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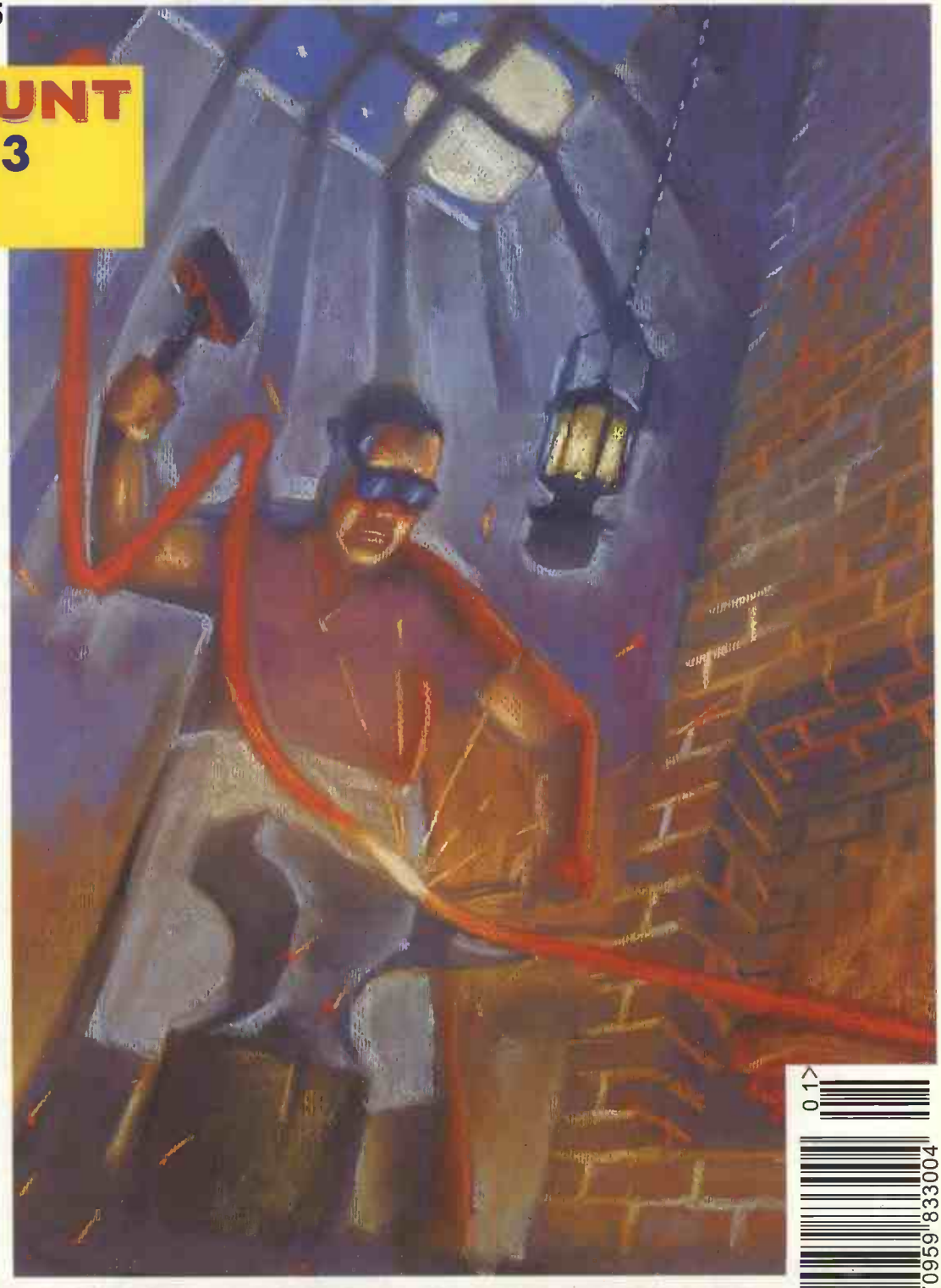

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40% DISCOUNT – PCB DESIGN SOFTWARE

***EW+WW* readers are offered a 40% discount on Powerware's recently launched upgrade to the windows-based pcb design package – Quickroute 3. See page 47 for our review of Quickroute and page 50 for details of this exclusive offer.**

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Between amplifier and speaker. Ivor Brown examines interactions between the audio power output and its load.

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Data logging via LPT1. Pei An describes a subsystem for logging 12bit analogue data via the printer port.

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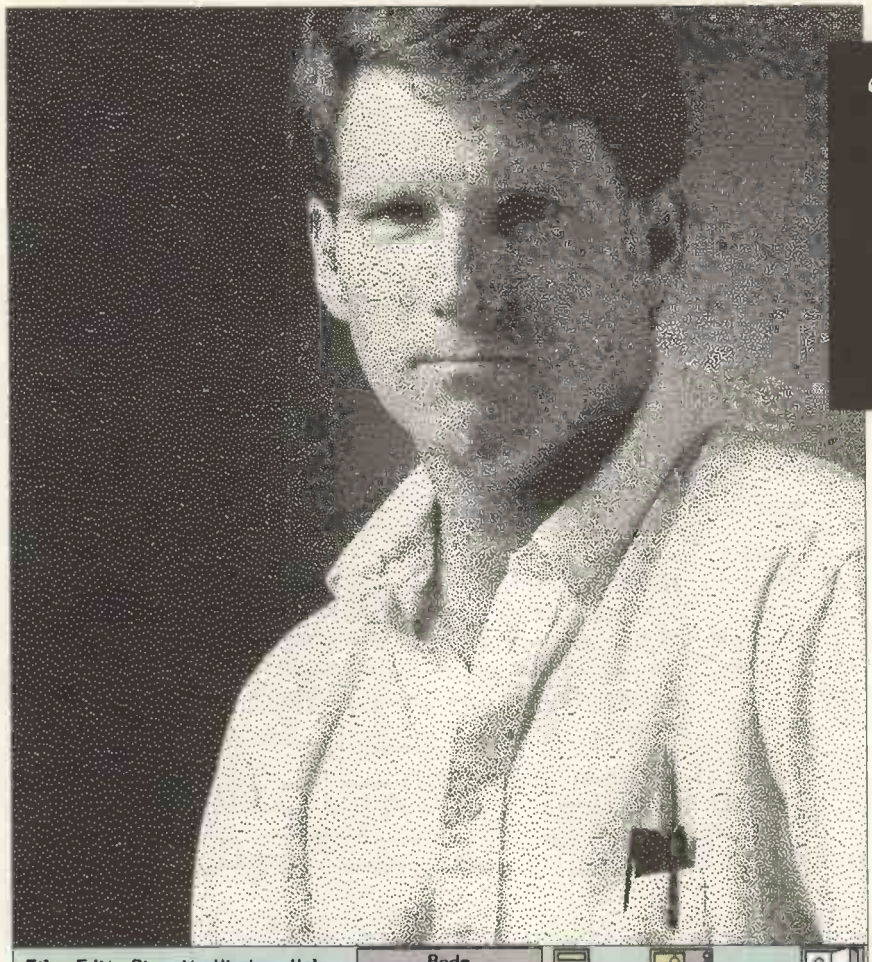

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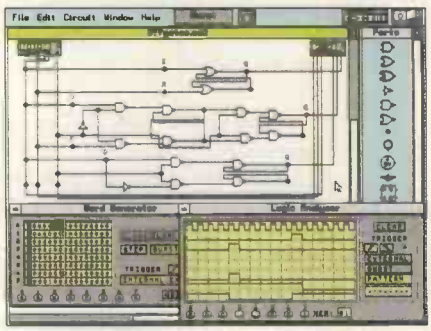
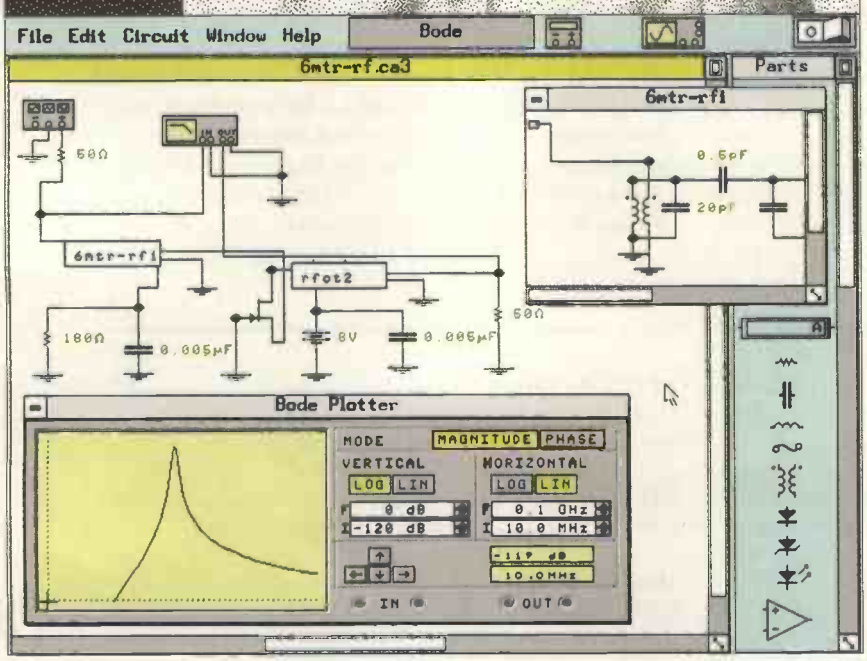


Digital Module: Analyzing a logic circuit.



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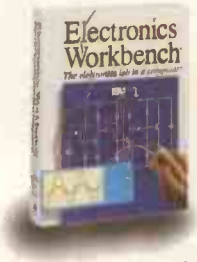

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CONSULTANT
Frank Ogden

DESIGN & PRODUCTION
Alan Kerr

EDITORIAL ADMINISTRATION
Jackie Lowe
081-652 3614

ADVERTISEMENT MANAGER
Richard Napier
081-652 3620

DISPLAY SALES EXECUTIVE
Malcolm Wells
081-652 3620

ADVERTISING PRODUCTION
Christina Budd
081-652 8355

PUBLISHER
Mick Elliott

EDITORIAL FAX
081-652 8956

CLASSIFIED FAX
081-652 8956

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Build or borrow?

We're used to differences between the UK and mainland Europe, but it was a surprise that last week's meeting between the Semiconductor Manufacturers Association (SMA) and the Parliamentary Information Technology Committee (PITCOM) should highlight quite such a fundamental gap between the official attitudes of the UK and mainland Europe towards microelectronics.

Both the UK and Continental Europe know microelectronics is important, but from there on attitudes diverge. Whereas Europe sees a necessity to protect and support European companies and European technology, the UK sees microelectronics as something done outside Europe which we need to attract inside if we are to get our share of the jobs and investment it involves.

Many a Brit has squirmed at Jessi (Joint European Sub-Micron Silicon Initiative) meetings when our European partners have referred to the 'modest' or 'limited' contribution coming from the UK.

Jessi, as the main body in the EU devoted to all aspects of silicon research – design, process technology and production equipment – is generally credited with revitalising the European semiconductor industry just as the collaborative VLSI Project brought Japan's chip industry to the fore, and the Sematech industry consortium halted the slide in US chip competitiveness.

Yet Britain's contribution to Jessi is only a tenth of Holland's, and Holland's is only 40 per cent of the individual contributions of Germany and France.

The SMA/PITCOM meeting last week explained this stinginess. There was nothing about improving the UK's competence in microelectronics; it was all about how best to spruce up the place to attract inwardly investing Japanese, American, Korean and Taiwanese microelectronics companies into the country.

Industry figures from US companies dangled tasty morsels before parliamentarians, such as the fact that between 100 and 150 new chip factories – each costing a billion dollars – will be built in the world in the next decade to a decade and a half.

MPs' eyes may glaze over at the mention of technology but figures like $100 billion to $150 billion have a good chance of attracting their attention. Compared to $2000 for asking a question it's serious money. All we have to do to get it, said chip men, is to adjust the tax laws a little, give grants on the basis of investment size rather than employee count (these plants won't employ many people), and skew the education system more towards the needs of these potential investors.

Is that good old Anglo-Saxon pragmatism, or is it a bit sad?

David Manners

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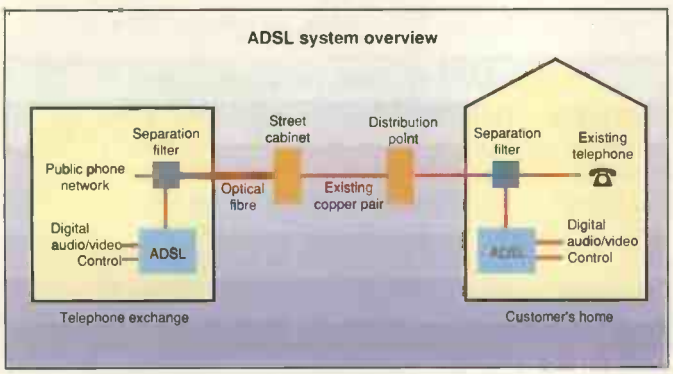
REED BUSINESS PUBLISHING

Video down the 'phone line

BT believes it has found the technology to transmit a 1.5Mbit/s digital video channel down 92 per cent of its 24 million copper telephone lines. As a result, next summer, 2000 households in Suffolk will get the chance to use interactive tv services such as home shopping, a multimedia version of TV Times and video-on-demand.

The asymmetric digital subscriber loop (adsl) technology which will achieve this is still in its infancy. US developers like Aware, Westell International and Analog Devices are now working on software and components which could both increase the capacity and cut the cost of adsl systems.

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



The BT system will be based on an adsl transceiver which Westell and Atlantech have been marketing for over 12 months. The principle of adsl is that a 16kbit/s data signal from the subscriber to the exchange is used to access a one-way broadband signal distributing video. It uses the carrierless amplitude and phase (CAP) modulation scheme originally developed by AT&T Paradyne for the US cable tv industry. CAP relies on effectively halving the 2.048Mbit/s channel data rate by splitting the data stream between two carriers which are $\pi/2$ out of phase with one another. BT believes that Westell's FlexCAP system can support 2Mbit/s transmission to the subscriber over 6km lengths of telephone grade copper pair. A newer adsl technology which increases the forward data rate to a theoretical 6.9Mbit/s and is based on a discrete multi-tone modulation (dmt) technology has also been evaluated by BT. This technology, pioneered by Aware, Westell and analogue-to-digital converter specialist Analog Devices, though technically superior to CAP, is too expensive and power thirsty for BT's application.

Dmt uses orthogonal frequency division modulation techniques and divides the signal between around 100 narrow carriers. The low data rates on each carrier reduce the dispersion in the cable and the use of Reed-Solomon forward error correction makes the dmt scheme more resilient to poor line quality and interference. Aware has developed a more efficient digital signal processing algorithm, which is claimed to reduce the cost of the dmt modulation scheme.

Analog Devices, which is already in prototype production of its dmt chipset, is currently incorporating the new dsp software, known as Wavelet technology, into its dsp-based adsl design. "Wavelet technology will significantly increase the bit rate of dmt and make it more cost effective", said a spokesman for Analog Devices. He likened the improvement in signal processing efficiency, with little loss of precision, to that achieved with a fast Fourier transform compared to the discrete Fourier transform. Westell is also developing an adsl system based on the new technology and the first Wavelet-based dmt systems should be available in 1995.

French propose voice identification as PIN replacement

Researchers and scientists at CEPT, the France Telecom and La Poste research centre, have embarked on a project of voice authentication, which could replace the PINs used in bank card transactions. Voice authentication not only offers better security, it can be used for remote verification typically performed over the phone or at an atm. Unique voice parameters are stored in the form of spectral coefficient and fundamental frequencies calculated for every 20ms of speech.

"We use evolution of the frequency and the spectral coefficient in the human voice," said Isabelle Milet, senior researcher at CEPT.

The system consists of a low pass filter and an a-to-d converter digitising on 8kHz to 14 bits on an ISA card designed by Acsys of France. It is installed in a pc also equipped with a Dialogic ISA card that performs a Fast Fourier Transform on the voice spectrum using a Motorola *DSP56000* chip.

The system's response time in the lab is less than one second. The product is at its prototype stage at present and CEPT is currently on the look-out for companies which would like to use this product. "For the moment we are not completely ready to sell the licence although we are showing the prototype. But we are looking for companies which are interested in it", said Milet.

Survey shows fast electronics growth

Electronics and computers was the fastest growing sector by a long way in the British engineering industry, according to the Engineering Employers' Federation (EEF) in its biannual survey of British engineering. This showed sales growing by 24 per cent over the past 18 months, measured at constant purchasing power.

The EEF says the sector is likely to continue its steady growth until the end of 1995, boosting sales by a further 14 per cent.

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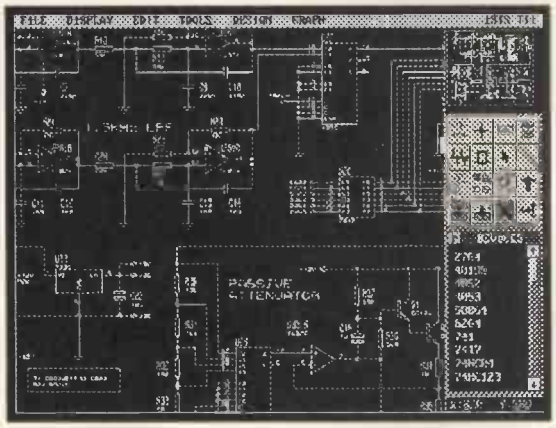

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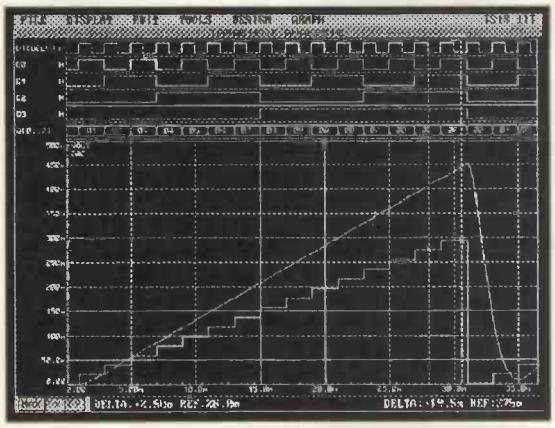

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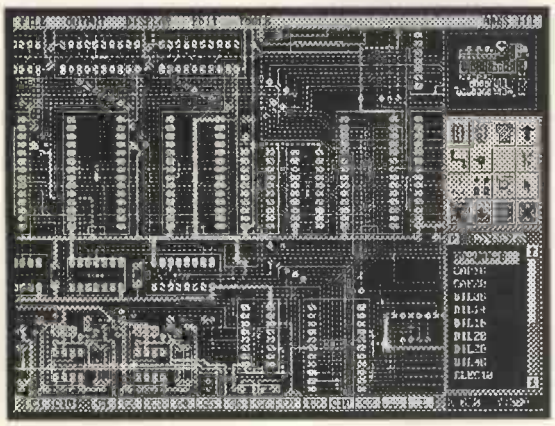

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CIRCLE NO. 104 ON REPLY CARD

Alternative to bar codes announced

Cambridge-based Scientific Generics has introduced a technology that is claimed to solve the main problems that relate to identification technology. There are two main methods currently in use: bar coding and electronic tags. Bar codes are cheap but need line of sight equipment to read and are not well suited to industrial use. Electronic tags (radio frequency devices) do not require line of sight technology but are too expensive for most applications.

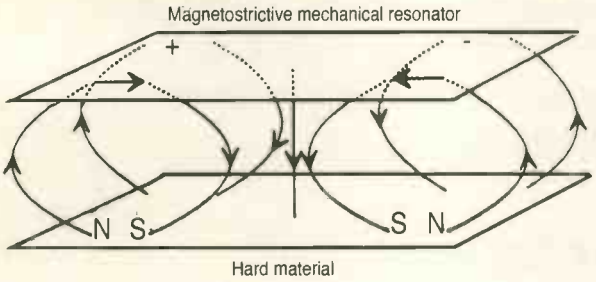
The Scientific Generics technology, Programmable Magnetic Resonance, like electronic tags, can be read without using line of sight scanning technology and irrespective of orientation of the scanned item. It involves tags with a hard magnetic strip and a mechanical resonator packaged together. These can be encoded at the point of issue, making it flexible to use. It is also claimed to be cheap to manufacture: each tag costs only a few pence rather than the $1 US typical of existing rf tags.

The tags are also suitable for industrial use. Providing the capsules containing the magnetic material are robust enough, the tags can be used over a wide temperature range as well as being able to withstand high pressure and adverse chemical environments.

In addition to the tag, the system consists of a conventional magnetic strip reader/writer and a remote reader that generates and detects alternating magnetic fields from a distance, using coils. Readers can be hand-held, or more sophisticated gate readers used for automatic sorting.

The tag's hard magnetic material can have a magnetic pattern written conventionally, like the strip on a credit card. This part is placed close to the mechanical resonator and sets up relatively high dc magnetic fields in the resonator. The resonator can resonate or ring at a number of well-defined frequencies, where frequency equals (resonance number × velocity of sound in material) divided by twice the length of the resonator. There is a characteristic spatial vibration pattern for each resonance.

Tagging via 'programmable magnetic resonance'. When set in motion by an ac electromagnetic field from the reader, the resonator rings at a number of programmable, well defined frequencies.



Hexfet power MOSFETS make technological leap

International Rectifier claims to have made the biggest advance in Hexfet power Mosfet technology since the original device was developed in 1979. Its Generation V range of products offer "lowest cost per Amp," claims Gene Sheridan, IR's director of strategic product marketing.

The next generation of IR's vertical dmos (VDMOS) process has enabled IR to slash power drain source on-resistance ($R_{DS\text{-}on}$) on the p-channel by up to 80 per cent, and reduce it by up to 50 per cent on the n-channel. This means the n-channel $R_{DS\text{-}on}$ level comes in at 8 milliohms and p-channel $R_{DS\text{-}on}$ levels will be at 30 milliohms early next year. The n-channel option is out first as it addresses the bigger volume users of the product. New p-channel products will be available this month.

The first series of products will offer ratings of 55V and below. The low-voltage sector is the largest in the market at 40 per cent. Its aims are automotive applications and 12V, portable and laptop computers.

International Rectifier will offer products in three packages including its new Micro 3 SOT-23 package which has a profile of less than 1.1mm. The portable 'phone market is one of its targets.

Protel plans Windows EDA client

San Jose developer Protel Technology is joining Intergraph in supporting Windows-based eda tools. Protel is proposing a client/server architecture for desktop eda and is developing a set of Windows-based eda tools which will run on existing client/server computer systems. This will enable eda tools to run in a shared environment and potentially opens the way for high performance eda tools from more than one vendor running under Windows.

Protel's EDA Client will enable a single pc-based client to run eda editors such as symbolic editors, special editors for Verilog, VHDL and pld/fpga design, and layout editors for pcb design. Servers will provide more complex functions such as analogue and digital simulators, design compilers and synthesis tools, design verification and analysis tools. For design vendors there is no need to include a graphical user interface on these server tools, as they will be accessed through the client system.

The user can not only run more complex eda tools in a cost effective Windows environment, but can also mix and match tools from a number of vendors. Cost savings could result when a single simulation or compiler tool licence can be accessed through a workgroup of clients although the eda vendors will probably be quick to plug any loophole.

Protel believes that increasing PC performance and the emergence of multi-tasking Windows and Windows NT environments opens the way for multi-user eda in a pc-based client-server environment. Intergraph is perhaps the only eda vendor committed so far to the view that the Windows NT environment will quickly start to replace the Unix environment as the main platform for both high and low cost eda. Earlier this year it introduced its Veribest family of chip and pcb design tools which run under Windows NT as well as Unix.

Jeff Edson, Intergraph's vice-president, predicts that the emergence of Windows 95 and Windows NT will threaten Unix as the main eda platform. "Unix will become an insignificant percentage of the market within two years; 20 per cent is insignificant," he added.

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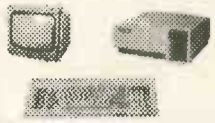

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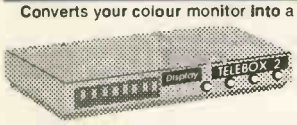

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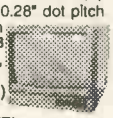

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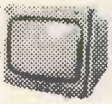

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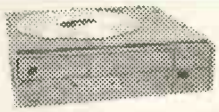

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Japanese propose chip production consortium

Fears for their future competitiveness in chipmaking have propelled the top Japanese semiconductor companies to propose an industry/government/academic consortium to develop production equipment for the Gigabit chip generation.

The move is a response to the successes of America's Sematech consortium, the European Jessi collaboration and the European and Taiwanese industry/university collaborations in developing advanced chip processes and equipment. Japan's top companies are looking for a similar arrangement to defray the heavy financial burden. Senior executives from ten leading Japanese chip firms, including NEC, Toshiba and Hitachi, met earlier this month to discuss the formation of the new collaboration.

Although Japan has an ongoing industry/government programme called Sortec which provides for shared R&D effort in advanced technology, this is thought to be the first collaborative effort to be focussed on chipmaking since the famous 'VLSI Project' of the early 80s which resulted in the Japanese putting a 256K dram into production a full year before the Americans. It also led directly to the exit of all US companies (except TI and Micron) from the dram business in the recession year of 1985 and to Japan's market share lead in chips after 1989.

Sematech, the American industry/government consortium was set up in 1988. It worked first on developing new processes and then in developing new process equipment. The US regained market share ascendancy in 1992 and has been increasing its lead ever since. In the meantime, Europe has formed a similar consortium called Jessi which developed the half micron process currently being used by Europe's leading chip companies, Taiwan's ERSO has delivered a half micron process to Taiwan's leading chip companies, and Korea's KIET has delivered world-competitive processes to Samsung, Hyundai and Lucky Goldstar.

The new Japanese consortium is expected to start its work in 1996. It will be aimed at developing cost-effective equipment to be used in chip-making from 2000AD onwards and manufacturing chips with feature sizes down to 0.18 micron.

DBS dish for computer data receipt

Hughes Network Systems is planning to introduce a direct broadcast satellite dish specifically for receiving computer data. The product will be called DirecPC and will be aimed at businesses who want to receive large amounts of computer data at T1 line speeds of about 1.5Mbit/s. According to Hughes, businesses will be able to receive large software updates directly from software vendors and they will also be able to download large multimedia files that would take hours using standard modem connections.
The DirecPC receiver will cost about $1,500 and carry a $16 monthly charge. Hughes hopes that DirecPC will follow the success of the DirecTV satellite digital TV broadcast system introduced earlier this year into the US with partners RCA-Thomson and Hubbard Broadcasting.

Integrated VR system for consumer market

IBM and UK virtual reality (vr) specialist Virtuality are expected to launch the world's first integrated vr system for the consumer market by 1996. The system will consist of IBM's hardware, including a head-mounted motion sensor, joystick, tracker and graphics cards, and Virtuality's software. "Individual packages and software like this already exist, but we want to integrate it and make it easier to use and more reliable," said Jan-Paul Boos, the new markets manager at IBM Europe.

The package will be sold either as a PC upgrade kit, priced around $5,000, or already integrated in a PC. Currently the two companies are delivering the more expensive version of this product, called Elysium, to beta sites. It costs between $10,000 and $70,000, depending on the model version, and is aimed mainly at companies and software designers who would like to develop software for their own 'full experience' VR model used for promotion or advertising.

"Elysium itself is not targeted at the consumer market, more in entertainment products for public use. But we are looking into dropping the price down for consumers to use it on their PCs," said Richard Holmes, the design director at Virtuality.

Possible European standard for digital TV access

Moves by British broadcasters to establish a conditional access (CA) format for digital TV services could broaden into a new European CA standard. Major terrestrial broadcasters from across Europe have discussed the development of a common digital CA standard in a meeting of the European Broadcasting Union.

The BBC, ITV companies and Channel 4 decided to work on their own CA system in September after the DVB, Europe's digital broadcasting standards body, failed to agree to a compulsory common CA system. Terrestrial broadcasters believe that the compromise that resulted from that meeting and is based on a code of practice, is unworkable.

The DVB has already agreed a non-compulsory common interface for CA systems. Sources close to the latest talks say they could result in a common Europe-wide CA format to go with this interface.

Multiple advances in disc drives

Advanced disc-based storage systems have been unveiled by 3M, Toshiba and Hitachi. 3M has shown a high-capacity CD-ROM disc that doubles the storage capacity of existing high density CD-ROM discs through the use of an innovative dual layer technology. The disc has a second layer that can be accessed by a CD-ROM drive read head by shifting the angle of reflected laser light as it shines through the top layer. 3M says its discs can store as much as 6Gbytes – ten times the current 600Mbyte capacity of cd discs. The disc was developed with help from Philips and Sony.

Toshiba, like Panasonic, is to sell rewritable optical disc systems based on phase change technology. The company has developed a double-sided 3.5 inch disc that can store 1.3Gigabytes. First products are expected towards the end of 1995.

Toshiba and Hitachi have both announced 2.5 inch magnetic hard disc drives with capacities of over a gigabyte. The drives are due on the market early next year. ■

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RESEARCH NOTES

Jonathan Campbell

Optical ether could become reality

Communication completely in the optical domain across the whole of the UK, in a network implemented on a step by step basis, could become a practical reality if proposals by two researchers in the department of electronics and electrical engineering, University College, London, succeed. Using the principle of an optical 'ether', they have developed (*IEE Proc Optoelectronics*, Vol 141, No 5) a novel method that will allow a network to be constructed piece-meal yet have the capacity to service 25 million customers with over 3Mbit/s capacity each.

There has already been much speculation about building a network, national or global, in which the complete transmission path is in the optical domain.

But up to now proposed methods have relied on simultaneous implementation across the whole network, with the ultimate number of nodes in the network set at the beginning of the project.

M Sabry and JE Midwinter's idea is to exploit unused bandwidth in existing optical fibres to develop an optically routed broadband network that could evolve 'gracefully' from today's point-to-point transmission, on some routes only, to an extensive multi-node network.

An inner core would be a highly interconnected mesh, with each of the nodes providing routing within the mesh as well as access to one or more ring-type networks making up an outer core. These in turn would be connected to end users via pons (passive optical networks).

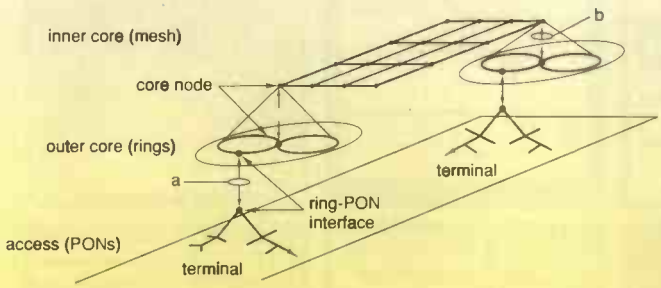
First step in building the network would be to replace present optical fibre regenerators with erbium-doped fibre amplifiers (edfas). This would remove the need for rigid data formats imposed by regenerators and would turn the fibre links into transparent lossless data pipes.

Existing transmission links that would form the inner core would then be upgraded to use dense-wavelength division multiplex (d-wdm) transmission, but with each carrier operating to standard interfaces. Later these interfaces would be replaced by transparent wavelength switches, which, once rings are added, would allow transmission between rings entirely in the optical domain.

The achitecture allows the network to be developed from the top down using a common set of components throughout.

"As our proposed evolution of the whole network relies on a single modular set of building blocks, the development can stop at any stage", say the researchers. It could just as easily restart and be used to interface to the existing public switched network as necessary.

Fig. 1. Hierarchical view of the proposed optical network architecture.
a) represents ATM cell injection and switching;
b) ring and mesh fibres have equal status for subframe switching.



Giant steps for memory boosts

Giant magnetoresistance (gmr) could be the key to a new generation of fast ram devices and permanent magnetic random access memory (mram) chips.

Many fundamental problems still need to be solved. But Barthélémy, Fert, Morel and Steren at the Laboratoire de Physique des Solides, Université Paris-Sud, France – one of the original groups involved in identifying giant magnetoresistance – believe that such breakthroughs should be possible and that gmr holds the key to advances in information storage and logic operations (*Physics World*, November, 1994, pp. 34-38).

In giant magnetoresistance, the resistivity of a layered structure of alternate magnetic and non-magnetic materials dramatically changes when a sufficiently high magnetic field is applied. The effect seems to be because the magnetisations in the alternate layers are anti-parallel.

At zero field the resistivity is at a maximum, but in high fields the magnetisations become parallel to the applied field and resistivity drops startlingly.

Mram chips would be constructed by cutting an array of magnetic plots into the top layer of a three-layer structure. Resistivity of each plot would depend on whether its magnetisation was parallel or antiparallel to the reference.

Information could be written by flipping the magnetisation of the plots – by running an electrical current in a nearby wire – and information could be read by sensing the resistivity of each plate. But the memory would be permanent because the magnetic plots would need no energy to retain their orientations. According to the group, such very high density memories will be feasible within a few years.

The new fast ram devices would make use of a development that adds magnetic clusters into a non-magnetic-layer/magnetic layer structure. While the dynamics of magnetic processes in multi-layers are governed by the relatively slow motion of domain walls, reversal of magnetisation in a small particle can be very fast – possibly as rapid as 10^{-11} or 10^{-12}s. This speed, combined with gmr sensitivity, would be the main attraction of the new devices.

Jamming could undermine GPS

Low-power rf jamming could wreak havoc with commercial global positioning satellite (gps) systems, according to researchers at Georgia Tech. As a result we must be much more careful about how gps is used, they warn.

GPS technology is increasingly becoming part of daily life, giving position or velocity information for a large number of uses.

So the findings by Georgia Tech's John Daher, Joseph Harris and Mark Wheeler that commercial gps receivers show worryingly large susceptibilities to in-band interference by anyone with unfriendly intent (*IEEE AES Systems Magazine*, October, 1994, pp.21-25) could have wide implications.

Two commercial receivers were tested in detail for the study: both exhibited low susceptibility thresholds to in-band continuous wave signals, and both could also be over-driven with an out-of-band signal.

The Georgia researchers also assessed just what it would take to jam a receiver. Using measured in-band susceptibility levels of –112dBm for threshold, –103dBm for las ('lost all satellites') and a typical gps antenna gain of 3dBi showed that a gps receiver could be vulnerable to jamming from low power transmitters tens of km away.

Fortunately the receivers showed good rejection of out-of-band and pulsed interference signals, so unintentional interference is much less likely than the intentional jamming threat.

But the researchers say that, based on their measured gps receiver vulnerabilities, careful consideration must now be given to the applications for which gps is chosen, particularly as the technology has been considered as a replacement for the current instrument landing system (ILS) and future microwave landing system (MLS), putting it in a safety-critical position.

Even where commercial use was not safety-critical, inaccurate data could still be very costly for users.

Interference mitigation techniques must be be investigated to improve system reliability, while interference rejecting antenna techniques such as the military controlled response antenna (crpa) should be evaluated for commercial applications, say the researchers.

Meanwhile, they say, signal processing and filtering techniques should be incorporated into commercial receivers.

Jamming of GPS – which is not too difficult – could cause significant problems.

● Also at Georgia Tech, software that scans research databases and automatically builds up profiles of activity in particular technical areas has been developed.

Toak software can analyse databases – such as the international electronics database, Inspec; the engineering, computer, and business indexes, and US Patents – and show how much an emerging technology is linked with various applications such as checking whether a new plastic is being used in electronics packaging or aerospace. Toak can also map relationships between sectors, identifying component technologies that contribute to advances in a target area, or can be used to compare research and development activity by state or nation. Such an analysis recently showed that Japan and the US are going head-to-head across most electronics assembly technologies.

Border controls that cross the line?

Robot search planes equipped with laser scanners and advanced electronic systems able to track large numbers of 'infiltrators' sounds like a scene from *The Terminator*. In fact it is the language and emergent technology of US immigration control.

Over 1000 people a day are drawn to enter the US illegally, many dodging border patrols and crossing from Mexico into the San Diego region of California

But border guards could soon be the last thing on the mind of frontier-crossing Mexicans if Bill Wattenburg, a senior research scientist at the University Foundation, California State University, is right.

He makes a case (*Science*, Vol 265, pp.1184-1185) for dusting all popular border-hopping routes with a material that fluoresces when illuminated with laser or uv light.

Using fdi – fluorescent tagging of infiltrators – Wattenburg, who has been an enthusiast of fdi for 25 years, says that anything touching the fluorescent material could be easily identified.

Amongst the many systems that could be used for tracking the illegal immigrants, Wattenburg likes the look of a portable system developed by Lawrence Livermore National Laboratory to track the trajectory of a rifle bullet. Fitted to a robot plane and used with a laser scanner, Wattenburg says that the system could be used to monitor a strip at least 10km long and keep track of even 100 moving 'targets'.

He estimates the cost of a six-month fairly-low tech experiment based around San Diego as $500,000. Compared with figures he reproduces for the drain on the state of illegal immigrants – "...while needy families of legal immigrants and native born alike will be denied the full government assistance they require to better their lives" – that sounds like a bargain. If the system could be automated, then no-one, it seems, need get their hands dirty.

Microwave sticking point

Knowing to the second when an adhesive has hardened is not easy, but it can be vital for efficient and safe production of glued structures and components. The problem is that production time is wasted through leaving a joint to cure for longer than is necessary, while trying to handle a joint too early can result in even more serious consequences.

But Abayomi Olusanya and colleagues at NPL, Teddington, are developing a technique to monitor the dielectric permittivity of curing adhesives that should indicate precisely when the bond has formed.

Unlike dielectric methods that have been tried before, the NPL technique uses microwave frequency, so removing much of the uncertainty that has been associated with the approach.

During the cure of thermosetting adhesives, cross-linking of molecules takes place. But initiators taking part in the reaction can cause confusing conductivity effects which have upset previous methods.

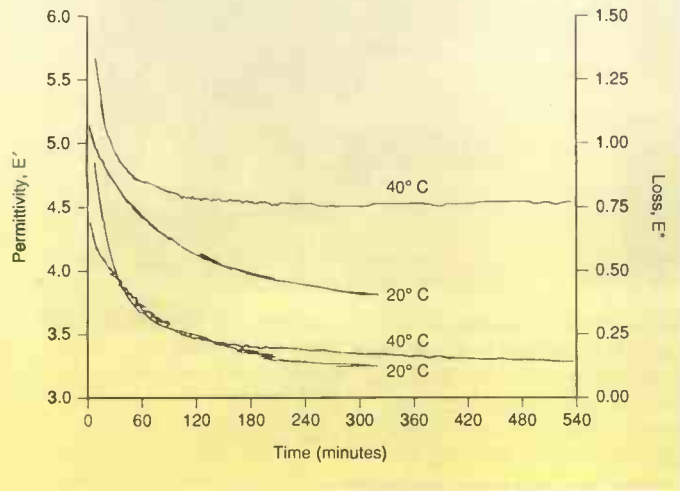
Using microwave frequencies effectively 'freezes' the initiator ions, removing them from the signal, while electrode polarisation problems are avoided too.

NPL says the method is particularly suited to applications where access to the bonded areas is restricted. Only one surface needs to be accessible and a use being envisaged in the early trials is in monitoring adhesive joints in bridges. On site, a large area would be coated with adhesive, with a small area left exposed to fit the electrode.

The method could also be suitable for measuring the degree of mixing in two-part adhesives, if the overall dielectric properties can be related to the individual dielectric properties of the constituent components.

Dynamic dielectric properties of adhesives during cure are obtained through analysis of the adhesives' reflection coefficient using a reflectometric coaxial sensor. At present, NPL is studying how changes in permittivity relate to mechanical properties.

Dielectric permittivity can be used to indicate precisely when an adhesive has set. These curves are for bridge adhesive SKIA PBA 31 LT cured at 20°C and 40°C.



Microscopic improvement is good news for chips

A US applied physicist has reported success in using an atomic force microscope (afm) to half the size of semiconductor gates – opening up the possibility of much faster transistors.

Cal Quate of Stanford University has been using the afm – up to now used mostly as an imaging tool – to create silicon dioxide masks less than 100nm wide on semiconductor chips.

Conventionally, the afm, which has an extremely fine tip, is used to trace surface topography on an atomic scale. But Quate has used it to apply a current to a substrate coated with amorphous silicon, forming a line of silicon dioxide which protects the underlying silicon during etching.

A series of mosfets was produced by normal manufacturing methods, with the afm used to make the gate.

Reports are (*Science*, Vol 266, p.543) that Quate and his team fabricated several devices with gates between 0.7-0.2μm. Transistors are said to have behaved much the same as conventional mosfets.

What has generated so much excitement is that afm techniques are in their infancy and can only get better, while conventional lithographic methods are pretty near their limits.

Voltage follower that gives a lead

A voltage follower that promises much higher chip performance – irrespective of which semiconductor technology is used – has been designed by W J Su and supervisor F J Lidgey at the School of Engineering, Oxford Brookes University, Oxford.

With conventional buffers, compromises have to be made between high speed and high accuracy. But by using an existing op amp negative feedback approach augmented by a feedforward technique to make the op amp float, the researchers look to have avoided the usual speed-accuracy trade-off (*Electronics Letters*, Vol 30, No. 22).

System design allows the input signal to propagate either via the op amp, with 100% nfb used to correct the difference between input and output, or through the additional feed-forward circuits – the source/emitter followers. The result is to float the op amp with all its nodes at virtually the same signal voltage as the input. So capacitances within the op amp have little effect on overall frequency response and a large bandwidth can be achieved.

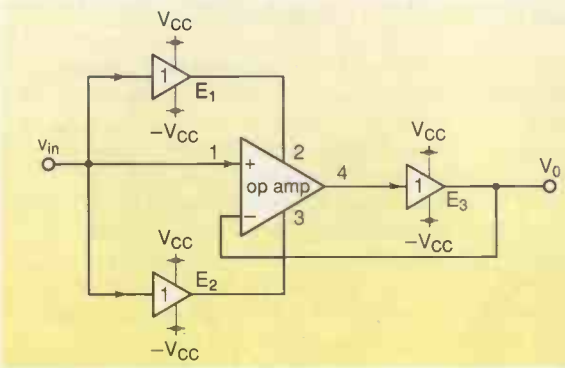
Near zero response delay time of the floating op amp to an input also means the main speed limitations of conventional followers are overcome.

The researchers say that simulations are demonstrating that more than 1300V/μs slew rate and 60ns settling time can be obtained relatively easily. At the same time, bandwidth is increased, from around 35MHz in conventional feedback buffers, to approximately 220MHz with a phase margin of 53°. Improved accuracy comes from the enhanced gain and faster speed of the core floating amplifier.

Some additional circuitry has had to be used, reducing the input and output voltage headroom, and the power supply current is higher. But the researchers insists that these cost penalties are: "relatively minor compared with the substantial benefits that the circuit exhibits".

The work is being supported by Analog Devices and the main application of the design is thought to be monolithic integrated circuit voltage followers. ■

Combined negative feedback buffer augmented with two source/emitter followers speeds performance.



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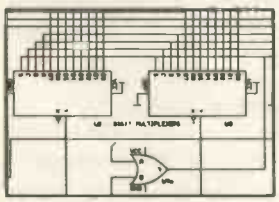

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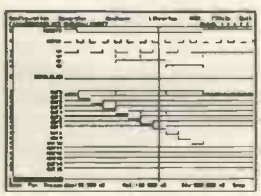

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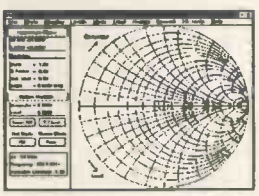

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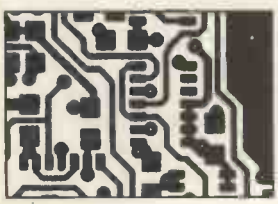

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Ironing out DISTORTION

As is often the case with articles on audio subjects, Douglas Self's* recent series on amplifier distortion caused a great deal of interest worldwide. Building on Doug's work, Edward Cherry offers an in-depth look at distortion, and discusses how to reduce it.

At the outset I should state that there appears to be something of a philosophical difference between our approaches to distortion inside the feedback loop. There is nothing wrong with tackling the distortion of various stages on an individual basis, but my approach to designing a high-quality amplifier is to choose a simple topology based on common-emitter amplifying stages and apply negative feedback to reduce distortion. Variations in circuit topology (other than push-pull operation) rarely give better than a ten-fold reduction in distortion on a production basis; feedback, however, can reduce distortion almost indefinitely[1].

The beautiful thing about feedback is that it reduces all distortions simultaneously. If there is enough feedback to fix the major sources of distortion, the minor sources will be taken care of automatically. However, as Self points out, feedback cannot correct distortions arising outside the feedback loop.

Common-emitter stages have theoretical advantages over common-collector amplifiers[3] and, in my opinion, have important practical advantages too. The theoretical basis for my positions regarding both feedback and common emitter stages rests ultimately on the work of Bode[4], who points out that a common-collector stage can be considered as a common emitter stage with a kind of local feedback, which rarely accomplishes anything for the stage within the local loop and usually makes matters worse for the stages outside it. In part, the complex behaviour of distortion in Self's amplifiers is attributable to the local feedback.

Incidentally, I take it as now being universally accepted that there is no basis for linking transient, interface and phase intermodulation distortions to large amount of feedback.

This commentary addresses audible distortions only; that is, nonlinearities which generate distortion products in the audible frequency range 20Hz to 20kHz. It is not concerned with nonlinearities that generate ultrasonic distortions, for I am not in the business of trying to '...please any passing bat...'![2]

Distortion products can arise as harmonics of a single input frequency, or from intermodulation between two or more simultaneous inputs, in which case the distortion products lie above and/or below the input frequencies. Ultrasonic distortion in an amplifier may, of course, be an indicator of trouble at audible frequencies, but not necessarily; what matters is the presence or absence of audible distortion products.

Indeed, I believe the 20kHz upper limit should be reduced, because practically no-one can hear distortions at even 15kHz. As an easy demonstration of this assertion, readers could compare the sounds of a 5kHz square wave with an accurate 1:1 mark-space ratio and a 5kHz sine wave of 1.273 times the peak-peak amplitude. The square wave contains a 5kHz component of the same amplitude as the sine wave, but it also contains a component at 15kHz of one-third the amplitude. Shift the frequency up or down to find your own frequency limit for audible distortion. Be honest, and remember that this is the equivalent of 33% third-harmonic distortion!

Distortion analysed

Figure 1 is Self's Fig. 1a from *EW&WW* of August 1993, with some minor changes. I analysed it in May 1982[5] and a main purpose of this commentary is to point out that many of Self's conclusions have a rigorous mathematical basis. In addition, given that an audio amplifier is to be of this basic topology (and it would not be my first choice), then the simple changes shown in **Fig. 2** give substantial improvement at little cost. Figures 1 and 2 can, of course, be flipped upside down, with n-p-n and p-n-p transistors interchanged; the analysis is identical. It is actually the flipped version that is considered in Reference 5.

Nonlinearity, sensitivity and distortion

Distortion can be considered as variation of incremental gain from point to point on the signal waveform. In other words distortion is caused by nonlinearity in parameters like ß and g_m. For example, when an amplifier is driven near the point of clipping, its incremental gain falls at the waveform peaks; these

**Distortion in power amplifiers. D Self. August, 1993 to March, 1994.*

Definition of terms

The analysis in Reference 5 is in terms of the effective values of ß and g_m for each stage.

Current amplification factor ß of a transistor is defined formally as the gradient of a graph of collector current I_C versus base current I_B. Nonlinearity in ß is any departure of the graph from a straight line, from any physical mechanism whatsoever. In practice there are many such mechanisms, and the magnitude of ß falls at both large and small currents.

Mutual conductance g_m is similarly defined as the gradient of a graph of collector current I_C versus base-emitter voltage V_{BE}. Nonlinearity in g_m is any departure of the graph from a straight line. One physical mechanism for such a departure is the exponential I_C *versus* V_{BE} characteristic, inherent in bipolar junction transistors and which results in the well-known formula

$$g_m \approx qI_C / kT \qquad (1)$$

where kT/q is approximately 25mV at room temperature.

Although these definitions of ß and g_m are most often applied to an intrinsic transistor (a transistor from which parasitic elements such as base-spreading resistance r_B have been removed), basic formulae like Eq. 1 can be adapted to a complete transistor:

$$g_{m(\text{eff})} \Rightarrow \frac{1}{r_B/\beta + kT/qI_C} \qquad (2)$$

For the Darlington transistor in **Fig. 3**,

$$g_{m(\text{eff})} \Rightarrow \frac{1}{[r_{B(a)}/\beta_{(a)} + kT/qI_{C(a)} + r_{B(b)}]/\beta_{(b)} + kT/qI_{C(b)}} \qquad (3)$$

where the (a) and (b) subscripts identify parameters of the individual members. The formulae can even be adapted to include local emitter degeneration

$$g_{m(\text{eff})} \Rightarrow \frac{1}{kT/qI_C + R_E} \qquad (4)$$

or a resistance R_S between base and emitter:

$$\frac{1}{\beta_{\text{eff}}} = \frac{1}{i_C/i_S} = \frac{1}{\beta} + \frac{R_E}{R_S(1 + qI_CR_E/kT)} \qquad (5)$$

neglecting r_B for simplicity. Equation 5 is a particularly useful analytical trick, since it allows the output resistance of one transistor or stage (a kind of source resistance) to be incorporated into the effective ß for the next.

In all such cases, 'base' and 'collector' currents and 'base-emitter' voltage are measured at the effective terminals of the device, and include the currents in shunt resistances or the voltage drops across series resistances. Nonlinearity in effective g_m then includes any nonlinear component of voltage drop across r_B, and therefore involves the nonlinearity in ß; note that ß occurs in Eqs. 2 and 3.

peaks are amplified less than the rest of the input waveform, and the output is 'squashed'.

Sensitivity is the ratio of a percentage change in some parameter like ß or g_m to the resulting percentage change in incremental gain. At any point on a signal waveform the instantaneous voltage and current in, say, the output transistors of an amplifier can be found. Hence the fall in, say, ß at the signal peaks can found from the known nonlinearity. Then, if sensitivity to changes in ß is known, the gain compression can be calculated and ultimately distortion can be predicted quantitatively. For example, if the gain compressions at the positive and negative peaks of a signal waveform are γ' and γ'' respectively, the second and third harmonic distortions are:

$$D_2 \approx \frac{\gamma' - \gamma''}{8} \qquad (12a)$$

$$D_3 \approx \frac{\gamma' + \gamma''}{24} \qquad (12b)$$

Figure 4, reproduced from Reference 5, shows the sensitivity of the overall gain of Fig. 1 to changes in parameters, as functions of frequency on logarithmic scales. Many of the labelled points have the physical significance of quantities like mid-band loop gain considered above. Numerical values are calculated for,

- $ß_2$ = 100 (typical);
- $ß_3$ = 5000 (a typical Darlington);
- g_{m1} = 4mA/V (typical for bjts operating at a few hundred microampères without emitter

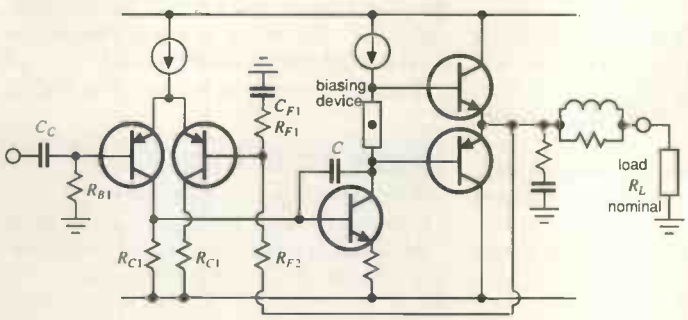

Fig.1. Outline circuit of many audio power amplifiers. This was Self's Fig. 1a.

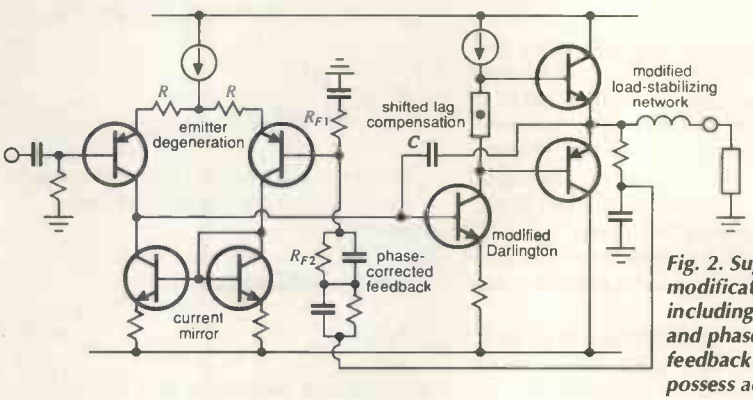

Fig. 2. Suggested modifications of Fig. 1, including a current-mirror and phase correction in the feedback loop are shown to possess advantages.

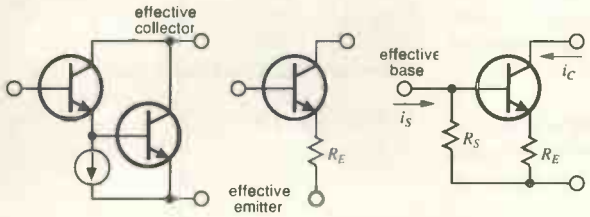

Fig. 3. Some examples of sub-circuits for which effective ß and gm can be defined: Darlington, emitter degeneration, and degeneration plus shunt resistance.

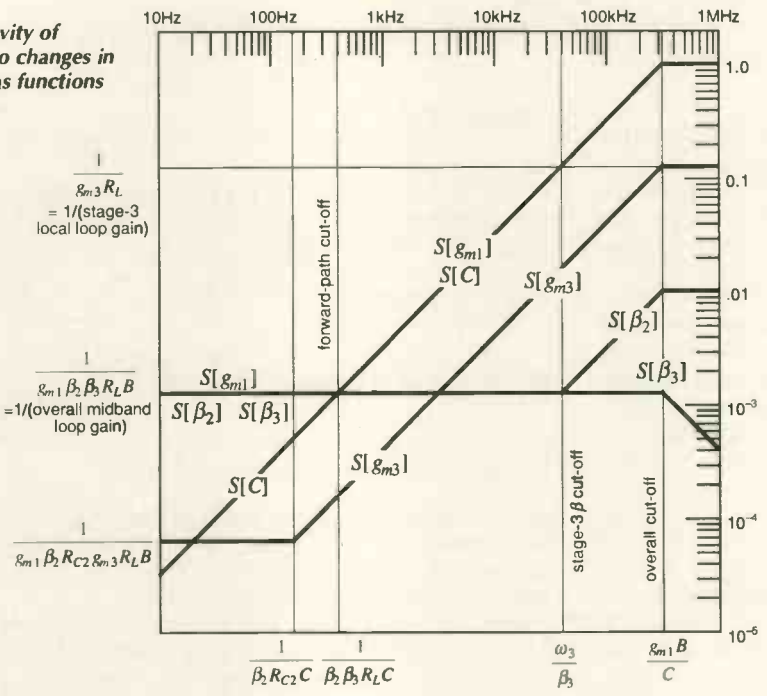

Fig. 4. Sensitivity of overall gain to changes in parameters, as functions of frequency.

degeneration, or at larger currents with degeneration);
– g_{m2} = 10mA/V (corresponding to about 68Ω in the second-stage emitter as part of the protection circuitry, Eq. 4);
– g_{m3} = 1A/V (corresponding to about 0.68Ω ballast in each of the third-stage emitters, Eq. 4);
–ω_3 = 1Gr/s (typical for a Darlington which consists of a reasonably fast first member and a slow second member);
– C = 100pF (to set the overall 3dB bandwidth at 300kHz, Eq. 8);
– B = 0.05 (corresponding to an overall mid-band gain around 20, perhaps R_{F1} = 2k2Ω and R_{F2} = 47kΩ);
– R_{C2} = 100kΩ (a guess, but it hardly affects the results)
– R_L = 8Ω nominal load.

First stage

As stated by Self, signal amplitude in the first stage increases in proportion to frequency above the forward-path cut-off $1/ß_2ß_3RLC$ where overall loop gain falls away. The nonlinearity of g_{m1} is therefore more strongly exercised as the frequency increases. Simultaneously, sensitivity to changes in g_{m1} increases with frequency as shown in Fig. 4. Therefore, overall distortion rises with frequency, either as the square or cube, depending on details.

In my opinion, Self's discussion of input stages is over-kill. Despite the rapid increase of distortion with frequency, the simple long-tailed pair with emitter degeneration shown in **Fig. 5** contributes vanishingly small audible distortions. Fancy topologies are simply not required.

Emitter degeneration. If a feedback amplifier is fed with a fast-rise mid-frequency square wave, the peak-to-peak input to the first stage is twice as large as the square wave itself. Therefore, if an amplifier is not to go into slew-rate limiting (alternatively, is not to generate hard transient intermodulation distortion) when fed with a full-amplitude fast-rise square wave, its input stage must be designed not to clip on a signal twice the amplitude of rated mid-band sinusoidal input to the complete amplifier[6]. This result is independent of the overall bandwidth and slewing rate.

Taking the numerical values in Fig. 5 as an example, g_m is about 4mA/V and maximum output is about 8mApk-pk; the stage clips at about 2Vpk-pk input. Therefore, this circuit should not be used as first stage in an amplifier rated at more than 1V pk-pk input (350mV rms) for full output. Suppose this input is at 6.67kHz, so that the third harmonic is at 20kHz (our upper limit for audible distortion), and suppose that other parameters are as in Fig. 4. Then:

– Overall loop gain at 6.67kHz is 46.
– Therefore, the differential component of the input to the first stage is overall input/loop gain, which is 22mVpk-pk.
– Therefore, signal current in each transistor is 1/2×g_{m1}×(differential input) = 47μApk-pk.
– Compression of incremental gain at this peak current is 0.0075%, found from the formula for effective g_m of a degenerated long-tailed pair:

$$g_{m(\text{eff})} \Rightarrow \frac{2}{kT/qI_{C(\text{left})} + kT/qI_{C(\text{right})} + 2R_E} \quad (13)$$

– But, from Fig. 4, sensitivity towards changes in g_{m1} at 6.67kHz is 0.022.
– Hence the (equal) compressions γ' and γ'' of overall gain at the signal peaks = sensitivity * first-stage compression = 0.00016%.
– Hence overall third harmonic at 20kHz associated with first-stage nonlinearity is 0.000013%, from Eq. 12b. This is at least a factor of ten smaller than the wildest suggestion I have ever seen as a target figure for an 'ideal' amplifier.

An analytical approach allows one to classify for certain the nonlinearities in an ampli-

Basic equations

With the notation shown in the Definitions panel, the main features of the small-signal response of Figs 1 and 2 are determined by just four components:

– the overall feedback resistors R_{F1} and R_{F2} via the overall feedback factor B:

$$B = \frac{R_{F1}}{R_{F1} + R_{F2}} \quad (6)$$

– the first-stage mutual conductance g_{m1}
– the second-stage lag-compensating capacitor C.

Overall mid-band gain

$$A_{\text{mid}} \approx \frac{1}{B} \quad (7)$$

Overall high-frequency 3dB cut-off

$$\omega_{3\text{dB}} \approx \frac{g_{m1}B}{C} \quad (8)$$

Forward-path mid-band gain

$$G_0 \approx g_{m1}\beta_2\beta_3R_L \quad (9)$$

Forward-path high-frequency 3dB cut-off

$$\omega_0 \approx \frac{1}{\beta_2\beta_3R_LC} \quad (10)$$

Mid-band loop gain

$$A_L \approx g_{m1}\beta_2\beta_3R_LB \quad (11)$$

where,
– $ß_1$ (not used here), $ß_2$ and $ß_3$ are the effective current amplification factors of the transistors, including the effect of any series or shunt resistors, in the first, second and third stages respectively;
– g_{m1}, g_{m2} and g_{m3} are the effective mutual conductances of the transistors (including the effect of resistors);
–ω_1, ω_2 (neither actually used here, but used in the JAES paper) and ω_3 are the projected gain-bandwidth products of the transistors;
– for later use, R_{C2} is the equivalent resistance (ideally infinite) of the second-stage current-source load.
The substance of these results is the same as given by Self.

fier as significant or insignificant contributors to audible distortion. Repeating the above calculation using a 20kHz input (the popular 20kHz thd test beloved of spec. men), gives the third harmonic as 0.0004% – not nearly so impressive, and even casting doubt on the intermodulation performance with real programme material. But this harmonic is, of course, at 60kHz and of itself has nothing to do with audible distortions.

In one of the more savage forms of the IEC total-difference-frequency intermodulation test, the input consists of two equal-amplitude sine waves at approximately 10kHz and 15kHz. For Fig. 5, the total of the audible intermodulation products near 5kHz is 0.000008%; the inaudible products near 25, 30, 35, 40 and 45kHz are larger, but it is the audible distortion that matters. To chase anything better than Fig. 5 would be folly.

The theoretical requirement that a complete amplifier should be able to accept a full-amplitude fast-rise square-wave input is unnecessarily severe; real programme material (even the output from a digital synthesiser) is subject to some form of band limiting. Typically I relax the requirement by about a factor of two. This increases 20kHz third-harmonic distortion of 6.67kHz by four, to 0.00005%.

However, I do regard some form of emitter degeneration as mandatory in a bjt first stage (perhaps not with fets). If there is none, a bjt long-tailed pair clips at about 100mV pk-pk input. Ideally, therefore, an amplifier that uses an undegenerated first stage should be designed to operate with no more than 50mVpk-pk or 18mV rms input. Self's Fig.1a of August 1993 would certainly generate hard transient intermodulation distortion in the square-wave test.

Including adequate emitter degeneration in the first stage carries the penalty of a slight increase in noise. The numerical situation for Fig. 5 is confused because total noise is dominated by the current mirror. However, if mirror noise could be eliminated (it can, but not in an amplifier of the Fig. 1 type), then the thermal and shot noise noise referred to the input of Fig. 5 would be 5.6nV/√Hz. After removing the emitter degeneration and adjusting the quiescent current to give the same gain, the noise drops about 1dB to 5.0nV/√Hz.

Current mirror. Self correctly points out that a current mirror in the first stage, rather than a simple resistance load: doubles the first-stage gain g_{m1} and therefore the overall feedback if nothing else is changed; doubles the available output current and hence the slewing rate; and improves the common-mode rejection.

Far more importantly, it raises the source impedance seen by the second stage. I shall show that distortion associated with output-stage ß nonlinearity is inversely proportional to effective ß of the second stage. Raising the source resistance for the second stage by using a current mirror in the first stage has the potential for increasing $ß_2$ (Eq. 5) and reducing this distortion by orders of magnitude.

Increased first-stage gain is a mixed blessing because it may provoke high-frequency instability. Self's suggested remedies are to double the first-stage emitter degeneration resistors, which is wasteful if these are already adequate, and it increases the noise; and to double the compensating capacitor C, which loses the slewing-rate improvement.

However, there is a third solution, which I strongly recommend: halve the value overall feedback factor B (halve R_{F1}, for example), thereby doubling the overall mid-band gain and halving the signal input voltage required to produce full output from the amplifier. This has two advantages in that it halves the common-mode voltage present in the first stage, and thereby halves a second-harmonic distortion mechanism associated with finite common-mode rejection; and it reduces the likelihood of clipping in the first stage, thereby reducing the incidence of hard transient intermodulation distortion.

Halving the input voltage required for full output, from something like the typical 0.6-1.0V of modern transistor amplifiers to 0.3-0.5V, makes the transistor amplifier more like the earlier Leak/Mullard/Quad vacuum-tube amplifiers. Why was this 'standard' ever changed? My best amplifiers are designed with 300mV sensitivity.

Second stage

Distortion in the second stage originates from three quite distinct types of nonlinearity:

– distortion associated with variation of effective $ß_2$ from point to point on the signal waveform, as the instantaneous current and voltage in the transistor change;
– distortion associated with similar variation of effective g_{m2} (this turns out to be very small);
– distortion associated with variation of the collector-base capacitance.

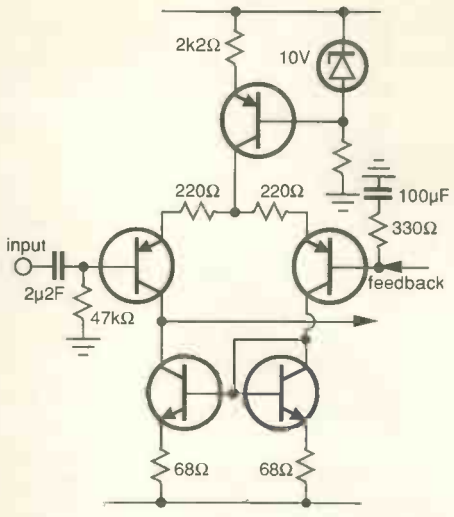

Fig. 5. First stage, with emitter degeneration, current source and current mirror, and with typical numerical values as used by the author.

Distortion associated with $ß_2$. Nonlinearity in $ß_2$ models the changes in current gain at high and low collector voltages and currents. Figure 4 shows that sensitivity to these changes is constant over most of the audible band of frequencies, but increases somewhere near the top of the band at the ß cut-off frequency $ω_3/ß_3$ of the output transistors (40kHz for the assumed data). However, sensitivity is inversely proportional to $ß_2$; distortion from this nonlinearity can be reduced simply by making $ß_2$ large – for example, by using a Darlington. Sensitivity and distortion are not affected by the choice of lag compensation C, and hence are independent of both the forward-path and overall high-frequency cut-off.

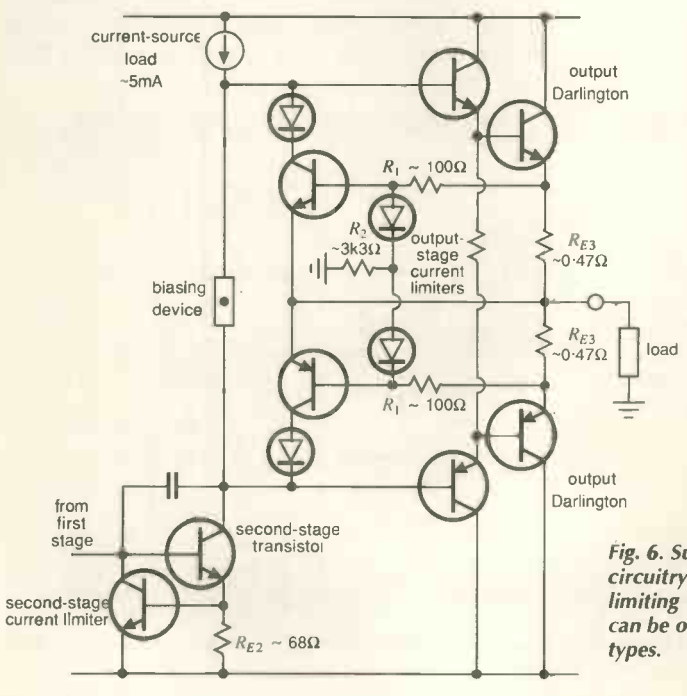

Fig. 6. Suggested protection circuitry for Fig. 1. Current-limiting transistors and diodes can be ordinary small-signal types.

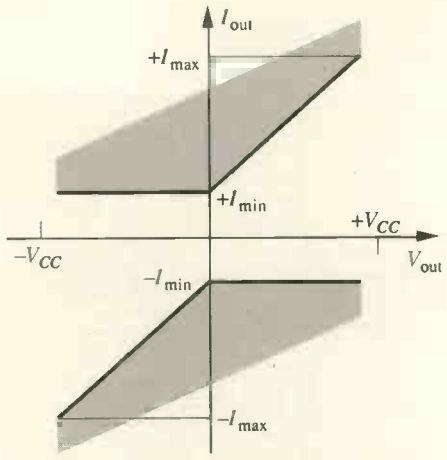

Fig. 7. Combinations of output voltage and current available with nonlinear fold-back protection of the output stage. For a nominal 50W, 8Ω amplifier, I_{min} should be around 1.5A and I_{max} around 6A.

Distortion associated with g_{m2}. Nonlinearity in g_{m2} models the exponential I_C versus V_{BE} characteristic intrinsic to a bjt, and the voltage drop across r_B. Sensitivity to changes in g_{m2} is extremely small (it is not even shown in Fig. 4) so the distortion associated with this nonlinearity is small too. Additionally, g_{m2} does not appear in the sensitivities of the other parameters. Therefore, emitter degeneration in the second stage, which reduces effective g_{m2}, has no effect on overall distortion and might at first appear pointless.

However, emitter degeneration can improve high-frequency stability. A significant amount of such degeneration is normally provided inadvertently, as part of the current-limiting protection circuitry. Second-stage degeneration therefore costs nothing in components, it does no harm, and it may do some incidental good.

Collector-base capacitance. Collector-base capacitance c_{CB} of the second-stage transistor is basically in parallel with the lag-compensating capacitor C and adds to its value. Collector-base capacitance is inevitably nonlinear, and has something like an inverse-square-root dependence on collector voltage.

Figure 4 shows that sensitivity to changes in C (hence c_{CB}) increases in proportion to frequency over the whole of the amplifier passband, and reaches unity at ω_{3dB}. The only way of reducing sensitivity towards C while retaining the basic amplifier topology is to increase overall cut-off frequency. Contrary to intuition, it does not help to use a larger value of C (the idea being that the nonlinear transistor capacitance would represent a smaller part of the total), nor does it help to increase ß (by using a Darlington); it does help to use a cascode for the second stage (Self's Fig. 4d of October 1993) or his modified Darlington (Fig. 4c same ref). Either removes signal voltage from the collector of the first member.

Output stage

In a push-pull class-B stage, the values of $ß_3$ and g_{m3} for the n-p-n and p-n-p transistors individually apply well into each half of the signal waveform, where only one transistor is conducting. In the overlap region near the middle where both transistors are conducting, or in a class-A stage, the values of $ß_3$ and g_{m3} are appropriately-defined averages.

Distortion associated with $ß_3$. Sensitivity to changes in $ß_3$ is constant throughout the amplifier passband. This rather surprising result is confirmed by experiment. Distortion associated with nonlinearity in $ß_3$ does not increase above the forward-path cut-off frequency ω_0 as loop gain falls away, nor does it increase above the ß cut-off frequency $\omega_3/ß_3$ of the output transistors.

Sensitivity to changes in $ß_3$ is inversely proportional to both $ß_2$ and $ß_3$. Increasing either reduces distortion without jeopardising stability. Notice particularly that $ß_2$ corresponds to the effective current gain of the second stage. Even in the ideal situation of very high transistor ß (such as a Darlington) and very large quiescent current I_C, $ß_{eff}$ given by Eq. 5 cannot exceed R_S/R_E. For example, in Self's Fig. 1a of August 1993, the first-stage collector-load resistors are 2.2kΩ and second-stage quiescent current is about 6mA. In a real circuit there would almost certainly be a resistor of 50-100Ω in Tr_4 emitter, as part of the protection circuitry. Hence $ß_{eff}$ of the second stage cannot exceed about 30. In contrast, if there is a current mirror in the first stage, the source resistance as seen by the second stage is large and $ß_{eff}$ approaches ß of the transistor. Herein lies the greatest advantage of the first-stage current mirror.

Self considers nonlinearity associated with $ß_3$ as a nonlinearity in the input resistance of the third stage and hence as a nonlinear loading on the second stage. He goes on to consider the benefits of an emitter-follower buffer between the second and third stages. This is perfectly valid, but I prefer to consider such a buffer as an extra member in the third-stage Darlington where it increases $ß_3$. Note that the 220Ω resistor in Self's Fig. 4f of October 1993 reduces effective ß (Eq. 5 again!); his Fig. 4e would be my preferred option. However, because sensitivity to changes in $ß_3$ is inversely proportional to both $ß_2$ and $ß_3$, it would do just as much good (and probably be simpler) to put the extra transistor into a second-stage Darlington.

Distortion associated with g_{m3}. Nonlinearity in g_{m3} models the exponential I_C versus V_{BE} intrinsic to transistors (twice over, because the transistor is usually a Darlington, Eq. 3); it models the nonlinear voltage drop across r_B associated with ß (also twice over in Eq. 3); and it models cross-over distortion. Effective g_{m3} includes the local emitter degeneration that is associated with emitter ballast resistors (Eq. 4).

Sensitivity to changes in g_{m3} increases in proportion to frequency, starting from a very small value which depends of all things on the equivalent resistance R_{C2} of the second-stage current-source load. The only way of reducing sensitivity towards g_{m3} while retaining the basic amplifier topology is to increase the overall cut-off frequency. Changing the emitter degeneration in any stage does not help, nor does increasing ß of any transistor.

Notice that cross-over distortion is predominantly associated with nonlinearity in g_{m3}. Self considers the cross-over region in detail; he points out the near impossibility of eliminating cross-over nonlinearity, stresses that the overall feedback is relatively ineffective in an amplifier of the topology of Fig. 1, and concludes that cross-over nonlinearity is the greatest source of distortion in a 'blameless' amplifier. In short, his observations confirm the theoretical prediction.

However, a great improvement can be achieved by slightly changing the amplifier topology: move the second-stage compensating capacitor C so that it encloses the third stage as shown in Fig. 27. Sensitivity to changes in g_{m3} becomes constant throughout

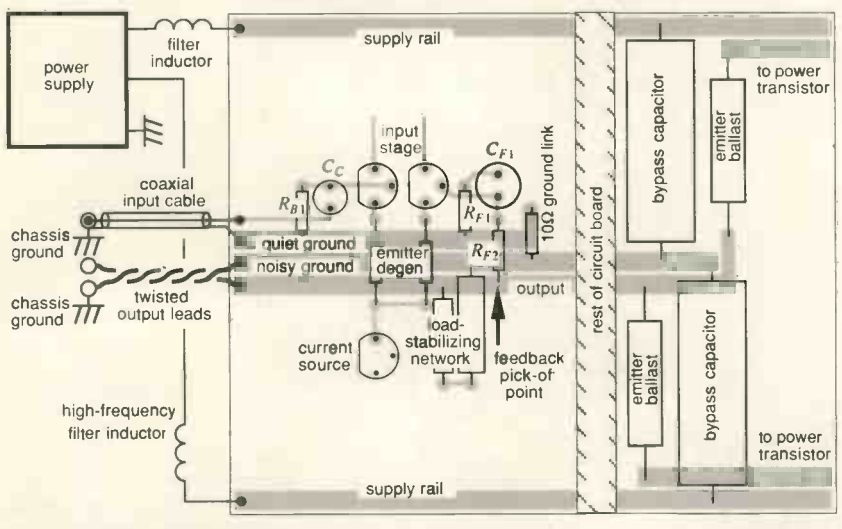

Fig. 8. Circuit-board and chassis layout for low distortion, showing separate tracks for noisy and quiet circuitry, separately grounded.

the passband, instead of increasing with frequency.

Many people believe that moving C provokes high-frequency oscillation, but this is not my experience and I strongly recommend the change. My amplifiers always incorporate a judicious amount of emitter degeneration in the second stage and a properly-designed load-stabilising network. If an amplifier oscillates when C is moved, it usually oscillates at several megahertz (far above the frequency of unity overall loop gain) and will usually continue to oscillate if the overall feedback can somehow be removed. The oscillation is a local parasitic. Try adding capacitors of around 50pF between collector and base of the first member of the output Darlingtons, using the shortest possible leads. Try shortening all leads to the output transistors. Try a small resistor in series with C, in theory about 20% larger than the second-stage emitter-degeneration resistor.

Nested feedback loops

Self makes brief reference to multiple feedback loops, nested one inside the another, but this of course is to depart from the basic amplifier topology under consideration. He also mentions multi-pole roll-off.

Nested feedback loops in general, and my own nested differentiating feedback loops in particular, offer a very great improvement in amplifier performance. However, the designer of a nested-loop amplifier needs to understand what he is about: time constants must be in correct ratios or the whole becomes impossible to stabilise. This is not to say that nested feedback circuits become more critical towards component tolerances – far from it – but the nominal values do need to be right.

Interested readers might refer to Reference 7, which describes how two nested differentiating feedback loops can be added to an amplifier of Self's basic topology, leading to an order-of-magnitude reduction in distortion. Loop roll-off is at a three-pole rate.

Protection

Self's class-B amplifier (February 1994) includes no protection – no doubt the circuit as printed was never intended to be a complete design. This amplifier would almost certainly be destroyed by even a momentary short-circuit of the output terminals; it requires current limiting in the output stage (probably of the fold-back variety) and also in the second stage, as in **Fig. 6**.

Despite what I may have published in the past, I have in recent years become an advocate of nonlinear fold-back limiting for the output stage. The circuit is not complicated, and it gives better protection than either simple limiting or linear fold-back limiting without restricting an amplifier's ability to drive reasonable reactive loads, so much so that it may even be possible to dispense with fuses in the supply rails.

Figure 7 shows the accessible regions of the load *VI* plane; the applicable design equations are,

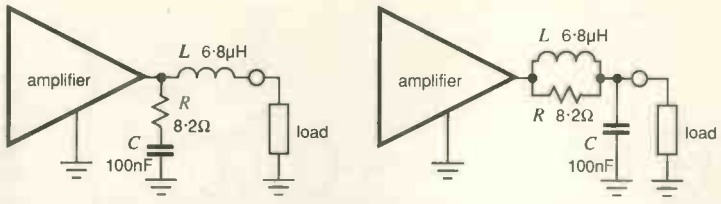

Fig. 9. Two forms of Thiele's load-stabilising network, with component values for 8Ω nominal load and 200kHz cut-off, where the output transistors see a nominally constant load.

$$I_{min} \approx \frac{V_{BE}}{R_{E3}} \qquad (14a)$$

$$I_{max} \approx \frac{1}{R_{E3}}\left[V_{BE} + V_{CC}\left(\frac{R_1}{R_2}\right)\right] \qquad (14b)$$

where V_{BE} is about 0.7V and R_1 should be somewhere around 100Ω.

I regard current limiting as mandatory in the second stage. If the load is short circuited and the input signal goes negative, the second stage is turned hard on, fighting against the limiter for the p-n-p half of the output stage. A simple current limit is sufficient for the second stage, and I set this at rather more than twice the quiescent current.

$$I_{limit} \approx \frac{V_{BE}}{R_{E2}} \qquad (15)$$

Typically, this quiescent current is a few milliampères, so R_{E2} becomes 50-100Ω. This provides just about the optimum level of emitter degeneration for high-frequency stability, as referred to above.

Distortions outside the feedback loop

Hum and distortion currents. Correct layout of an amplifier pcb is essential, to isolate hum and distortion currents in the output stage from the low-level wiring. **Figure 8** shows the approach I adopt[8] to reduce both conductive and inductive coupling.

Note first the use of separate quiet and noisy ground tracks on the pcb, connected to chassis ground at separate points. Power-supply ground is connected to the chassis at yet another point. I don't believe in single-point grounding! Within the power supply, the transformer centre-tap and filter-capacitor grounds are all joined together as described by Self, and then a single lead comes out from this junction.

Quiet and noisy ground tracks run parallel to each other on the pcb, and as close as possible to minimise the area between them. Magnetic fields associated with the large currents in the output stage induce voltages between these tracks, proportional to this area.

Connect the quiet ground track to chassis ground at the input socket, via the shield on the input coaxial cable. This track carries the input-resistor and feedback-resistor ground currents and, depending on circuit details, may carry the ground currents from intermediate-level stages. Also, the vector area of the loop formed by the input lead/coupling capacitor/input transistors/local emitter-degeneration resistors/feedback capacitor/feedback resistor is zero; follow this loop, and note how the

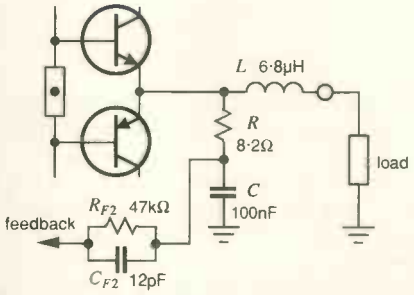

Fig. 10. Modified load-stabilising network which incorporates the overall feedback network. Values are for 8Ω nominal load and 200kHz cut-off.

areas enclosed on its left and right sides are equal. All these components hug the quiet ground track.

Connect the noisy ground track to chassis ground at the output terminal, via the twisted output leads. This track carries the ground currents from the supply bypass capacitors, the load-stabilising network (if any), and it may also carry ground currents from intermediate-level stages. The ground ends of the bypass capacitors are connected to this track as close together as possible, and no connections are made to the track between these two capacitors.

Similarly, the emitter ballast resistors in the output stage are connected to the output track as close together as possible, and no connections are made to this track between these two resistors. The output track runs parallel and close to the noisy ground track. The feedback pick-off point from this track is located between the ballast-resistor connections and the output connection, as is any load-stabilising network.

Mutual inductance should be zero between signal wiring and the loop formed by the bypass capacitors, emitter ballast resistors, power transistors and associated wiring. This corresponds approximately to setting zero vector area for the loop; in Fig. 8 the ballast resistors and bypass capacitors form a figure-of-eight, and the tracks to collector and emitter are spaced as closely as possible. Self's recommendation of twisted power-supply leads (February 1994) is really not enough; harmonic currents flow in all components of the loop, including the wiring to the power transistors. I recommend small filter inductors of a few microhenries in the positive and negative supply rails, to confine high-frequency components of supply current to the figure-of-eight loop on the pcb where the wiring layout is well defined[8]; these inductors should be

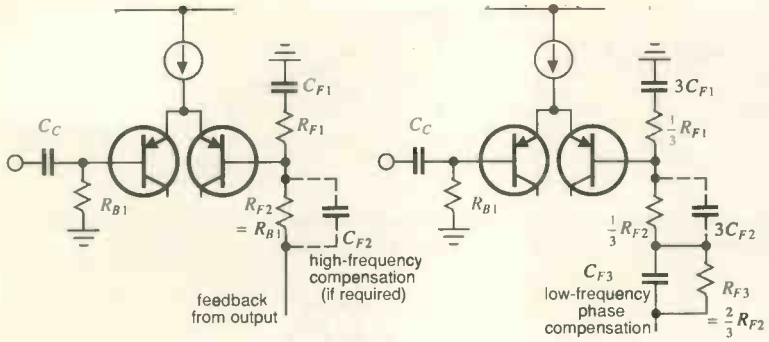

Fig. 11. Simple and phase-compensated low-frequency cut-off: as important at low frequencies as at high frequencies.

mounted well away from the pcb itself, so that their magnetic fields do not interact with the input stage.

The two ground tracks on the pcb are linked via a 10Ω resistor, so that the bypass capacitors are effective for the low-level stages. This resistance is practically short-circuit in comparison with the impedances in typical low-level stages (it is smaller than the reactance of a 10nF capacitor at all frequencies up to 1MHz), but is open-circuit in comparison with the impedances in high-level stages. Signal components of current in the low-level stages can cross into the noisy ground track, but currents in the high-level stages cannot cross into the quiet ground.

Further thoughts on distortion

Here are a few ideas, unrelated to Self's articles.

First, there was another outstanding series on audio amplifiers by Peter Baxandall in *Wireless World*, beginning in January 1978[9]. Sixteen years on, these articles are still well worth reading.

Load-stabilising networks. Thiele[10] has proposed an *LRC* network to be connected between an amplifier and its load, to reduce the problem of high-frequency instability when the load is capacitive and also to reduce the problem of radio-frequency pickup on the loudspeaker leads. The output transistors are in principle presented with a constant-resistance load, and in practice are protected from the worst excesses of high-frequency variation in loudspeaker impedance.

Figure 9 shows two forms of Thiele's circuit. Parameter inter-relations are:

$$R = R_{L(\text{nominal})} \qquad (16)$$

$$\frac{1}{RC} = \frac{R}{L} = \omega_X \qquad (17)$$

where ω_x is the network cut-off frequency, usually corresponding to 100-300kHz. Figure 9b, with 100nF connected directly across the load looks crazy – more like an unstabilising network – but it is correct and has some advantages.

It is amazing how few published circuits are correctly designed (Self's are not). Usually they appear to be based on Fig. 9a, but they include a resistor in parallel with L as shown in Fig. 1, and the values are all wrong anyway!

Figure 10 is a modified load-stabilising network[2,7] which incorporates the overall feedback network and has two advantages over Fig. 9 in that the network does not introduce a 3dB loss at the cut-off frequency ω_x; and that radio-frequency interference picked up on the loudspeaker leads is isolated from the feedback point by a two-pole filter (isolation in Fig. 9 is single-pole).

If the amplifier without feedback has just one dominant pole, and if the overall loop gain without the network falls through unity at ω_x, then the overall response is made phase-linear by choosing

$$R_{F2}C_{F2} = \frac{(\sqrt{3}-1)}{\omega_X} \approx \frac{0{\cdot}7}{\omega_X} \qquad (18)$$

In practice, where the amplifier has second-order poles, C_{F2} is selected around this value to give the best square-wave response. The inductor should be air cored and mounted with nylon or other non-conducting screws, well away from any metalwork to avoid nonlinear eddy-current losses. I usually mount it on the pcb, in the under-populated area near the first-stage tail current source.

Low-frequency phase compensation. There is a simple modification to the overall feedback network which linearises the phase of the low-frequency cut-off and improves the square-wave response.

In **Fig. 11a** the low-frequency cut-off associated with the feedback network is

$$\omega_{low} = \frac{1}{R_{F1}C_{F1}} \qquad (19)$$

There is an additional fall-off associated with R_{B1} and C_C, usually small in comparison. If ω_{low} is chosen corresponding to 5Hz, a 20Hz square wave is reproduced with about 40% tilt and looks nothing like a square wave. This is the result of phase nonlinearity; all the Fourier components in the waveform are reproduced within 3% of their correct amplitudes.

However, if C_{F3} in Fig. 11b is chosen,

$$R_{F3}C_{F3} = \frac{2}{\omega_{low}} \qquad (20)$$

the phase is linearised and a 20Hz square wave is reproduced with essentially zero tilt. In practice, C_{F3} should be somewhat smaller than this theoretical value, to compensate also for the phase associated with C_C. The nearest preferred-value resistors are close enough.

Maximal flatness of frequency response is incompatible with phase linearity, at both low and high frequencies. Linearising the high-frequency phase inevitably results in a small drop in gain. Linearising the low-frequency phase inevitably results in a small peak (1dB at 1.6Hz for the values given). Given the choice between flat frequency response and phase linearity, everybody opts for the latter at high frequencies (that is, for best square-wave response). Why not at low frequencies?

Capacitor types. Most readers will know that polyethylene-terephthalate (Mylar) capacitors exhibit nonlinear effects at audio frequencies, associated with the dielectric relaxation time, and should be avoided in high-quality amplifiers. Polycarbonate capacitors are recommended for values up to a few microfarads. A problem is C_{F1}, of the order of 100μF. Nonlinearity in C_{F1} results in distortion that increases at low frequencies.

Ten years ago I made a study of the capacitors available in Australia. The surprising result was that ordinary cheap aluminium electrolytic capacitors were remarkably linear, far better than most tantalum types. I have made no measurements on more modern components, but the underlying chemistry has not changed, so it is unlikely that the situation has changed. Be sure to use the capacitors in the correct polarity – the positive side of both C_C and C_{F1} towards the transistor bases if amplifier polarities are as in Figs 1 and 2. ■

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INTERFACING
piezoelectric cable

Piezoelectric coaxial cable offers a very simple means of sensing pressure or vibration. But since the signal produced is high impedance, a carefully designed interface is needed. Barry Gillebrand looks at some of the alternatives.

Piezoelectric coaxial cable is similar to standard coaxial cable with the inter-electrode insulation replaced by a piezo-electric composite elastomer. When compressed or flexed, the cable generates a voltage between the centre conductor and the screen braiding which varies directly with the applied radial pressure.

The voltage/pressure relationship is linear over several decades. However, being a signal source of high Thévénin impedance – typically several megohms – care is needed when designing interface circuits or preamplifiers.

Consider the design of an amplifier to operate at audio frequencies with a gain of 40. The simplest approach, **Fig. 1**, would be to design a non-inverting voltage amplifier using a low-noise op amp such as the *TLA071*. This has a unity-gain bandwidth of 3MHz, so this would give a flat response up to around 70kHz.

It is necessary to provide a path to signal earth for the input bias current of the amplifier, in this case some 200pA. The resistor needed for this purpose appears as the principal component of the input impedance, so it should not be less than 20MΩ if the piezo cable is not to be loaded excessively.

A current of 200pA flowing through 20MΩ generates an offset voltage of 4mV, which combined with the inherent input offset voltage of 3mV and the contribution of the gain-defining components will produce an output offset of 0.2V; if offset trim is not used, ac coupling may be advisable in the next stage. The principal advantage of this circuit, apart from its simplicity, is the excellent low-frequency response, which extends down to dc.

A more compact voltage amplifier with the advantage of lower noise figures can be constructed using an fet/bipolar cascode pair of transistors. The circuit of **Fig. 2** has an input impedance of 20MΩ and exhibits a high degree of linearity due to the negative feedback used. Gain can be varied without affecting the operating point of either transistor by changing the feedback coefficient using the ac-coupled resistor R_2. The gain of this amplifier is:

$$A_v=(R_1+R_2)(R_s-R_2)/R_sR_2$$

assuming reasonably high values of g_m and h_{fe} for the two transistors: I assumed 5mA/V and 100 respectively for this calculation, giving an open-loop gain of some 105dB with the specified load resistors.

The output impedance of this collector-coupled circuit may be a little high for some applications; there are a number of commercial instrumentation amplifiers of hybrid microcircuit construction which can be used 'out of the box'. Unipulse's *U300* series has an output impedance of 0.1Ω and and a fixed gain of 200, variable gain versions are also available.

One of the most suitable types of amplifier

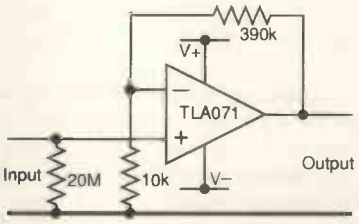


Fig. 1. A non-inverting buffer based on a low-noise op-amp is the simplest way of interfacing the high-impedance of piezoelectric cable.

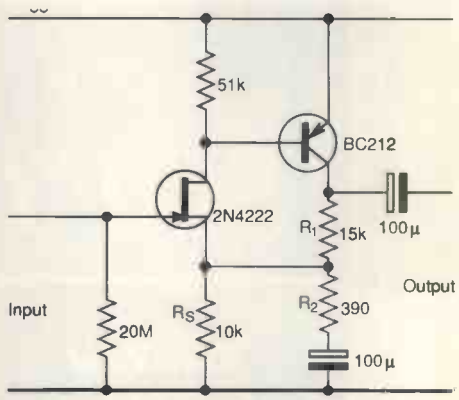


Fig. 2. Fet/bipolar cascode amplifier with input impedance of 20MΩ and excellent linearity due to negative feedback.

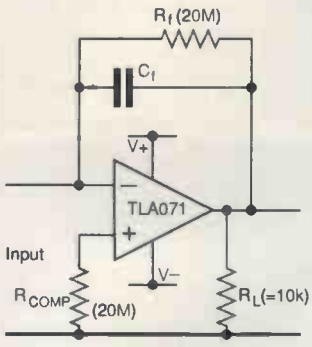


Fig. 3. One of the best types of amplifier for interfacing is the charge amplifier since it amplifies the input signal but not dc offsets.

Barry Gillebrand is managing director of Quantelec.

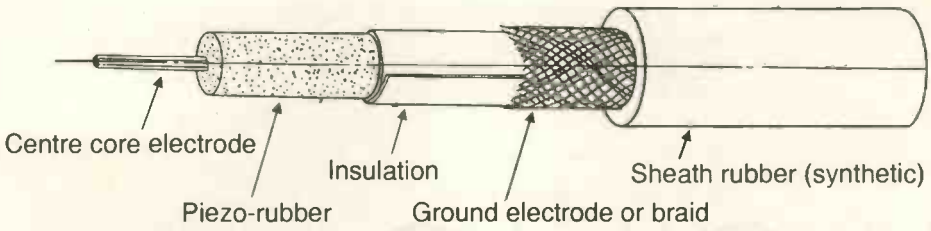


Fig. 4. Piezoelectric coaxial cable produces an output voltage in response to compression or vibration.

for use with piezoelectric cable is the charge amplifier, which exhibits a higher gain to the generated signal than for any dc offsets. A circuit is shown in **Fig. 3**, where the operational amplifier is again a low noise, low offset fet input type such as the *TLA071*. For a continuously-varying signal, the transfer function of such an amplifier is :

$$V_o=Q/C_f(1+1/(jR_fC_f))$$

where Q is the rms value of the charge generated by the variations in pressure. Note that the effect of the feedback resistor needed to provide a leakage current path for the op amp is to introduce a single pole which limits the gain at low frequencies; for the *TLA071* this could be 20MΩ for an offset of less than 5mV at the output.

For a 1000pF piezo cable, a feedback capacitor of 25pF would give the required gain, since the charge variation generated by the cable is proportional to its total capacitance. The lower 3dB point of this amplifier would then be 318Hz.

Getting charge amplifiers to work at very low frequencies poses a problem because of the large values of feedback resistor needed with typical feedback capacitor values. In this instance, the desired gain could be split between the charge amplifier and a conventional second stage to improve the low-frequency response; assuming the charge amplifier had unity gain, the feedback capacitor would then be 1000pF and the lower 3dB point at 8Hz.

As with any engineering design, a compromise must be reached between the various characteristics of the amplifier: maximum permissible offset versus low-frequency performance, closed-loop gain and resistor availability and physical size here. With such large values of resistance and small capacitors, this is an obvious contender for a custom hybrid, if numbers permit. The principal advantage of using a charge amplifier compared with a voltage amplifier as the first stage is that the piezoelectric cable is driving a virtual earth point, so the input impedance of the op amp does not load the transducer.

Piezoelectric coaxial cable is a durable, water-resistant form of pressure sensor ideally suited to outdoor applications such as vehicle detection and vibration analysis. With a little care in the design of the input stages of the conditioning amplifier, it is also linear and responds well to both high and low frequencies. Though still a very new material (first described by Hisao Banno of NTK Ceramics in 1986), it should find widespread application in industry over the next few years. ■

Further reading

Banno, O., *et al*: *Piezoelectric & Acoustic Properties of Piezoelectric Composites*, *Japan Journal of Applied Physics*, Vol. 26.

For an excellent analysis of charge amplifiers, see Mark Serridge & T.R. Licht: *Piezoelectric Accelerometer & Vibration Preamplifier Handbook*, Bruel & Kjaer, 1987.

For further information on piezoelectric cable, contact Quantelec at 50 Market Square, Witney, Oxon OX8 6AL. Tel : 0993 776488, fax 0993 705415.

ELECTRONICS WORLD
SOFT INDEX

A computerised index of *Electronics World+Wireless World* magazine is now available. It covers the five years 1990 to 1994 – volumes 96 to 100 – and contains over 1400 references to feature articles, circuit ideas and applications, with a synopsis for each. The software is easy to use and very quick. It runs on any IBM or compatible PC with 512K ram and a hard disk. Each disk is scanned before shipping with the current version of Dr Solomon's Anti-Virus Toolkit.

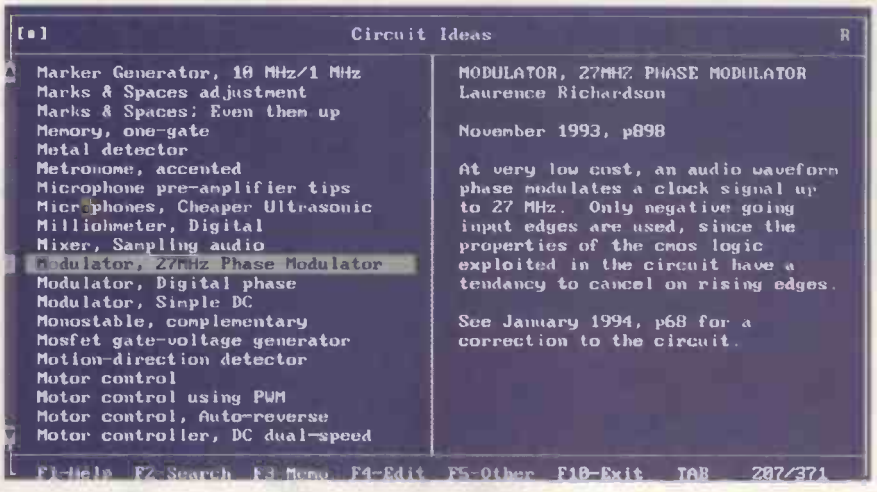
For the UK, the five year index is priced at £20. Please specify $5^1/_4$ or $3^1/_2$in format. This price includes UK postage and VAT. Add an extra £1 for overseas EC orders or £5 for non-EC overseas orders.

Photo copies from back issues of *EW+WW* are available at 50p per page plus VAT (in EC) and a flat postage charge of 50p UK, £1 rest of EC, and £2 rest of world. For enquiries about photo copies send an sae to Video Interface Products.

Please allow up to 28 days for delivery. Cheques should be made payable to **Video Interface Products,** not *EW&WW* or Reed Business Publishing.

Please post your request to **Video Interface Products Ltd, 1 Vineries Close, Cheltenham GL53 0NU, UK.**

As with feature articles, each circuit idea in the EW+WW computerised index is accompanied by a brief description.



Peter Marlow explains the rationale behind the new computerised index for *Electronics World*.

EW+WW SOFT INDEX

Electronics World and Wireless World now has a computerised index. It runs on IBM and compatible personal computers, and covers the five years from 1990 to 1994 – volumes 96 to 100. It contains over 1400 references to feature articles, circuit ideas and applications, each accompanied by a synopsis.

Why put the index of *EW&WW* on to a data base? A computer is particularly good at searching quickly through large quantities of data. Applied to an index it can save a lot of time looking for information.

Minimum requirements for the soft-index are an IBM or compatible pc running DOS 3.0 or higher, with at least 512K of ram, and a hard disk with 600Kbyte available. A floppy disk drive for either $5^1/_4$in 360K or $3^1/_2$in 720K disks is needed to load the software. The program runs under DOS, and supports a mouse; it will also run via windows.

Features

The soft-index is easy to use with plenty of on-screen help. Finding information is a two stage process; first you select a subject in the 'Table of Contents' then you examine the subject index. The subject index is like a card file: an index is displayed on one side of the screen while information relating to the chosen index item appears on the other side.

Many useful features have been built into the software. They are accessible by using function keys and menus. Rapid searches of data on one or more words are possible. A Memo function enables comments to be added to index data. A blank subject index is also provided for your own notes and information.

The *EW&WW* Soft-Index should be treated like a book. A purchased copy should be run on only one machine at any one time but a copy may be taken for backup.

Now for a look at the main features of the software. Full instructions for use are contained in the manual supplied with the disk.

The *EW&WW* Soft-Index is shipped on disk in 'archived' form. There is a simple install procedure which results in the program and data being 'expanded' and copied to a directory called EW on drive C, taking up 600Kbyte. Other directory names and drives can be used.

To run the program, type 'EW' followed by Enter.

Table of contents

After a short pause, the Table of Contents screen appears. A list of subjects is displayed on the left hand side of the screen. The Information section provides details about *EW&WW* such as editorial information, magazine subscription rates and back copies. It also contains licensing details, etc. 'Notes' is, as the name suggests, for your own notes, information and contacts, etc.

To access the subject index, move the highlight bar to that subject using the up or down arrow keys and the PgUp or PgDn keys. Then press the Enter key. A faster way of positioning the highlight bar is to type the first letter of the desired subject. If a mouse is in use, it can be 'clicked' on a subject to select it.

On the far left of the screen there is a 'progress bar' which shows where you are, relative to the rest of the file. The little arrows at the top and bottom of the progress bar are for scrolling up and down with the mouse. The mouse can be 'clicked' on any part of the progress bar to go to different parts of the Table of Contents.

The 'Help line' at the bottom of the screen identifies the various functions available. Pressing the F1 key brings up a page of more detailed help information.

The Search function is perhaps the most useful feature. Pressing F2 enables word searches to be made of either the individual subject indexes (very fast) or all of the data.

On the right of the screen is a welcome message. This is a scratchpad area for your notes which can be modified or edited by pressing the 'forward arrow' key. Text can be typed anywhere in the box. Press escape to exit.

The current line number and total lines of the Table of Contents are displayed at the bottom right of the screen.

To exit the program press F10, or 'click' the mouse on the small square in the top left of the screen.

The Subject Index shown belongs to *Circuit Ideas*. The screen has two halves with an index displayed on the left of the screen and data on the right. Moving the highlight bar to a different index entry automatically brings up the correct data as with a card file. A rapid way of positioning the highlight bar is to spell the desired item on the keyboard.

As before, the 'Help line' at the bottom of the screen identifies the various functions available. Pressing the F1 key brings up a page of more detailed help information.

F2 is the Search function which is more extensive than in the Table of Contents screen. It allows searches on up to five words, either together (and) or individually (or). Entries can also be marked with an arrow for quick access later.

Information in the top left of the screen shows that two items are marked and that you are pointing at the first one. The + and – keys are used to 'travel' between marks. In the opposite corner an 'R' symbol shows that the file is 'read-only'; it cannot be edited as the information is fixed. However, using the Memo function, F3, a note can be added in the bottom line of the data. In the *Notes* subject index data can be added or altered using the edit function, F4.

'Other' functions on the F5 key enable printing. Individual records can also be stored in an ascii text file for incorporation into word-processed documents – particularly useful for company names and addresses.

To return to the Table of Contents press F10 or escape, or 'click' the mouse on the small square in the top left of the screen. The computer remembers where you are for next time.

I hope that you will make suggestions about the content, presentation and software operation. The plan is to publish a new issue of the disk annually, and any useful amendments will be incorporated. ■

Advances in ADCs

Growth in demand for a-to-d converters – in particular for devices with lower power consumption – has triggered the introduction of a number of new designs over the past two years. Simon Parry reports.

The last two years have witnessed a hive of activity in the world of analogue-to-digital converters. Designers have been spurred to develop novel architectures by demand from growing audio and communications applications.

Sigma-delta oversampling, interpolation, residue or subranging and pipeline techniques have come to the fore as engineers have striven to gain performance without significantly impacting either chip area or power consumption. The old a-to-d converter king – the flash device – has hardly had a look in because of its high power-consumption.

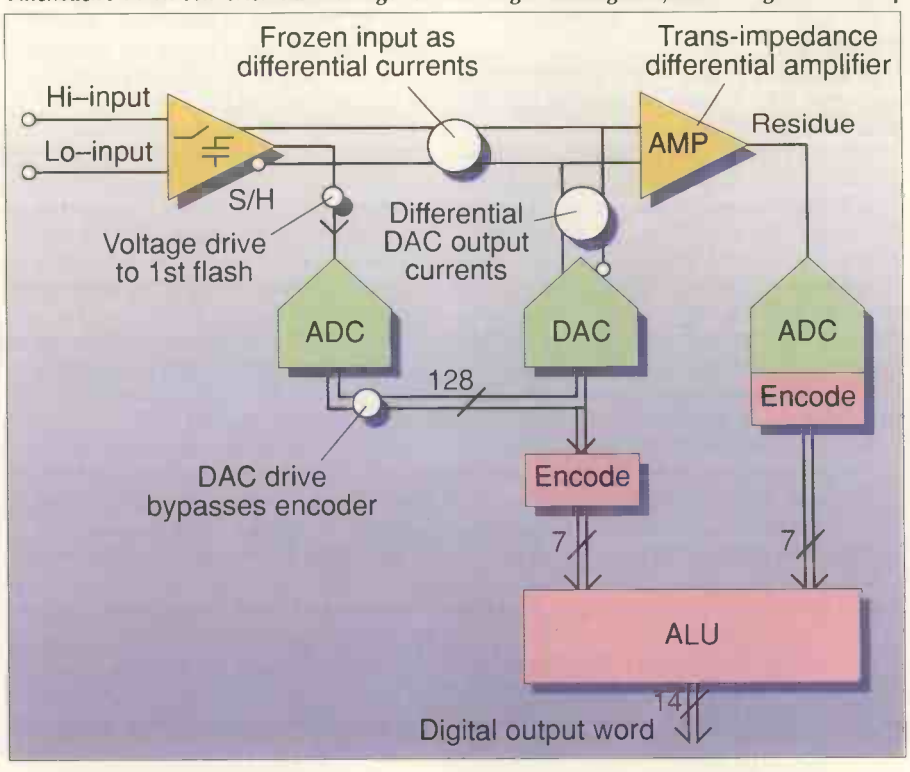
The last few months have proved exceptional in the number of analogue-to-digital converters being either launched onto the market or nearing completion in the lab. And the trend is due to continue. Again next year, the International Solid State Circuits conference has two whole sessions dedicated to data converters.

Thankfully Europe – the UK included – is not being left behind in the innovation stakes. Philips engineers continue to drive forward the technology and in recent months world-class analogue-to-digital converters have been designed by UK engineers.

For instance: Data Conversion Systems, a small UK chip design company based in Cambridge, has developed a world-beating analogue-to-digital converter. Engineers there have designed a 500kHz analogue-to-digital converter with a 14-bit resolution employing novel techniques which ensure a compact circuit.

Mark Pinchback, lead designer, reports that samples of a prototype two-chip implementa-

Pinchback's 14-bit converter is a two stage residue design attaining its dynamic range with overlap



tion have just been received and are currently undergoing testing.

"The testing is going well," said Pinchback. "It will be difficult characterising the complete performance of the analogue-to-digital converter because it is in two parts at the moment but we should have a single monolithic chip early next year."

The single chip should be the industry's first monolithic converter to offer a 14-bit 500kHz performance, says Pinchback. It will cost significantly less and have a higher reliability than the hybrid converters currently available. "DCS is hoping to have the part on the market by October next year," he said.

According to Pinchback current monolithic 14-bit analogue-to-digital converters do not have a sampling rate above 100 to 200kHz. "This device is at least twice as fast," he said. "Also, although the initial target is 500kHz, the design is capable of going to 2MHz."

The converter employs the classic two-stage residue architecture in which the output digital word has been converted from the analogue input in two halves. However, clever circuit design and the incorporation of signal processing techniques have meant dramatic silicon area savings and an enhanced dynamic range, says Pinchback.

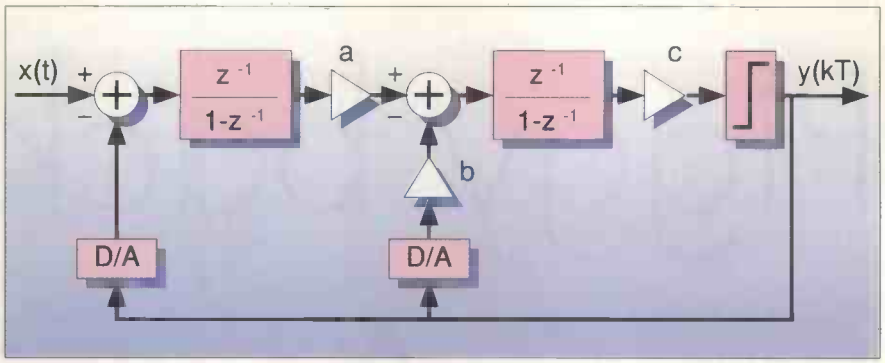
In a residue converter a first-stage analogue-to-digital converter produces a coarse approximation of the input analogue sample. The digital output is reconverted into the analogue domain and subtracted from the analogue input. The residue, or conversion error, is amplified and then converted by a second 'fine' converter. The coarse and fine digital outputs are combined to produce the final digital output word.

However, a conventional residue converter needs an overlap between the first and second digital samples to correct for system errors. This overlap reduces the overall dynamic range; two 8-bit analogue-to-digital converters with an overlap of 2 bits would be needed for a 14-bit output.

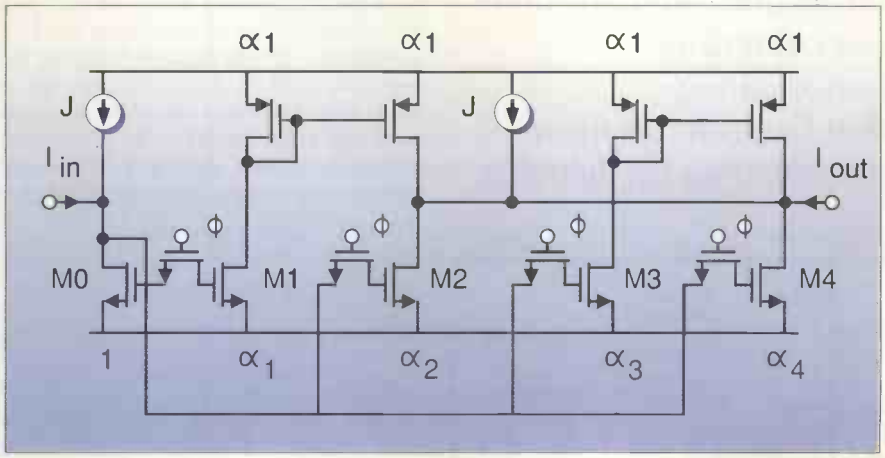
The breakthrough in the DCS converter is to use only 7-bit converters combined with digital signal averaging to give a true 14-bit dynamic range on the output. A two-bit overlap has to be employed – just like any other residue converter – but the signal averaging over a number of samples suppresses the noise floor 'recovering' the dynamic range.

The arithmetic logic unit, alu, performs the signal averaging on the fine second-stage conversions. The user of the analogue-to-digital converter is able to control the alu to trade-off resolution against speed. At 500kHz 16 samples are averaged but at 2MHz only four are used.

Other features of the converter include a DCS-developed 'flashDAC' approach inside the converter and novel dual-mode front-end sample-and-hold amplifier. In the flashDAC circuit, the digital output from the initial



The basic Sigma-delta modulator, above, employed by the Linkoping University team utilises a memory-cell with low noise clocking scheme(below).



coarse analogue-to-digital conversion is reconverted into an analogue residue signal without first being encoded. This combined with a subranging architecture for the coarse flash analogue-to-digital converter ensures there are no pipeline delays allowing the complete system to provide continuous 14-bit conversion at up to 2MHz signal input.

Initial simulations predicted a signal-to-noise ratio of 82dB at 500kHz and 78dB at 2MHz. The two-chip prototype and the final monolithic device are being built in a 2µm Bicmos process from SGS-Thomson Microelectronics.

At Philips, engineers are preparing samples of a folding-interpolating analogue-to-digital converter to offer a 10-bit dynamic range at up to 100MHz sampling. Importantly, the device, the *TDA8762*, will consume half the power of the company's present *TDA8760* chip; that is about 350mW.

The *TDA8762* is more akin to the Philips *TDA8718* part which uses folding-interpolating techniques to attain 8-bit resolution at a 650MHz sampling rate rather than the *TDA8760* which is a two-stage residue converter.

Folding-interpolation techniques use re-entrant transfer functions that allow the designer to reduce the number of input stages for two reasons. First quantisation levels are interpolated between comparator stages and analogue preprocessing of the input signal means each comparator stage detects more than one level of the input.

The major advantage of this approach is the reduction of input capacitance, because of the reduced number of input stages, allowing a higher performance.

"We've seen samples of the *TDA8762* with a performance better than 9.6 bits with a 40MHz input," said Claude Giraud, Philips product manager. "A slight modification of the architecture should take us to 100MHz."

At Linkoping University in Sweden a team of engineers has designed an 11-bit sigma-delta modulator (the basic building block of sigma-delta data converters) using switched-current circuit techniques rather the conventional switched-capacitor schemes.

Although the performance of the modulator is unremarkable – it attained a 60dB signal-to-noise-and-distortion ratio with a 2kHz input and an oversampling ratio of 128 – the circuit is notable for its use of switched-current techniques.

The modulator was designed in a 0.8µm digital cmos process and operates from a 3.3V supply. A modified memory cell provided reduced power consumption and chip area and eliminated some of the noise sources associated with the switched-current technique. ■

Piggy backing for

Piggy-backing two or three balanced-driver audio ICs improves both noise and common-mode rejection ratio performance. Ben Duncan has been investigating the benefits.

Analog Device's *SSM-2142* balanced driver IC is a cross-coupled circuit simulating a signal transformer with a centre-tapped secondary. It suits both audio and instrumentation.

To avoid degrading low-frequency common-mode rejection – particularly in instrumentation – the manufacturer cautions that output dc blocking capacitors must not be used. This is because the close matching, needed to avoid degrading the output common-mode rejection, is doubtful with the electrolytic-sized values required, of at least 100µF. However, real *2142*s commonly exhibit high dc output offsets, commonly ±10mV to above ±100mV.

There is no offset null capacity. To avoid the expense of selecting components, and assuming the input can be direct coupled, then the

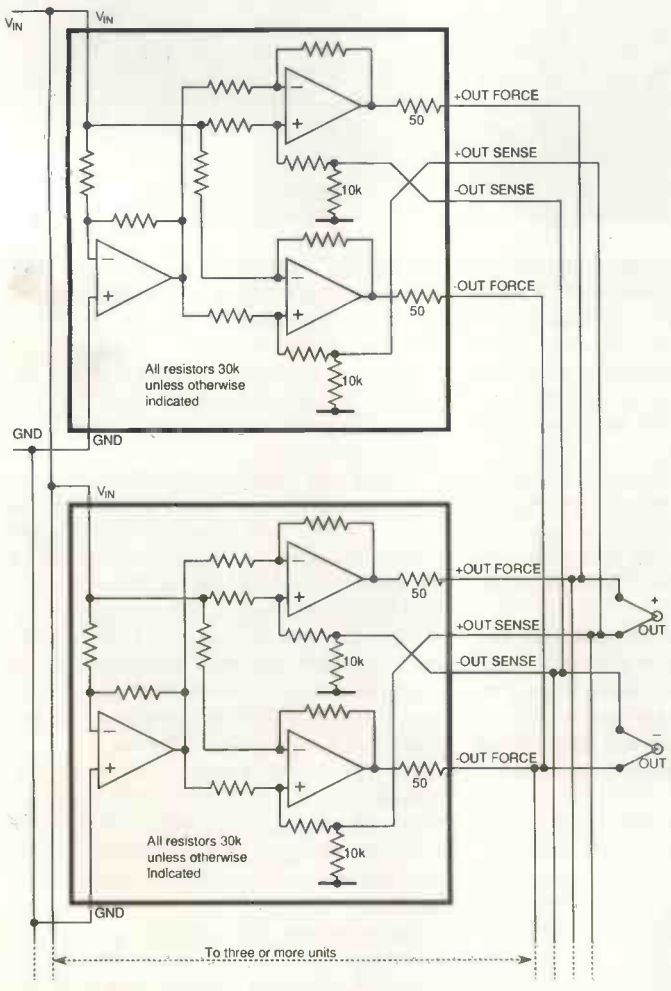
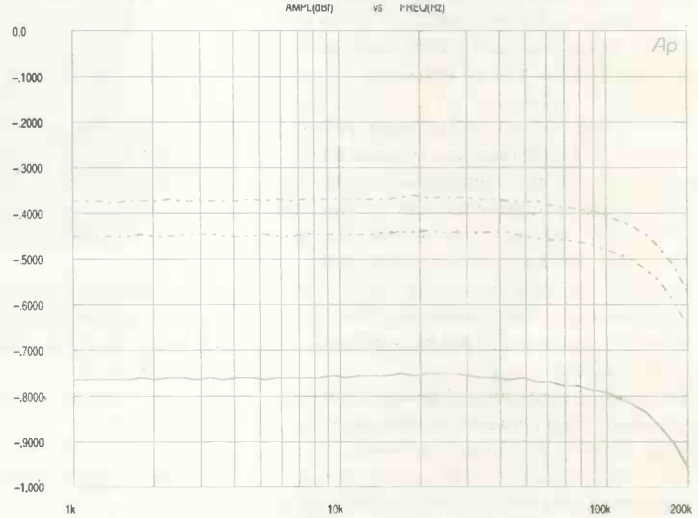


Fig. 1. To improve noise and cmr performance, ICs are simply piggy backed and soldered to each other.

Fig. 2. Bandwidth remains almost constant whether the IC is used alone, bottom curve, or with two or three devices piggy-backed, middle and top curves. Parallelling devices does improve the transfer function by taking gain nearer to unity.



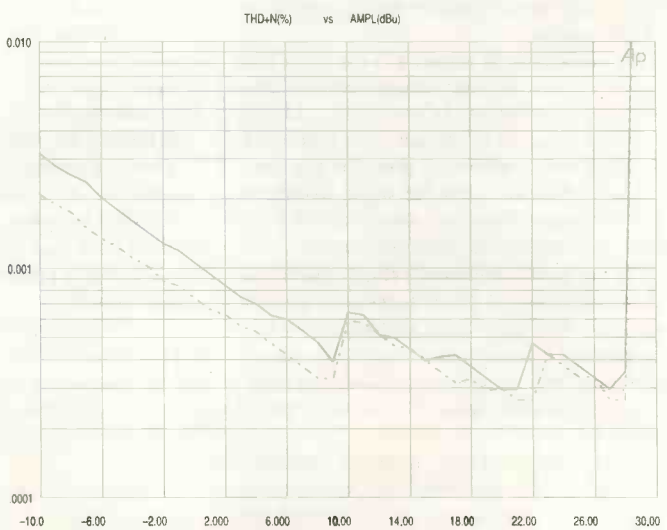


Fig. 3. Plotting distortion against output level shows an improvement at lower levels, where noise is dominant.

performance

preceding op amp stage can be offset by an amount that best nulls the *2142*'s outputs, both to each other – differential mode – and to ground – common-mode. Abusing an op amp's offset null facility in this way can degrade dc stability over temperature – a point that might need consideration in critical dc applications.

The *SSM-2142* is about 10dB noisier than is the norm in professional audio for the same balanced circuit. This is because the balanced circuit is commonly built, 'discretely' with an *NE5532* op-amp. But the *2142* has the advantage of a good common-mode rejection without trimming, and a major saving in pcb real estate. Its signal-to-noise ratio can be improved by parallelling, **Fig. 1**. The ics are simply stacked vertically, ie, soldered leg to shoulder.

Assuming dc offsets are matched within a few percent, outputs won't appreciably load each other, and each sees some fraction of the output loading. This may reduce total harmonic distortion in some cases.

Figure 2 shows how bandwidth is unaffected while unity gain is ever more closely approached, with one then two and three paralleled units driving 600Ω. Direct-current offset was measured at this stage, using a number of samples.

Differential offset hardly varied but common-mode offset could increase cumulatively. Direct-current offset was measured at this stage, using a number of samples. Grading would overcome this.

Figure 3 shows %THD+*N* plotted against level. The lowest point represents pure percentage total harmonic distortion. There is no difference here between one or three stacked devices. The curves on the left are by contrast mainly noise ('+*N*'), and the divergence shows that three stacked *2142*s (lower curve) are quieter. **Figure 4** confirms this: above 500Hz, where the ear is most sensitive, the stack of three is about 3.5dB quieter, as would be predicted.

Figure 5 shows how output common-mode rejection below 10kHz progressively improves by up to 15dB at low frequency, as units are stacked. The BBC test method was used; see the *SSM2142* data sheet. Higher stacks have not been tested, as the law of diminishing returns sets in rapidly with parallel channels.■

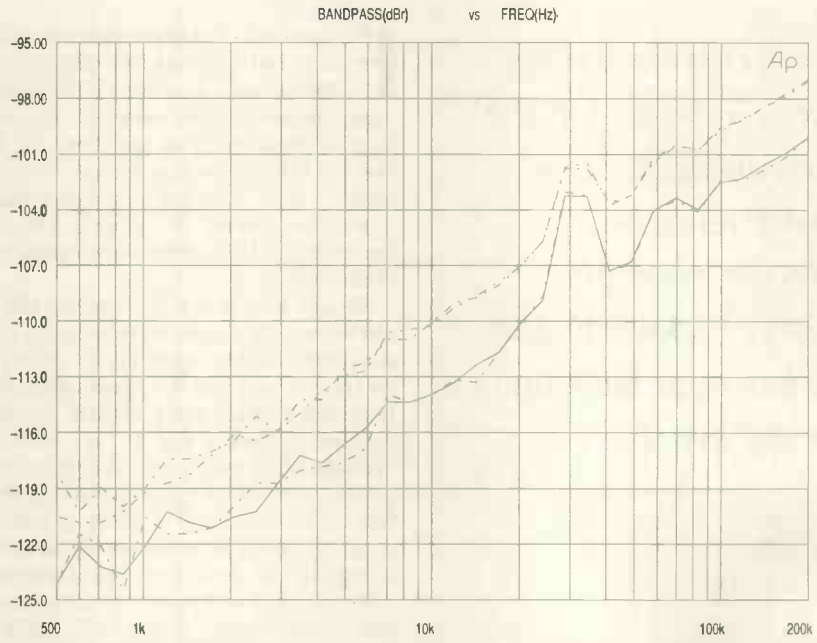


Fig. 4. Above about 500Hz, where the ear becomes more sensitive, stacking three ICs lowers noise by about 3.5dB.

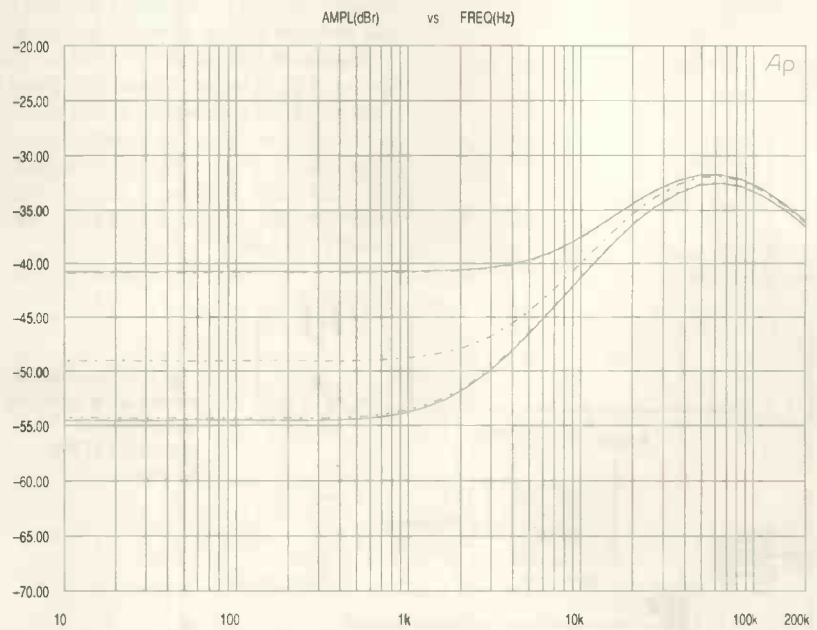


Fig. 5. Output common-mode rejection below 10kHz improves progressively by up to 15dB when three devices are stacked, bottom curve.

DETECTION before the diode

Its effect is similar to that of a diode, but the electrolytic detector clearly relies on different principles. George Pickworth puts the barretter back on the test bench.

Fessenden's electrolytic barretter, patented in 1903, was the first practical continuous wave detector. Yet for such an important device in the history of radio, technical information on its operation is scarce. Only by building modern replicas of various models and carrying out experiments ('Detection before the diode', *EW+WW*, December, pp.1003-1006) can its operation be understood.

The end result of using the barretter **Fig. 1** is half wave rectification of rf current, an effect simulated today by substituting a diode for the barretter and applied current source. With rf current forward-going through the diode (positive at point *A*, **Fig. 2**), the device conducts and half cycles are dissipated. When polarity reverses (rf current positive at *B*) the half cycles are blocked by the diode and routed through the load resistor to appear as positive-going half cycles on the oscilloscope.

Rectification in this way, by dissipation of unwanted half cycles, is very inefficient in that the barretter damps the tuner heavily and poor sensitivity results. But it leaves the wanted half cycles untouched, so non-linearity of rectification does not effect the end result and it can perform well as an am demodulator.

In a biased-carborundum crystal-set configuration (**Fig. 3**), where wanted half cycles pass through the detector and unwanted half cycles are blocked, it can not be made to operate. So an advantage of the crystal-set configuration – that the load on the tuner is virtually that presented by the headphones, around 30kΩ at 1.0kHz – is lost.

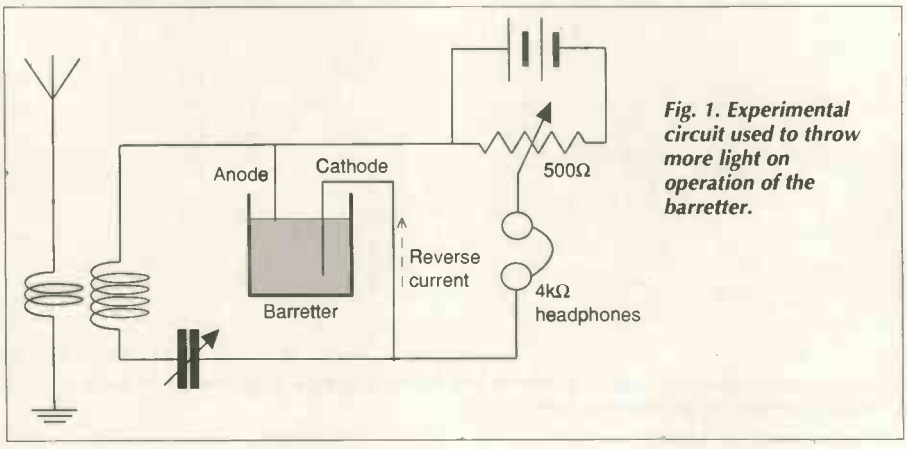


Fig. 1. Experimental circuit used to throw more light on operation of the barretter.

Setting up experiments

Its effect is similar to that of a biased diode, but obviously the barretter operates in a quite different manner. To find out exactly how, we must experiment.

Using a typical receiver circuit, Fig. 1, and a function generator as a signal source, the potential of the applied dc can be adjusted until the waveform across the load resistor (observed on the oscilloscope) shows that negative-going half cycles are greatly attenuated. The trace consists principally of positive-going half cycles which, significantly, are found to be riding on the applied dc.

If the applied dc is reduced to a level where forward current no longer flows, ie below 1.7V, rectification ceases and both negative and positive-going half cycles are then displayed. Optimum forward current is 10-400μA (applied dc 1.9-2.75V) depending on amplitude of the rf input **Fig. 4**.

But amplitude of the positive-going half cycles is only 25% of that achieved with the diode, even though input power is substantially the same. Possibly, when rf current is positive at *B* (Fig. 2), some of it passes through the barretter as reverse current and less passes through the load resistor. With the diode, reverse current is minimal.

Moreover, while negative-going half cycles (as displayed by the oscilloscope) are present but small with the diode, with the barretter they are significant and their magnitude increases with frequency, causing the forward/reverse current ratio to deteriorate progressively, ultimately becoming 1:1 (see panel entitled Oscillograms).

One explanation for the effect is that when rf current is positive at point *A*, the barretter's forward resistance is much higher than the diode's, causing significant current to flow back through the load resistor and appear, to the oscilloscope, as negative-going half cycle.

Presumably, the barretter's forward resistance increases with frequency, causing more 'reverse' current to flow through the load resistor, in turn increasing the magnitude of the negative-going half cycles. On the other hand, with optimum applied dc, the amplitude of the positive going half cycles remains more or less constant.

This can be simulated with a diode, **Fig. 5a**, where, by choosing appropriate resistors, Fig.

5b, the end effect is approximately that of the hybrid barretter. In the circuit R_2 shunts the load resistor and has a value of about 4.0kΩ – about the same as the headphones.

Blocking diode

What can we learn from blocking the reverse current? Putting an *AA119* diode, with corresponding polarity, in series with the barretter at point *X* on the cathode line, Fig. 3 causes forward current to cease when the applied dc is less than 1.7V. The output waveform disappears completely but the rf half cycles riding on the steady forward current can be observed much more readily on the oscilloscope.

Provided the applied dc has a potential high enough to allow a small forward current to flow through the barretter, the waveform displayed by the oscilloscope is virtually identical, in shape and magnitude, to that for the diode alone. Reduction in amplitude of negative-going half cycles ('reverse current') could be due to the blocking diode somehow reducing the barretter's forward resistance to that of the diode.

Water voltameter

We have seen how, at first glance, the barretter looks behave as a biased diode. But closer inspection reveals it to be a unique device, with operation based on polarisation and the movement of ions in an electrolyte during electrolysis. A brief look at the water voltameter could give some insight here.

The water voltameter does not ordinarily produce a current, but uses an applied current to decompose water into oxygen and hydrogen. The volume of gas produced is in direct ratio to the applied current and can easily be measured. Oxygen is released at the anode and hydrogen at the cathode.

When the applied voltage is less than 1.7V, current flows through the voltameter for only a short time, before stopping.

In simple terms, the electrodes became covered with a gaseous insulating film having a maximum dielectric strength of about 1.7V. So, when potential of the applied dc is less than 1.7V, current falls quickly to virtual zero through polarisation.

In this state, the electrodes are in effect hydrogen and oxygen, so when connected to an external load the internal current flows from the oxygen to the hydrogen electrode. Indeed, the voltameter seems to behave as if it were a capacitor and can be made to discharge, briefly, through an external load with the anode behaving as if it were the positive pole of a current source.

Interestingly, the hybrid barretter maintains a steady 40μA through a microammeter (resistance 650Ω) for 5h after the applied dc is switched off.

Using a digital voltameter to measure applied dc and an analogue microammeter to measure current, Fig. 6a, the experimental water voltameter develops a constant 0.2V even after standing for long periods with the applied dc switched off. Unless precautions are taken, this will cause a reverse current, heavy enough to damage the microammeter.

Starting with 0.2V as the base line, 0.1V increments in the applied dc cause a momentary rise in current to a peak value, Fig. 6b, which after 10s declines. The drop is attributed to the dielectric strength of the gaseous film stabilising at a new level after each incre-

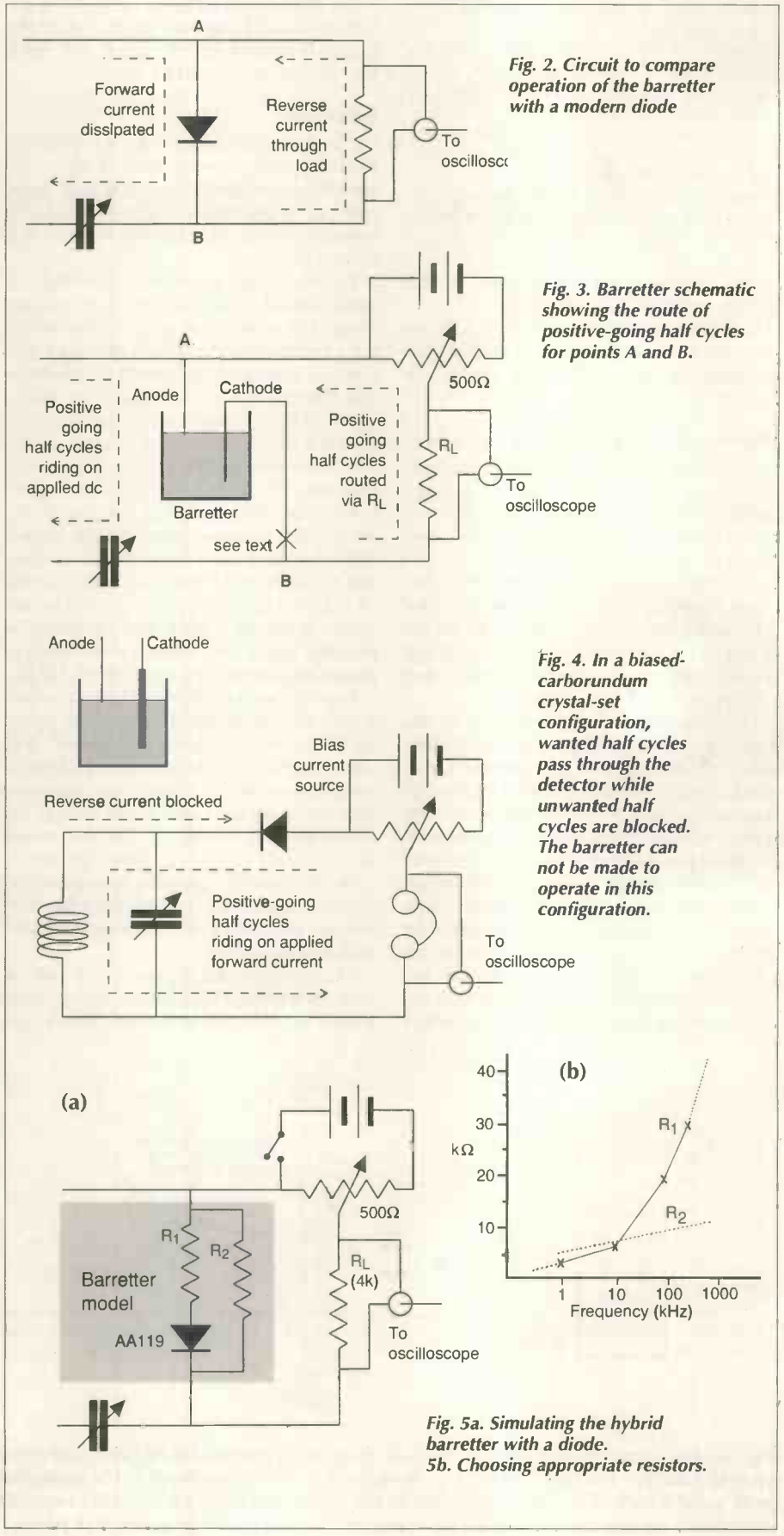


Fig. 2. Circuit to compare operation of the barretter with a modern diode

Fig. 3. Barretter schematic showing the route of positive-going half cycles for points A and B.

Fig. 4. In a biased-carborundum crystal-set configuration, wanted half cycles pass through the detector while unwanted half cycles are blocked. The barretter can not be made to operate in this configuration.

Fig. 5a. Simulating the hybrid barretter with a diode.
5b. Choosing appropriate resistors.

Reliability

All the replica barretters needed time to settle down, when some chemical reaction was probably taking place between the electrodes and the electrolyte

After the hybrid barretter was 'tuned' to give best results and allowed to settle, it behaved reliably and consistently during the period of experiment.

Ultimately lead sulphate deposits on the cathode became so great that the barretter ceased to function. But, so far, no deterioration has occurred with replicas employing platinum or mercury cathodes.

mental increase in applied dc. After 10s the decline continues, but very slowly.

When the applied dc reaches about 1.7V, as anticipated, the characteristics of the voltameter change and current increases more or less linearly with applied dc.

Water voltameter *v* hybrid barretter

In the hybrid barretter, current rises immediately to its steady value, **Fig. 6c**, and any slight increase in applied potential causes it to rise and remain steady at a new high point. The behaviour approximates to the lower values dotted curve for the voltameter, Fig. 6b.

Not surprisingly, the characteristic curve for a Delaney lamp detector is found to be similar to that for the water voltameter as the only real physical difference is that the Delaney lamp electrolyte is 20% nitric acid.

Differing electrical characteristics of the Delaney lamp detector, compared to the barretter, are attributed to the barretter's tiny anode. Only a minute current is required to polarise the anode, so stabilisation occurs too quickly to be observed on the microammeter. For the water voltameter, a momentary reversal of applied current causes instant de-polarisation, followed quickly by re-polarisation. This may well apply to the barretter where the smaller anode requires less current to re-polarise after conducting positive half cycles.

Such an interpretation is consistent with frequency response being related to the surface area of the minute electrode and explains why commercial barretters with much smaller electrodes than my replicas had a higher frequency response.

The difference could also account for the Delaney lamp being less efficient than the simple Fessenden type barretter, and much less efficient than the BIWC design.

Sensitivity

We could attempt to measure sensitivity of barretter type receivers in terms of input voltage. But more informative is to keep input power fixed and compare the amplitude of the output waveform of the barretter with that of a modern diode.

The wave train generator is essentially a shock-excited oscillator based on a capacitor charged to a given potential and discharged via a thyristor. So the amount of energy present in a wave train is assumed to be reasonably constant over a wide range of loads.

Substituting a point contact diode for the barretter increases the peak potential of the wave train envelope by a factor of five. At a frequency of 10kHz and a wave train repetition rate of 200Hz, the peak potential of the wave train envelope needs to be around 500mV to produce a moderately loud tone. But at 100kHz and a wave train repetition rate of 1.0kHz, a peak potential of 200mV is adequate. Apparently increased sensitivity is probably due to diaphragm resonance and to the ear being most sensitive at around 1.0kHz.

Incoming wave trains would have to contain a large amount of energy for the peak potential of the envelope to be 200-500mV. But, spark transmitters were capable of packing an enormous amount of energy into each wave train and as we have seen, could damage the minute barretter electrode, and was one reason for employing the Delaney lamp detector.

Overall, we have seen that the replica barretters, which represent c.1905 experimenters' devices, were very inefficient when compared with modern diodes.

But, the fact that a tone could still be obtained with high impedance magnetic headphones when the forward/reverse current ratio fell to about 1.2:1 at 500kHz throws new light on early rectifier type detector. ■

Oscillograms

To produce oscillograms, the hybrid barretter was used with a 5.0-500kHz wave train generator having a wave train repetition rate variable from 50Hz to 1.0kHz. (My miniature spark transmitter was used only in a railway tunnel because of the possibility of causing interference.)

The wave train generator is inductively coupled to the receiver tuner, so some degree of interaction is inevitable. True simulation of energy arriving via em waves would only be possible by to using opto/electric devices, but this was not warranted for this study.

Starting at 10kHz, the wave train generator's output frequency was incrementally increased to 200kHz and the ratio of positive-going half cycles to negative half cycles was measured directly from the oscilloscope screen and plotted on a graph.

My ancient magnetic headphones were used to evaluate the purity of the tone, and note any effect they might have on output waveform when compared with a non-inductive load resistor. It was found that replacing the headphones with a 8.0kΩ non-inductive load had only a marginal effect on waveform at frequencies below 50kHz.

But, at frequencies above 50kHz the headphones had some effect on the envelope shape, an influence that increased with frequency and was presumably caused by the back emf generated by the electromagnets.

For more information on oscillograms and related experiment, please see previous issue.

Acknowledgements

Thanks to goldsmith A Thornton, Kettering, who made and fused the platinum electrodes into the glass ware. Thanks also to the DTI for kindly allowing me to continue my barretter experiments in the railway tunnel.

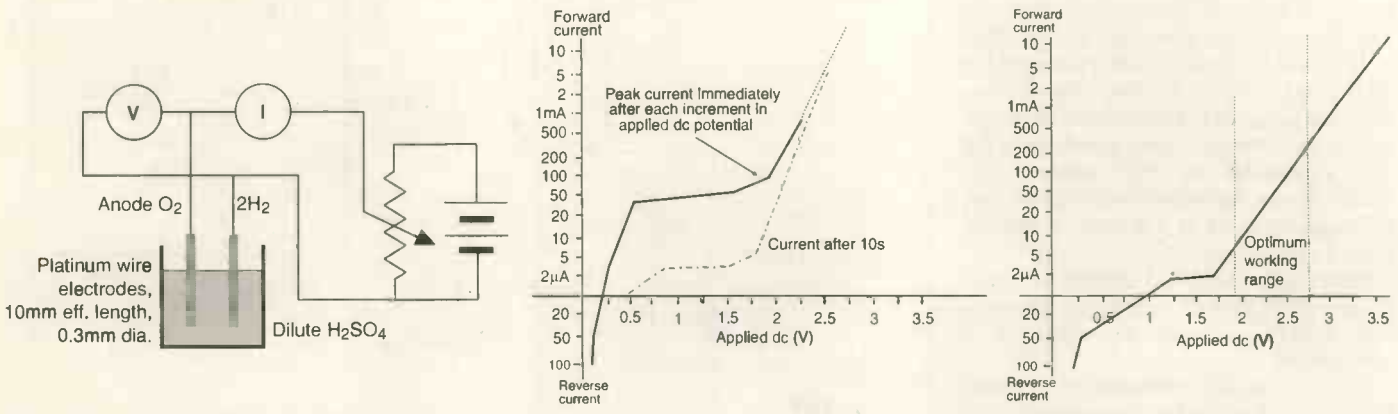


Fig. 6a. The experimental water voltameter develops a constant 0.2V even after standing for long periods with the applied dc switched off. 6b. Starting with 0.2V as the base line, 0.1V increments in the applied dc cause a momentary rise in current to a peak value which after 10s declines. 6c. In the hybrid barretter, current rises immediately to its steady value and any slight increase in applied potential causes it to rise and remain steady at a new high point.

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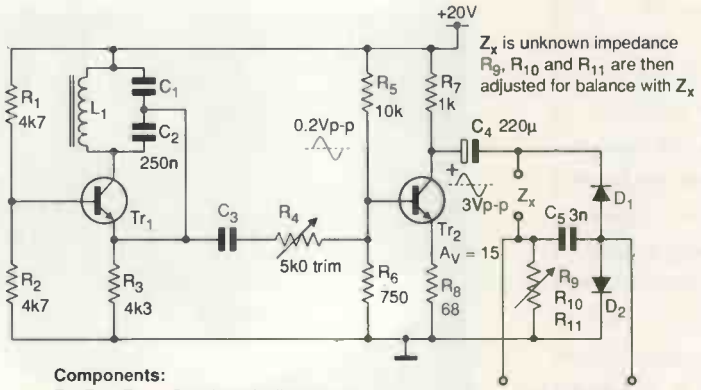
Quick and accurate impedance measurement

Having a large number of inductors with unknown impedances, I designed this instrument to allow fast, but accurate measurement up to around 10kHz; at 20kHz it is in error by about 20%.

A sinewave from the Colpitts oscillator goes to Tr_2, being adjusted to give 3V pk-pk at Tr_2 collector by R_4; aside from an amount of buffering, Tr_2 produces a gain of 15 with an output impedance of just less than 10kΩ. Capacitor C_4 provides dc isolation.

An unknown impedance goes in series with $R_{9,10,11}$, three pots wired series. Back-to-back diodes allow C_5 to charge in only one direction, so that when the three pots are adjusted for lowest voltage across C_5, the unknown is in balance with the pots, which can be either measured or calibrated.

Frank Simonsen
Burnaby
British Columbia
Canada



Components:

R_9, R_{10}, R_{11}	Consists of 3 linear pots in series 100R, 1k0, 10k0
C_1,C_3	2µ non-polarised
Tr_1, Tr_2	2N3703 general purpose transistors
L_1	100mH
D_1, D_2	1N914 silicon diodes (If greater accuracy is required diodes should be balanced)

Quicker than an LCR bridge for a large number of measurements, this meter is very accurate to about 10kHz.

Double an op-amp's output current

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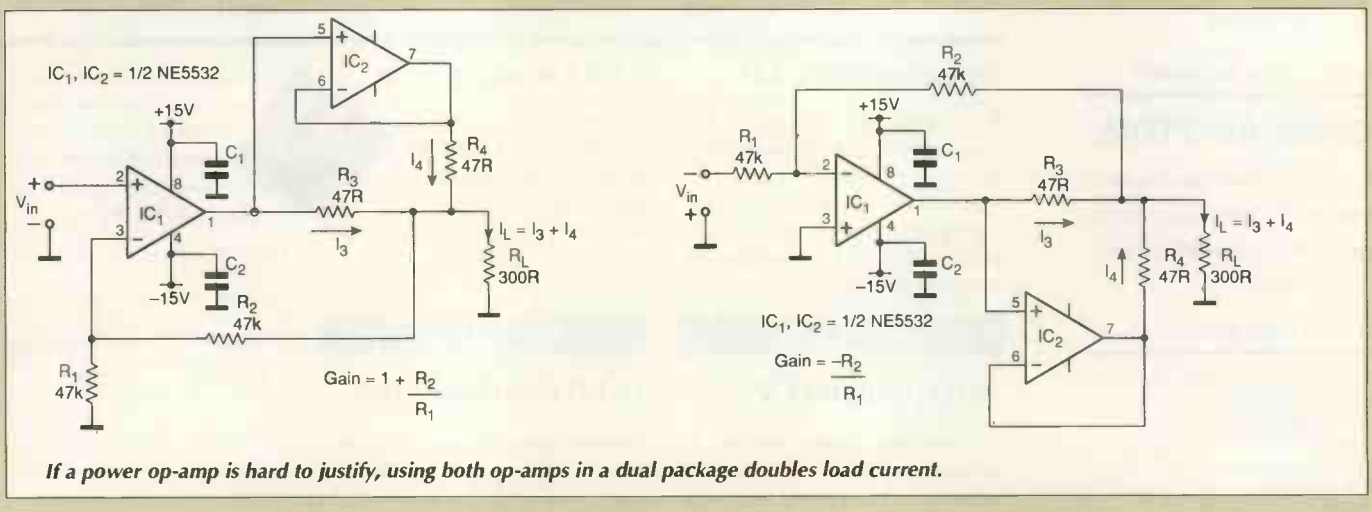
Using both the op-amps in a dual package drives a 24V pk-pk signal into a 300Ω load, doubling the output current of a single device. The arrangement serves when the use of a higher-power op-amp cannot be justified. Circuit arrangements for both polarities are shown.

Op-amp IC_2 is in the feedback path of IC_1, load current being shared in the ratio $I_4/I_3 = R_3/R_4$, currents matching if resistor values are equal. Normal output current of 20mA is increased to 40mA.

Ms Hema and V Manoharan
Naval Physics and Oceanographic Laboratory
Cochin
India



If a power op-amp is hard to justify, using both op-amps in a dual package doubles load current.

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Our judging criteria are ingenuity and originality in the use of modern components – with simplicity particularly valued.

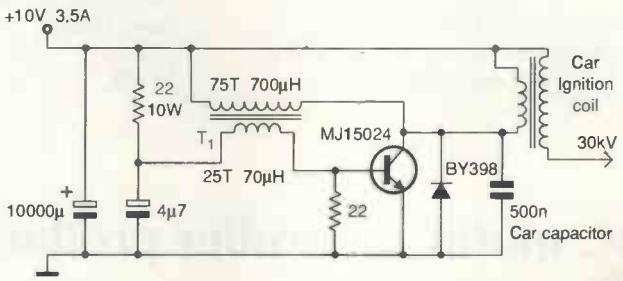
Generating high voltages

A car ignition coil and a blocking oscillator will generate pulses of about 30kV.

A separate coil must be used for the blocking action, since ignition coils only have two windings, the one shown being made from 1mm enamelled copper wire on a laminated iron rod measuring 7 by 7 by 57mm. The circuit oscillates at 1.5kHz from a supply of just a few volts up to over 12V, limits being set by the coil insulation and the transistor voltage rating. Use a 5-7°C/W heat sink for continuous working.

This arrangement will sustain a steady 0.5in spark at a 10V supply voltage.

D Di Mario
Milan
Italy



Blocking oscillator and car ignition coil produce 30kV pulses.

Frequency doubler/quadrupler for square waves

Twice and four times the input square-wave frequency are produced by two ICs and a few passive components.

Inverter 1 in **Fig. 1** integrates the square wave to produce a triangle, which is amplified by inverter 2 and squared by inverter 3. Since waveform B in **Fig. 2** is 90°phase-shifted with respect to A and C is in phase with B, C is effectively phase-shifted by 90° at one frequency. Exclusive-oring C with the input produces D, which is $2f_{in}$. An identical process is carried out by inverters 4, 5 and 6 and the second ex-or to produce $4f_{in}$.

L Szymanski
Stamford
Lincolnshire

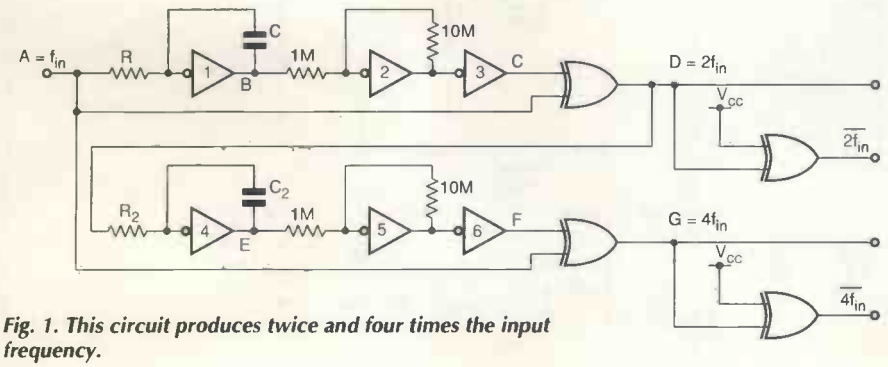


Fig. 1. This circuit produces twice and four times the input frequency.

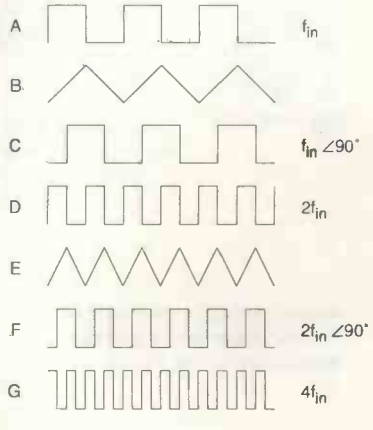


Fig. 2. Waveforms seen in the circuit of Fig.1.

Synchronised power-rail switching

Plugging a mixed-rail board into an already-powered system can cause power-supply problems, particularly where digital supplies must be applied either before or simultaneously with the analogue rails. Surges can appear in the main system. This circuit overcomes the problem.

At the core of the circuit is the Dallas *DS1233 EconoReset* microprocessor reset IC (*EW+WW* November, 1992) to delay the analogue rail. After a delay of 200ms after the 5V input in Fig. 1 rises above the IC's threshold, it turns Tr_2 on and switches on the 15V supply via Tr_1.

For dual-polarity rails, the circuit of Fig. 2 switches both together. Since the negative rail uses a p-channel fet, isolated switching is provided by the opto-isolators. Connecting the isolator leds in series is an effective fail-safe arrangement; if one led fail goes open-circuit, the other is blocked and neither supply comes on. To switch the 5V at the same time as the 15V supplies, connect a p-channel mosfet with its source to the 5V input, its drain as the output and drive the gate with Tr_3.

Steve Winder
Ipswich
Suffolk

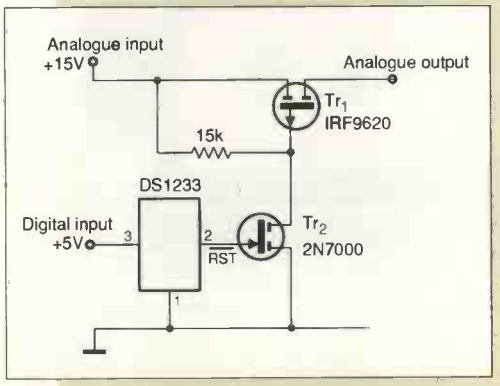


Fig. 1. Dallas EconoReset delays application of analogue supply until 5V digital rail is established.

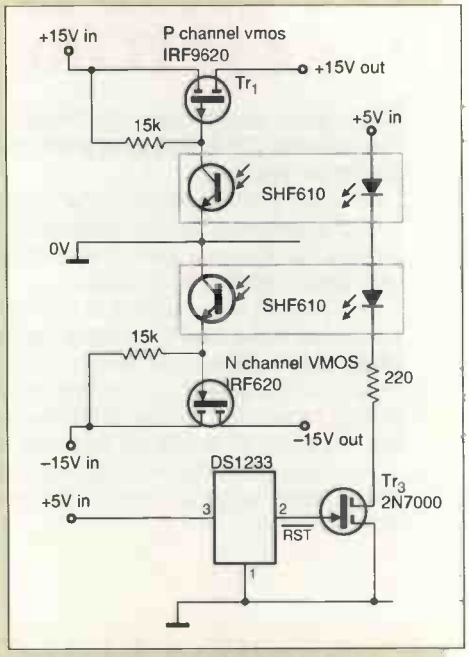


Fig. 2. Dual supply switched by the same method, but with opto-isolators to take account of the p-channel device on the negative rail.

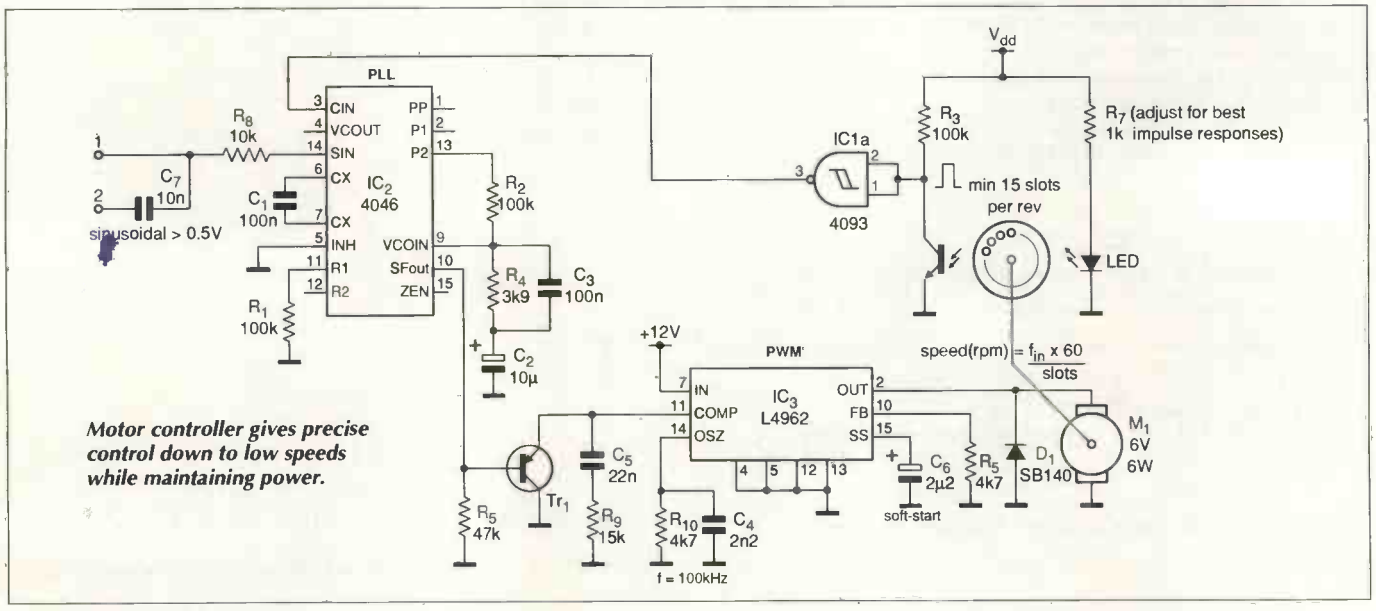
DC motor controller produces high torque at low speeds

Using a pulse-width modulator and phase-locked loop, this circuit smoothly controls a 6V dc motor, according to the input frequency, down to low speeds at high power.

Feedback to the pll is by way of a slotted disc on the motor shaft, a difference between the fed-back pulses and the input frequency causing a voltage variation into the pll and a change in pulse width driving the motor. Phase-locked-loop switching frequency is determined by the values of R_{10} and C_4.

Values of the filter components R_4 and C_2 must be selected to suit the motor in use, C_3 smoothing the pulses and reducing noise. To check operation, connect one channel of a dual-beam oscilloscope to the signal input and the other to pin 3 of the *4046*.

Ernst Schmid
Munchen
Germany



Motor controller gives precise control down to low speeds while maintaining power.

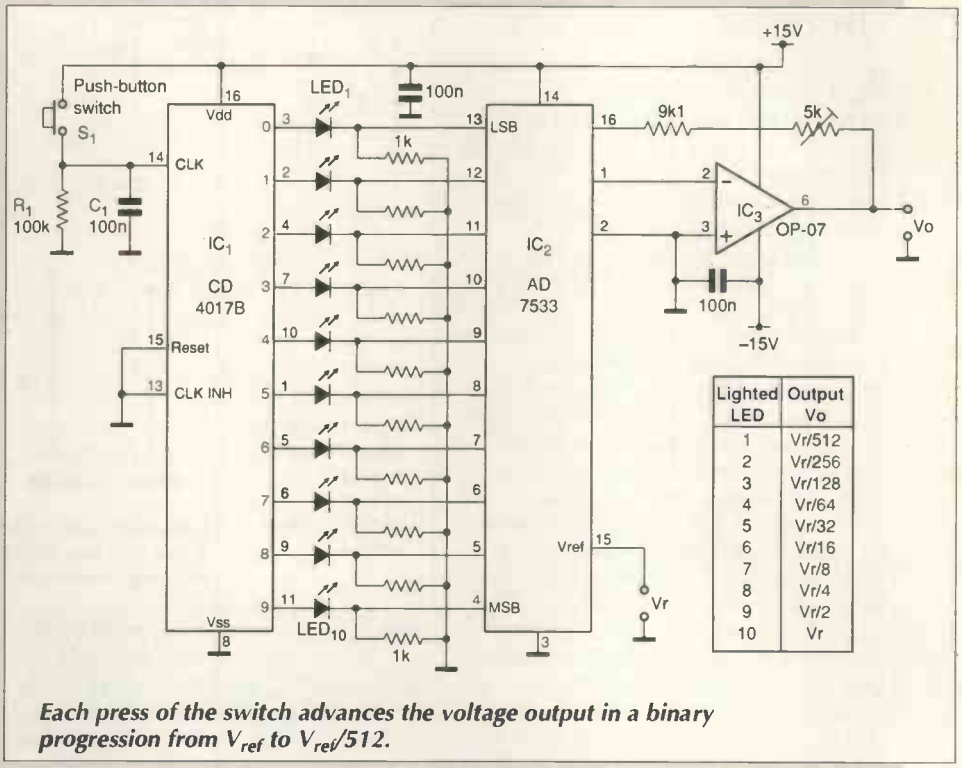
Binary voltage generator

Each operation of the push-button switch produces at the output a binary fraction of a reference voltage.

Ten decoded outputs of the *4017B* cmos counter IC_1 are input to an *AD7533* 10-bit digital-to-analogue converter, current from which is converted to voltage by the *OP-07* low offset op-amp. Only one input to the d-to-a is high at any time.

Each time the switch is pressed, the counter advances by one count and the output voltage doubles until $V_o = V_{ref}$, in the progression *V*/512, *V*/256, *V*/128, etc. Debouncing is provided by R_1 and C_1. Adjust the op-amp gain by the potentiometer until $V_o = V_{ref}$ when the counter msb is high.

M S Nagaraj
ISRO Satellite centre
Bangalore
India



Lighted LED	Output Vo
1	Vr/512
2	Vr/256
3	Vr/128
4	Vr/64
5	Vr/32
6	Vr/16
7	Vr/8
8	Vr/4
9	Vr/2
10	Vr

Each press of the switch advances the voltage output in a binary progression from V_{ref} to V_{ref}/512.

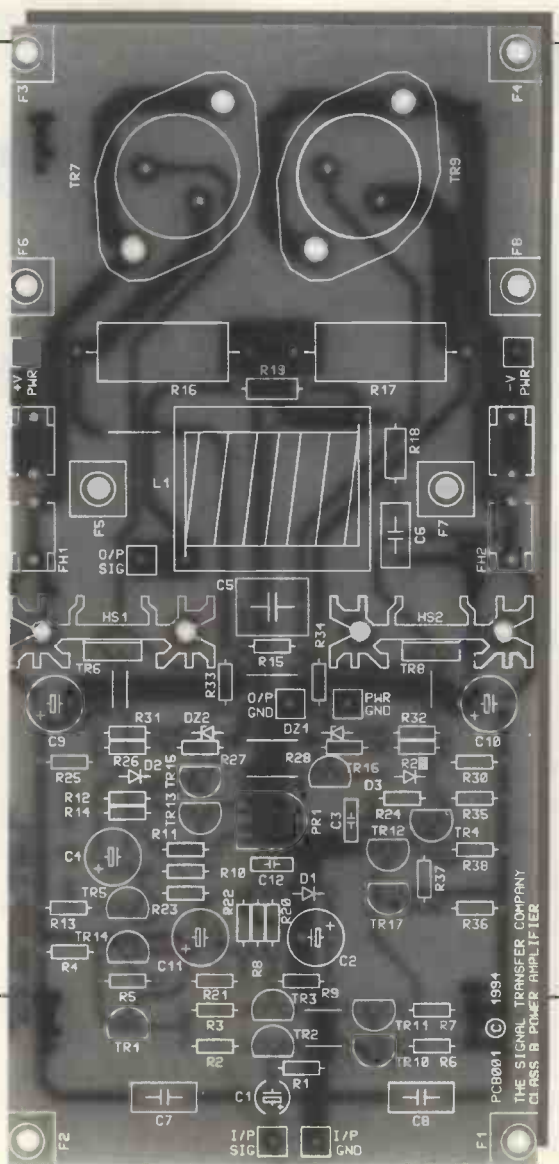
PCBs for Douglas Self's power amplifier series

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

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

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

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



Mosfet drive booster

Originally used for mosfet gate drive in a radio-control motor-speed controller, this circuit boosts gate drive from a 5-6V supply to 8-10V, providing an improvement in mosfet on resistance. Its advantages are that it needs no separate boost regulator, it produces no extra switching noise, and it takes up little space, particularly if a tantalum bead capacitor is used.

Figure 1 shows the circuit, with the switching transistor represented as a mechanical switch. When the switch is on, C charges to $V_s - V_{on} - 2V_d$, where V_{on} is the switch on resistance and V_d is a diode voltage drop. When off, V_{out} becomes $2V_s - 2V_d$. Loading the circuit does, of course, cause a drop in V_{out}, but attention to leakage paths maintains the level for many seconds or even minutes. Hold time is affected by capacitor value.

Load capacitance C_{in} and equivalent output resistance in the practical circuit of **Fig. 2** slow the mosfet's switch-on to around 4µs for a 2nF capacitor, but this has the effect of reducing rf noise and avoids simultaneous conduction in push-pull circuits, although some switching losses do occur. Turn-off time depends on the on resistance of Tr_1 and is smaller, at less than 100ns. It is possible to extend the technique to triplers, but losses are then increased.

A New
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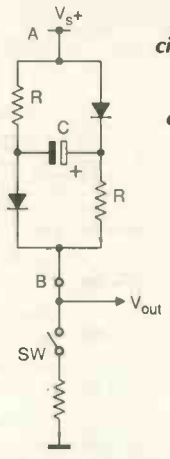
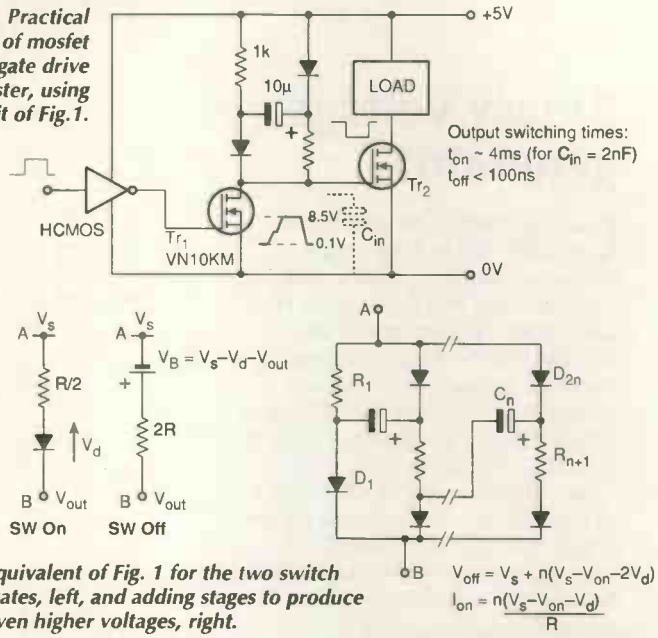


Fig.1. Charge on capacitor adds to supply voltage to increase output voltage.

Fig.2. Practical circuit of mosfet gate drive booster, using circuit of Fig.1.



Equivalent of Fig. 1 for the two switch states, left, and adding stages to produce even higher voltages, right.

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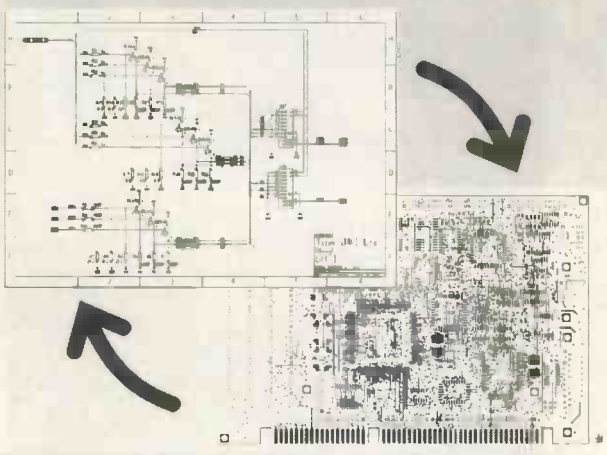

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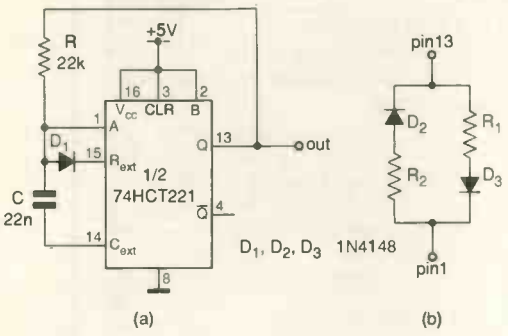
Spare monostable as oscillator

With an eye to cost and space saving, a spare one-shot multivibrator from, say, a *74HCT221* makes a convenient oscillator.

Capacitor *C* charges and discharges at the same rate, duty cycle is around 50% and *T=RC*. Substituting the diode network in (b) for the charging resistor *R* provides other duty cycles, *C* charging through R_1 and discharging through R_2 to give $T=(R_1 + R_2)C/2$ and a duty cycle approximately $R_1/(R_1 + R_2)$.

Logic families such as *HC* or *LS* series will work, but different periods and duty cycles can be expected. *HC* types need a 10kΩ resistor in series with pin 1 to avoid spurious transitions.

Giorgio Delfitto
University of Padova
Padova
Italy

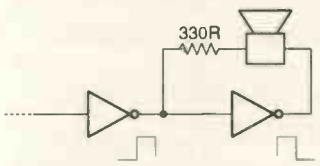


Unused monostable employed as a free-running oscillator with 50:50 duty cycle. Circuit in (b) replaces R to produce asymmetrical waveforms.

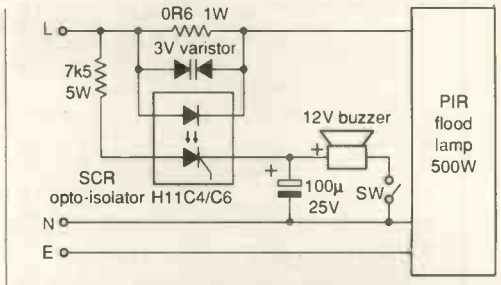
Something for (perhaps) nothing

If you need more output from a speaker driven by a logic gate, drive it differentially across input and output so that the cone moves twice as far as when it is taken to the rail. If, as is often the case, an inverter is spare, nothing more is needed.

Michael A Covington
Artificial Intelligence Center
University of Georgia
Athens, Georgia



Frugal method of obtaining more from a logic-driven speaker.



Completing the deterrent effect of a PIR alarm and floodlamp, this circuit also switches on a buzzer.

Buzzer for PIR alarms

To enhance the warning given by an infrared alarm, this circuit activates a buzzer whenever the light is illuminated.

When current is drawn by the 500W floodlamp, the voltage dropped by the 0.6Ω series resistor turns on the scr opto-isolator, which rectifies the ac to drive a 12V dc buzzer. In the original, a 7.5kΩ, 1W dropping resistor was needed, but this value depends on the type of buzzer in use. All components can be fitted in the indoor wiring to the lamp.

H P Ho
University of Nottingham
Beeston
Nottinghamshire

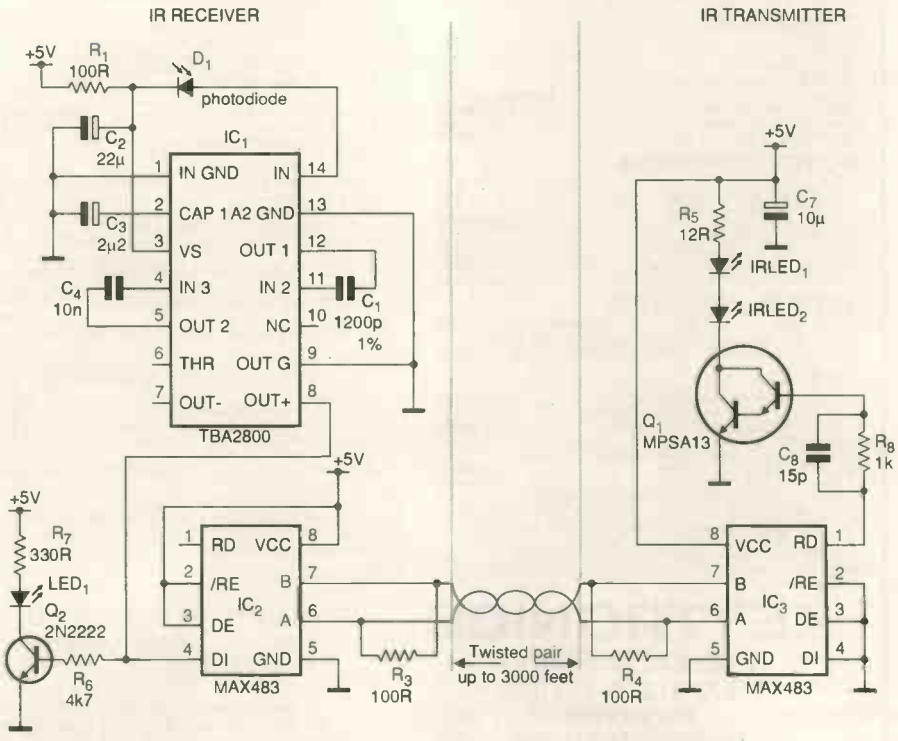
Infrared control from another room

To control a video recorder or satellite receiver by infrared, but rather more remotely than usual, this circuit uses a twisted-pair from a local infra-red receiver to the remote controller.

A local infra-red controller sends infra-red pulses to a receiver, where they are recognised by a photodiode, amplified and converted to ttl-level signals by the *TBA2800* preamplifier. These signals drive a *MAX483* RS-485 differential transmitter, which feeds a twisted-pair cable up to about 3000ft long leading to an RS-485 receiver on a *MAX483* in the other room, where the signals are again converted to ttl and used to drive the two infra-red emitters. This infra-red beam controls the remote equipment and, since it is identical with that from the original transmitter, the circuit will work with almost any type of remote control.

W H Ip
University College of Swansea

Infrared link repeater controls equipment such as a video recorder from a distance of up to 3000ft.



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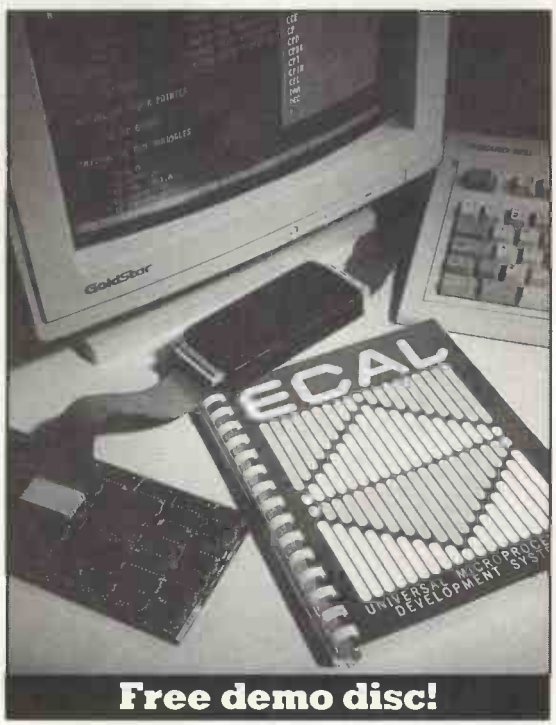

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CIRCLE NO. 118 ON REPLY CARD

Circuits by design

3: Input and output

Delving deeper into the workings of modern spice software, Owen Bishop examines the special analysis tools designed to help engineers predict precisely how well their circuit will perform.

Those of you who took the trouble to assemble the basic common-emitter amplifier described last month may well have been disappointed. To make a comparison with the simulation, we set up the 'final' design on a breadboard and found that, as predicted, V_{bb} is 0.62V. But the output voltage at the collector of Q_1 is far from the required 6V. Consequently, the collector current is much less than the predicted 0.88mA.

Since the amplifier is not at its operational point, before attempting to supply an input signal to it, the design needs further refinement. The most likely cause for the discrepancy is that the gain of the transistor on the breadboard is higher than the gain of the model transistor used in the simulation. This is a general-purpose transistor and, while it is reasonable to use it for non-critical circuits, including switching circuits, it is not a precise enough model for amplifiers.

An amplifier is intended to amplify, and any lack of precision in one part of the circuit may be amplified in other parts. If the present circuit is to be brought to a suitable operating point, then V_{bb} must be set much more precisely than is possible by taking two convenient E12 resistors from the component box. But selecting high tolerance resistors does not solve the problem. There is a spread of gain between transistors, even between transistors of the same type. We need to make the circuit less dependent on the exact gain of the transistor and on the values of the base biassing resistors. In short, the circuit needs to be better stabilised.

One approach to stabilisation is to introduce an emitter resistor, **Fig. 1**. If this is a 1kΩ resistor and the current through it is to be 0.88mA, the emitter voltage is 0.88V. Given that the emitter-base voltage of a transistor in its operating region is approximately 0.6V, the voltage V_{bb} is about 1.48V. Using the values

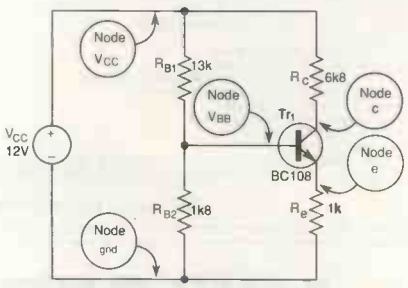


Fig. 1. Adding an emitter resistor to last month's basic amplifier circuit stabilises it; feedback makes its gain independent of the transistor's gain.

```
*  CBD03 - Common-emitter amplifier - stabilising
V Vcc     -out:gnd    +out:vcc    Ex=None   Of=12.00000
R Rb1     p1:vcc      p2:vbb      v=13.0000k
R Rb2     p1:vbb      p2:gnd      v=1.80000k
R Rc      p1:vcc      p2:c        v=6.80000k
R Re      p1:e        p2:gnd      v=1.00000k
> Q1      qbc108.lib  collector:c  base:vbb  emitter:e
```

Fig. 2. In this netlist of the stabilised amplifier of Fig. 26, we specify a model of a BC108 transistor in place of the general-purpose npn transistor used last month.

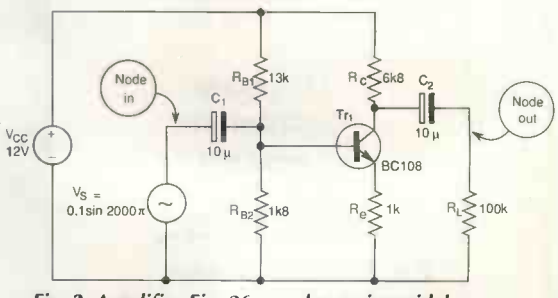


Fig. 3. Amplifier Fig. 26 now has a sinusoidal input signal applied to it, a load resistor, and a pair of coupling capacitors.

SpiceAge has no node limit

In Owen's first article, the brief background note about *SpiceAge* on the first page stated that the package could handle up to 60 nodes. This is not the case. The node capability of all but the Level 1 *SpiceAge* package is only limited by the amount of ram available in the host PC.

of the biassing resistors of Fig. 1 achieves this fairly closely.

Netlist **Fig. 2** shows the new circuit, including a *BC108* transistor. Place a probe on Node c to measure voltage. A dc quiescent analysis shows that the collector current is 793µA, voltage at c is 799mV and the base voltage is 1.45V, all reasonably close to the required values. These values are commensurate with those obtained when the circuit is breadboarded. The purpose of the emitter resistor is to provide negative feedback, to make the circuit relatively independent of the gain of the transistor.

Exploding the transistor model shows that the value of the *VCCS* is 0.992. Applying the formula quoted last month, this produces a current gain of 0.992/(1–0.992)=124. To investigate the extent to which the quiescent state of the circuit is influenced by gain, set up the program to sweep the gain over a wide range. Explode the model and edit its specification, adding 'vs=0.98 vf=0.995' This is the equivalent of sweeping the gain from 50 to 200.

Running a quiescent sweep produces a graph in which the collector voltage ranges linearly from 6.77V for a gain of 50 to 6.57V for a gain of 200. The voltage changes by only 0.2V for a fourfold increase of gain. The circuit would operate in much the same way with many differing types of transistor with differing gains. For the same sweep, the collector current ranges between 769µA and 799µA. This analysis clearly and quickly illustrates that the gain of the amplifier is virtually independent of the gain of the transistor. Such demonstrations emphasise how valuable simulation software can be as a teaching aid.

Delete the start and finish values from the model, Implode the model, and cancel the value sweep.

Transient behaviour

To provide input to the circuit and to obtain output from it, add a capacitor to couple the circuit to an alternating voltage source and a second capacitor to couple it to a load resistor, **Fig. 3**. The voltage source has a sine waveform at 1kHz, and 100mV amplitude. Place probes to measure input voltage (at node V_{in}) and output voltage (at node V_{out}).

In the Time Selector dialogue box make Start time zero, Stop time 5ms and Step time 50µs. This samples the waveforms 100 times during 5 complete cycles. A Transient analysis gives **Fig. 4**. The curve of lesser amplitude is the input sine wave. The curve of greater amplitude is the output sine wave. It is 180° out of phase, as expected. Its amplitude is about 6.14 times that of the input signal, so the voltage gain of the amplifier is much lower that the gain of transistor. This is because of the feedback effect of R_e, and is the price to be paid for stability.

Figure 4 shows that the output swing is only ±600mV, so there is scope for increasing input signal amplitude without introducing distortion. **Fig. 5** shows the result of a value sweep carried out for the amplitude source generator.

Fig. 4. A transient analysis of the amplifier of Fig. 1. at 1kHz shows the output signal 180° out of phase with the input signal, and a gain of just over 6.

The range is specified by adding 'vs=100m vf=500m' to the netlist statement. At the sweep dialogue box select Value sweep, the Linear option, and 10 steps. It is clear that input amplitude may be as great as 300mV without introducing distortion.

Editing the netlist again, to set the value sweep range from 500mV to 1.5V, produces the transient analysis of **Fig. 6**. Clipping affects the outputs of greatest amplitude. The maximum amplitude without clipping is about ±900mV.

Frequency response

To investigate the behaviour of the circuit over a useful working range of frequency, the single transient analysis with 100mV input is repeated with the generator frequency increased to 1MHz. At the Time Selector box the values were altered to Start time = 0, Stop time = 5µs, Step time = 50ns. Remember that the 's' for 'seconds' must not be included when keying in the times. This again allows enough time for 5 complete cycles. The curves are virtually identical to Fig. 4, except that the gain is not quite as high. High frequency response is acceptable. Next, check on response at the low end of the audio range, 30Hz. Specify Start time = 0, Stop time = 166.66667ms, and Step time = 1.66667ms. Once again, curves similar to Fig. 4 are obtained, though the gain is only 5.85. Obviously there is a falling off of gain below 30mHz and above 1MHz.

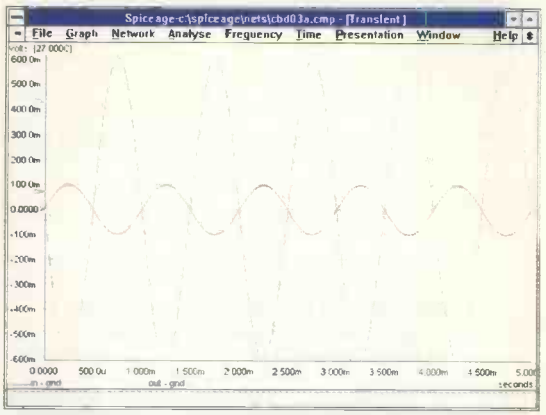
To study the how gain varies over a range of frequencies, use the Frequency response analysis. In the Frequency selection box set the Start frequency to 25Hz, the Stop frequency to 1MHz and the number of steps to 200. The result is **Fig. 7**, showing a fall in response at both ends of the range, but a level response between about 500Hz and 100kHz. Note that the scale is logarithmic on the x-axis. The scale on the y-axis is in volts but it can optionally be graduated in decibels.

Increasing gain

Voltage gain obtained from this circuit, at about six, is disappointing. This low figure is due to the feedback effect of the emitter resistor. The resistor not only compensates for differences in the gain of different transistors but at the same time 'compensates' for voltage changes caused by the audio signal. To a large extent it cancels out the signal, resulting in low gain.

The cure for this is a by-pass capacitor connected across the resistor. Its passes the audio-frequency signals to ground, so that their effect is not fed back to the base terminal of the transistor and they are not attenuated. **Fig. 8**

Fig. 7. Frequency response curve from 25Hz to 1MHz is level from 500Hz to 100kHz.



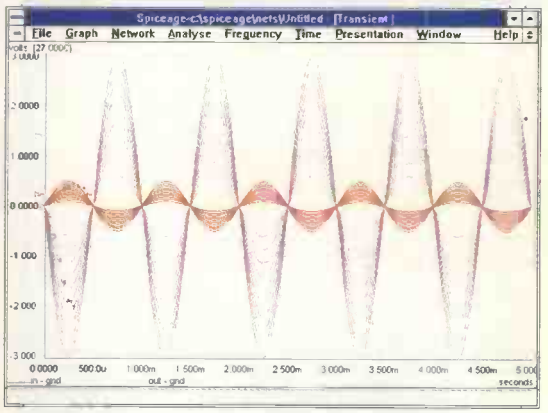


Fig. 5. Simulators make it easy to sweep circuit parameters over a chosen range and observe the results. Here, input amplitude is swept from 100mV to 500mV in 10 linearly spaced steps. There is no distortion in the output signal.

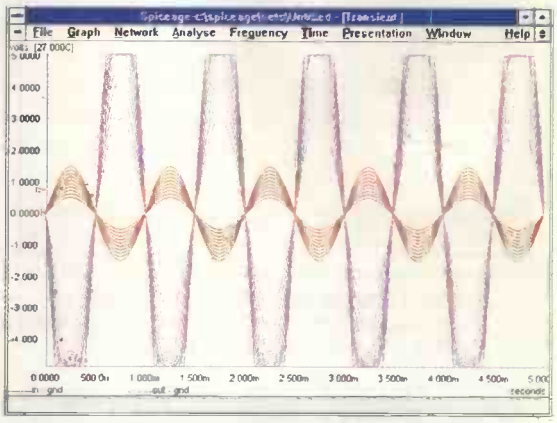
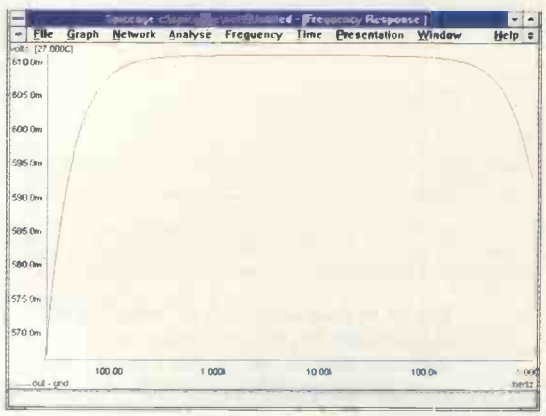


Fig. 6. Sweeping input amplitude from 500mV to 1.5V shows clipping of the output signal when input exceeds ±900mV.



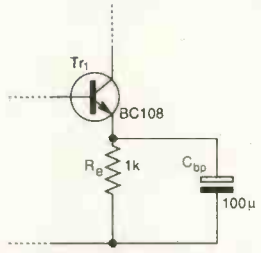


Fig. 8. Circuit Fig. 28 is given negative feedback by the addition of a by-pass capacitor across R_E.

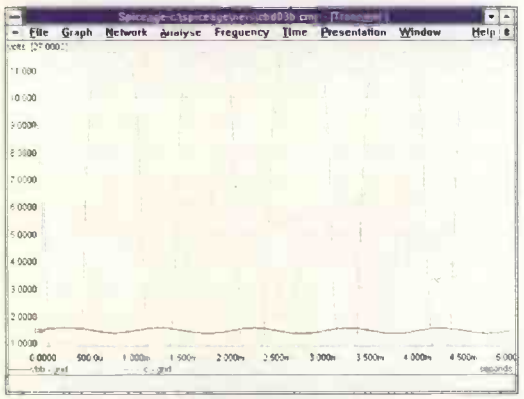


Fig. 9 The effect of the by-pass capacitor is to increase gain so much that the output signal is severely clipped.

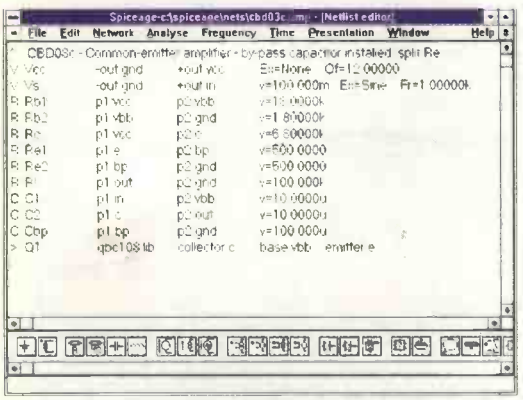


Fig. 10. In this netlist of Fig. 28 plus the by-pass capacitor, R_E is tapped half-way by listing it as two resistors, R_{e1} and R_{e2}, each of 500.

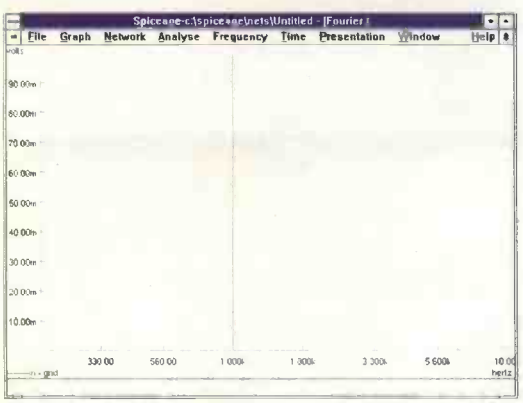


Fig. 11. Fourier analysis of the output of the signal generator produces a frequency spectrum which confirms that it is pure sine wave, frequency 1kHz, amplitude 100mV.

shows the location of the by-pass capacitor. Add this to the netlist and re-run the transient analysis; the result is **Fig. 9**.

Gain is increased to such an extent that the output signal is clipped in both directions. Obviously, we need a response somewhere between that of Fig. 4 and that of Fig. 9. The solution is to connect the by-pass capacitor across part of R_e. This is done in the netlist by setting up two resistors, R_{e1} and R_{e2}, as **Fig. 10**. For successive analyses, edit the values of R_{e1} and R_{e2}, keeping their total equal to 1kΩ. Starting with R_{e1}=R_{e2}=500Ω, gain becomes 11.9. With R_{e1}=300Ω and R_{e2}=700Ω, gain increases to almost 19. With R_{e1}=200Ω and R_{e2}=800Ω, gain is 26, but the output signal is slightly asymmetrical. In the positive direction it swings to 2.61V (above its approximately 6V central value). In the positive direction it swings down to –2.74V. There is no clipping and distortion is relatively insignificant. With R_{e1}=100Ω and Re_2=900Ω, gain is nearly 45, and there is no clipping, but asymmetry becomes more apparent (+3.96, –4.96). In the end, we settled for R_e=250Ω, R_e=750Ω, with almost perfect symmetry and a gain of 22. This is a great improvement on the original gain of 6 for this circuit.

As a final check, run the analysis with probes set to measure power in R_c, R_{e1} and R_{e2}. Power in R_c and R_{e1} is negligibly low, but in R_{e2} it peaks at 850mW. Provided that a 0.6W resistor is used for R_{e2}, there should be no over-heating.

Fourier analysis

Although the output curve in Fig. 4 appears to be a sine wave, there is no way in which this can be confirmed by visual inspection. A Fourier analysis is used to check on its purity, or otherwise. Various options are available in *SpiceAge*; the following analysis has Autodisperse deselected so that we can set the frequency range to start from 100Hz and extend up to 10kHz. In this application, there is little of interest outside this range.

Phase plot and decibel plot are also deselected, to obtain a straightforward graph, indicating the amplitudes of the component waves. The Lines to origin option is selected so that the display appears as a frequency spectrum which, in general, is the most easily understood and informative of the various display modes. The grid option for graph plotting is deselected because vertical lines that coincide with grid lines are printed out as inverse dashed lines, practically indistinguishable from ordinary grid lines, which makes the printout almost impossible to read.

As a check on the purity of the signal from the sine wave generator, set up a circuit consisting of the same generator, with a 100kΩ resistor connected across its output. Run a Transient analysis – always required prior to a Fourier analysis – and obtain a pure (?) sine wave.

In all the analyses, the transient is timed to last for five cycles. It is important to run for several cycles and for a time that is an exact number of cycles. This 'informs' the Fourier

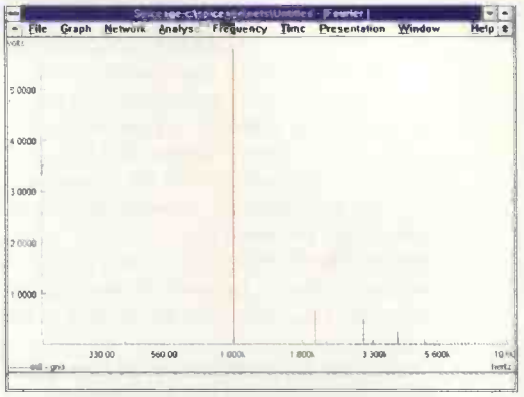


Fig. 12. In contrast to Fig. 36, the amplifier has reasonably high gain but is distorted by the presense of the first 400 harmonics.

analysis that the waveform is repetitive and prevents spurious harmonics being generated by the odd fraction of a cycle at the end of the run. Then run a Fourier analysis, **Fig. 11**; this shows a single frequency: 1kHz, amplitude 100mV, so there are is no doubt that the input to the circuit is a pure sine wave.

Returning to the circuit of Fig. 10, with R_{e1}=250Ω and R_{e2}=750Ω, place Probe 1 to measure the voltage at node 'in'. Run a Transient analysis, followed by Fourier. The result is the same as in Fig. 9, a pure 1kHz sine wave. Change the probe to measure output voltage at node 'out'.

Mathematica conventions

User variables and names are lower-case *Mathematica* variables and objects begin with upper-case letter, e.g. Pi, E.

Equals symbol usage is as follows

= assigns a value to a variable immediately

:= assigns a value to a variable each time the statement is encountered

== mathematical equality of two sides of an equation.

Round brackets are used in equations while square brackets define commands. Curly brackets include lists and the percentage symbol represents for the result of the previous calculation.

Multiplication is implied when two variables have a blank space between them; or use an asterisk.

The statement,

f[variable1_, variable2_, variable 3_]

is a function of the variables. It can be assigned a value calculated by one or more routines, enclosed in round brackets and separated by semi-colons. To call the function give its name and variable values.

The Fourier analysis shows the 1kHz signal coming through strongly, amplitude 2.2V, but with the first harmonic of 2kHz showing slightly, amplitude 29mV. For comparison, edit the netlist to make R_{e1}=50Ω and R_{e2}=950Ω and re-run the tests. Now the 1kHz signal is accompanied by a number of harmonics, mainly the 1st and 2nd harmonics, as shown in **Fig. 12**. Gain is high but distortion is becoming unacceptable.

Computation routines

This is intended as an example of the way *Mathematica* is used when there is a preferred routine for circuit design. We carry out a sequence of calculations in order, the result of one calculation providing values to insert into the next.

In this example, we assume that the values of the resistors have already been chosen so as to bring the circuit to a suitable operating point. We could, of course, have previously set up a routine for deciding on these values and this too would be an example of a computation routine. Here we intend to study the response of the circuit of Fig. 3 (plus the by-pass capacitor of Fig. 8). We are looking at its response to small-amplitude ac input, so consequently we can replace the capacitors with short circuits, as in **Fig. 13**. Resistor R_e is shown as a variable resistor, with the wiper connected to ground representing the by-pass capacitor. In the calculations, R_e is the resistance between the wiper and the positive end of the resistor. The output impedance of the source is so small that it is ignored.

The sequence of calculations is set out in **Fig. 14** as a *Mathematica* notebook. In effect, it is a program for calculating ic. On the first few trials, take the steps one at a time. This gives insight into the way the software works, and also makes it possible to monitor the validity of the results. It is also better to use step-wise evaluation when checking the program for bugs. At each stage, click on the appropriate line, then on the Evaluate button.

The first seven stages invite the user to key in the relevant values. Each stage calls up a dialogue box into which values are entered (incidentally, there is no need to click on the 'OK' button; just press Enter). After each input command there is a command of the form ' rb1= %'. This assigns the entered value to the variable. The value for re is 1000, or less to make it partly by-passed.

The next six stages set out the sequence of calculations:

- v_{bb} is the base voltage
- r_b is the resistance of r_{b1} and r_{b2} in parallel
- i_{cq} is quiescent collector current, assuming that base-emitter voltage v_{be} is 0.7V.

Having found i_{cq} in this way we then use its value to calculate a more precise figure for v_{be}. This is used in the next stage to re-calculate i_{cq} more precisely. Strictly, this pair of equations should be iterated until there is no change in the values of i_{cq} and v_{be}. But in practice the values converge only slowly and a single iteration gives values precise to about three significant figures. av is the voltage amplification of the circuit.

All except the last of these equations is followed by a semi-colon. The effect of this is to suppress the output of the result, which is held in memory and used in the next equation. When debugging the program the semi-colons are omitted, so that the result of each stage can be checked. There is no semicolon on the last equation so the value of av is displayed as the last equation is evaluated. After that, click and evaluate icq, calculated earlier but displayed at the end.

This may be as far as you need to go but it is now possible to input a signal voltage, and have the corresponding output voltage displayed.

These equations are not a circuit simulation, so no account is taken of the inner workings of transistors. It is possible to input an unrealistically large signal voltage such that the calculated output voltage exceeds v_{cc}, or is less than v_{be}. Also it is essential that r_b is less than $0.1 \times h_{fe} \times r_e$. The final statement in the program checks that results are valid and realistic. In effect it is saying ' IF rb < 0.1 × hfe × re AND vout < 12 AND vout > vbe THEN print 'True', ELSE print 'False'. If 'False' is displayed at the end of a run, the results are not acceptable.

For more frequent use, evaluate the Notebook in a single operation; Select Action, then Evaluate notebook. The computer runs rapidly through the sequence, demanding input and displaying all results on the way.

The program is easily modified to suit the user. It may be that only the values of re and vin are to be varied. Having to input the same values of the other variables each time is a tedious matter.

Figure 15 shows how the program may be streamlined. In the first block of the program we create a procedure called 'output', listing the symbols for all the variables that are to be input to it. This is followed by all the equations, separated by semi-colons and enclosed in round brackets. After this, the user is asked to enter re and hfe.

Next comes the 'output' procedure with a bracketed list of the values of the variables. Clicking on and evaluating this causes the procedure to be run, using these values and also the two previously input values. The result is the value of the final equation, av. We have also asked for '1000 icq' so as to display the quiescent collector current in milliamps. The routines for finding the vout for any given vin are the same as before. This Notebook, too can be run in a single operation.

The procedure is easily modified for calculations based on other circuit equations, or to accept other input values, or to return other outputs. In contrast to the predetermined routines of simulation programs, even with their options, Mathematica offers a flexible , DIY approach to computation. Next month we look at some Mathematica routines specially written for electrical and electronic engineering.■

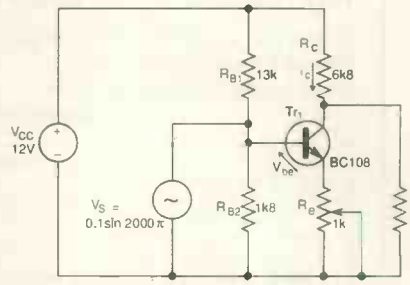


Fig. 13. The circuit of Fig. 28 redrawn without capacitors, to facilitate its analysis by Mathamatica.

```
Input["Base resistor 1"];
rb1 = %;
Input["Base resistor 2"];
rb2 = %;
Input["Emitter resistor"];
re = %;
Input["Collector resistor"];
rc = %;
Input["Load resistor"];
rl = %;
Input["Supply voltage"];
vcc = %;
Input["hfe"];
hfe = %;

vbb = rb2 vcc/(rb1 + rb2);
rb= rb1 rb2/(rb1 + rb2);
icq = (vbb - 0.7)/(rb/hfe + re);
vbe = Log[icq/10^-14]/38.647;
icq = (vbb - vbe)/(rb/hfe + re);
av = -(rl rc)/(rl + rc)/((0.026/icq) + re)
icq
Input["vin"]
vin = %;
vout = 12 - icq rc + av vin

If[rb < 0.1 hfe re && vout<12 && vout>vbe, True,
 False]
```

Fig. 14. The first section of this program asks for the relevant values to keyed in from Fig. 13. The next section calculates icq, the quiescent collector current. Then, given V_{IN}, the routine calculates V_{OUT}. The last line checks that the values used are valid ones.

```
output[rb1_,rb2_,rc_,rl_,vcc_,hfe_] :=
(vbb = rb2 vcc/(rb1 + rb2);
rb= rb1 rb2/(rb1 + rb2);
icq = (vbb - 0.7)/(rb/hfe + re);
vbe = Log[icq/10^-14]/38.647;
icq = (vbb - vbe)/(rb/hfe + re);
av = -(rl rc)/(rl + rc)/((0.026/icq) + re))

Input["Emitter resistor"];
re = %;

Input["hfe"];
hfe = %;

output[13000,1800,6800,100000,12,124]

1000 icq


Input["vin"];
vin = %;

vout = 12 - icq rc + av vin

If[rb < 0.1 hfe re && vout<12 && vout>vbe, True,
 False]
```

Fig. 15. Here the program of Fig. 14 is simplified by defining a procedure, called output, to find icq. Given values of R_E, h_{FE} and V_{IN}, output calculates V_{OUT}.

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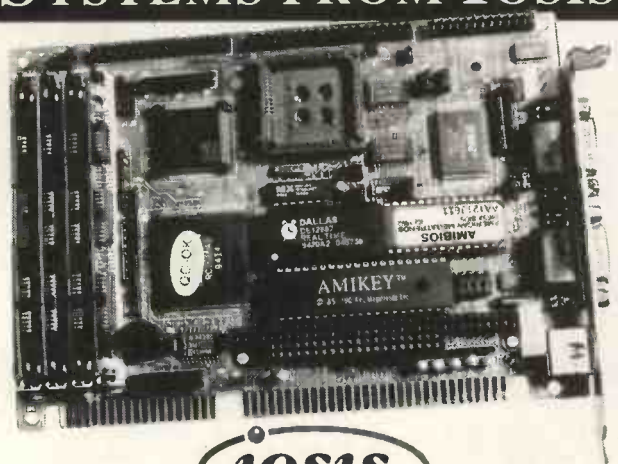

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HP 8410 – A – B – C Network Analyzer 110Mc/s to 12GHz or 18GHz – plus most other units and displays used in this set-up – 8411a – 8412 – 8413 – 8414 – 8418 – 8740 – 8741 – 8742 – 8743 – 8746 – 8650. From £1000.
Racal/Dana 9301A – 9302 RF Millivoltmeter – 1.5-2GHz – £250-£400.
Racal/Dana Counters 9915M – 9916 – 9917 – 9921 – £150 to £450. Fitted FX standards.
Racal/Dana Modulation Meter type 9009 – 8Mc/s – 1.5GHz – £250.
Marconi RCL Bridge type TF2700 – £150.
Marconi/Saunders Signal Sources type – 6058B – 6070A – 6055A – 6059A – 6057A – 6056 – £250-£350. 400Mc/s to 18GHz.
Marconi TF1245 Circuit Magnification meter + 1246 & 1247 Oscillators – £100-£300.
Marconi microwave 6600A sweep osc., mainframe with 6650 PI – 18-26.5GHz or 6651 PI – 26.5-40GHz – £1000 or PI only £600. MF only £250.
Marconi distortion meter type TF2331 – £150. TF2331A – 200.
Tektronix Plug-Ins 7A13 – 7A14 – 7A18 – 7A24 – 7A26 – 7A11 – 7M11 – 7S11 – 7D10 – 7S12 – S1 – S2 – S6 – S52 – PG506 – SC504 – SG502 – SG503 – SG504 – DC503 –DC508 – DD501 – WR501 – DM501A – FG501A – TG501 – PG502 – DC505A – FG504 – 7B80 + 85–7B92A
Gould J3B test oscillator + manual – £200.
Tektronix Mainframes – 7603 – 7623A – 7613 – 7704A – 7844 – 7904 – TM501 – TM503 – TM506 – 7904 – 7834 – 7104 – 7623 – 7633.
Alltech 757 Spectrum Analyser – 001 22GHz – Digital storage + readout – £2000.
Marconi 6155A Signal Source – 1 to 2GHz – LED readout – £400.
Barr & Stroud Variable filter EF3 0.1Hz – 100kc/s + high pass + low pass – £150.
Marconi TF2163S attenuator – 1GHz. £200.
Farnell power unit H60/50 – £400 tested. H60/25 – £250.
Racal/Dana 9300 RMS voltmeter – £250.
HP 8750A storage normalizer – £400 with lead + S.A or N,A Interface.
Marconi TF2330 – or TF2330A wave analysers – £100-£150.
Racal/Dana signal generator 9082 – 1.5-520Mc/s – £500.
Racal/Dana signal generator 9082H – 1.5-520Mc/s – £600.
Tektronix – 7S14 – 7T11 – 7S11 – 7S12 – S1 – S2 – S39 – S47 – S51 – S52 – S53 – 7M11.
Marconi mod meters type TF2304 – £250.
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Systron Donner counter type 6054B – 20Mc/s – 24GHz – LED readout – £1k.
Racal/Dana 9083 signal source – two tone – £250.
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Tektronix TM515 mainframe + TM5006 mainframe – £450 – £850.
Rhodes & Schwartz power signal generator SLRD-280 – 2750Mc/s – £250-£600.
Ball Efratom rubidrum standard PT256B-FRKL – £1000.
Farnall electronic load type RB1030-35 – £350.
Racal/Dana counters – 9904 – 9905 – 9906 – 9915 – 9916 – 9917 – 9921 – 50Mc/s – 3GHz – £100-£450 – all fitted with FX standards.
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Marconi TF2092 noise receiver. A, B or C plus filters – £100-£350.
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Fluke 8506A thermal RMS digital multimeter. £400.
Philips panoramic receiver type PM7900 – 1 to 20GHz – £400.
Marconi 6700A sweep oscillator + 6730A – 1 to 2GHz – £500.
HP8505A network ANZ + 8503A S parameter test set + 8501A normalizer – £4k.
Racal/Dana VLF frequency standard equipment. Tracer receiver type 900A + difference meter type 527E + rubidium standard type 9475 – £2750.
HP signal generators type 626 – 628 – frequency 10GHz – 21GHz.
HP 432A – 435A or B – 436A – power meters + powerheads – Mc/s – 40GHz – £200-£1000.
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Marconi TF2370 spectrum ANZ + TK2375 FX extender 1250Mc/s + 1st gen – £1.5k.
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HP 8671A syn microwave 2 – 6.2GHz – £2k.
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HP 8640B S/G AM-FM 512Mc/s or 1024Mc/s. Opt 001 or 002 or 003 – £800-£1250.
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Philips PM5390 RF syn – 0.1 – 1GHz – AM + FM – £1250.
Philips PM5519 Colour T.V. pattern generator – £250.
S.A. Spectral Dynamics SD345 spectrascope 111 – LF ANZ – £2500.
Tektronix R7912 Transient waveform digitizer – programmable – £400.
Tektronix 496 Analyzer 1kHz – 1.8GHz – £3.5k.
Tektronix TR503 + TM503 tracking generator 0.1 – 1.8GHz – £1k – or TR502.
Tektronix 576 Curve tracer + adaptors – £900.
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Tektronix 1502/1503 TDR cable test set – £1000.
Tektronix 7L5 LF analyser – 0 – 5Mc/s – £800. OPT 25 – £1000.
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Tek 2445 150Mc/s oscilloscope – £1400.
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Fluke 510A AC ref standard – 400Hz – £200.
Fluke 355A DC voltage standard – £300.
Schlumberger 5229 Oscilloscope – 500Mc/s – £500.
Solartron 1170 FX response ANZ – LED dislay – £280.
Wiltron 610D Sweep Generator + 6124C PI – 4 – 8GHz – £400.
Wiltron 610D Sweep Generator + 61084D PI – 1Mc/s – 1500Mc/s – £500.
Time Electronics 9814 Voltage calibrator – £750.
Time Electronics 9811 Programmable resistance – £600.
Time Electronics 2004 D.C. voltage standard – £1000.
HP 8699B Sweep PI YIG oscillator .01 – 4GHz – £300. 8690B MF – £250. Both £500.
Schlumberger 1250 Frequency response ANZ – £2500.
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CIRCLE NO. 124 ON REPLY CARD

PCBs through windows

At the heart of Powerware's range of windows-based pcb design products is a core called Quickroute. This core has recently been upgraded to version 3.0, chiefly to cope with larger pcb designs, as Martin Cummings explains.

Quickroute started life as shareware. Two software generations later, it is being sold as a professional cad package. It provides pcb layout and schematic capture in several forms depending on the complexity of your design and price you want to pay. *Quickroute* runs under windows, and it gets top marks for efficient programming. There can be few windows programs these days that can be supplied on one 720Kbyte disk.

One of the chief advantages of the windows platform is that *Quickroute* can output to any device, printer or plotter that comes with a windows driver. It can handle up to eight layers plus two silk screens and all versions come with an autorouter, although the competence of the router varies as you go up through the product range.

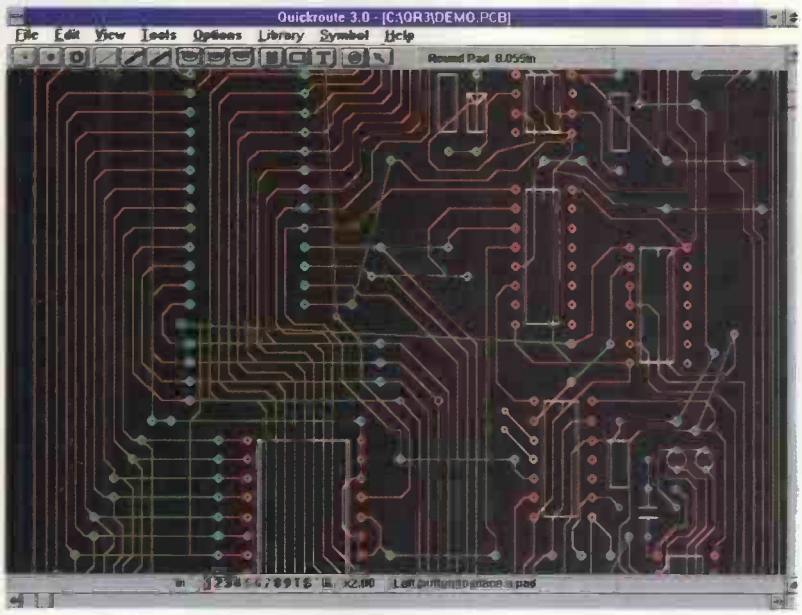
The trouble with windows is that it makes you think you can use software right from the word go. This is not the case. Don't misunderstand me, *Quickroute* is easy to use, and makes full use of the standard windows interface, but if you don't read the manual the interesting designs on the screen will not live up to expectations when you try to do something clever.

The *Quickroute* screen is exactly what you would expect. Design area is surrounded by a menu bar, button bar, scroll bars and a status line at the bottom of the screen. Current working environment is reported on the status line. It show the current layer, the *xy* coordinates of the cursor and the current zoom level.

Cursor snap can be set from 1in down to 0.001in and a reassuring click is given at each step of the cursor. A grid can be displayed and in keeping with previous versions of *Quickroute* you have no control over the resolution, it adjusts itself according to the zoom level. This sounds disconcerting but is surprisingly acceptable in practice.

There are two ways to zoom. A pull down menu provides a selection of magnification factors from 0.15 to 10 in seven steps. The function keys provide a single key press alternative to the menus. You can also type in your own magnification factor in the unlikely event that one of the standard ones is unacceptable.

While zooming is not a problem, the ability to define a zoom rectangle would be a natural way to zoom and could be a useful addition.



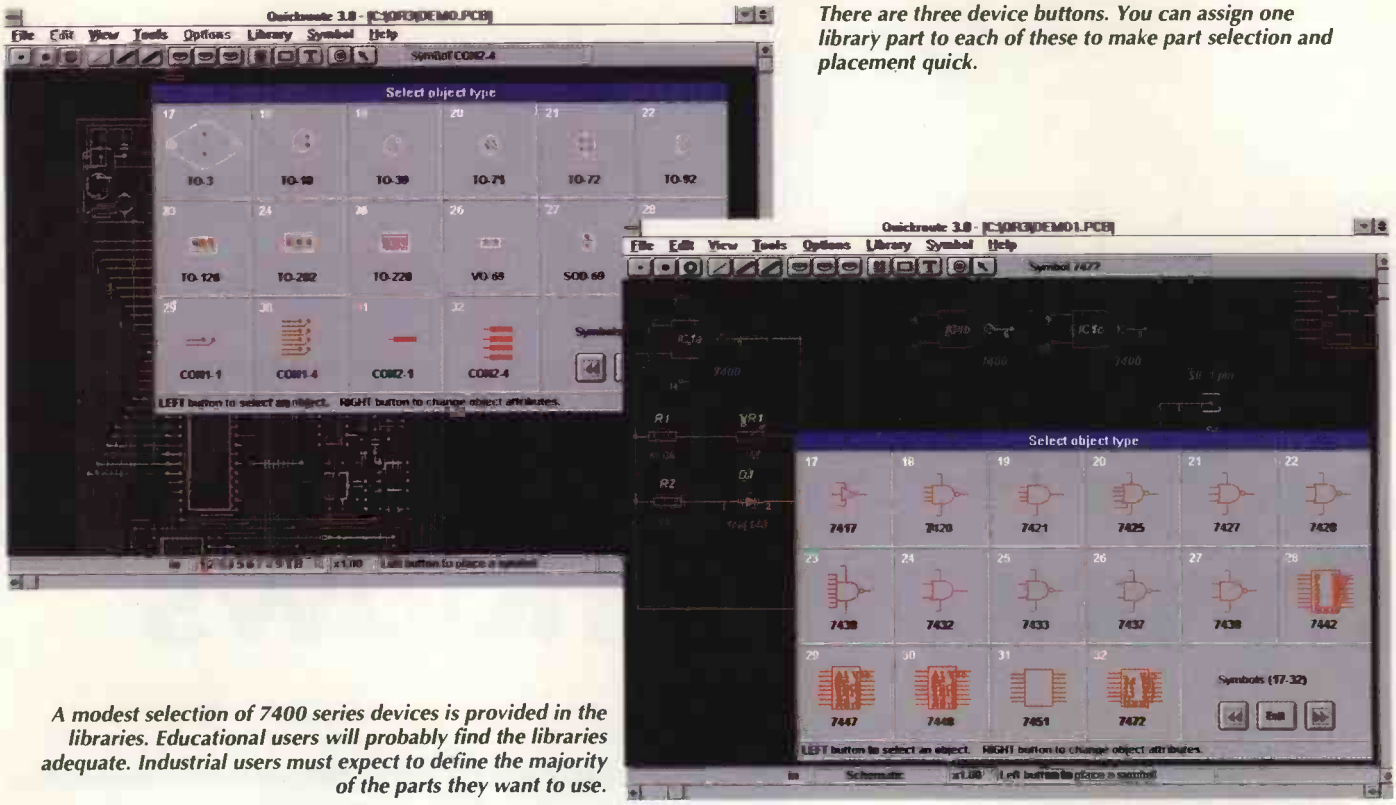
Quickroute 3 has evolved as a pcb layout tool but it can also turn its hand to schematic design.

Capturing schematics

In any design, the first step is to draw a schematic, so why is it that the tutorial in the manual jumps into pcb layout and treats schematic capture as an afterthought? Eventually chapter 4 of the manual gets round to explaining how to design the circuit.

The first decision is that of library choice. *Quickroute* is supplied with a number of libraries. For analogue designs and simple logic work (ie not/nand/nor etc) the supplied libraries will usually suffice, there is a good chance you will find what you want. If you need to go beyond the *741* op-amp and the most popular *7400* series devices then be prepared to define the part yourself.

An up to date listing of library components is provided via a read-me file. It is useful to have a print of this to



There are three device buttons. You can assign one library part to each of these to make part selection and placement quick.

A modest selection of 7400 series devices is provided in the libraries. Educational users will probably find the libraries adequate. Industrial users must expect to define the majority of the parts they want to use.

hand. Nevertheless a little exploration is necessary to confirm exactly what you want to select.

Select a library then one of the three symbol buttons from the button bar. You can allocate a different symbol to each of the three, so you could have say a transistor, a resistor and a capacitor at your finger tips. Beyond this you have to assign another symbol to one of the buttons. A right mouse click on a button opens an array of possibles to choose from.

Quickroute versions are scaled in price versus features.

	Designer	Designer+	Curriculum	Pro	Pro+
Schematic capture	no	no	max 20 symbols	max 250 symbols	max 500 symbols
Layers	8+2 silk screen	8+2 silk screen	8+2 silk screen	8+2 silk screen	8+2 silk screen
Design size (pads)	10 000	20 000	10 000	30 000	40 000
Design size	30x30 inches	30x30 inches	30x30 inches	30x30 inches	30x30 inches
Autorouter speed comparison	1	2.2 times faster	1	5.2 times faster	5.2 times faster
Rat's nest generation	no	no	yes	yes	yes
Global nets	no	no	no	no	yes
Gerber export	no	yes	yes	yes	yes
Gerber import	no	no	no	no	yes
Price (single user ex vat)	£99	£149	£149	£199	£299

Symbol placement

Placing a symbol on the design area is pleasing. The symbol moves smoothly with the cursor and you can see exactly where it will end up. Top marks to *Quickroute* for correcting a major frustration of their earlier product. But the sting in the tail is that symbol rotation is a little cumbersome.

Symbols are automatically numbered as they are placed and a dialogue box allows you to enter a value for each device. Multi gate devices are well handled. For example calling up a *7400* will place all four gates on your design. Use the one you want and the spares are ready and waiting for future employment. The power pins are also shown to remind you to connect V_{cc} and ground.

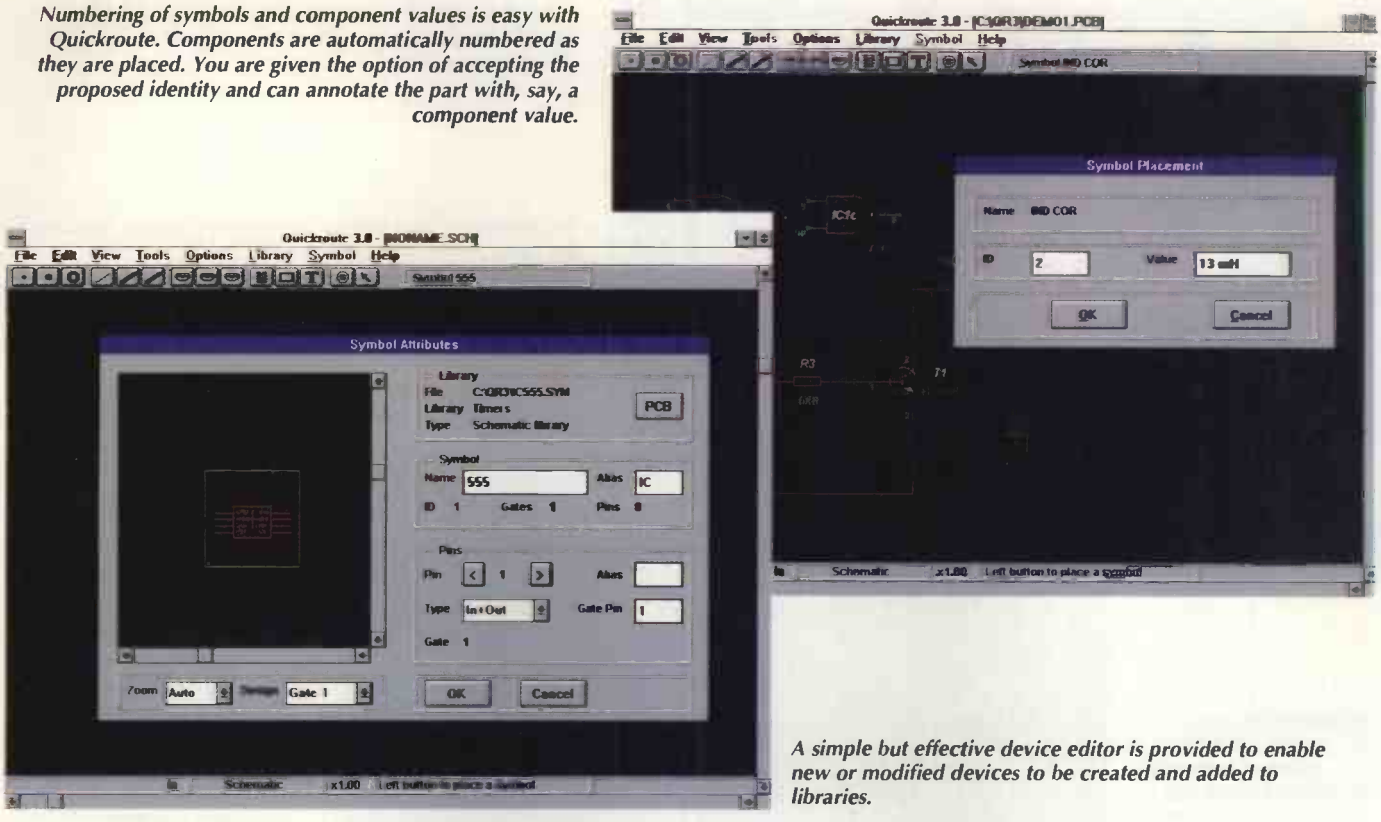
When performing editing operations, *Quickroute* does a complete redraw after each move. The speed of redraw is acceptable even on my relatively slow *386SX* machine, but if you are swiftly moving several items in quick succession there is just enough delay to become noticeable. There is a 'turbo draw' feature which omits the fill operation. This speeds the redraw by approximately 30 percent. The worst redraw I experienced was reduced from 10 seconds to 7 by the turbo setting.

Care must be taken when entering connections on a schematic. The connections must be entered on the rat's nest layer which all sounds reasonable, however it is also necessary to select the track type netlist. Without this what appear to be connections turn out to be nothing of the sort. This is one of the reasons why manual reading is essential to avoid the pitfalls.

Once the design is complete the *Pro* and *Pro+* versions of the software will scan the design and draw up a netlist ready for the autorouter. If you are working on more complex designs global nets and buses are possible although only on the *Pro+* version.

Global nets are connections that are not actually shown on the schematic but exist electrically. Two nets at either side of a schematic can be given the same name to avoid the need to draw a line all the way around. Multiple power

Numbering of symbols and component values is easy with Quickroute. Components are automatically numbered as they are placed. You are given the option of accepting the proposed identity and can annotate the part with, say, a component value.



A simple but effective device editor is provided to enable new or modified devices to be created and added to libraries.

connections also lend themselves to global nets. By this same simple technique buses can be constructed.

Design rule checking

Net lists can be exported from *Quickroute* and because they are nothing more than ascii text, can be read by any text editor. The manual has a section called 'Design Rule Checking'.

Many other programs perform design rule checking and can detect shorts, inputs left open circuit, outputs tied together and unconnected power pins. In the case of *Quickroute*, design rule checking means printing out the netlist at various stages in the design and checking the connections you want are still there.

Perhaps a little disappointing but we are told that some form of automated design rule checking is on the horizon for future versions.

PCB layout

With schematic and netlist in hand we can now begin the pcb layout. Both *Designer* and *Designer+* versions are perhaps aimed more at the layout than the schematic capture. While the drawing of schematics is possible you cannot transfer the design directly into the layout part of the software and nets must be entered manually.

On the higher specification versions, moving from schematic to layout could not be easier. Select the menu item called 'generate rat's nest' and the components appear with the point to point rat's nest connecting the relevant pins. Components can then be placed in a more suitable arrangement prior to routing.

When moving a component a ghost of the component follows the cursor as you decide where to put it and once the decision is taken the new rat's nest connections are drawn in. When you click to say I want it here, a mini menu pops up with the selections move, copy or multi-copy. This is another example of something that sounds trivial, but can become irritating when a lot of moving around has to be done. I would prefer to enter move mode or copy mode so the decision is taken at the start of the exercise and only once.

If components are being placed manually it is a similar exercise to schematic drawing. Select a suitable library, and the component needed then place it on the layout. A vast array of pad types is provided and you can define your own if you want. Track widths from 0.01in up to 0.6in are provided and you can define your own track down to 0.001in if needed. They can be placed on any of the 10 layers which are clearly identified by colour and the status bar at the bottom of the screen. Track drawing is assisted by turning on or off polylines and 45° options so the whole exercise is as painless as it can be.

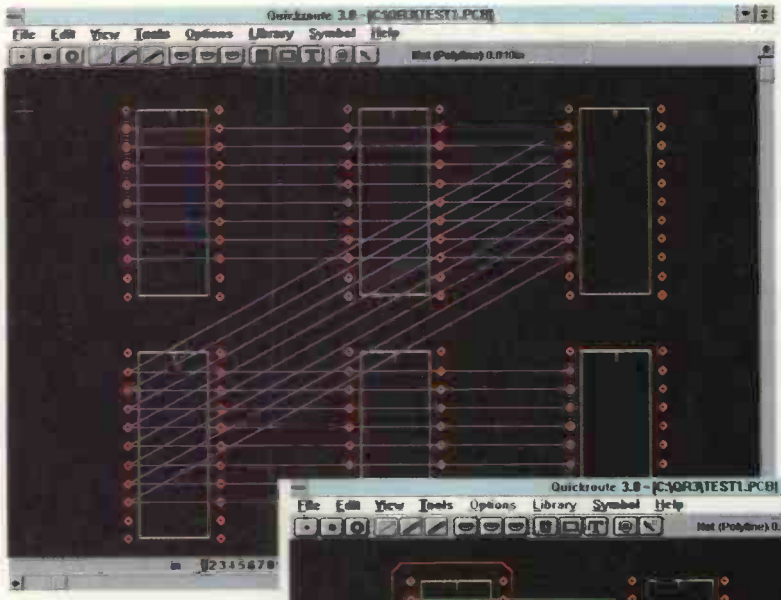
The layout can be annotated with text. Vector text is constructed from lines and can be exploded into its individual tracks although I have yet to think of a reason why I might do that. Alternatively, any font available through windows can be used given the output device can print it. Even wingdings could be used to brighten up a dreary pcb.

Autorouteing

All versions of the software include an autorouter of some form. The autorouter needs a netlist to tell it what to route, and on the up market versions of the software this is generated from the schematic.

In the case of the *Designer* and *Designer+* versions, the netlist has to be entered manually by drawing the rat's nest. I entered some connections so that I could exercise the autorouter and the procedure is reasonably easy but it does beg the question: would it have been just as easy to lay the tracks?

The autorouter, while not really configurable, does have one or two adjustments you can make. There are two grid settings for the routing algorithm, normal at 0.05in and fine at 0.025in. The clearance to be left around pads can also be specified to allow for manufacturing tolerances. The autorouter will route on any or all of the eight layers, you specify which ones by defining them as visible or not.



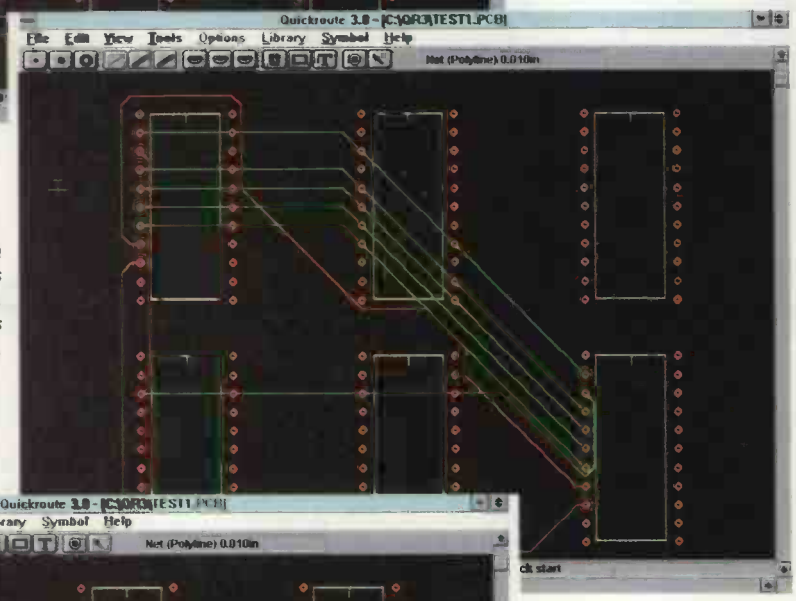
A bus structure on a small memory array was used to exercise the autorouter. This is the rat's nest showing the desired connectivity.

The track size used will be the current selection made on the menu.

Once unleashed, the autorouter carries on to the bitter end which may, or may not result in all tracks being routed. If the constraints are such that not all can be completed the autorouter reports its success level and leaves rat's nest lines to indicate the outstanding connections to be made.

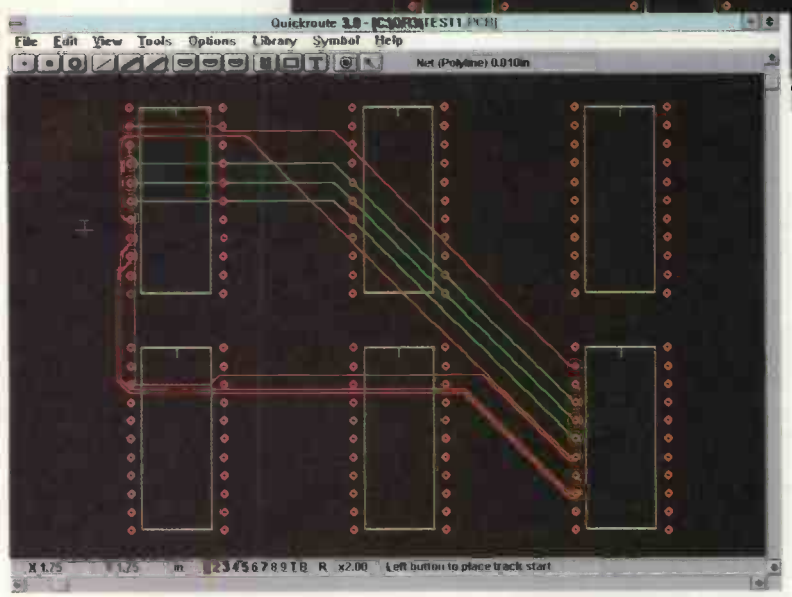
There is no provision to perform a partial route, which would allow for example power tracks to be routed at 0.1in width, then 0.015in to be used for the signal tracks.

For reasons not fully explained, the algorithm avoids routing through component identities on the silk screen layer. These can be turned off to avoid this effect, and it can even be used to advantage because sometimes it is necessary to avoid routing in certain areas and rectangles, circles and other shapes can be placed on the design to achieve this effect.



The autorouter was set to a pad clearance of 1.0 and its normal grid setting, whereupon it produced this result.

Performance of the autorouter was difficult to judge. So many factors can influence the perceived speed of operation, and according to the literature the speed varies depending upon which version of *Quickroute* you buy. It gave some curious results on occasions, this may have been due to misuse or incorrect setup but it did not leave me feeling I had discovered an invaluable tool. My conclusion was that perhaps with effort and patience I would eventually get it to work.



Curious results were encountered from the autorouter, in this case switched to its fine grid setting and the minimum permitted pad clearance. Be prepared to plough through the manual and experiment a little in order to achieve useful results from the autorouter.

Multicoloured prints

Printing or plotting the design is remarkably simple. The output is produced at whatever zoom level is currently selected for the screen. Bearing in mind that you can specify any multiplier as a zoom factor there is plenty of flexibility to fine tune the size of prints.

Any combination of layers can be output using the same idea, just make sure they are selected to show on the screen. Normally all items are printed in solid black to ensure best quality for the final design, however the software will support colour printers, if for example you wanted a multicolour print for checking.

Package options

With the exception of the *Designer* version, *Quickroute* will output Gerber and NC-drill files for automated manufacturing, and the *Pro+* version has the ability to import Gerber designs, perhaps for checking or to integrate parts of a design produced elsewhere.

There is a special version of the program aimed at the educational market known as *Curriculum*. This includes a 'teacher page' which provides password protection for most of the configurable features of the package. It evens

Software source

All versions except curriculum: Powerware,14 Ley Lane, Marple Bridge, Stockport SK6 5DD. Tel./fax: 0161 449 7101. The curriculum version is only available to educational establishments and is supplied by: Economatics Education Ltd, Tel. 01742 561122.

denies access to the netlisting and autorouteing tools which seems a little draconian, but the general principle is a good idea.

Maximum design area is a more than adequate 30 inches square. Compared to most windows software, *Quickroute* is quite modest in its requirements for memory. However if you plan to design large circuits and particularly if you are using the autorouter be prepared to plug in more ram.

Quickroute 3 bears a lot of similarities to its earlier generations. The windows interface has been tidied up a little and standardised. Powerware has deliberately moved up market with this product. It costs more than previous versions but is a more competent package. The competence has been extended not so much in additional features but by giving the capacity to deal with larger designs. The top of the range autorouter is also a much more competent algorithm than on previous releases.

There has been a clear effort to address the educational market and this is probably a sensible move. *Quickroute* shows evidence of evolving from that area rather than the industrial sector, and there is no doubt that the software provides an attractive, easy to use and well structure teaching aid for schools and colleges.

In the mainstream of cad software it will face some stiff competition. But the Powerware organisation is small, responsive and provides good technical support – factors that should be considered when buying software. ■

SPECIAL OFFER

40% discount – Quickroute 3 Designer
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As an introductory offer – exclusive to *EW+WW* readers – Powerware is offering the *Designer* version of its latest *Quickroute 3.0* package for £59.40. Apart from having its manual on disk rather than as a hard copy, this package is identical to the full-priced software, which normally retails at £99 – excluding vat, postage and packing.

In addition, Powerware is also offering 25% discount on the *Pro+* version of *Quickroute 3.0,* bringing the price to £224 from the usual price of £299. See reader reply card between pages 56 and 57.

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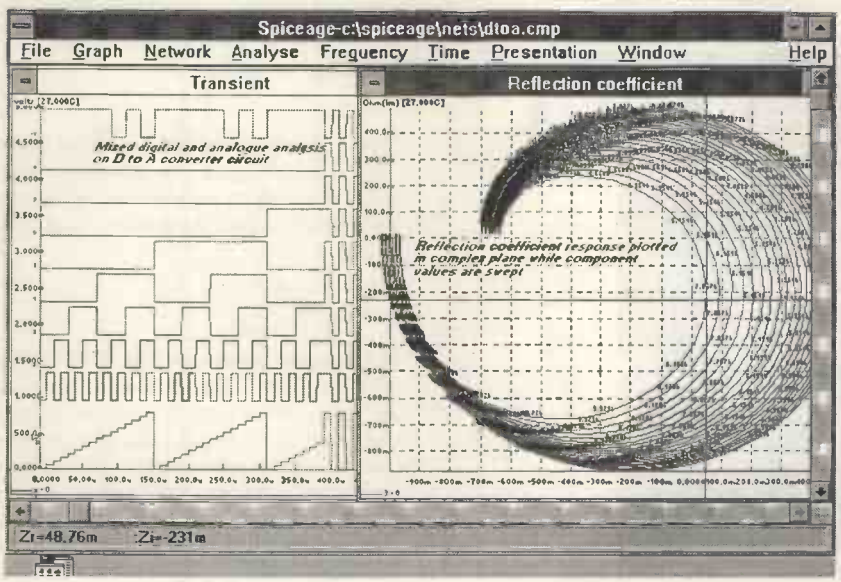

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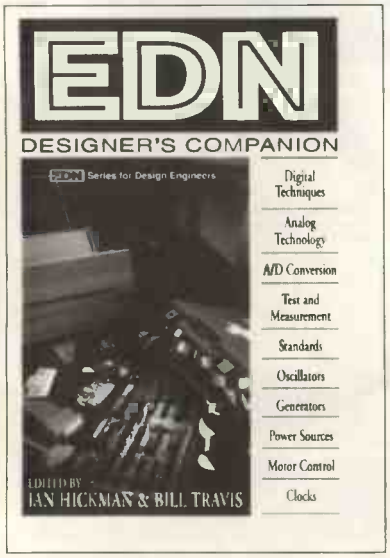
LOWER THD from vintage designs

***Frequency-dependent negative resistors produce anti-aliasing and anti-imaging filters with low sensitivity to component values. Rick Downs shows how fdnr filters can replace inductors in this extract from the book* EDN Designer's Companion.**



EDN Designer's companion is available by postal application to room *L333 EW+WW, Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS.*
Cheques payable to Reed Book Services please. Credit card orders accepted by 'phone (081 652 3614).
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Price £25.00+£2.50 postage.

Digital audio designers are always concerned about noise, total harmonic distortion, and phase linearity. To prevent aliased noise and distortion, using an anti-aliasing filter, designers have to limit the bandwidth of the audio signal before the signal reaches the analogue-to-digital converter.

Using an anti-imaging filter, designers must also limit the bandwidth of the audio signal that comes from the digital-to-analogue converter. However, maintaining an acceptable phase linearity for a filter that matches the dynamic range of converters with resolution of 16 bits or more can be a challenging task. More and more digital audio designers now use early filter architecture, the Generalised Immitance Converter (gic), to meet this challenge. Immitance is a combined term that covers impedance and admittance.

The immitance converter in **Fig. 1** is a two port network whose input impedance is:

$$Z_{in(s)} = \frac{Z_{1(s)}Z_{3(s)}Z_{5(s)}}{Z_{2(s)}Z_{4(s)}}$$

where s is the standard Laplacian operator. When impedance Z_{s2} is capacitive and the other impedances are resistive, the immitance converter's input impedance simulates an inductor.

$$Z_{in(s)} = \frac{sR_1C_2R_3R_5}{R_4}$$

An inductor can also be simulated by making Z_4 capacitive and the remaining impedances resistive. When any two of the numerator impedances are made capacitive, such as Z_1 and Z_5, and the other impedances resistive, the converter simulates a frequency dependent negative resistor.

$$Z_{in(s)} = \frac{R_3}{s^2C_1R_2R_4C_5}$$

The impedances that are made capacitive have different circuit implications. If Z_1 is made a capacitor, and the filter is ac coupled, the operational amplifier, IC_1, will not have a bias-current return path, which could affect the filter's operation. When Z_3 and Z_5 are made capacitive, the op-amps in the immitance converter all have a bias-current return path.

Audio designers generally make Z_1 and Z_5 capacitive, letting R_2=R_3 (Z_2=Z_3) to be set to minimise the effect of op amp gain bandwidth mismatch. If you configure the converter with capacitors for Z_3 and Z_5 you should employ

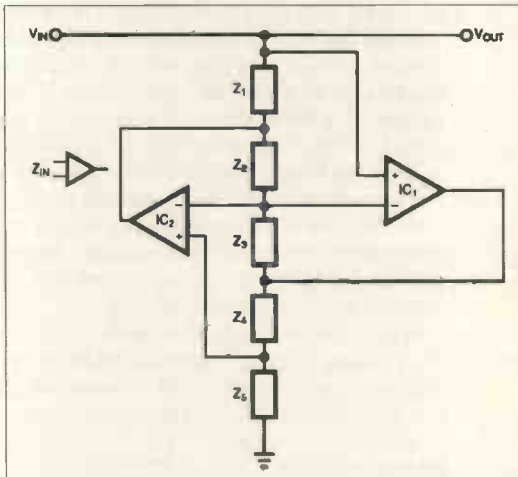

Fig. 1. A Generalised Immitance Converter can simulate the immitance of any passive component, including a frequency dependent negative resistor. Designs incorporating these avoid the need for inductors.

op-amps with well-matched gain bandwidths.

An active gic filter can simulate the transfer characteristic of a passive *LC* ladder network (**Fig. 2a**). An *LC* ladder network is transformed into an active gic filter by multiplying each ladder network element by 1/s. The transformation changes all the inductors to resistors, the capacitors to fdnrs, and the resistors to capacitors (**Fig. 2b**). R_1=R_2 and C_3=C_5 can be set, allowing the filter to be trimmed using only one component, R_4.

Because the gic filter simulates an *LC* filter, it has a lower sensitivity to component value variations than other *RC* active filters, such as the familiar Sallen and Key topology[1]. In addition, the gic filter topology lets you design high-order filters having unity gain, whereas the Sallen and Key filter topology (**Fig. 3**) often requires gain greater than unity to derive real resistor values.

To illustrate an active gic filter design, consider the filter requirements in a practical digital audio record and playback channel. Modern digital-audio channels oversample the recorded bandwidth using a sampling rate that is more than four times the Nyquist rate. Multiplying the recommended standard 48kHz audio sampling rate by a factor of four produces the standard 192kHz four times the oversampling rate.

Such a high oversampling rate eliminates the need for 'brick-wall' attenuation slopes

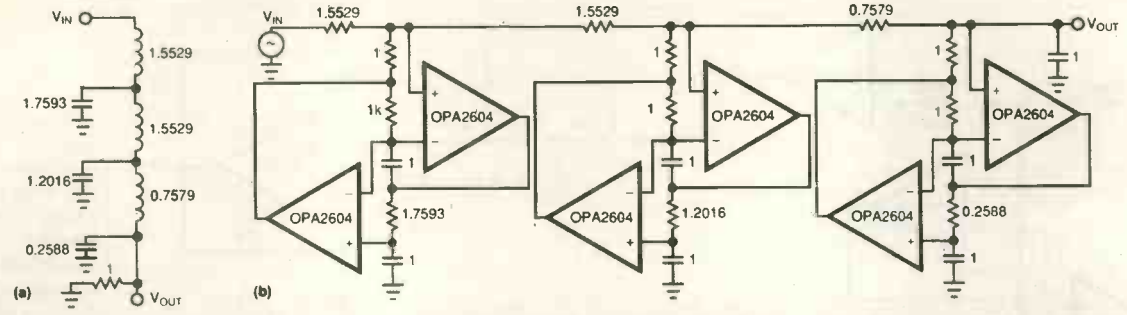

Fig. 2. An LC ladder network generates any low pass filter polynomial (a). The filter's transfer function can be simulated by multiplying each element by 1/s and the filter can be realised using GICs (b).

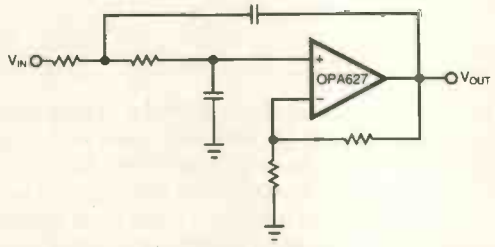

Fig. 3. Active-filter handbooks contain many 'cookbook' designs for the familiar Sallen and Key lowpass architecture. The topology often requires greater then unity gain to realise the filter.

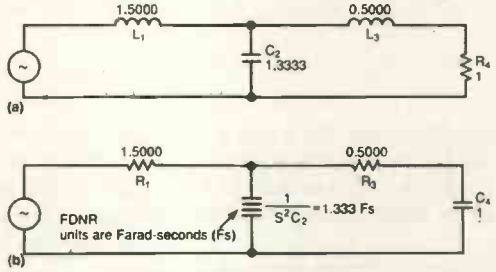

Fig. 4. A normalised third-order Butterworth filter serves as a building block for the anti-imaging filter (a). The normalised GIC filter is obtained by multiplying each normalised element value by 1/s (b).

and eases limitations on allowable phase distortion, hence allowing you to use lower-order – and lower-cost – anti-aliasing and anti-imaging filters. It also avoids the manufacturing difficulties and non-linear group delay associated with brick-wall filters.

Playback devices aid anti-imaging

The anti-imaging filter's attenuation requirements benefit from the attenuation provided by three factors in the playback channel. The first factor is the attenuation characteristics of the digital interpolation filter that precedes the digital-to analogue converter. The filter interpolates the data between samples and removes most of the signal energy above 20kHz to prevent aliasing by the dac.

The second factor is the frequency response and linearity of the output power amplifier. The power amplifier bandwidth is generally restricted to 20kHz, which attenuates higher-frequency energy. A note of caution, however: noise shaping dacs have significant out-of-band energy, which, if the amplifier becomes non-linear, can create intermodulation products.

The third factor is the frequency response of the dac. Because a dac maintains the analogue value of each digital sample, the converter exhibits the frequency response of a zero-order-hold filter given by

$$H_{(f)} = \frac{\mathrm{Sin}\left(\pi \frac{f}{f_s}\right)}{\left(\pi \frac{f}{f_s}\right)}$$

where f is the signal frequency and f_s is the sampling frequency. The converter's $\sin(x)/x$ frequency response can provide 20dB or more of attenuation in the imaging frequency range, which is 20kHz removed from the sampling rate.

The anti-aliasing filter's attenuation requirements are more stringent. Because there is no digital filter or power amplifier preceding the analogue to digital converter, the filter cannot benefit from the additional attenuation provided by these devices. In addition, because the digital samples from the adc are represented mathematically as impulses in discrete-time, an adc does not exhibit the zero-order-hold response of a dac.

To match the dynamic range of a 16 bit analogue to digital converter, the anti-aliasing filter should theoretically provide 96dB of attenuation to keep alias responses below the quantisation noise level. In practice, however, the amplitude of audio signals in the 10 to 20kHz range is significantly less than the adc's dynamic range. Therefore, designers often use 65dB as an adequate rule of thumb for the anti-aliasing filter's attenuation. Because of the help it gets in the playback channel, the anti-imaging filter can be of a lower order than the anti-aliasing filter.

Although there is a wide choice of low pass filter polynomials, many of these polynomials are not suitable for audio applications. Chebyshev and elliptic filters have steep cutoff characteristics, but their large pass-band ripple can be troublesome in some audio applications.

The Butterworth filter has a maximally flat pass-band response, but its cutoff characteristics are less steep. Because these three types of filters do not exhibit constant group delay, pass-band frequencies experience unequal time delays that can cause excessive overshoot, and ringing in the transient response.

The Thompson filter, also known as the Bessel filter, has constant group delay, which provides excellent transient response in audio and dsp applications. However, the Thompson filter's cutoff characteristics are even less steep than the Butterworth filter's, so a high-order Thompson filter would be needed to achieve the same stopband attenuation as a lower-order Butterworth filter. Because oversampling relaxes the attenuation requirements, a 40kHz Butterworth filter can often be used. It has group delay from 20Hz to 20kHz, tolerable in most four times over-sampled digital audio systems.

Using a 192kHz four times oversampling rate, a lkHz Butterworth anti-aliasing filter must attenuate the aliasing components in the 160 to 170kHz frequency range by the rule of thumb 65dB. The order of a Butterworth filter is determined using the following equation:

$$10^{\frac{K_s}{20}} = \sqrt{1+\left(\frac{\omega_s}{\omega_c}\right)^{2n}}$$

where K_s, is the stopband attenuation in decibels, ω_s is the minimum stopband frequency, ω_c is the 3dB cutoff frequency, and n is the filter order. Solving for n yields

$$n = \frac{\log\left[\sqrt{10^{\frac{K_s}{20}}-1}\right]}{\log\left[\frac{\omega_s}{\omega_c}\right]}$$

Substituting $K_s = 65$, $\omega_s = 2\pi \times 160$kHz, and $\omega_c = 2\pi$ x 40kHz yields $n = 5.4$. This means a sixth-order anti-aliasing Butterworth filter would be needed. Because the anti-imaging filter benefits from attenuation, due to the dig-

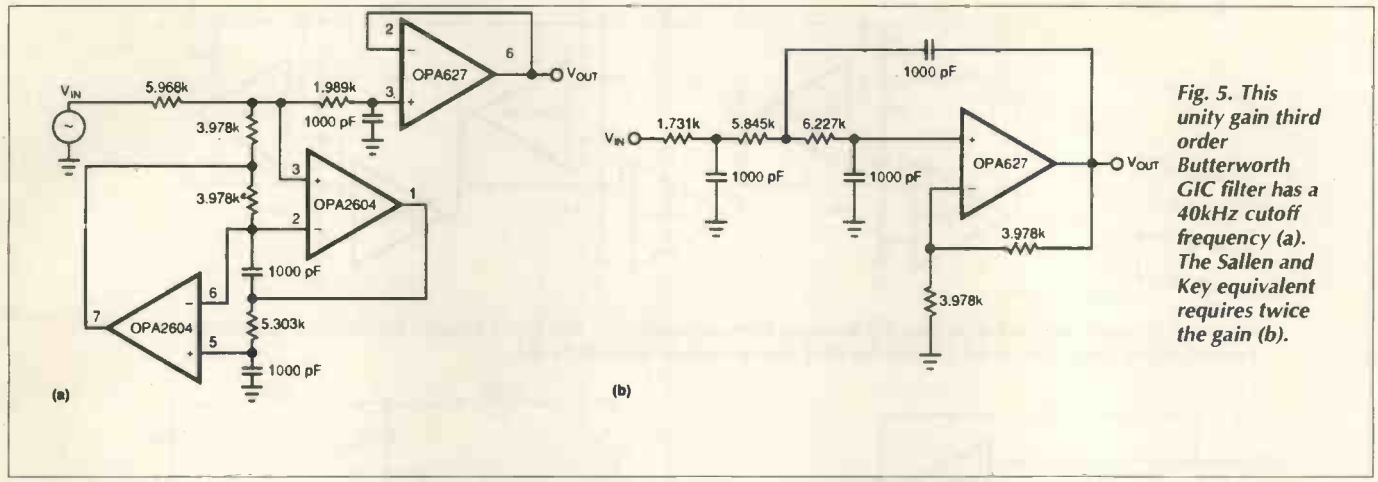

Fig. 5. This unity gain third order Butterworth GIC filter has a 40kHz cutoff frequency (a). The Sallen and Key equivalent requires twice the gain (b).

ital interpolation filter, the dac's $\sin(x)/x$ response, and the restricted power amplifier bandwidth, a third-order Butterworth filter, producing 36dB of attenuation at 160kHz, should suffice.

Figure 4a shows the component values for a third-order Butterworth *LC* filter with a normalised cutoff frequency of 1 rad/s. The component values from standard filter tables such as those found in Reference 1 can be extracted. To realise the filter using a generalised immitance converter, each component is first transformed by multiplying each value by $1/s$. Therefore, L_1 becomes R_1, C_2 becomes a frequency dependent negative resistor having the value $1/(s^2C_2)$, L_3 becomes R_3, and the terminating resistor, R_4, becomes C_4, Fig. **4b**.

By setting the fdnr values, referring to Fig. 1: $Z_1=R_1$, $Z_2=R_2$, $R_1=R_2=1$; $Z_3=C_3$, $Z_5=C_5$, $C_3=C_5=1/s$; and $Z_4=R_4=1.333\Omega$. As with the circuit in Fig. 2b, a single resistor R_4 determines the value of the frequency dependent resistor.

Next the normalised cutoff frequency must be scaled to the desired cutoff frequency of 40kHz. Frequency scaling can be achieved simply by dividing all capacitor values by $\Omega_n = 2\pi\times40\text{kHz}$. Finally, large capacitor values must be scaled by an impedance scale factor to realise practical circuit elements.

The impedance scale factor is

$$Z_n = \frac{\text{normalised C value}}{\text{normalised C value}}$$

Choosing the desired C value to be 1000pF yields an impedance scale factor of

$$Z_n = 7.23\times10^3$$

Multiplying all resistor values and dividing all frequency-scaled capacitor values by 7.23×10^3 produces the final filter shown in **Fig. 5a**. Because the output impedance of the filter is high, the output should be buffered via an op-amp voltage follower. **Figure 6** shows amplitude and phase response of the final filter.

The measured noise-and-distortion and noise levels show a contrast between a third order Butterworth filter design based on the familiar Sallen and Key architecture, Fig. **5b**, and this gic realisation. Noise-and-distortion measurements were performed using a 1kHz test signal and a 22Hz to 80kHz measurement bandwidth. The noise-only levels were measured in the same bandwidth when the circuit inputs were grounded. The immitance converter realisation achieves a 96dB noise and distortion level that matches the dynamic range of a 16bit digital audio system.

The Sallen and Key realisation achieves a –93dB noise and distortion level that is adequate for most consumer audio systems, but is 3dB higher than the gic realisation. Simple output noise level was measured in dBu, referenced to 0dBu, which equals 0.775 V (rms). The immitance converter filter, at –104dBu, has 7dB less noise than the Sallen and Key filter. Because the Sallen and Key filter must have greater-than-unity gain to realise the component values, its noise gain is higher than that of the unity gain gic filter.

A sixth order gic filter can be designed in one of two ways to achieve the 65dB anti-aliasing filter requirements. The previous design procedure can be extended using standard element values for a normalised sixth order Butterworth filter, or two of the previously designed third order gic filters can be cascaded to achieve the sixth order polynomial.

A design based on a normalised sixth order Butterworth filter is sensitive to gain bandwidth mismatches between all the op amps in the circuit, however. If this approach is used, op amps with high gain bandwidth products should be employed. Cascading two third order gic filters achieves acceptable results and has fewer op amp matching difficulties.■

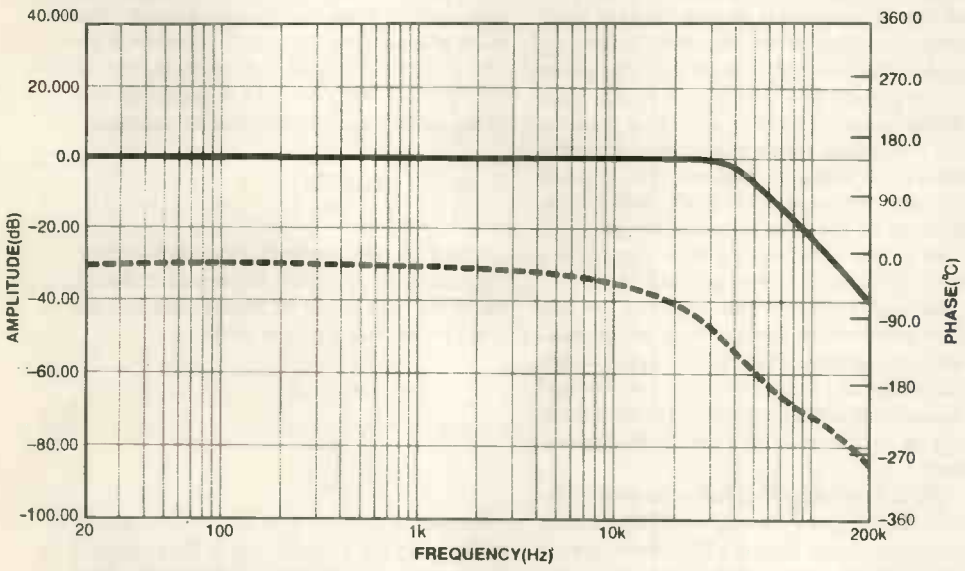

Fig. 6. The amplitude (solid line) and phase response (dotted line) of the third order Butterworth GIC filter are nearly constant throughout the audio frequency band.

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3. R Downs, *'High-Speed A/D Converter Lets Users Reap Benefits of Oversampling'*, Burr-Brown Update, Vol XIV, No. 2, May 1988, p. 3.
4. R Downs, *'A Low-Noise, Low-Distortion Design for Anti-Aliasing and Anti-Imaging Filters'*, Burr-Brown Application Bulletin AB26, February 1991.

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Best rf article '95

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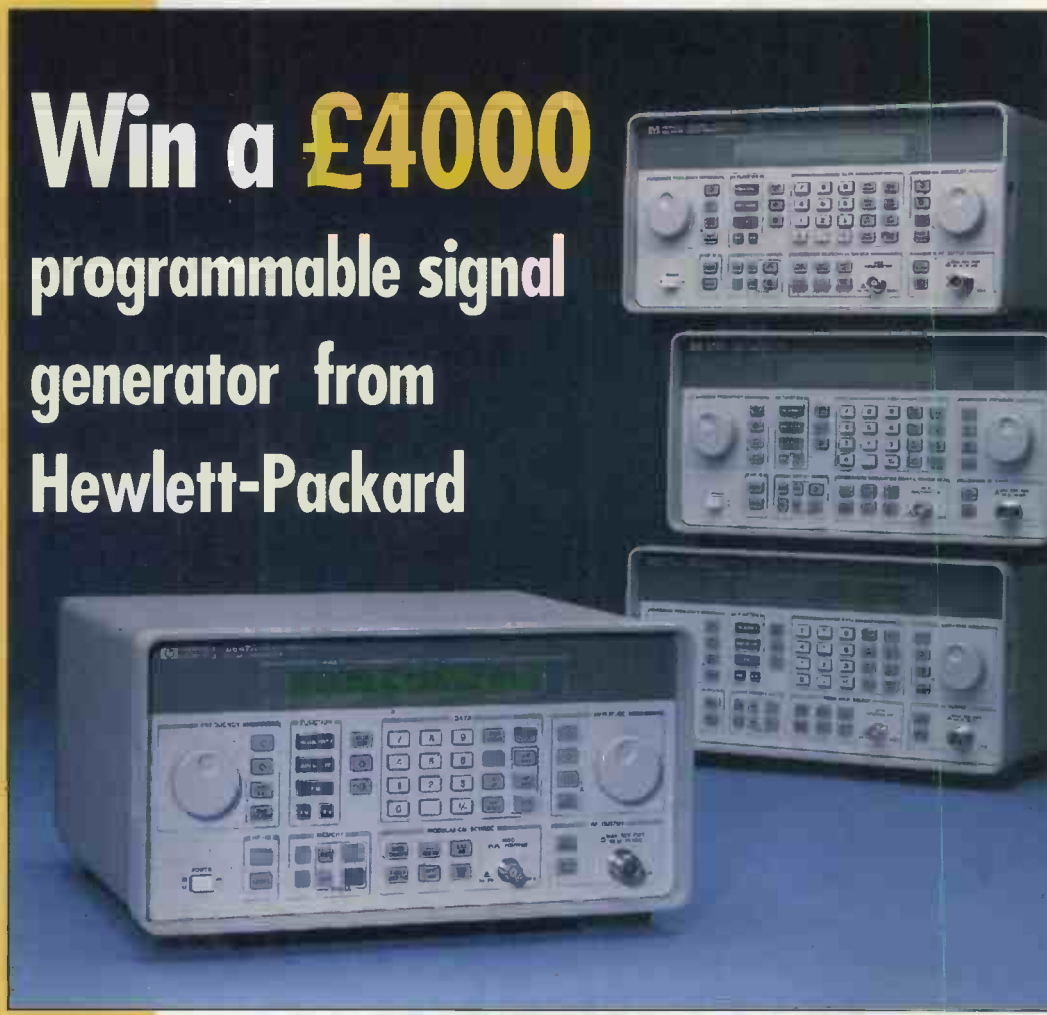

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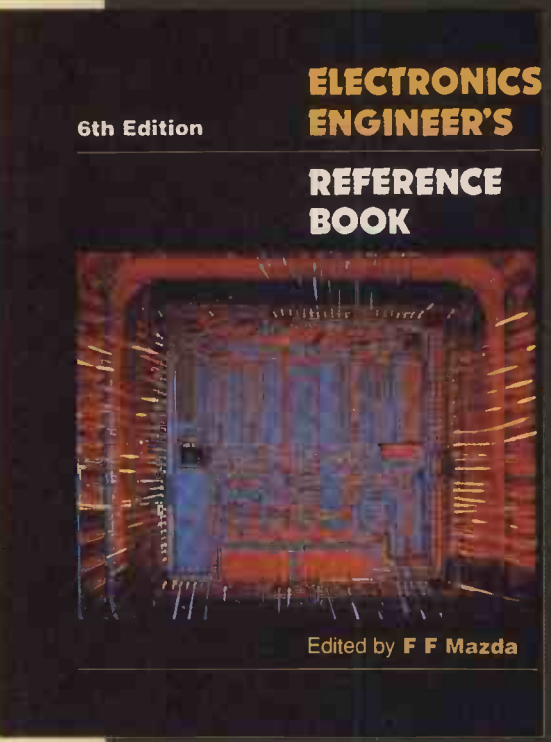

Fraidoon Mazda has worked in the electronics and telecommunications industry for over twenty years, and is currently Product and Operations Manager, Generic Network Management, with Northern Telecom. He is the author of six technical books (translated into four languages) and the editor of the Communications Engineers Reference Book published by Butterworth-Heinemann.

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TRANSMITTING AM

With today's high-speed solid-state switches, it is possible to design a scalable 'outphasing' transmitter for am broadcast that combines many desirable performance and manufacturability features. John Burnill explains.

This article owes its origin to a conversation I had with another broadcast engineer about a decade ago. We were debating whether an efficient, scalable and easily reproducible solid-state am broadcast transmitter could be designed and manufactured.

At that time, the technology was very much valve oriented, with its attendant disadvantages – in particular components such as high voltage high power coils. Time has passed and major manufacturers have adopted various solid-state designs which largely meet these desirable goals. But the design I arrived at is different way of attaining these goals. More importantly, it is possibly a better way.

AM transmitters – past and present

Descriptions of past and current transmitter designs that follow are only outlines. Individual implementations differ quite widely. A manufacturer may favour triodes and another tetrodes for example. With most transmitters, the most complex elements are usually the control circuits, and not the audio and rf stages.

Of the various modulation methods, high level modulation is the most widely known method, **Fig. 1**. It relies upon the fact that the output of a class C – or other high efficiency radio-frequency amplifier – is linear with supply voltage. The modulator modulates the power supply.

One half the dc power input to the rf amplifier must be provided by the modulation amplifier. Conventional transmitters achieve 50 to 60% overall efficiency at 100% modulation. These efficiency levels can be improved upon by using higher efficiency rf amplifiers – harmonic peaking, class D, etc. – and pwm audio stages.

Advantages of high level modulation are that it is widely understood and easy to set up – a good selling point. Its disadvantage is the high-power audio modulator, involving a transformer, choke and extra high power valves.

Low level modulation with a linear amplifier is very inefficient, 30% being typical. There are numerous modulation schemes which fall into this low efficiency category. Screen grid modulation is one of these. The principle advantage is simplicity, but the low efficiency rules out high-power use of low level modulation.

Doherty modulation involves a form of high efficiency linear amplifier, **Fig. 2**. Two output stages are linearly summed via a quarter-wave impedance-inverting section . The valve carrier amplifier provides the carrier and negative modulation while the peak amplifier provides

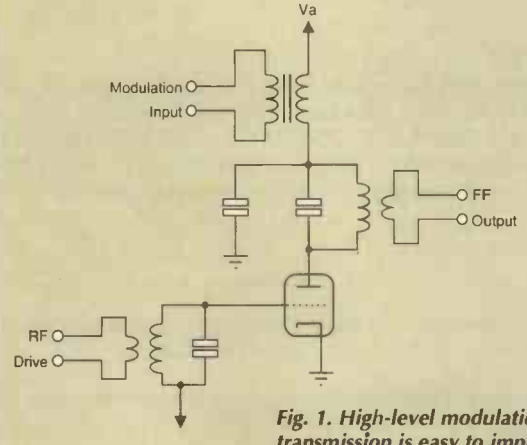

Fig. 1. High-level modulation for am transmission is easy to implement but needs a transformer, choke and extra high-power valves.

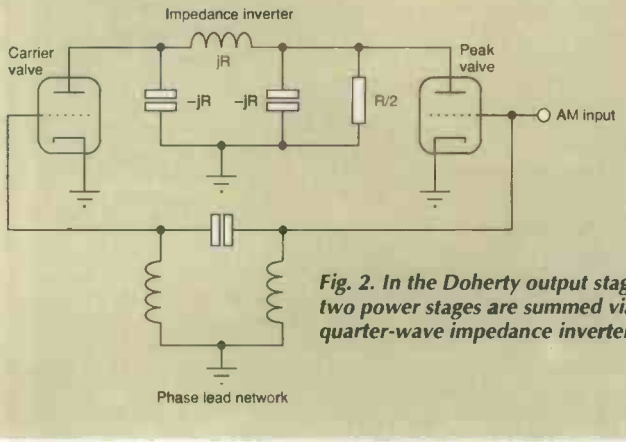

Fig. 2. In the Doherty output stage, two power stages are summed via a quarter-wave impedance inverter.

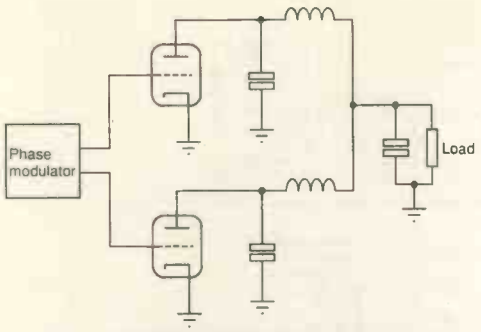

Fig. 3. In the Ampliphase am transmitter, two phase-modulated carriers are summed together at the output. Modulation involves varying the phase between the carriers.

positive modulation by impedance modulating the carrier amplifier.

The Doherty system is efficient – 60% being typical – and is not too complex. Its main advantages are respectable efficiency and relative simplicity. There are few disadvantages, which is probably explains its popularity.

Outphasing, also known as Ampliphase and Chereix, is shown in **Fig. 3**. Two phase-modulated carriers are summed at the output. By varying the phase between the carriers, amplitude modulation can be produced.

The idea is simple but practical implementations have needed additional complications to achieve reasonable efficiency and linearity. Ampliphase transmitters in particular needed amplitude modulation of the two carriers in the stages driving the output amplifiers. This led to a complicated and difficult to set up transmitter, which has discouraged many people from using it. Efficiency of this configuration is on a par with Doherty.

Digital modulation, **Fig. 4**, is only applicable to solid state transmitters. The digital transmitter uses many power amplifier modules in parallel – 160 for the 100kW Harris transmitter. Each module can be on or off under the control of an analogue-to-digital converter.

The output modules are in effect rf power d-to-a converters. Efficiency is very good, at more than 80%.

Outphasing

This description covers an up to date (solid state) outphasing am broadcast transmitter. I chose outphasing since it allows modern techniques to be used to produce a simple yet efficient transmitter. The design falls neatly into two blocks, namely the output stages and the phase modulating exciter.

Figure 5 shows an output stage for combining the two phase-modulated carriers. It has two features that make it suitable as a high-efficiency outphasing output stage. Firstly it uses class D voltage switching amplifiers. These operate efficiently because the inductors present a high impedance at the harmonic frequencies. Secondly it allows summing of two output stages with no interaction between the amplifiers.

The first four equations can be used to calculate the component values of the output stage for the desired power level at peak carrier when the outputs are in phase. Voltage V_p is the peak to peak output swing from the class D amplifier.

$$V_1=V_2=(\sqrt{2}V_p)/\pi$$

$$Q=R_l/(\omega L/2)$$

$$\omega=1/\sqrt{(L/2C)}$$

$$V_{out}=(V_1+V_2)R_l/(\omega L)$$

It is worth noting that in the practical output stage, one of the inductors needs to be varied to balance the power from each amplifier. Otherwise the inductors are fixed and the capacitor is varied to effect tuning of the output stage. By varying only the capacitor the bandwidth remains constant whatever frequency is used,a useful feature if negative feedback is used.

Figure 6 shows one of the class D amplifiers. It is a modified totem pole configuration chosen because it needs no output transformer. Since the transformer shown is at a low power level, it is not difficult to implement. It can be designed out but I have not shown this modification since it has not been tested.

The *BYV27* diode is a fast power type since it is part of the output bootstrap circuit which

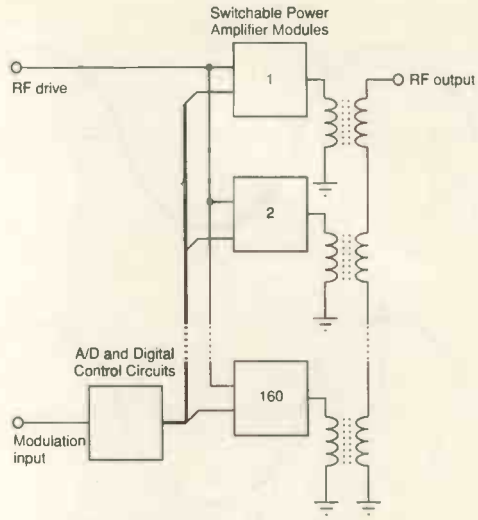

Fig. 4. Digital generation of the am carrier is only practicable using solid-state devices. It involves multiple power amplifier modules working in parallel.

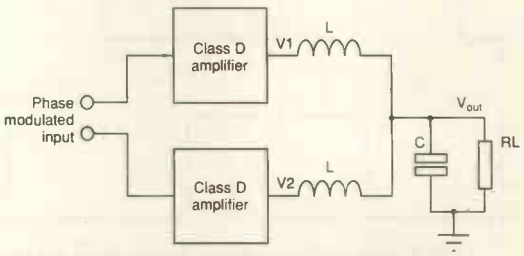

Fig. 5. Outphasing output stages produce two phase-modulated carriers that are summed together. Efficiency is high. The inductors present a high impedance to the class-D voltage switching amplifiers at harmonic frequencies.

has to switch at the output frequency. Mosfets were chosen for the output stage since they are cheap and can be made to switch very quickly (tens of nanoseconds) with simple driver stages.

The next equation below shows the maximum power output for a given device maximum V_{ds} and I_{ds} for a class-D voltage switching power amplifier.

Bear in mind that in the am transmitter, for a

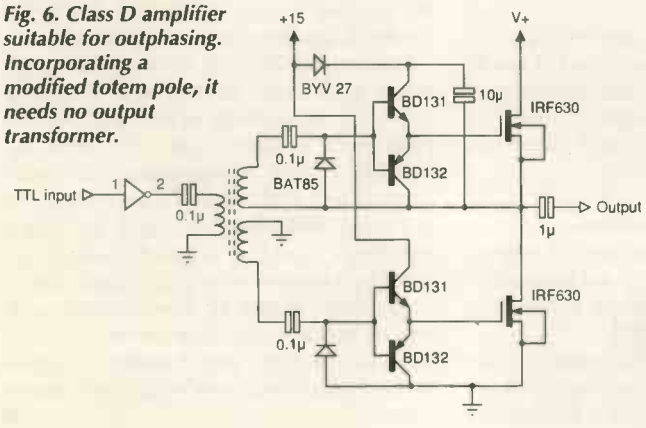

Fig. 6. Class D amplifier suitable for outphasing. Incorporating a modified totem pole, it needs no output transformer.

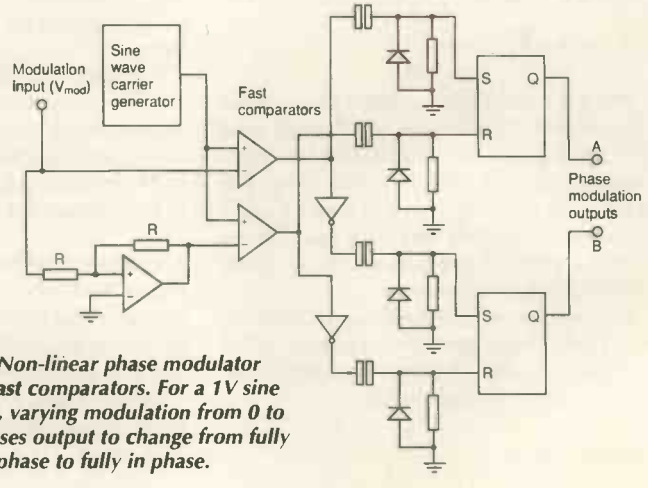

Fig. 7. Non-linear phase modulator using fast comparators. For a 1V sine carrier, varying modulation from 0 to 1V causes output to change from fully out of phase to fully in phase.

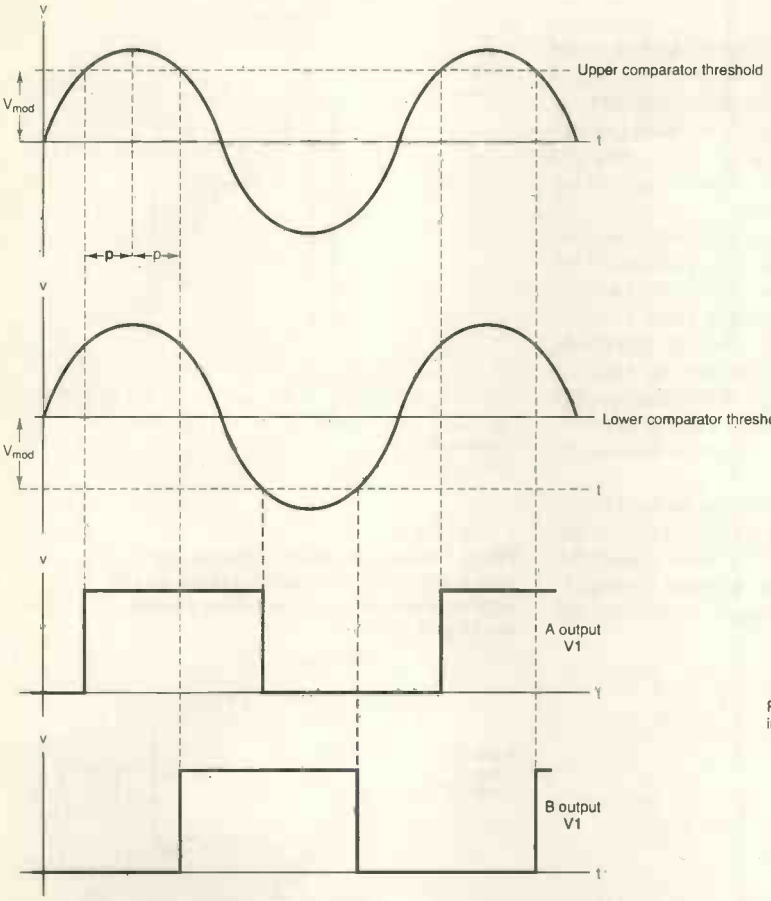

Fig. 8. Waveforms produced by the phase modulator of Fig. 7.

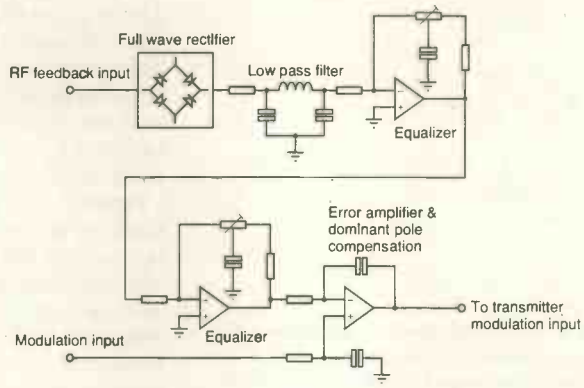

Fig. 9. This negative feedback scheme is applicable to any am transmitter provided that its bandwidth is not unduly restricted.

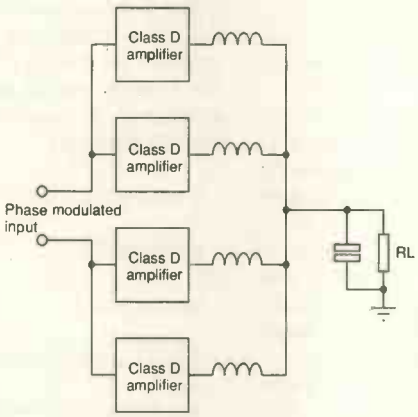

Fig. 10. Parallelling output stages of the outphasing transmitter to boost power levels. Multiple inductors are needed but each only handles a portion of the output power.

pair of output stages, each power amplifier must deliver twice the carrier power at modulation peaks. Additionally, in the real world, the maximum V_{ds} and I_{ds} must be derated somewhat to cope with potential variations in mains supply voltage and variations in loading conditions.

$$P_{out}=V_{ds}.I_{ds}/\pi$$

Outphasing the exciter

The equation below shows the output voltage as a function of V_1, V_2 when they are phase modulated by $+p$ and $-p$ degrees respectively. Linear phase modulation produces non linear amplitude modulation of V_{out}.

$$V_{out}\propto(V_1+V_2)\cos p$$

What is needed is a non linear phase modulator. **Figure 7** shows what was devised using two fast *LT1016* comparators and two edge-triggered set-reset flip flops. If the carrier sine wave is 1V peak varying V_{mod} from 0V to 1V causes the outputs to vary from out of phase (100% negative modulation) to in phase (100% positive modulation).

Figure 8 shows the waveforms generated by the phase modulator. The equation below derived from Fig. 8, shows how the carrier phase varies with modulation.

$$p=\cos^{-1}V_{mod}$$

Substituting this equation into the previous one shows that the output voltage is now linear with modulating voltage.

$$V_{out}\propto(V_1+V_2)V_{mod}$$

In the exciter, the sine-wave generator is more complicated than the phase modulator. It consists of tuned circuits for extracting a clean sine wave from the square wave generated by the phase-locked-loop IC. This IC is referenced to a crystal and has an amplitude stabilisation circuit.

In hindsight a 'digital' method of sine wave generation would have been better leading to a one adjustment frequency setting of the exciter, namely the phase-locked-loop divider ratio.

Note that this scheme is capable of producing dsb modulation with any level of carrier suppression. It also follows that the sine-wave carrier could be phase or frequency modulated in the exciter enabling all forms of modulation to be implemented at low level.

Negative feedback solution

Figure 9 outlines a negative feedback scheme. It is of course relevant to any am transmitter, providing the bandwidth of the actual transmitter is not unduly restricted.

The main item to note is the audio high-frequency boost applied in the negative feedback path (zeros). This is necessary for stability because of the poles in the forward path introduced by the transmitter. These come about because of bandwidth restrictions of the output matching network, harmonic filters and the antenna.

Dominant pole compensation was used with an f_t of 50kHz. The low pass filter must have as high a corner frequency as possible to avoid introducing unnecessarily low poles in the overall loop frequency response. I used 500kHz in the experimental exciter. The use of a full wave rectifier reduces the filtering needed.

Results

I built a 10W transmitter to test these ideas. It worked as expected, linearity being such that the negative feedback was unnecessary. The feedback was tested and proved stable.

Efficiency could not be measured at this low power level as the dissipation of the power mosfets – which were overly large for this power level – consisted only of the static dissipation of their output capacitance being charged and discharged. To get meaningful efficiency measurements means building a transmitter capable of at least kilowatt levels.

For high power levels, output stages need to be paralleled. **Figure 10** shows one way to do this. It needs multiple inductors, but each needs handle only a part of the total rf current in the output tuned circuit so can be relatively small. ■

APPLICATIONS

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Probing more deeply

An oscilloscope probe is more than just a piece of wire for applying a test signal to an oscilloscope input: it is an integral part of the measurement system. This statement is the lead in to one six articles from Fluke's 60-page A4 book, *ABCs of oscilloscopes*.

Characteristics that make different probes more suitable for specific tasks are discussed below. Active probes include active electronic components to provide amplification. Passive probes without such active components usually provide input attenuation.

Shielding. An important task of the probe is to ensure that only the wanted signal appears at the oscilloscope input. A piece of wire for example would act as an antenna producing lots of unwanted interference.

Some of this noise could even be injected back into the circuit under test. Good screening is needed, provided by screened cable earthed at the bnc connector to the scope and at the circuit under test.

Probe bandwidth. Just like oscilloscopes, probes ave a finite bandwidth which has to be allowed for. If we take a 100MHz oscilloscope and a 100MHz probe the combined response is less than 100MHz. Capacitance of the probe adds to the input capacitance of the oscilloscope, lowering the system bandwidth and the maximum displayed risetime.

Loading effect. Measured voltage is rarely exactly the same as when no connections are made. Each probe has an input impedance, consisting of resistive, capacitive and inductive components. Connecting such a probe will affect the circuit under test because of the extra load imposed by the probe. Therefore the probe characteristics must be considered when looking at the measurement results, as well as the test circuit impedances. Some probes have no series resistance. Basically, these consist of a length of cable and a test tip.

Within their operating frequency range or useful bandwidth, there is therefore no attenuation of the signal. These probes are referred to as 1:1 or ×1 probes. They have a loading effect on the circuit because they connect the oscilloscope input impedance and their own capacitance (including the cable capacitance) straight across the test point. Capacitive loading effect becomes more significant as frequency increases.

A 1:1 probe typically has an input capacitance from around 35pF to over 100pF, depending on the cable type and length and probe body construction etc. This will load the circuit under test with a low impedance. A 1:1 probe with an input capacitance of 47pF has a reactance of only 169Ω at 20MHz, making this probe unsuitable for work at these frequencies. The loading effect can be reduced by adding impedance in the probe in series with the scope input impedance. However, this means that a voltage divider is formed.

The diagram shows a simplified probe equivalent circuit. Resistors R_p and R_s form a 10:1 divider, where R_s is the 'scope input impedance. The compensation capacitor, C_{Comp} is adjusted to match the probe to the oscilloscope. This ensures that the correct frequency response curve is maintained at the probe tip. Frequency response of this type of probe is much wider than that of a 1:1 probe.

Compensating the probe. With the several adjustable capacitors and resistors, the correct response over a wide frequency range can be achieved. Most of these are set in the factory when the probe is manufactured.

There is only one trimmer capacitor for the user to adjust. This is called the low frequency compensation capacitor, and it should always be adjusted to match the oscilloscope input with which the probe is used. Adjustment is easy, using the front panel signal output of most oscilloscopes. This is marked 'Probe Adjust', 'Calibrator', 'CAL', or 'Probe Cal.', and provides a squarewave voltage output.

The effects of over and under compensation are very noticeable at higher frequencies. The displayed amplitude of the 1MHz sinewave is now very inaccurate. There can be slight differences in input capacitance between channels, so you should always compensate the probe on the channel you want to use it with.

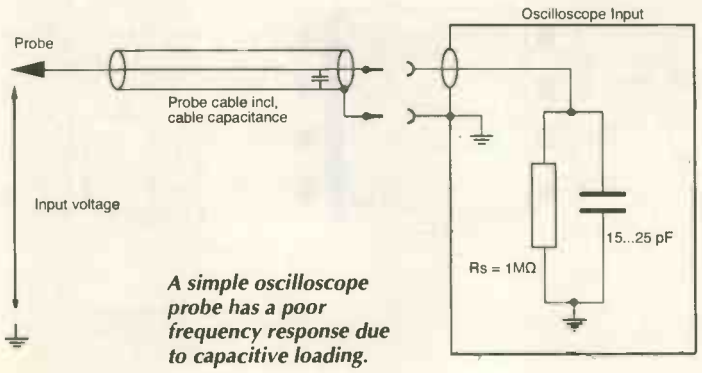
Probe read-out. Modern oscilloscope probes are equipped with a coding system that enables the oscilloscope to recognize the kind of probe is connected. Consequently, the oscilloscope can compensate the vertical deflection indication and all amplitude measurements to avoid confusion. When probes without such recognition system are used, the user has to rescale all waveforms and measurements to compensate for the probe attenuation.

Ground lead inductance. A series tuned circuit can be formed with the ground lead inductance and the combined probe and oscilloscope input capacitances. The tuned circuit is damped by the input resistance.

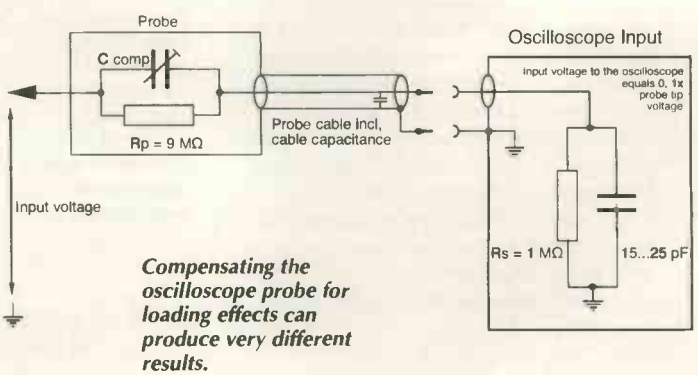
Like any other tuned circuit, this one will also ring if a step voltage is applied to it. Excessive ground lead inductance can also reduce the displayed risetime. Keep your ground lead as short as possible, especially with high frequencies and fast-rise time signals. Before connecting the probe's ground contact, make sure it is not going to short circuit any part of the system under test.

Further articles cover basics, storage instruments, analogue-versus-digital instruments and new developments. There are also measurements and exercises at the back of the book to help new and potential users understand what oscilloscopes can and cannot do.

Fluke UK, Colonial Way, Wayford, Hertfordshire WD2 4TT. Tel. 01923 240511, fax 0923 225067.



A simple oscilloscope probe has a poor frequency response due to capacitive loading.



Compensating the oscilloscope probe for loading effects can produce very different results.

Active scsi terminator aids reliable data transfer

The small-computer-systems interface standard recommends that both ends of the line are actively terminated to maximise data integrity. One chip – Unitrode's *UCC5614* – provides nine lines of active termination for an scsi parallel bus. In addition, as described in the data sheet, the device is ideal for high performance 3.3V scsi systems.

Key features contributing to such low operating voltage are the 0.2V drop out regulator and 2.7V reference. Reduced reference voltage was needed to accommodate the lower termination current dictated in the scsi-3 specification. During disconnect, supply current is typically only 0.5A, which makes the IC attractive for battery powered systems.

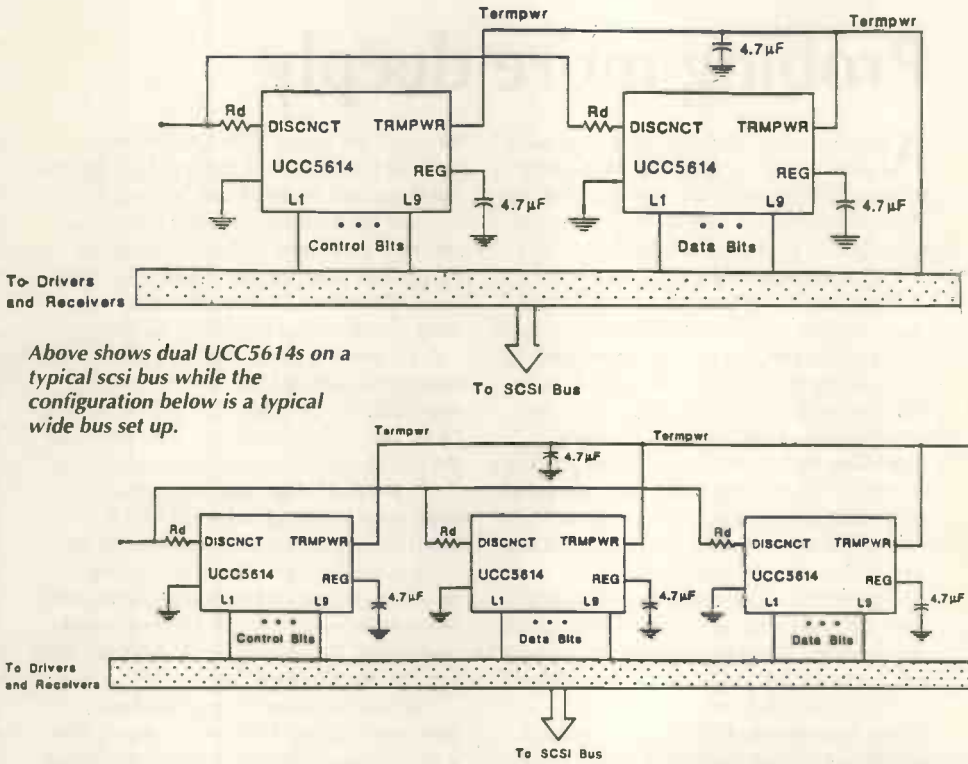
Ultra low channel capacitance of 1.8pF eliminates effects on signal integrity from disconnected terminators at interim points on the bus.

The device is programmable for either 110Ω or 2.5kΩ termination. Standard scsi bus lengths use 110Ω termination while shorter buses use 2.5kΩ termination. When driving the ttl-compatible DISCNCT pin directly, the 110Ω termination is connected when the DISCNCT pin is driven low, and disconnected when driven high. When the DISCNCT pin is driven through an impedance between 80 and 150kΩ, the 2.5kΩ termination is connected when the DISCNCT pin is driven low, and disconnected when driven high.

The power amplifier output stage allows the *UCC5614* to source full termination current and sink active negation current when all termination lines are actively negated.

Internal circuit trimming is used, first to trim the 110Ω termination impedance to a 7% tolerance, and then most importantly, to trim the output current to a 4% tolerance, as close to the max scsi-3 specification as possible. This maximises noise margin in fast SCSI operation.

Other features of the device include thermal shutdown and current limit.



Above shows dual UCC5614s on a typical scsi bus while the configuration below is a typical wide bus set up.

Unitrode UK, 6 Cresswell Park, Blackheath, London SE3 9RD. Tel. 0181 318 1431, fax 0181 318 2549.

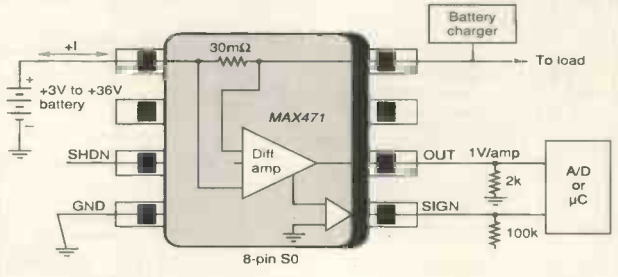
Current sensing for portable, battery powered equipment

As described in Maxim's Analog Design Guide 3/8, the *MAX471* is a dedicated single-IC, bidirectional, high-side current-sense amplifier for portable pcs, palmtops, and other battery-powered systems. It reduces board space, design time, and cost by eliminating precision resistor networks, precision amplifiers and a current-sense resistor.

Using the eight-pin device in series with the positive battery terminal and the load, all that is needed are gain-setting and pull-up resistors. The result is what is claimed to be the world's simplest circuit that measures battery charge and discharge current – including magnitude and polarity.

The device incorporates an internal 30Ω sense resistor and measures battery currents from zero to three amps. With a 2kΩ gain setting resistor, the resulting output gain is 1V/A. For applications that require increased design flexibility, the *MAX472* functions with an external sense resistor. Both devices operate from three volts to 36V, draw less than 100µA over temperature and include a five amp shutdown feature to conserve power. ■

Maxim, 20 A Horseshoe Park, Pangbourne, Berks RG8 7JW, Tel. 01734 845255, fax 01734 843863.



Consuming less than 100µA, the 8-pin MAX471 *measures battery charge and discharge rates accurately using a minimum of components.*

Putting theory to work in low power design

Now that they have given us the tools to design the circuit, Norm Dye and Helge Granberg step back and look at the matching process. In this article they show how to determine the desired source and load impedances.
***From the book* RF Transistors: principles and practical applications.**

RF Transistors: Principles and practical applications is available by postal application to room ***L333 EW+WW, Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS.***
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Basic steps in designing a low power amplifier are: select a transistor; choose a bias point; then, finally, design a circuit offering the proper impedances at both input and output to the transistor.

Each step depends to a large degree on the application. Is the circuit narrow band or broad band; must it have low noise figure? All determine how the transistor will be impedance matched.

The matching process itself is made up of two distinct parts: determining the desired source and load impedances (which we will look at in this article); and designing networks that present these impedances to the transistor (final part, next month).

Maximum gain at a single frequency is achieved by conjugate matching both input and output of the transistor, while lowest noise figure is obtained by presenting the proper source impedance to the transistor that produces minimum noise. So trade-offs between gain and noise figure are controlled by selecting source and load impedances that produce the desired gain and noise at the same time.

For suitable performance over a band of frequencies, source and load impedances that give proper overall circuit results as frequency varies must be presented to the transistor.

Whether or not S_{12} is ignored obviously depends on its magnitude and the accuracy required in a design. Wherever possible, S_{12} should be assumed to be 0 – at least in the initial design effort. Where S_{12} cannot be ignored, software such as *Touchstone* or *Mmicad* should be used, particularly if the design requires controlled noise and gain performance over a band of frequencies.

Working through specific examples is the best way to understand these procedures and here we are going to show how to determine source and load impedances for maximum gain at a single frequency – both with and without consideration of S_{12}. A third example will illustrate how to achieve a specified amount of gain, for the same conditions as in examples 1 and 2, and a fourth goes through the procedures necessary for a broadband design. Finally, we show how to design for low noise while maintaining adequate gain.

Example 1.
Narrow band – match for optimum gain, $S_{12} = 0$.

First, assume $S_{12} = 0$, that the source and load impedances are 50Ω, and that we are interested only in maximum gain at a single frequency. Our frequency of interest is 1GHz, so a suitable transistor is the *MRF571*. Bias will be 6V and 50mA because the manufacturer's data sheet shows that f_t is near its peak at 50mA and values of scattering parameters are given for this particular bias point. From the data sheet we find:

$S_{11} = 0.6$ at an angle of 156°
$S_{22} = 0.11$ at an angle of –164°
$S_{12} = 0.09$ at an angle of 70°
$S_{21} = 4.4$ at an angle of 75°

$\Gamma_s = S_{11}^* = 0.6$ at an angle of –156° and $\Gamma_L = S_{22}^* = 0.11$ at an angle of +164° (**Fig. 1**). G_s, the 'gain' contributed by the input circuit is calculated from $G_S = (1 - |\Gamma_s|^2)/|1 - S_{11}\Gamma_s|^2$ to be 1.56 or 10 $\log_{10}$1.56, 1.93dB.

Likewise, for the output circuit where the magnitude of S_{22} is 0.11, the 'gain' (G_L) contributed by the output circuit match is calculated from $G_L = (1 - |\Gamma_L|^2)/|1 - S_{22}\Gamma_L|^2$ to be 1.01 or 0.05dB. Obviously, in this example the transistor is essentially matched in the output, and little is gained from further matching.

Looking at $|S_{21}|^2$ we can determine the gain contributed by the device itself from $G_o = |S_{21}|^2$ as $G = (4.4)^2$ or $10\log(4.4)^2 = 12.9$dB. So the total gain expected from conjugate

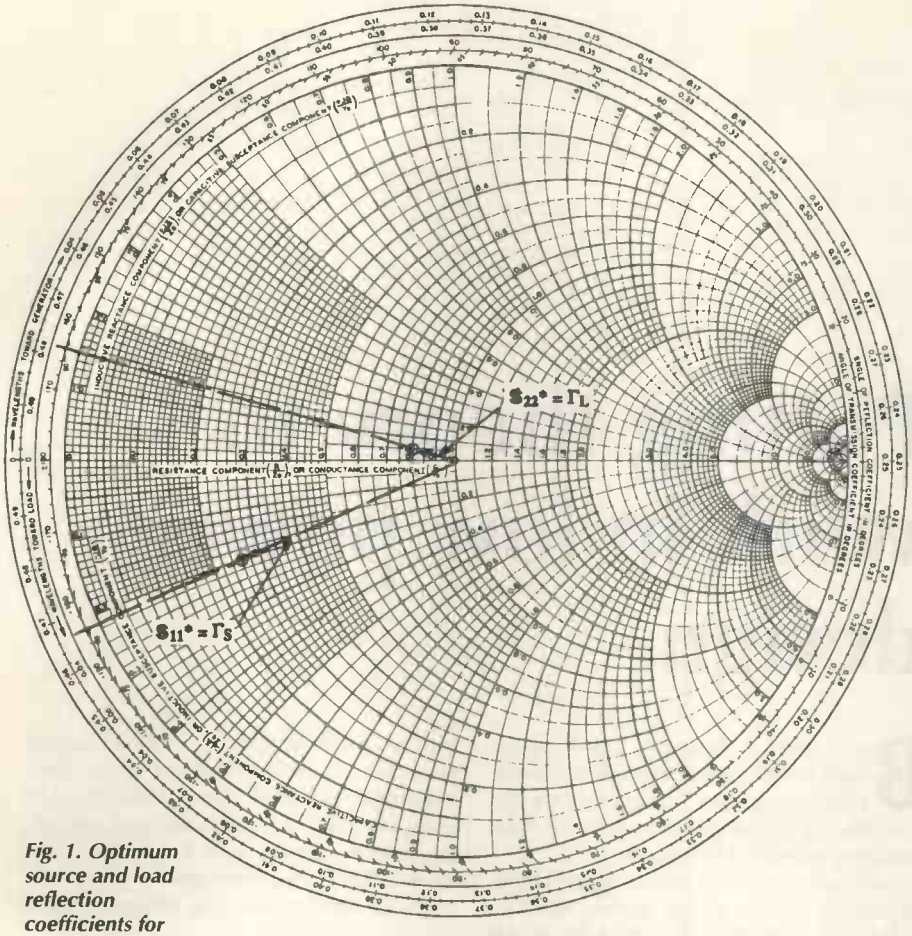

Fig. 1. Optimum source and load reflection coefficients for example 1.

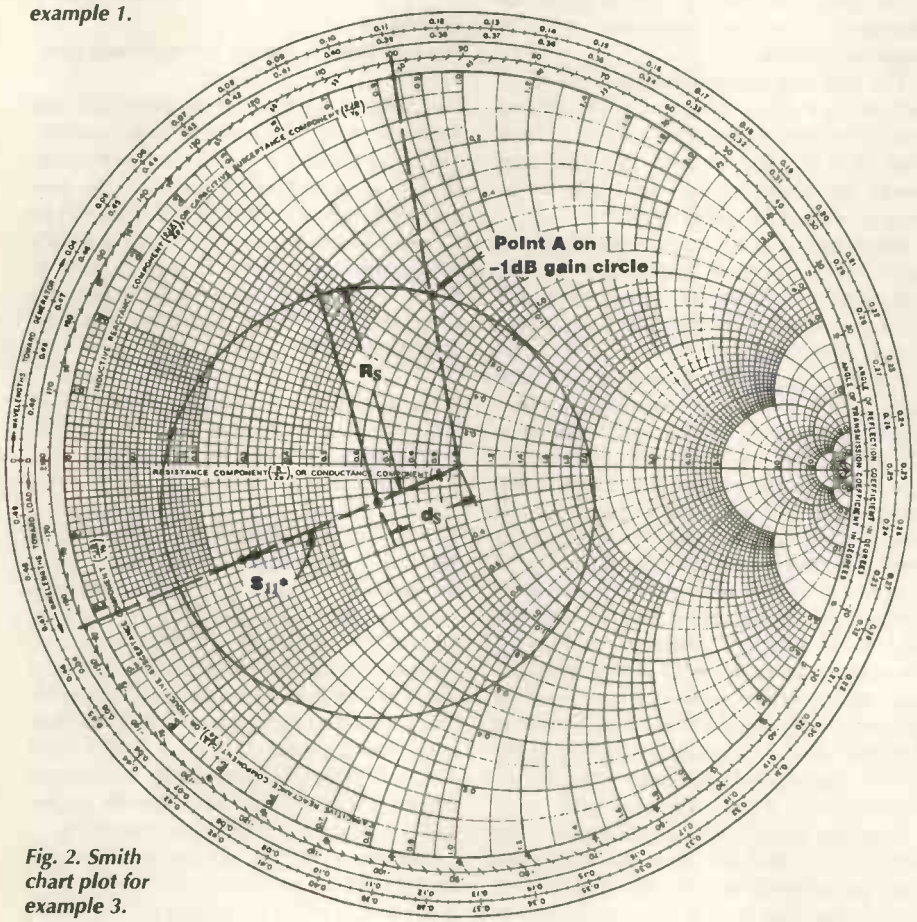

Fig. 2. Smith chart plot for example 3.

matching is $G_{TU} = G_s + G_o + G_L = 1.94 + 12.9 + 0.05 = 14.9$dB.

Example 2.
Narrow band – match for optimum gain, S12 ≠ 0.

To understand the effect of assuming that $S_{12} = 0$, calculate the gain of the same amplifier with the more precise value of $S_{12} = 0.09$ at an angle of 70°. The formulas are more complex while, in addition, optimum output impedance depends on input impedance and vice versa. The maximum available gain, $G_{p,max}$ – or sometimes mag – is given, if the device is unconditionally stable, by the formula:

$$G_{p,max} = \frac{|S_{21}|}{|S_{12}|}\left(K - \sqrt{K^2 - 1}\right)$$

where K is the Rollett stability factor given by:

$$K = \frac{1 + |D_s|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{21}||S_{12}|}$$

and $D_s = S_{11}S_{22} - S_{12}S_{21}$ as stated previously.

By using the values of the scattering parameters, the first step is to calculate D_s using the relationship :

$$\begin{aligned} D_s &= (0.6 \angle 156°).(0.11 \angle -164°) \\ &\quad - (0.09 \angle 70°).(4.4 \angle 5°) \\ &= 0.066 \angle -8° + 0.396 \angle -35° \\ &= 0.46 \angle -23°. \end{aligned}$$

Then calculate K from $K = (1+|D_s|^2 - |S_{11}|^2 - |S_{22}|^2)/2|S_{21}||S_{12}|$ to verify that the transistor is unconditionally stable (value of $K > 1$):

$$K = \frac{1 - 0.36 - 0.0121 + 0.212}{(2)(0.09)(4.4)}$$

$$= \frac{0.840}{0.792} \cong 1.07$$

In the present example, for the S parameters given and subsequent calculated value of K, $G_{p,max}$ becomes:

$$G_{p,max} = \frac{4.4}{0.09}\left(1.07 - \sqrt{0.145}\right) = 33.7$$

It is easy to see how much more complex the situation becomes when S_{12} cannot be assumed equal to zero. In the instance given, the difference in gain is approximately 0.4dB and it appears marginal to have assumed $S_{12} = 0$. We could have determined the magnitude of the error at the outset by using $1/(1+U)^2 < G_t/G_{tu} < 1/(1-U)^2$ as follows. First, calculate U from:

$$U = \frac{|S_{11}||S_{21}||S_{12}||S_{22}|}{\left(1 - |S_{11}|^2\right)\left(1 - |S_{22}|^2\right)}$$

$$= \frac{(0.6)(4.4)(0.09)(0.11)}{\left(1 - (0.6)^2\right)\left(1 - |0.11|^2\right)} = 0.041$$

Then from:

$$G_p = \frac{1 - |\Gamma_S|^2}{1 - |\Gamma_{IN}|^2} \bullet |S_{21}|^2 \bullet \frac{1 - |\Gamma_L|^2}{|(1 - S_{22}\Gamma_L)|^2}$$

we can determine the limits of possible error for assuming $S_{12} = 0$. The lower limit is $1/(1 + U)^2 = 0.919$, while the upper limit is $1/(1 - U)^2 = 1.108$. Expressed in decibels, these limits become + and –0.36dB.

We can also see the effect that S_{12} will have on the optimum source impedance by letting the load reflection coefficient remain the conjugate of S_{22}, but then calculating Γ_s from:

$$\Gamma_{IN} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L}$$

will show that Γ_s becomes 0.55 at an angle of –159°, only slightly changed from the previous value for Γ_s when $S_{12} = 0$: 0.6 at an angle of –156°.

Example 3.
Narrow band, specified gain < optimum gain, $S_{12} = 0$.

Before going to broad band circuit design, assume – for the same transistor and frequency used in examples 1 and 2 – we want a specified value of gain less than maximum, say 1dB. Example 1 has already shown that optimum match for the load results in only 0.05dB gain increase. We also know that the transistor gain is 12.9dB, so must create a match on the input such that the gain contribution from the input network is approximately –1.0dB.

To create the –1dB gain circle for the input network we can use the following equations:

$$d_s = \frac{g_s|S_{11}|}{1 - |S_{11}|^2(1 - g_s)}$$

$$R_s = \frac{(1 - g_s)^{\frac{1}{2}}(1 - |S_{11}|^2)}{1 - |S_{11}|^2(1 - g_s)}$$

where

$$g_s = \frac{G_s}{G_{s,max}}$$

and

$$G_s = \frac{1 - |\Gamma_s|^2}{|1 - \Gamma_s S_{11}|^2}$$

From example 1, $G_{s,max}$ is 1.94dB or, as a number, 1.56. So g_s becomes $g_s = G_s/G_{s,max} = 0.79/1.56 = 0.506$; $d_s = 0.37$ and $R_s = 0.545$. Remember in plotting the –1dB gain circle (Fig. 2), d_s is expressed in terms of the magnitude of S_{11}. Any point on the –1dB gain circle will provide the desired value of Γ_s but for reasons of convenience in matching, the point selected is point *A*. At this point, $\Gamma_s = 0.44$ at an angle of 100°.

Computing Γ_L from:

$$\Gamma_{OUT} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S}$$

which for the case of $S_{12} = 0$ becomes $\Gamma_{out} = S_{22}$, since we know that Γ_L is the conjugate of Γ_{out}, $\Gamma_L = 0.11$ at an angle of +164°, the same value as in example 1.

Example 4. Broad-band design, S12 = 0.

Our goal is to design the *MRF571* into a broad-band circuit having 14dB of gain oper-

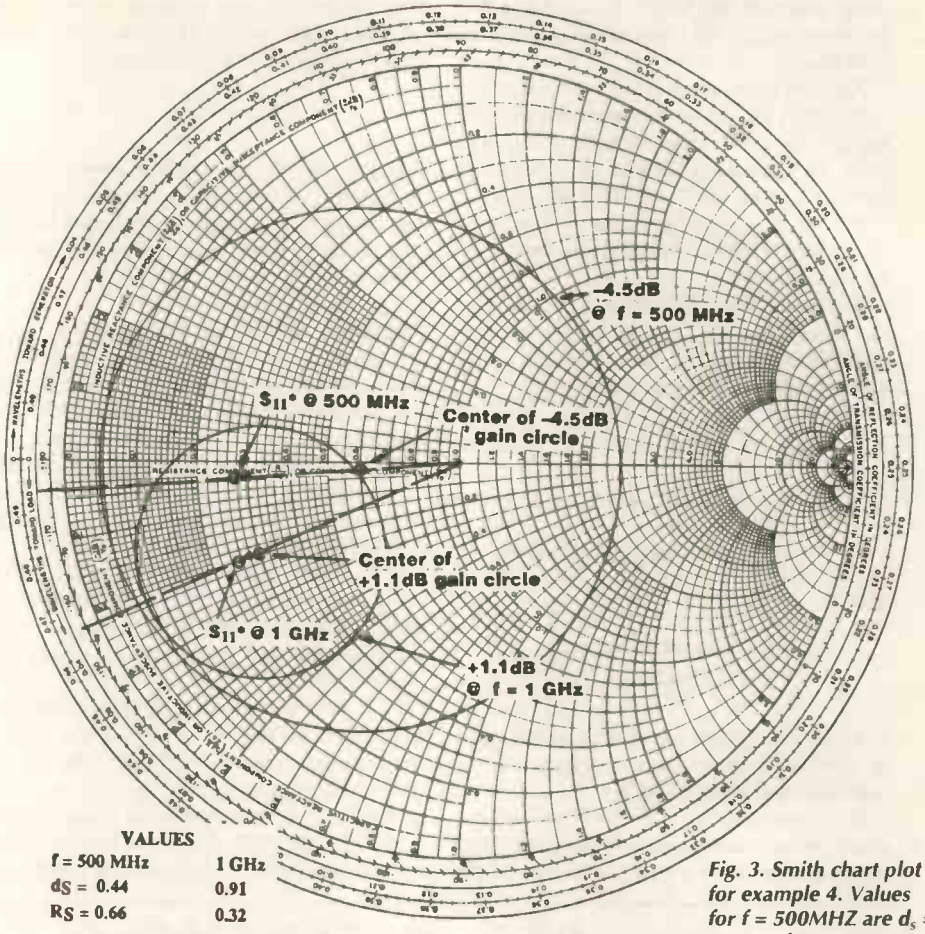

Fig. 3. Smith chart plot for example 4. Values for f = 500MHZ are d_s = 0.44 and R_s = 0.66. Values for f = 1GHz are d_s = 0.91 and R_s = 0.32.

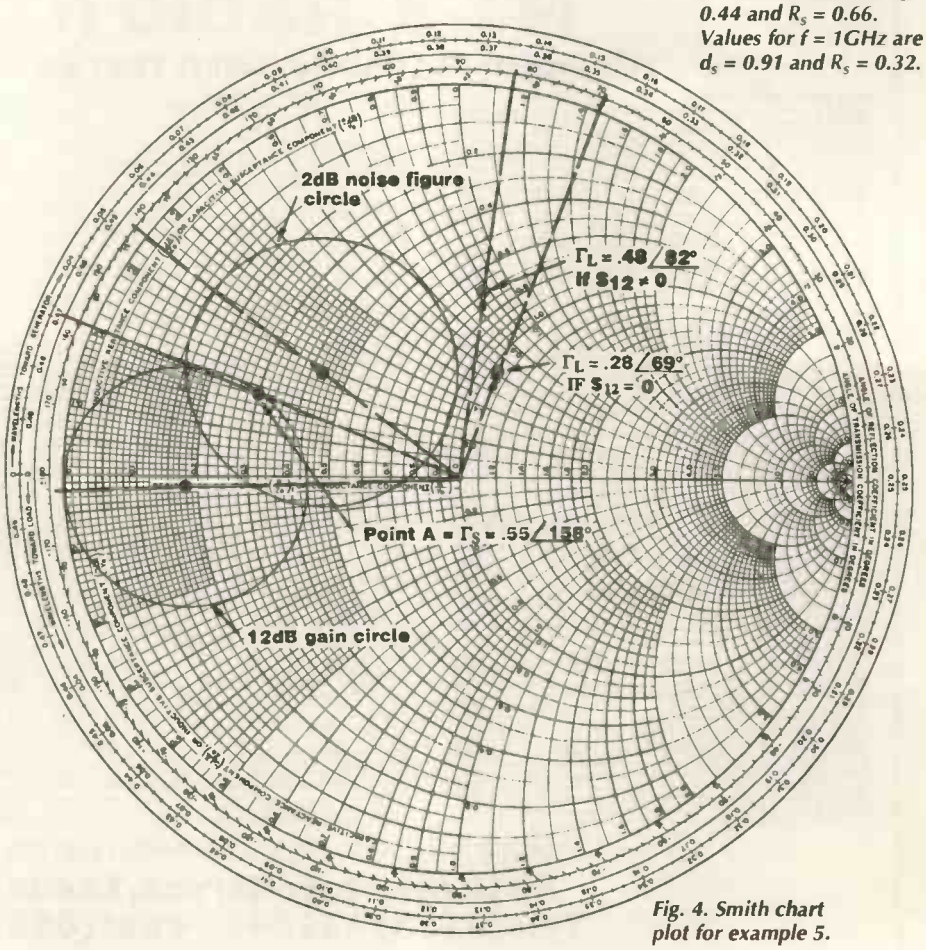

Fig. 4. Smith chart plot for example 5.

ating from 500-100 MHz. The amplifier is to be driven from a 50Ω source and is to drive a 50Ω load.

Assume, again, the bias to be 6V and 50mA – less than the maximum available gain, as seen in example 1. From the table of *S*-parameters, $|S_{21}|$ results in 70.6 or 18.5dB gain at 500MHz, and 19.3 or 12.9dB gain at 1000MHz. So matching circuits must decrease the gain by 4.5dB at 500MHz and increase it by 1.1dB at 1GHz.

The plan is to put all the 'gain' or 'loss' into the input matching network, an assumption that makes drawing of the gain circles easier using the formulas for available gain:

$$g_a = G_A/|S_{21}|^2,\ C_1 = S_{11} - D_s S_{22}^*$$

$$R_a = \frac{\left[1-2K|S_{21}S_{12}|g_a + |S_{21}S_1\right.}{\left|1+g_a\left(|S_{11}|^2 - |D_s|^2\right)\right.}$$

and

$$C_a = \frac{g_a C_1^*}{\left|1+g_a\left(|S_{11}|^2 - |D_s|^2\right)\right|}$$

Computer programs can also be useful, such as *Mmicad* which was used in determining the gain circles plotted in **Fig. 3**.

In the next issue of *EW + WW* we will use these circles to determine matching networks that place the source impedance on the –4.5dB circle at 500MHz and at the same time on the +1.1dB circle at 1GHz. Then the load impedance can be calculated using:

$$\Gamma_{OUT} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_S}$$

The process of choosing the proper source reflection coefficient is an iterative one and can be carried out manually or through the use of a computer optimisation program such as *Mmicad*.

Example 5. Designing for low noise

Up to now, gain has been the only concern. But many low power amplifiers must also be low noise so the effect that matching has on noise figures must be taken into account.

Only the source reflection coefficient affects noise figure, though it is true that the load reflection coefficient will affect the source reflection coefficient when $S_{12} \neq 0$.

If $S_{12} \neq 0$, then the simplest step to take is to choose the value of T_s that gives the desired trade-off between gain and noise figure, then use that value of T_s to calculate Γ_L from the above equation.

In this example, once again, assume the transistor is the *MRF571*. Frequency is 1GHz, but this time change the bias to 6V and 5mA – a more appropriate value for a low noise application. Another reason for these choices is that the *MRF571* data sheet has gain and noise figure contours plotted for the conditions stated. Also, further assume that gain must be at least 12dB, and the noise figure not greater than 2dB. New values of *S*-parameters are:

$S_{11} = 0.61$, angle +178°
$S_{12} = 0.09$, angle +37°
$S_{21} = 3.0$, angle +78°
$S_{22} = 0.28$, angle –69°

Examining a plot of the 2dB noise figure circle (**Fig. 4**) and the 12dB gain circle taken from the *MRF571* data sheet, shows that the 2dB noise circle intersects the 12dB gain circle in two places. Either value of G_s will result in both the desired gain and noise figure.

Selecting a value for G_s on the 12dB gain circle between these two points will produce a noise figure even lower than 2 dB.

Take the point *A* shown in Fig. 4. The value for T_s is estimated to be 0.55 at an angle of 158°. All that is left is to determine the value of Γ_L from $\Gamma_L = [S_{22} + (S_{12}S_{21}\Gamma_S/\{1 - \Gamma_S S_{11})\}]^*$.

The formula clearly shows that for the case when $S_{12} = 0$, $\Gamma_L = S_{22}^*$. So, if $S_{12} = 0$, $\Gamma_L = 0.28$ at an angle of 69° .

Taking the value of S_{12} into account, $\Gamma_L = 0.48$ at an angle of 82°. These values are also shown in Fig. 4 as *B* and *C* respectively. ■

Next article: Finding the circuits that will realise the impedances we want.

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PIR DETECTOR Made by famous UK alarm manufacturer these are hi spec, long range internal units. 12v operation. Slight marks on case and unboxed (although brand new) £8 **REF: MAG8P5**

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CIRCLE NO. 127 ON REPLY CARD

LETTERS

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

The message not the medium

Stephen Dyke may have somewhat idealised recollections of the *ZX Spectrum* and its youthful programmers (*Letters*, November), but the point he makes is an important one. Readers may remember that at the time Sir Clive was marketing the *ZX* range of computers, the BBC had a series of programmes with the title *Young Scientist of the Year*. Groups of pupils presented their science and technology projects and a panel of judges eventually chose a winning group based on the progress they had made.

Not very exciting tv. But nor did it try to be 'zany', by which I mean it did not patronise its audience. Sir George Porter, no mean chemist, was so intrigued by one problem raised on the program that he investigated it himself.

When it vanished from our screens, I was told by one of the judges that they had found that parents were spending money on equipment to help with the projects, which was felt to be undesirable. The BBC has also claimed that it became increasingly difficult to find schools willing to enter. Judging by the school stories which regularly appear in my local paper, I find this hard to believe.

The real reason may be more to do with the perceptions of science and technology which the media have manufactured for mass consumption. It is not difficult to find some 'researcher' to tell us that spending too much time with computers is addictive, anti-social, even pathological. I've yet to hear of a dedicated young musician who did not spend a great deal of time alone practising on an expensive instrument, often bought by parents. A youthful violinist next door is a great deal more anti-social than a computer nerd. But of course that is art.

For the most part science and technology on television and radio have become subordinated to programmes about the effects of science and technology – not at all the same thing. In the BBC2 series *White Heat* it was considered appropriate that the majority of the contributors were historians, cultural critics, sociologists and the like.

Too many scientists and technologists pander to this approach by telling their colleagues they must be 'better communicators'. According to this view, watchers and listeners are merely passive sponges and must never be required to engage their intellects. If that is true, where do the audiences for *Radio 3*, *The Late Show* and *Kaleidoscope* come from?

Having seen blue collar unemployment rise because many of our manufacturing industries have been exported to the third world, we now face the prospect of increasing competition from the huge white collar sector of countries like India. It is not enough that we are a nation of trained and competent technology users; someone has to beat the drum for our being a nation of science and technology innovators.

Les May
Rochdale

Amplified disagreement

I would like to thank Doug Eleveld for his kind comments (*Letters*, *EW + WW*, November) on my recent amplifier writings. However, I am now placed in the position of disagreeing with much of the rest of his letter.

He asks, if a complementary mosfet common-source output stage is driven by a current-source, "then do we not have perfect dc linearity in the output?" My answer can only be no, we most certainly do not. Power fets are rather non-linear devices and driving them with a current source – which will in itself have some non-linearity – does not alter this. As I explained in the series *Distortion in amplifiers*, part III[1], it is in any case almost unheard of for an output stage to be driven by a true current source: off-hand I can think of no examples.

Commonly we talk of voltage-amplifier-stages (vas) with current-source collector loads. But this verbal shorthand glosses over the fact that an explicit dominant-pole capacitor is almost always required. Even if it appears to be missing the C_{cb} of the vas transistor, it is carrying out much the same function: giving local negative feedback around the vas which has its output impedance reduced in the classical manner. Since the feedback element has an impedance that falls with rising frequency, the feedback factor increases and the vas output impedance decreases as the frequency rises.

In a typical amplifier the vas collector impedance might be tens of kilo-ohms at very low frequencies, but it may well have dropped to below 1kΩ at 20kHz. I do not think this can be meaningfully called current-drive, or come to that, voltage-drive either.

The value of the vas current-source can place an unexpected extra limit on the maximum positive-going slew-rate[2]. But this is not the most important limit, which for conventional architectures is still set by the input tail-current.

I cannot follow Eleveld's arguments concerning oscillation being less likely with current-drive. Were it possible to make such an amplifier, I would expect the stability to be very doubtful if the vas pole was undefined, or rather defined only by the internal mosfet capacitances. These are both large and variable.

Biasing the amplifier suggested by Eleveld does indeed seem tricky. This appears to be a major problem with common-emitter output stages. Eleveld does not state if the design is intended to operate in class A or B (or some mixture of the two) but the biasing scheme shown does not seem appropriate to either.

It cannot work for class B, because the summed diode voltage applied to the low-pass filter would seriously vary with the output signal level. Real power diodes will only give very approximate clipping, and they do not make any detectable change to linearity of the output stage. (I tested this by Pspice simulation, using *2SJ50/2SK135* power fets and *MR752* 6A diodes.)

If class A is intended, biasing should work providing the diodes shown are replaced with the usual small-value resistors, as in pure class A. The sum of voltages across the two resistors will be constant, and in fact the low-pass filter is redundant[3].

This is about as far as I plan to go in guessing how Mr Eleveld's idea is supposed to work. In putting an idea that could well be useful in the public domain, it might well be wise to explain its intended operation in more detail.

Douglas Self
London

References

1. D Self. Distortion In Power Amplifiers, Part 3, *EW + WW*, Oct 1993, p.821.
2. D Self. High Speed Audio Power. *EW + WW*, Sept 1994, p762.
3. D Self. Distortion In Power Amplifiers, Part 8. *EW + WW*, Mar 1994, p227.

Fine point

May I say how much I enjoyed reading George Pickworth's article on electrolytic detectors (***Detection before the diode***, *EW + WW*, December, p1003-1006)?

Perhaps I can throw some light on the term 'Wollaston wire', which Pickworth notes is often used in reference to platinum wire. The term refers to a particular process for producing very fine wire.

"By Wollaston's procedure, a platinum rod is covered with a close-fitting silver tube, and the composite rod is drawn through wire dies. After the silver has been etched

Still waters run deep...

Having read of the fantastic improvement that low resistance loudspeaker cables can make, I remembered I had some lengths of copper water pipe spare. So I substituted these for the cable (wife a little upset with floorboards up). I was very disappointed with the results. I mentioned my disappointment to a homeopathic friend and he explained that as the pipes had been previously used for water which had run over rocks, and so still had the memory of the rock water, I could only expect an improvement with rock music.

R A Ellis.
London

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The minimum obtainable diameter of 0.5µm shows that the process is more than able to meet the specification for the BIWC cell.

J S Linfoot
Oxford

Blowing the whistle on referees

Pete Davis (*Letters*, December) queries suggestions of a scientific establishment conspiracy and states a naive belief in respect for objective evidence. This is not my own experience.

There may be no formal conspiracy, but many scientists display attitudes which seem to differ little from those of extreme religious fundamentalists for whom their holy book is the incontestable truth.

Where is the claimed objectivity of science when more open minded-scientists risk their careers merely because they make suggestions or try experiments to test the validity of someone else's contrary idea without in any way expressing a strong personal belief?

The practice of refereeing papers for professional journals, although necessary, is inherently wide open to bias by bigots or those with vested interests. Too often new concepts – later fully accepted – are suppressed for decades. At least one has had to finance a new journal because no other would publish. Why cannot responsible editors expose ideas to full discussion as soon as possible? A percentage of editorial pages could be allocated to potentially controversial papers, refereed only for logic and sound method.

The current approach is largely self defeating as many young scientists grow up with an unquestioning acceptance of establishment ideas. Some no longer even understand the need to question.

It now seems fashionable to knock science. Unfortunately, now an ill-informed general public is left unable to distinguish between the claims of ignored or rejected novelty and a flood of unanswered, simplistic and half-baked pseudo science.

R G Silson
Herts

Un-square wave

Many writers, including Ian Hickman (*Design Brief: Cautionary tales for circuit designers*, November, pp.926-930) state that a unity mark-space ratio square wave may be obtained by dividing an asymmetrical pulse train by two.

This is not exactly true. Referring to cmos technology, tp_{LH} is typically slower than tp_{HL}, attributed to use of complementary devices with differing carrier mobility in the output stage. Our 'square wave' is actually a slightly asymmetric trapezoid.

This can easily be seen on a spectrum analyser by the presence of low-amplitude even harmonics ("Spectrum analysis on the cheap", Nick Wheeler, *EW + WW*, March 92, p.205) and occasionally can be seen on a good oscilloscope.

In most cases the distinction does not matter. But I found I could not generate a spectrally pure sine wave by a combination of low-pass filtering and notch filters on the odd harmonics.

Nick Wheeler
Sutton

Interface signal

C J D Catto's design (*Circuit Ideas, Interfacing a signal riding high*, November, p. 921) is most useful. Using differential inputs to one single operational amplifier avoids the drawback in the analogue-computer approach of inverting one input to mix it with the other, imposing different frequency characteristics on the two inputs thus wrecking the cmrr at high frequency.

It is also the only approach that will accommodate inputs well outside the supply rails by attenuating the inputs, since superior circuits for responding only to differential inputs cannot be used directly.

However the published circuit needs minor corrections such as a blob between R_5 and R_6 and another change concerning R_{16}, while trimmer R_{17} in the text should read trimmer P_1. The first stage differential gain from the input terminals is also not 13.5 but about 1/18.

To obtain the quoted common-mode rejection of 90dB, ten resistors must restrict their total deviations with time and temperature to one part in 31,600 – or 31.6ppm. But these are to be added algebraically, taking R_6 with R_8 as a parallel pair and R_9 and R_{10} likewise.

Further, the layout of R_1–R_4 and their screening should be symmetrical, with wide spacing, and the inputs be via terminals/sockets in a metal panel. Where $R_1 + R_3 =$ 1.36MΩ, an extra 1pf across these end-to-end would have a reactance of 33180MΩ at 50Hz and the phase of that input would be shifted by 1/2400 radian or so. Consequently the cmrr would not adjust above about 68dB. As well as permitting 250V input, use of two or more resistors in series at the input is valuable in keeping down such end-to-end capacities of the resistor chains.

Bernard Jones
London

Radar search

I am searching for basic information on development of radar, as I have uncovered that its principle and public knowledge of it was firstly revealed in *EW + WW* by Mr R L Smith-Rose in 1945.

How was this device invented, and where was it first successfully built? I understand it was actually first built in 1935, by Sir Robert Watson-Watt. So who was really credited as the inventor of the first radar system?

I also learn that the first radar in the US was built in 1936, by Major Corput and Watson (not the same Watson-Watt). Does any reader have technical information on what kind of rays or source of rays (radio-frequency, wavelengths, type of oscillator etc.) were used that first time, or on the receiving antennas employed?

I am planning on an article on the impact of radar in naval traffic and civil aviation in Norway, and would be most grateful for any help.

Roy Albrigtsen
Feltspatkroken 3
N-4028 Stavanger
Norway

Relaxing pursuits

Your *Comment*, (Abuse of the licence fee, November) points out that the BBC is neatly fulfiling its remit with five national and one local station.

The other 'local' stations are usually vehicles for commercial advertising. As a product of, and stimulus for, the market economy, no doubt they will emulate the BBC and eventually distribute dab, or something like it, to local transmitters via satellite. Commercial media outlets are steadily being underpinned by multi-national conglomerates, so it will probably not take long.

It is worth noting that any broadcast reception must be licensed in the UK, even if via a dish aligned to a foreign satellite. Those of us already acquiring digital music via the German TV Sat 2 are about to lose out, as it is being sold to the Swedes and re-located. The same 16 stations can be received from the lower-powered Kopernicus, albeit via a much larger dish, with a fat fee for local planning permission. The only alternative will be to pay for it from Astra early next year. Even so this is still a lot cheaper than buying cds.

On the subject of audio, I am currently completing a total of five Doug Self amps for split filters and sub-woofers and would dearly like to see an update of the practical active filter published by *EW + WW* in 1978, a version that used over 30 op-amps in the full implementation.

Hugh Haines
Sunderland

Fighting words

I was interested to read outgoing editor Frank Ogden's comments concerning the demise of the British electronics industry (*Comment*, *EW + WW*, October).

But *EW + WW* must take some of the blame. In the past, editorial has been extremely critical – to the point of vitriolic – of the Ministry of Defence's involvement in the electronics industry and has busily promoted pacifism. As a respected voice, obviously the magazine must have had some influence.

Defence cuts have savaged the armed forces and consequently, the MoD is no longer the massive purchasing power that it was, while the electronics industry has lost a major customer. Now we see the consequences.

Defence can be a large customer of home industry and it is my belief that the large scale reduction in the British armed forces over many years has lost the country a large manufacturing base covering many technologies.

Interestingly, because of the lack of British entrepreneurial initiative, we are now rated as a third world country by foreign industrialists setting up manufacturing bases here to take advantage of our current economic state. Fortunately, but probably not in my life-time, the trend will be for Britain to climb its way back to the top.

Colin Long
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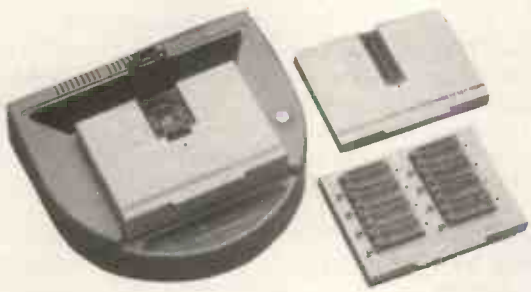

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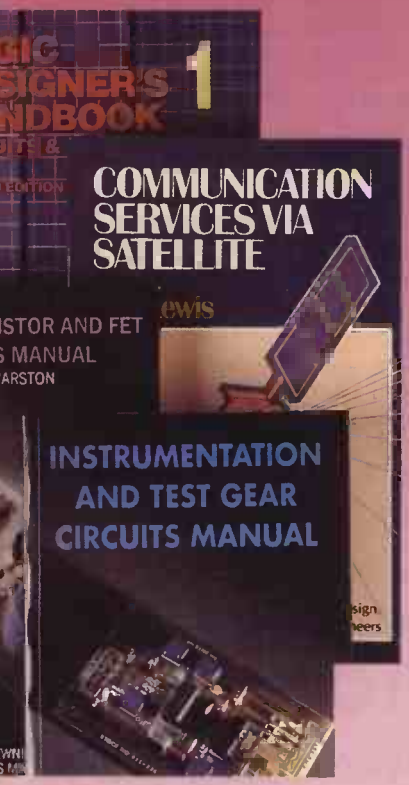

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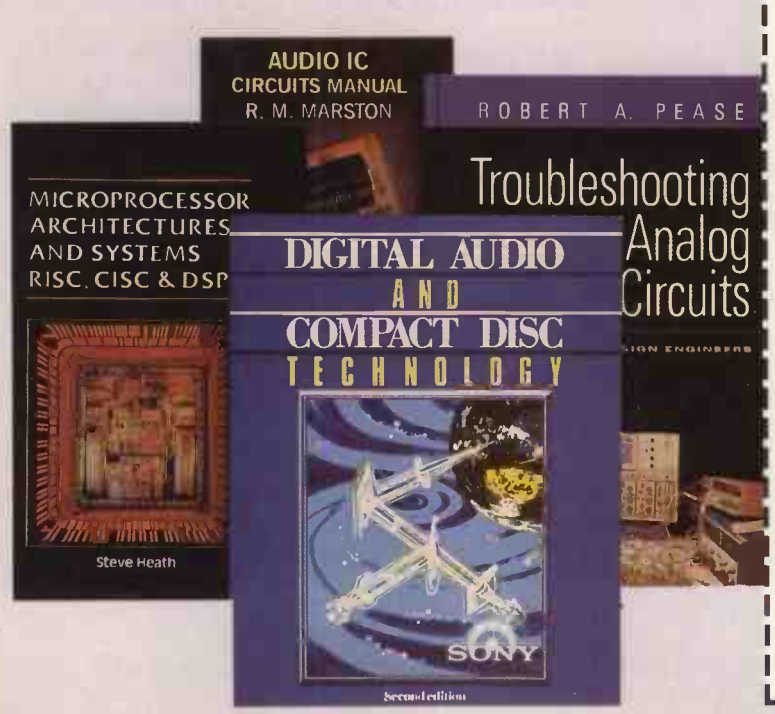

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Analysing waveforms in the practical world

In the second part of his examination, Ian Hickman uses real equipment to check how practical waveforms match up to their theoretical shapes.

I have shown how relative amplitudes and phases of the harmonics of a square wave relative to the fundamental component can be correctly shown (Figure 6b from last month, reproduced here as **Fig. 1b**). But it may seem odd at first sight that amplitude of the fundamental itself is 0.637, or 63.7%, or –4dB. 63.7% of what? Certainly not of the amplitude of the square wave.

Summing $1 - {}^1/_3 + {}^1/_5 - {}^1/_7 + {}^1/_9 \ldots$ to infinity gives 0.782 as the amplitude of the square wave whose fundamental component is unity. Taking the reciprocal gives the fundamental component of a ±1V square wave as ±1.278V. This brings us to the question of the reconstruction of sine waves from their digital representations.

Take a digital audio system built around a compact disc, digital audio tape or digital compact cassette system – or even for the present purpose of explanation a straight-through wire connection (**Fig. 2a**). Assuming the wire connection, and a commonly employed sampling frequency such as 44.1kHz, then the wave form (**Fig. 2b**) with exactly 48 steps per cycle, represents a frequency of just over 900Hz. Clearly, after low-pass filtering to knock off the ripple, it will be a virtually exact replica of the original, 100%, a gain of unity or 0dB down.

We know a digital sampling system can represent frequencies (almost) right up to the 'Nyquist rate', or half the sampling frequency. When the input to the A-to-D is just below the Nyquist rate, at first sight the D-to-A output (**Fig. 3a**) does not look very much like the input signal. But this is because it contains the image frequency component, which is just *above* the Nyquist rate.

However, when that output waveform is passed through a low pass filter with a steep cut-off at the Nyquist frequency, the image component is suppressed (along with the sharp edges representing harmonics of the wanted and image frequencies) leaving just the original sine wave. Assuming that the sine wave amplitude is ±1V peak to peak, we can see (**Fig. 3b**), that the square wave out of the D-to-A varies in amplitude between zero and ±1V peak to peak.

At its maximum, the fundamental component of the square wave will be, as noted earlier, ±1.278V peak to peak. The wanted frequency and its image are equal in amplitude, so after the low-pass filter has suppressed the image and all the harmonics, the amplitude of the wanted signal will be 1.278/2 or ±0.637V peak to peak. While the fundamental component of the 48 step representation of the sine wave shown in Fig. 2b barely differs in amplitude from the original, a signal at – or at least very close to – the Nyquist rate will be, after filtering, only 63.7% of the true amplitude. Or 3.9dB down, as predicted by the sin *x* upon *x* curve.

Similarly, a frequency at a quarter of the sampling rate (or half the Nyquist frequency) will be 0.9dB down. The expression sin *x* upon *x* predicts the loss at any frequency and explains why better digital audio reconstruction chips include a 'sin *x* upon *x* compensation filter'. Such a filter will correct the mild high frequency roll-off which the signal would

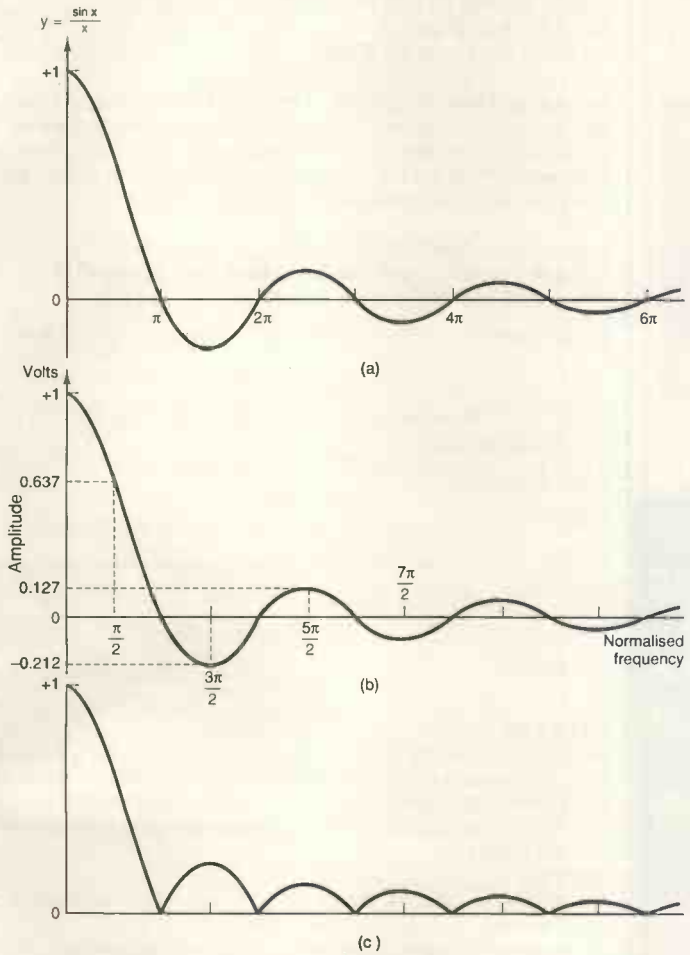


Fig. 1a. Sin x upon x.
1b. Value of y gives the amplitude and phase of the various frequency components of a square wave.
1c. Modulus of 1a.

otherwise suffer, for all those golden ears who would otherwise be disappointed by the reduced level at 20kHz.

Asymmetrical fit

The sin x over x curve actually fits the spectrum of all sorts of asymmetrical square waves and pulse trains. In fact, an equal mark:space ratio pulse train is simply a square wave with a dc component **(Fig. 4a)**. If now the pulse width of such a train is kept the same, but the space between pulses increased slightly, the frequency of the fundamental will be slightly lower than shown in Fig. 1b, so that its second harmonic will now fall just to the left of the first zero of y **(Fig. 4b)**. The waveform now has some even harmonic components. If the space between pulses were further widened to twice the pulse width, the frequency would fall to two thirds of that shown in Fig. 1b. The spectrum would then have both odd and even harmonic components, though the 3rd, 6th, 9th... harmonics would all be zero.

Reducing the mark:space ratio further to 25:75%, still with the same pulse width, reduces the fundamental frequency to half that shown for the sin x over x curve. It would be the 4th, 8th, 12th... harmonics that would be zero.

Testing theory

Of course theory and practice must tie up – mustn't it? Just to make sure, I built a circuit to check **(Fig. 5a)**, using fast *74ACT* devices clocked at a leisurely 4MHz, and giving out a 250ns pulse every µs **(Fig. 5b)**.

The waveform spectrum **(Fig. 5c)** showed a null at the fourth harmonic, looking fairly convincing even on a linear scale. But note here that we are using a log scale – so it is actually well over 50dB down on the fundamental, or less than 0.3%.

Feeding the values of frequency of the fundamental, second and third harmonics – namely $\pi/4$, $\pi/2$ and $3\pi/4$ – into the formula gives values of amplitude for these components of 0.90, 0.64 and 0.30. From these figures, the second harmonic calculates out as 3dB down, and the third as 9dB down on the fundamental. This ties up pretty closely with the spectrum achieved (Fig. 5c), which also shows the higher harmonics falling into the pattern shown in Fig. 1c – allowing for the vertical scale being logarithmic, not linear. So the sin x upon x curve is an envelope or locus which passes through the values of amplitude of the component frequencies of any pulse waveform.

Reduced pulse width

The foregoing examples illustrate that if the width of the pulses or 'marks' is held constant but the width of the intervening spaces is increased, so lowering the

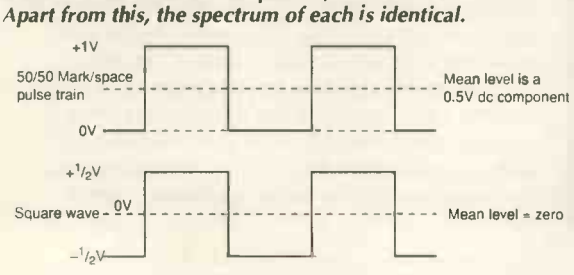


Fig. 4a. The only difference between a square wave and a unipolar pulse train with unity mark/space ratio is that the latter has a dc component; the former does not. Apart from this, the spectrum of each is identical.

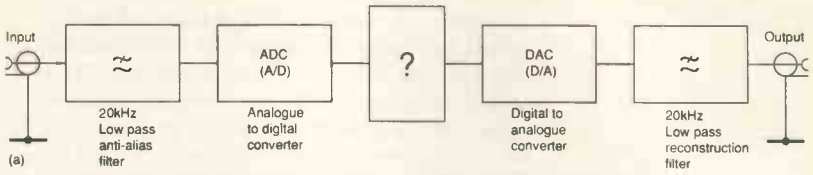
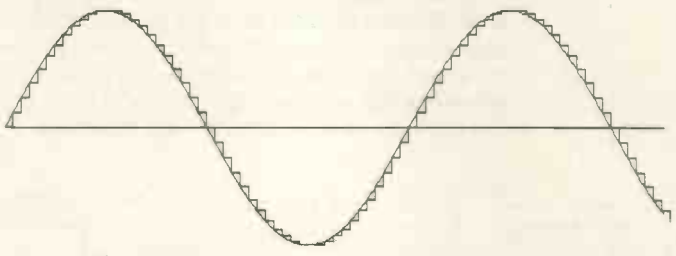


Fig. 2a. Basic digital audio signal chain. The ? block could represent a compact disc, digital audio tape or digital compact cassette storage.



2b. A sine wave of around 1kHz as it might appear at the output of the D-to-A converter, before low pass filtering. Shown with the original sine wave.

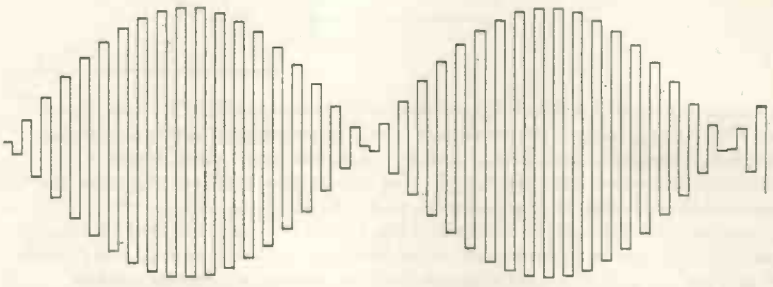
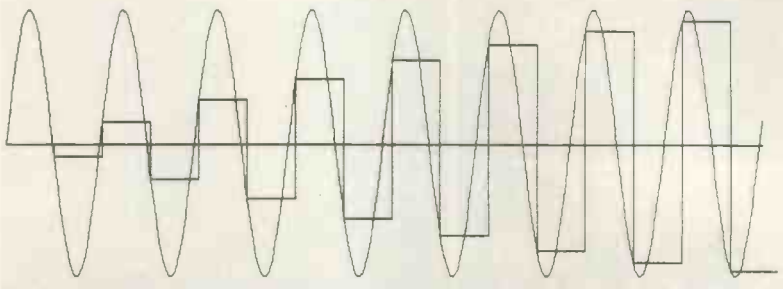
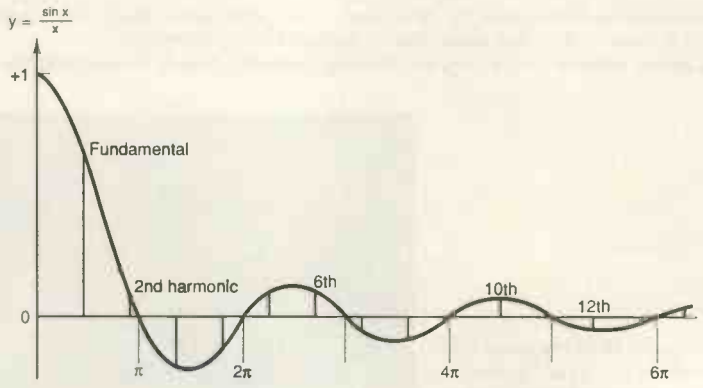


Fig. 3a. Waveform of the output of the D-to-A converter for a frequency just below the Nyquist frequency – not at first sight very much like the original sine wave.



3b. The left hand end of waveform 3a expanded in time and showing the original sine wave as well.



4b. If the width of the pulses in 4a is kept the same but the distance between them is increased slightly, then the frequency is slightly lower. The second harmonic will no longer fall on a zero of the sin x upon x function.

frequency, the positions of the the zeros of the sin x upon x function are unchanged but the spectral lines become more numerous and closer together. In this way, the frequency of the waveform determines the spacing of the spectral lines within the first loop of the sin x upon x function (and elsewhere), the lines being, of course, equally spaced everywhere.

But what happens if the prf (pulse repetition frequency) is fixed while the pulse mark:space ratio is reduced by increasing the space at the expense of the mark? Because the prf is fixed, the position of the lines on the x or Ω axis will remain the same. But as the second (and other even harmonics) must start to appear, the first zero crossing of the sin x upon x function must move to the right – and further and further to the right as the mark:space ratio becomes smaller. As the pulse narrows down to infinitesimal width – and its amplitude is greatly increased if it is still to convey a finite amount of signal power – the first zero-crossing moves out towards infinite frequency. The spectrum of such a 'delta function' consists of a component at the fundamental frequency plus all its harmonics, all in phase and all of equal amplitude.

The effect can be demonstrated, approximately, by differentiating a square wave and then picking out just the positive-going edges with a fast diode **(Fig. 6a)**.

The 10MHz square wave was obtained from the same generator as used for the 4MHz clock in the above circuit, but this time using its much faster *SN74128* 50Ω line driver output. Time-constant *CR* of the differentiator is 4.7ns or nearly 5% of the square-wave's period. So the resultant pulses won't be exceedingly narrow – just 4.7ns wide at 37% of full amplitude giving a near-ideal square-wave drive – and will be a different shape to that so far considered. Nevertheless, the resultant spectrum **(Fig. 6b)** is clearly going in the right direction.

Alternate harmonics (the even ones) are slightly smaller than odd harmonics because of incomplete suppression of the negative-going spikes by the diode. If the diode were shorted, feeding both positive and negative spikes to the spectrum analyser, even harmonics would disappear completely. Last month (Fig. 5, December) we showed that spikes of both polarities involved no even harmonics.

A train of narrow pulses, ie a comb of near-equal amplitude sine waves like this, has numerous uses – the lo (local oscillator) drive in a superhet receiver, for instance. The lo is arranged to run at a very much lower frequency than the band of interest, feeding narrow spikes to the mixer. Any signal within the band width defined by the receiver's front-end band pass filter is likely to be near enough to one of the comb of 'local oscillator frequencies' to be received. This permits a sensitive narrow-band (and hence low-noise) receiver to cover the whole band at once without the delay taken with a single swept lo approach – handy if you are sweeping a room for radio bugs.

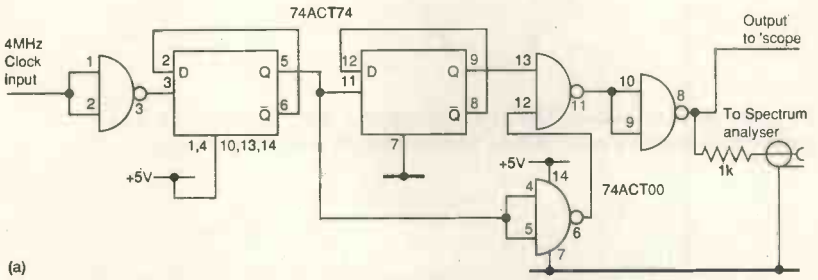
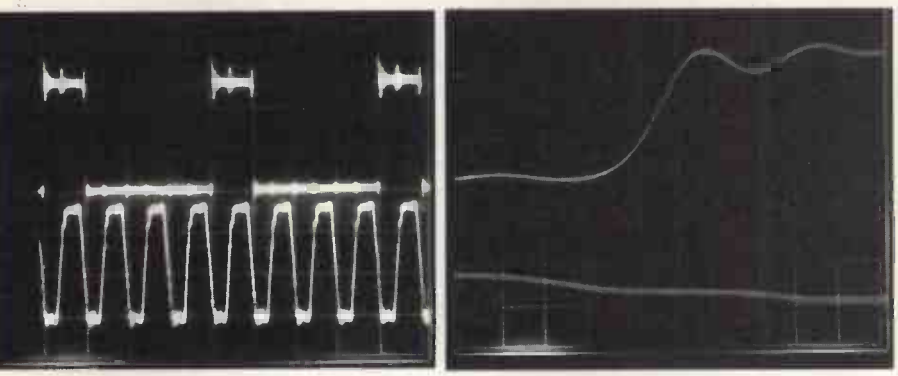
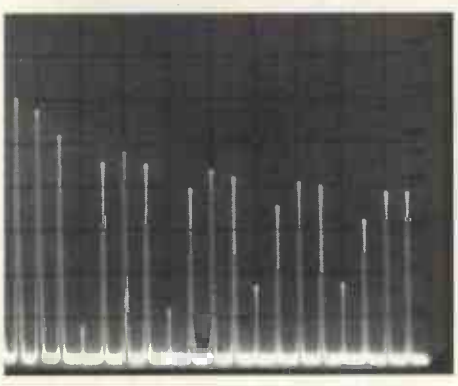


Fig. 5a. Circuit used to produce a pulse train with an accurate 25:75% mark:space ratio. The fourth section of the Nand gate is used to provide a delay to match that through the second flip-flop. Before adding it, a narrow sneak pulse or glitch occurred at the output when the first flip-flop output went positive, triggering the second flip-flop's output to change from 1 to 0. This glitch was due to the propagation delay through the second flip-flop before its output fell to the 0 level, resulting in a couple of ns during which the Nand gate inputs at pins 12 and 13 were both positive. To get the ringing down to the level shown, considerable care in connecting the probe was necessary, discarding its usual earth lead in favour of a short thick wire strap to circuit ground. Similar precautions were taken in connecting the signal to the spectrum analyser.



5b. Waveforms associated with 5a. (left) 4MHz cmos input clock (lower trace) and 1MHz 25% duty cycle pulses (upper trace) both at 2V/div, 250ns/div. (right) Leading edge of (left) at 1ns/div, showing pulse rise time to be less than 2ns.



5c Spectrum of the pulse train produced by circuit 5a, display centre frequency 10MHz, 2MHz/div horizontal, ref level 10dBm, 10dB/div vertical, if band width 30kHz, video filter off.

Sine wave continuum

Now let us look at the case where not only does the pulse-width tend to zero (pushing the frequency of the first zero of the sin x upon x function out towards infinity), but its frequency tends to zero also. The lower the frequency of the narrow 'lo' pulses, the closer together are the spectral lines. Ultimately the spectrum becomes an infinite number of sine wave components all of the same amplitude, so closely packed together they represent a continuum.

In Fourier analysis, sine wave components are assumed to exist from time equals minus infinity until plus infinity. That is fine as long as the square wave (or whatever) being represented also exists permanently. But it can lead to some interesting, though idle, speculation when the prf falls to zero. In this case the components represent an isolated pulse occurring at time $t = 0$. The myriad of frequency components of negligible amplitude can be pictured buzzing away, so phased that their sum is zero at every instant, except when the magic moment $t = 0$ arrives. At that point they are all uniquely in phase, producing just the one isolated, infinitely-narrow, never-to-be-repeated spike.

No pulse paradox

Suppose a programmable pulse generator has been set up to give a single narrow delayed pulse so that, say, $t = 0$ is set for 3.00pm next Tuesday afternoon. What would happen to all those components if, just before that time, there was a power cut so that unexpectedly the pulse did not after all occur? There is really no paradox here. The existence of harmonic components of a single isolated pulse prior to that pulse is only a mathematical convenience – such precursors are 'non-causal'. If all those components could suddenly be 'turned on', each at the same amplitude and the correct phase (all at zero, positive-going), they would add up to give that single isolated pulse. They would also never all simultaneously come in phase again to produce another pulse. So, they might as well be all turned off again. As 'post-cursors' they are also ineffective or non-causal.

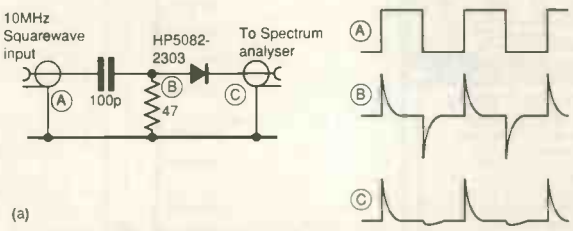
Key experiment

But that doesn't mean their combined effect does not linger on, as can be proved with a simple experiment.

Many homes contain an audio spectrum analyser, of the filter-bank variety, with considerably better frequency resolution than the usual third-octave filters.

Hold down the loud pedal with a couple of bricks, and take off the front (or raise the lid, if you run to a grand piano) and the experiment can begin. Wait for complete quietness (let $t = 0$ be the middle of the night if necessary) and clap loudly, once.

All the piano strings will be heard to sound, each responding to its particular frequency component of the near-delta-function. The sound soon dies away, since the strings exhibit a finite Q. If they were not losing energy in the form of sound waves we would not be able to hear them. But if the Q of the individual filter bank channels were near infinite, the post-cursors would indeed continue, almost, for ever. ■



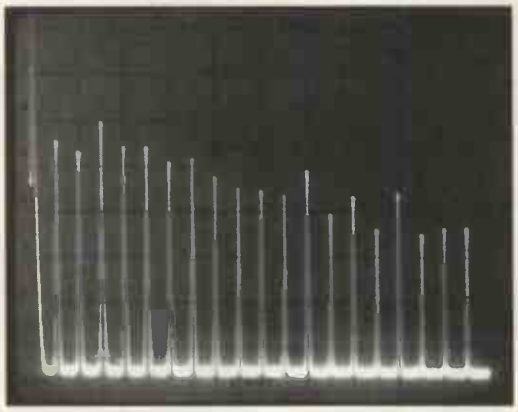


Fig. 6a). Producing some narrow pulses at a prf of 10MHz 6b). The spectrum of 6a approximates a sea of equal amplitude 10MHz-spaced spectral lines stretching out towards infinite frequency. Centre frequency 100MHz, 20MHz/div horizontal, 10dB/div vertical, if band width 100kHz, video filter max.

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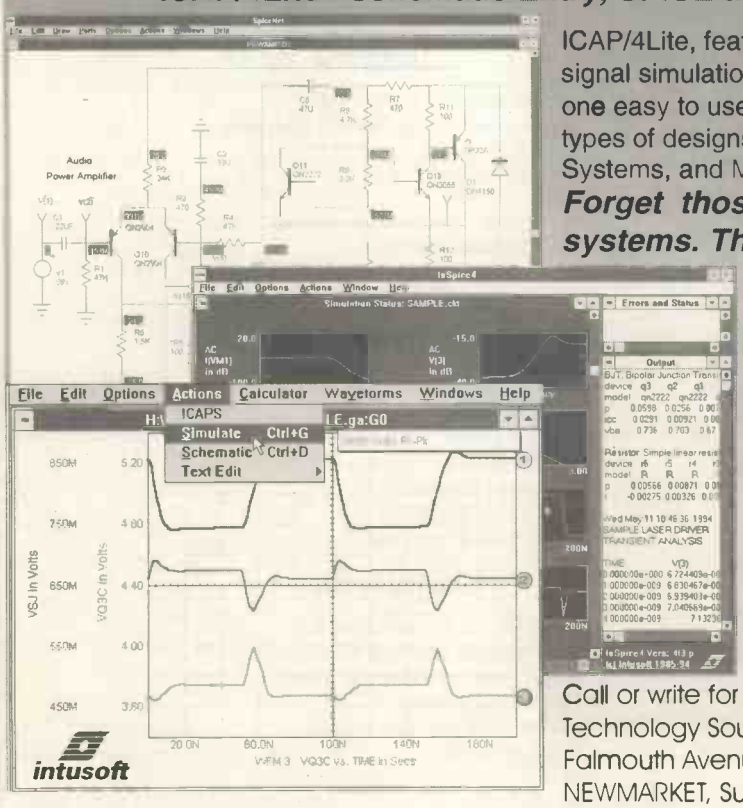

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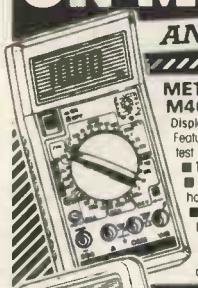

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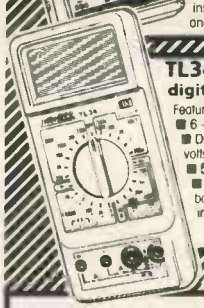

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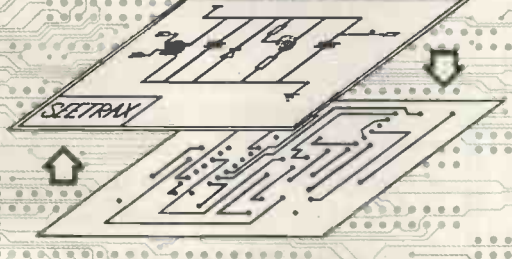

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ACTIVE

Asics

0.6µm cmos. AMI has a new 0.6µm cmos digital asic process, two libraries being announced – *AM16G* gate array and *AM16S* standard cell. Each individual pad can be driven independently by 3V or 5V and the core can be either. Performance figures for 5V are 135ps for a fanout of 2 and 200ps for a fanout of 2 with 2mm interconnect; for 3V, 185ps and 280ps, all for a 2-input Nand. Sabre Advanced Microelectronics Ltd. Tel., 01420 22004; fax, 01420 22008.

Dense, fast asics. Toshiba's new 0.4µm asics are 70% denser than earlier 0.5µm cmos devices and offer 20% shorter delay times. The *TC200* series holds 750,000 usable gates with delay times of 0.19ns at 3.3V. *TC200G* are gate arrays, *TC200C* cell-based ICs and *TC200E* embedded arrays; development orders for all types are being taken. Toshiba Electronics (UK) Ltd. Tel., 01276 694600; fax, 01276 691583.

Microprocessors and controllers

User-defined microprocessors. *MiniRISC* family processors by LSI Logic enable system-on-a-chip design by producing scalable, user-defined processors. The first three cpu cores in the MiniRISC *R4000* family are *CW4001/4010/4100* for low-power embedded use in the 25-250mips range. To make an application-specific processor, an engineer selects the core on performance/price and then uses the *MiniSIM* architectural simulator to evaluate other configurations, incorporating or modifying existing processor blocks, creating his own floating-point unit, cache, bus interface or memory management. Finally the whole system is integrated to produce the chip. The MiniRISC family has the support of a suite of tools consisting of the simulator, *Verilog/VHDL* models, a prom monitor, system verification and evaluation board. LSI Logic GmbH. Tel., +49 8945 836130; fax, +49 8945 836138.

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

Audio a-to-d converter. Analog Devices's *AD1877* 5V, 16-bit analogue-to-digital converter has a fourth-order modulator, three-stage decimator, voltage reference and lock divider. It directly interfaces with microprocessors and dsps, having two serial output ports with eight user-defined modes for master or slave operation. Dynamic range in the 20Hz-20kHz input bandwidth is 92dB and s:n ratio is 88dB. Analog Devices Ltd. Tel, 01932 253320; fax, 01932 247401.

Discrete active devices

GaAs fets. Improved versions of Mitsubishi's *MGF0900B* series L and S band power GaAs fets, the *MGF0906B/7B* give Class A operation at output powers of 37/40dBm at 2.3GHz. Gain at 2.3GHz is 11/10dB. Mitsubishi Electric UK Ltd. Tel., 01707 276100; fax, 01707 278692.

Digital signal processors

52Msample/s down-converter. Harris has the *HSP50016* digital down-converter with a 52Msample/s maximum input sample rate, which will extract a narrow baseband signal from a wide-band IF. It has a synthesiser, a mixer and two low-pass filters, tuning centre frequencies to a resolution of less than 0.009Hz to mix with a LO signal of over 102dB spurious-free dynamic range. Micromark Electronics Ltd. Tel., 01628 76176; fax, 01628 783799.

Memory chips

32-bit cmos srams. Organised as 128K by 32, EDI's *EDI8L32128* 4Mb cmos sram accesses in 17ns and is contained in a Jedec 68-pin PLCC, which allows expansion to 512K by 32 with no change to the pcb. Four chip-enable lines control one byte to allow selection of 256K by 16 or 512K by 8 organisation. EDI (UK). Tel., 01276 472637; fax, 01276 473748.

Mixed-signal ICs

Analogue switches. Siliconix has a new family of general-purpose quad analogue switches using a high-voltage silicon-gate cmos process, which affords an on resistance of 45Ω, leakage of 20pA, t_{ON} of 120ns and a power consumption of 0.7nW. The six-member range consists of *DG201/202B, DG308/309B and DG211/212B*, all of which show a smaller signal loss, greater accuracy, better sample rate and fewer transients than the earlier series. Siliconix/Temic Marketing. Tel., 01344 485757; fax, 01344 427371.

Remote control. Goldstar *GM3043/4* remote-control transmitter ICs contain everything needed to interface a keypad to a Tx diode in a 20-pin SOP working on 2-3.3V. The *3043* is used for up to 32 function keys and has an led driver, while the *3044* takes 32 keys and three dual-action keys, both handling up to 256 custom codes selected by an external diode. Since both operate with 16-bit ppm codes, the diode is not essential. Also available are the *GM3274/6* preamplifiers to amplify the incoming photo-diode signal and produce ttl/cmos output; the *3276* has a trap to eliminate the effects of high-frequency fluorescent lights. Flint Distribution. Tel., 01530 510333; fax, 01530 510275.

Optical devices

Linear optocoupler. C P Clare's *LOC110* opto-coupler uses two phototransistors, one of which provides feedback to compensate for the infra-red led's time and temperature characteristics, so that the other transistor gives an output proportional to the led drive current. Bandwidth is over 200kHz. C P Clare International nv. Tel., +32 12/39 04 00; fax, +32 12/23 57 54.

Light pipes. *Optopipes* by Dialight are light pipes, attached to a circuit board to convey the light output of a surface-mounted led to the required viewing location. Variations include 3mm and 5mm round pipes, square

Bright led panel lamps. *Dialight's T1 3/4 and T3 1/4* led panel lamps are provided with colour caps matched to the wavelength of the led light to give higher brightness than is usual and form a long-life alternative to filament bulbs. They come in red, ultra-red, green and yellow, fitted with midget flange, wedge, bl-pin and bayonet bases. Dialight. Tel., 01638 665161; fax, 01638 660718.

Higher-power lasers. *NDL7401P* laser diodes by NEC give an output of 3.5mW, against the 1mW of the company's earlier types, although power consumption is the same, as are the other analogue characteristics. Spectral width is 1.3nm, so that fewer repeaters are needed over long distances. No cooling is required. NEC Electronics (UK) Ltd. Tel., 01908 691133; fax, 01908 670290

types and one to fit over an RJ45 jack for lan use. Dialight. Tel., 01638 665161; fax, 01638 660718.

Oscillators

Skewless clock generator. A very intelligent chip, the Micro Linear *ML6510 Super PACMan* automatically removes clock skew without any extra feedback wires. Eight independent outputs determine the length of the track they drive on the pcb by measuring reflection time from the end of the track, and adjust the timing of the output. Timing adjustment is an analogue function, so that sub-nanosecond variation, including that introduced by manufacturing tolerances, is manageable. Ambar Components Ltd. Tel., 01844 261144; fax, 01844 261789.

Programmable logic arrays

26,000-gate fpga. AT&T's *ATT2Cxx* ORCA field-programmable gate arrays, based on a 0.5µm three-level metal cmos process, run at clock speeds up to 150MHz, the *2C26* having 26,000 gates. Pin-pin and clock-out delays are less than 11ns and, using *PREP* benchmarks, *2C26* delivers an average capacity of 104 instances and a speed of 50MHz. The devices are compatible with the original ORCA types. *2C26* offers 384 i/os, 36,864b of user ram and 2304 flip-flops in the core logic cells. AT&T Microelectronics. Tel., 01734 324299; fax, 01734 328148.

Power semiconductors

600V triac. Isocom's *IS6000* series of optically coupled triacs features a 600V withstand voltage to enable control of rms voltages up to 240V with a safety factor. Triggering in the range is 3-30mA and forward current 60mA dc or 100mA rms, with a peak of 1.5A. Total power dissipation is 350mW. Isocom Components Ltd. Tel., 01429 863609; fax, 01429 863581.

PASSIVE

Passive components

Electrolytics. For general-purpose and high-cv use, the *Nichicon SA and SR* electrolytic capacitors measure 7mm long and 4–8mm in diameter. Values cover the 0.1–220µF range at ±20% tolerance and at voltages from 6.3V to 50V. Permissible ripple is 240mA or 130mA rms, depending on value and voltage. Nichicon (Europe) Ltd. Tel., 01276 685393; fax, 01276 686531.

Solid capacitors. Sanyo's *OS-CON* range of solid capacitors use an organic semiconductor electrolyte, which enables the capacitors to replace aluminium electrolytics 20 times their value at frequencies over 1MHz. Frequency range is 10kHz-10MHz, and there is negligible impedance change within the operating temperature range of –55°C to 105°C. ESR is around a tenth of that of tantalum types and the units withstand reverse voltage of 20% of rated voltage. Jayex Components Ltd. Tel., 01734 810799; fax, 01734 810844.

Dual capacitors. Due to a reported resurgence of valve amplifiers, LCR has discerned a need for power-supply filter capacitors in the form of two electrolytics in the same can. Those now available are rated at 450V dc and 500V dc in dual values of 8, 16, 32 and 16+32µF in 450V types and 50µF in the 500V version. LCR offers a short-form catalogue. LCR Components. Tel., 01495 723131; fax, 01495 722321.

Connectors and cabling

IEC PCB mains inlets. *Multifit* board-mounted mains inlets by Schurter are in UL94 V-0 thermoplastic and take both 7mm and 9.05mm pin spacings. They can be snapped into place with no further fixing required, or will take self-tapping screws for added security. The inlets have a full range of approvals and protective covers are available. Radiatron Components Ltd. Tel., 01784 439393; fax, 01784 477333.

Waterproof connectors. Miniature connectors in Tajimi's *RO4* series are in the screw-latch, circular form and are waterproof to a depth of 20m. They are available in a variety of plugs, jack sockets and bulkhead receptacles in diameters from 12mm to 20mm with up to 12 gold-plated, copper-alloy contacts. Quiller Switches Ltd. Tel., 01202 417744; fax, 01202 421255.

PCB interconnectors. Connectors and headers on a 2mm pitch and 4mm height off-board by Harwin are designed for space saving, being around 35% shorter than the 2.54mm types. The series includes single and dual row types from 2 to 16 ways, with crimp terminals, a tool being available. Terminals are in tin-plated phosphor bronze, rated at 650V rms and 2A at 50°C. Contact resistance is 30mΩ and minimum insulation resistance 100MΩ. RS Components Ltd. Tel., 01536 201234; fax, 01536 405678.

Side-entry mains plug. Designed for side entry, the *Bulgin PX0686/SE* mains plug is rewireable and has a rigid body shell in UL94V-*O* flammability-rated nylon incorporating a screw-down cable clamp and strain-relief sleeve/grommet; connection is to screw terminals. It is rated at 10A and complies with IEC-320-2-2; approvals are pending for UL, CSA, BEAB, SEV and SEMKO. Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095.

Displays

Graphic LCD..Resolution of 240 by 128 and fluorescent back-lighting are features of Datavision's *DGF-24128* graphic lcd module. A built-in *T6963C* control circuit, a rom character generator and 8K ram simplify interfacing. Viewing area is 132mm by 76mm. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

1/4VGA LCD. Hitachi's new colour lcd module uses stn techniques to allow makers of small instruments the choice of a colour display. Resolution is 320 by 240, or the equivalent of 1/4 VGA, and the modules provide eight or more colours, with a backlight. Dimensions are 160-by-120-by-8.5mm and mounting holes are compatible with the existing *LM69xx* monochrome displays. Hitachi Europe Ltd. Tel., 01628 585000; fax, 01628 585200.

AC plasma display. With a typical brightness of 180cd/m^2 and a 20:1 contrast ratio, the Jayex Components 10in ac gas-discharge plasma display provides a 640 by 400 matrix on a 0.33mm pitch. Viewing angle is 160° and service life over 50,000h. It is designed for use in touch-sensitive screen overlay systems. Jayex Components Ltd. Tel., 01734 810799; fax, 01734 810844.

Filters

Very small EMI filters Designed for heavily populated pcbs, TDK's *ACF* series of very thin surface-mounted emi chip filters measure 3.2 by 1.8 by 2.5mm (*ACF321825*) and 4.5 by 1.8 by 3.2mm (*ACF451832*). Each is a T-filter with microbead inductors and a multilayer chip capacitor, the external structure being of ferrite. Frequency ranges are 17-650MHz and 12-600MHz, insulation resistance 1GΩ minimum and dc resistance 0.15Ω maximum. TDK UK Ltd. Tel., 01737 772323; fax, 01737 773810.

Instrumentation

Accelerometer calibration. An *Automated Accelerometer Calibration System* from Endevco provides pneumatic shock exciters from 10g to 100,000g with a range of modules, vibration from 1Hz to 50kHz, high-speed data acquisition and an automatic database. The instrument is pc-compatible for analysis. Consisting of controller, shock and vibration modules, the system controls up to four excitation sources. Features include auto display of results, manual or computer-controlled modes and frequency-domain analysis. Endevco UK Ltd. Tel., 01763 261311; fax, 01763 261120.

Indicators/controllers. Indicators from Control Transducers also act as controllers, units in each of the three ranges containing all electronics relevant to the type of transducer in use: pressure, force and weight and displacement. There is a hardware version at low cost and a micro-based type to work as a complete control unit. Features include open-collector/relay output, min/max detection, dual scale, linearisation and autozero. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

Power meter. In one 144mm by 144mm panel mounting, the Seltek *UPM3001* power meter measures V, A, power factor, kVA, kVArh, Hz, kWh, max and min values, combining these with harmonic analysis up to the 25th for voltage and current. Current-transformer and voltage inputs are provided. Features include a menu presentation on the lcd, and a memory for half-hourly storage of energy profiles and harmonic samples for downloading to a pc. Seltek Instruments Ltd. Tel., 01920 871094; fax, 01920 871853.

Literature

Rack slides. Widney has a new product selector describing a comprehensive range of cabinet and

Hardware

Cooling system. Meech Exair *3200* compressed-air vortex tubes provide cooling rates to dies, solders and adhesives and in electrical enclosures of up to 10,200Btu/h. Cooling is achieved by heat exchange between the 80lb/in^2 compressed air spinning down the tube at up to 10^{6}rev/min and air returning at slow speed through the high-speed airstream. No moving parts are involved. Meech Exair Ltd. Tel., 01993 706700; fax, 01993 776977.

rack telescopic slides, which take loads of up to 3000kg. Widney Enclosures Ltd. Tel., 0121 327 5500; fax, 0121 328 2466.

Printers and controllers

Thermal line printers. Three models of *FTP600* 24V thermal line printer from Jayex produce high-density, high-resolution output at 480, 640 and 832 dots/line in 56mm, 80mm and 104mm printing width and are suitable for printing bar codes. The units are available as mechanisms only, mechanisms with interface board or as a set with microcontroller unit and gate array for complete driver control. Jayex Components Ltd. Tel., 01734 810799; fax, 01734 810844.

Production equipment

Solder-paste printer. For accuracy to within 0.05mm, Panasonic's *Panasert SPP-G1* digitally controlled solder-paste printing machine has a built-in visual recognition system for precision and to allow rapid screen exchange. Both board and screen pattern are rotatable through 45° and 90° to reduce printing variation and continuous high-quality print is achieved by automatically cleaning the underside of the screen. A board positioner automatically adjusts for different board thicknesses. Panasonic Industrial (Europe) Ltd.

Drop-out simulator. Schaffner EMC offers the *NSG1003* simulator, which will duplicate all the disturbances that are liable to occur in a mains supply, carrying out all the tests in EN61000-4-11, IEC-4-11 and others. It provides voltage drops down to 200µs, drops cycles or half cycles, varies the voltage and simulates over and under voltage. The instrument can be linked via RS-232 to a pc controller running *WIN1003*, the Windows-based instrument control package. Schaffner EMC Ltd. Tel., 01734 770070; fax, 01734 792969.

Fastest dsos. Far exceeding other digital storage oscilloscopes in its capture rate, Tektronix' *TDS700A TruCapture* range of instruments handles 400,000 acquisitions per second – better than the current fastest analogue oscilloscopes – using the *InstaVu* Acquisition technique. Features included in these instruments are far too numerous to mention here, but there is a seven-inch colour lcd with a 0.01in dot size, a graphical user interface, data storage on a 3.5in floppy and hard copy, waveform maths on board and everything to give an unprecedented level of waveform measurement, data handling and ease of operation. *TDS784A* has a 1GHz bandwidth and 4Gsample/s and *TDS744A* 500MHz and 2Gsample/s. Tektronix UK Ltd. Tel., 01628 486000; fax, 01628 474799.

Tel., 01344 853277; fax, 01344 853803.

Enamel-wire stripper. Series 4's *EC2 Enamel Wire Stripper* takes the insulation of enamelled copper wire in gauges from 29awg to 8awg with no damage to the wire. A rotary steel or carbide insert is adjustable to match the wire diameter and is locked in position by a chuck. Special inserts allow stripping close to a coil bobbin or other components. Series 4 Ltd. Tel., 01703 866377; fax, 01703 866323.

Hot-melt masking. *ITW Dynatec's 022SP* hot-melt masking system accurately applies a quick-setting Seal and Peel mask material that cures in minutes and is simply peeled off after use. The system incorporates a hot-melt Melt-on-Demand tank process which eliminates the waste inherent in the use of cold fluid. The material can be used in a variety of thicknesses on conformal coatings including paralene, silicone, urethane, acrylic and epoxy. Bayotech Ltd. Tel., 01933 411125; fax, 01933 411126.

EMI/RF gaskets. *Shieldseal IGS* is a James Walker product that forms an emi/rf/environmental seal directly onto a component such as an enclosure or onto a backing sheet for transfer to components. A pressurised dispensing head moves under computer control in *X*, *Y* and *Z* axes to follow complicated paths, forming seals in *Shieldseal* room-temperature-vulcanising, two-part elastomer and filled with silver-plated ceramic microspheres or carbon graphite. Gasket thickness can be varied throughout the sequence and beads down to 0.4mm are possible on materials including metal, chromate or plastics with painted, plated or vacuum-deposited metallic finish. James Walker & Co, Ltd. Tel., 01483 757575; fax, 01483 755711.

Power supplies

Power switch. Siliconix' *Si9718CY* reverse blocking switch is meant for use in battery-disconnection application in dual-battery notebook computers, replacing two mosfets and their drive circuits. It allows the computer to switch from one battery pack to the other before cells become completely discharged. Micromark Electronics Ltd. Tel., 01628 76176; fax, 01628 783799.

Surface-mount dc-dc converter. *SM series* surface-mounted, high-density converters by Abbott Electronics provide up to 280W at power densities of up to 50W/cubic inch over the –55°C to 100°C range. Input voltages from 18V to 36V and 150V to 400V are acceptable for the 28V and 270V versions, providing outputs of 3-28V DC at currents of 40-10A. Abbott Electronics Ltd. Tel., 01233 623404; fax, 01233 641777.

Internal ups. Network Connectors has a range of *Mini UPSs* designed to fit a 5.25in bay in a pc. After a fault on the supply line, the device sounds an alarm, whereupon the ups keeps the pc running for around six minutes to allow an orderly shutdown. Both 300VA and 400VA versions are TUV-approved, transferring at 4–8ms, and incorporate surge suppressors, line filters, short and overheat protection and provide a 10h battery charger. The units operate with *Autoguard* dos data protection or Novell data protection software. Network Connectors Ltd. Tel., 01275 848421; fax, 01275 844234.

Radio communications products

LNA/mixer. The *RF2418* uhf receiver front-end ic from Anglia is a low-noise, 700MHz–1GHz amplifier/mixer for uhf digital and analogue receivers and digital and spread-spectrum communications, the only extras needed being local oscillator and filtering. It contains the lna, a second rf amplifier, dual-gate GaAs fet mixer and IF output buffer driving 50Ω. Anglia Microwaves Ltd. Tel., 01277 630000; fax, 01277 631111.

Input-voltage switches. *STR83145* and *STR84145* from Allegro are latched, universal input-voltage switches which sense the applied ac line voltage, switching the rectifier and capacitors from a voltage-doubler configuration for inputs under 141V and a full bridge for inputs over 149V. The devices contain timing, control and drive circuitry and a high-current triac. User errors when misconnecting jumpers or switching to the incorrect voltage range are eliminated, and the switched-mode power stage need only cope with a smaller range of inputs. During voltage variations, a latch holds the full-bridge mode to avoid the doubler taking over. Switch-over voltage is set during manufacture and an adjustable delay ensures start-up in the full-bridge mode. The two devices offer on-state currents of 10A (*83145*) and 12A. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.

Quadrature modulator. Anglia offers the *RF2422* low-cost quadrature modulator ic, which provides universal direct modulation for am, fm, pm or compound carriers in the 0.9GHz-2.5GHz range. It contains a 90° carrier phase-shift network, carrier limiting amplifiers, two matched doubly balanced mixers, summing amplifiers and an output amplifier driving 50Ω. Supply is a single 5V rail. Anglia Microwaves Ltd. Tel., 01277 630000; fax, 01277 631111.

Switches and relays

Double-tier switches. Arranged in a double tier, *JTP2230* switches by Jeil Products have switch buttons on 12.5mm centres with 3.5mm diameters, the whole standing 22.6mm above the pcb and measuring 8.4mm wide. At 12V, 50mA, the switches have a life of 100,000 operations. Operating force can be 100, 160 or 260gf. Seven models are available with different stem lengths. Pedoka Ltd. Tel., 01493 440047; fax, 01493 440048.

Versatile relays. Schrack *RT* relays are only 15.7mm in height, but offer wide versatility and high capacity. Coils may be monostable or bistable in ac or dc working, consuming 400mW, and have either one or two cadmium-free changeover contacts rated at 16A. The relays are immersion-proof to IEC 68/2.17, are isolated to 5kV/8mm and meet VDE protection category 2. Pinning is standard and the units are mounted on the pcb or in DIN rail sockets, designed for the *RT* series. Schrack Ltd. Tel., 01868 1211; fax, 01868 2221.

Transducers and sensors

Zero hold-power solenoid. The *PED Series 66* solenoid by BLP Components uses magnetic latching to provide continuous hold without power. It is pulse-operated and coils come in 5V, 6V, 12V, 24V and 48V types for each four power ratings from 3W to 20W. Magnet holding force is typically 0.48Kgf. BLP Components Ltd. Tel., 01638 665161; fax, 01638 660718.

Heat imaging. *Vero's* new thermochromic boards allow engineers a 'first-pass' view of the temperature distribution on a populated plug-in board. In Eurocard, Hard Metric (Futurebus+) and a universal version, the boards indicate by a colour change temperature differentials down to 1°C; each board has a colour chart for quick reference. A transparent front panel preserves the normal operating airflow while allowing inspection. Vero Electronics Ltd. Tel., 01489 780078; fax, 01489 780978.

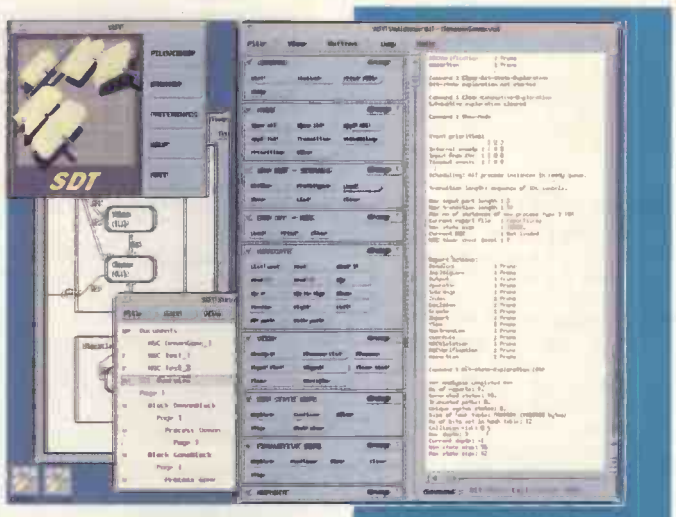
Temperature sensors. *IRA-E401* is a range of infra-red pyroelectric sensor by Murata, provided with optical filters for flame detection and remote sensing. *IRA-E500/600* series sensors detect infra-red emissions from the human body and have filters for various applications. *IMD* devices have in-built signal processing, the *B101* version having analogue and digital output. Murata Electronics (UK) Ltd. Tel., 01252 811666; fax, 01252 811777.

Displacement transducers. Internal electronics in Monitran's new range of transducers provide a 4-20mA output for use in control loops. Measuring ranges available are ±2.5mm to ±600mm, in various styles of device. There is also a pressurised type designed to withstand a pressure gradient of 6000lb/in^2 across its body. Monitran Ltd. Tel., 01494 816569; fax, 01494 812256.

Linear potentiometer. From Data Instruments, the *Short Longfellow* position transducer is available in strokes from 30mm to 156mm in five ranges, with independent linearity better than ±0.1%. Resistance is 1.5kΩ/in travel and the unit is rated at 0.75W/in electrical travel. Its 1in square anodised aluminium housing has a *MystR* plastic resistive element which gives low noise and a life of 10^9 operations. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

DC motors/gearboxes. Bodine has expanded its range of small 24V dc motors and gearboxes. Those now available include permanent-magnet motors and brushless dc types with 15-250W power ratings. There is also a range of four parallel-shaft and three right-angle gearboxes in a wide range of ratios. All are totally enclosed, providing a typical performance of 2500rev/min at a torque of 40Nm. At 12V, the motors give full torque at half speed. Bodine Electric Company. Tel., 01252 811800; fax, 01252 811801.

Ministepper drive. Two ministep motor drives, the Digiplan *CD60M/80M*, are under 50mm wide and are said to provide 50% more power than previous narrow units. Units are in 10HP modules, so that eight can be contained in one 3U 19in rack. The drives step at 400-4000 steps/rev. at high accuracy; motor current at up to 6A/7.8A is programmable down to 60% peak. Both drive and motor are protected against heat and shorts. Parker Hannifin plc, Digiplan Division. Tel., 01202 699000; fax, 01202 600820.



Software development. *Validator* is a new function of *SDT2.3*, a software development tool by the Swedish company TeleLOGIC for designers of embedded microprocessor systems. *SDT* is based on the ITU/CCITT language SDL and has tools for specification and design, verification, validation and maintenance, plus code generation. Validator automatically validates the SDL specification, including any C code included. Reflex Technology Ltd. Tel., 01494 465907; fax, 01494 465418.

Vision systems

Video mixer. Replacing six ttl adders and two 12-bit multipliers, Harris's *HSP48212* 12-bit digital video mixer mixes two sources, providing frame addition, fade in/out, video switching and high-speed multiplying. Its features include 12-bit pixel data, 12-bit mix factor and rounding between 8 and 13 bits. A programmable pipeline delay of up to seven click cycles helps to control misaligned data. Micromark Electronics Ltd. Tel., 01628 76176; fax, 01628 783799.

COMPUTER

Computer systems

Single-chip PX/XT. The *LH72501* from Sharp provides most of the functions of a PC/XT on one chip, using the NEC *V20HL* processor. Included is the serial/parallel i/o, a three-channel counter/timer, an interrupt controller, dma controller, clock, bus controller and power management for 3V or 5V operation. There are also keyboard, printer and PCMCIA card interfaces and memory controllers for dram, sram and pseudo-sram. At 5V, the processor runs at 16MHz. Sharp Electronics (Europe) GmbH. Tel., +49 40/23 76-0; fax, +49 40/23 76-25 10.

Tough, portable PC. Containing six ISA slots and a 9in colour lcd screen in a 15in by 16in by 4in case, the *Fieldworks FW7500* portable pc withstands 100g shock and vibration to Mil Spec 810C. It uses an IBM 75MHz *Blue Lightning* 486 card with an optional co-processor, 4Mb of ram, expandable to 64Mb, and a 130Mb hard disk drive, expandable to 550Mb. Power supply is suitable for use world-wide, being self-setting, and internal and external battery packs are available. Workstation Source Ltd. Tel., 01734 759292; fax, 01734 757522.

Data acquisition

PC data. Amplicon's *DAP 800/1/2/3* boards, each of which is a complete data acquisition system on a pc card, is a 'very low cost' solution, which means it costs less than £1000. Four models, *DAP 800/1/2/3* and *DAP 801/1*, comprise the range: 100/1 has a 10MHz cpu, samples at 75ksample/s and has a 256K ram; 800/2, 16MHz, 105ksample/s and 256K; *803* similar to *800/2*, but with 1024K ram; and *801/1* 10MHz, 75ksample/s and 256K. All have eight 12-bit analogue inputs, two 12-bit analogue outputs, 8-bit digital i/o ports, a programmable-gain amplifier, internal and external clock and hardware and software triggering. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 01273 570215.

Mini modem. Datastripe's *OEModem* is a BABT-approved miniature internal modem, measuring 135 by 43 by 11mm and contained in an insulated case. It may be connected by flying lead or plugged into a mother board in the host pc. Two versions exist: *DS022* supports V21, V22, V22bis; and *DS023* additionally supports V23 (1200/75 full duplex and 1200/1200 half duplex), V21 fax, V27ter and V29 fax. Datastripe Ltd. Tel., 01276 682921; fax, 01276 683016.

Data communications

Dual uart. From California Micro Devices, the *CM16C552* contains two '*550A* uarts with integrated fifos and serves two serial i/o ports simultaneously, also having a bi-directional parallel printer port. It is compatible with other '*552*s and controls transmit, receive, line status and data set interrupts, and includes an internal programmable baud-rate generator, with fax/modem capability. Sabre Advanced Microelectronics Ltd. Tel., 01420 22004; fax, 01420 22008.

Multiple i/o interface. By means of the *IMS PCL-746* 115kb/s multi-interface, four-port serial comms card, each port is configurable to RS-232, RS-422 or RS-485. The card has four 16C550 uarts with an on-chip fifo buffer to reduce the load on the cpu and speed things up, particularly in Windows. All four ports can be set as normal pc serial COM1–COM4 ports or set to share the same interrupt. It is compatible with Arnet four-port cards supporting SCO UNIX/XENIX. Programming software is included. Integrated Measurement Systems Ltd. Tel., 01703 771143; fax, 01703 704301.

Mass storage systems

PCMCIA flash storage. Claimed to be the world's smallest removable mass storage system, Sundisk's *CompactFlash* uses 32-bit flash techniques and a single-chip ATA controller. It weighs 0.5oz and is around the size of a book of matches, providing full PCMCIA-ATA functions. Capacities of 2, 4, 10 and 15Mbyte are available – double with Stacker. The unit fits a standard PCMCIA Type II adapter card. Sundisk Corporation. Tel., +1 408 562 0500; fax, +1 408 562 0503.

1.8in PCMCIA disk drive. Maxtor offers the latest in the *MobileMax* series of hard disk drives: the first 1.8in unit to fit 5mm Type II PCMCIA slots. Capacities are 85 and 121Mb and the units operate from 3.3V with a shock specification of 1000G (non-operating). Spin-up time is less than 1.5s and MTBF 300,000h. Ambar Systems Ltd. Tel., 01296 435511; fax, 01296 394183.

Software

PictureBook v.2.2. Digithurst announces an enhanced version of the multimedia package, *PictureBook*, a Windows 3.1-compatible package meant for the creation of electronic books. New features include better object handling for greater flexibility when making publications containing digital video, and the incorporation of text into pages in a scrollable form to allow the use of large virtual pages. *PictureBook* will now integrate information from services such as Internet into electronic publications, rather than simply teletext, as in the earlier version. Taking information from broadcast and network sources enables the package to be used as front-end for electronic newspapers or on-line publications. Version 2.2 can be run with a range of video capture and scanning devices without the use of a *MicroEye* card for moving image display. Digithurst Ltd. Tel., 01763 242955; fax, 01763 246313.

PC letter writer. Apparently, engineers are not the only ones accused of being virtually illiterate. CCA Software has introduced *Effective Letters*, which is a software package designed to put a veneer of sophistication on letters to grieving bank managers and plaintive customers, since electronics companies are among those whose standard of writing is considered by the Adult Literacy Unit to be the worst. It contains more than 400 letters, in nine categories, on subjects such as how to fire someone, how to avert the wrath of creditors and many others of a more mundane nature. CCA Software. Tel., 0161 480 9811; fax, 0161 429 6028.

DSP software development. NEC has a new suite of software development tools by Atair for its *μPD7701x* family of fixed-point digital signal processors, which runs on a PC under Windows 3.1. There is *Workbench*, a 'housekeeping' utility, *Debugger* for in-circuit emulation and *Simulator*, which uses a gui for function verification without a need for hardware. NEC Electronics (UK) Ltd. Tel., 01908 691133; fax, 01908 670290.

Control design/development. Integrated Systems's *Matrixx* family of visual design and development tools for control systems designers will now run under Windows NT on a pc, providing design, test, graphical simulation and modification aid throughout the whole design cycle and producing code and documentation automatically. The family includes *SystemBuild* modelling and simulation, *Xmath* object-oriented maths analysis and visualisation, *AutoCode* code generator, *DocumentIt*; and *RealSim* prototyping computers. Integrated Systems Inc. Ltd. Tel., 01707 331199; fax, 01707 391108. ■

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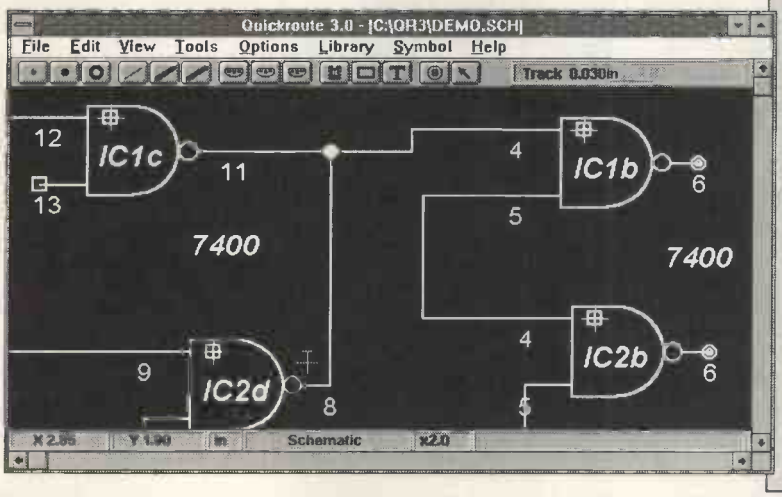

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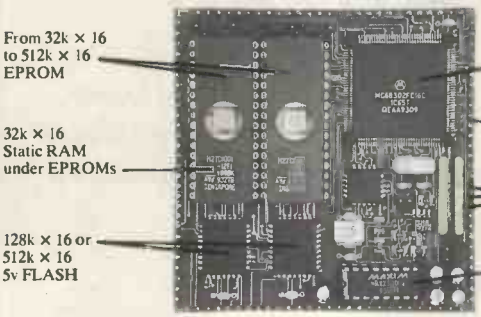

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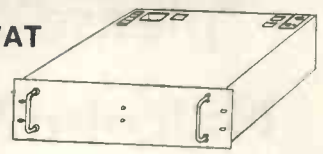

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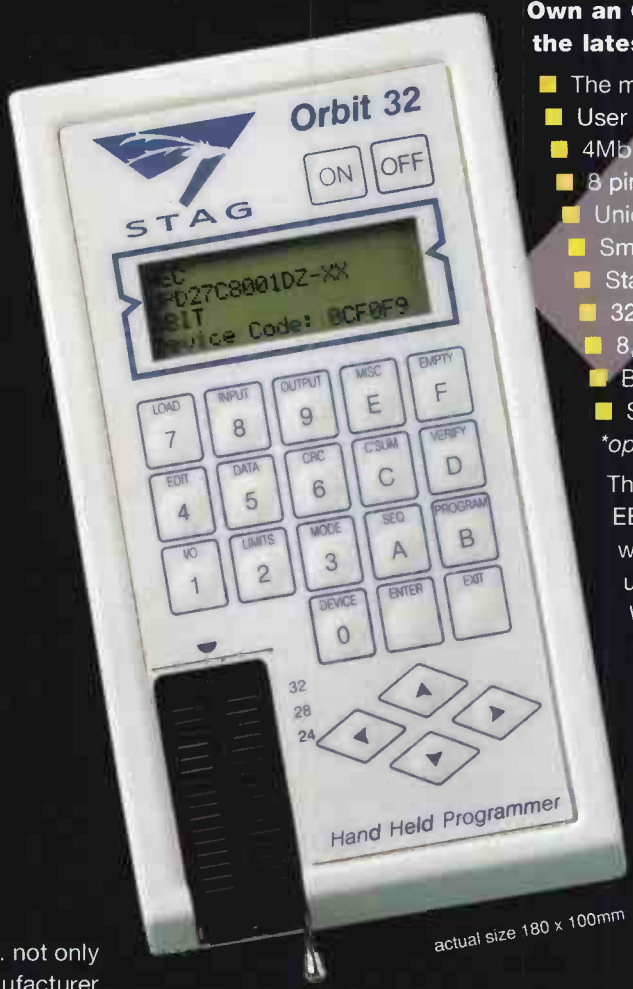

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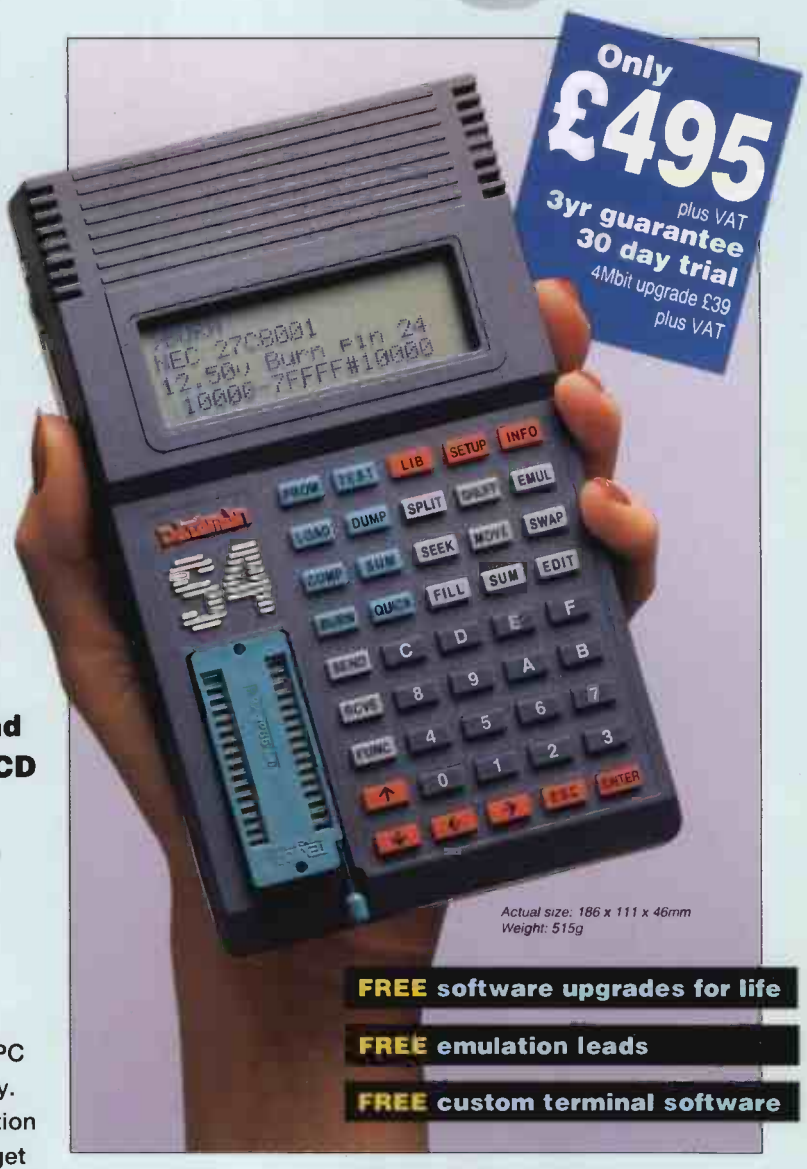

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