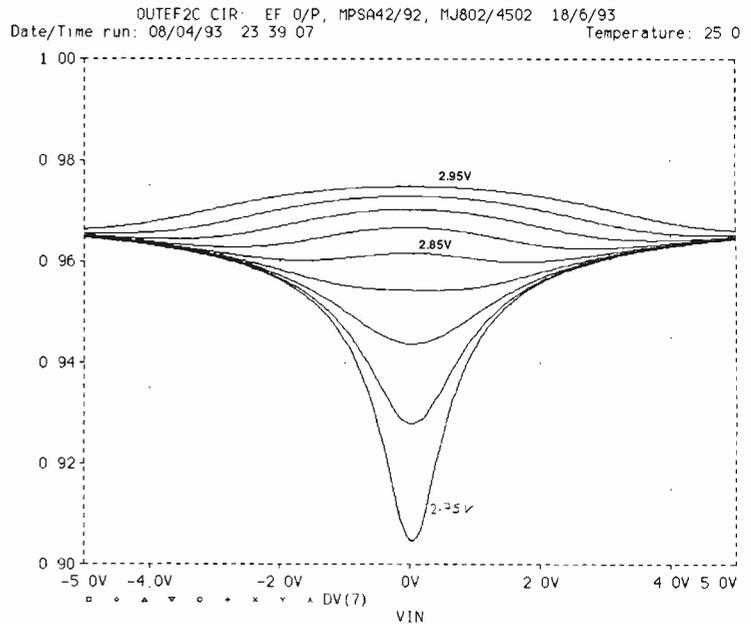


Fig. 7. Emitter follower crossover region gain deviations, ±5V range.

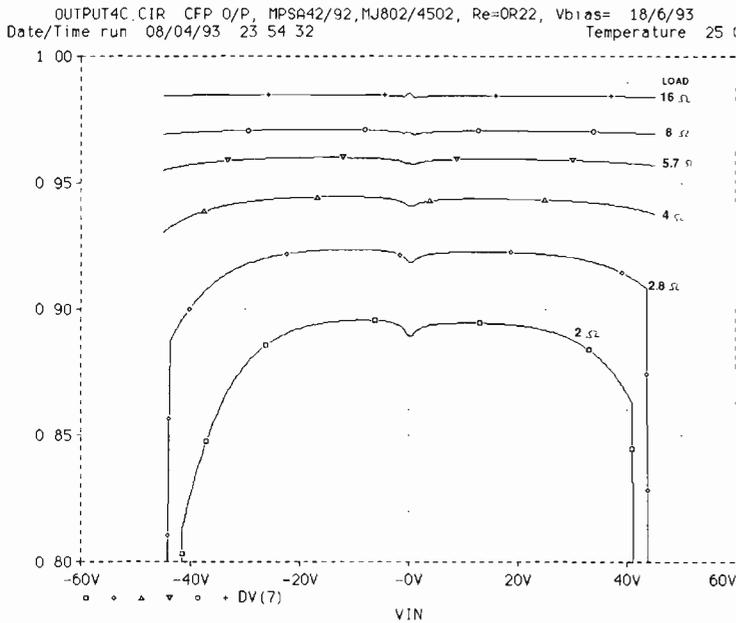


Fig. 8. Complementary feedback pair gain vs output.

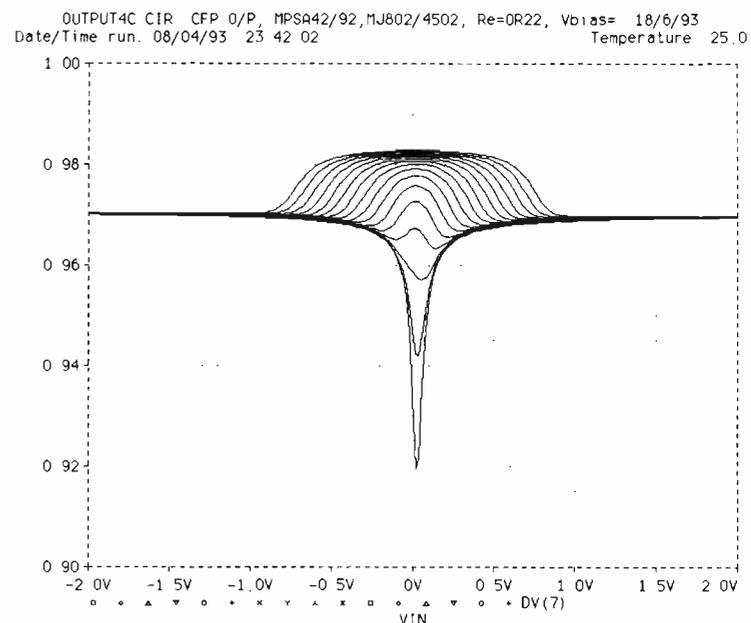


Fig. 9. CFP crossover region ±2V, Vbias as a parameter.

Like emitter follower outputs, the drivers are conducting whenever the outputs are, and so special arrangements to keep them in class A seem pointless. This stage, like emitter follower type I, can only reverse bias the driver bases rather than the outputs, unless extra voltage rails outside the main ones are provided.

The output gain plot is shown in Fig. 8. Fourier analysis shows that the CFP generates less than half the large signal distortion of an emitter follower stage. (See Table 1) Given also the greater quiescent stability, it is hard to see why this topology is not more popular.

The crossover region is much narrower, at about ±0.3V (Fig. 9). When under biased, this shows up on the distortion residual as narrower spikes than an emitter follower output gives.

The bad effects of g_m doubling as V_{bias} increases above optimal (here 1.296V) can be seen in the slopes moving outwards from the centre.

Quasi complementary outputs

The original quasi complementary configuration⁷ was almost mandatory, as it was a long time before pnp silicon power transistors matched the performance of the npn versions. The standard version shown at Fig. 3b is well known for poor symmetry around the crossover region, as shown at Fig. 10. A close-up of the crossover region (Fig. 11) reveals an unhappy hybrid of the emitter follower and CFP, as might be expected, and that no setting of bias voltage can remove the sharp edge in the gain plot.

A major improvement to symmetry may be made by using a Baxandall diode⁸ as shown in Fig. 3c. This stratagem yields gain plots very similar to those for the true complementary emitter follower at Figs 6, 7, though in practice the crossover distortion seems rather higher. When a quasi Baxandall stage is used closed loop in an amplifier in which distortion mechanisms 1 and 2, and 4 to 7 have been properly eliminated, it is capable of better performance than is commonly believed. For example, 0.0015% (1kHz) and 0.015% (10kHz) at 100W is straightforward to obtain from an amplifier with a negative feedback factor of about 34dB at 20kHz.

The best reason to use the quasi Baxandall approach today is to save money on output devices, as pnp power transistors remain

	Emitter Follower	CFP	Quasi Simple	Quasi Bax	Triple Type 1	Simple mosfet	Quasi mosfet	Hybrid mosfet
8Ω THD:	0.031%	0.014%	0.069%	0.050%	0.13%	0.47%	0.44%	0.052%
Gain:	0.97	0.97	0.97	0.96	0.97	0.83	0.84	0.97
4Ω THD:	0.042%	0.030%	0.079%	0.083%	0.60%	0.84%	0.072%	0.072%
Gain:	0.94	0.94	0.94	0.94	0.92	0.72	0.73	0.94

Table 1. This summarises the spice curves for 4 and 8Ω loadings. Each was subjected to Fourier analysis to calculate THD results for a ±40V input.

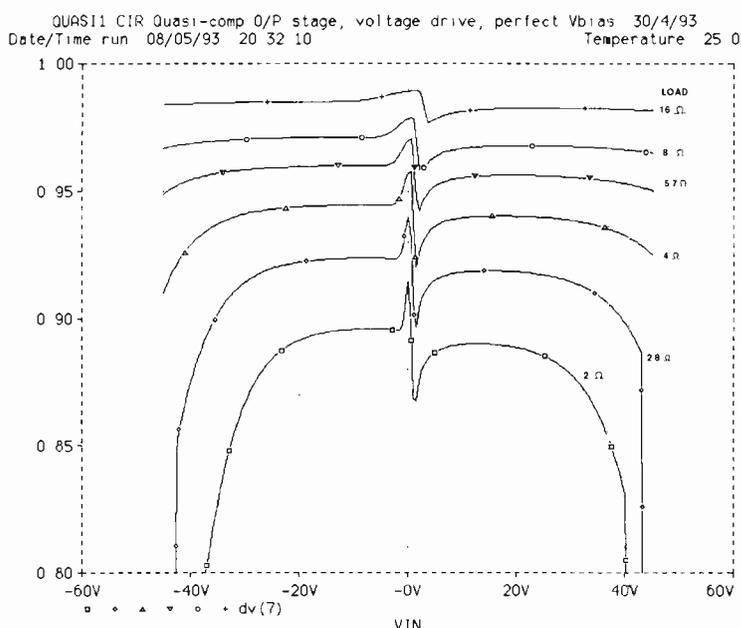


Fig. 10. Quasi complementary large signal gain vs output load resistance

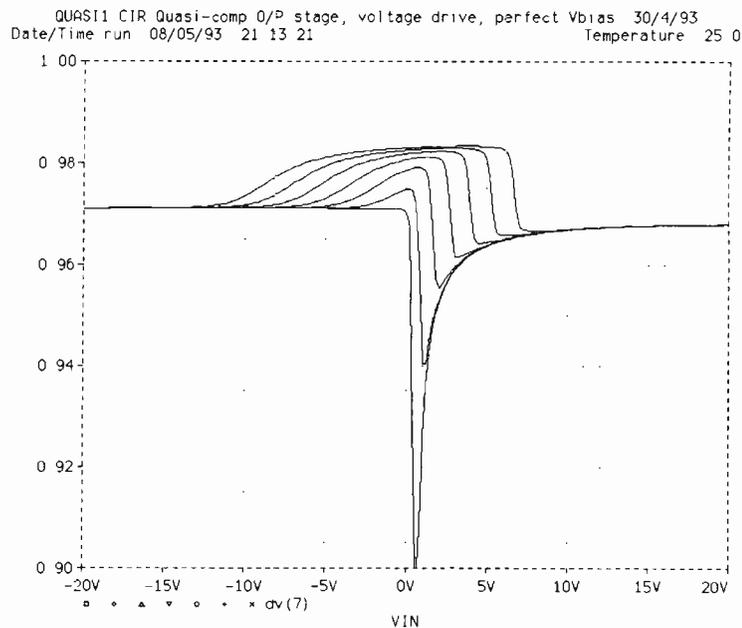


Fig. 11. Quasi crossover region ±20V, Vbias as parameter.

somewhat pricier than npns. Given the tiny cost of a Baxandall diode, and the absolutely dependable improvement it gives, there seems no reason why anyone should ever use the standard quasi circuit. My experiments show that the value of R_1 in Fig. 3c is not critical; making it about the same as R_c seems to work well.

Triples

With three rather than two bipolar transistors in each half of an output stage the number of circuit permutations possible leaps upwards. There are two main advantages if output triples are used correctly: better linearity at high output voltages and currents; and more stable quiescent setting as the pre drivers can be arranged to handle very little power, and remain almost cold in use.

However, triples do not automatically reduce crossover distortion, and they are, as usually configured, incapable of reverse biasing the output bases to improve switch-off. Fig. 4 shows three ways to make a triple output stage – all of those shown (with the possible exception of Fig. 4c, which I have just made up) have been used in commercial designs. The circuit of 4a is the Quad 303 quasi complementary triple. The design of

triples demands care, as the possibility of local HF instability in each output half is very real.

Power fet outputs

Power mosfets are often claimed to be a solution to all amplifier problems, but they have their drawbacks: poor linearity and a high on-resistance that makes output efficiency mediocre. The high frequency response is better, implying that the second pole P2 of the amplifier response will be higher, allowing the dominant pole P1 be raised with the same stability margin, and in turn allowing more over-all feedback to reduce distortion. However, the extra feedback (if it proves available in practice) is needed to correct the higher open loop distortion.

To complicate matters, the compensation cannot necessarily be lighter because the higher output resistance makes the lowering of the output pole by capacitive loading more likely.

The extended frequency response creates its own problems; the HF capabilities mean that rigorous care must be taken to prevent parasitic oscillation, as this is often promptly followed by an explosion of disconcerting violence. Fets should at least give freedom from switchoff troubles as they do not suffer from charge storage effects.

Three types of fet output stage are shown in Fig. 5. Figures 12 to 15 show spice gain plots, using 2SK135/2SJ50 devices.

Most fet amplifiers use the simple source follower configuration in Fig. 5a; the large signal gain plot at Fig. 12 shows that the gain for a given load is lower, (0.83 rather than 0.97 for bipolar, at 8Ω) because of low g_m . This, with the high on resistance, noticeably reduces output efficiency.

Open loop distortion is markedly higher; however large signal non-linearity does not increase with heavier loading, there being no equivalent of “bipolar gain droop”. The crossover region has sharper and larger gain deviations than a bipolar stage, and generally looks pretty nasty; Fig. 13 shows the difficulty of finding a ‘correct’ V_{bias} setting.

Fig. 5b shows a hybrid (ie bipolar/fet) quasi complementary output stage⁹. The stage is intended to maximise economy rather than performance, once the decision has been made (probably for marketing reasons) to use fets, by making both output devices cheap n-channel devices; complementary mosfet pairs remain relatively rare and expensive.

The basic configuration is badly asymmetrical, the hybrid lower half having a higher and more constant gain than the source fol-

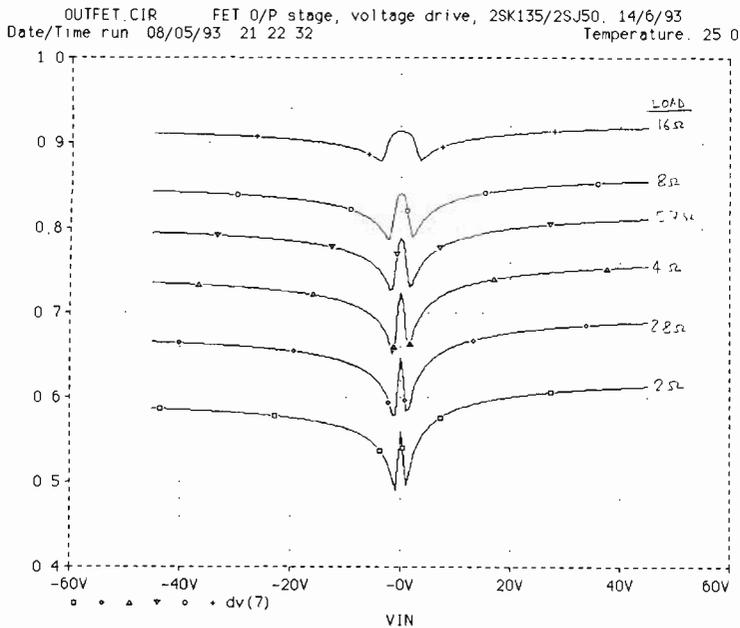


Fig. 12. Source follower fet large signal gain vs output.

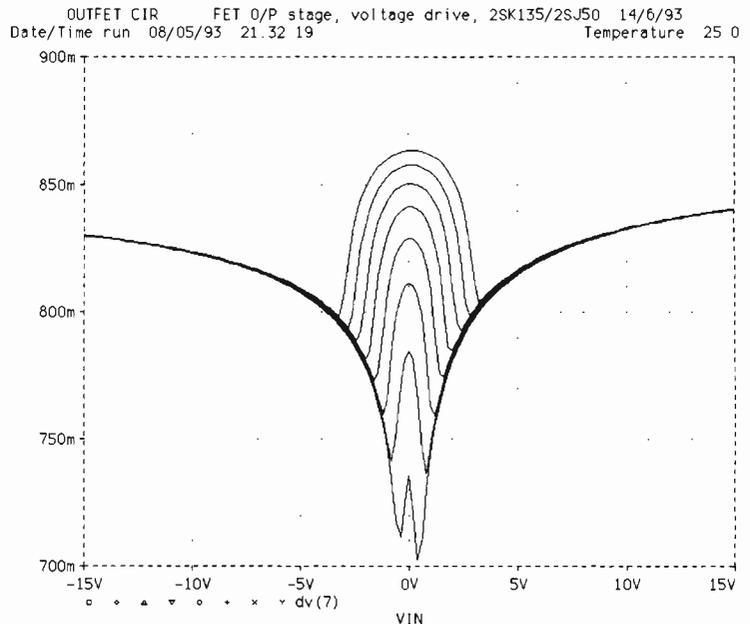


Fig. 13. Source follower fet crossover region ±15V range.

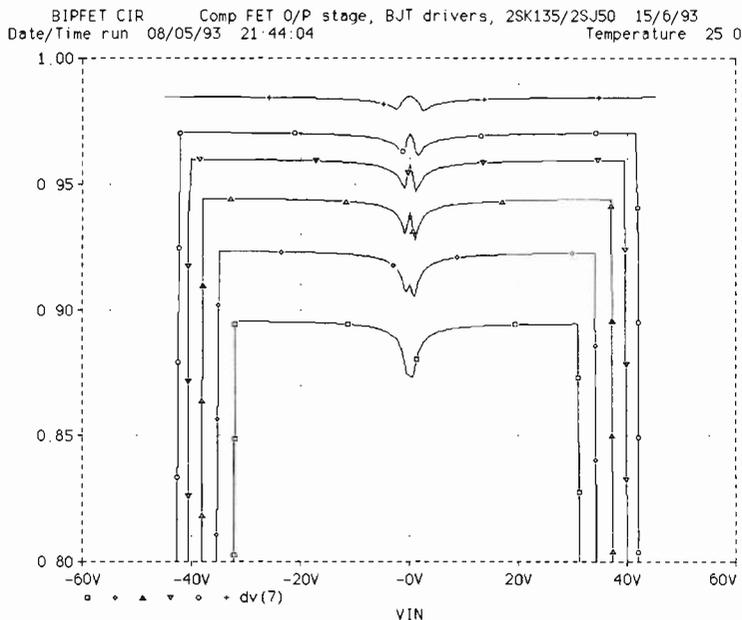


Fig. 14. Complementary bipolar fet gain vs output.

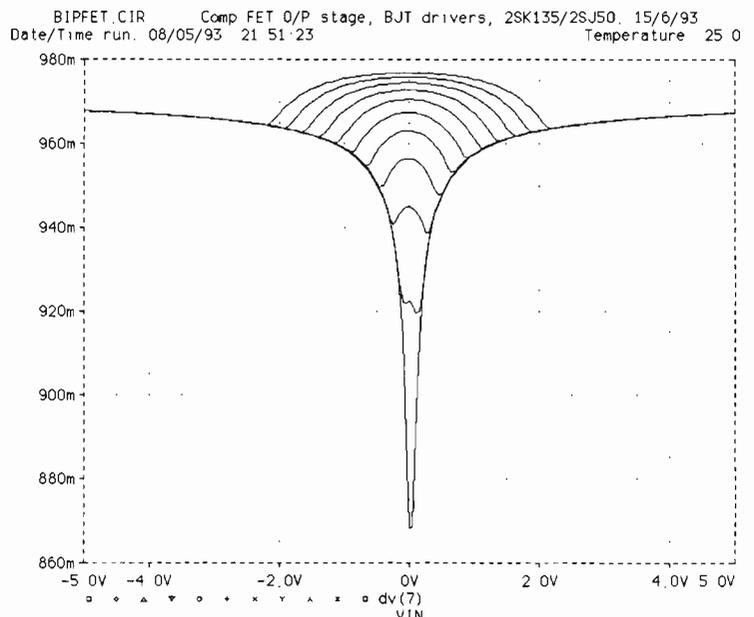


Fig. 15. Complementary BJT fet crossover region ±15V range.

lower upper half. Increasing the value of R_{e2} gives a reasonable match between the gains of the two halves, but leaves a daunting crossover discontinuity.

The hybrid full complementary stage in Fig. 5c was conceived¹⁰ to maximise performance by linearising the output devices with local feedback and reducing I_Q variations due to the low power dissipation of the bipolar drivers. It is highly linear, showing no gain droop at heavier loadings (Fig. 14) and promises freedom from switch-off distortion. But, as shown, it is rather inefficient in voltage swing. The crossover region (Fig. 15) still has some dubious sharp corners, but the total crossover gain deviation (0.96-0.97 at 8Ω) is much smaller than for the quasi hybrid (0.78-0.90) and so less high order harmonic energy is generated.

Next month: Controlling large signal non-linearity, crossover, and switch-off distortion.

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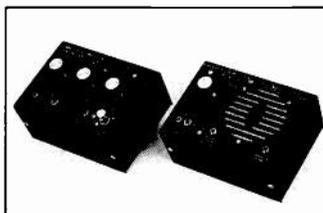
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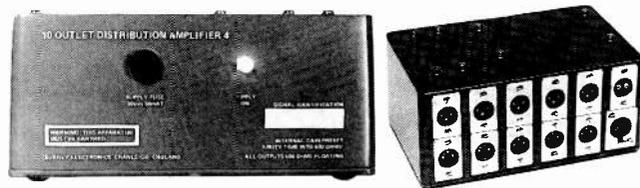


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George Pickworth traces the evolution of the rotary discharger, from simple hemispherical electrode to the complexity of Marconi's 300kW transmitter

The spark that gave RADIO to the world



The spark-electrode assembly, the discharger, was of vital importance to the spark transmitter, in the same way as the thermionic valve was to the valve-type transmitter.

Early dischargers were simply a pair of hemispherical electrodes housed in a sound deadening wooden box. But this design was superseded, in all but low power installations, by Wien's quenched gap – adopted almost universally by the Telefunken company – and by rotary type dischargers taken up by the Marconi company.

Evolution of the spark transmitter is essentially that of the discharger. In this article, we will concentrate mainly on the rotary discharger, of which there are two basic types: plain, using disc type electrodes to facilitate cooling and so reduce electrode erosion; and those employing radial or transverse electrodes whose role was to initiate and then quench a discharge after a given number of oscillations. These are referred to as either interrupters, or quenched-spark type dischargers.

As they were self cooling, rotary dischargers were inherently suited to high power installations. Their development stemmed from experimentation with different methods of

quenching the discharge, and from early hemispherical types to highly complex rotary designs – culminating in Marconi's 1914, 300kW synchronous discharger

Resonating selectivity

The term "spark transmitter" described what appeared to be a momentary electric discharge, and it became the generic name for this type of transmitter. The discharge was oscillatory with a duration corresponding to the train of oscillations.

Perhaps "arc" would have been a slightly more exact description, but this would have caused confusion with quenched-arc continuous-wave systems. Spark transmitters were generalised as wave train transmitters. Amplitude of the wave train declined steeply – especially with early systems.

This was the principal reason for developing continuous wave systems: "syntony", a term invented by radio pioneer Lodge to describe receiver selectivity by resonance, where oscillations progressively build up in amplitude in the receiver tuner.

With spark transmitters, resonance, and therefore syntony, is largely ineffective. One effect of quench-type dischargers was to

increase the number of significant waves in each train and so allow a small measure of syntony.

Even where syntony was not vital, for example in maritime systems which did not require a long range and where the frequency span was large, quenching still greatly improved transmitter efficiency.

Quest for long range

Early in the history of radio, before the relationship of wavelength and transmission distance was appreciated, long range was synonymous with high power.

Engineers knew that high effective power could best be obtained by storing energy and suddenly releasing it by spark discharge – the basic principle of spark transmitters.

Much of Marconi's early work in the quest for long range was with "untuned" systems, where energy was stored in the antenna and released by a spark discharge – virtually as an

The Caernarfon three turn, 2.0m diameter primary circuit inductor – a single layer helix of stranded insulated wire wound around a hollow 300mm tube. The three section antenna coupling coil was supported on a square section wooden beam.

electromagnetic pulse. The pulse, being aperiodic, has no frequency, so cannot be tuned (syntony was out of the question).

Nonetheless, an EM pulse can contain an enormous amount of energy: inducing a current pulse in the receiver antenna; triggering the coherer (a sensitive relay); activating a morse-inker (powered by its own battery), and so registering a dot for each pulse.

Grouped pulses were used for Morse signalling (Fig. 3).

Spark system principles

A capacitor suddenly discharging across a spark gap and through an inductor causes an oscillatory discharge at a frequency set by the values of the capacitor and inductor. (Fig. 1).

In early spark systems, DC pulses generated by an induction coil periodically charged the capacitor. But because of the short duration of the pulses, capacity value was limited to that which could be charged (typically to 15kV) within the pulse duration – only a fraction of the coil's open circuit potential.

Theoretically, an open circuit allows the potential across the secondary of an induction coil to rise to infinity when the primary circuit is interrupted. So, to avoid damaging the coil through a broken charging circuit, protection spark gaps were placed across the coil's secondary winding. Induction coils were graded by maximum safe open circuit spark length, for example 6in or 12in.

Transformer systems give a much longer duration of each negative- or positive-going AC half cycle than an induction coil pulse. Their longer charging period enables more energy to be stored, while still allowing the capacitor to charge to typically 15-30kV.

When the potential reaches the point where the air dielectric between the spark-gap breaks down – typically 4.0kV/1.0mm gap – ionisation dramatically reduces the resistance across the gap. Every discharge initiates a wave train, with the period between each train setting the tone heard with rectifier type receivers.

Induction coil-type transmitters had their discharge rate set by the interrupter – typically a few hundred Hz.

For the early transformer systems, employing fixed hemispherical electrodes, discharge repetition rates were set by the AC frequency (bearing in mind that there are two discharges per cycle). High frequency alternators, typically 500Hz, were generally used with transformer systems.

In rotary dischargers, drive speed and the number of electrodes set the discharge rate.

Shock excitation

Instead of discharging the capacitor directly through the tuned circuit, as with simple spark systems, it can be discharged through an additional coil with very low inductance and resistance, inductively coupled to the tuned circuit. Quickly quenching the discharge so that the capacitor discharges virtually a single unidirectional pulse, shocks

More energy from tuned systems

Much more energy could be stored in a capacitor proper than in an antenna. So later tuned systems discharged the capacitor through an inductor to generate a train of exponentially declining oscillations. Frequency was set by the value of the capacitor and inductor (see box).

All the radiated energy first had to be stored in the capacitor so it had a large value. But to tune the system to the maritime frequencies – 500kHz and 1.0MHz – the inductor value was low, typically a few turns of heavy gauge wire.

the tuned circuit into oscillation at its resonant frequency. This was the effect exploited by Prof Wien and adopted by the Telefunken system under the name "quenched spark", Fig. 1a.

The quenched-arc system had its roots in Elihu Thompson's 1892 "wave generator" which consisted of an arc lamp burning DC and shorted by an inductor and capacitor in series. The oscillator generated continuous oscillations, so the arc appeared to burn continuously – hence the name. But this too was quenched and re-ignited every rf half cycle (Fig. 2). Operation of the quenched arc is complicated and still not fully understood. But a simplified explanation is that the choking coils, and in some cases a variable resistor, limit the rate at which the capacitor can charge. Once the arc is struck, its resistance falls to a low level. Current is then drawn from the capacitor faster than it is replaced, and potential falls to a point where the arc is quenched.

Resistance across the gap then reverts to a high level and the capacitor re-charges. The cycle is repeated at a rate set by the resonant frequency of the inductor/capacitor. Potential at which the arc is quenched is set by the gap-width, but to ensure quenching, ionised gases have to be scavenged.

Spark or quenched arc?

As a general rule, early systems powered by high potential pulsed DC, generated by an induction coil, can be considered as spark systems.

Ordinary transformer-type spark transmitters with hemispherical electrodes also operated as true spark systems because the spark-gap was set so wide that discharge did not occur until the capacitor had charged to operational potential.

Marconi's interrupter-type rotary dischargers also made use of shock excitation. These were mechanical, so much slower than the Wien quenched gap, which was a magnetic device. Quenching seems to have occurred after a number of oscillations, depending on design of the interrupter and resonant frequency of the exciter circuit. However, subsequent oscillations declined very rapidly in amplitude after the initial pulse.

A similar shock effect would also have been significant with early spark transmitters employing loosely-coupled antenna circuits.

Marconi's transmitters, with their multi-spark interrupter rotary-dischargers, seem to have operated primarily in the shock

excitation mode. But they may well have operated in quenched-arc mode during each negative- or positive-going half AC cycle when, in effect, the source current is DC. Systems employing fixed gap dischargers and energised by smooth high potential DC invariably operated in the quenched arc mode.

At longer range, syntony was virtually nil. The amplitude of the waves declined too rapidly for resonance to be effective, though the coherer could still be used as the detector (Figs. 4 and 4a).

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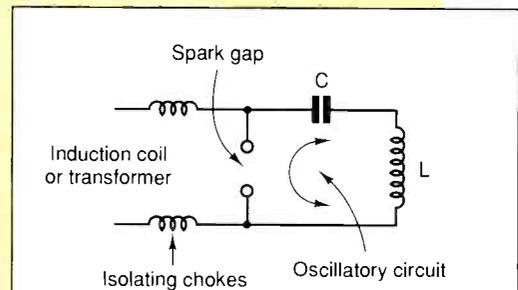


Fig. 1. Basic spark oscillator.

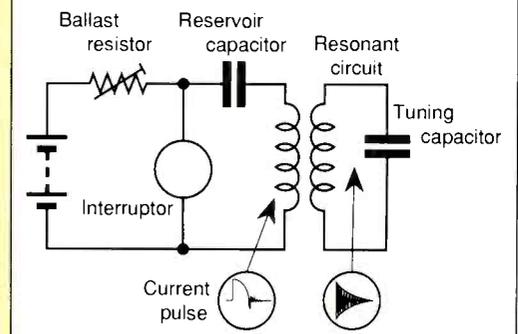


Fig. 1a. Wien's quenched spark adopted by Telefunken.

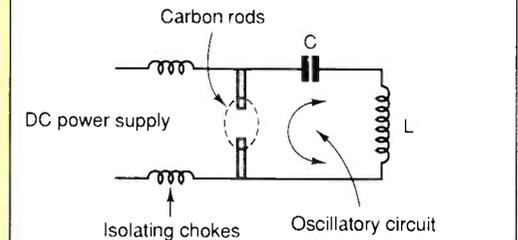


Fig. 2. Quenched arc oscillator generating continuous oscillations, with the arc quenched and reignited every rf half cycle.

Loose antenna coupling

Ionisation dramatically lowers the resistance across the spark-gap. But as the gap is in series with the capacitor and inductor, it dissipates appreciable energy. The logical way to minimise the loss is immediate transfer of the energy by EM induction to a tuned-secondary circuit, normally the antenna system.

The horizontal part of the antenna has capacitance with earth, or ship deck, and in conjunction with the antenna coupling coil, creates a resonant circuit. The antenna's low capacitance value meant the secondary winding normally had many more turns than the primary. But, with small ships where the length between the masts is limited, antenna capacity was increased by employing several parallel wires.

When the antenna circuit was precisely in tune with the primary circuit, the amplitude of the oscillations in the antenna circuit rose less steeply and declined more slowly than in the primary circuit, radiating pear-shaped wave trains. Although the amplitude of pear-shaped oscillation trains declined as a result of energy being radiated, the reduction in number of significant waves was much less pronounced than with exponentially declining wave trains, allowing a very limited degree of syntony (Fig. 5, 5a, 5b and 5c).

Beats occurred if the primary and secondary circuits were not tuned to precisely the same frequency (Fig. 5d) but this problem was avoided by shock excitation.

Quenching

Even a loosely-coupled antenna coil allowed some energy to be re-transferred to the primary circuit and dissipated by the spark-gap. The logical approach was to quench the discharge so that the spark-gap reverted to its high resistance state. By quickly quenching the discharge, the incidence of beat notes was greatly reduced.

In Wien's quenched gap, the discharge was probably quenched after less than six oscillations, so the secondary circuit was virtually shock excited. To all intents, this prevented generation of beats.

One way of quenching was to direct an air-blast across the gap to disperse the ionised gases. As energy was transferred to the secondary circuit, amplitude dropped to the point where the air-blast prevented re-ignition of the arc after being quenched at the end of an rf half cycle. But the method posed practical problems and was not widely used.

Prof Wien's simple and elegant approach, developed in 1906, was to use the spark's own magnetic field to drive the discharge radially to a wide part of the gap where it was automatically quenched. This "quenched-gap" was adopted almost universally by the Telefunken Company (Fig. 6a).

Essential feature of the Wien quenched gap was its pile of pairs of discs, each pair separated by mica washers to create a spark gap. The discs all had a circular groove close to the periphery, and their flat surfaces were parallel

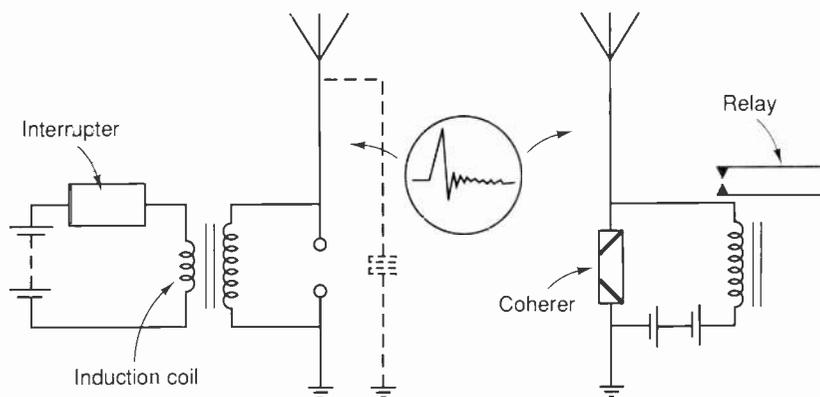


Fig. 3. Marconi's untuned system. Energy stored in the antenna was released by a spark discharge.

History replicated

The early radio pioneers produced "oscillograms" using a special camera focused on a fairly wide horizontal spark gap. During discharge, the photographic plate would drop rapidly so tracing the image on a moving plate.

To produce the oscillograms shown in this article, I constructed small scale reproductions of each transmitter. The oscillograms, recorded with scope and camera, are essentially the same as early graphic representations of wave forms produced by the early workers and confirm what they had intuitively deduced. The results also give an indication of the philosophies of those pioneers and their approach to problem solving.

GP

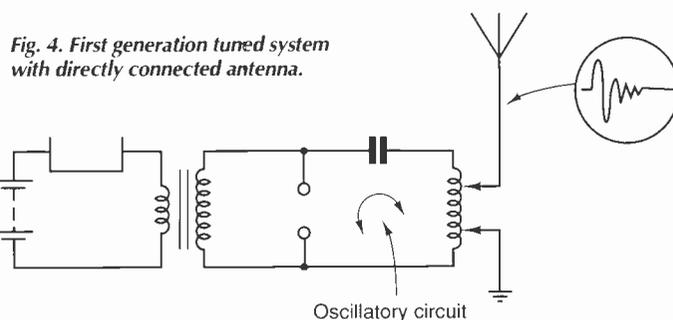


Fig. 4. First generation tuned system with directly connected antenna.

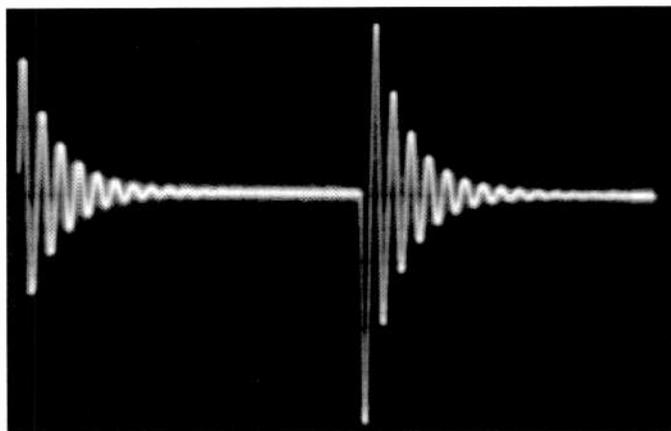


Fig. 4a. Wave train radiated by early maritime spark transmitter. Directly coupled antenna heavily damps the oscillations.

and silver plated to prevent formation of pimples that would short the discs (Fig 6).

During discharge, the magnetic field that developed around the axis of the pile drove the sparks radially to the groove where gap-width was greater and the discharge was extinguished. The number of pairs of discs in cascade depended on power requirements: for 12.5kV this was typically 14.

The capacitor was discharged virtually as a single unidirectional pulse, shocking the tuned circuit into oscillation at its resonant frequency.

Marcconi experimented with a similar device but rejected it in favour of his rotary-type

quenched-discharger where the gaps widened after each discharge.

Interrupter-type rotary dischargers

Early Marconi synchronous rotary dischargers consisted of a hub with four radial electrodes which rotated between a pair of fixed rod-type electrodes. The hub was driven by a shaft extending from its complementary alternator so the two rotated synchronously – hence the name.

Discharge commenced while the gaps were still narrowing and potential of a half cycle was increasing. Further narrowing as the electrodes rotated, reduced resistance across the

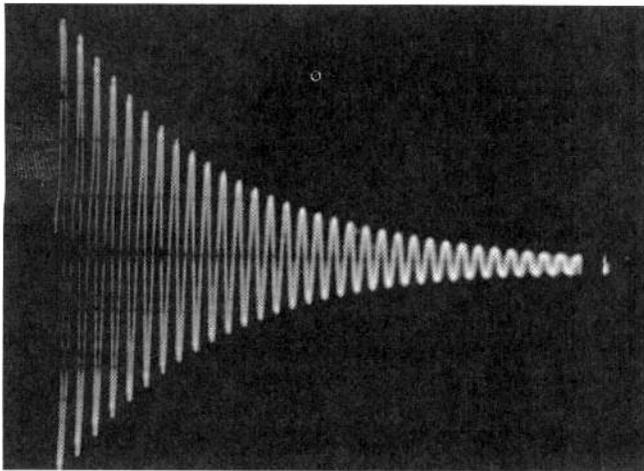


Fig. 5a. Oscillation train in a primary circuit. Antenna (secondary circuit) is loosely coupled.

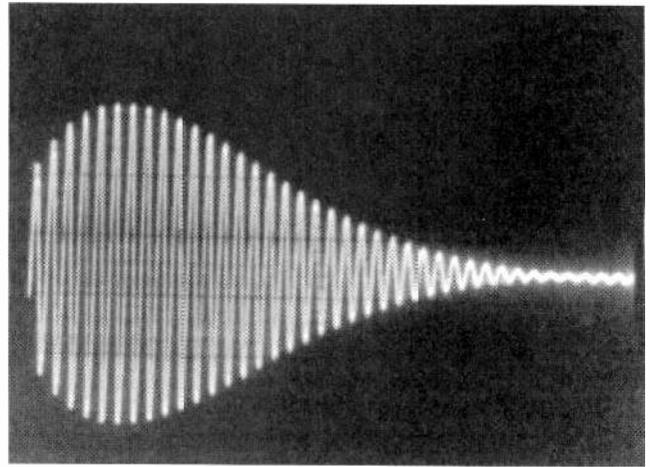


Fig. 5b. Oscillation train in a loosely coupled antenna circuit.

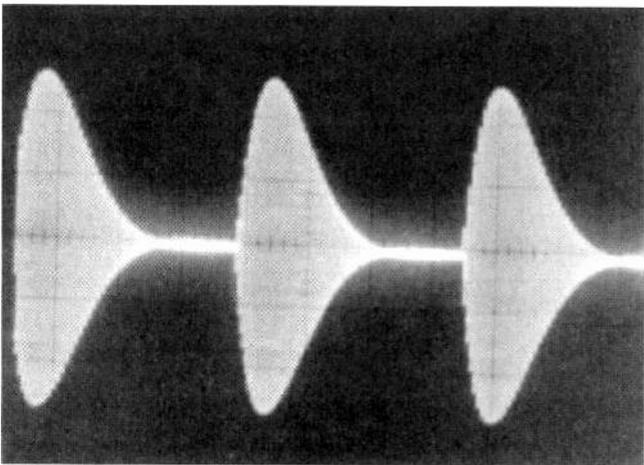


Fig. 5c. Wave trains radiated by a loosely coupled antenna.

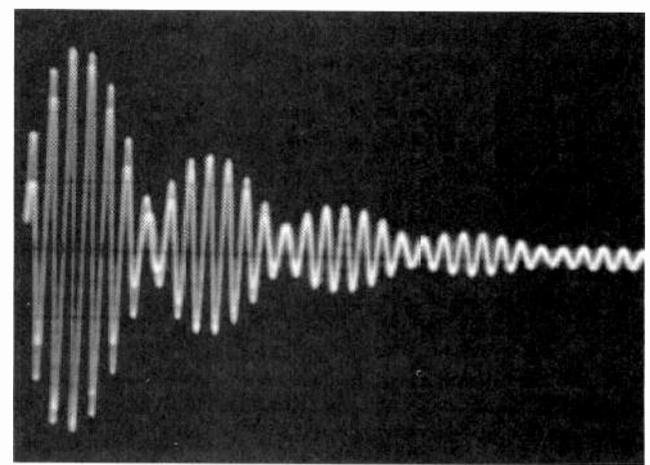
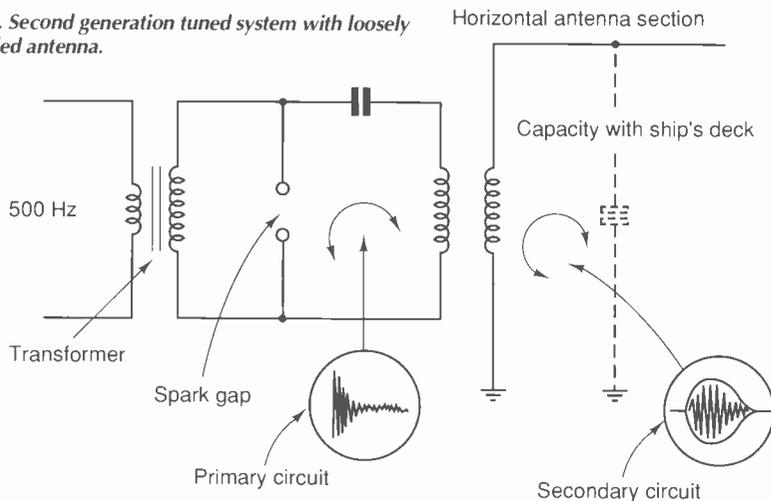


Fig. 5d. Oscillations in secondary circuit when not tuned to precise frequency of primary circuit.

Fig. 5. Second generation tuned system with loosely coupled antenna.



gaps as the amplitude of the oscillations declined, and while energy was being transferred to the secondary circuit. The gaps then widened and the draught created by the rotating electrodes dispersed ionised gases, quenching the discharge and returning the gaps to a high resistance state (fig. 7b).

Rotary dischargers originally had four radial electrodes – corresponding to the alterna-

tor's four pole pieces, – aligning four times during each revolution. Rotation speed was typically 25rev/sec (1500rev/min). Each revolution generated two cycles, so there were four discharges/rev, 100 discharges/sec. But the corresponding tone produced by rectifier receivers was too low to be heard through "static" interference. (Fig. 7b)

Multiple-spark dischargers

To raise the pitch of the tone, Marconi increased the number of radial electrodes to 16 or 32, depending on the model. But the rugged four pole piece alternator was retained so that instead of one discharge per ac half cycle, there were now four or eight discrete discharges (Fig. 8a).

Furthermore, the magnitude of successive discharges varied according to the point on each AC half cycle where discharge occurred (Fig. 8b).

Operation was primarily in the shock excitation mode. But it could also have been in the quenched arc mode because during each negative- or positive-going ac half-cycle, the power supply was in effect dc. The 16 radial-electrode model gave 16 discharges/rev – at 25rev/sec 400 discharges/sec – while the 32 electrode model gave 800 discharges/sec, producing a pleasant tone with rectifier type receivers. But because energy available from each AC half cycle was distributed among several wave trains, effective power was less than if the energy had been radiated as a single wave train.

Maritime systems had the alternator driven by a DC motor running off the ship's DC supply. Terrestrial installations used a petrol

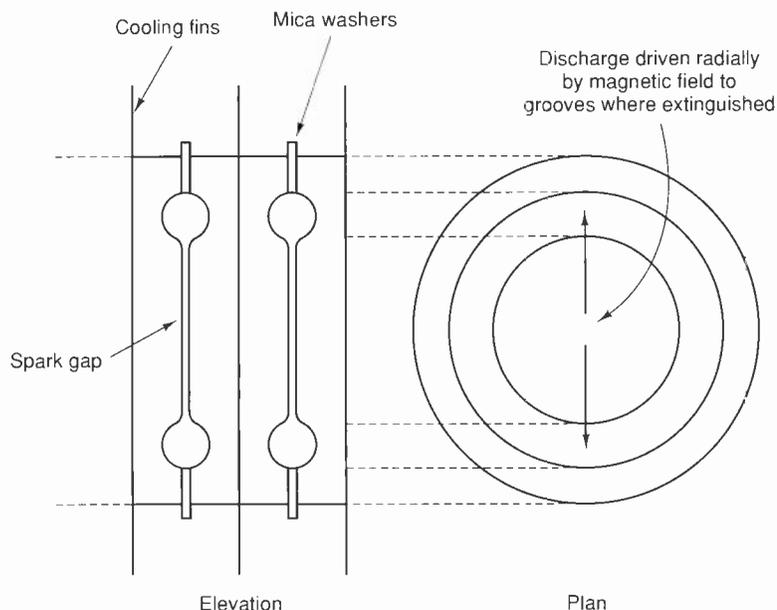


Fig. 6. Quenched spark system – Wien's quenched gap – showing two spark-gap sections.

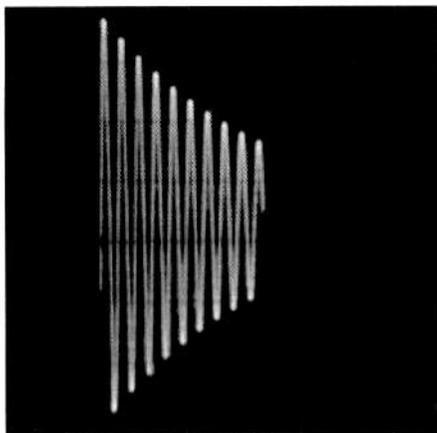


Fig. 6a. Oscillation train in primary circuit with a quenching type discharger.

Fig. 7a. below. Quenched spark system schematic with its rotating electrodes.

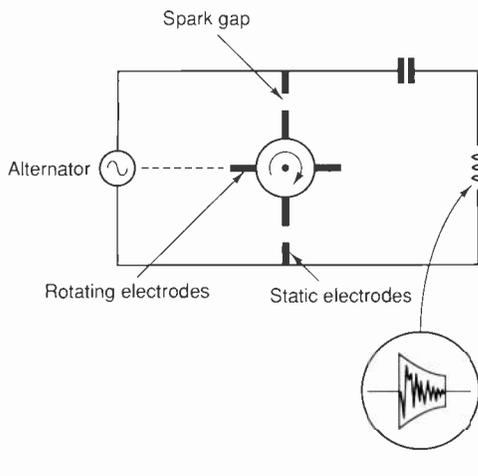
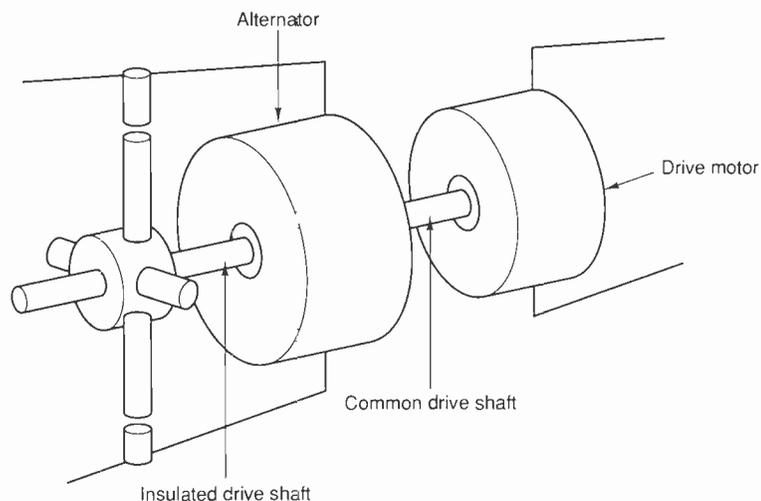


Fig. 7b. below right. Rotating electrodes.



engine, or mains power and a transformer – in which case the drive motor was synchronised with the mains supply.

The electrodes rotated in synchrony with the alternator, though this was not essential and there was no noticeable difference in performance with transformer systems and an asynchronous drive motor. Indeed, the multiple discharger had its roots in an asynchronous version developed by Tesla for use with his "Magnifying Transmitter".

Only a limited degree of sympathy was possible, but this was of no great significance with early maritime systems. More important was that the discharger was efficient and, in conjunction with thermionic valve regenerative receivers, remained in service with maritime systems until the 1930s when interference caused to other radio users could no longer be tolerated.

Marconi-Mackie disc discharger

A scaled up version of the smaller maritime multiple-spark discharger, capable of han-

Detectors

The gradual rise in amplitude of the wave train greatly reduced spark jamming. But because the pulse effect was dramatically reduced, the coherer was almost useless as a detector.

Fessenden, another giant in radio history, developed a liquid detector which was also a relay type. With a battery and with rf current present, it produced a roaring sound in headphones. Massie's poor contact or microphonic detector was also of the relay type and produced a similar sound to the liquid detector.

Marconi's famous magnetic detector also produced a sound in headphones. Being a rugged magnetic/mechanical device it was far better suited to shipboard use than the liquid or microphonic detector, though it was less sensitive.

Rectifier-type detectors, including the silicon detector and the audio valve, were experimented with by other pioneers. But unlike relay types and the magnetic detector, rectifier detectors produced a sound with a pitch corresponding to the wave train repetition rate. Pear-shaped wave trains produced an almost musical note, as each wave train gently caused the earpiece diaphragm to vibrate in sympathy with the wave train repetition rate. But the first wave of exponential wave trains virtually "hammered" the diaphragm into vibration.

With all rectifier-type detectors, and the magnetic detector energy to produce the sound was derived from the em waves and this limited sensitivity; this in turn brought about the development of highly sensitive headphones.

dling considerable power, is the Marconi-Mackie discharger. Largest example seems to have been 60kW machines employed by maritime relay stations.

The assembly was installed in a brick "silence-cabin" with the drive shaft and flexible coupling extending through the wall to the alternator.

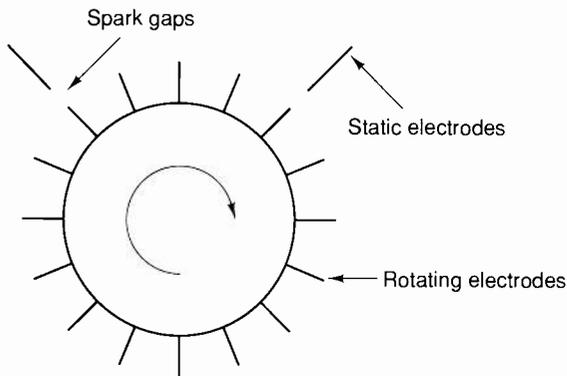


Fig. 8a. Marconi's multiple spark discharger.

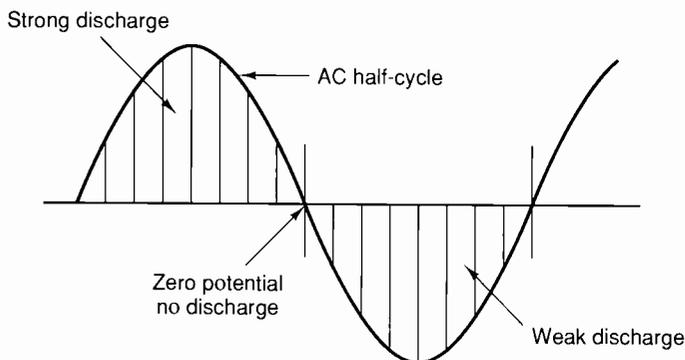


Fig. 8b. Relationship of discharges to AC half-cycles – synchronised version.

Rod-type static electrodes were eventually superseded by rotating side discs 300mm diameter and 30mm thick. The main disc, or hub, was of solid copper 900mm diameter and 30mm thick with 24 radial electrodes each about 200mm long. At its normal speed of 25rev/sec the 24 radial electrodes gave 600 discharges/sec.

Caernarfon synchronous discharger

Numerous operator manuals were produced to explain the workings of maritime radio systems. But very little meaningful technical data were published on the super stations. They were one-off installations, and because of their strategic importance during the World War I and the fact that they were used to signal to

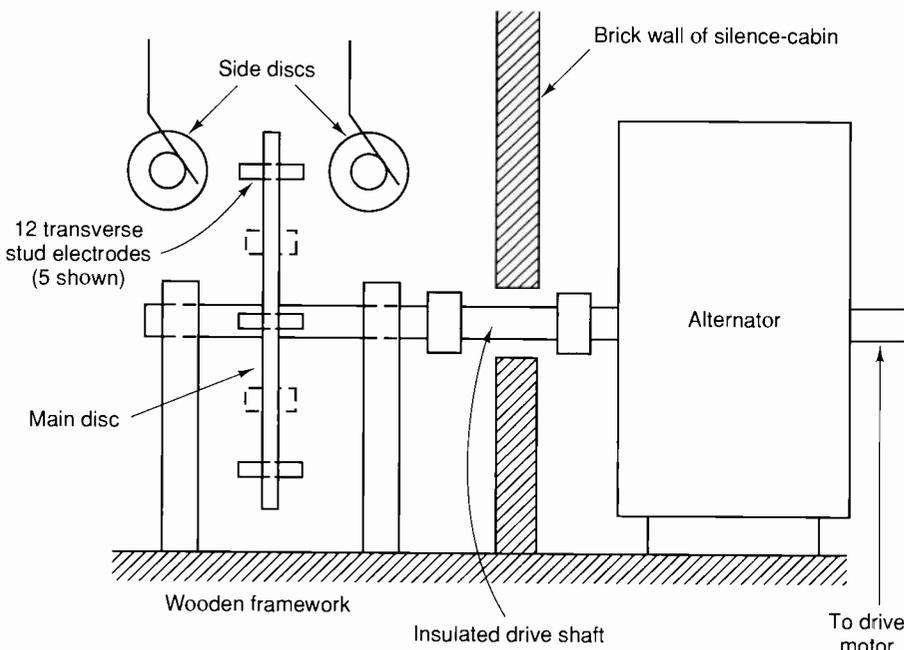


Fig. 9. Caernarfon synchronised discharger.

submerged submarines during that war, they were shrouded in secrecy.

So, to fill gaps in known specifications, some assumptions have to be made.

The 300kW Caernarfon 21.4kHz transmitter had a discharger that evolved from of the early maritime-synchronous-discharger. But instead of four radial electrodes, 12 copper stud electrodes, 10mm long, extended transversely from both sides of the periphery of a 900mm diameter steel disc. These aligned with a pair of slowly rotating copper side discs, each 300mm diameter.

The steel disc was coupled to the alternator by a non-conductive shaft extending through the brick wall of the silence cabin, which housed the machine, to the 300kW single phase alternator driven by a three-phase motor drawing power from high voltage mains (Fig. 9)

Contemporaneous notes state that the alternator – and so the discharger – ran at 25rev/s (1500rev/min) and that AC frequency was 150Hz. But they give little further technical information. The assumption is that the alternator had 12 pole pieces and the discharger had 12 electrodes, giving 12 discharges/rev or 300 discharges/s. To pack maximum energy into each wave train, there was one discharge per ac half cycle.

Wave trains would have been pear-shaped and contain about 25 waves of significant amplitude. Each wave had a period of 46µs, so the duration of the wave train was 1.175ms. Period between each wave train was 3.3ms. Keying was by breaking the capacitor charging circuit, and the arc which developed when the circuit was broken, was quenched by an air blast at 2.0 psi.

The three turn, 2.0m diameter primary circuit inductor consisted of a single layer helix of stranded insulated wire wound around a hollow tube 300mm in diameter. The multi-turn, three section antenna coupling coil of stranded insulated wire was supported on a square section wooden beam that allowed the coil to slide axially inside the primary coil, and thereby adjust the degree of coupling.

The antenna was an inverted L and comprised 10 parallel wires each 1300m long supported by 10 tubular steel masts 130m high over an earth screen (9.0m above the ground. Antenna resistance was given as 0.62Ω and maximum antenna current as 220A.

Efficiency from alternator to antenna was given as 30% in one document – but 60% in another.

Two dischargers were completed in 1914, one of which seems to have been a stand-by. Although perfectly satisfactory in service, syntony was poor and in 1916 they were replaced by the timed-spark transmitter which radiated continuous but undulating waves. ■

Marconi continuous-wave transmitter will be covered in a future article.

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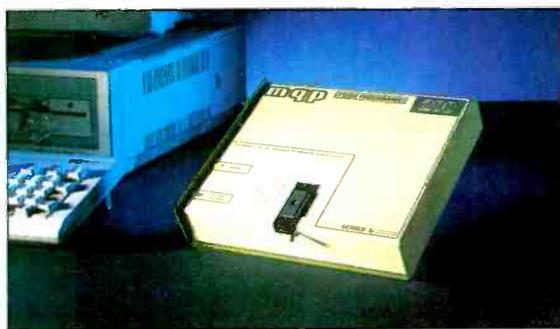
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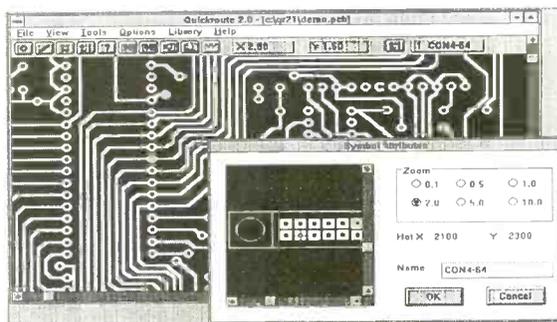
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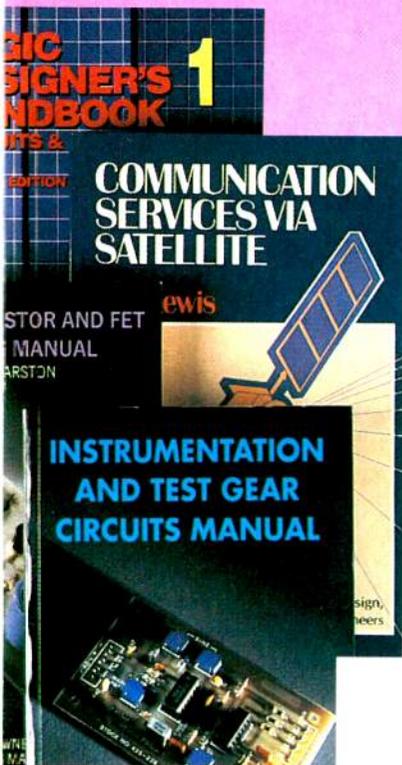
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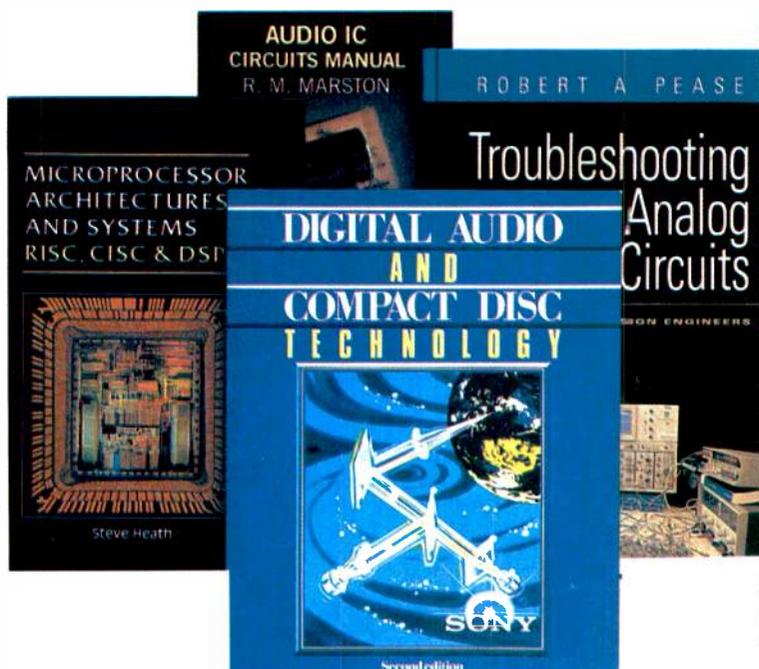
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One of the most useful means of specifying a linear device is to use scattering parameters, commonly referred to as *S*-parameters, in reality voltage reflection and transmission coefficients when the device is embedded into a 50Ω system.

Magnitude of the input reflection coefficient $|S_{11}|$ is directly related to input VSWR by the equation

$$\text{VSWR} = (1 + |S_{11}|)/(1 - |S_{11}|)$$

Likewise magnitude of the output reflection coefficient $|S_{22}|$ is directly related to output VSWR (Fig. 1). Square of the magnitude of the input-to-output transfer function $|S_{21}|^2$ is also the power gain of the device, referred to on data sheets as "insertion gain". Note that $|S_{21}|^2$ is the power gain of the device when the source and load impedances are 50Ω. An improvement in gain can always be achieved by matching device input and output impedances – almost always not 50Ω – to

50Ω through matching networks. The larger the linear device, the lower the impedances and the greater is the need to use matching networks to achieve useful gain.

Another gain specification shown on low power data sheets is "associated gain", with the symbol G_{NF} . It is simply the gain of the device when matched for minimum noise figure. Yet another term is "maximum unilateral gain", G_{Umax} . As expected, G_{Umax} is the gain achievable by the transistor when the input and output are conjugately matched for maximum power transfer (and $S_{12} = 0$). G_{Umax} can be derived using scattering parameters:

$$G_{Umax} = |S_{21}|^2 / \{(1 - |S_{11}|^2)(1 - |S_{22}|^2)\}$$

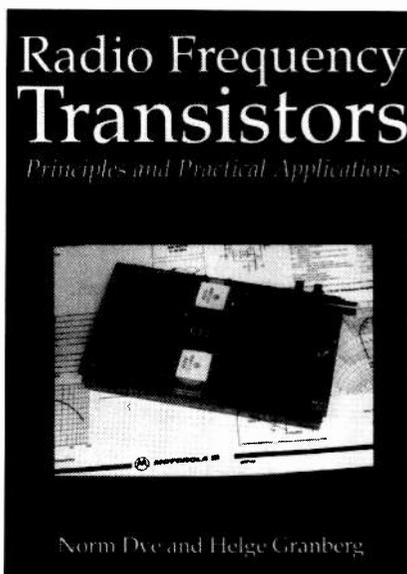
Simply stated, this is the 50Ω gain increased by a factor which represents matching the input, and increased again by a factor representing matching the output.

RF low power transistors are often used as low noise amplifiers, leading to several transistor data sheet parameters related to noise figures. NF_{min} is defined as the minimum noise figure that can be achieved with the transistor. Achieving this NF requires source impedance matching which is usually different from that required to achieve maximum gain. So design of a low noise amplifier is always a compromise between gain and NF .

A useful tool in aiding this compromise is a Smith Chart plot of constant gain and noise figure contours, drawn for specific operating conditions – typically bias and frequency (Fig. 2). These contours are circles, either totally or partially complete within the confines of the Smith Chart.

If the gain circles are contained entirely within the Smith Chart, then the device is unconditionally stable: if portions are outside, then the device is "conditionally stable", and the device designer must investigate instabilities, particularly outside the normal frequency range of operation.

Where the data sheet includes noise parameters, a value will be given for the optimum



Intercept point

The intercept point is a useful concept in that it allows the value of distortion for any signal level to be determined.

Second order distortion products can be shown mathematically to have amplitudes that are directly proportional to the square of the input signal level. Third order distortion products have amplitudes that are proportional to the cube of the input signal level. The conclusion is that a plot of each response on a log-log scale (or dB/dB scale) will be a straight line with a slope corresponding to the order of the response.

Fundamental responses will have a slope of 1, the second order responses will have a slope of 2, and the third order responses a slope of 3. Note that the difference between fundamental and second order is a slope of 1, and between fundamental and third order is a slope of 2. So for second order distortion, a 1dB change in signal level results in a 1dB change in second order distortion, but a 2dB change in third order distortion. Using the curves shown in the figure, if the output level is 0dBm, second order distortion is at -30dBc and third order distortion is at -60dBc. Change the output level to -10dBm and the second order distortion should improve to -40dBc (-50dBm) but third order distortion will improve to 0dBc (-90dBm).

Thus, a 10dB decrease in signal has improved second order distortion by 10dB, and third order distortion by 20dB.

The intercept point is defined as the point on the plot of fundamental response and second (or third) order response where the two straight lines intercept each other. It is also that value

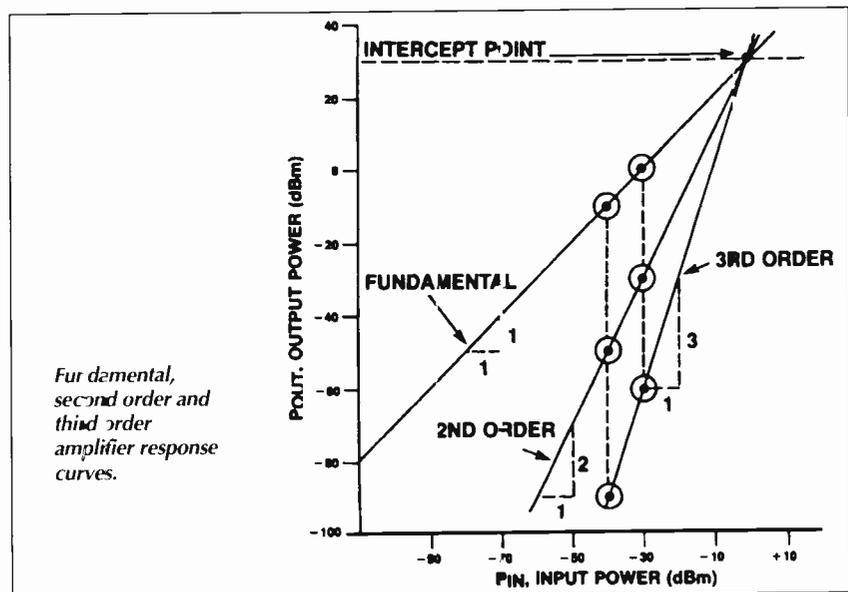
of signal (hypothetical) at which the level of distortion would equal the initial signal level.

For example, if at the point of measurement, second order distortion is -40dBc and the signal level is -10dBm, then the second order intercept point is 40dB above -10dBm or +30dBm. In the figure, +30dBm is the value of output signal at which the fundamental and second order response lines cross. The beauty of the intercept point is that once it is known, the value of distortion for any signal level can be determined, provided it is in a region of operation governed by the mathematical relationships stated - typically IMDs greater than 60dB below the carrier.

Similarly, to determine third order intercept point, measure third order distortion at a known signal level. Then,

take half the value of the distortion (expressed in dBc) and add to the signal level. For example, if the signal level is +10dBm and the third order distortion is -40dBc, the third order intercept point is the same as the second order intercept point or 10dBm + 20dB = 30dBm. Both second order and third order intercept points are illustrated.

In general, the intercept points for second and third order distortion will be different because the non-linearities that create second order distortion are usually different from those that create third order distortion. However, the concept of the intercept point is still valid: the slopes of the responses are still 1, 2, and 3 respectively. All that needs to be done is to specify a second order intercept point different from the third order intercept point.



input reflection coefficient to achieve minimum noise figure, Γ_0 or sometimes Γ_{opt} . But remember, matching this value of input reflection coefficient means far less gain is likely to be available than is achievable by the transistor.

Input reflection coefficient for maximum gain is normally called Γ_{MS} while the output reflection coefficient for maximum gain is normally called Γ_{ML} .

Another important noise parameter is noise resistance, R_n , expressed in ohms. Sometimes in tabular form, this value may be normalised to 50Ω, in which case it is designated r_n . The significance of r_n can be seen in the formula:

$$NF = NF_{min} + \{4r_n|\Gamma_S - \Gamma_0|^2\} / \{(1 - |\Gamma_S|^2)(1 + |\Gamma_0|^2)\}$$

which determines noise figure NF of a trans-

istor for any source reflection coefficient Γ_S if the three noise parameters NF_{min} , r_n and Γ_0 (the source resistance for minimum noise figure) are known (Table 1).

The locus of points for a given NF turns out to be a circle - the NF_{min} circle being a point. So by choosing different values of NF , a series of noise circles can be plotted on the chart.

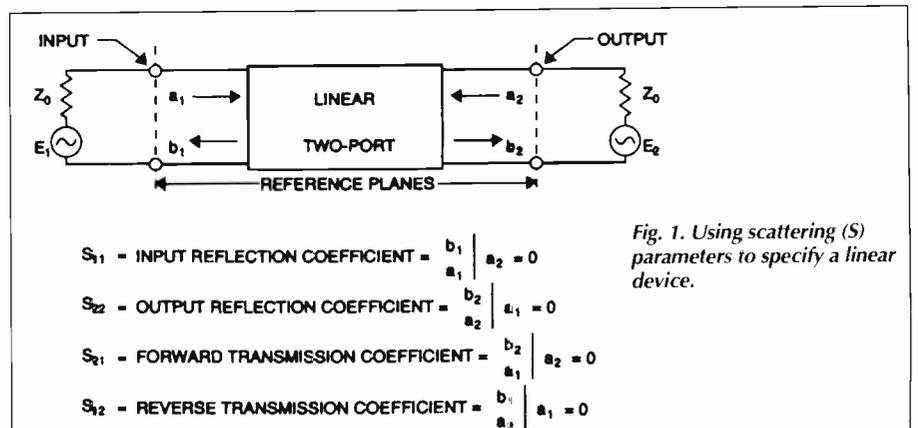


Fig. 1. Using scattering (S) parameters to specify a linear device.

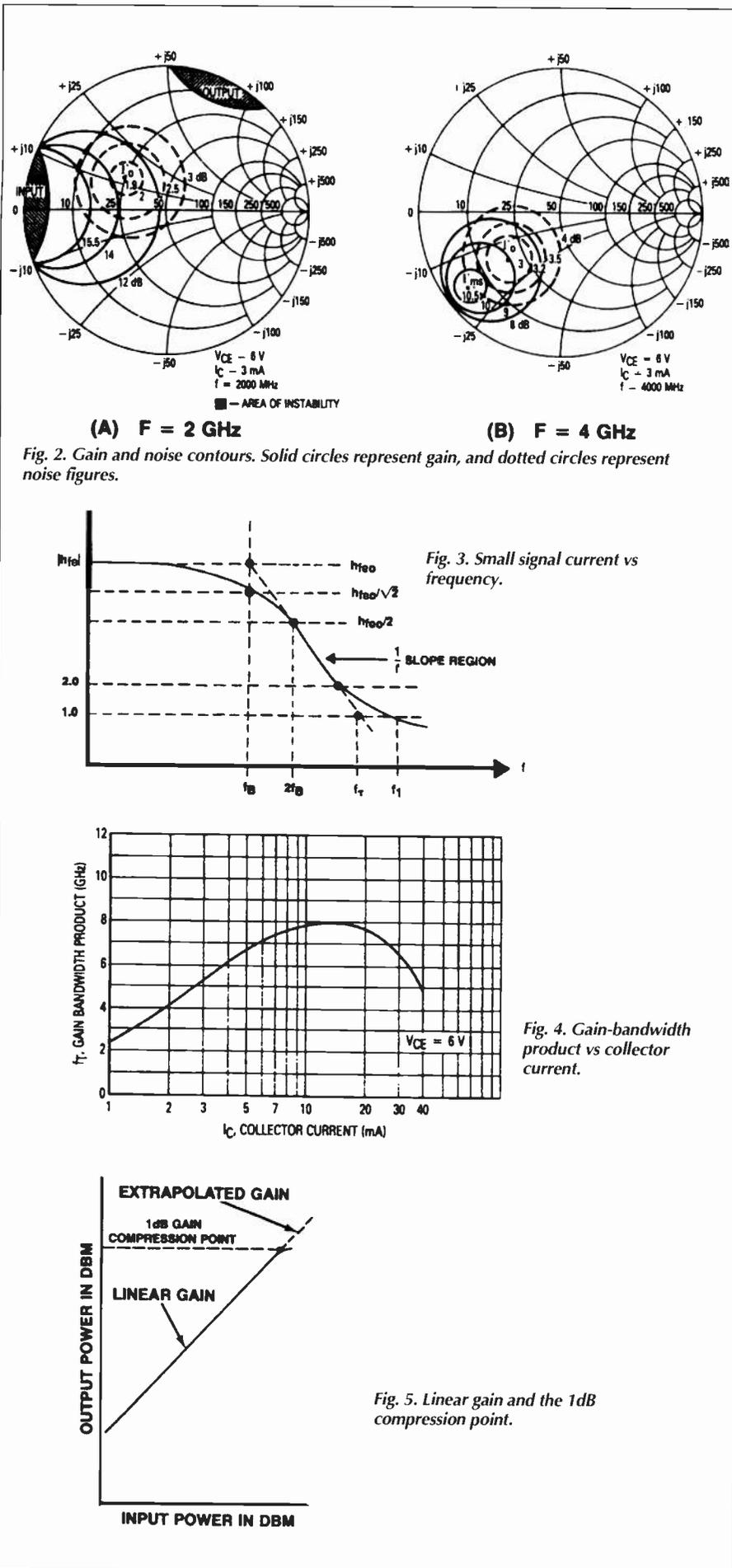


Table 1. Typical noise parameters for the MRF942 transistor.

V_{CE} (V_{dc})	I_C (mA)	f (MHz)	N_{Fmin} (dB)	G_{NF} (dB)	Γ_o (mag, ang)	R_N (Ω)	$N_{F50\Omega}$ dB
6	3	1000	1.3	16	$0.36 \angle 94$	17.5	1.7
		2000	2.0	11	$0.37 \angle -145$	15.5	2.6
		4000	2.9	8.0	$0.50 \angle 134$	21.5	4.3
15	3	1000	2.1	19	$0.25 \angle 150$	13	2.6
		2000	2.7	14	$0.26 \angle 173$	16.5	3.1
		4000	4.3	9.0	$0.48 \angle -96$	47	5.4

r_n can be measured by measuring noise figure for $\Gamma_s = 0$ and applying the above equation.

A parameter found on most rf low power data sheets is commonly called the current gain-bandwidth product, f_T . Sometimes it is referred to as the cut-off frequency, because it is generally thought to be the product of low frequency current gain and the frequency at which the current gain becomes unity. While this is not precisely true (Fig. 3), it is close enough for practical purposes. It is true that f_T is an excellent figure-of-merit, becoming useful in comparing devices for gain and noise figure capability.

High values of f_T are normally required to achieve higher gain at higher frequencies – other factors being equal.

To the device designer, high f_T specs mean decreased spacings between emitter and base diffusions, and shallower diffusions – factors that are more difficult to achieve in making an rf transistor.

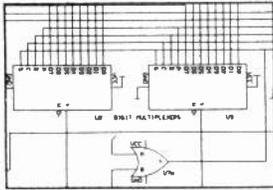
The complete rf low power transistor data sheet will include a plot of f_T versus collector current. Such a curve (Fig. 4) will increase with current, will flatten, and then begin to decrease as I_C increases thereby revealing useful information about the optimum current with which to achieve maximum device gain.

Another group of characteristics associated with linear (or class A) transistors has to do with the degree to which the device is linear. Most common are terms such as “ P_{1dB} , 1dB gain compression point” and “third order intercept point” (ITO). P_{1dB} , 1dB gain compression point is simply the output power at which the input power has a gain associated with it that is 1dB less than the low power gain. In other words, the device is beginning to go into saturation, where increases in input power fail to realise comparable increases in output power (Fig. 5).

Importance of the 1dB gain compression point is that it is generally accepted as the limit of non-linearity tolerable in a “linear” amplifier and points to the dynamic range of the low power amplifier. On the low end of dynamic range is the limit imposed by noise; on the high end is the limit imposed by “gain compression.”

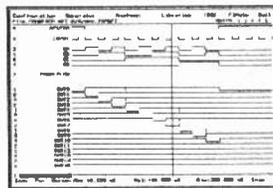
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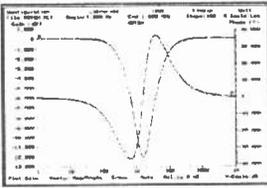


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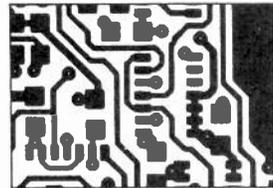
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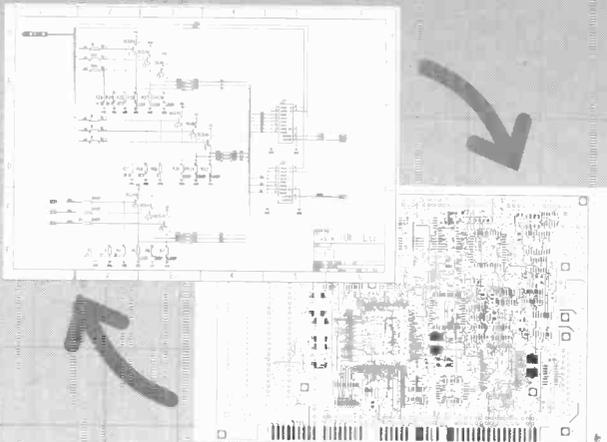
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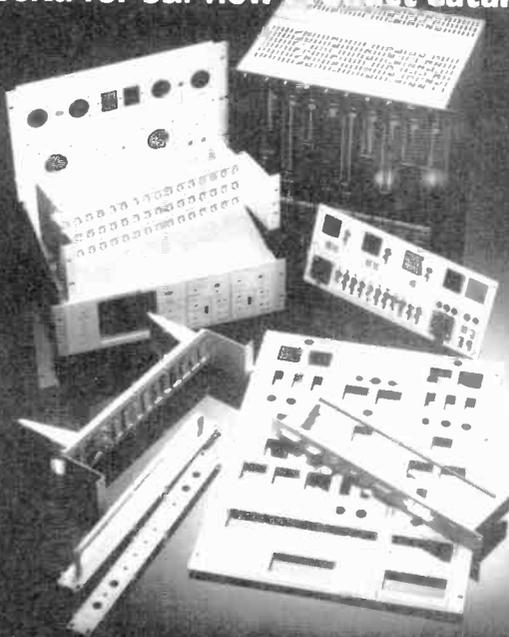
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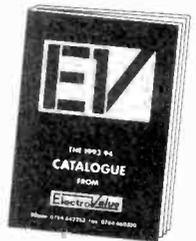
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Asics

8000-gate PLD. *EPF8820* is a member of Altera's *Flex8000* family of high-density programmable logic devices in 0.8 micron technology. *EPF8820* contains 820 registers and has 152 user i/o pins in 208-pin quad flat pack, 192-pin PGA and 192-pin ball grid array packs. Average benchmark speed is 40MHz Altera Ltd, 0628 488811.

A-to-D & D-to-A converters

Codec. In Mitel's *MT9125* adaptive differential pulse-code modulator, two 64kb/s PCM channels are compressed into two 32kb/s ADPCM channels and ADPCM to PCM. The ADPCM algorithm conforming to CCITT G.721 and ANSI T1.303-1989. It also supports a 24kb/s (3-bit word) algorithm CCITT G.723. The device needs only 50mW for dual-channel working. Mitel Semiconductor, 0291 430000.

Discrete active devices

Automotive fets. Low-voltage power mosfets from IR are particularly suited to the car electronics market. The 60V *IRFP064* has an on resistance of 9m Ω , and a current rating of 70A, while the *IRFZ48* handles 50A with an on resistance of 18m Ω . Two further devices, *IRFZ46* and *IRFP048*, are meant for use in traction and transmission control, UPSs, DC-DC converters and motor control. International Rectifier, 0883 714234.

SM, p-channel mosfet. New in the Supertex family of p-channel enhancement-mode mosfets with gate thresholds of 1V maximum is the *LP0701LG*, a surface-mounted version of the TO92 *LP0701N3* and *LP0701ND* die. Drain-source breakdown is 16.5V and on resistance is better than 1.5 Ω at 5V and 300mA. Kudos Thame Ltd, 0734 351010.

Adjustable zener. Two external resistors program the Zetex *ZR431* surface-mounted shunt regulator in the range 2.5-20V. No-load current

consumption is 35 μ A and the device handles 2W maximum. Maximum output current is 100mA and temperature stability is 50ppm/ $^{\circ}$ C. Zetex plc, 061-627 5105.

Digital signal processor

16-bit 52Msample/s digital filter. Harris's *HSP43216* is a digital half-band filter with 16-bit precision. It incorporates a 67-tap filter processing 16-bit data with 20-bit coefficients and gives better than 90dB stop-band attenuation and 0.0005dB pass-band ripple, with 1.24-1 filter shape factor. Decimation or interpolation by 2 and quadrature up or down conversion are on-chip, as is coefficient generation. Harris Semiconductor (UK), 0276 686886.

16-bit DSP. TI's newest addition to its *TEC320C* DSP family is the *TEC320C52*, which has a 16-bit, fixed-point DSP and an array from the 0.8micron *TGC1000* series to provide 10K usable gates, 3V or 5V operation and 25ns or 35ns processing speed. The core executes up to 40Mips, has 4Kword of rom and 1Kword of dual-access ram. Texas Instruments, 0234 223252.

Linear integrated circuits

Low-noise amplifier. With a noise figure of 2.5dB, bandwidth of 1.8-3GHz, gain of 17dB and 10dBm output power, Harris's *HMP-130203* is a microwave monolithic gallium arsenide cascaded amplifier, designed for the 1.8/1.9 and 2.4GHz industrial, scientific and medical bands. Anglia Microwaves Ltd, 0277 630000.

Digital volume control. With a dynamic range of 110dB and THD of better than 0.001%, Crystal's *CS3310* stereo digital volume control is claimed to be the industry's best performer and is on a single chip, needing no extras. Zero crossing volume changes avoid noisy transitions and the log-law characteristic gives good control at low levels. The three-wire interface controls two independent channels and allows daisy-chaining of multiple units. Crystal Semiconductor Corp., (512) 445 7222 (US).

Pager receiver. Sensitivity of -130dBm is claimed for GEC Plessey's *SL6609* pager receiver IC, which offers better integration, lower power consumption and smaller outline than its predecessor, the

SL6649-1. *SL6609* uses direct conversion and needs no external ceramic filters, operating up to 350MHz at 512, 1200 or 2400baud with 70dB adjacent-channel rejection. A 1V, 5mA regulator is included. GEC Plessey Semiconductors, 0793 518582

One-chip car radio. ITT's car audio processor (CAP) integrates stereo decoder, baseband processing, fast-tuning synthesiser, AM tuning for 455kHz or 10.7MHz IF, AM IF processing and AM stereo demodulation. The CAP will handle both analogue and digital audio input and has a programmable digital audio interface. Flexibility is conferred by the DSP core which, with standard software, provides a complete solution but which can be easily adapted to requirements. S:N ratio is 85dB and THD less than 0.01%. ITT Semiconductors, 0932 336116.

SVGA video ICs. National's *LM1205* 130MHz RGB video preamplifier, together with the *LM2419* CRT driver, forms a complete video channel in 1024 by 768 SVGA non-interlaced monitors, including VESA formats. Video adjustment is DC to allow digital drive from a microcontroller to adjust contrast, channel gain for colour temperature setting and cut-off. *LM2419* amplifies the 4V *LM1205* output to 50V with a bandwidth of 65MHz. National Semiconductor, 0793 697415.

Switched-mode controller. Siliconix has designed its *Si9114* switched-mode control chip for use in communications equipment. It has the elements of the standard *Si9100* series and increased functions and performance, including a 500kHz switching frequency. Soft start, internal start-up and latched shut-down are included and, using telephone-line voltage, flyback or forward converters can be built to switch at up to 1MHz. Siliconix/Temic, 0344 485757.

Charge pump/dual op-amp. Replacing two separate devices in less board space, TI's *TLE26E2* consists of a negative-rail charge pump and a pair of op-amps in one 16-pin wide-body SM package. The inverting charge pump supplies 100mA, used to power the op-amp negative rail and other sub systems, and each op-amp gives over 25mA with micropower requirements of less than 600 μ A. When the negative rail is not needed, it can be shut down. Texas Instruments, 0234 223252.



Fast cmos. The first 20 members of Toshiba's 150MHz cmos logic family are now being sampled. Propagation delay is 3.5ns and switching noise 0.8V, and the devices will interface between 3V and 5V systems, by means of a signal-voltage shift. Output drive is 8mA. Toshiba Electronics (UK) Ltd, 0276 694500.

Logic building blocks

Disk-drive chip. Allegro's *A6980CJT* Superservo provides drive, management and control of the voice coil and spindle-motor power actuation subsystems for hard-disk drives. Internal circuitry gives start-up and microcomputer-assisted run modes with no need for external components; all current-sensing and diode protection is internal. Allegro MicroSystems, 0932 253355.

Video clock synthesiser. Amega's *AMCC S4503* cmos video clock synthesiser is designed for clock distribution, providing multiple clock frequencies up to 300MHz from one crystal, producing independently selected outputs in the 10-300MHz range. A pair of positive-referenced ECL outputs with only 250ps jitter operate to 300MHz and a second pair provide complementary 24mA outputs at up to 80MHz. Amega Electronics, 0256 843166.

Microprocessor supervisors. Maxim's *MAX805L* and *MAX813L*

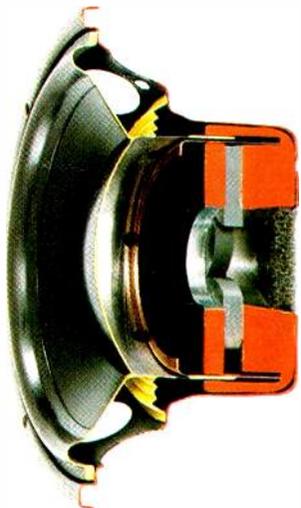
generate active-high resets when a microprocessor's supply voltage drops below 4.65V, in any phase of operation, this type of reset being needed by Intel devices, as opposed to active-low resets. A watchdog timer monitoring software operation issues a reset whenever activity on a selected i/o line lapses for over 1.6s. There is battery back-up for other circuitry in the system. Maxim Integrated Products Ltd, 0734 845255.

Memory chips

PCMCIA memory. Atmel announces the AT28C16-T, a low-power eeprom for the attribute memory of PCMCIA cards in a TSOP package measuring only 0.045in in the Z dimension and needing only 0.165 square inches of board. The IC needs a 5V supply at 30mA (10µA in standby) and it is reprogrammable in byte-sized increments. Write cycle occupies 1ms. Atmel (UK) Ltd, 0276 686677.

PCMCIA memory development. Highland has put together a development kit to allow Mitsubishi PCMCIA memory cards to be

Woofers. Low and mid-range loudspeaker drive units from Morel use shielded neodymium magnets for high flux density and techniques to linearise impedance and reduce distortion. A copper shorting device cancels voice-coil modulation, prevents impedance increase with frequency and eliminates position modulation. Units in the range 5in to 12in are made. 0284 828926



programmed from a PC. It includes a Mitsubishi 1Mbyte sram card, Elan J101 card reader/writer, application software and all documentation and accessories. Highland Technology, 0444 236000.

Mixed-signal ICs.

Television chip-set. Philips has a complete chip-set for image enhancement and flicker reduction in multi-standard, improved-definition television applications. It improves the quality of pictures on standard 50/60Hz transmissions and allows the design of a single, alignment-free Pal/Secam/NTSC board for use anywhere in the world. The set consists of the TDA9141 colour decoder/sync. processor, the TDA9150B or TDA9151B deflection controller, TDA8755 video A-to-D converter, SAA4950WP memory controller, SAA4940H noise reducer and SAA159WP back-end chip. Philips Semiconductors, 8071 580 6633

Wave-table synthesis. CS8905 and CS9203 by Sequoia are wave-table synthesiser chips offering the performance needed for professional keyboards and electric pianos. They are compatible with Roland GS enhancements to the General MIDI spec and with the MPC Level 2 Extended Multitimbral spec defining performance in Windows. Sequoia Technology Ltd, 0734 311822.

Power semiconductors

3A regulators. Semtech's LM1576/2576 3A switching regulators are interchangeable with National Semiconductor chips. They come in a range of versions including 3V and 3.3V types with fixed outputs from 3V to 15V and variable outputs from 1.23V to 37V, from inputs of 4-40V. Semtech Ltd, 0592 773520.

C-band power fets. 30W GaAs fets by Toshiba are designed to replace several lower-power transistors, giving 45dBm 1dB point total power and gains of between 10dB and 6dB, depending on frequency. Bandwidth is up to 800MHz, the six devices covering the 3.7-8.5GHz range. As an example, the TIM7785-30L operates at 7.7-8.5GHz, with intermodulation distortion of -43dBc. Toshiba Electronics (UK) Ltd, 0276 694600.

Current-mode PWM controllers. Unitrode's UCC12/3806 BicMOS PWM controllers have dual, 1A fet-driving outputs, a true differential input current sense amplifier and a well defined voltage threshold for turn-on. Maximum operating current is 1.4mA and delay from current-sense inputs to outputs is 125ns. Start-up current is 50µA. Unitrode UK Ltd, 081 318 1431.



Passive components

EMI suppression inductor. In a 1.6mm by 0.8mm by 0.8mm package, Murata's BLM11A12 chip inductor for EMI suppression has a current rating of 200mA DC and 0.5Ω maximum resistance. Impedance at 100MHz is 220Ω. Other models in the BLM series have 5-2000Ω impedances. Murata Electronics (UK) Ltd, 0252 811666.

Tantalum chips. Surface-mounted tantalum chip capacitors in Panasonic's KE series have no lead frame and are therefore only about half the size of the previous TE type, measuring 2 by 1.25mm. Values in the first to be made are 4.7, 6.8 and 10µF in ratings of 10, 6.3 and 4V. More in the series are due in April 1994. Panasonic Industrial Ltd, 0344 853827.

EMI/ESD protection. Murata's VFR303 combines electromagnetic interference suppression with protection against electrostatic discharges and is meant to protect cmos and TTL circuitry. It will handle ESD surges of up to 30kV and attenuates radiated interference by 30dB at 250MHz. Surtech Interconnection Ltd, 0256 51221.

SM inductors. Vishay's LPC range of inductors, which cover the 47-6800µH range, have wrap-round terminals printed on the ceramic base for increased reliability, solderless wire/pad bonding being used for invulnerability to reflow solder temperatures. The surface-mounted devices are rated at 7-10mA. Vishay Components (UK) Ltd, 0915 144155.

Connectors and cabling

Microwave cable assemblies. Flexible 18GHz cable assemblies by ITT Cannon can be cut to any size and terminated by the company with any current type of connector. They can be formed into the desired shape by the user. Diameter is the same as in the RG-405/402 semi-rigid cables, but losses are lower. An FEP outer is fitted. ITT Cannon/Sealelectro, 0256 473171.

Data-network connectors. M/A-COM Greenpar connectors are available for data networks including Ethernet and PC networks. Connectors are made in nickel-plated brass, the internal PC types being of die-cast material. They fit all common cables, come in BNC, N and Twinax

types, can be assembled with a low-cost tool and the N and BNC versions have an earth tagged terminal. M/A-COM Sales Ltd, 0344 869595.

Smart-card connectors. Connectors accepting ISO 7816 smart cards are made by ITT Cannon. They have 16 contacts for up to eight pads on the card, the contacts being gold over nickel-plated copper alloy to give 100mΩ contact resistance. Connectors are either in the lower-cost friction version or in the landing variation. RS Components Ltd, 0536 201234.

Card-edge connectors. Palladium-nickel edge connectors in Thomas and Betts's Holmberg range are capable of over 5000 mating cycles. The design accommodates a full range of board thicknesses and the contoured mating surface concentrates the force in the correct area. Individual contacts are replaceable. Operating conditions are 600-1500V DC and 3-5A. Thomas & Betts Ltd, 0582 608816.

Audio connectors. A range of professional audio connectors from Neutrik, distributed here by Verospeed, now includes right-angled versions of the XLR and XLR-compatible mains connectors, audio jack plugs and sockets. Verospeed, 0703 644555.

Filters

Band-pass filters. Covering most of the IFs currently in use, Chronos high-performance elliptic band-pass filters offer out-of-band rejection of 45dB, insertion loss of 1.5dB maximum and 1.2dB VSWR in the pass-band. Mounting is by 8-pin relay header or in leaded or non-leaded surface mounting. Chronos Technology Ltd, 0989 85471.

Hardware

Keyboard components. Keys and indicators by Rafi are for use in low-profile, membrane keyboards and flat data-entry equipment. The RF15 and RF19 series of board-mounted devices includes non-illuminated, spot-illuminated and fully illuminated keyswitches with several indicator options with standard or multi-coloured leds. Rafi (GB) Ltd, 0737 778660.

Instrumentation

Pulse-response measurement. IMP is short for Impulse response Measurement and Processing. It consists of a software/hardware package to be used with a PC to perform measurements on loudspeakers or associated equipment. Its output pulse drives the speaker, whose acoustic output is picked up by a microphone, the

output of this being fed back to the IMP board, digitised and taken to the PC by way of its printer port. The IMP software then analyses the data. Falcon Acoustics Ltd, 0508 78272.

Oscilloscopes. New Tektronix oscilloscopes include the *TDS320*, a 100MHz digital real-time instrument that allows sampling at five times its analogue bandwidth, 500Msamples on both channels without aliasing. It has the TDS user interface a 1K record length and auto measurement. The *TAS400* series comprises four analogue units with bandwidths from 60MHz to 200MHz Feedback Instruments Ltd, 0892 653322.

TV signal generators. Three new instrument families from Fluke cover most requirements in television and video testing. Features include 16:9 test patterns for wide-screen equipment; Y/C outputs for Super-VHS and Hi-8 recorders and satellite receivers; and video program system and program delivery control for testing automated video recorder operation. *PM5414V* offers basic tv and video test for Pal and NTSC; *PM5415* versions cover more extensive testing; and *PM5418* models offer the widest selection of functions and will test Secam equipment. Fluke (UK) Ltd, 0923 240511.

Laser analyser. Melles Griot/Photon Control offers the 13 *SKP 001 BeamAnalyzer*, which measures width, 3D intensity profile, power and position of a CW laser beam in real time, using a tomographic technique. Working with a PC, it will handle lasers in the 400-1050nm range with beam widths between 3µm and 5µm. The system consists of a plug-in card and software. Melles Griot Ltd, 0223 420071.

26.5GHz signal analyser. Rohde & Schwarz's *FSMS* is a spectrum and network analyser working in the 100Hz-26.5GHz range, finding application in the digital techniques used in mobile and directional radio, satcoms and microwave module and component test. Features include -142dBm inherent noise, -103dBc phase noise at 1kHz from carrier and some flexible software. Rohde & Schwarz UK Ltd, 0252 811377.

2MHz function generator. Variable-symmetry sine, square and triangular waveforms from 0.002Hz to 2MHz from Thurlby Thandar's *TG215* are digitally indicated in frequency, amplitude and DC offset. At low frequencies, resolution and speed of display up-date are preserved by arranging for the counter to read the VCO control voltage rather than signal period. Output is 20V peak from 50Ω or 600Ω and there is a TTL/cmos output. An external voltage-control input is provided.

Thurlby Thandar Instruments, 0480 412451.

Interfaces Optical Tx/Rx modules. Claimed by Toshiba to be the world's smallest and fastest, optical data transmitter and receiver modules are less than one-sixtieth the size of conventional systems but, with a transmission speed of 10Gbit/s, will handle the equivalent of 100 HD television channels or 150,000 telephone lines simultaneously. Five GaAs mesfet chips in 0.5micron technology comprise the set, the transmitter including multiplexer, laser diode driver and laser diode and the receiver a demultiplexer, pre-amp, decision IC and photodiode. Toshiba Corporation, fax (03)3456-4776.

Literature

Video/imaging ICs. Harris's video and imaging product guide describes 850MHz amplifiers, 8GHz transistor arrays, signal converters and processing chips such as FIR filters, convolvers, histogrammers, multipliers and buffers. There are also video switches, sync generators, PLLs and chroma processors. Harris Semiconductor (UK), 0276 686886.

T & M catalogue. Among the 29 oscilloscopes in the Hitachi Denshi 1993/4 catalogue is the *VC-704* four-channel, 100Msamples/s, 150MHz digitising instrument, which has its own printer, 64Kbyte to 2Mbyte sram, auto set-up, go/no-go comparison and window trigger. Other equipment described includes signal processor and analysis software, television measuring instruments, counters, plotters and function generators. Hitachi Denshi (UK) Ltd, 081-202 4311.

Data analysis tutorial. A free interactive software tutorial from National Instruments, *Analysis Advisor* covers data analysis techniques for use with LabVIEW graphical programming software and LabWindows automatic code-generation. The package deals with DSP, statistics, simulation and time-domain analysis, with demonstrations in each session. Requirements are a 386/33 or better, *Windows 3.1*, 8Mb memory and a VGA or SVGA. National Instruments UK, 0635 523545.

Emulator tutorial. In-circuit emulators in Pentica's *Alimo* series are the subject of a free tutorial disk. It shows screen displays concerned with high-level debug and trace qualification and indicates how they are used. Typical problems are covered in detail in a simulated emulation session. Pentica Systems Ltd, 0734 792101.

IGBTs. Toshiba's range of second-generation insulated-gate bipolar



transistors is the subject of a new brochure, which covers all the company's devices, including a 600A, 1200V module and a number of devices protected against overcurrent, over temperature and drive undervoltage. Toshiba Electronics (UK) Ltd, 0276 694600.

Power supplies

55W SMPS. A new series of cascad switched-mode power supplies, the *PUP* from Amplicon, gives 55W continuous power from single, dual or triple outputs. Input voltages are 85-264V, efficiency 65%, hold-up time 10ms and line regulation at full load ±0.5%. Amplicon Liveline Ltd, (Free)0800 525335.

Batteries. NiCad and nickel metal hydride cells made by GP Batteries use foam or sponge type electrodes to allow a maximum of active

Shielding tape kit. An engineering kit from 3M consists of eight foil shielding tapes in a dispenser and is intended for design, prototype and manufacture of small quantities where a few inches are needed. Each roll is 1.9cm wide and 3.66m long, the kit being priced to allow one for each engineer 3M United Kingdom plc, 0344 858000.

1.3GHz counter. Although a hand-held instrument, Thurlby Thandar's *PFM1300* digital frequency meter covers the 5Hz-1.3GHz range and uses reciprocal counting to give 0.001mHz resolution at low frequencies, the reciprocal being computed for display. A low-pass filter can be selected. Thurlby Thandar Instruments, 0480 412451.



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material. NiCad 1/3 AA cells have 180mAh capacity, AA cells 850mAh and 7/5AF cells 2Ah, while the nickel metal hydride types in 2/3AA give 550mAh, AA 1.2Ah and 7/5AF 2.3Ah. The company now has ISO 9002 approval. GP Batteries (UK) Ltd, 0278 445995.

In-cable DC-DC converter. Development of Microspire's SC series of 15W DC-DC smart converters has produced the latest 55mm by 18mm by 28mm diameter version, which is designed to be mounted inside a cable for use in towed arrays and similar marine and defence applications. Outputs are from $\pm 5V$ to $\pm 15V$ from inputs of 28V or 170V. Microspire UK Ltd, 0227 740368.

EMI suppressors. Surface-mounted EMI suppressors in the ACC range by TDK have current ratings between 1.5A and 3A and use a ferrite chip to allow wide-band operation with the correct conductor thickness for low resistance. Eleven versions have impedances from 60 Ω to 750 Ω and with resistance of 0.05 Ω and 0.04 Ω in the 1.5A and 3A types. TDK UK Ltd, 0737 772323.

Radio communications products

Directional coupler. Working in the frequency ranges 400-1500MHz and 1600-2000MHz, Murata's LDC series of surface-mounted directional couplers are claimed to be the world's smallest. They are meant for cellphone application, are shielded against interference and offer -3dB to -17dB coupling. Nominal impedance is 50 Ω , insertion loss 0.2dB and emissible power 3W. Murata Electronics (UK) Ltd, 0252 811666.

GPS receiver. Navcore MicroTracker is Rockwell's new credit-card-sized five-channel global positioning system receiver, which has an average power usage of 670mW. One option available with the new design is RTCM SC-104 compatible differential GPS, which improves error from 100m to 5m most of the time. It is designed to work with a passive antenna. Time to first fix is 20-30s from a warm start and dynamic tracking avoids the effects of foliage, buildings and vibration. Rockwell International, 010 33 93 00 33 01.

Mobile data terminal. Solo has a range of mobile data terminals for use by transport and distribution organisations and field workers needing access to a central database. The MDU800 series operates on the Cognito communications network, being easily configurable and taking a variety of input and output devices, particularly for use in vehicles. It is designed for use by unskilled people. Solo Electronic Systems Ltd, 0202 676859.

Switches and relays

Solid-state relays. High-voltage Form C relays are introduced by the American company Coto Wabash.

They provide 3750V RMS isolation, linear AC/DC operation and low thermal EMF characteristics with break-before-make operation. 200V, 200mA and 400V, 150mA versions are available. Coto Europe, +31-45-320838.

Reed relay. The Coto Wabash model 2342 changeover reed relay in 2C form occupies a board space of 5.3 by 20.6mm and is encapsulated in a steel shell. The switches come in 3V and 5V versions, are rated at 3W at up to 100V DC and peak AC, and maximum switched current is 0.25-0.5A continuous. Rhopoint Components, 0883 717988.

Transducers and sensors

Linear displacement sensors. Fast linear displacement sensors from Control Transducers are now reduced in length to give a length to measuring range of 1:1 and a 15kHz frequency response. No RFI shielding is needed, accuracy is within 0.15% and there is a range of signal processors. FS5000 operates over the -50 to 125°C temperature range. Control Transducers, 0234 217704.

Digital potentiometer. Based on an optical encoder, Control Transducers's DP100BT digipot produces 10, 20, 50 or 100 two-phase square-wave counts per revolution to provide a maximum of 400 binary 2bit code changes per rev. A free design guide is on offer, which advises on interfacing. Control Transducers, 0234 217704.



Computer board level products

Video card. TI's TIGAVIDEO card captures and displays high-res. images in Windows 3.1 in real time. It is supplied with free Adobe Photoshop Light software for image manipulation, capturing images from RGB, Pal, Y/C(SVHS, HI-8mm) inputs, still images being captured in .TIFF and .BMP format. The card provides screen resolutions from 160 by 160, 360 by 288 and 768 by 576 in 16-bit form. Texas Instruments, 0234 223252.

Development and evaluation

DSP Eurocard. MPE's TMS320C31 CPU board has 33Mflop capability in transient analysis, spectrum analysis, filtering and waveform generation. On a 3u high, 160mm deep card, the board includes up to 512K zero-wait-state static ram, two serial ports with RS232 or 2/4-wire RS485, 24 TTL i/o lines, a 15 μ s, 12-bit A-to-D converter and a socket for a 1MByte eprom. It



has access to all the MPE Powerboard i/o family and is suited to high-level languages such as Forth or C. MicroProcessor Engineering Ltd 0703 631441.

Fast firmware up-grades. Flash memory in PEP's VM30 computer allows complete firmware code change in situ with no hardware handling. 1Mbyte of 5V flash eprom on the VMEbus unit takes new code from disk or modem in a few minutes with little technical effort being needed. VM30 now operates at almost 12Mips and, with a 68882 coprocessor, at 4.5Mflops. PEP Modular Computers (UK) Ltd, 0273 441188.

DSP starter kit. Texas Instruments has a kit which includes everything needed to make a start in DSP, including a PC-linkable board, assembler and debugger software and documentation. Two versions are based on TMS320C26 or TMS320C51 DSPs. Polar Electronics Ltd, 0525 377093.

Software

PC signal analysis. Intelligent Instrumentation has the Signalyzer on offer - a data-acquisition, analysis and display software package running under Windows. From 1 to 32 channels of data at sample rates up to 10MHz are acquired in bursts to the PC ram or streamed to disk, the software then supporting both time and frequency-domain analysis, with FIR and IIR filter analysis being an extra. The software supports the existing range of PCI data acquisition boards. Intelligent Instrumentation, 0923 896989.

Emulator. The PC card-based Universal Emulator from Flash Designs offers a fast way of developing embedded C or Assembler code for all 4/8/16-bit microprocessors. Using the Flash DMA technique for accessing all registers and symbols allows very rapid program development and debugging. No monitor is required. The software will work with all popular assemblers and IAR/Keil C cross-compilers. It can also be used within Windows to allow comprehensive integrated debugging. Animate, continue to next break and snapshot are all included. Turbo-trace gives logic analyser functions such as tracing memory reads and writes in real time.

32-channel A-to-D card. An addition to IMS's PC-Lab range is the PCL-813 32-channel, 12-bit A-to-D on a 2/3 length card. Opto-isolators give over 500V DC protection and input range, A-to-D triggering and data transfer are all software-controlled. Analogue input ranges supported are $\pm 5V$, $\pm 2.5V$, $\pm 1.25V$ and $\pm 0.625V$, gain being set individually. Integrated Measurement Systems, 0703 771143.



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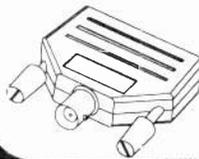
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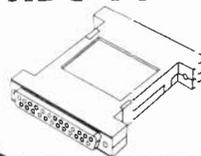
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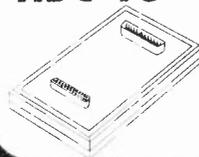
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DESIGN BRIEF

WHITE NOISE— WHITE KNIGHT?

Ian Hickman shows how to make noise work for you and looks at the principles behind an economical semiconductor noise source operating up to 1000MHz.

Fig. 1a. Simplified circuit diagram of a wideband rf noise generator employing a temperature limited diode.

Fig. 1b. Suggested rf noise generator using a Schottky diode.

Noise is an unwanted guest in all analogue circuits, so it seems perverse to want to make yet more noise; but the reason is simply summed up in the old adage "Know thine enemy".

A calibrated noise source is a great convenience when developing a low noise circuit, and – although other methods do exist – is almost essential for the accurate determination of the noise figure of a low noise amplifier or receiver.

The classical approach to the design of a wideband noise generator is to use a temperature limited thermionic diode. This is one where all the electrons emitted from the cathode are attracted directly to the anode: there is no cloud of electrons forming a space charge surrounding the cathode. The effect is usually achieved with a pure tungsten filament cathode, fed with a smooth dc current which can be adjusted to set the anode current to the desired value.

The resultant noise is described as shot noise since the electrons forming the anode current fall onto the anode like lead shot onto a corrugated iron roof. If there is no residual gas in the valve to be ionised, and the anode voltage is not so high that secondary emission results, the rms fluctuation (noise) current i is related to the dc anode current I , according to the formula:

$$i^2 = 3.18 \times 10^{-19} \times I \times df$$

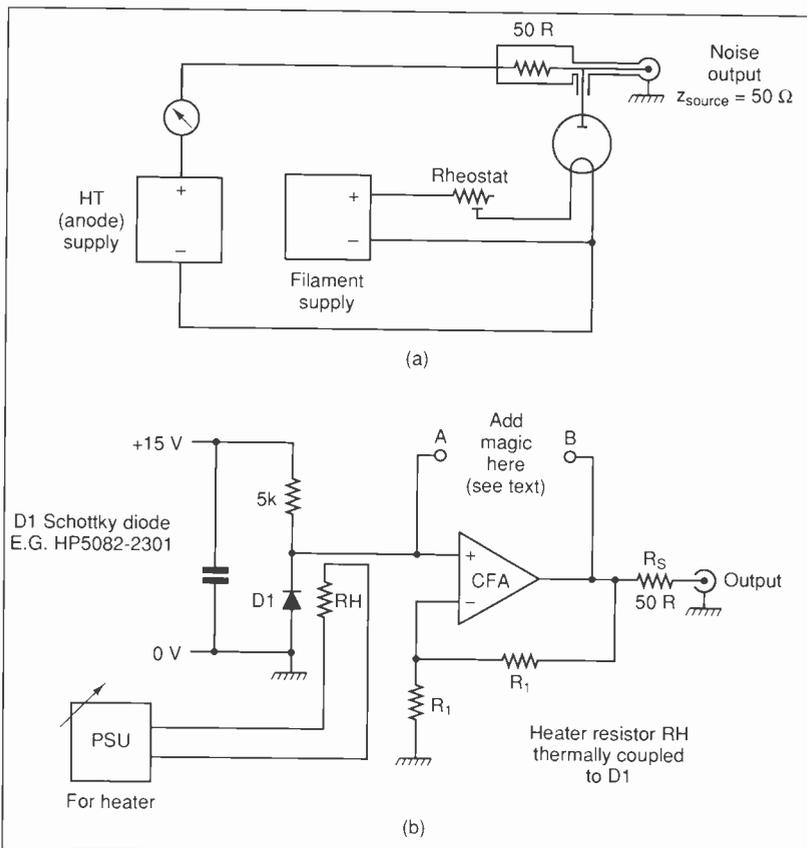
where I is in amperes and df is the bandwidth of interest. By contrast, the noise current i flowing in a short-circuit across a resistor R , is given by:

$$i^2 = (1.59 \times 10^{-20} \times df)/R$$

so that a noise diode passing an anode current of I gives as much rms noise current as an equivalent resistor $R_{eq} = 0.05/I\Omega$

Noise current available from a noise diode permits construction of a noise source whose output forms a known, absolute standard. **Figure 1a** shows the arrangement, where the anode current is passed through a 50Ω resistor. This forms a source of noise matched to the circuit under test, the magnitude of the noise power relative to that of a 50Ω resistor being precisely known.

In theory, df can extend from 0Hz up to any frequency, although in practice an upper limit is set by the shunt capacitance to ground of the noise diode anode, as well as other factors. Rhode and Schwarz are probably the best known European manufacturer



of noise sources of this kind, with a specified operating range up to 1000MHz.

Measurement

Measurement of receiver noise is simple. With the noise generator connected to the receiver under test and the diode anode current zero, the noise output from the receiver is noted, using a suitable measuring instrument. Then 3dB of attenuation is inserted between the receiver output and the measuring instrument, and the output from the noise generator increased to restore the previous level.

Noise supplied by the noise generator is equal to the receiver's own front end noise. A meter in the noise generator monitors the anode current and is directly calibrated in dB above thermal, with corresponding values of kTB also given.

Practical considerations in manufacture of the noise diode limit the maximum output that can be achieved to around 15dB above thermal. If the aim is to measure the noise figure of a receiver higher than this, the output of the noise generator must be amplified first. Clearly the amplifier used for this purpose will need to have a good signal-to-noise ratio.

A solid state alternative to the thermionic diode noise generator was described by TH O'Dell¹. Here, the source of noise was the flow of reverse leakage current induced by the creation of hole – electron pairs in a photodiode, by photons illuminating the same.

It was stated that the reverse photodiode current is subject to the same relation between dc and its shot noise component as the saturated thermionic diode. But, as the practical limit to the available current, at 100µA, is only about 1% of the maximum current possible with the thermionic diode, we must work at an impedance level of around 5K and provide a 10:1 turns ratio transformation to 50Ω. As just 1pF of stray capacitance across a 5K source would give a -3dB point of 32MHz, a tuned transformer was used, this having the additional advantage of absorbing the self capacitance of the diode as well as that of the transformer.

Substitute diode

Not having a sample of the HP5082-4220 photodiode to play with, I wondered whether an HP5082-2301 (1N5165) Schottky diode could be used. The paint was scraped off and the diode's reverse leakage current at $Vr = 15V$ found to be totally unaffected by any practical level of illumination.

But, there is another mechanism for inducing hole-electron pairs in a semiconductor – heat. The leakage of the sample diode at $Vr = 15V$ and 25°C was 0.056µA, against the manufacturer's quoted maximum of 0.300µA. Raising the temperature of the device by holding it close to, but not touching, a soldering iron, produced a leakage current of more than 100µA, at an estimated temperature of between 200 and 250°C. This is beyond the device's top rated temperature of 150°C, but on being left to cool, the current fell to less than a microamp in seconds and right back to 0.056µA within a few minutes.

Thus this scheme might form the basis of a practical noise generator, the utility of which would clearly be much increased if it were broadband, rather than

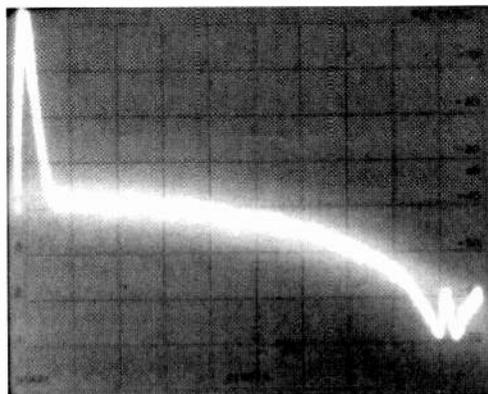


Fig. 2a. Output spectrum of National Semiconductors MM5437 digital noise generator (output 2, pin 5), with top-of-screen reference level -20dBm, resolution bandwidth 3kHz, video filter maximum, 10dB/div vertical, 20kHz/div horizontal, and centre frequency 100kHz. The output of the MM5437 was not low pass filtered, so some 180kHz clock breakthrough is visible in the first zero of the spectrum.

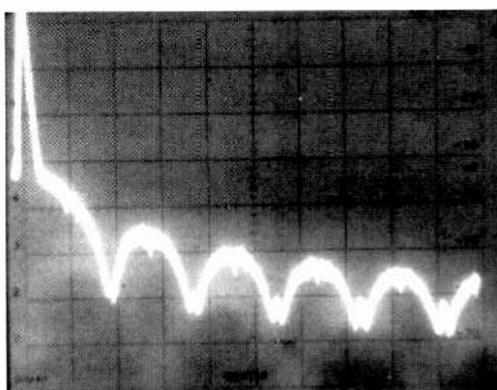


Fig. 2b. As Fig. 2a except resolution bandwidth 10kHz, 100kHz/div horizontal, and centre frequency 500kHz.

limited to a spot frequency¹. Figure 1 shows such a possible scheme.

In addition to the diode's self capacitance of under 0.3pF at 15V reverse bias, there is the input capacitance of the amplifier to consider, typically several pF, limiting the useful top frequency to less

Fig. 3a. Circuit used to test the base-emitter junction of a 2N918 as an rf noise source.

Fig. 3b. Circuit of a noise source using the base-emitter junction of a BFR90A in reverse breakdown.

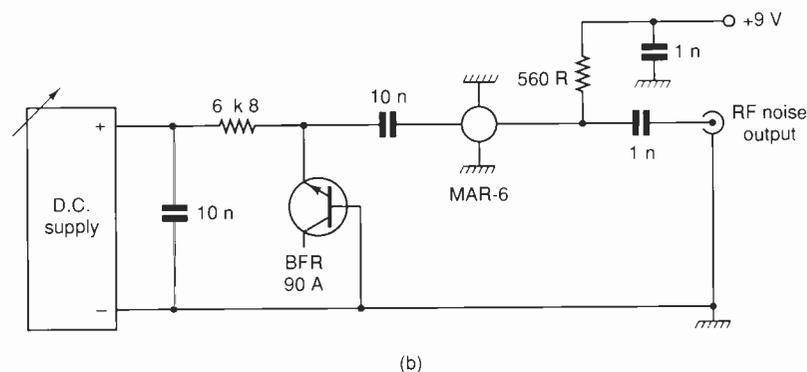
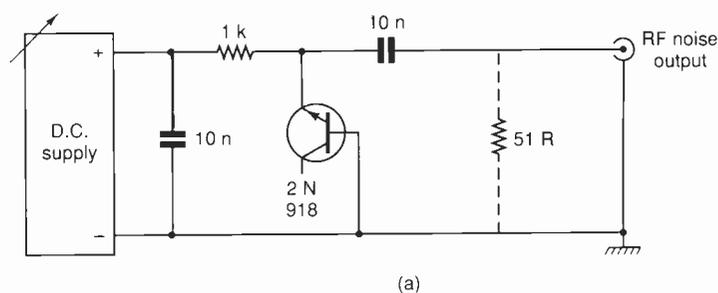


Fig. 4a. Output of the noise source of Fig. 3a: lower trace, diode current zero; middle trace, diode current 156 μ A; and upper trace, diode current 310 μ A (10dB/div vertical, reference level -33dBm, centre frequency 500MHz, span 1000MHz, IF resolution bandwidth 1MHz, video filter maximum).

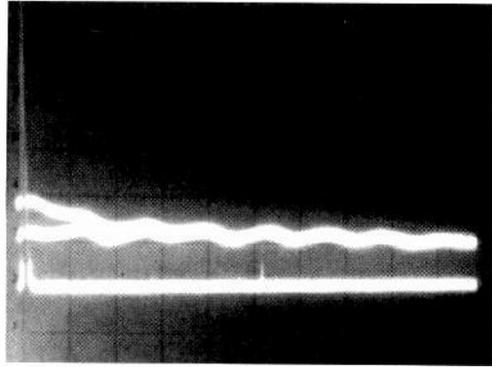


Fig. 4b. As Fig. 4a, but a 51 Ω load resistor added to the source.

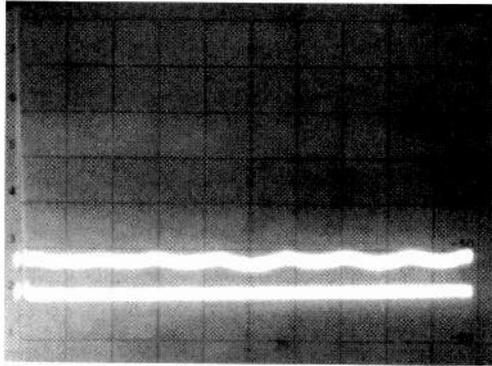
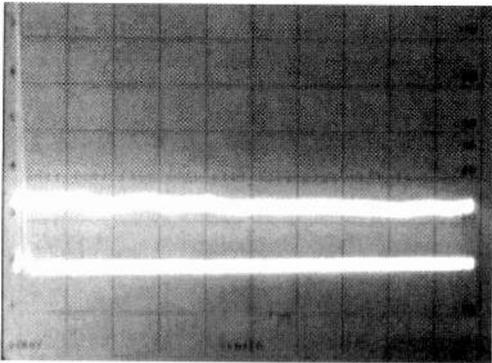


Fig. 4c. Output of an amplified BFR90 zener source: lower trace, diode current zero; upper trace, diode current 1.3mA (10dB/div vertical, reference level -33dBm, centre frequency 500MHz, span 1000MHz, IF resolution bandwidth 1MHz, video filter maximum).



than 10MHz. But, given a modern low noise current feedback op-amp with a bandwidth at a non-inverting gain of two approaching 1GHz, the high frequency attenuation of the diode noise due to the total shunt capacitance could be avoided by cancelling it with an equal shunt negative capacitance to ground². Here, it just involves connecting the equal capacitance between points A and B in Fig. 1b.

Noise measurements are also necessary in audio circuits, and Fig. 1 of reference 3 shows three different audio frequency noise generators, while pointing out that they can conveniently be replaced by a simple digital noise generator, the National Semiconductors MM5437, shown in Fig. 3 of that article. This also shows the typical output noise in the time domain, that is waveforms.

Figure 2a shows the output of the MM5437 in the frequency domain, indicating that the noise level is virtually flat to 60kHz and only 3dB down at about 100kHz. Note, however, that this is with no low-pass filtering of the pseudo random digital output from the chip, so that the amplitude distribution is not Gaussian.

The omission of any low-pass filtering also makes the first zero of the output at 180kHz very obvious, with a spectral line due to clock breakthrough sitting

in its middle. The spectrum repeats, at diminishing amplitude, around multiples of twice the clock frequency, so the device can be used as a noise source, over limited bandwidths, at frequencies well above that of the clock, up to low radio frequencies (see Fig. 2b).

Higher frequencies

For higher radio frequencies, while a noise source based on a temperature limited noise diode source is convenient for lab measurements, it is too large and expensive for building into equipment. One based on the shot noise in an illuminated photodiode or a heated Schottky diode is likely to be limited to operation at frequencies in the lower VHF range.

A number of manufacturers specialise in producing compact semiconductor noise sources for OEMs to build into equipment, probably the best known maker being an American company⁴.

Figure 1b of reference 1 shows a zener diode used as a semiconductor noise source, with the comment that this can work up to vhf, given a suitable diode. A 2N918, running the base emitter diode in reverse breakdown with the collector left disconnected, with some qualifications, works well and has been used as a noise diode in the built-in test equipment of a military tactical radio system, as a go/no-go test of the front-end and IF modules.

Figure 3a shows a circuit using a 2N918 thrown together to reproduce those results, the performance being as in Fig. 4a. The lower trace is with the diode current at zero: it shows the noise floor of the spectrum analyser, with a local tv station just making itself visible at around 510MHz.

The top-of-screen reference level is -33dBm so the trace is at -93dBm, and (assuming the noise bandwidth of the analyser's 1MHz IF resolution bandwidth filter is equivalent to a 1MHz wide brickwall filter) this corresponds to -153dBm/Hz. Thermal weighs in at -174dBm/Hz, equating to a noise figure of 21dB - rising to 22dB at 1GHz.

This is good for a spectrum analyser, since these are always designed primarily for linearity rather than sensitivity; 25dB is a not untypical noise figure. With a diode current of 156 μ A, the power delivered to the analyser is about 12dB above the noise level, or about 33dB above thermal - a lot more than one can get from a noise generator using a thermionic noise diode. If the current is increased to 310 μ A, the noise level above 300MHz is unchanged, but the level rises below that frequency. Further increases of current see the picture alternating smoothly between the two upper traces.

The noise is definitely falling off by 1500MHz, so the total noise delivered to the 50 Ω load presented by the spectrum analyser's input is roughly -60dBm or a bit less, around 0.001 μ W.

With the base emitter breakdown voltage of 4.5V, 156 μ A represents a power input to the diode of 700 μ W, giving a fairly low efficiency as a noise generator, but still much higher than a thermionic

noise diode. The variations of around ± 1 dB in noise output were a mystery at this stage. Their periodicity is around 150MHz, corresponding to a round journey in co-ax of around 65cm, but the circuit was connected to the analyser by a lead of no more than 6cm, including the BNC plug.

Figure 4b shows the effect of connecting a 51Ω resistor in parallel with the noise output; the variations in level have been largely damped out, especially up to 300 or 400MHz. The level has apparently fallen by 4 or 5dB, but is so close to the analyser noise that the latter is contributing to the indicated level. The true fall is probably nearer 6dB, which is what one would expect if the diode acted as a perfect constant noise-current generator.

So we have a matched 50Ω noise source, albeit at lower power. Where the noise power was $i^2 \times 50$, with the extra resistor in parallel it becomes $i^2 \times 25$, the rms noise current i being constant. Not only is noise power halved, but half of what there is, is dissipated in the additional 50Ω resistor, hence the 6dB drop in output.

Another *2N918* was tried in this circuit with generally similar results, except that there was no value of operating current that would avoid some rise in output below 300MHz. As the *2N918* is a very ancient device, an obvious next step was to try a more modern transistor. A *BFR90A* was therefore connected in circuit, but found to give a much lower output than the *2N918*, barely above the analyser's input noise. A 20dB broadband amplifier stage, using a Mini Circuits⁵ *Mar 6* type amplifier was therefore added, as in **Fig. 3b**, the resulting output then being about 13dB above analyser noise (**Fig. 4c**).

Variations in output level

The variations of output level with frequency are quite low right up to 1000MHz. This is not due to the different device but to certain other circuit changes. The 1k resistor in **Fig. 3a** was an 1/8W miniature carbon film axial lead type. Such resistors are made with a film of about 1% of the nominal value, required final value being achieved by making a spiral cut in the film.

Such resistors thus have an appreciable inductance, though it is often possible to ignore this due to its low Q. The output shown in **Fig. 4c** was achieved with a different feed resistor, **Fig. 3b**, namely a 6k8 sub-miniature solid carbon type, with a further substantial improvement by selecting the optimum supply voltage to the amplifier. With the aid of a heater, the whole noise generator circuit was raised to +75°C, with no measurable change in output noise at any part of the 0 to 1000MHz range.

While the circuit of **Fig. 3b** offers the basis of a potentially useful noise source, its output level is fixed, not readily adjustable from thermal upwards as in the case of a thermionic diode noise source.

However, this limitation is easily circumvented by the addition of a step attenuator. As the attenuation is increased from zero, the noise delivered to the circuit under test is reduced, in principle only reaching

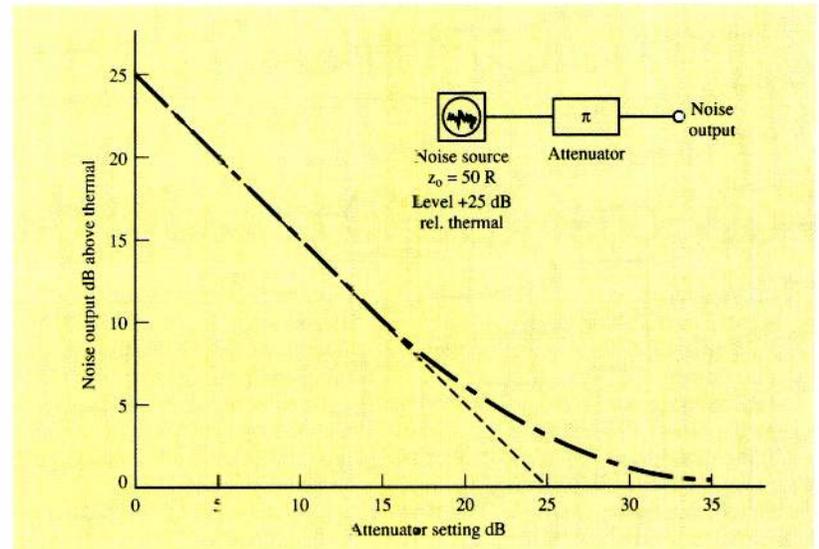


Fig. 5. Showing the output from a noise source 25dB above thermal via an adjustable variable attenuator, against the setting of the attenuator.

thermal when the added attenuation is infinite. In practice, as **Fig. 5** shows, 31dB is sufficient to reduce a noise level of 25dB above thermal to a mere 1dB above thermal, low enough to test any amplifier or receiver operating at room temperature, with the possible exception of a parametric amplifier.

A noise output of 25dB above thermal was mooted, as it is sufficient for most applications (and certainly much more than obtainable from most thermionic diode noise sources), but well below the level available from the circuit of **Fig. 3b**. This allows for the fitting of a fixed 9dB pad at the output of the *Mar 6* amplifier, which would be enough to provide a source with a return loss of 18dB even if the output vswr of the amplifier were infinity. In fact, the output vswr of the *Mar 6* is 1.8:1 maximum up to 2GHz, corresponding to a return loss of 11dB, so the addition of a 9dB pad at the output would provide a noise source with an output vswr of less than 1.08:1 – a return loss of around 30dB.

It seems clear that a very economical semiconductor noise source operating up to 1000MHz, with a calibrated adjustable output flat to within ± 1 dB and suitable for the laboratory measurement of receiver noise figures and other purposes, could be constructed for little more than the cost of the variable attenuator: or no more than a few pounds if an attenuator is already to hand.

References

1. "Self-calibrating noise source using silicon," TH O'Dell, *EW + WW*, February 1993, pp 170-172.
2. "Negative approach to positive thinking," Ian Hickman, *EW + WW*, March 1993, pp 258-261.
3. "Making a right white noise," Ian Hickman, *EW + WW*, March 1992, pp 256-257.
4. Noise/Com, E.49 Midland Avenue, Paramus, New Jersey, 07652, USA. UK agent Densitron Microwave, 4 Vanguard Way, Shoeburyness, Essex, SS3 9SH. Tel: 0702-294255, fax: 0702-293979
5. Mini Circuits, a division of Scientific Components Corporation, PO Box 350166, Brooklyn, New York, 11235-0003, USA. UK agent BFI Electronics, BFI Ibexa House, Burnt Ash Road, Quarry Wood Estate, Aylesford, Kent, ME20 7NA. Tel: 0622-882467, fax: 0622-882469.

APPLICATIONS

Low-cost 198 kHz radio data receiver

Phase-modulated data from BBC Radio 4 transmissions can be received via a single IC from GEC Plessey and a common, low cost op-amp.

As described in note AN86, the SL6659 is a general-purpose FM radio IC. It contains a limiting IF strip and FM detector, both of which are used in this application. Further, the device contains a signal-strength indicator and a mixer stage. It needs a supply of between 2.5 and 7.5V at less than 2mA.

Within the note are equations detailing how a 1.8V peak-to-peak phase signal is obtained from the $\pm 22.5^\circ$ phase modulation using integration.

Carrier at 198kHz received on the tuned antenna should be adjusted for maximum at test point TP₁ via the trimmer capacitor, Fig. 1. Because of the frequency of the broadcast, the rf signal is affected by interference from switch-mode supplies in televisions and monitors. Without the crystal filter following the fet buffer, the

receiver fails to work consistently at a metre away from a switch-mode supply. With the filter however, the distance at which problems start falls to a few centimetres.

To reduce Q, a resistor is placed in series with the crystal. Sourcing this series resonant 198kHz crystal may prove difficult; an address is given later.

Limiting within the SL6659 turns the signal into a 198kHz square wave, available at pin 1. Phase shifting takes place in the quadrature tuned circuit and the resulting signal is fed to the demodulator. At this stage, the signal is compared with the in-phase signal to produce a demodulated output at pin 3.

Amplitude and source impedance of this signal are 10mV pk-pk and 40k Ω respectively. Buffering via the transistor is needed in order to drive the Sallen and Key filter.

Tuning of the quadrature circuit is set by DC voltage at the collector of the transistor. When adjusted for resonance at 198kHz,

output at 198kHz jumps from 1.2 to 4V. The inductor slug is then tuned to set output voltage at TP₂ to 3V DC.

A critical component is the quadrature coil. Since the inductor has a high Q of 200, temperature drift, winding settling and other ageing effects become significant. As a result, careful design is needed. An RM6/160 was used in the prototype.

Sallen and Key filter

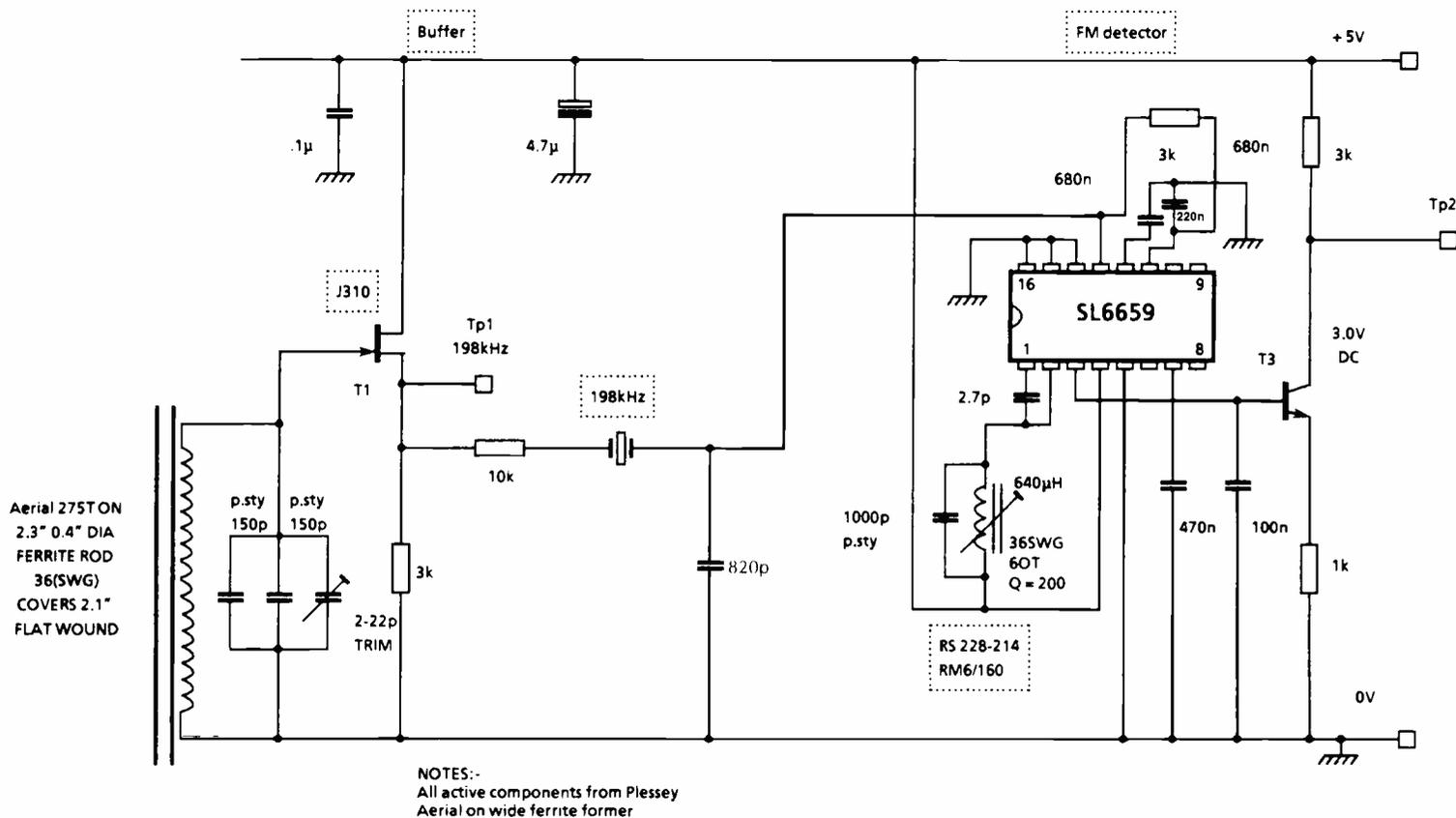
Initial low-pass filtering removes interference, which could cause distortion and rapid DC shifts on the output, Fig. 2. Next the signal is amplified 47 times. Gain is adjustable by altering the 47k Ω resistor, which may be necessary to compensate for the quadrature-coil Q factor.

A clean signal, as in the upper part of Fig. 3, should now be present at TP₄. If the signal is distorted, check the Sallen and Key filter bandwidth, the crystal series resistor or the 330 μ F capacitor.

Output from the integrator will be between

Radio 4 198kHz Phase modulation

Fig. 1. This low-cost radio-data receiver consumes only a few milliamps. Potential applications for radio-data information are numerous, among them high-accuracy clock/timer/calendars that never need adjusting.



1 and 3V pk-pk with a DC component that can drift between 1.5 and 4V. This drift is removed by a comparator which produces the final rectangular wave signal. A small piezo-electric sounder over the output helps alignment.

Output data comprises two frequencies of 12.5 and 25Hz. From these, both data and clock signals can be recovered.

Clock recovery is best carried out using a microprocessor to detect edges within the phase waveforms. When the clock is

recovered, the data is sampled to see if there is a falling or rising edge within the middle of each 40ms clock period. A falling edge represents logic one and a rising edge logic zero.

Crystals can be obtained from AEL Crystals Ltd, Worth Corner, Turners Hill Road, Crawley, West Sussex, RH10 7SL. Tel: 0293 882299.

GEC Plessey Semiconductors, Cheney Manor, Swindon, Wiltshire SN2 2QW. Tel: 0793 518000.

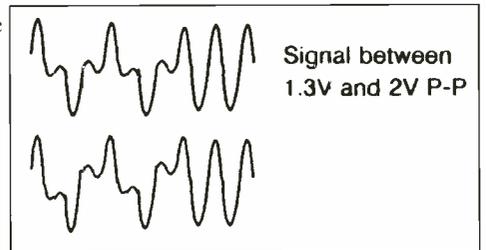


Fig. 3. Radio Data output at TP4 should be undistorted as in the top trace. If not, remedies are suggested in the text.

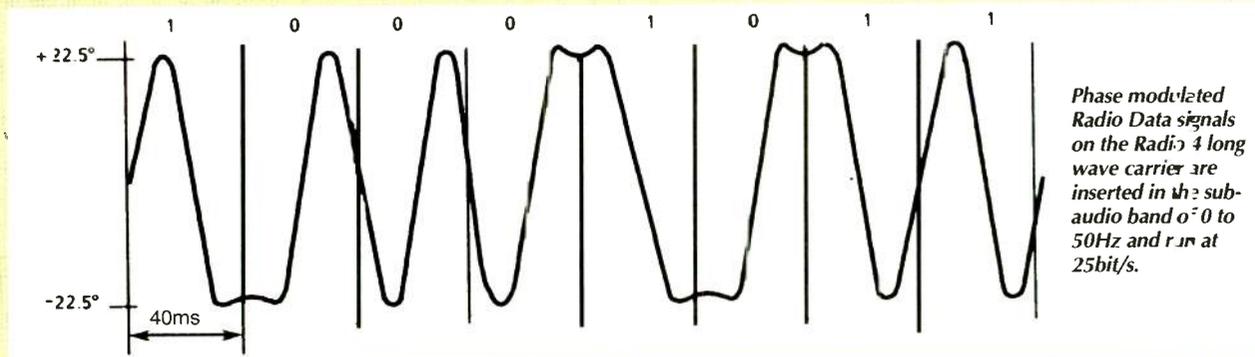
Radio Data

From February 1988, BBC Radio 4 on long wave shifted in frequency from 200kHz down to 198kHz. This carrier now also contains phase-modulated Radio Data as shown here. To avoid

the data affecting the audio information, it is slotted into the 0 to 50Hz band below the audio range at 25bit/s.

One block is transmitted every two seconds, with block zero starting at the

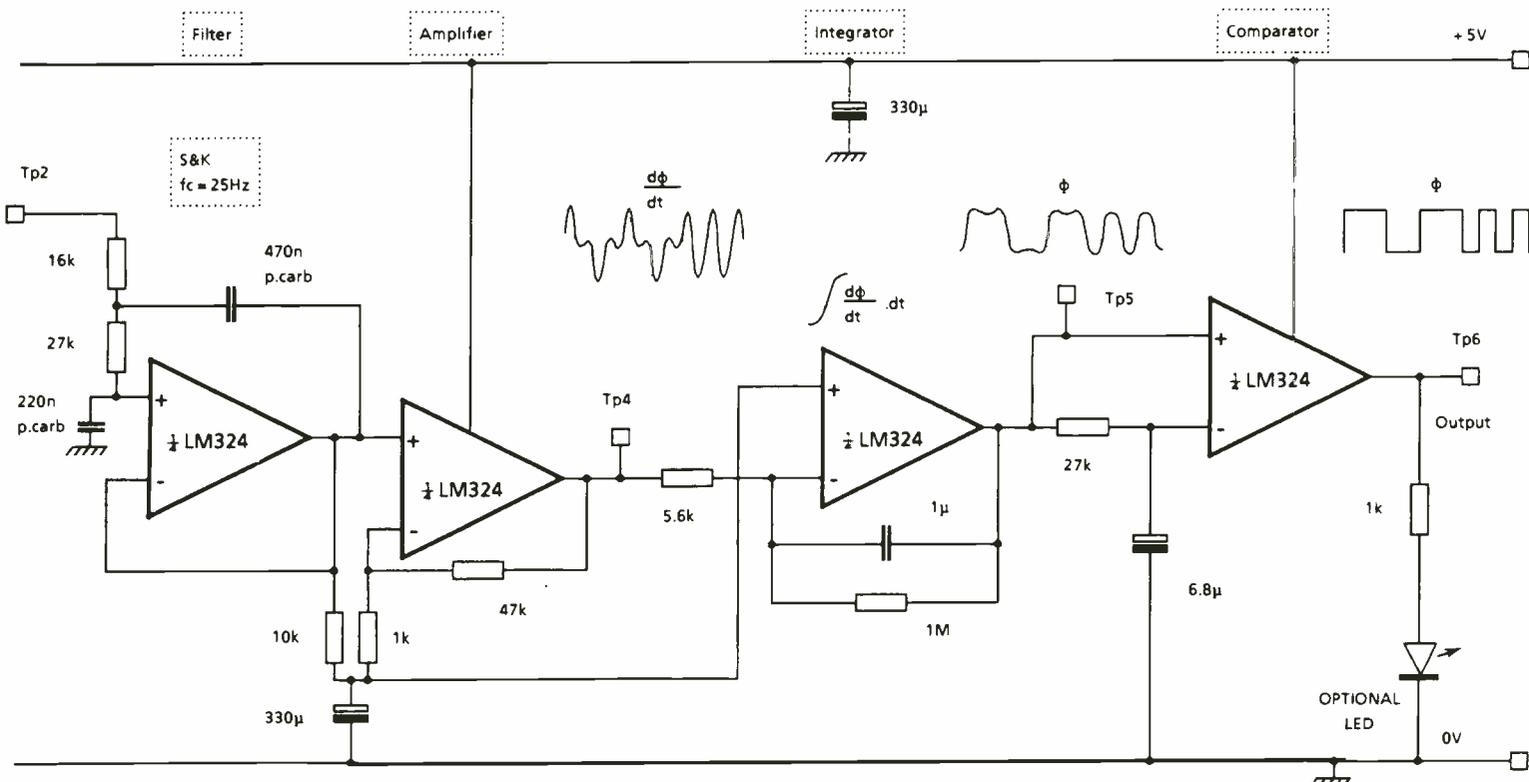
top of the minute. Block 29, the last block, contains time of day, month, year, leap year and local offset information. Each block also contains a cyclic redundancy check word.



Phase modulated Radio Data signals on the Radio 4 long wave carrier are inserted in the sub-audio band of 0 to 50Hz and run at 25bit/s.

Radio 4 198kHz Phase modulation

Fig. 2. Raw information from the Radio Data receiver is converted to a rectangular waveform ready for clock and data recovery by this filter, amplifier, integrator and comparator.



Current conveyor circuits

Unlike op-amps, current conveyors do not have their bandwidth restricted by feedback. According to LTP Electronics' data sheet on the CC1101, they extend familiar op-amp functions to 100MHz bandwidths. Additionally, they remain stable with both inductive and capacitive loads.

These circuits illustrate that current conveyors need few additional passive components. Many circuits can be implemented by adding just one or two resistors so the problems of resistor matching are reduced too.

Bandwidth of the CC1101 dual conveyor is 100MHz and its slew rate is 2000V/ μ s. Each element comprises three terminals. The one marked X is a virtual-ground current-input/voltage-output. This i/o line exhibits low impedance. Terminal Y is a high-impedance voltage input while terminal Z provides the current output.

From Fig. 1 you can see that current mirrors replicate X-node current at the Z output, which has a high impedance. Port relationships of the conveyor are simply $V_X=V_Y$, $I_Y=0$ and $I_Z=I_X$.

Ideally, voltage at the Y input follows voltage at the X node and input current at node X is copied exactly by current at node Z. In practice though, the buffer between X and Y has an output impedance of about

10 Ω up to 10MHz. Above this frequency, the impedance starts to rise, reaching 25 Ω at 100MHz. This reduces accuracy of the V_Y to V_X transfer, and hence the accuracy of I_X to I_Z , as frequency rises.

The data sheet discusses voltage compliance and maximum current considerations. It also describes using two

conveyors as a single conveyor. This improves the voltage buffer by more than two orders of magnitude and the current buffer by one order of magnitude.

LTP Electronics, 2 Quarry Road, Headington, Oxford OX3 8NU. Telephone 0865 200767.

Fig. 2. As current conveyors work without feedback, they do not suffer from the same bandwidth limitations as op-amps. These circuits illustrate that they can reduce component count too, relative to conventional op-amps.

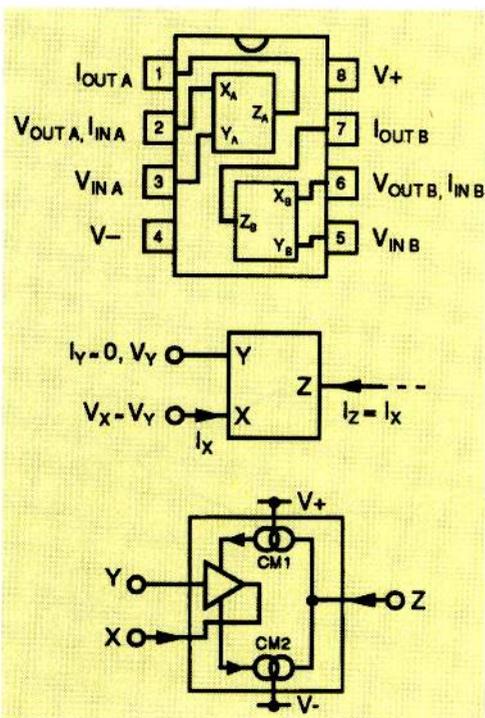
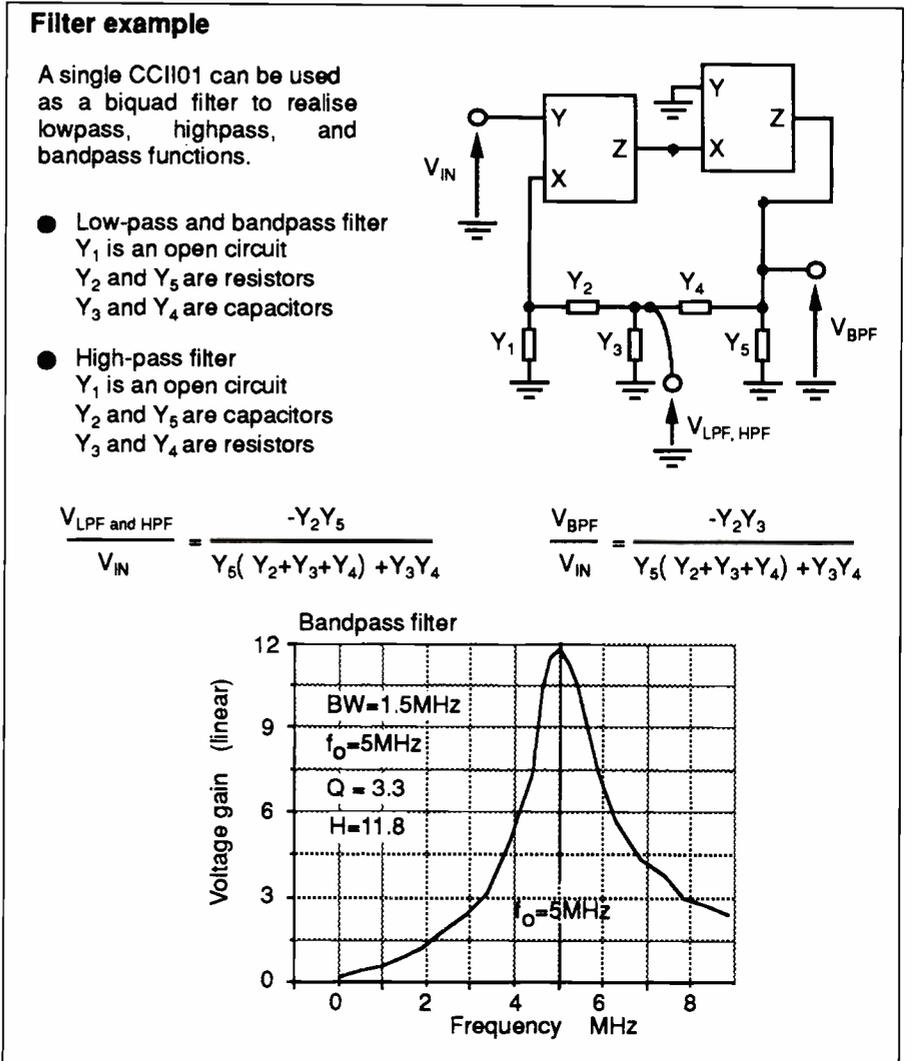
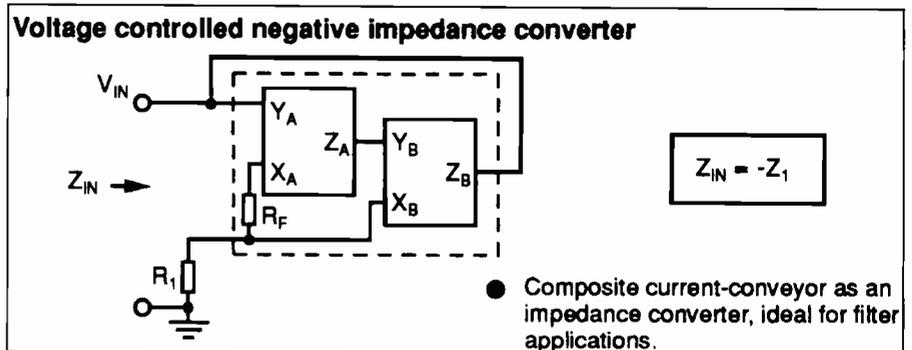
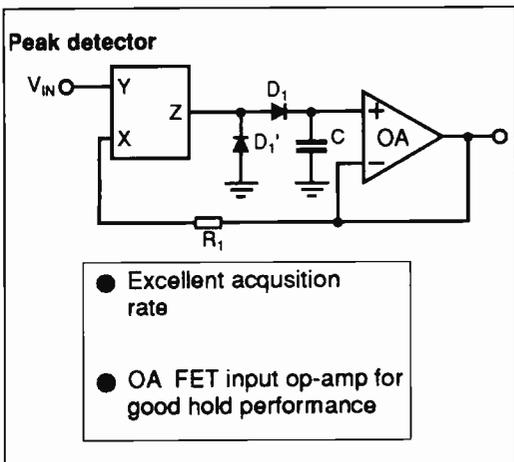
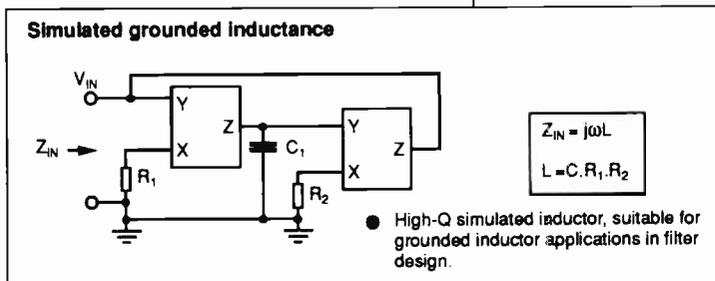
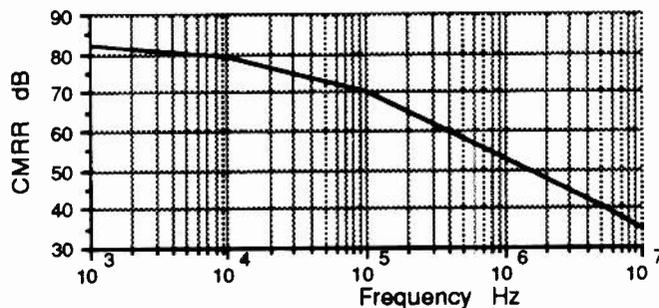
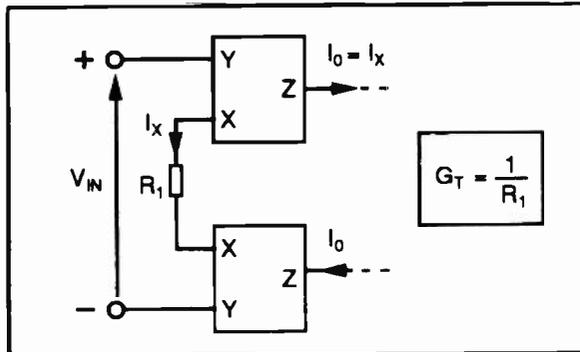


Fig. 1. Current conveyors are a combination of voltage and current-mode devices offering low component count and high bandwidth. Voltage at Y is followed by voltage at X and input current at X is copied by current at Z.



Fully balanced differential amplifier

- Differential inputs
 $I_x = \frac{V_{IN}}{R_1}$
- Common-mode inputs
 $I_x \approx 0$
- No matched components
- Typically CMRR=53dB at 1MHz

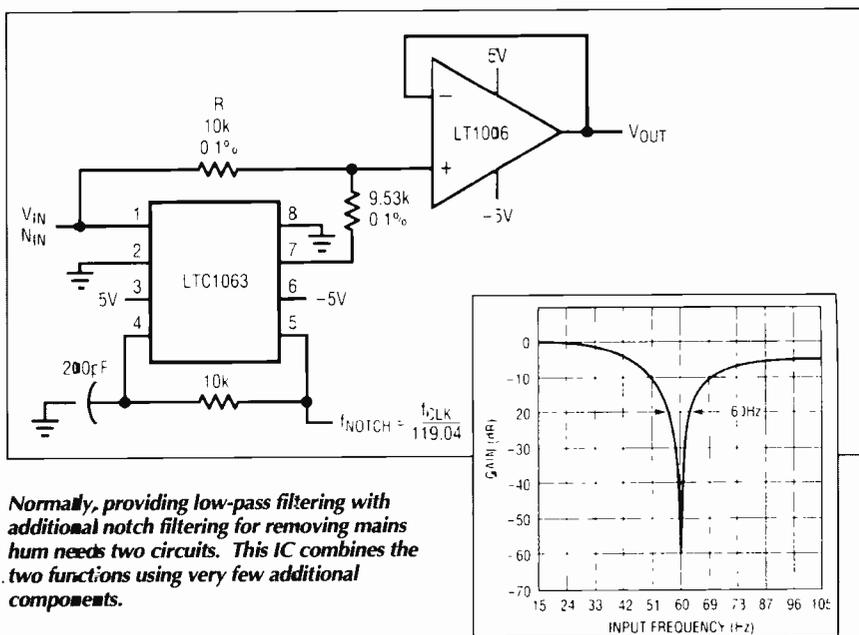


Lowpass/notch filter

This circuit is a sample from the *Linear Technology Chronicle*, Vol 4, No 4. Some applications require two transfer functions superimposed. An A-to-D converter for example may need low-pass filtering at its input to remove LF noise together with notch filtering to reduce the effects of mains hum. Normally this would involve two circuits but the solution shown uses only one IC followed by a buffer.

The LTC1063 is a fifth-order Butterworth filter tuned via a signal from an built-in RC controlled oscillator or external clock. It remains accurate down to DC, where it exhibits less than 1mV offset, and can be tuned continuously up to 50kHz.

To provide the notch, the upper two resistors are chosen to cancel exactly at the frequency where phase shift is 180°. This is at 1.19 times the corner frequency of the low-pass filter and is not adjustable. It is set by the internal transfer function of the filter. A 50.42Hz low-pass frequency for example sets the notch at exactly 60Hz (this circuit was designed for American mains hum of course).



Normally, providing low-pass filtering with additional notch filtering for removing mains hum needs two circuits. This IC combines the two functions using very few additional components.

Linear Technology, 111 Windmill Road, Sunbury-on-Thames, Middlesex TW16 7EF. Telephone 0932 765688.

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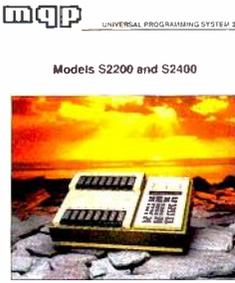
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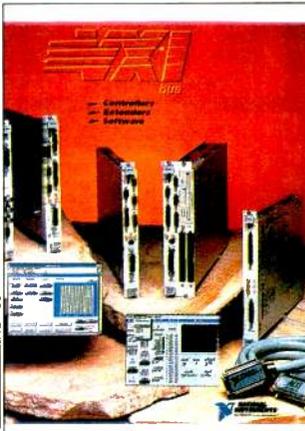
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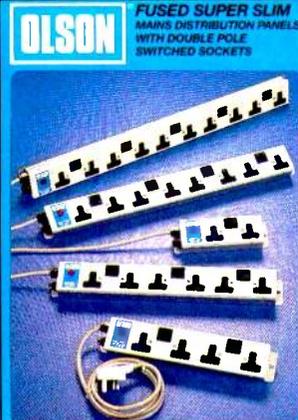
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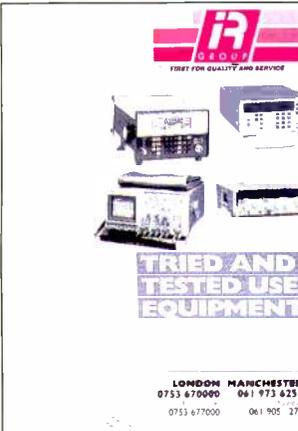
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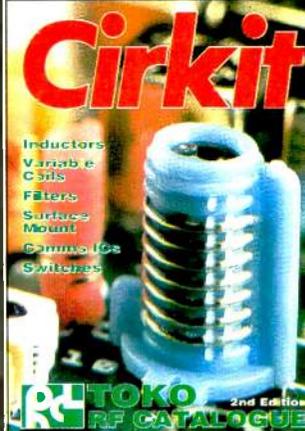
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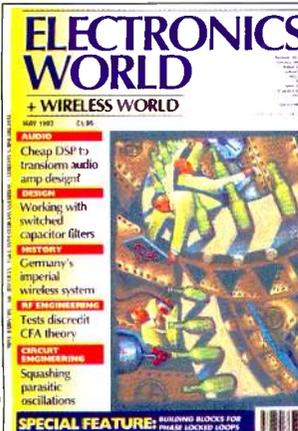
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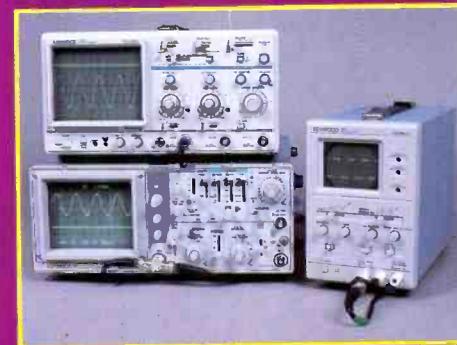
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FUNCTION GENERATORS - MX2020: 0.02MHz to 2MHz sweep function generator with frequency readout, output waveforms include sine, square, triangle, skewed sine, pulse and TTL, lin & log sweep, DC offset and symmetry. FG2020B: 0.5Hz to 500kHz function generator providing sine, square and triangle waveforms.



PS303 - £154.65, PS303D - £283.86, PS2243 - £119.26

POWER SUPPLIES - The PS series of low cost bench power supplies offer single or dual output with output protection. PS303: single DC power supply, 0-30V 3A. PS303D: dual tracking DC power supply 2x0-30V 3A. PS2243: DC power supply, 0-24V 3A.



MX9000 - £468.83

MULTI INSTRUMENT - The MX9000, suitable for a broad range of applications, combines four instruments including: 1. Triple output power supply with LCD offering 0-50V 0.5A, 15V 1A, 5V 2A, with full overcurrent protection. 2. An 8 digit LED 1Hz-100MHz frequency counter with gating rates of 0.1Hz, 1Hz, 10Hz, 100Hz providing resolution to 0.1Hz plus attenuation inputs and data hold. 3. A 0.02Hz to 2MHz full featured sweep function generator producing sine, square, triangle, skewed sine, pulse and a TTL output and lin or log sweep. Outputs of 50Ω and 600Ω impedance are standard features. 4. Auto/manual 3.5 digit LCD multimeter reading DCV, DCA, ACV, ACA, resistance and relative measurement with data hold functions.



MIC-4070D - £123.38

MIC-2060PA - £65.80

LCR METER - The MIC4070D LCD digital LCR meter provides capacitance, inductance, resistance and dissipation measurement. Capacitance ranges are from 0.1pF to 20,000µF plus dissipation. Inductance ranges from 0.1µH to 200H plus a digital readout of dissipation. Resistance ranges from 1mΩ to 20MΩ. Housed in rugged ABS case with integral stand complete with battery and probes.

CLAMP METER - The clamp meter provides digital readout and the following ranges, ACV to 600A, ACV to 750V, DCV to 1000V, Resistance to 2kΩ. Peak detector holds the max rms value. Audio continuity for short circuits.

CREDIT CARD ORDERS BY PHONE/FAX

To order any of the above items please send this form with your cheque/credit card details to: SAJE Electronics, 117 Lovell Road, Cambridge, CB4 2QW. Please send me the following instruments:-

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Emulation

See your code running before committing yourself to an EPROM. With S4's powerful and easy-to-use internal emulation system, download your code to S4, press 'EMULATE', and your target system runs in real time as if an EPROM was plugged in to the socket. Use S4's 'EDIT' command to make minor alterations to your code and see the changes happen *immediately* - just one reason why S4 is used by the world's car manufacturers to develop advanced engine management systems in real time! With S4 emulation there's no need for trailing cables or external power sources; earth loop problems are a thing of the past. S4 even emulates RAM.

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The Company

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