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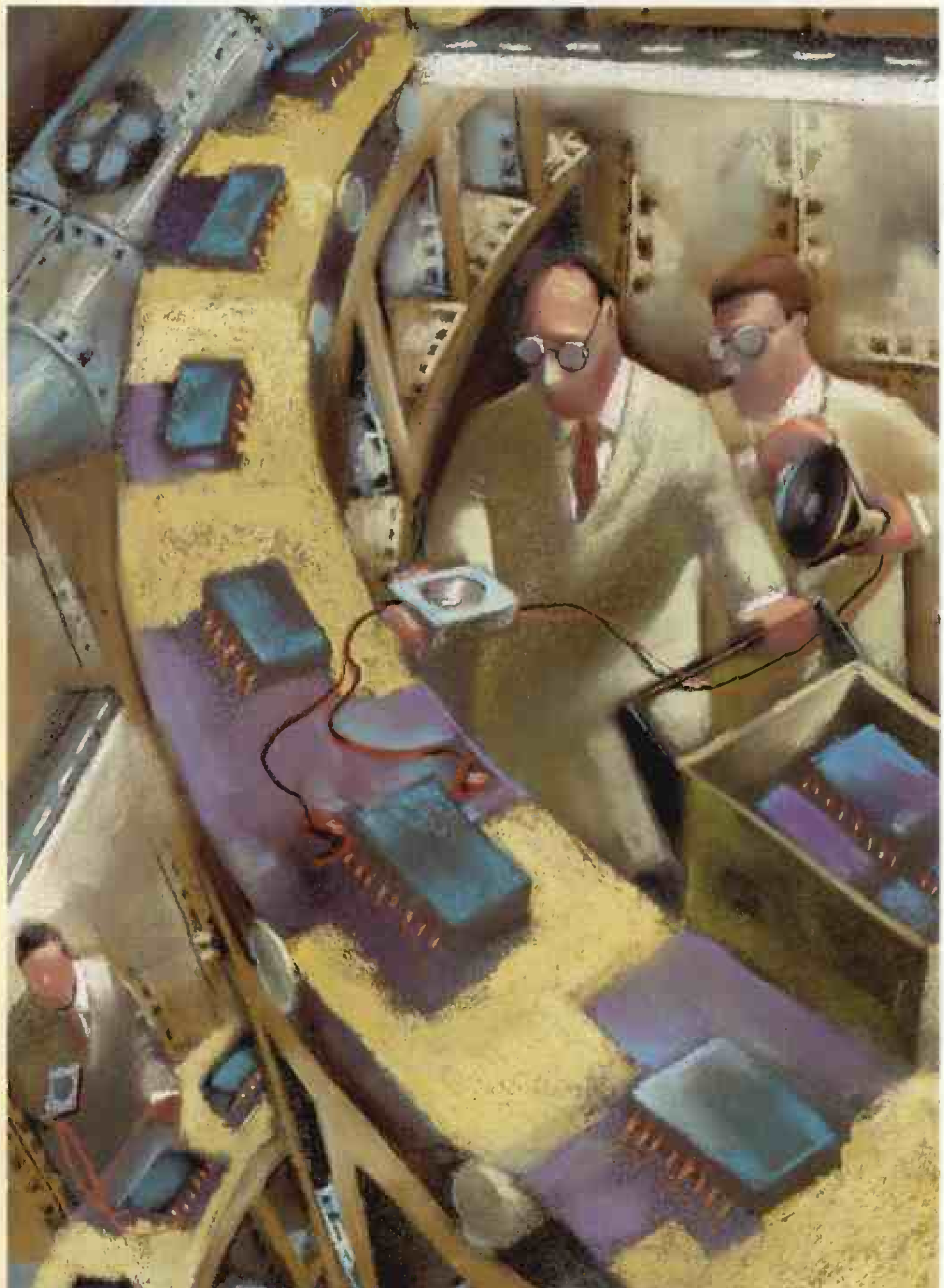

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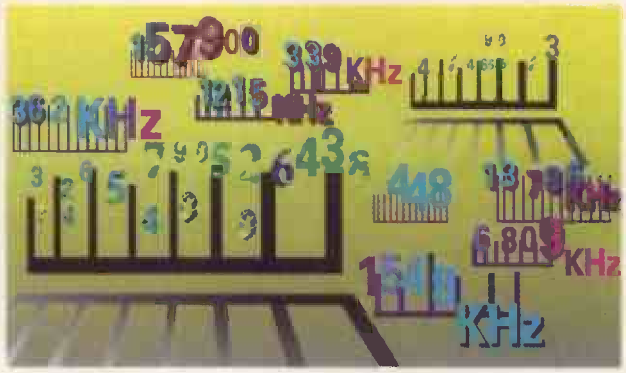

REGULARS

In next month's issue: Digital television. In 1991 a MAC future for TV broadcasting seemed certain. Two years on it is now digital television which will bring 100's of channels into the living room of the next century. We present both the systems and the silicon. Also next month, the start a new series on using programmable logic.

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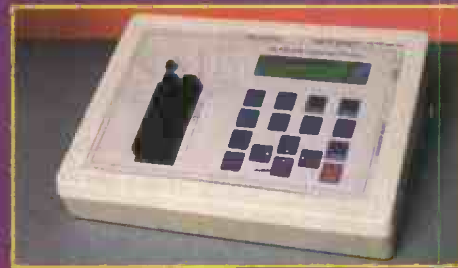

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A communications revolution

The world is about to change irrevocably. At face value, the UK now has, or will have, an extra 14 satellite TV channels controlled by News International and its associates. Whatever one may say about the quality of programming, this new group will broaden viewer choice. What we are really seeing is the start of a revolution.

For those of us who have grown up with three or four channels of terrestrial television, it is hard to conceive how we could comfortably deal with 20 or 30 channels. New digital compression technology promises to increase this viewer choice by an order of magnitude within ten years.

With hundreds of channels to choose from, individual programming will become highly specialised. The most appropriate analogy seems to be with the world of publishing. If one considers the four existing terrestrial channels as daily newspapers, then the new satellite services will become specialist magazines. Not inferior substitutes for something which already exists but a new communications medium for a specialist audience. If you want to find your particular leisure activity – for instance sailing and windsurfing – on the box whenever you want it, then the new TV explosion will almost certainly offer it to you.

But the communications revolution means more than that. Satellite video channels can transmit a large amount of data very quickly to the viewer's or user's home. When local high density storage (video recorder?) and some computing power is available at the receiving end, then many things become possible. Entertainment programmes, electronic newspapers and magazines may be downloaded off-peak for later viewing or reading – a sort of multimedia interactive video service. And since this magazine, your daily newspaper and most of the other things you read and watch are already originated electronically, this is only a very small step from the world today. The existing telephone network will allow the viewer and reader to place their specific programme requirements with the local TV control centre.

The concept of panoramic choice is difficult to grasp; the concept of global broadcasting seems equally so. We conventionally think that our programme material should originate, in the main, from our own country. Satellite broadcasting represents true internationalism by its nature and foreign doesn't always mean inferior. Ted Turner's *CNN News Service* is highly regarded the world over, mostly exceeding anything which our own national networks can provide in the field of global current affairs. High quality international leisure and education services will also become available outpacing our own national networks in both scope and resource. No matter what we may think about such an idea at present, we will eventually accept and use them.

Global communications have so far acted as an humanising influence; brutal regimes can no longer hide their activities from the world's television screens. But communication without borders carries significant risk: those who would control global networks wield immense power unfettered by individual nations or electorates. This makes me uneasy. One could conceive of a new kind of tyranny where communications moguls might practice blatant propaganda or, even worse, subtle persuasion against governments and people. Broadcast technologists are now able to deliver this power to their unanswerable masters.

We must ensure that we address the moral, ethical and democratic issues of the communications revolution.

Frank Ogden.

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Time runs out for digital tv breakthrough

There is hardly any time left to stop tv broadcasters, electronics manufacturers and viewers in the next century looking back on 1993 as the year in which the ITC lost the UK a unique opportunity to lead Europe into a digital future.

History has left Britain in a unique position. It has a pair of frequencies in the middle of the UHF tv broadcasting band that are not used for broadcasting and can be used to kick start Britain into the digital tv age.

Unfortunately the already ageing Broadcasting Act 1990 obliges the Independent Television Commission to try to earn money for the treasury by selling off these frequencies for a fifth tv service, to be called Channel 5. This will use the same analogue technology as is used for today's four tv channels. It will provide one new programme channel for around 70% of the population. Analogue Ch 5 will also cause interference to many millions of VCRs, satellite receivers and video games that use the frequencies to connect with a tv set.

Research work recently carried out by the BBC suggests that if the same two frequencies were used with digital technology, they could provide 97% of the population with eight new channels and less risk of interference.

If granted, the licence for analogue Ch 5 will run for 10 years – though it cannot be granted until 1994 at the earliest so the analogue route blocks the digital option until well into the next century. It is hard to find anyone, except those who hope to make quick money out of it, who favours analogue Ch 5. In the long term the digital option will generate far more revenue for the treasury. But it has yet to see the wisdom of waiting a couple of years for the technology to be ready for consumer use.

In late June the ITC published a consultation document on the broad issue of digital television. Although heavy reading, the document gives the clear message that digital tv technology makes far more efficient use of available frequency spectrum.

When the ITC published the document, Peter Rogers, the ITC's deputy chief executive, refused to discuss the issue of allocating the Ch 5 frequencies for digital tv. He said that the ITC had separated the issues of digital tv and Ch 5, and would be publishing a second document, specifically dealing with Ch 5, in a few weeks time.

The ITC published its Ch 5 document in mid-July and by the ITC's own admission the content was deliberately neutral. It simply listed holding the Ch 5 frequencies free for digital tv as one of three options. The ITC says it believes it has discharged its duties.

A lost digital multi-channel opportunity will cause few tears to be shed by the Ch 3 and Ch 4 stations, who are already losing advertising to the satellite stations.

The ITC has invited anyone interested in Ch 5 to comment on the two documents* published so far, by October 15. But without wider explanation of the issues, it is a safe prediction that many of those who should be interested in commenting will not realise it until they are years too late.The Broadcasting Act may bar the ITC from taking sides, and blocking analogue Ch 5. But it does not bar it from ensuring that the mass media understand the issues and inform the public of their right to comment. If the ITC does grant a licence for analogue Ch 5 then it must stand accused by future generations of failing to ensure that the public had the best possible chance to understand what the UK stood to lose.

Barry Fox

**Copies of both documents are still available from the ITC at 33 Foley Street, London W1P 7LB.*

The joint European x-ray telescope, called Jet-X, has completed its vibration and decompression tests at British Aerospace Space Systems in Stevenage. The picture shows the 551kg telescope being prepared for the decompression tests. These were carried out in a 3.5m diameter by 6.4m autoclave evacuated from 2 to 1bar in 80s. Video film recorded no distortion of the structure or movement of the thermal insulation. The satellite is due for launch in 1995 by the Russian Proton rocket. Its four-day orbit will be highly elliptical with a perigee of 1000km and an apogee of 200,000km.

BT moves on System X replacement

BT has named AT&T and Alcatel, the world's largest telephone equipment manufacturers, as its first suppliers for the digital telephone exchange technology that will replace System X exchanges at the turn of the century.

The decision not to support any development of the broadband switch technology known as ATM (asynchronous transfer mode) in Britain by ignoring a bid from GPT and Siemens is likely to come as a major blow to hundreds of engineers at GPT.

GPT is already struggling to adapt its switching business in anticipation of a run down of System X exchange orders from BT now that its UK network is largely complete. In 1991 GPT, the UK's largest telecommunications manufacturer, shed more than 1000 workers from its Liverpool switching site.

The disappointment will be increased because it was thought that the company's technical partnership with its German parent Siemens had secured its future by developing new technologies like ATM and the SDH transmission system.

Earlier this year GPT/Siemens was named as one of BT's two SDH suppliers and also won orders for broadband computer network technology known as MANs. But the failure to be named as a main ATM supplier is a blow to GPT, which supplied virtually all of BT's existing digital telephone exchanges. Siemens has already secured ATM orders in Germany and the US.

The AT&T ATM switches will be used in BT's national telephone network while Alcatel's will support BT's participation in the European ATM trial. BT will also introduce ATM switches into the SuperJanet academic network in 1994.

Richard Wilson, *Electronics Weekly*.

Pirates ride the waves

The annual report of the Radiocommunications Agency, since 1990 an Executive Agency of the DTI, tells that radio piracy is on the increase again. In the year 1990/91, the RA prosecuted 145 people for radio piracy. The Broadcasting Act 1990 then came into force and gave the RA new powers to target people who finance pirate stations and advertise on them, as well as those who broadcast. This deterred some operators and the RA prosecuted only 67 in 1991/92. But now they are broadcasting again. In the year 1992/93 the RA made 536 raids, and secured 68 convictions.

Look out for broken glass while bonding: This glass sheathed microwire (top left) was originally developed for a military application and consists of a fine core filament of copper, gold, or silver of 3 to 10μm diameter encased in a glass sheath with an outside diameter of 15 to 30μm. A short high energy burst of ultrasonics breaks the glass and forms the bond with a substrate allowing manufacture of microsensors in low conductivity fluids. The bared conductor may be used as a microprobe, for instance in measuring membrane potentials. Chemring 0705-735457.

Video scramble for pot of gold

When Rupert Murdoch talked recently of his "global vision" and plans to develop a common standard for digital tv transmission and the delivery of video pictures by telephone line, he is really talking about the pot of gold to be earned from scrambling pictures.

The electronics industry has already developed the digital compression technology that makes transmission and delivery possible, and in a remarkable demonstration of world wide cooperation the industry has almost finished five years work on setting a common standard. But nobody has yet been able to agree on the encryption system to be used to ensure that only those viewers who pay to watch are able to watch.

News Datacom, a small security company owned by Rupert Murdoch's News International, has already developed an encryption system. If Murdoch can get the world to adopt this, NI will earn vast royalties through the next century.

When Rupert Murdoch began broadcasting his Sky service from Luxembourg's Astra satellite in February 1989, he chose the existing analogue Pal tv system, and broadcast clear – without scrambling. A year later Sky started to scramble some transmissions, to earn money from subscriptions.

Broadcasters had previously judged it impossible to scramble analogue Pal, with sufficient security to prevent piracy and without degrading the picture when unscrambled.

Thomson Consumer Electronics of France updated an old idea from Westinghouse in the US, which involved chopping up the lines of the picture before transmission and putting them together again in a receiver decoder. Professor Adi Shamir of the Weitzmann Institute of Science in Israel had already been working on ciphers for NI's News Datacom. Using smart card technology, TCE and NDC jointly developed Videocrypt for Sky. The system uses digital encryption to control the analogue line chopping.

Murdoch has snubbed TCE and signed a deal between what he calls the "best brains". These are NDC, NTL (formerly the Independent Broadcasting Authority's research laboratory), and US telecommunications company Comstream. A separate deal ties in British Telecom and its cellular phone network Cellnet.

The deal is nicely timed. The moving picture experts group of the ISO and IEC recently sealed a first standard, MPEG-1, for handling digital tv pictures at low data rate (1.5Mbits/s) and the telecomms industry is already looking at ways of sending MPEG-1 data streams down telephone lines. The group is on the point of finalising its second standard, MPEG-2, for broadcasting digital tv at higher data rates (usually 4 or 8Mbit/s) from terrestrial or satellite transmitters.

Although Murdoch has dropped TCE, News Datacom had worked with TCE on a digital version of Videocrypt for use with DirecTV, the new digital satellite tv service planned by aerospace company Hughes in the US. Because the previous deals between NDC and TCE were specific to projects, Murdoch is free to offer NDC's encryption as the common standard for use with both MPEG standards. The potential rewards for NDC are huge.

Any manufacturer wanting to make a Videocrypt decoder for use with Sky first gives TCE an up-front payment of £200,000 to cover royalties at £4 each on the first 50,000 decoders made. After that the royalty remains £4, of which half goes to NDC and half to TCE. Firms which cannot afford to manufacture in bulk, pay £35 for a ready-made circuit board. Until recently they paid £40. This includes the £4 royalty split between NDC and TCE.

There are 3 million Videocrypt decoders in use round the world, meaning that NDC has already earned £6m in royalties. If Murdoch can make NDC's encryption a world standard for the next century, the potential rewards are enormous.

Although patents on scrambling technology, such as chopping picture lines, have limited life, the encryption algorithms developed by NDC are covered by copyright, like computer software. NDC can thus protect it rights and draw royalties for decades into the next century.

NTL is developing multi-channel video compression and multiplexing technology for Murdoch. The firm will be looking at new methods of carrier modulation as well as a method of statistical multiplexing that will make more efficient use of the satellite transponder bandwidth.

BF

Resources pooled to make solid state discs

America's third largest disc-drive maker Quantum is teaming up with Silicon Storage Technology (SST), a Silicon Valley flash chip designer, to produce solid state memory discs.

The companies have signed an agreement where Quantum will gain exclusive rights to use SST's flash chips in PCMCIA memory cards with IDE and AT interfaces. It is projected that 12 million portable computers will have PCMCIA compatible slots by 1996.

James Prasad, a product manager at Quantum, said there are several reasons why Quantum has chosen SST's technology.

First, it only requires a single power supply of 5V for programming and reading. This contrasts with the chips made by Intel, the world's largest flash supplier, which need two power supplies. This, says Prasad, means that Quantum flash devices will dissipate little power.

Secondly, the sector erase size on the SST chip is fully compatible with the AT bus used in most PCs.

Finally, the SST flash chips have a much thicker oxide layer than Intel's chips (40nm as opposed to 10nm), which makes them easier to make and results in a smaller die size. The flash drives should be available in the second half of next year.

TFT cell scores 3V goal

Mitsubishi has used its thin film transistor (TFT) static ram cell, first developed for a 4Mbit part, to build a 1Mbit device.

The result is a chip that operates from a 3V supply while consuming just 0.05μA on standby. Power consumption rises to 16mA for the chip when active.

The device is organised as 128K x 8bit and samples will be available later this year. A fourth polysilicon layer forms the TFT yielding soft error counts of 0.03 on cycle times of 300ns.

New opportunities for environmentally sound electronics: small scale electricity generation projects such as the Caennes Bay wind farm in Wales are providing new opportunities for development in process control engineering. This wind farm uses rented equipment from Livingstone Hire as part of the high voltage safety commissioning procedure.

Flat screen TV leads Matsushita's display drive

Matsushita Electric Industrial says it plans to take 10% of the £140 million a year worldwide display market by the end of the century. And the product it believes will spearhead this push is a flat screen television launched in Japan this month. Priced at £1835 the television has a 14in screen and is less than one third the depth of a comparable CRT. Initially about 1000 units a month will be produced.

The display is based on Matsushita's flat vision technology unveiled in 1985, and uses a similar principle to that of a CRT; a fluorescent surface is illuminated by incident electrons.

A matrix of wires is used as the electron source instead of a single gun. The electrons are focused onto confined sections of the display by electrostatic deflection circuits. Removing the gun and the conventional, but bulky, magnetic deflection circuits have let Matsushita get the depth down to less than 10cm.

The firm claims its approach provides the brightness of an LCD combined with the picture quality of a CRT while being considerably simpler to build than an LCD. The firm plans to build larger, wall mounted versions and to improve picture quality suitable for HDTV systems.

Europe snubs US global plans

Europe's largest telephone operators have shunned US plans to create a global mobile telephone network using low orbit satellites.

STET, which owns Italy's public telephone network, is the only European company to commit any money to Motorola's Iridium project to build the world's first global mobile telephone system, which has raised $800m from backers in the US, Japan, China, and Russia.

Europe's largest operators, Deutsche Bundespost Telekom, BT and France Telecom, are signatories of Inmarsat, the international satellite operator, which plans to launch a rival system by the end of the decade.

RESEARCH NOTES

Blue laser burns itself into storage record books

Scientists have demonstrated a blue laser optical recording system that can write and read data at what they claim is a world record density of 2.5 gigabits per square inch on a removable magneto-optic disk. That's the equivalent of 6.5 gigabytes on a double-sided 5.25-inch optical disk.

Researchers have long known that blue light, because of its shorter wavelength, is potentially capable of writing more data on a given disk area. It is simply that blue light can be focused on to a smaller spot than infra-red: 0.4μm, as compared to 0.8μm, about a quarter of the area. But while the theory is straightforward, the practical problems are immense. What researchers at IBM's Almaden Research Centre in California have shown is that the high storage density predicted for blue light systems can now be achieved under realistic conditions.

For their latest demonstration, the IBM team wrote a pattern of data bits on a rotating glass disk coated with a film of magneto-optic material optimised for blue light. They then read back the pattern at a realistic rate of 2Mbyte/s and were able to distinguish the signal from the background noise as accurately as in current commercial products.

In rewritable magneto-optic recording, a mark is made when a short, high-intensity laser pulse very quickly heats a small spot of the magneto-optic coating in the presence of a magnetic field. Within a few milliseconds after the laser pulse, the material cools, locking in the magnetic orientation, which in turn affects the polarisation of any light that subsequently passes through it.

To read the data, a low intensity laser illuminates the track. A detector and its associated electronics then reads the changes in polarisation and converts them into data pulses. Erasure is achieved by re-heating the area with the high intensity laser while the magnetic field is reversed.

IBM's team generates its blue light by passing the output of an infra-red gallium arsenide diode laser through a specially designed frequency doubler. Careful feedback controls enable this process to operate efficiently at the power levels required. The blue laser technology has been licensed to Coherent Inc who will initially be targeting industrial, medical and printing applications.

Blue lasers have long proved attractive as a means to increase disk storage desnsity. Now IBM claims to have the beginnings of a real blue laser system.

The successful high density data recording experiments have depended, not just on blue laser technology, but also on the parallel development of new encoding techniques, new magneto-optic materials and a new way of mechanical tracking.

As the marks written on a disk become smaller, it becomes increasingly difficult to keep the head positioned accurately over the data track. Current optical disks use grooves for the purpose, but the hope is that these will be eliminated because of the additional noise they cause. The latest experiments successfully used a "sample" servo to track a grooveless disk by rapidly analysing the presence of periodic patterns to determine whether or not the head is accurately positioned over the data.

Top people are always in a spin

EW + WW can now reveal the secret to staying at the top of the pile. To succeed in the stock market, politics, courtship or the media, you have to behave just a little unpredictably. At least that is the conclusion of a study by two researchers at the University of Illinois in Urbana.

Computer simulations by physics professors Alfred Hubler and David Pines show that the best strategy in any kind of chaotic situation is to be neither completely orderly nor totally unpredictable. According to Hubler: "a competitor who imposes a weakly chaotic dynamic on a chaotic environment stays leader for the longest."

Using mathematical terms to define variables such as a competitor's predictability and changes in the environment, the researchers charted the results of several hypothetical long-term interactions. Their simulations suggested that the most productive arrangement for two competitors is to exist in a leader-follower relationship. The arrangement avoids conflict, whether in the home, in the workplace or between nations.

Hubler and Pines' conclusion is hardly surprising. But what is interesting is the finding that such stable relationships last longest when a small degree of unpredictability is introduced by one or both participants. If a person, or organisation, behaves too predictably, a competitor can figure out a pattern and attempt to short-circuit it, in turn, destabilising the relationship.

The logical advice to companies, therefore, is to avoid making your corporate strategy too transparent or unchanging.

On the other hand, says Hubler: "if behaviour is too chaotic, unpredictable and non-reproducible, it may alarm competitors who were formerly content."

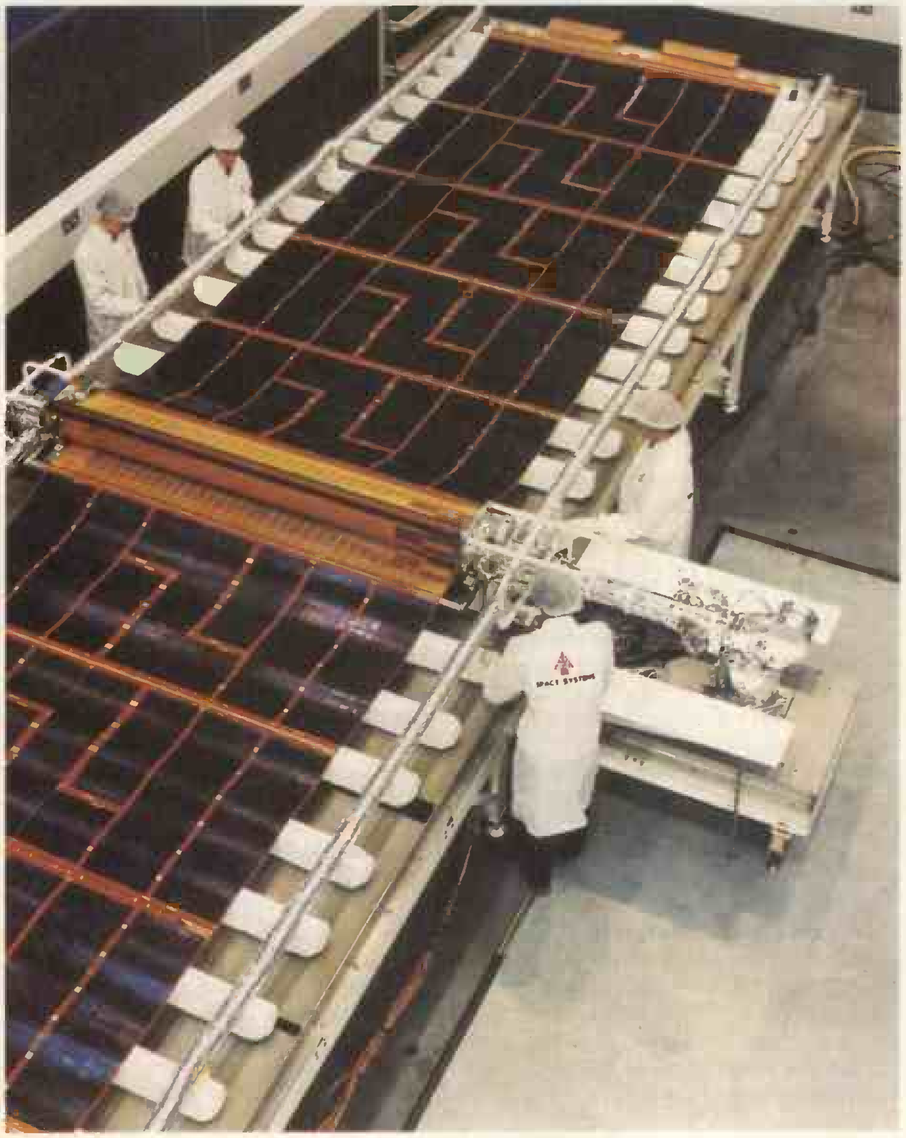
Such formerly happy competitors may then energise all their efforts in a powerful, but dangerous over-reaction. It's the sort of thing that leads to price battles, stock market collapses and ultimately, wars.

According to the computer simulation, a small degree of unpredictability is desirable for two reasons. Not only does it stabilise your position at the top, it also offers the tools to respond more quickly to changes in the economic or political climate.

Space service will tackle Hubble's wobbles

When the Space Shuttle takes off – probably in December – to service the Esa/Nasa Hubble space telescope, one of its tasks will be to replace the pair of solar panels supplying the telescope with power.

The solar arrays consist of about 50,000 silicon cells covering a surface area of 70m^2. They generate about 5kW of electrical power when new, yet must be flexible enough to be rolled up for storage, but rigid enough to remain stable when unfurled in space. To meet this need they are constructed of huge sheets of fibreglass-reinforced Teflon held in place with metal struts,



Unfortunately, after the telescope was launched in 1990, its pointing capability was affected by jitter caused by expansion and contraction of the solar arrays every 90 minutes as the craft passed in and out of sunlight. Temperature variations of 200°C caused the tips of the panels to bend out of true by up to 30cm.

The problem was solved provisionally by anticipating the temperature changes and issuing commands for the telescope to change its direction to counter the changes caused by the flexing solar panels. But this extremely cumbersome procedure occupies a large amount of on-board processing power.

Effects of the flexing have been made worse by the fact that the expansion and contraction is jerky. The new panels, with fewer moving parts, are being carefully designed to minimise the problem. Each support boom is covered with an aluminium layer supported by 900 plastic discs forming an accordion-like structure. The structure will reduce thermal gradients by a factor of 20 while, at the tip of the arrays, a complex system for countering expansion has been replaced by one of frictionless springs. The deployment drum will also be immobilised by an electrical brake.

If all this seems a bit over the top, it is important to realise the degree of stability necessary. The telescope's precision pointing system demands that residual torque from the solar cells must not exceed 0.002Nm – equivalent to the weight of half a paper-clip held at arm's length!

As well as ridding Hubble of its wobble, the astronauts will be fitting an optical device to correct the aberration of the primary mirror. Space testing and servicing of the device will make use of the faint object camera (foc), The foc, one of the main contributions from Esa, the European Space Agency, is one of several light sensors on the telescope, and consists of a powerful image intensifier able to detect stars of magnitude 29. That is five billion times too low to be seen by the naked eye and is roughly equivalent to a candle on the moon. In conjunction with the telescope optics, the foc is sensitive enough to distinguish between a pair of coins at a distance of 200km. At the moment, one of its two detectors is showing signs of deterioration and will also need to be fully re-tested

Final testing at British Aerospace of the roll-out solar array for the Esa/Nasa Hubble space telescope. The first servicing mission is scheduled for December.

Geniuses are made not created

At a time when British industry is desperately trying to seek out (and retain) its brightest talent, it is illuminating to ask – as psychologists at Florida State University have - exactly what are the ingredients of talent or giftedness.

Florida's Anders Ericsson, professor of psychology, may have an answer. He has been studying world-class experts in music, chess, sport and the arts to learn what talent really means. It seems that all we need to emulate feats – such as the amazing ability to recite pi to 32 000 digits – is practice.

Reassuringly, he is discovering that experts may not be any smarter than the rest of us. Indeed, to judge from Ericsson's findings, giftedness has as much more to do with environment than with whatever "clever" genes we inherit. The common factor that emerges when top people are studied is not their IQ but their dedication. As Ericsson casually observes with delightful understatement: "By the time Mozart was 7, he had had a lot of practice."

Unfortunately, there may be a certain age after which you might as well give up trying to become a megastar – in the artistic professions at least. A common feature of many famous musicians and sportspersons is that they started practising consistently from a very early age. Genius at the Mozartian level appears to be difficult to create later in life if the foundations have not been laid during childhood and adolescence.

Fortunately, for those of us in more mundane engineering or academic professions, you don't necessarily have to start your PhD thesis at 18 months of age. Apparently, brilliance in these areas can flower much later in life.

To illustrate the relevance of sheer hard work and practice, professor Ericsson cites the example of students who were put through a rigorous training programme to help them memorise numbers. After a week of practising an hour a day, a typical college student could recite the value of pi to about 12 or 13 digits. After two years' practice, he or she could manage about 80. Ericsson is now continuing his research into the ability of the 35-year old psychology graduate and his 32 000 digit pi – taking about three hours to recite: he's hoping to reach 100 000.

Ericsson says that people are not born with this sort of exceptional memory. Being gifted in this – or indeed other areas – seems to be largely a matter of motivation.

Ericsson doesn't imply that everyone can become a genius. There are undoubtedly genetic differences that make some people more likely to succeed than others. But the assumption that talent is largely in the genes is not supported by the evidence.

The key difference between ordinary people and highly talented people is the extent to which they are motivated to practise, and it is not just the quantity of practice, but its quality. Endless repetition will not create talent: the sort of practice that makes for genius is well organised and aimed at a specific goal.

Truly talented people, says Ericsson, regard properly directed practice as so essential that they come almost to enjoy what other people find an unrewarding grind. They are also motivated to seek out other experts who will direct their practice effectively. As yet no-one knows why some people have the necessary motivation to become geniuses and others, with equal intelligence, don't. Pushy parents, I'd guess.

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

Diamonds without pressure

Chemists at Pennsylvania State University have devised what they say is the first method for making diamond from a soluble, solid-phase non-diamond material. The process works at normal atmospheric pressure and at relatively low temperatures.

Natural diamonds are created under conditions of extremely high temperatures and pressures deep inside the Earth, and most existing industrial processes employ similarly energy-intensive approaches. The Penn State chemists, led by Patricia Bianconi, have adopted a less aggressive and therefore potentially cheaper technique.

They have developed a polymer, called polyphenylcarbyne, which looks like brown sugar and has a natural tendency to turn into diamond when heated because some of its atoms are already in a diamond-like configuration. The polymer is also the first proto-diamond material that can be dissolved in solvents. It opens up the possibility of being able to coat objects with the polymer solution and then, by heating, to cover the object with a protective coat of wear-resistant artificial diamond.

Bianconi speculates that such artificial diamond could have widespread applications in the electronics industry. Not only does it have the highest melting point of any known material, it also has a very high thermal conductivity. She says that it might be possible to draw diamond components directly onto a chip by coating it with the polymer solution and then directing localised heat at it by means of a laser beam.

Bianconi and her team are now refining the chemistry to reduce the amount of impurity, especially graphite. ■

Penn State protodiamond polymer after processing (40 x). Black areas are microscopic diamond coated with graphite; transparent areas are crystalline diamond.

Ben Duncan analyses distortion performance and harmonic spectra of the UK's ten most popular audio power ICs. In a unique performance guide, he names the good, the bad – and the ugly

SPECTRALLY CHALLENGED:

The top 10 audio power chips

The world's first monolithic audio power amplifier IC was released by Plessey in the UK in 1968. But it took Clive Sinclair to make a success of it as the *IC12*[1]. Since then, IC power blocks have become the mainstay of portable audio and communications equipment.

So who needs reviews? Without them, purchasers will undoubtedly work out which products are poor value and eventually bad business draws to a close. But reviewing provides feed-forward as well as speeding the feedback loop, and quality of many end products – from cars to software – has certainly been forced upwards by continuous published scrutiny.

Electronic components have faced little of this. The open-ended application of component parts appears to make comparative review (beyond checking published specifications) next to impossible. But IC power op-amps are different in that they are made and used mainly for just one task – driving audio into loudspeakers.

They are used in professional audio for talk-back and headphone monitoring; in low budget audio, and by one maker as a power amp driver stage. Similarly, comms, and wherever a few tenths to tens of watts are needed, from a few Hz to 20kHz, are also likely applications.

In use, the small signal stages must be protected from positive feedback induced by asymmetric, Class A-B currents in the adjacent output section and from thermal distortion caused by widely fluctuating Class A-B junction temperatures. So design with low distortion and RF/LF stability is quite a challenge[2], but today there are over 100 IC power amplifiers to choose from, several dating back twenty years. A large number are Japanese and most of these are not readily available outside the orient. The remaining main players are Nat Semi in the US, and Philips and SGS-Thomson (representing Europe).

For practical testing, these companies' 50 or so devices have been whittled down to the most useful ten best sellers in the UK (**Table 1**).

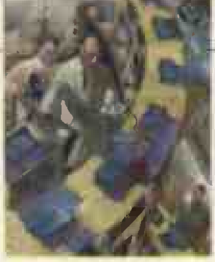


Table 1. Power op-amps under test.

Model	Maker	Nominal Po %THD, load
LM380	Nat Semi	3W, 3%,4Ω
LM383 T	Nat Semi	17W, 10%, 2
LM386 N4	Nat Semi	0.3W, 3%, 4
LM1875	Nat Semi	33W, 1%, 8
TBA820 M	SGS-Thom	1.6W, 10%, any
TDA1514	Philips	50W, 0.1%, 4
TDA2030	SGS-Thom	16W, 0.5%, 4
TDA2040	SGS-Thom	36W, 10%, 4
TDA2611	Philips	12W, 10%, 8
TDA2822 M	SGS-Thom	1W, 10%, 8

These 10 power op-amps were chosen as being presently the largest selling models in UK, from European & US makers. Maximum power outputs are maker's specification into lowest rated load at highest rated supply, or closest tabulated rating. They cannot be directly compared, as the loads and %THD reference points differ: for example, power ratings at 10% THD will be higher than those taken (more properly) at 0.1% THD.

Power capabilities range from hundreds of milliwatts (*TBA820M*) up to tens of watts (*TDA1514*).

Testing standards

The devices being tested vary greatly, and bold decisions were needed to make the processing reasonably straightforward while giving meaningful results. To begin with – and unlike most IC op-amps – over half the group of power ICs is devoid of a standard package or pinouts. The four units in the *Pentawatt* package are the exception. With consistent pinout, and dual polarity supplies and inputs, the *Pentawatt* is a universal receptacle for amplifiers configured as op-amps.

Most of the remaining ICs are not pure op-amps (note the requirement for a DC blocking capacitor on the *LM383*'s –ve input, for example, signifying internal input biasing). But they all employ mainly direct-coupled stages with overall NFB. The *LM380, 386* and *TDA2611, 2822* have fixed gains to save on parts count, ranging from +26 to +39dB. This is too much gain for use at common line levels, but allows most transducers to be connected without intermediate preamplification – a particularly useful facet in ultra compact, low budget equipment. On the other hand, it prevents testing at a uniform gain.

Gain can be adjusted on all the remaining devices, and – with the exception of the *TBA 820* – it is set by an external, shunt feedback

Figs. 1 to 10 show %THD (on left) vs input level in dBμ (horizontal axis) vs output level in Log watts (on right). %THD is the solid line. Transfer function in watts is dashed. By selecting onset of clip or 0.1% THD (whichever is highest) as the reference, power outputs can be meaningfully compared between these figures. Note that some devices employ 30V and others 15V, giving target powers of about 21 and about 4W, assuming 2V saturation losses. The comparison suggests some devices (eg LM 383T and TBA 820) have either rather optimistic power specifications or have entered thermal shutdown when swept – a fact not registered by thermal monitoring. Retests from cold show that spectral results are essentially unaffected by thermal conditions.

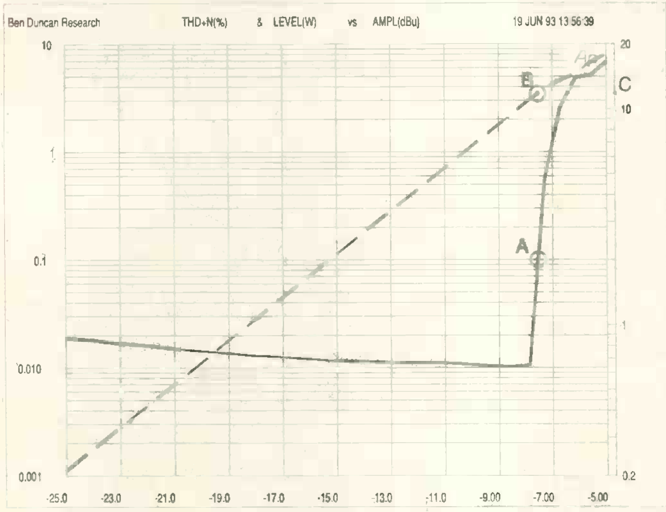


Fig. 1. TDA1514

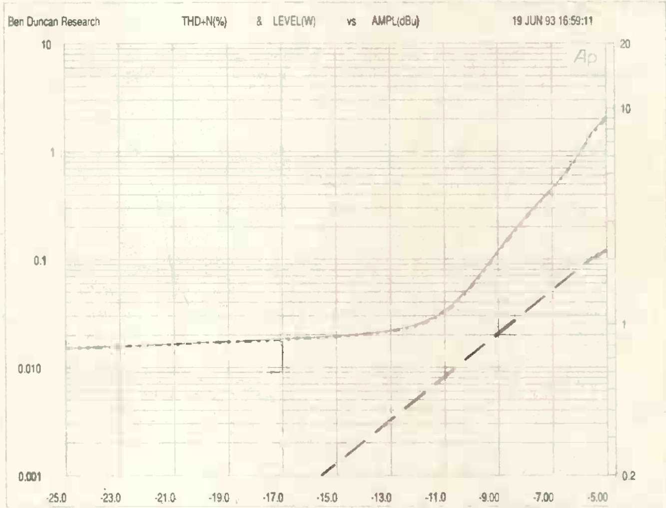


Fig. 3. LM383 T

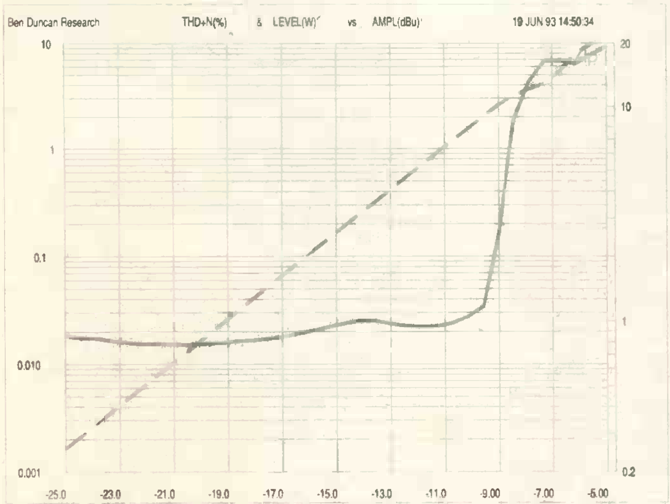


Fig. 2. TDA2030

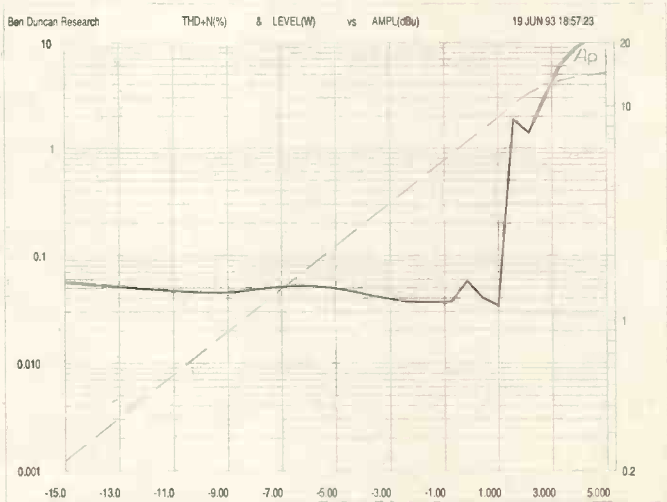


Fig. 4. LM1875

Fig. 5 to 10. The dynamic THD and transfer plots on this page show a variety of pathologies. All plots are at 5kHz.
In Fig. 8, the distortion is barely out of the noise floor (at –31dBu input drive) before it begins to rise again, into a very soggy asymmetric clip, rather like a single ended valve amplifier. The transfer functions for all the others bend as clip is entered.
The almost flat distortion curves in Figs. 6, 9 and 10 are characteristic of crossover distortion.
In Fig. 5, distortion (and noise) are commendably low for signals 20dB below clip, but %THD rises at a steady rate thereafter.

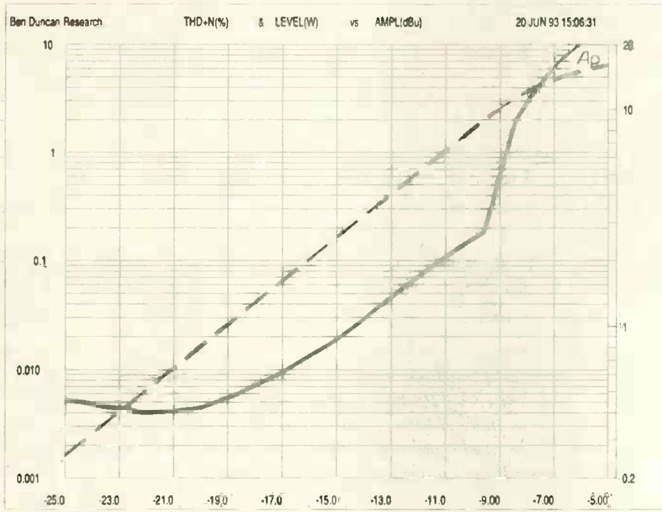


Fig. 5. TDA2040

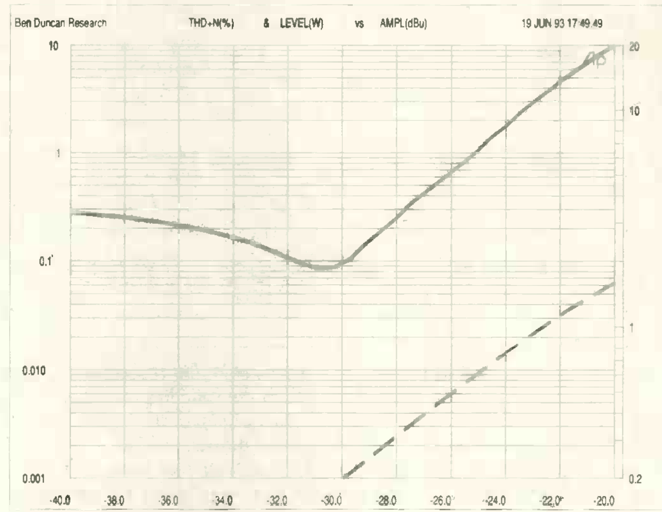


Fig. 8. TBA820 M

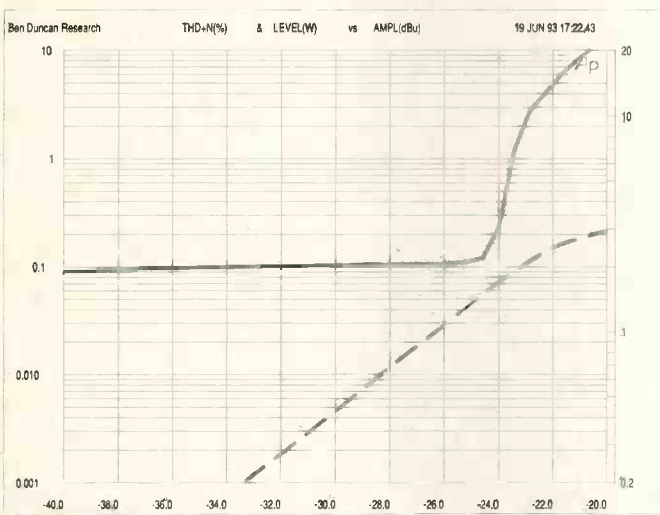


Fig. 6. TDA2611 A

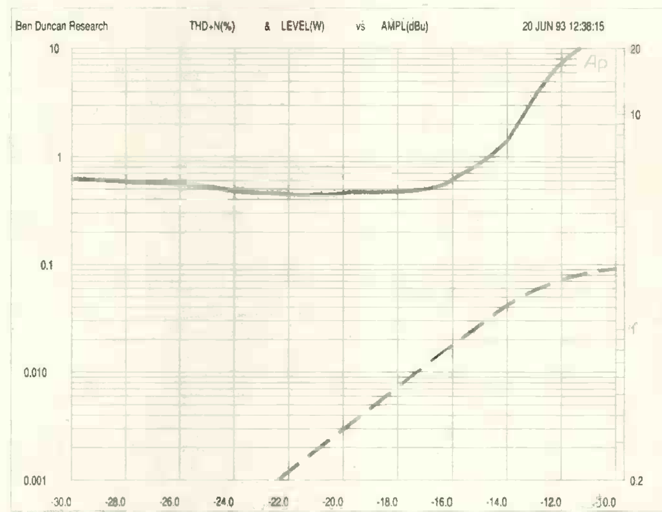


Fig. 9. LM386 N4

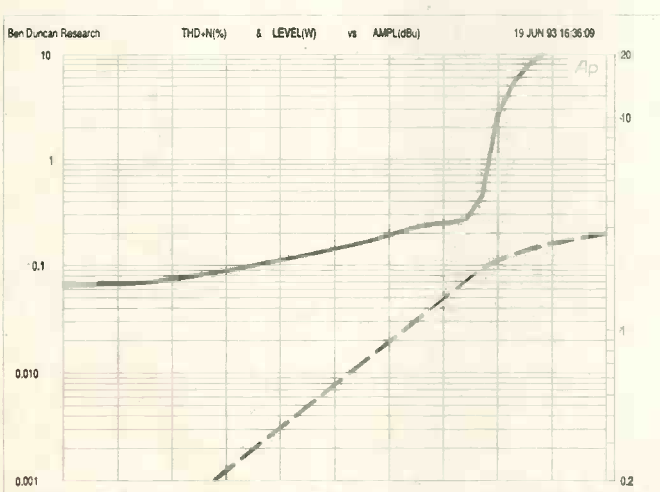


Fig. 7. LM380

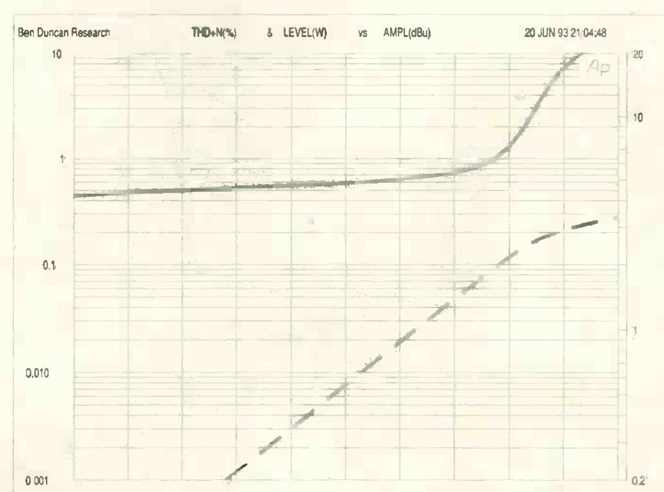


Fig. 10. TDA2822 M

divider. The *LM383* (**Fig. 23**) employs unusually low resistor values here to enhance PSRR. Instead of fixing the gains of the flexible devices at one arbitrary level, gain is allowed to follow the maker's evaluation circuit, excepting the *LM383* and *LM1875*, which are set at 20dB.

The conditions are reasonable in that all the devices have gains which are close to those most used in practice, considering the ICs' span of power ratings. For example, the higher power devices are less liable to be used in portable battery-powered systems, and rather more likely to be driven at line level, requiring less gain.

All devices are operated in the non-inverting mode as this is the sole option for the non op-amp types (**Fig. 22**), excepting the *TDA1514*. As the lowest gain is +20dB, common mode distortion will be expected to be a minor feature.

Most of the lower power ICs work on single rails, while the higher output models employ dual supplies. Maximum supply voltages vary from ±30V (*TDA1514*) down to +15V (*TDA2822 M*). The test supply is fixed at the lowest device's maximum: +15V for single rail devices and ±15V for dual rail devices. Under these conditions, the loop gain of the devices having significantly higher maximum voltages will be depressed about 10 to 15%. This affects % THD above 5kHz, but only marginally against the background of loop gain unit-to-unit tolerance of 30 to 50%, and the gain differences already discussed.

Test circuits

Looking at the test circuitry in more detail, on all the single rail circuits, the output DC blocking capacitor is standardised at 1000μF. Some circuits specify – and most received – an RF filter capacitor at their inputs. On several ICs, it is mandatory, despite being driven by just 1.5m of screened cable from a 25Ω source. Some units need a DC blocking capacitor at their inputs to preserve their biasing.

Figs. 11 to 21 show harmonic spectra and noise. Harmonic levels are slightly changed when testing preceeds from switch on, as opposed to being pre-warmed by quiescent current. If further changes were evident over logarithmic time intervals, the audiophile's claim that equipment sonics change over a long period of 'warm up' would be potentially validated, subject to further psychoacoustic translation. And while the broad pattern of the harmonics were only slightly changed with temperature, this might amount to a disproportionate audible change.

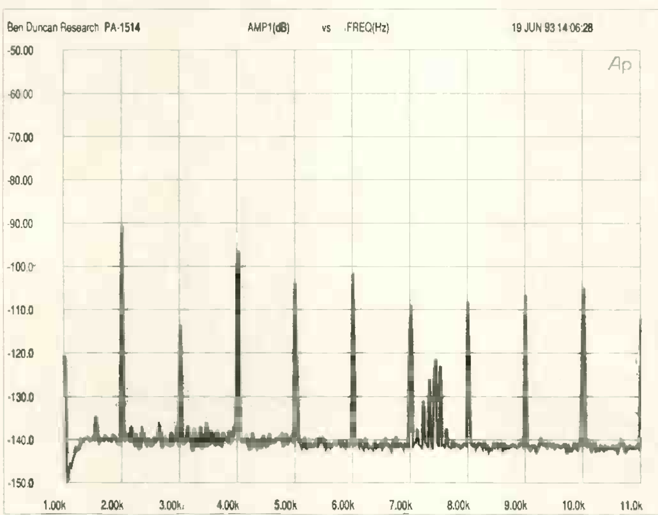


Fig. 11. TDA1514

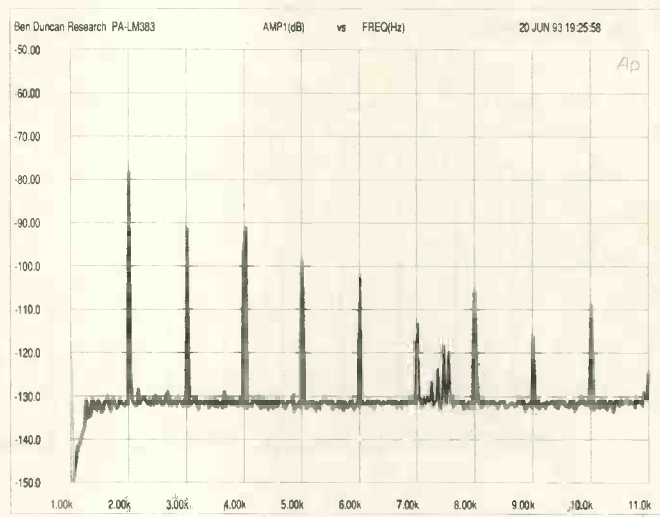


Fig. 13. LM383 T

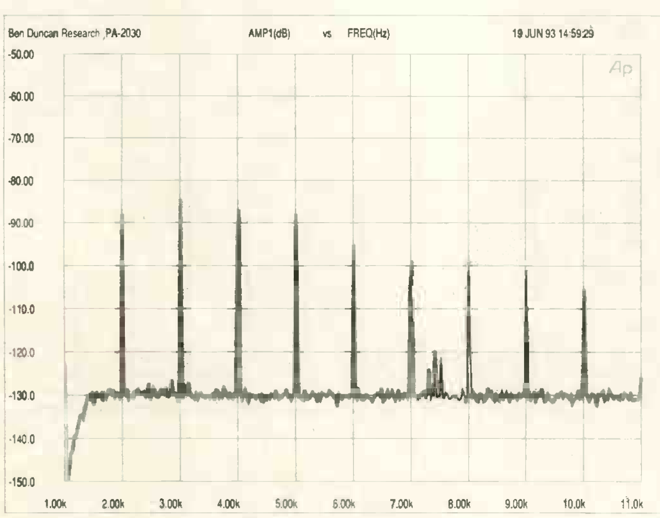


Fig. 12. TDA2030

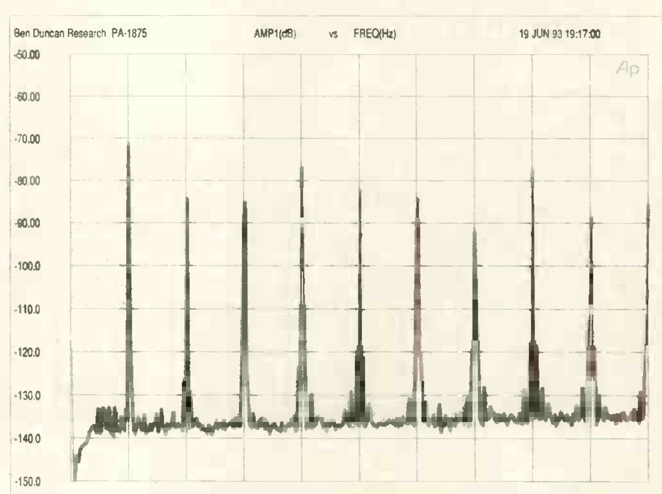


Fig. 14. LM1875

Figs. 15 to 20. In this group the TDA2040 is a good example of monotonically reducing harmonics, with the exception of the 10th and 11th. This kind of structure is more pleasing to the ear, and more like the harmonics in euphoric musical sounds than the humped pattern of the LM380 in Fig. 17. At over twenty years, LM380 is the oldest device in the tests, and the only DIL unit having provision for copper heatsinking "wings" to be soldered to 6 of its pins. But the even more ugly, roller-coaster harmonic pattern of a much newer device like the LM386 (Fig. 19) shows that audio quality evolution cannot be counted upon.

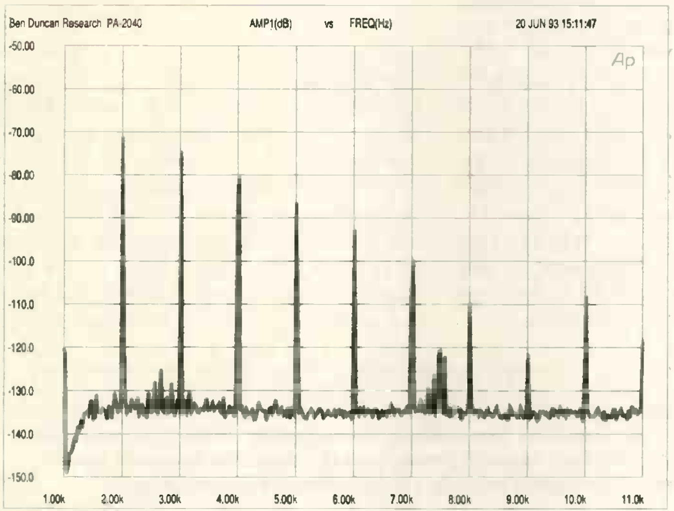


Fig. 15. TDA2040

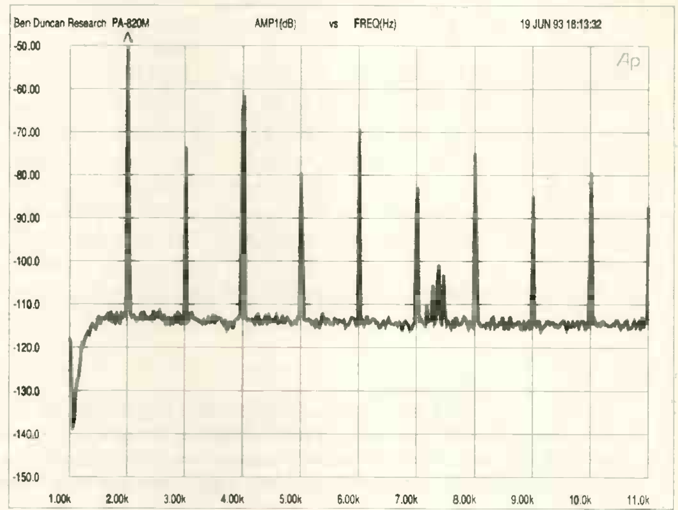


Fig. 18. TBA 820 M

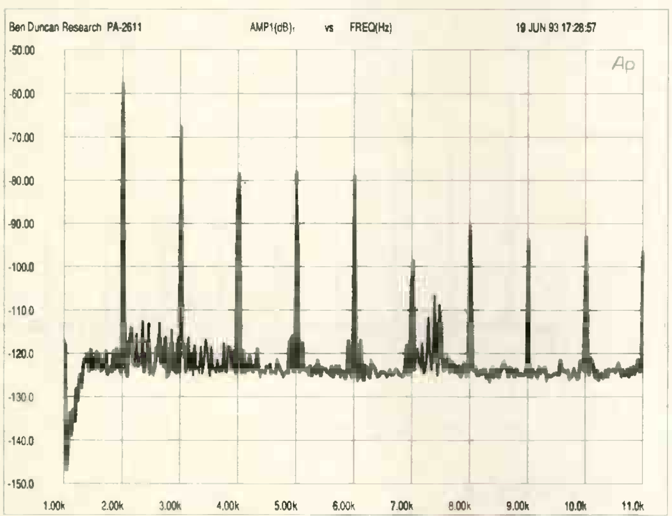


Fig. 16. TDA2611 A

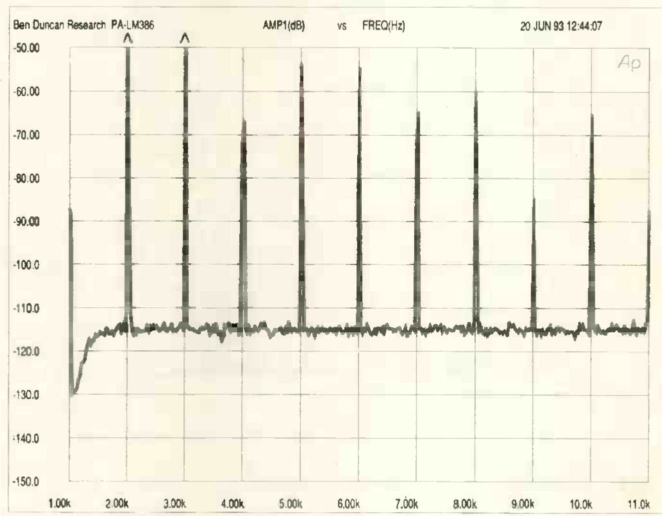


Fig. 19. LM386 N4

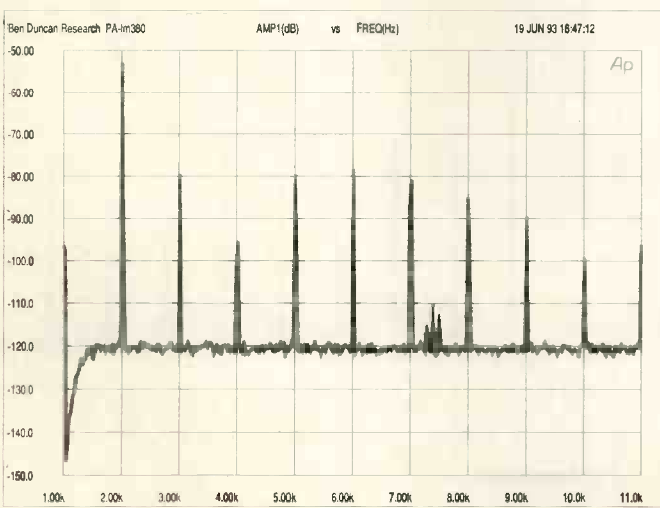


Fig. 17. LM380

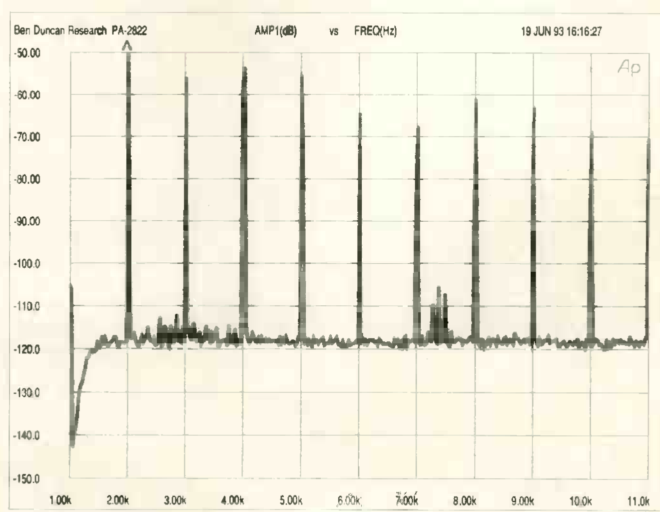


Fig. 20. TDA2822 M

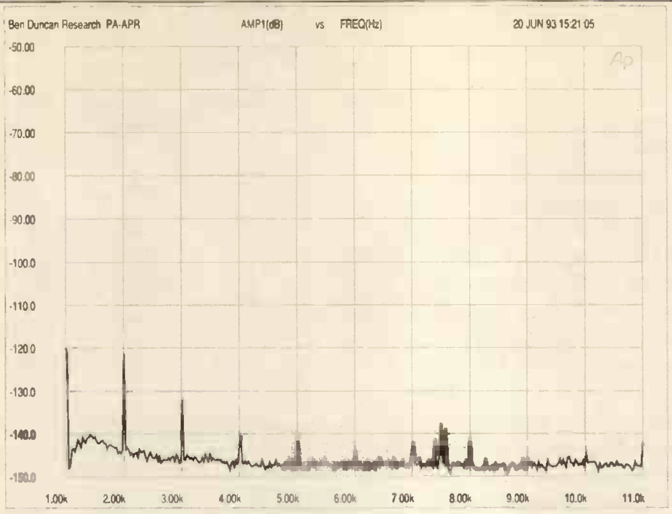


Fig. 21. AP residue

The *TDA2611*'s stability margin is such that the Audio Precision analyser's mildly reactive loading had to be isolated with 1k. The *TDA1514* works well despite its modest supply decoupling. Past experience led me to increase the supply decoupling on other units beyond the maker's recommendations.

The 470μF rail decoupling capacitors are all Low-Z at HF types. Power is taken from a regulated supply with a reasonably low source impedance, below 10mΩ (including leads & connections), from below 100Hz to above 10kHz. In practice, many battery and low budget mains supplies will be considerably higher.

The circuits are built on Veroboard explicitly made for analogue prototyping and solely available from RS (433-911). As each node has five holes, inter-track capacitance is limited to a few pF. RF stability and low THD are assured by noding all the main grounds to a common point, and by running solder along power, output and ground tracks to keep resistance down. Hook-up is with 1/0.6mm solid core wire with a maximum length of 50mm.

Test procedure

All testing is performed into an 8Ω load, with about 1nF of load capacitance. As a preliminary to spectral testing, each DUT (device under test) is checked by first performing a bandwidth sweep. Several have secondary (HF) crossover distortion, slew limiting and/or RF oscillation above 20kHz. Results of the tests led to some of the fixes described above.

Next, %THD is plotted dynamically, against level. The DUT is driven from 15 or 20dB below clip, up to and beyond clip, at a spot frequency of 5kHz. The characteristic of a

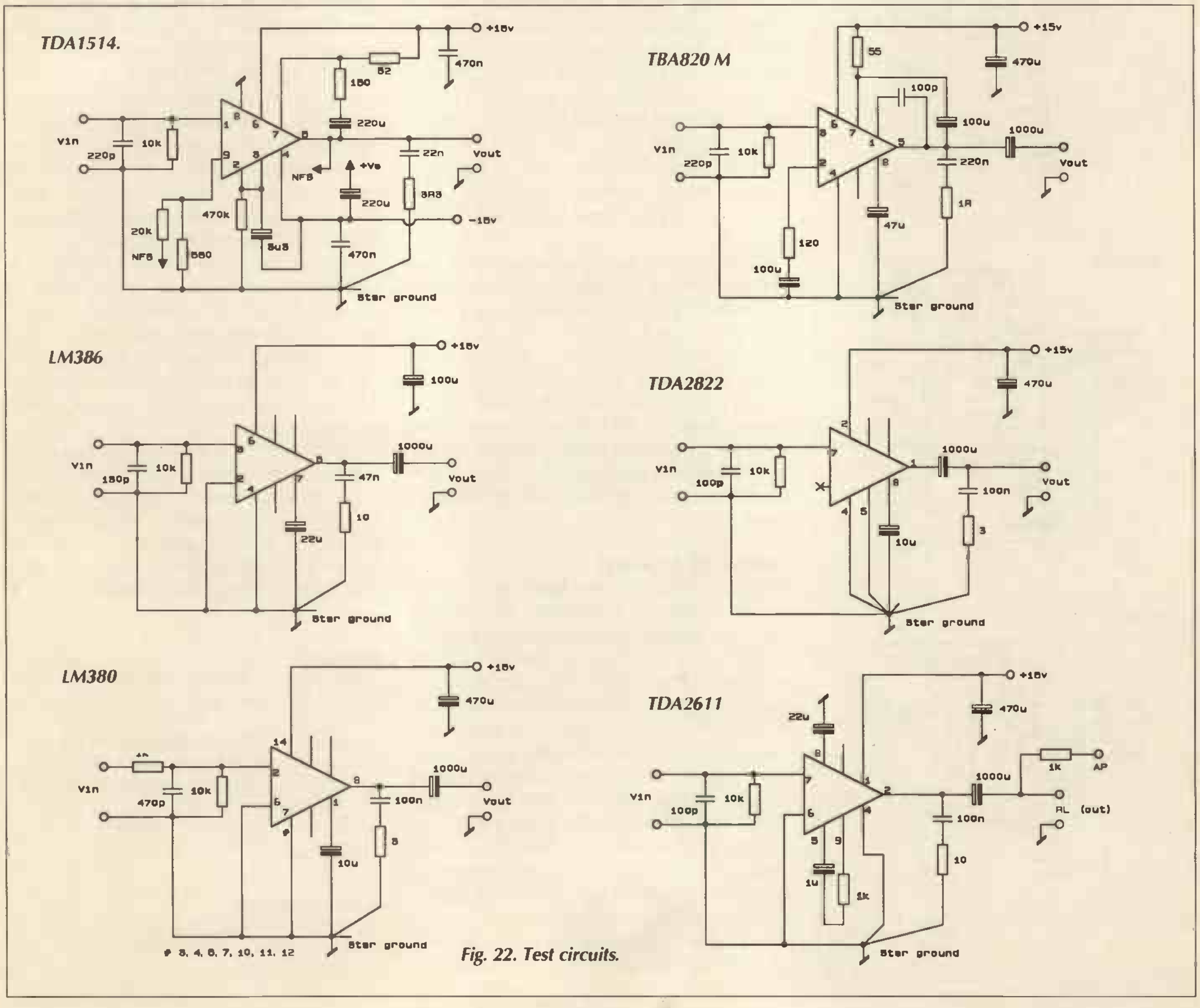
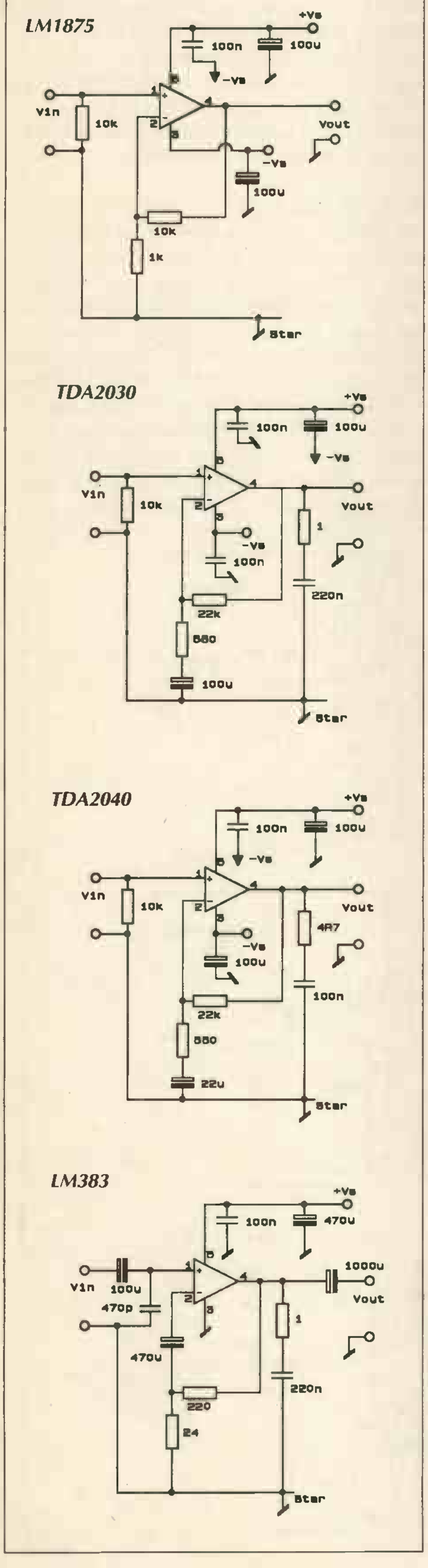


Fig. 22. Test circuits.

Fig. 23. Test circuits, all Pentawatt package.



high-performance, clean device with high NFB is a "lazy V" or sideways "L" shape. The noise floor at first drops linearly to reveal the real %THD, and the THD then rises abruptly, beyond clip. Looking at the results in **Figs. 1-10**, the *TDA1514* and *LM1875* come closest to this. The same Figs also plot the transfer function, in watts, with a log scale so that the power can be related to a given %THD and drive level.

On Fig. 1, for example, 0.1% at *A* on the THD graph corresponds to point *B* (draw a straight vertical line above), which is around 11W. Looking down shows that input drive is around –7.5dBu. The AP's %THD residue is typically below 0.0007% up to clip, much lower than even the best DUTs.

As in previous tests[3], the devices' harmonic spectra are then plotted, using the Audio Precision dual domain test set's DSP facility, set at 48kHz sample rate to provide a tight, 3Hz bandwidth. All the units had spectra (**Figs. 11-20**) much higher than the AP residue (**Fig. 21**) and their own noise, so there is none of the uncertainty that plagues measuring the better op-amps. Processing is speeded up by averaging just 16 samples for certainty, and all testing was done with DUTs placed in a steel shielding tray. The results of Figs. 1-10 were used to set the test level at between 1 and 2dB below clip – the latter defined clearly in most cases by the point where the %THD graph turns irrevocably upwards.

All the devices heated up when tested, and whereas external heatsink temperatures would be lower with program instead of a sinewave, and with a real loudspeaker as a load, peak junction temperatures could be much the same. Some tests were run immediately at switch on, and while several showed greater power on output, %THD and spectral results were only marginally better, if different at all, particularly with the 8pin DIL devices, which have small junctions and high thermal resistance.

From best to worst

Ranking devices by absolute level of their spectra (**Table 2**), it is fitting that Philips' *TDA1514* appears as the top performer, as it's trumpeted as a "high performance Hi-Fi amplifier", as well as the being the newest in the group (1991). But that two of the oldest (circa 1975-8) devices came next best shows that evolution in audio quality is not to be relied upon. If the harmonic pattern is taken to be more critical to the ears than the absolute levels of spectra, then the clear leaders are *TDA1514* and *TBA820M*, with their even-order dominance. On this basis, asymmetric clipping and even the *820*'s primitive, single-ended output stages are beneficial!

Next most likely to be preferable on sonic grounds are the units with monotonically

Table 2. Spectral ranking generally ordered by 2nd harmonic level. Cleanest devices with least residue are at the top.

Model	Package	Harmonic pattern	Gain * = fixed
TDA1514	SIL	Even	+30
TDA2030	Pen	Mono	+30
LM383 T	Pen	Mono	+20
LM1875	Pen	Cplx	+20
TDA2040	Pen	Mono	+30
TDA2611	SIL	Mono	+38*
LM380	DIL 14*	Cplx	+34*
TBA820 M	DIL 8	Even	+34
LM386 N4	DIL	Cplx	+26*
TDA2822 M	DIL 8	Cplx	+39*

Notes:
Pen = *Pentawatt* = five leg T0 220
Even/Odd = even/odd harmonics dominate
Mono = harmonics slope off monotonically
Cplx = complex pattern
* = fitted with 2in^2 heatsink.

descending spectra (*TDA2030, 2040, 2611* and *LM383*). Soft clipping may contribute to good sonics, where the amplifier's swing is forced (by design, battery technology and budget constraints) to be inadequate. When this is the case, the *LM386, TBA820, TDA2822* and to a lesser extent the *LM2040* are expected to sound least painful in overdrive. Much hearing damage is caused by compression or signal clipping, not high SPLs *per se*, so it is hoped that designers of equipment driving headsets and earpieces (particularly domestic replay equipment likely to be used by children) are making notes... Ideally, audio designers will all be studying Douglas Self's treatise (series running in this issue) on fundamental bipolar amplifer distortion pathologies too.

Looking now at Fig. 10, the *TDA2822M* is strictly unacceptable as an audio device, on the grounds that THD consistently above 0.1% from a device with overall feedback is bound to be fatiguing and irritating. It could prove fatal (or at least needlessly stressful) if used to drive an air traffic controller's or pilot's headset, for example.

Finally, how do ICs compare to real power amplifiers? Well compared to the professional PA amplifiers measured so far, the top five ICs are returning similar absolute levels, though with different spectra. The remaining five are probably not much different to most tight budget commercial amplifiers. ■

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Acknowledgement

The author would like to acknowledge the assistance given by Audio Synthesis and Macro Marketing.

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1 x DC Voltage Reducer. 12v-6v. Order Ref: 916.
1 x 10 amp 40v Bridge Rectifier. Order Ref. 889.
Lightweight Stereo Headphones. Moving coil so superior sound. Order Ref: 896.
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2 x nicad Constant Current Chargers. Easily adaptable to charge almost any nicad battery. Order Ref: 30.
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2 x White Plastic Boxes. With lids, approx. 3" cube. Lid has square hole through the centre so these are ideal for light operated switch. Order Ref: 132.
2 x Reed Relay Kits. you get 8 reed switches and 2 coil sets with notes on making relays and other gadgets. Order Ref: 148.
1 x Big Pull Solenoid. Mains operated. Has ½" pull. Order Ref: 871.
1 x Big Push Solenoid. Mains operated. Has ½" push. Order Ref: 872.
1 x Mini Mono Amp. 3W into 4 ohm speaker or 1W into 8 ohm. Order Ref: 268.
1 x Mini Stereo 1W Amp. Order Ref: 870.
1 x In-Flight Stereo Unit is a stereo amp. Has two most useful mini moving coil speakers. Made for BOAC passengers. Order Ref: 29.
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2 x Lithium Batteries. 2.5V penlight size. Order Ref: 874.
2 x 3m Telephone Leads. With BT flat plug. Ideal for 'phone extensions, fax, etc. Order Ref: 552.
1 x 12V Solenoid. Has good ½" pull or could push if modified. Order Ref: 232.
4 x In-Flex Switches. With neon on/off lights, saves leaving things switched on. Order Ref: 7.
2 x 16V 1A Mains Transformers. Upright mounting with fixing clamps. Order Ref: 9.
2 x Humidity Switches. As the air becomes damper, the membrane stretches and operates a micro switch. Order Ref: 32.
5 x 13A Rocker Switch. Three tags so on/off, or changeover with centre off. Order Ref: 42.
2 x Flat Solenoids. You could make your multi-tester read AC amps with this. Order Ref: 79.
1 x Suck or Blow-Operated Pressure Switch. Or it can be operated by any low pressure variation such as water level in tanks. Order Ref: 67.
1 x 6V 750mA Power Supply. Nicely cased with mains input and 6V output lead. Order Ref: 103A.
2 x Stripper Boards. Each contains a 400V 2A bridge rectifier and 14 other diodes and rectifiers as well as dozens of condensers, etc. Order Ref: 120.
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2 x 5 Aluminium Fan Blades. Could be fitted to the above motor. Order Ref: 86.
1 x Case. 3½x2¼x1¾ with 13A socket pins. Order Ref: 845.
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4 x Luminous Rocker Switches. 10A mains. Order Ref: 793.
4 x Different Standard V3 Micro Switches. Order Ref: 340.
4 x Different Sub Min Micro Switches. Order Ref: 313.

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2m 3-Core Lead. Terminating with flat pin instrument socket, £1, Order Ref: 879. Ditto but with plug on the other end so that you could use this to extend an instrument lead, £1.50, Order Ref: 1.5P10.
20W 5" 4 Ohm Speaker mounted on baffle with front grille, £3, Order Ref: 3P145. Matching 4 ohm 20W tweeter on separate baffle, £1.50, Order Ref: 1.5P9.

LCD 3½ Digit Panel Meter
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VU Meter. Illuminate this from behind becomes on/off indicator as well, 1½" square, 75 each, Order Ref: 366.
Edge-Wise Panel Meter. Ideal when short of panel space, only 40x14mm, also have built-in LED, 500uAf.sdd, scaled 0-5, £1 each, Order Ref: 131.
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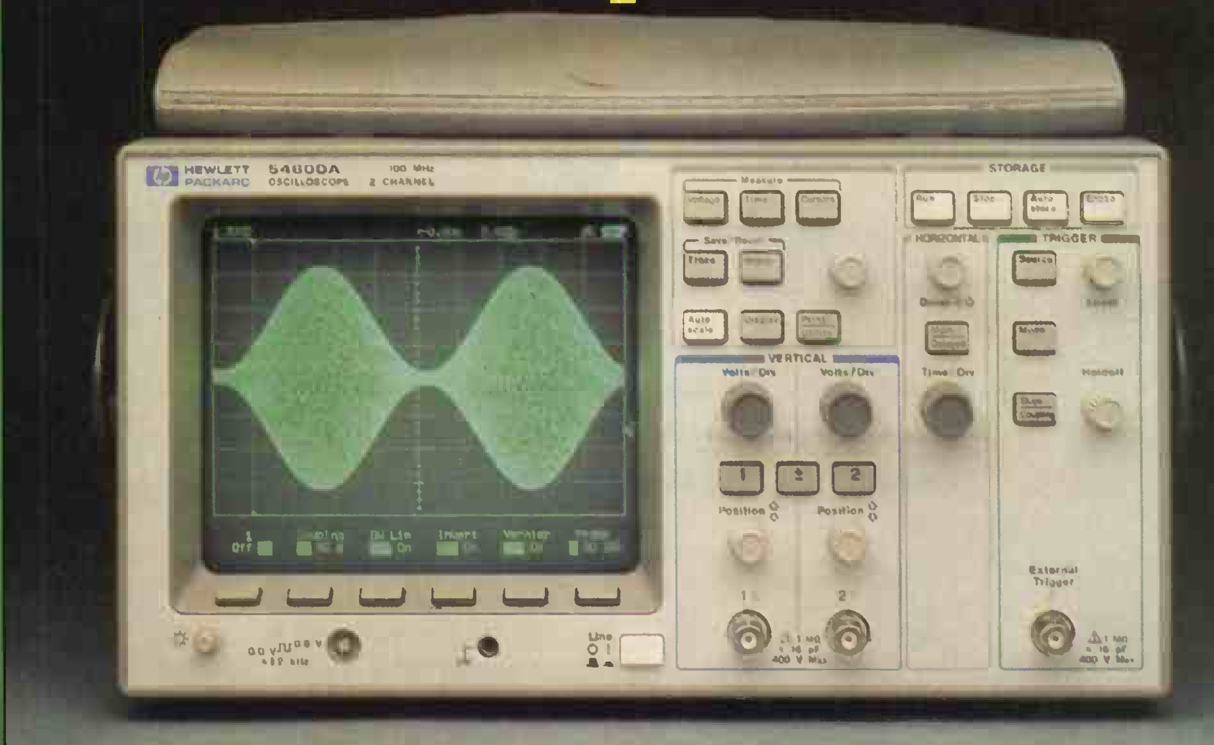

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Sliding into a software specialty

Drawing together various graphs and plots from different sources into a coherent slide presentation is notoriously difficult. Allen Brown assigns Stanford Graphics to the task.

One of the key aspects of any good presentation is confidence. For that, a coherent and lucid aural presentation of the facts supported by well produced visual aids is essential – gone are the days when an engineer could present information simply by using half a dozen transparencies and a few felt tip pens.

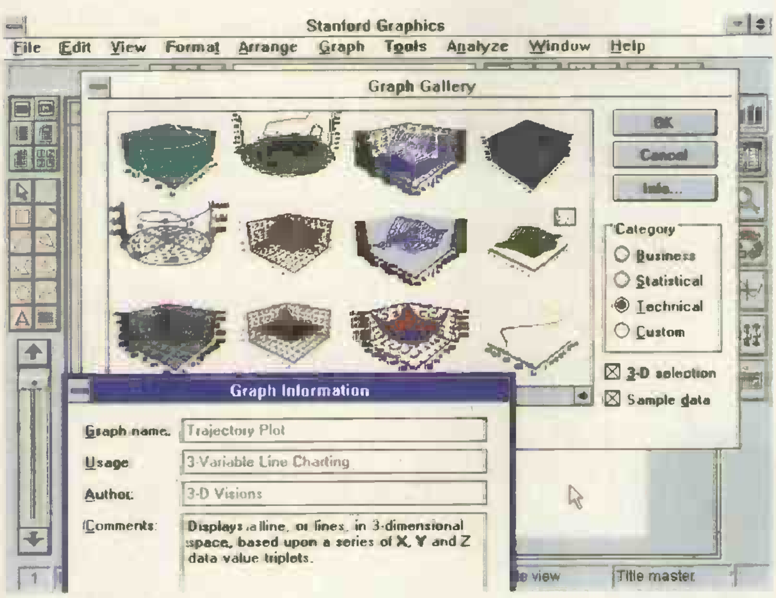
Fortunately presentation software – such as *Stanford Graphics for Windows from* 3-D Visions of California – is making the task of generating quality visual aid material now relatively easy.

Stanford Graphics is an alternative to the well known *Harvard Graphics*, and allows the user to create a set of slides with a consistent format, irrespective of source. The slides may be included in documents or used for audio visual purposes (or both).

Images can be imported from other sources (including TIF and PCX formats) and so slides can incorporate graphs, images and text. Using the slide sorter enables slides to be displayed as an array for final checking.

The finished product could be a set of 35mm colour transparencies for projectors (several companies will now convert EPS image files into 35mm) and a set of accompanying hand outs, including the slide contents, which can be produced from a laser printer. When generating text slides the user has access to the Style Master which allows the creation of several templates which are used to standardise the presentation format.

Graph gallery in the technical section illustrating the 3-D plots available.



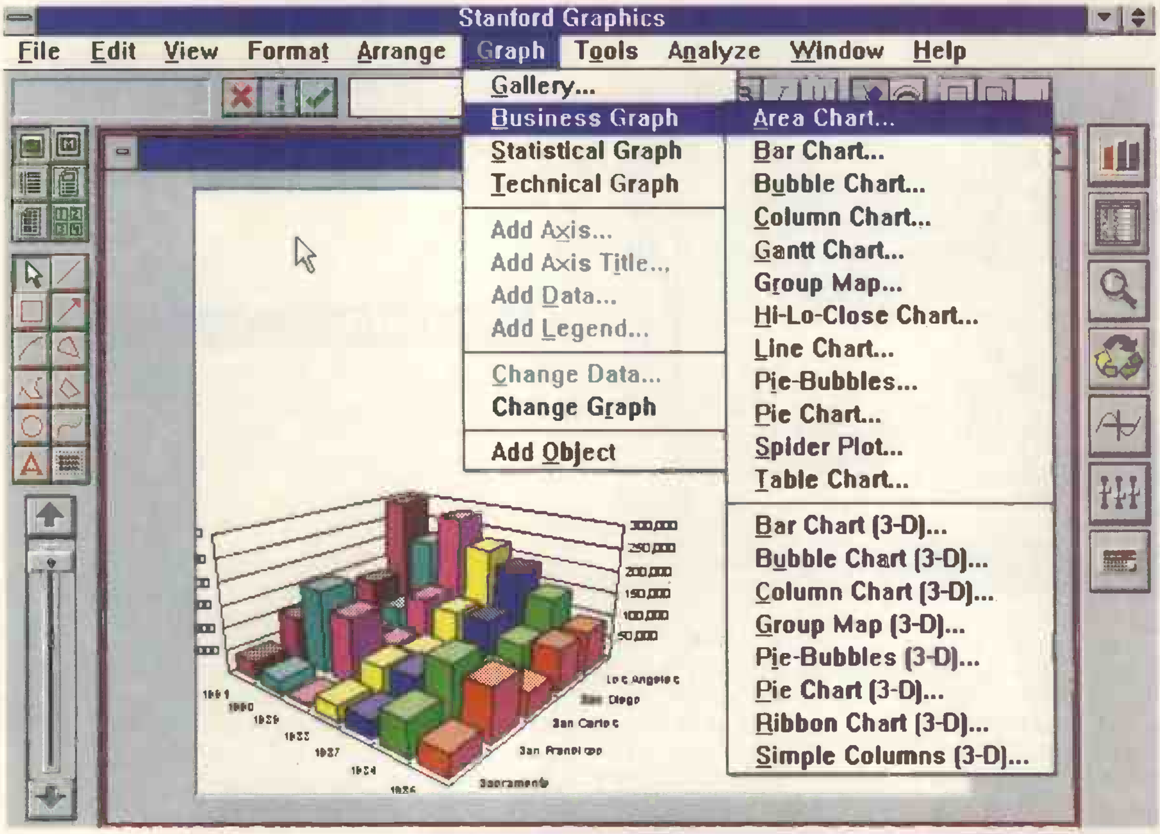
Graphing

Prime feature of *Stanford Graphics* is its large array of graphing formats, grouped into the four categories of business, statistical, technical and custom. A "gallery" enables the user to have a quick look at each option by browsing through the possible designs. Each group offers an impressive range of designs, including 3-D options: some are well known (bar and column graphs); others not so. For example, apparently the Spider plot allows a number of system parameters to be rated on an equal scale, enabling a user to track any parameter moving outside an acceptance envelope. Among the useful business options is the Gantt chart, frequently used to plan projects (unfortunately it still does not predict slippages!).

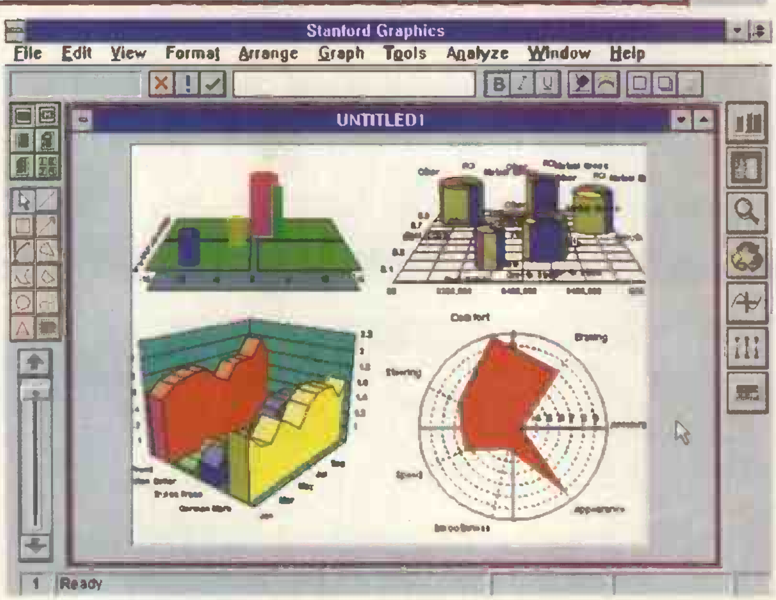
Statistical options include histograms, scatter plots, star plots and error-bar plots, all useful for plotting very basic statistical information. But anyone needing to perform any serious statistics would probably turn to a statistical package such as *SPSS* or *StatGraphics*. These already have an impressive range of graphical representations and since they run under *Windows*, it is highly unlikely that an engineer or statistician would deliberately choose *Stanford Graphics* for statistical processing. But the technical graphing options are very useful and include polar plot, shadow-contour, Smith chart and a surface plot.

Importing data

Importing data from spreadsheets such as Microsoft Excel or Lotus 1-2-3 presents no problem. But if other sources are



Drop-down menu shows the variety of 3-D choices that are offered within business graph options.



3-D and 2-D business graphs include a group map, pie-bubbles, ribbon charts and spider plot.

used, data must be in an ascii file format with a minimum of two arrays (X and Y). Unfortunately, a sequence of data (Y co-ordinates) can not be plotted directly from a data acquisition card or data file, as X co-ordinate values must be supplied. The limitation can be irritating as all data files must be pre-processed to add the scaled X array data.

Imported data is loaded into a spreadsheet-like table, which can be edited, and the spreadsheet associated with each graph can be accessed at any time by clicking on the appropriate icon. As with spreadsheet programs, editing options allow data cells to be manipulated, including a paste function which enables a mathematical function to be applied to a cell or a range of cells (the user's guide is not very informative on this feature, and its use is confusing and needs to be cleared up on a future version). Overall, playing around with the spreadsheet facility makes it clear that technical data should be fully prepared in another package prior to importing it for plotting.

On the graphs themselves, reasonable control is exercised over labelling, axis definition and orientation of the plots, all performed through typical *Windows* dialogue boxes. Through format and graph menus, labelling size, orientation, font and positioning can all be specified, and axis scales can be changed to reflect logarithmic requirements, either dBs or linear. Even the 3-D pie charts can be scaled logarithmically – giving true meaning to " ...lies and statistics".

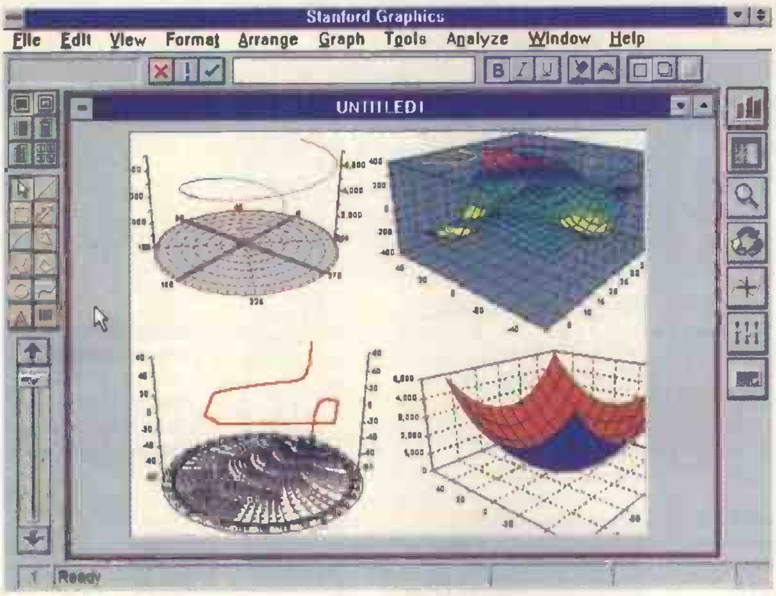
Analysis

Data, once imported and graphed, can be analysed using curve fitting (regression analysis), interpolations, statistics and an FFT.

USER'S GUIDE

Documentation is a detailed and well laid out – though hefty – soft-bound guide. Much of it is taken up with the standard interface features of *Windows*, while the all-important "getting started" section provides an adequate overview of features.

To help the new user, the package powers up to a "quick start" menu, guiding access to standard features. The tutorial included is well thought out, and the resulting impression of this carefully planed introduction to the package is a very favourable one for the first time user.



Smith charts are one of the options in the technical graphs section.

SYSTEM REQUIREMENTS

33MHz 486-PC with 8Mbyte ram
S-VGA monitor
Laser printer

SUPPLIER DETAILS

Stanford Graphics £395+VAT, is available from Adept Scientific Micro Systems Ltd, 6 Business Centre West, Avenue One, Letchworth, Herts SG6 2HB. Tel: 0462 480055

Curve fitting choices are quite reasonable including one for surface fitting which carries out a regression process on a surface plot. Another attractive feature is the graphics equaliser, allowing coefficients to be adjusted after regression modelling analysis has been carried out on a plot. The result is the original curve plus the error difference curve, generated by the coefficient adjustments. It's a sort of "what if?" analyser, though the pictorial graphics equaliser icon only seems to appear in a super-VGA graphics mode.

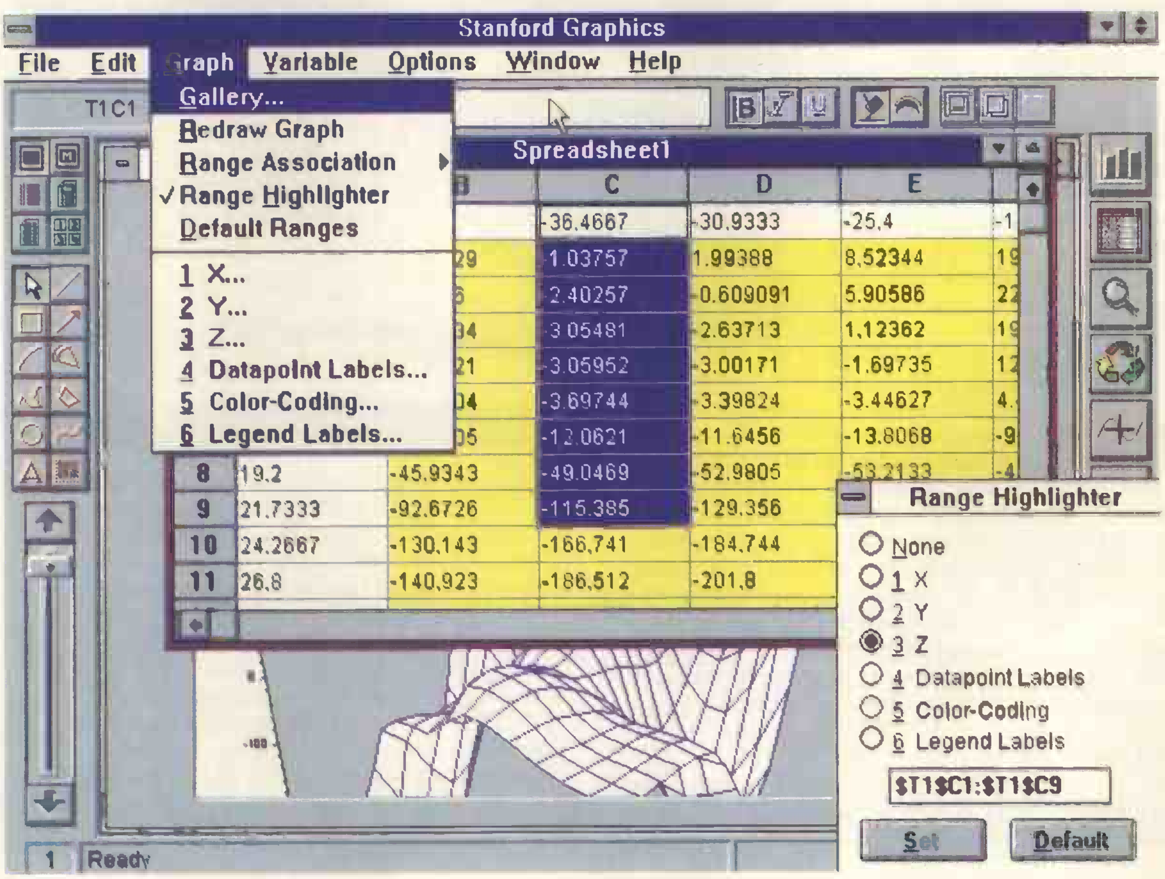
Speed limitation

For generating presentation slides, hand-outs and professionally drawn graphs for importing into desk top publishing packages, *Stanford Graphics* scores highly. Its bewildering array of graphing options leaves the user spoilt for choice.

But there are a few problems, and the main criticism of the package, apart from its inability to plot a single array of data, is speed. Under certain conditions, plotting rate on a 33MHz 486-PC with 8Mbyte ram can be painfully slow – spectral plots for example. Redrawing when the mouse button is released is also irritating and time consuming and "Out of memory" messages keep appearing, even when there may be 3Mbyte of unused ram left.

Running on a 16MHz 386-PC does not bear thinking about and, though possible, should be avoided. To run this package efficiently needs a fast PC and a fair amount of memory. But once equipped, *Stanford Graphics* is easy to work with and very effective for producing presentation information ■

Each graph in Stanford Graphics has a spreadsheet attached to it which can be edited.



Distortion in power amplifiers

The typical circuit which provides most of the voltage gain and all the voltage drive to the output stage of an audio amplifier seems likely to contribute a significant distortion component. Detailed analysis contradicts this. Good design can reduce its contribution to below the noise floor. By Douglas Self.

3: the voltage-amplifier stage

The voltage-amplifier stage (or VAS) has often been regarded as the most critical part of a power-amplifier, since it not only provides all the voltage gain but also must deliver the full output voltage swing. This is in contrast to the input stage which may give substantial transconductance gain, but the output is in the form of a current. But as is common in audio design, all is not quite as it appears. A well-designed voltage amplifier stage will contribute relatively little to the overall distortion total of an amplifier, and if even the simplest steps are taken to linearise it further, its contribution sinks out of sight.

As a starting point, **Fig. 1** shows the distortion plot of a model amplifier with a Class-A output, (±15V rails, +16dBu out.) The model is as described in previous articles. No special precautions have been taken to linearise the input stage or the VAS and output stage distortion is negligible. It can be seen that the distortion is below the noise floor at low frequencies; the distortion slowly rising from about 1kHz is coming from the voltage amplifier stage. At higher frequencies, where the VAS 6dB/octave rise becomes combined with the 12 or 18dB/octave rise of input stage distortion, we can see the accelerating distortion slope typical of many amplifier designs.

The main reason why the voltage amplifier stage generates relatively little distortion is because at LF, global feedback linearises the whole amplifier, while at HF the voltage amplifier stage is linearised by local negative feedback through C_{dom}.

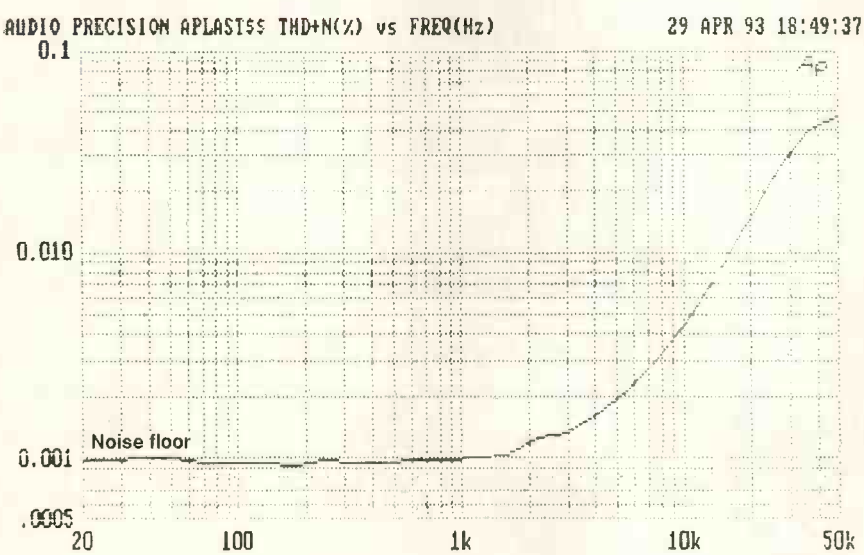


Fig. 1: THD plot for model amp showing distortion below noise floor at low frequency, and increasing from 2kHz to 20kHz. The ultimate roll-off is due to the 80kHz measurement bandwidth.

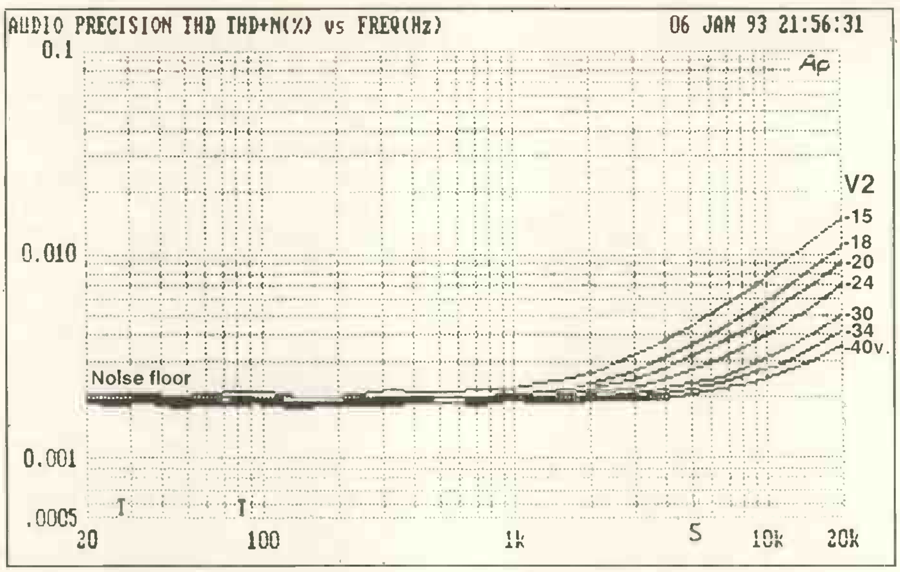


Fig. 2: The change in HF distortion resulting from varying the negative rail in the VAS test circuit. The voltage amplifier stage distortion is only revealed by degenerating the input stage with 100Ω resistors.

Examining the mechanism

Isolating the voltage amplifier stage distortion for study requires the input pair to be specially linearised, or else its steeply rising distortion characteristic will swamp the VAS contribution. This is most easily done by degenerating the input stage which also reduces the open-loop gain. The reduced feedback factor mercilessly exposes voltage amplifier stage non-linearity. This is shown in **Fig. 2**, where the 6dB/octave slope suggests origination in the VAS, and increases with frequency solely because the compensation is rolling-off the global feedback factor.

Confirming that this distortion is due solely to the voltage amplifier stage requires varying VAS linearity experimentally while leaving other circuit parameters unchanged. **Fig. 3** shows achieves this by varying the VAS neg-

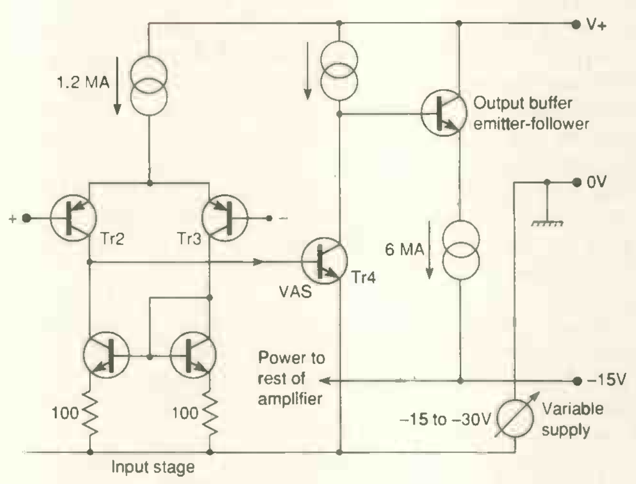


Fig. 3: Voltage amplifier stage distortion test circuit. Although the input pair mirror moves up and down with the VAS emitter, the only significant parameter being varied is the available voltage swing at the collector.

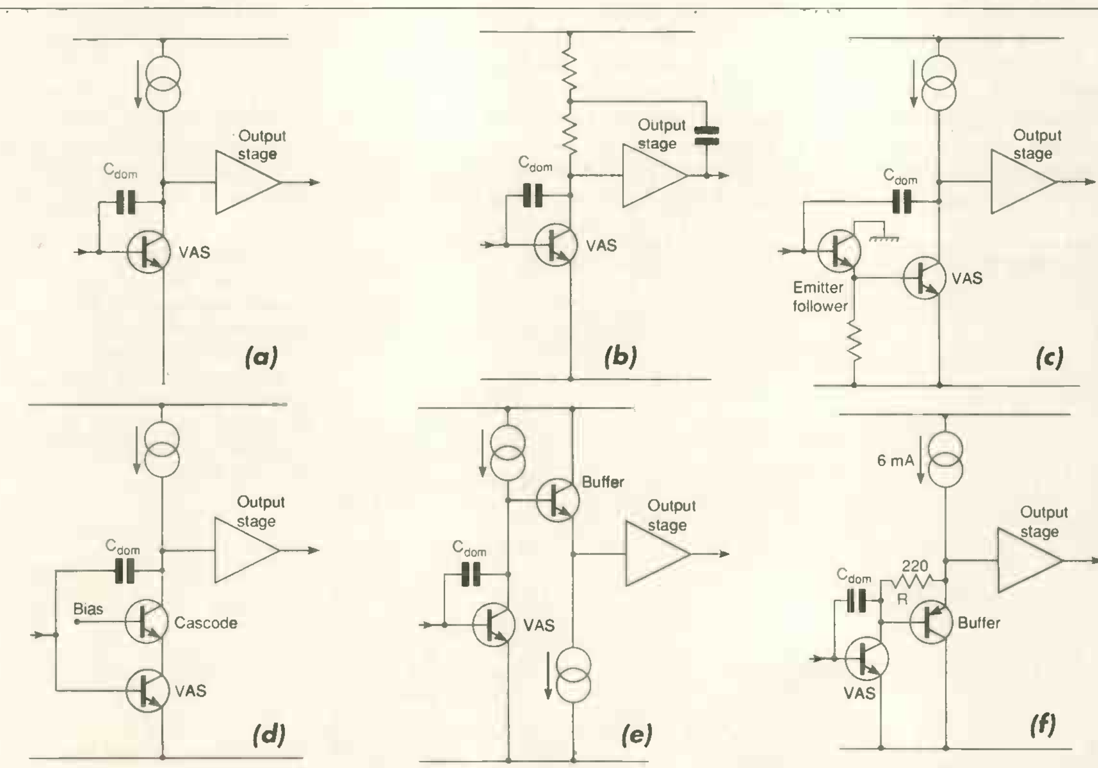


Fig. 4: Six variations on a voltage amplifier stage:
a. conventional current source VAS
b. conventional bootstrapped VAS
c. increase in local NFB by adding emitter follower
d. increase in local NFB by cascoding
e. one method of buffering VAS collector from output stage
f. alternative buffering arrangement uses bootstrapping resistor

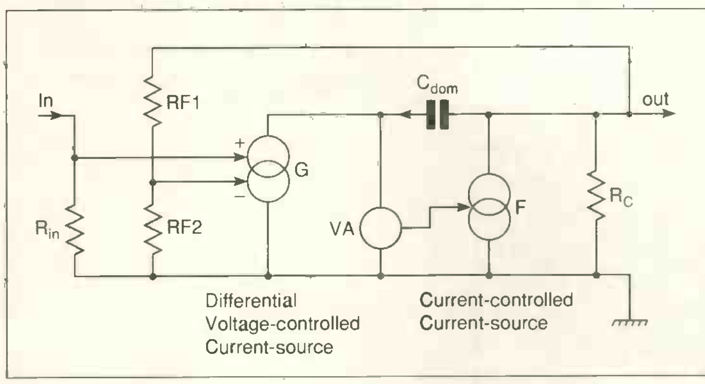


Fig. 5: Conceptual spice model of differential input stage (G) and VAS (F). The current in F is beta times the current in G.

ative rail voltage; this varies the proportion of its characteristic over which the voltage amplifier stage swings, and thus only alters the effective VAS linearity, as the important input stage conditions remain unchanged. The current-mirror must go up and down with the VAS emitter for correct operation, and so the V_{ce} of the input devices also varies, but this has no significant effect as can be proved by the unchanged behaviour on inserting cascode stages in the input transistor collectors.

The typical topology as shown in **Fig. 4a** is a classical common emitter voltage amplifier stage with a current-drive input into the base. The small-signal characteristics, which set open-loop gain and so on, can be usefully simulated by the spice model shown of **Fig. 5**, of a VAS reduced to its conceptual essentials. G is a current source whose value is controlled by the voltage-difference between R_{in} and R_{f2}, and represents the differential transconductance input stage. F represents the voltage amplifier stage transistor, and is a current source yielding a current of beta times that sensed flowing through ammeter V which, by spice convention, is a voltage source set to 0V.

The value of beta, representing current-gain, models the relationship between VAS collector current and base current. R_c represents the total stage collector impedance, a typical real value being 22kΩ. With suitable parameter values, this simple model provides a useful demonstration of relationships between gain, dominant-pole frequency, and input stage current outlined in the first article in this series. Injecting a small signal current into the output node from an extra current source also allows the fall of impedance with frequency to be examined.

The overall voltage gain clearly depends linearly on beta, which in real transistors may vary widely. Working on the trusty engineering principle that what cannot be controlled must be made irrelevant, local shunt NFB through C_{dom} sets the crucial HF gain that controls Nyquist stability. The LF gain below the dominant pole frequency P1 remains variable (and therefore so does P1) but is ultimately of little importance; if there is an adequate NFB factor for overall linearisation at HF then there are unlikely to be problems at LF where gain is highest. As for the input stage, the linearity of the voltage amplifier stage is not greatly affected by transistor type, given a reasonably high beta value.

Stage distortion

voltage amplifier stage distortion arises from a curved transfer characteristic of the common-emitter amplifier, a small portion of an exponential[1]. This characteristic generates predominantly second-harmonic distortion, which, in a closed-loop amplifier, will increase at 6dB/octave with frequency.

Distortion does not get worse for more powerful amplifiers as the stage traverses a constant proportion of its characteristic as the supply-rails are increased. This is not true of the input stage: increasing output swing increases the demands on the transconductance amp as the current to drive C_{dom} increases. The increased V_{ce} of the input devices does not measurably affect their linearity.

It seems ironic that VAS distortion only becomes clearly visible when the input pair is excessively degenerated – a pious intention to 'linearise before applying feedback' can make the closed loop distortion worse by reducing the open loop gain and hence the NFB factor available to linearise the VAS. In a real (non-model) amplifier with a distortive output stage, the deterioration will be worse.

The local open-loop gain of the VAS (that existing inside the local feedback loop closed by C_{dom}) should be high, so that the voltage amplifier stage can be linearised. This precludes a simple resistive load. Increasing the value of R_c will decrease the collector current of the transistor reducing its transconductance. This reduces voltage gain to the starting value.

One way to ensure sufficient gain is to use an active load. Either bootstrapping or a current source will do this effectively, though the current source is perhaps more dependable and is the usual choice for hi-fi or professional amplifiers.

The bootstrap promises more output swing as the collector of Tr_4 can soar above the positive rail. This suits applications such as automotive power amps that must make the best possible use of a restricted supply voltage[2].

These two active-load techniques also ensure enough current to drive the upper half of the output stage in a positive direction right up to the supply rail. If the collector load were a simple resistor, this capability would certainly be lacking.

Checking the effectiveness of these measures is straightforward. The collector impedance may be determined by shunting the collector node to ground with decreasing resistance until the open loop gain falls by 6dB indicating that the collector impedance is equal to the current value of the test resistor.

The popular current source version is shown in Fig. 4a. This works well, though the collector impedance is limited by the effective

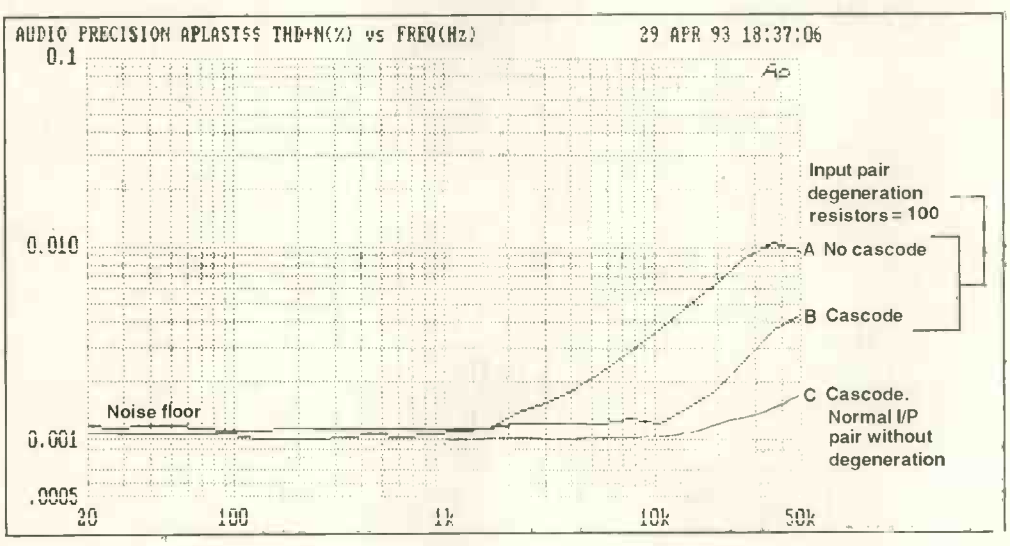


Fig. 6: Showing the reduction of VAS distortion possible by cascoding. The results from adding an emitter follower to the voltage amplifier stage, as an alternative method of increasing local voltage amplifier stage feedback, are very similar.

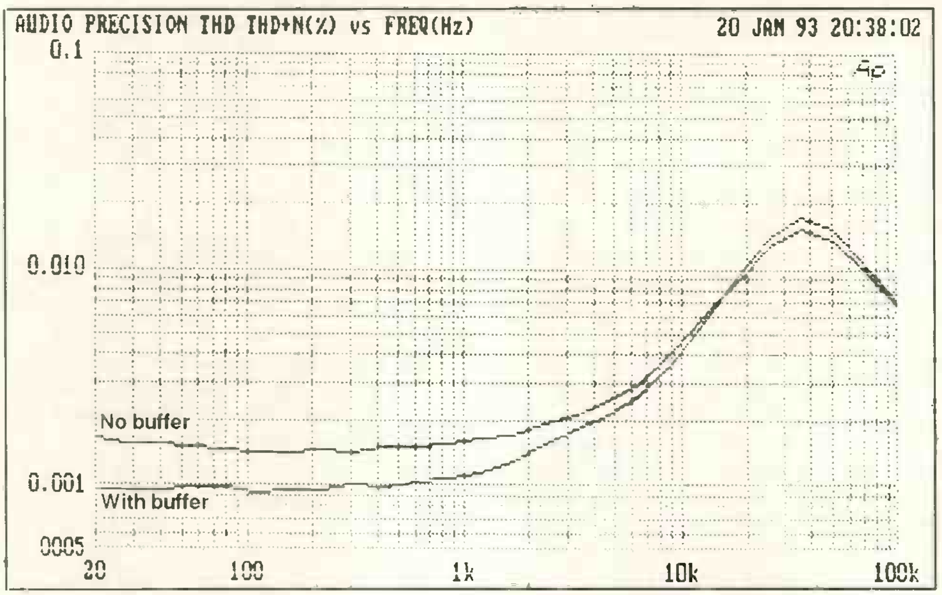
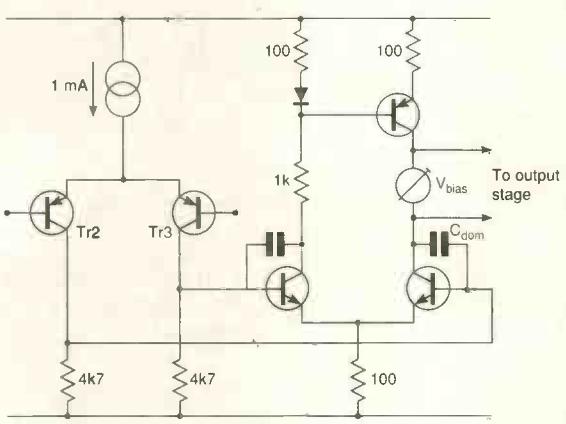


Fig. 7: The beneficial effect of using a VAS buffer in a full scale Class B amplifier. Note that the distortion needs to be low already for the benefit to be significant.



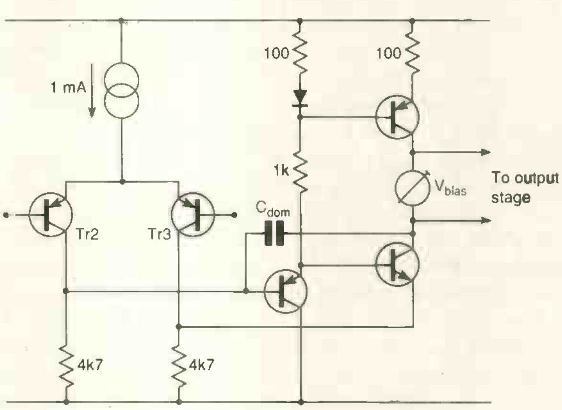


Fig. 8: Two kinds of balanced voltage amplifier stage: Type 1 gives more open loop gain, but no better open loop linearity. Type 2 the circuit originated by Lender.

output resistance R_o of the voltage amplifier stage and the current source transistors[3] which is another way of saying that the improvement is limited by Early effect.

It is often stated that this topology provides current-drive to the output stage; this is only partly true. It is important to realise that once the local NFB loop has been closed by adding C_{dom} the impedance at the VAS output falls at 6dB/octave for frequencies above P1. The impedance is only a few kΩ at 10kHz, and this hardly qualifies as current-drive at all.

Bootstrapping (Fig. 4b) works in most respects as well as a current source load, for all its old-fashioned flavour. The method has been criticised for prolonging recovery from clipping. I have no evidence to offer on this myself, but I can state that a subtle drawback definitely exists: LF open loop gain is dependent on amplifier output loading. The effectiveness of bootstrapping depends crucially on the output stage gain being unity or very close to it. However the presence of the output transistor emitter resistors means that there will be a load-dependant gain loss in the output stage significantly altering the amount by which the VAS collector impedance is increased. Hence the LF feedback factor is dynamically altered by the impedance characteristics of the loudspeaker load and the spectral distribution of the source material.

This has significance if the load is a quality speaker with impedance modulus down to two ohms, in which case the gain loss is serious. If anyone needs a new audio-impairment mechanism to fret about, then I offer this one in the confident belief that its effects, while measurable, are not of audible significance.

The standing dc current also varies with rail voltage. Since accurate setting and maintaining of quiescent current is difficult enough, an extra source of possible variation is decidedly unwelcome.

A less well known but more dependable form of bootstrapping is available if the amplifier incorporates a unity gain buffer between the VAS collector and the output stage as shown in Fig. 4f, where R_c is the collector load, defining the VAS collector current by establishing the V_{be} of the buffer transistor across itself. This is constant, and R_c is therefore bootstrapped and appears to the VAS collector as a constant current source.

In this sort of topology a voltage amplifier stage current of 3mA is quite sufficient, compared with the 6mA standing current in the buffer stage. The voltage amplifier stage would in fact work well with collector currents down to 1mA, but this tends to compromise linearity at the high-frequency, high-voltage corner of the operating envelope, as the VAS collector current is the only source for driving current into C_{dom}.

Voltage stage enhancements.

Fig. 2, which shows only VAS distortion, clearly indicates the need for further improvement over that given inherently by the presence of C_{dom} if an amplifier is to avoid distortion. While the virtuous approach might be an attempt to straighten the curved voltage amplifier stage characteristic, in practice the simplest method is to increase the amount of local negative feedback through this capacitance. Equation 1 in the first article shows that the LF gain (ie the gain before C_{dom} is connected) is the product of input stage transconductance, Tr_4 beta and the collector impedance R_c. The last two factors represent the VAS gain and therefore the amount of local NFB can be augmented by increasing either. Note that so long as the value of C_{dom} remains the same, the global feedback factor at HF is unchanged and so stability is not affected.

The effective beta of the stage can be substantially increased by replacing the VAS transistor with a Darlington, Fig. 4c. Adding an extra stage to a feedback amplifier always requires thought because, if significant additional phase-shift is introduced, the global loop stability may suffer. In this case the new stage is inside the Miller loop and so there is little likelihood of trouble. The function of such an emitter follower is sometimes described as 'buffering the input stage from the VAS' but its true function is linearisation by enhancement of local NFB.

Alternatively the stage collector impedance may be increased for higher local gain. This is could be done with a cascode configuration (Fig. 4d) but the technique is only useful when driving a linear impedance rather than a Class-B output stage with its non-linear input impedance.

Assuming for the moment that this problem is dealt with, either by use of a Class-A output or by VAS-buffering, the drop in distortion is dramatic as is the beta-enhancement method. The gain increase is ultimately limited by Early effect in the cascode and current source transistors, and more seriously by the loading effect of the next stage. But it is of the order of 10 times and gives a useful improvement.

This is shown by curves A, B in **Fig. 6** where the input stage of a model amplifier has been over-degenerated with 100Ω emitter resistors to bring out the voltage amplifier stage distortion more clearly.

Note that in both cases the slope of the distortion increase is 6dB/octave. Curve C shows the result when a standard undegenerated input pair is combined with the cascoded VAS; the distortion is submerged in the noise floor for most of the audio band, being well below 0.001%.

This justifies my assertion that input stage and VAS distortion need not be a problem; we have all but eliminated distortions 1 and 2 from the list of seven given in the first article.

A cascode transistor also allows the use of a high-beta transistor for the voltage amplifier stage; these typically have a limited V_{ceo} that cannot withstand the high rail voltages of a high-power amplifier. There is a small loss of available voltage swing, but only about 300mV, which is usually tolerable. Experiment shows that there is nothing to be gained by cascoding the current source collector load.

A cascode topology is often used to improve frequency response by isolating the upper collector from the C_{bc} of the lower transistor. In this case the frequency response is deliberately defined by a well defined passive component.

It is hard to say which technique is preferable; the emitter follower circuit is slightly simpler than the cascode version, which requires extra bias components, but the cost difference is minimal. When wrestling with these kind of financial decisions it as well to remember that the cost of a small-signal transistor is often less than a fiftieth of that of an output device, and the entire small-signal section of an amplifier usually represents less than 1% of the total cost, when heavy metal such as the mains transformer and heatsinks are included.

Benefits of voltage drive

The fundamentals of linear voltage amplifier stage operation require that the collector impedance is high, and not subject to external perturbations. Thus a Class-B output stage, with large input impedance variations around the crossover point, is the worst possible load. The 'standard' amplifier configuration stands tribute that it can handle this internal unpleasantness gracefully, 100W/8Ω distortion typically degrading only from 0.0008% to 0.0017% at 1kHz assuming that the avoidable distortions have been eliminated. Note however that the effect becomes greater as the global feedback factor is reduced. There is little deterioration at HF, where other distortions dominate[4].

The VAS buffer is most useful when LF distortion is already low, as it removes Distortion 4, which is – or should be – only visible when grosser non-linearities have been seen to. Two equally effective ways of buffering are shown in Figs 4e and 4f.

There are other potential benefits to VAS buffering. The effect of beta mismatches in the output stage halves is minimised[5]. Voltage drive also promises the highest f_T from the output devices, and therefore potentially greater stability, though I have no data of my own to offer on this point. It is right and proper to feel trepidation about inserting another stage in an amplifier with global feedback, but since this is an emitter follower its phaseshift is minimal and it works well in practice.

A VAS buffer put the right way up can implement a form of dc coupled bootstrapping that is electrically very similar to providing the voltage amplifier stage with a separate current-source.

The use of a buffer is essential if a VAS cascode is to do some good. Fig. 7 shows before/after distortion for a full-scale power amplifier with cascode VAS driving 100W into 8Ω.

Balanced voltage amplifier stage

When linearising an amplifier before adding negative feedback one of the few specific recommendations made is usually the use of a balanced voltage amplifier stage – sometimes combined with a double input stage consisting of two differential amplifiers, one complementary to the other. The latter seems to have little to recommend it, as you cannot balance a stage that is already balanced, but a balanced (and, by implication, more linear) voltage amplifier stage has its attractions. However, as explained above, the distortion contribution from a properly-designed VAS is negligible under most circumstances, and therefore there seems to be little to be gained.

Two possible versions are shown in **Fig. 8**; The first type gives approximately 10dB more open loop gain than the standard, but this naturally requires an increase in C_{dom} if the same stability margins are to be maintained. In a model amplifier, any improvement in linearity can be wholly explained by this o/l gain increase, so this seems (not unexpectedly) an unpromising approach. Also, as John Linsley Hood has pointed out[6], the standing current through the bias generator is ill-defined compared with the usual current source VAS. Similarly the balance of the input pair is likely to be poor compared with the current-mirror version. Two signal paths from the input stage to the VAS output must have the same bandwidth; if they do not then a pole-zero doublet is generated in the open-loop gain characteristic that will markedly increase settling-time

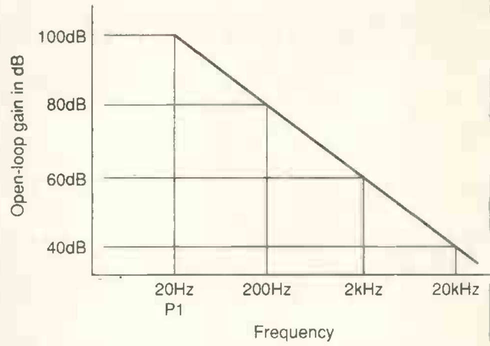


Fig. 9: Showing how dominant pole frequency P1 can be altered by changing the LF open loop gain. The gain at HF, which determines Nyquist stability and HF distortion, is unaffected.

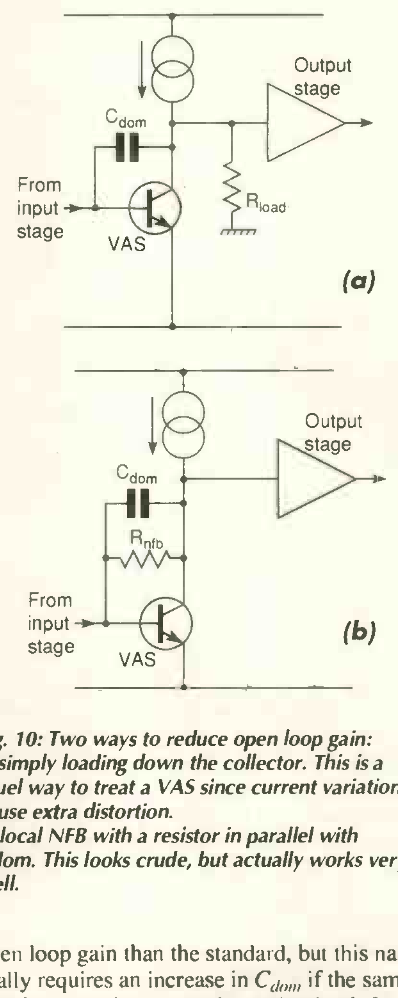


Fig. 10: Two ways to reduce open loop gain:
a. simply loading down the collector. This is a cruel way to treat a VAS since current variations cause extra distortion.
b. local NFB with a resistor in parallel with Cdom. This looks crude, but actually works very well.

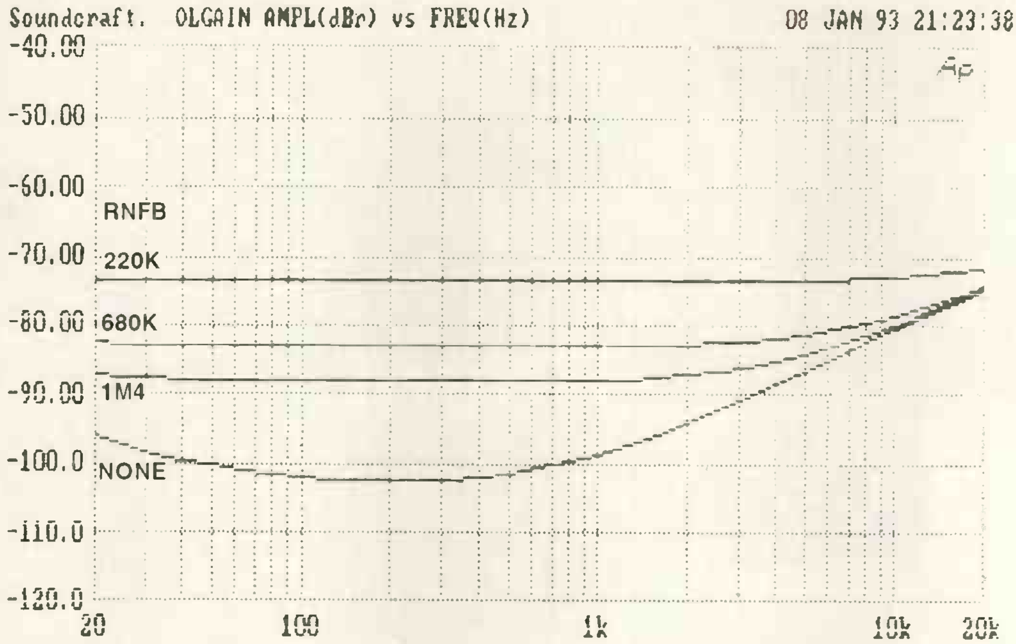


Fig. 11: The result of voltage amplifier stage gain reduction by local feedback; the dominant pole frequency is increased from about 800Hz to about 20kHz, with high frequency gain hardly affected.

after a transient. This seems likely to apply to all balanced voltage amplifier stage configurations.The second type is attributed by Borbely to Lender[7]. Fig. 8 shows one version, with a quasi-balanced drive to the VAS transistor, via both base and emitter. This configuration does not give good balance of the input pair since it depends on the tolerances of R_2,R_3, the V_{be} of the voltage amplifier stage, and so on. Borbely has advocated using two complementary versions of this giving a third type, but this would not seem to overcome the objections and the increase in complexity is significant.

All balanced voltage amplifier stages seem to be open to the objection that the vital balance of the input pair is not guaranteed, and that the current through the bias generator is not well-defined. However one advantage would seem to be the potential for sourcing and sinking large currents into C_{dom}, which might improve the ultimate slew-rate and HF linearity of a very fast amplifier.

Open loop bandwidth.

Acute marketing men will appreciate that reducing the LF open loop gain, leaving HF gain unchanged, must move the P1 frequency upwards, as shown in **Fig. 9**. Open loop gain held constant up to 2kHz appears so much better than open loop bandwidth restricted to 20Hz. These two statements could describe near identical amplifiers, except that the first has plenty of open loop gain at LF while the second has even more. Both amplifiers have the same feedback factor at HF, where the amount available has a direct effect on distortion performance, and could easily have the same slew rate. Nonetheless the second amplifier somehow reads as sluggish and indolent, even when the truth of the matter is known.

Reducing low frequency open loop gain may be of interest to commercial practitioners but it also has its place in the dogma of the subjectivist. Consider it this way: firstly there is no engineering justification for it and, secondly, reducing the NFB factor will reveal more of the output stage distortion. NFB is the only weapon available to deal with this second item so blunting its edge seems ill-advised.

It is of course simple to reduce open loop gain by degenerating the input pair, but this diminishes it at HF as well as LF. To alter it at LF only requires engineering changes at the VAS. **Fig. 10** shows two ways. 10a reduces gain by reducing the value of the collector impedance, having previously raised it with the use of a current source collector load. This is no way to treat a gain stage: loading resistors low enough to have a significant effect cause unwanted current variations in the VAS as well as shunting its high collector impedance, and serious LF distortion appears. While this sort of practice has been advocated in E&WW in the past[8], it seems to have nothing to recommend it. 10b also reduces overall open loop gain by adding a frequency insensitive component to the local shunt feedback around the voltage amplifier stage. The value of R_{NFB} is too high to load the collector significantly and therefore the full gain is available for local feedback at LF, even before C_{dom} comes into action. **Fig. 11** shows the effect on the open loop gain of a model amplifier for several values of R_{NFB}; this plot is in the format described in the first part of this series where error voltage is plotted rather than gain so the curve appears upside down compared with the usual presentation. Note that the dominant pole frequency is increased from 800Hz to above 20kHz by using a 220kΩ value for R_{NFB}; however the gain at higher frequencies is unaffected and so is the stability. Although the amount of feedback available at 1kHz has been decreased by nearly 20dB, the distortion at +16dBu output is only increased from below 0.001% to 0.0013%. Most of the reading is due to noise.

In contrast, reducing the open loop gain by just 10dB through loading the VAS collector to ground requires a load of 4.7kΩ which, under the same conditions, yields distortion of more than 0.01%.

It might seem that the stage which provides all the voltage gain and swing in an amplifier is a prime suspect for generating the major part of its non linearity. In actual fact, this is unlikely to be true, particularly with a cascode VAS/current source collector load buffered from the output stage. Number 2 in the distortion list can usually be forgotten. ■

Next month: the power output stages

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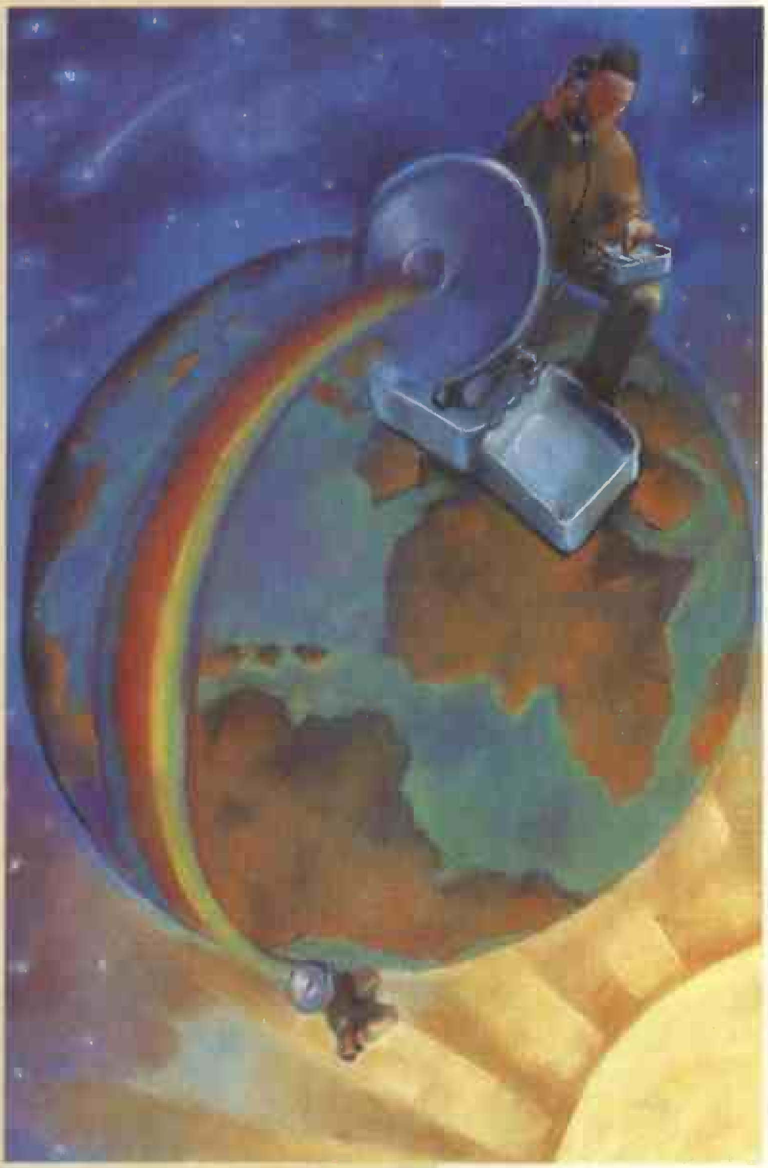


Voice link over spread spectrum radio

2: detailed circuitry

Spread spectrum radio is now used extensively for both military and civilian communications. James Vincent* describes the circuitry for a fully functional experimental voice link.

*James Vincent G1PVZ works for a major aerospace company.

Pseudo-random codes can be categorised as being linear or non-linear codes. Linear codes are generated using linear operations (which for binary pseudo-random codes is solely modulo-2 addition or subtraction). This essentially means only ex-OR gates are used in the shift register feedback path. A pseudo-random generator which does not use such techniques is termed non-linear.

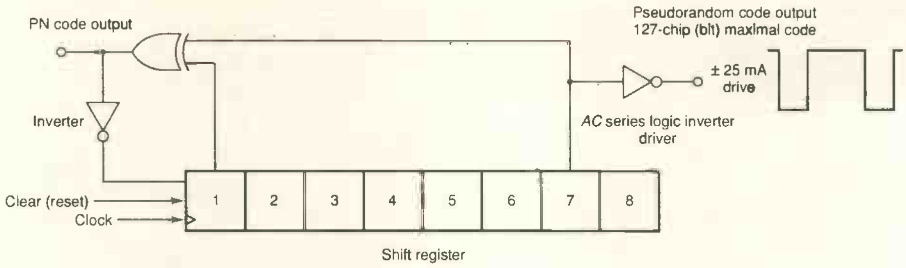
The most commonly used group of pseudo-random sequences used in spread spectrum are the maximal linear code sequences (sometimes called M-sequences or pn – pseudo-noise – codes). Maximal codes are the longest codes that a shift register of specified length can produce and have mathematical properties well suited to spread spectrum communications.

A maximal shift register pseudo-random generator consists of a shift register with selected outputs being exclusive-ORed and fed back into the shift register input. The circuit goes through a number of states (determined by the bits in the shift register at each clock pulse) before it repeats itself after a set number of clock pulses. The maximum number of states for a shift register of length m is 2^m, ie for a 7-stage shift register $2^7 = 128$ states. However the all-zero state is not allowable as the pseudo-random generator would lock-up as ex-ORing two logic 0 results in yet another logic 0 at the input. Therefore a maximal length pseudo-random code generator can produce a pseudo-random sequence 2^m-1 bits long before repeating itself.

To obtain a maximal sequence, the correct shift register outputs (tap points) must be found. These could be found be experimentation but this would be very time consuming! However tables of feedback connections are available[3].

A 7-stage (ie seven flip-flop) shift register can produce a maximal code of length $2^7-1 = 127$ bits (known as *chips* in spread spectrum terminology) long. The feedback tap points may be taken from the following stages:

[7,1] [7,3] [7,3,2,1] [7,4,3,2] [7,6,4,2] [7,6,3,1] [7,6,5,2] [7,6,5,4,2,1] and [7,5,4,3,2,1]



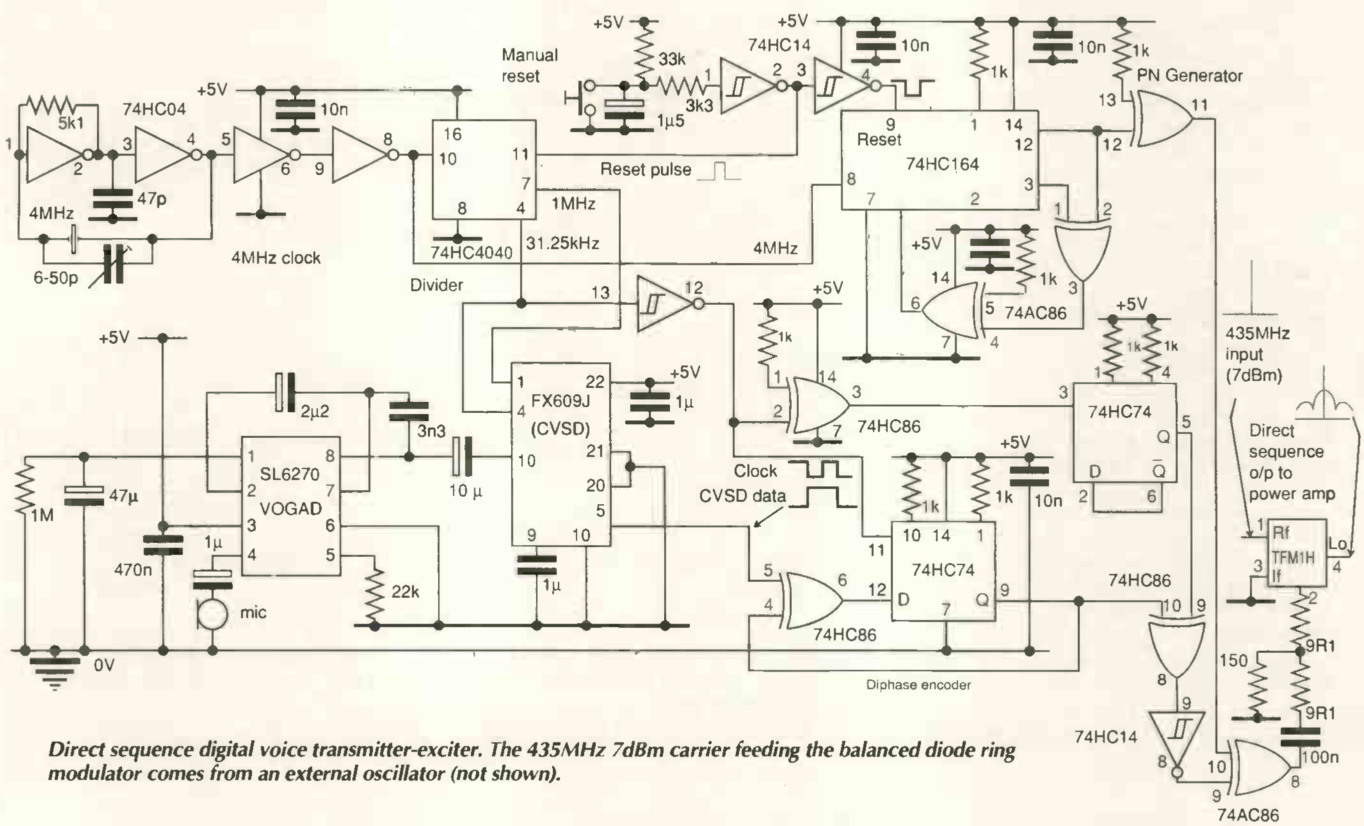
127 chip maximal length pseudorandom code generator

As the simplest circuit implementation is often desired, the first option of tapping the seventh and first stages is selected

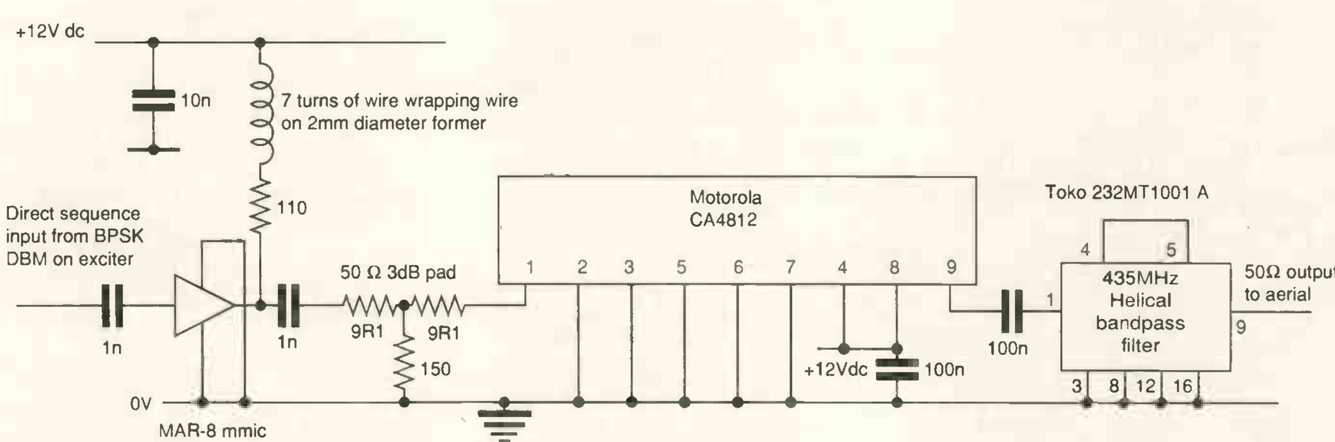
To avoid the all-zero lock up problem, inverting stages are inserted before the shift register input and at the output of the shift register. When the shift register is switched on, a reset pulse is initiated. This pulse initiates all shift register outputs to logic 0. This would normally lock up the pseudo-random sequence generator. However the input inverter injects a logic 1 so that the maximal sequence can commence. The output inverter ensures that maximal code output is inverted negating the effect of the anti-lock-up inverter at the input. The maximal code is also available at the output (A) of the modulo-2 adder, but the second inverter output is normally used to permit direct drive of the DBM in a direct sequence system

Receiver functional description

The 435MHz direct sequence (ds) signal is first amplified by a low noise amplifier[2] followed by a helical filter and further amplification by a low noise amplifier block (*MAN1-LN*) and a *MAR8*. The ds signal is mixed with a 7dBm 365MHz local oscillator in a down-converter. The ds signal now centred on an intermediate frequency of 70MHz is amplified (*MAR8*) and bandpass filtered (*PIF-70*), before being resistively split into three identical signal paths. In each signal path (late, on time or early) the 70MHz signal is amplified by a further *MAR6* amplifier. A DBM configured as a biphase shift keyer is driven by an early, on time or late pn code. The DBM is



Direct sequence digital voice transmitter-exciter. The 435MHz 7dBm carrier feeding the balanced diode ring modulator comes from an external oscillator (not shown).



Power amplifier and driver circuit for direct sequence transmitter.

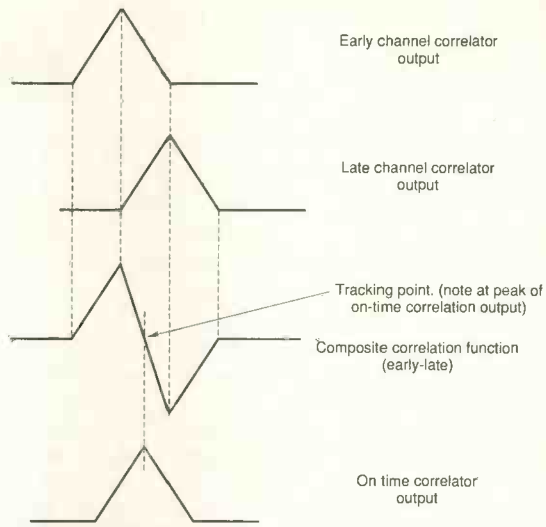
The delay locked loop

After initial acquisition, the spread spectrum receiver must maintain synchronisation by tracking changes in the transmitter's pn code clock. The circuitry required is known as a tracking loop as it tracks the transmitter's code clock frequency variations. Without a tracking loop synchronisation will be lost as the transmitter and receiver pn code clocks drift apart.

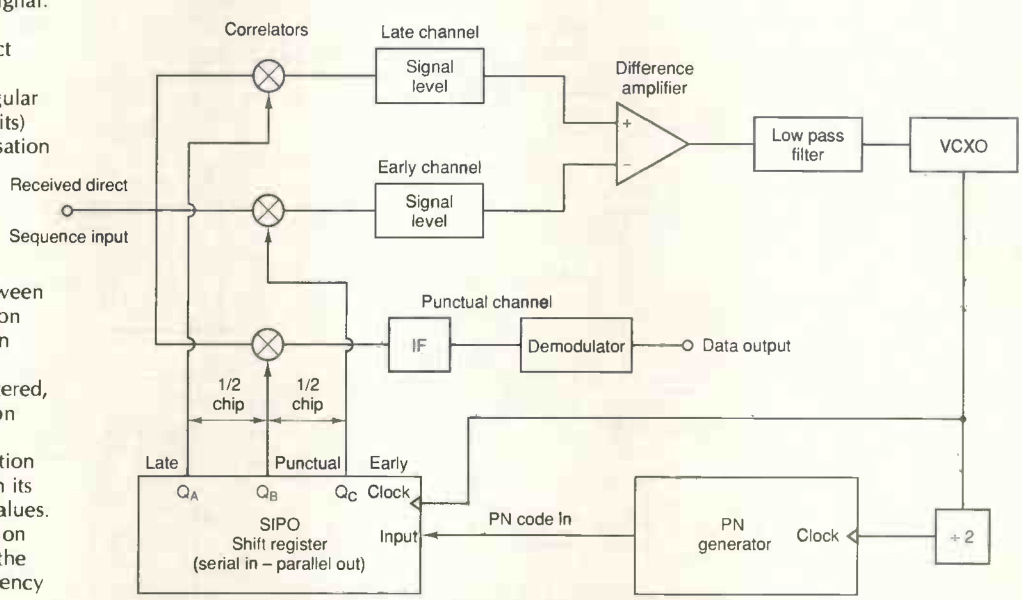
In a delay locked loop two identical pseudo-random or pn despreading codes are delayed with respect to each other. Each pn code is used in separate correlators (early and late) to despread (correlate) the received direct sequence signal. The result of correlation between an incoming direct sequence signal and the receiver pn code is a triangular function two chips (code bits) wide. Assuming synchronisation two correlated signals (each with a triangular correlation waveform) are produced with their correlation peaks separated by the delay between the early and late receiver pn codes. If the two correlation signals are summed in a difference amplifier and filtered, then a composite correlation function is produced. This composite correlation function has a linear region between its maximum and minimum values.

If this composite correlation function is used to control the receiver's code clock frequency (for example by driving a voltage controlled oscillator) then the receiver will track the transmitter's code clock at a point halfway between the maximum and minimum values of the composite correlation function.

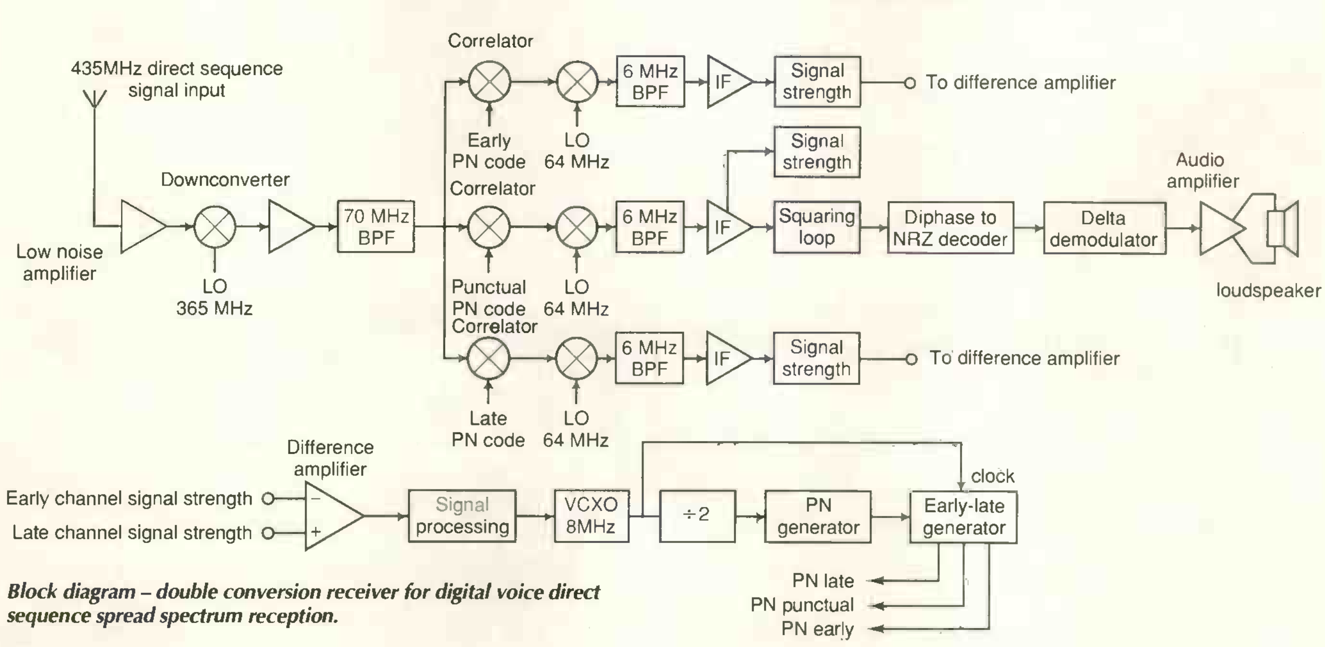
An optimum solution is to have a third on-time (punctual) pn sequence correlator channel for signal recovery, with early and late correlators simply providing tracking to keep the on-time channel in the middle of the correlation window. Such an approach provides an optimally correlated (despread) output signal for subsequent data demodulation.



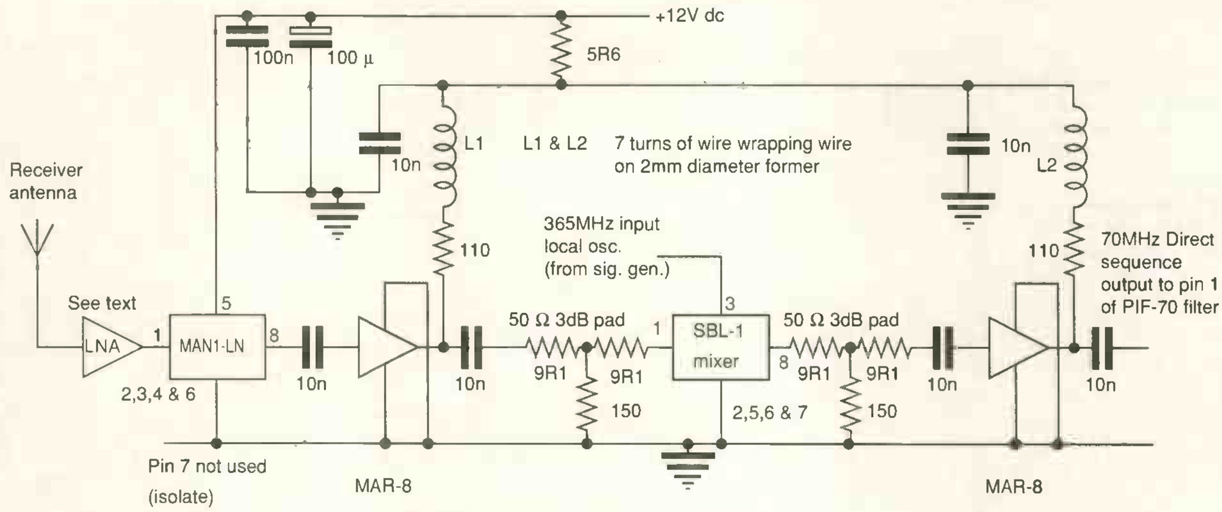
The composite correlation function and correlator waveforms in the 1/2 chip delay locked loop.



The 1/2 chip delay locked loop system. An analogue difference signal derived from the relative correlation levels of advanced and retarded PN code controls a high stability VXCO such that code synchronisation is maintained



Block diagram – double conversion receiver for digital voice direct sequence spread spectrum reception.

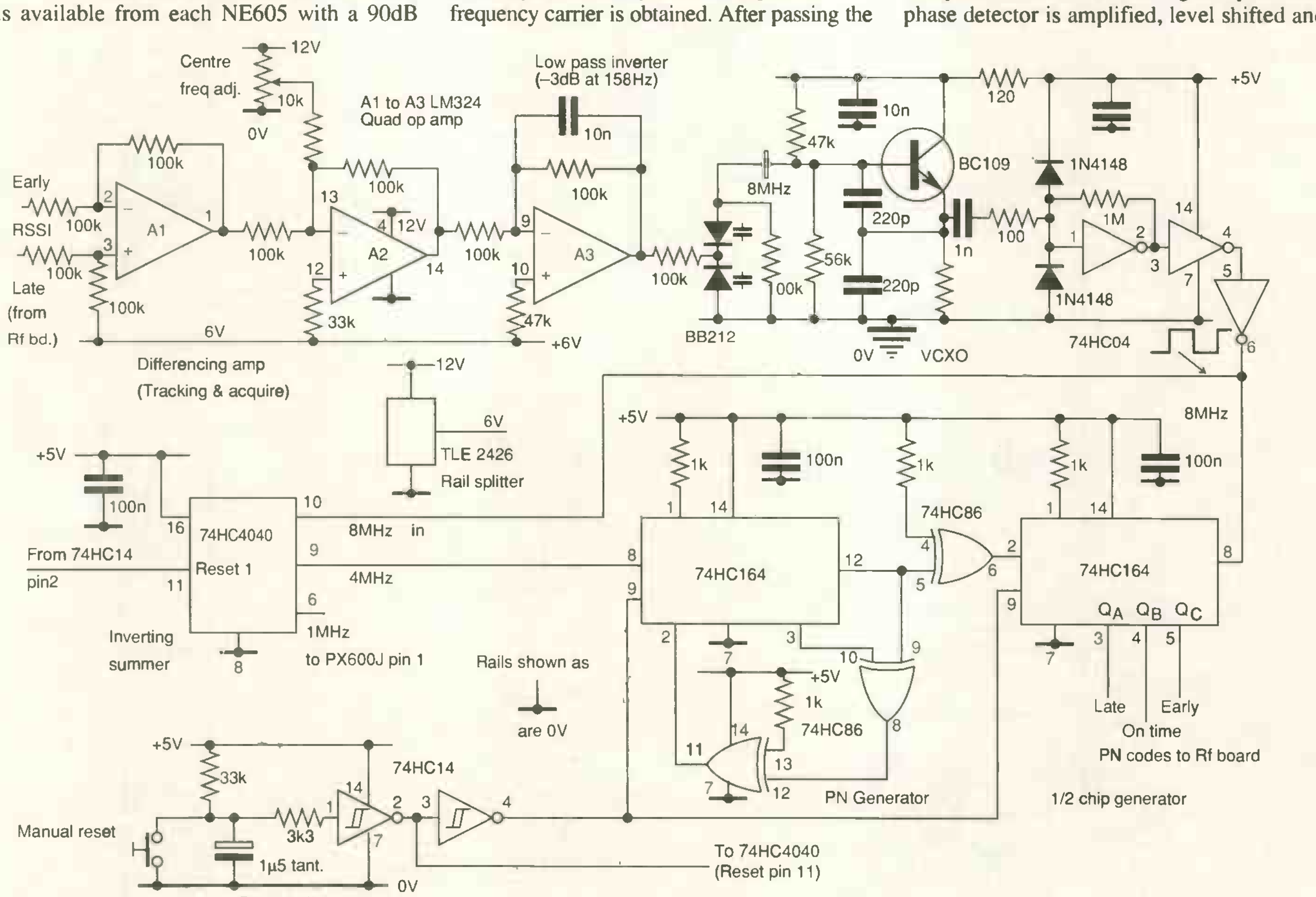


Receiver front end and down-converter. 365MHz local oscillator drive is derived from an external signal generator (not shown).

buffered by 50Ω pads and driven by an AC logic buffer as with the DBM used as a BPSK modulator in the transmitter.

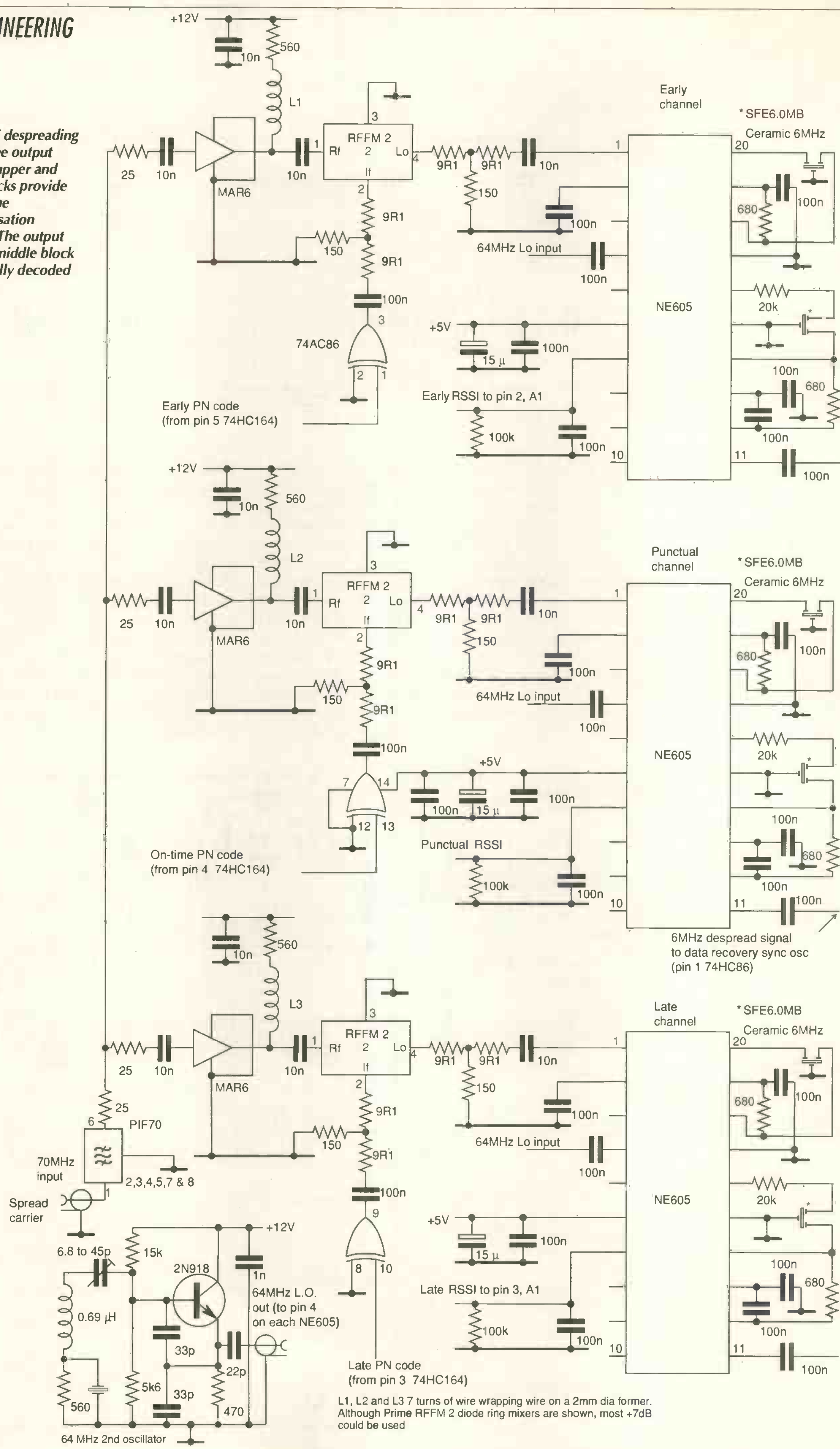
Assuming synchronism the despread output is injected into a *NE605* low power FM IF integrated circuit. The second local oscillator at 64MHz in conjunction with the on-chip mixer downconverts the despread signal (which contains the data in a BPSK format) to 6MHz. The *NE605* further amplifies the 6MHz signal and provides filtering using 6MHz ceramic filters originally designed for television sound strips.

A RSSI (received signal strength indicator) is available from each NE605 with a 90dB range logarithmic output. The RSSI outputs from the early and late channels go to the delay locked loop circuit. The despread data output from the on time (punctual) channel is further amplified by a MAR8 amplifier before being frequency doubled in a Mini Circuits RK3 doubler. As previously discussed the despread data signal has a biphase shift keyed (BPSK) format. The BPSK frequency spectra is similar to that of a double sideband suppressed carrier and as for DSBSC, carrier recovery is required to demodulate the signal. It can be shown mathematically[3] that by squaring or doubling a BPSK signal a twice frequency carrier is obtained. After passing the doubled signal through a 12MHz crystal used as an exceptionally narrow bandpass filter, the signal is applied to a synchronous oscillator. This versatile circuit (see section *The Synchronous Oscillator*) free runs at 6MHz and on application of the 12MHz signal synchronously locks to half of the input frequency, effectively regenerating the 6MHz carrier reference. This locked 6MHz output is buffered and amplified to produce a logic level 0, +5V output, which together with the signal output from the on-time (punctual) *NE605* IC is injected into a DBM configured as a phase detector. The voltage output of the phase detector is amplified, level shifted and

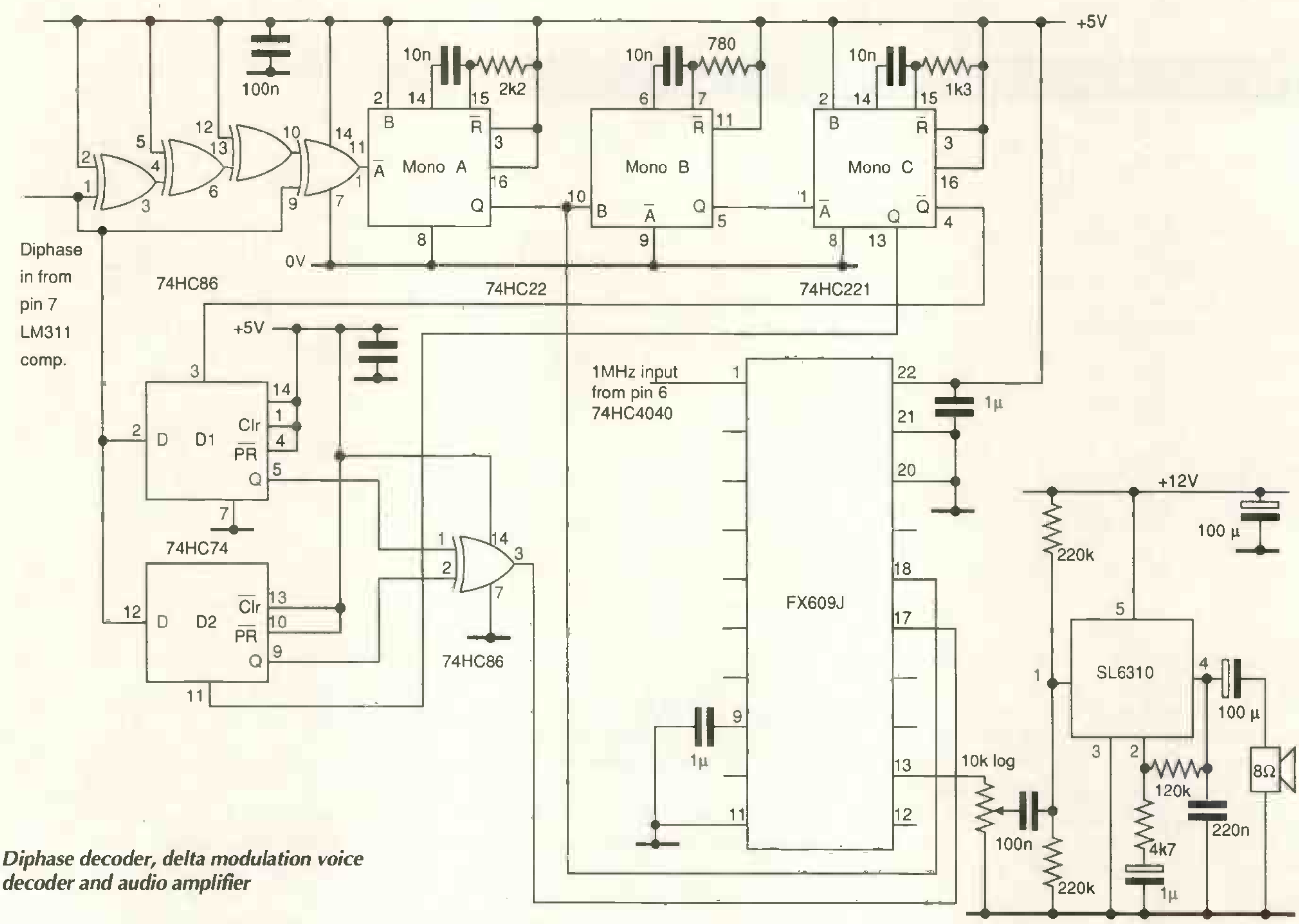


Direct sequence receiver delay lock loop and PN code 1/2 chip generator

70MHz IF despreading circuit. The output from the upper and lower blocks provide data for the synchronisation function. The output from the middle block is eventually decoded to audio



L1, L2 and L3 7 turns of wire wrapping wire on a 2mm dia former. Although Prime RFFM 2 diode ring mixers are shown, most +7dB could be used



Diphase decoder, delta modulation voice decoder and audio amplifier

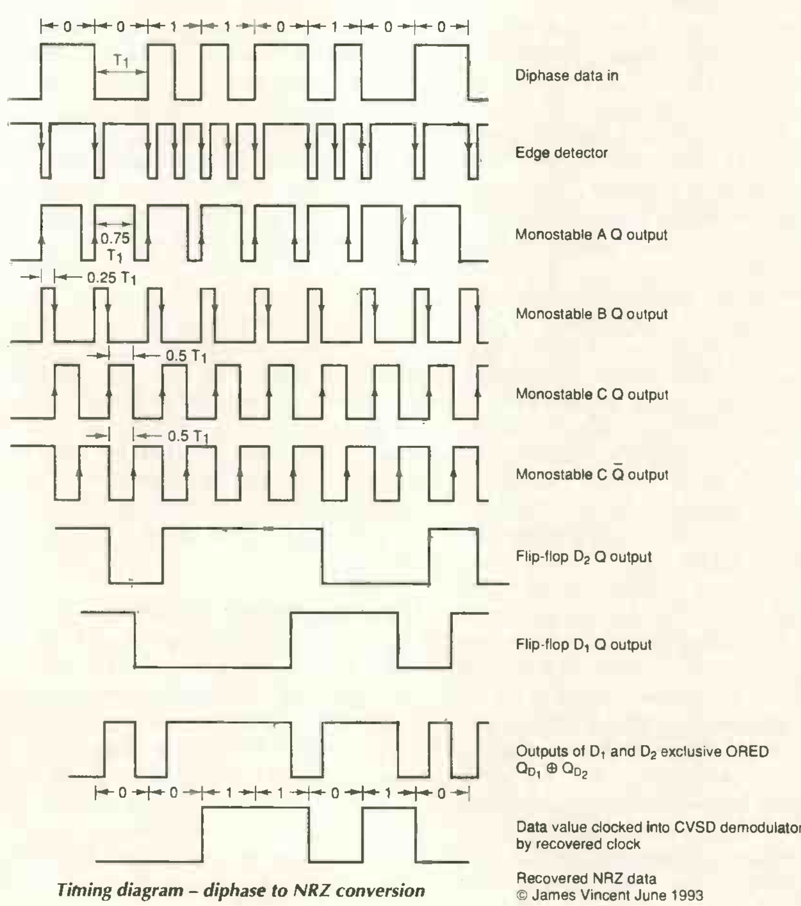
using a voltage comparator converted into standard logic levels.

The output from this squaring loop BPSK demodulator does not recover the original data polarity as the original phase of the signal is lost in the doubling process. This is why the data was diphase encoded at the transmitter so that the correct data polarity could be recovered at the receiver.

An edge detector configured from an exclusive-OR gates produces a negative pulse for both positive and negative edges of the comparator's diphase data stream output. The edge detector output triggers monostable *A*, one half of a dual monostable. (Note: all monostables are non-retriggerable). Monostable *A* is set to produce a positive output pulse with a duration of 75% of the diphase bit cell period. The Q output of monostable A triggers monostable *B* which produces a positive output pulse of duration 25% of the diphase bit cell period.

In turn the negative going edge of monostable *B* output triggers monostable *C* which produces a positive pulse with a duration of 50% of the diphase bit cell period. D-type flip-flop D_1 is clocked by the /Q output of monostable *C* and flip-flop D_2 by the Q output.

The positive edges of the Q and /Q outputs of monostable *C* occur before and after any mid-bit transition. Thus when D_1 and D_2 are clocked, their outputs will be different if the diphase encoded bit represents a one, or the same if the diphase encoded bit represent a



Timing diagram – diphase to NRZ conversion

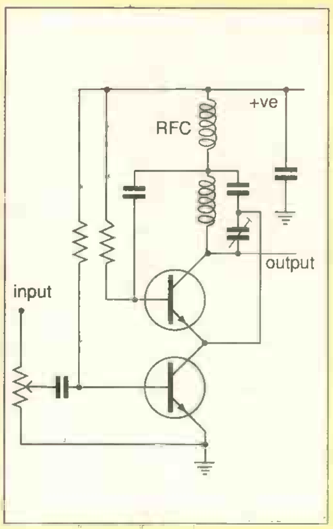
The synchronous oscillator

The synchronous oscillator is an elegant but little known circuit which can be used to advantage where a phase-locked loop (PLL) would normally be employed. The SO is a free-running oscillator which oscillates at a frequency determined by its LC tank with no signal applied to its input.

When a signal is applied within the SO's acquisition bandwidth the oscillator synchronises and tracks the input signal. The SO output amplitude is constant when locked to and tracking an input signal. A decrease in the input carrier-to-noise ratio reduces the SO's tracking bandwidth to maintain a constant signal-to-noise ratio at the SO's output. This characteristic allows a SO to acquire and track very noisy signals.

The SO can also act as a frequency multiplier or divider. In the direct sequence receiver, the SO locks to a noisy 12MHz signal and provides a stable 6MHz output. This function could be achieved using a PLL but the SO has many advantages[5,6,7,8] and, as it is based on only two transistors, is much simpler to implement.

A simplified explanation of operation is that the upper transistor acts as a Class C oscillator. The upper transistor only conducts for a very brief period of time; when the upper transistor conducts, there is a voltage across the lower transistor biasing it allowing it to conduct. At this time the input signal can then be injected to synchronise the oscillator. During the rest of the oscillator cycle input noise is unable to enter the oscillator as the lower transistor is reverse biased. This arrangement produces coherent amplification which is why the SO can extract signals from very low signal-to-noise inputs.



zero. If D_1 and D_2 outputs are exclusive OR-ed then the instantaneous NRZ data is obtained. The clock is recovered at the Q output of monostable *A*. It can be seen that missing or corrupted diphase data could cause monostable *A* to trigger on a mid-bit transition rather than a 'start' transition. This false synchronisation will be corrected on the next diphase encoded zero as monostable *A* will not be triggered.

The recovered clock and NRZ data is delivered to the delta modulator integrated circuit where it is converted back into audio and amplified to a loudspeaker.

The delay lock loop and code generation circuitry permits code correlation, synchronisation and tracking. The difference amplifier has its inverting and non-inverting inputs respectively connected to the early and late channel RSSI outputs. The difference amplifier is followed by a summing amplifier used to adjust the quiescent frequency of the voltage controlled crystal oscillator and a low pass filter. The output of the inverter drives the control input of the voltage controlled oscillator. The VCXO consists of a high stability AT cut crystal in a discrete transistor based oscillator with varicap frequency control. The oscillator's low voltage output is amplified by approximately 10,000 with a linear biased *HC* logic gate. This hard limits the buffer's output to standard logic levels. The VCXO provides a highly stable, repeatable output which has a 2kHz tuning range centred on 8MHz for a tuning voltage of 0 to 6V.

The VCXO output is divided by two to produce a 4MHz clock. This clock signal drives the 127 chip maximal pn generator. The output of this pn generator is re-clocked through a shift register by the original 8MHz clock. By extracting the three outputs from neighbouring outputs three identical pn codes are available (early, on-time and late) but with a half clock cycle difference between them. Thus the early code is one clock cycle (or "chip" in spread spectrum terminology) ahead of the late code. Each pn code generator output drives the relevant correlator (de-spreader). See section *Delay Locked Loop*.

In operation the VCXO is offset to a slightly higher frequency than the crystal clock in the transmitter, effectively producing a sliding correlator. Assuming that the receiver is in range and unsynchronised, the receiver code will slide past the transmitter code. At one point in time the two codes will match. This will result in correlation and the direct sequence signal will be despread. The early channel will be despread before the late channel and the early RSSI value will be considerably higher than the late uncorrelated channel. This difference signal after filtering steers the VCXO output towards the frequency of the transmitter clock. When the receiver and transmitter clocks and pn codes are synchronised the RSSI outputs from the early and late channels will be identical and the difference amplifier output will be zero. Should the receiver clock be retarded greater energy will in the late channel than the early channel, and the VCXO will be driven by the difference amplifier to increase its frequency. If the receiver clock is advanced greater energy will be in the early channel than the late channel and the VCXO will be driven by the difference amplifier to decrease its frequency. Thus the delay locked loop will maintain synchronism once the sliding correlator has caused the receiver to lock. The frequency offset is selected such that it will cause rapid synchronisation but remain within the capture range of the loop.

Construction and Testing

The direct sequence transmitter and receiver were constructed on a combination of Veroboard and double-sided printed circuit boards. The radio-frequency circuits were built on the double-sided pcbs, with the usual RF design techniques employed. The photograph of the completed transmitter- exciter and receiver shows the combination of construction techniques used.(See August *EW+WW*).

The system is designed around easily obtainable components and all inductors and filters are selected from either the Toko or Mini-Circuits range to avoid the difficulties of winding coils.

The set-up of the receiver requires a functioning exciter as a source of a direct sequence signal, hence the exciter is adjusted first. This involves setting the master 4MHz crystal oscillator with the aid of a frequency counter.

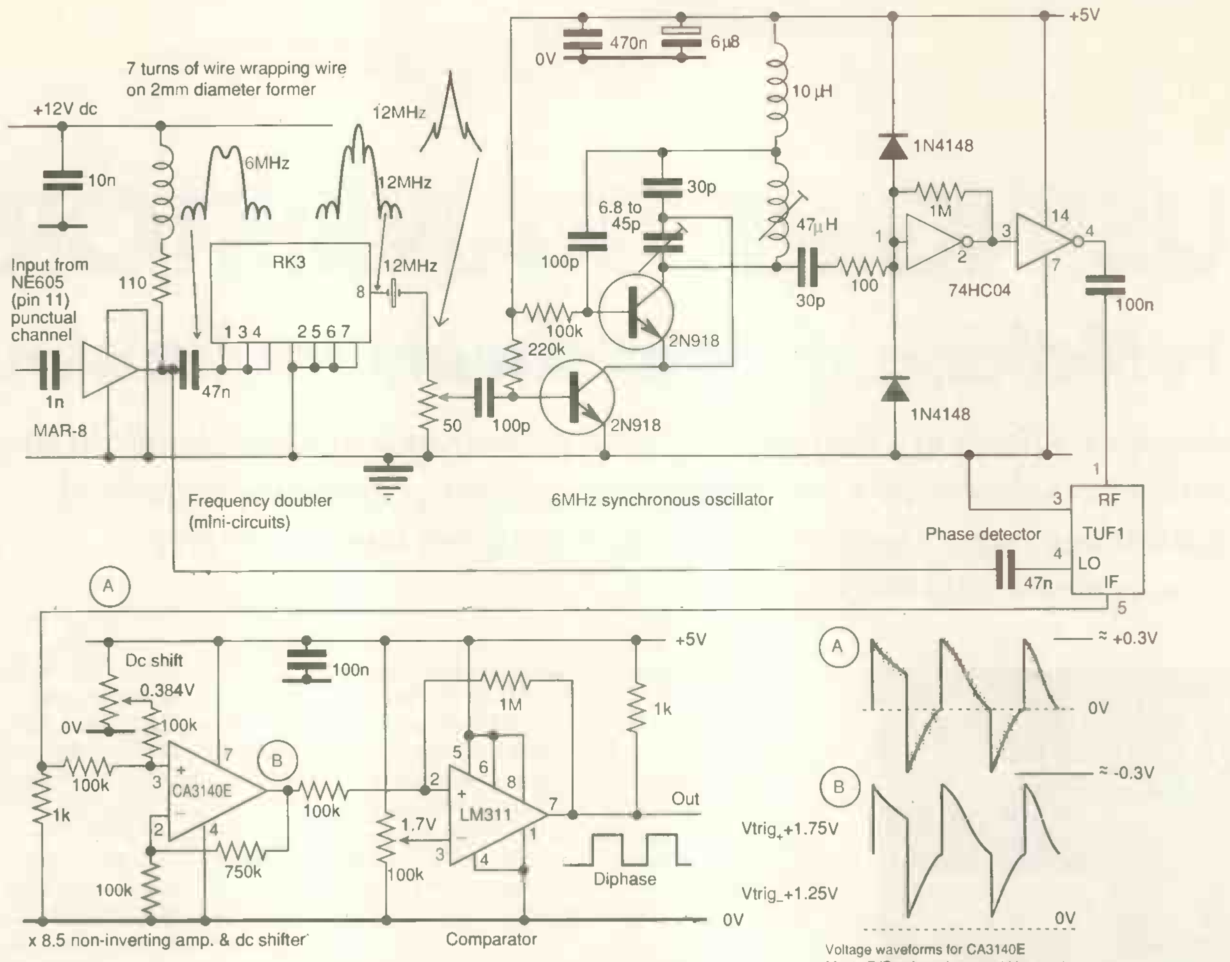
Now the receiver can be directly connected (with a suitable attenuator in-line to prevent overload) to the exciter output. Initially the 64MHz second local oscillator should be adjusted on frequency. The VCXO's frequency is set using the centre frequency adjust potentiometer, to be slightly higher or lower than twice the measured frequency of the transmitter's master clock.

The resonant circuit of the synchronous oscillator (SO) has to be adjusted until it free-runs at 6MHz. It is important to ensure that the SO oscillates at 6MHz (ie not a harmonic) and the input level potentiometer is set to the minimum input level which permits reliable operation.

The gain and comparator reference point potentiometers should be adjusted such that the phase detector recovers the diphase data stream with HC logic compatible levels.

The VCXO frequency is slowly adjusted until the sliding correlator and delay locked loop lock to and track the transmitter. If a spectrum analyser is available, a narrowband despread BPSK data signal will can be detected at the input to the *PIF-70* filter. A dual channel oscilloscope can be used to monitor and compare the transmitter and receiver (punctual) pn codes. If the receiver has synchronised then the two pn codes will line up and the receiver code will be seen to track the transmitter's code. If all is correctly adjusted then the synchronous oscillator will regenerate the 6MHz carrier with the data recovery circuit and delta-modulator i.c. recovering the audio. Various waveforms are shown on the circuit diagram to aid trouble shooting.

After this initial procedure the power and



Squaring loop carrier recovery and diphase data recovery circuits

pre-amplifiers can be added for free-space checks (provided radio regulations permit). Some minor adjustments particularly of the VCXO frequency may be required to ensure reliable acquisition and locking. If the VCXO frequency offset is too great then the receiver will initially acquire the signal, but will be unable to track it. A degree of trial and error may be necessary to arrive at a receiver clock offset which provides rapid synchronisation and reliable tracking performance. The prototype took less than two seconds from power-up to synchronise and would remain in lock provided the signal was not lost.

The Radiocommunications Agency (the UK radio regulatory authority) granted special authority to the author to experiment with spread spectrum techniques on the 70cm band under his amateur radio service licence.

At present the UK amateur radio licence does not permit the use of spread spectrum modulation. It is hoped that in the future the standard UK licence will permit spread spectrum modes of operation as is allowed in the USA by the Federal Communications Commission. ■

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8. United States Patents: 4 335 404, 4 274 067, 4 356 456.

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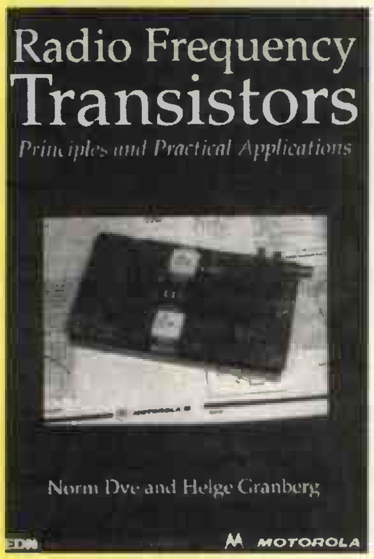
Digital Communication by Satellite. James J. Spilker Jr. Prentice-Hall. ISBN 0-13-214155-8.

TUF1, RK3 and MAR series devices are manufactured by Mini-Circuits (from Cirkit in the UK and Dole Electronics, Camberley) FX609J from Consiumer Microcircuits, Witham, Essex. Crystals are available from IQD Ltd, Crewkerne, Somerset. RFFM2 DBM comes from Walmore Electronics, London.

USING RF TRANSISTORS

1: Making rf data sheets make sense

How does a figure in a data sheet relate to performance in a real circuit? In an extract from their book Radio frequency transistors: principles and practical applications, Norm Dye and Helge Granberg explain how to interpret manufacturers' data sheets.



Norm Dye is Motorola's product planning manager in the Semiconductor Products Sector, and Helge Granberg is Member of Technical Staff, Radio Frequency Power Group (Semiconductor Products) at Motorola. Their rf transistors book includes practical examples from the frequency spectrum from 2MHz to microwaves, with special emphasis on the UHF frequencies .

RF Transistors: Principles and practical applications is available by postal application to room ***L333 EW+WW, Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS.*** Cheques made payable to Reed Books Services. Credit card orders accepted by phone (081 652 3614). ***288pp HARDBACK 07506 9059 3 Cost £19.95 + Postage £2.50***

Data sheets are often the sole guide to a product's capability and characteristics. Circuit designers, unable to talk directly with the factory, must rely on the data sheet for device information. So it is vital that user and manufacturer of rf products speak a common language. The problem is that what the semiconductor manufacturer is saying about an rf device is not always fully appreciated by the circuit designer.

DC specifications

Rf transistors are characterised by two types of parameters: dc and functional.

By definition, "dc" specs consist, of breakdown voltages, leakage currents, h_{FE} (dc beta). and capacitances. Functional specs cover gain, ruggedness, noise figure, Z_{in} and Z_{out}, S-parameters, distortion, etc.

Thermal characteristics do not fall cleanly into either category since thermal resistance and power dissipation can be either dc or ac. So thermal resistance is best treated as a special specification.

Table 1 is from a typical rf power data sheet showing dc and functions and specs. A critical part of selecting a transistor is choosing one that has *breakdown voltages* compatible with the supply voltage of an intended application. The design engineer must select a transistor that, on the one hand, has breakdown voltages which will not be exceeded by the dc and rf voltages appearing across the various junctions of the transistor; and on the other, has breakdown voltages permitting the "gain at frequency" objectives to be met by the transistor.

Mobile radios normally operate from a 12V source and portable radios use a lower voltage, typically 6 to 9V. Avionics applications are commonly 28V, while base station and other ground applications such as medical electronics generally take advantage of the superior performance characteristics of high voltage devices, operating with 24-50V supplies.

In making a transistor, breakdown voltages are largely determined by material resistivity and junction depths (**Fig. 1**), so breakdown voltages are intimately entwined with functional performance characteristics. Most product portfolios in the rf power transistor industry have families of transistors designed for use at specified supply voltages such as 7.5V, 12.5 volts, 28, and 50V.

Leakage currents (defined as reverse biased junction currents occurring prior to avalanche breakdown) are likely to be more varied in their specification – and also more informative.

Many transistors do not have leakage currents specified because they can result in excessive, and frequently unnecessary, wafer/die yield losses. Leakage currents arise as a result of material defects, mask imperfections, and/or undesired impurities entering wafer processing. Some sources of leakage currents are potential reliability problems: most are not. Leakage currents can be material-related such as stacking faults and dislocations, or can be "pipes" created by mask defects and/or processing inadequacies. These sources result in leakage currents that are constant with time. If initially acceptable for a particular application, they will remain so, and do not pose long term reliability problems.

But channels induced by mobile ionic contaminants in the oxide (primarily sodium) cause leakage currents that tend to change with time and can lead to increases that render the device useless for a specific application. Distinguishing sources of leakage current can be difficult, which is one reason devices for application in military environments require HTRB (high temperature reverse bias) and burn-in testing. But even for commercial applications – particularly where battery drain is critical or where bias considerations dictate limitations – a leakage current limit should be included in any complete device specification.

Dc parameters such as h_{FE} and C_{ob} (output capacitance) need little comment. Typically, for rf devices, h_{FE} is relatively unimportant for unbiased power transistors because the functional parameter of gain at the desired frequency of operation is specified. Note, though, that dc beta is related to ac beta **(Fig. 2)**. Functional gain will track dc beta particularly at lower rf frequencies. An h_{FE} specification is needed for transistors requiring bias, which includes most small signal devices normally operated in a linear (class A) mode. Generally rf device manufacturers do not like to have tight limits placed on h_{FE}, primarily because:

- There is a lack of correlation with rf performance;
- difficulty in controlling wafer processing;
- other device manufacturing constraints dictated by functional performance specs which preclude tight limits for h_{FE}.

A good rule of thumb for h_{FE} is to set a maximum-to-minimum ratio of not less than 3 and not more than 4, with minimum h_{FE} value determined by an acceptable margin in functional gain. Output capacitance is an excellent measure of comparison of device size (base area) provided most of the output capacitance is created by the base-collector junction and not parasitic capacitance arising from bond pads and other top metal of the die. Remember that junction capacitance will vary with voltage **(Fig. 3)** while parasitic capacitance will not. Also, in comparing devices, the voltage at which a given capacitance is specified should be noted, as no industry standard exists. The preferred voltage at Motorola is the transistor V_{cc} rating, ie 12.5V for 12.V transistors and 28V for 28 volt transistors.

Ratings and thermal characteristics

Maximum ratings **(Table 2)** tend to be the most frequently misunderstood group of device specifications. Ratings for maximum junction voltages are straightforward, and simply reflect the minimum values set forth in the dc specs for breakdown voltages. If a device meets the specified minimum breakdown voltages, then voltages less than this will not cause junctions to reach reverse bias breakdown and the potentially destructive current levels.

The value of BV_{CEO} is sometimes misinterpreted. Its value can approach or even equal the supply voltage rating of the transistor, and the question naturally arises as to how such a low voltage can be used in practical applications. First, BV_{CEO} is the breakdown voltage of the collector-base junction, plus the forward drop across the base-emitter junction with the base open. It is never encountered in amplifiers where the base is at or near the potential of the emitter. That is, most amplifiers have the base shorted or use a low value of resistance such that the breakdown value of interest approaches BV_{CES}.

Second, BV_{CEO} involves the current gain of the transistor and increases as frequency increases. So the value of BV_{CEO} at rf frequencies is always greater than the value at dc.

Maximum rating for power dissipation (P_d) is closely associated with thermal resistance (θ_{JC}). In reality, maximum P_d is a fictitious number because it assumes that case temperature is maintained at 25°C. But, providing everyone arrives at the value in a similar manner, the rating of maximum P_d is also a useful comparative.

Even so, several reasons dictate a conservative value be placed on θ_{JC}. Thermal resistance increases with temperature, and die temperature T_j is not a worst case number. Also, by using a conservative value of θ_{JC}, a realistic value is determined for maximum P_d. Generally, Motorola's practice is to publish θ_{JC} numbers approximately 25% higher than that determined

Table 1. Typical dc and functional specifications from an rf power data sheet.

	Symbol	Min	Typ	Max	Unit
Off characteristics:					
Collector-emitter breakdown voltage (I_C = 20mA dc. I_B = 0)	$V_{(BR)CEO}$	16	–	–	Vdc
Collector-emitter breakdown voltage (I_C = 20mA dc. V_B = 0)	$V_{(BR)CES}$	36	–	–	Vdc
Emitter-base breakdown voltage (I_E = 5.0mA dc. I_C = 0)	$V_{(BR)EBO}$	4.0	–	–	Vdc
Collector cut-off current (V_{CE} = 15V dc, V_{BE} = 0, T_C = 25°C)	I_{CES}	–	–	10	mAdc
On characteristics:					
DC current gain (I_C = 4.0A dc, V_{CE} = 5.0V dc)	h_{FE}	20	70	150	–
Dynamic characteristics:					
Output capacitance (V_{CB} = 12.5V dc, I_E = 0, f = 1.0MHz)	C_{ob}	–	90	125	pF
Functional tests:					
Common-emitter amplifier power gain (V_{CC} = 12.5V dc, P_{out} = 45W, $I_C(max)$ = 5.8A dc, f = 470MHz)	G_{pe}	4.8	5.4	–	dB
Input power (V_{CC} = 12.5V dc, P_{out} = 45W, f = 470MHz)	P_{in}	–	13	15	watts
Collector efficiency (V_{CC} = 12.5V dc, P_{out} = 45W, $I_C(max)$ = 5.8A dc, f = 470MHz)	η	55	60	–	%
Load mismatch stress (V_{CC} = 16V dc, P_{in} = Note 1, f = 470MHz, VSWR = 20:1, all phase angles	ψ^*	No degradation in output power			
Series equivalent input impedance (V_{CC} = 12.5V dc, P_{out} =45W, f = 470MHz)	Z_{in}	–	1.4 + j4.0	–	ohms
Series equivalent output impedance (V_{CC} = 12.5V dc, P_{out} =45W, f = 470MHz)	Z_{OL}	–	1.2 + j2.8	–	ohms

Notes:
1. P_{in} = 150% of drive requirement for 45W output @ 12.5V.
ψ^* = Mismatch stress factor – the electrical criterion established to verify the device resistance to load mismatch failure. The mismatch stress test is accomplished in the standard test fixture terminated in a 20:1 minimum load mismatch at all phase angles.

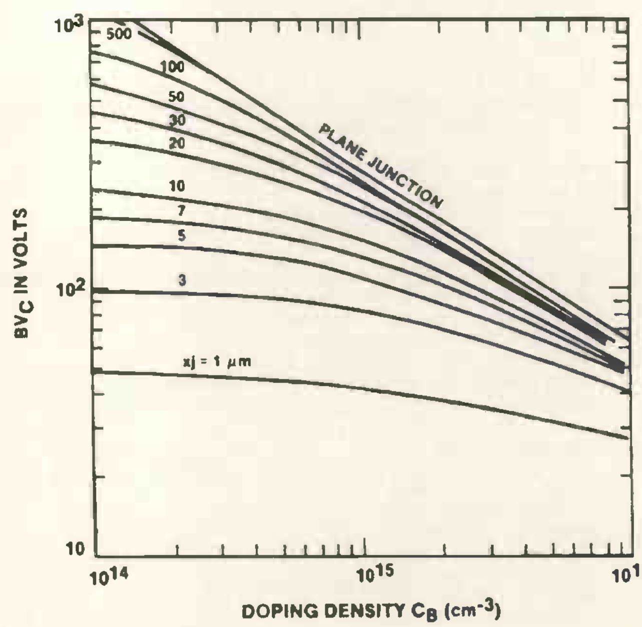


Fig. 1. Curvature and resistivity determine breakdown voltage.

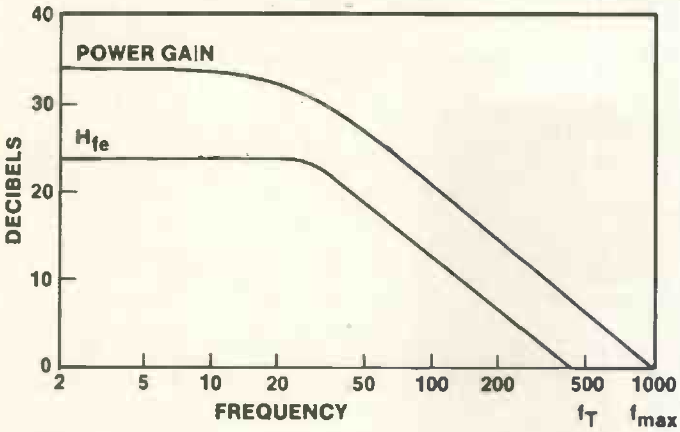


Fig. 2. Dc beta is related to ac beta.

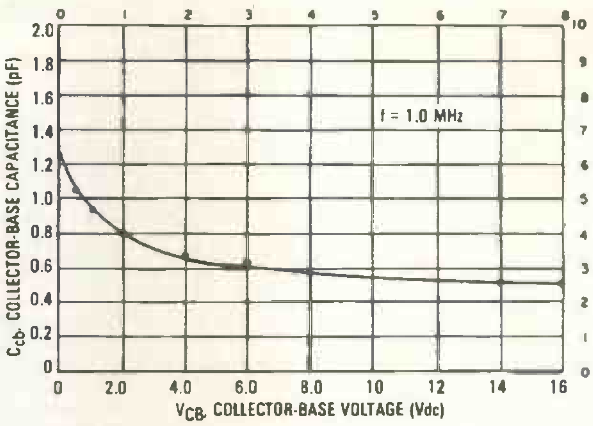


Fig. 3. Junction capacitance varies with voltage.

Now onto die temperature. Reliability considerations dictate a safe value for an all-Au (gold) system (die top metal and wire) to be 200°C. Once $T_j max$ is determined, along with a value for θ_{JC}, maximum P_d is simply:

$$P_d(max) = (T_j(max) - 25°C)/\theta_{JC}$$

Specifying maximum P_d for T_c = 25°C leads to the necessity to derate maximum P_d for any value of T_c above 25°C. The derating factor is simply the reciprocal of θ_{JC}!

Maximum collector current I_c is probably the most subjective maximum rating on the transistor data sheets. Different methods of determination lead to different maximum ratios. The only valid maximum current limitations in an rf transistor concern the current handling ability of the wires or the die. But power dissipation ratings may restrict current to values far below (what should be) the maximum rating. Unfortunately, many older transistors had their maximum current rating determined by dividing maximum P_d by collector voltage (or by BV_{CEO} for added safety). But this is not a fundamental maximum current limitation of the part. Many lower frequency parts have relatively gross top metal on the transistor die – ie wide metal runners – and the "weak current link" in the part is the current handling capability of the emitter wires (for common-emitter parts). Current handling ability of wire is well known, so maximum current rating may be limited by the number, size and material used for emitter wires.

Most modern high-frequency transistors are die-limited because of high current densities resulting from very small current carrying conductors. These densities can lead to metal migration and premature failure.

It is up to the transistor manufacturer to specify an $I_c max$ based on whichever of the two limitations, die or wire, is paramount. Circuit design engineers should consult semiconductor manufacturers for additional information if $I_c max$ is of any concern.

Storage temperature is another maximum rating that is frequently not given the attention it deserves. A range of –55°C to 200°C has become more or less standard. For the single-metal, hermetic-packaged type of device, 200°C creates no reliability problems. But a lower high-temperature limitation exists for plastic encapsulated or epoxy-sealed devices which should not be subjected to >150°C to prevent deterioration of the plastic material.

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Maximum ratings:

	Symbol	Value	Unit
Collector-emitter voltage	V_{CEO}	16.5	Vdc
Collector-emitter voltage	V_{CES}	38	Vdc
Emitter-base voltage	V_{EBO}	4.0	Vdc
Collector-current – continuous	I_C	12	Adc
Total device dissipation @ T_C 25°C	P_D	135	W
Derate above 25°C		0.77	W°C
Storage temperature range	T_{stg}	–65 to +150	°C

Thermal characteristics

	Symbol	Max	Unit
Thermal resistance, junction to case	$R_{\theta JC}$	1.3	°C/W

Table 2. Maximum power ratings of a typical rf power transistor, the Motorola MRF650.

Power transistor characteristics

Selection of a power transistor usually depends on frequency of operation, output power, desired gain, voltage of operation and preferred package configuration consistent with circuit construction techniques.

By necessity, functional characteristics of an rf power transistor are tied to a specific test circuit. Without specifying a circuit, the functional parameters of gain, reflected power, efficiency and even ruggedness, hold little meaning. Furthermore, most test circuits used by rf transistor manufacturers (even those used to characterise devices) are designed mechanically to allow for easy insertion and removal of the device under test (DUT). This mechanical restriction sometimes limits achievable device performance, explaining why performance by users frequently exceeds that indicated in data sheet curves.

Conversely, a circuit used to characterise a device is usually narrow band and tunable, resulting in higher gain than attainable in a broadband circuit. Unless otherwise stated, characterisation data such as P_O vs frequency can be assumed to be generated on a point-by-point basis by tuning a narrow band circuit across a band of frequencies. As such it represents what can be achieved at a specific frequency of interest provided the circuit presents optimum source and load impedances to the DUT.

Broad-band, fixed tuned test circuits are the most desirable for testing functional performance of an rf transistor. Fixed-tuned is particularly important in assuring everyone – manufacturer and user – of product consistency on a day to day basis.

Tunable, narrow-band circuits have led to device users and manufacturers resorting to the use of "correlation units" to assure product consistency over time. Fixed tuned circuits minimise (if not eliminate) the requirements for correlation. In so doing, they tend to compensate for the increased constraints placed on the device manufacturer. Against this, manufacturers prefer tunable test circuits because they allow adjustments to compensate for variations in die fabrication and/or device assembly. Unfortunately, gain is normally less in a broad-band circuit than in a narrow-band circuit, and this fact frequently forces transistor manufacturers to use narrow-band circuits to make their products more attractive when compared with competitors (specsmanship!). One compromise the transistor manufacturer can make is to use narrow-band circuits with all tuning adjustments "locked" in place.

For all these reasons, when comparing devices, the data sheet reader should observe carefully the test circuit in which specific parameter limits are guaranteed.

Ruggedness

RF power transistors give considerable weight to ruggedness.

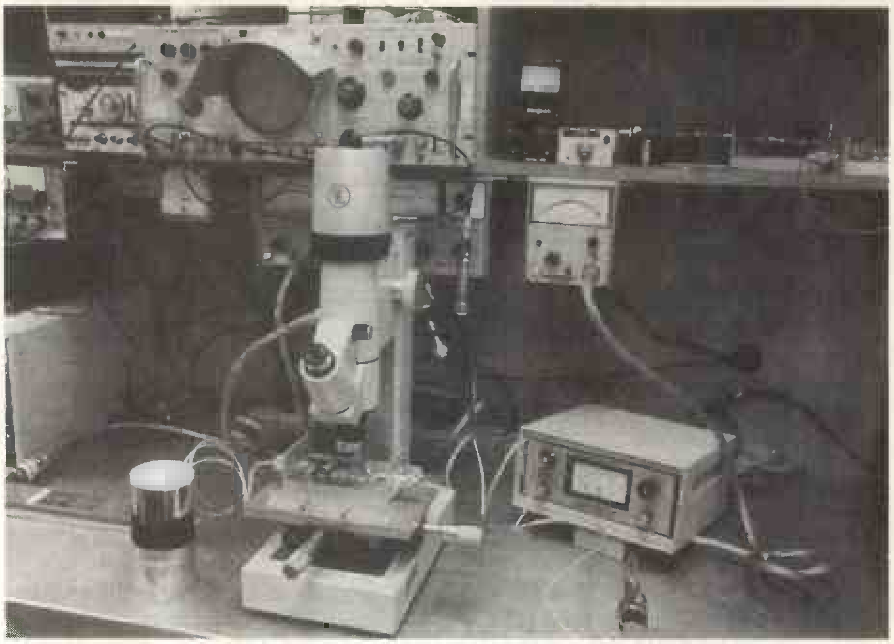
Ruggedness is the characteristic of a transistor to withstand extreme mismatch conditions in operation – causing large amounts of output power to be "dumped back" into the transistor – without altering performance capability or reliability.

Many circuit environments, particularly portable and mobile radios, have limited control over the impedance presented to the power amplifier by an antenna (at least for some duration of time). In portables, the antenna may be placed against a metal surface; in mobiles, perhaps the antenna is broken off or inadvertently disconnected from the radio.

RF power transistors must be able to survive such load mismatches without effect on subsequent operation. One realistic possibility for mobile radio transistors, though not a normal situation, is where an RF power device "sees" a worst case load mismatch (an open circuit, any phase angle) along with maximum V_{cc} and greater than normal input drive – all at the same time.

So the ultimate test for ruggedness is to subject a transistor to a P_{in} (rf) 50% above that necessary to create rated P_O; increase V_{cc} by about 25% (12.5V to 16V for mobile transistors); and then set the load reflection coefficient at a unity while its phase angle is varied through all possible values from 0-360°.

Ruggedness specifications come in many forms. Older devices (and even some newer ones) simply have no ruggedness specification. Others are said to be "capable of" withstanding load mismatches, while still others are guaranteed to withstand load mismatches of 2:1 VSWR, to infinity at rated output power. A few truly rugged transistors are guaranteed to withstand 30:1 VSWR at all phase angles with both overvoltage and overdrive: for all practical purposes 30:1 VSWR is the same as infinity VSWR.



Die temperature is measured using an infra-red microscope.

Again, the user must match circuit requirements against device specifications.

But, confusing the whole issue even further is the way semiconductor manufactures define what constitutes passing the ruggedness test, generally saying that after testing the DUT "shall have no degradation in output power." A better phrase would be "no measurable change in output power", though even this is not perfect. Unfortunately, the DUT can be damaged by the ruggedness test and still have "no degradation in output power."

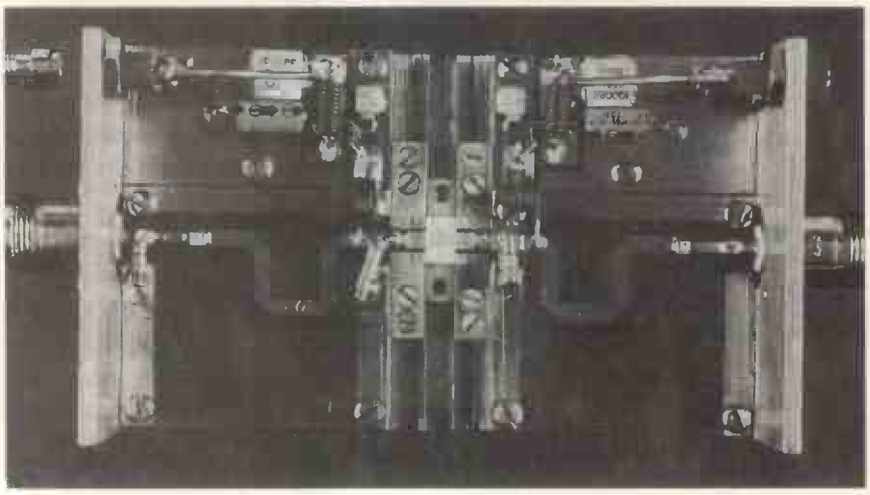
RF power transistors now consist of up to 1000 or more low power transistors connected in parallel. Emitter resistors are placed in series with groups of these transistors to give best control of power sharing throughout the transistor die.

Manufacturers know that a high percentage of an rf power transistor die (25-30%) can be destroyed with the transistor still able to deliver rated power at rated gain – at least for a while. If a ruggedness test destroys a high percentage of cells in a transistor, then a second ruggedness test (by manufacturer or user while in circuit) would probably result in additional damage and premature device failure.

A more scientific measurement of passing or failing a ruggedness test is called ΔV_{re}, the change in emitter resistance before and after the ruggedness test. V_{re} is determined to a large extent by the net value of emitter resistance in the transistor die. So if cells are destroyed, emitter resistance will change with a resultant change in V_{re}.

Changes as small as 1% are readily

Physical test jig for an rf power transistor.



detectable, with 5% or less normally considered an acceptable limit.

More advanced device specifications for rf power transistors use this criteria to determine success or failure in ruggedness testing.

Matching circuit

A circuit designer must know the input/output characteristics of the rf power transistor(s) selected, to design a circuit that matches the transistor over the frequency band of operation. Data sheets provide this information in the form of large signal impedance parameters, Z_{in} and Z_{out} (commonly referred to as Z^*_{OL}). Normally, these are stated as a function of frequency and are plotted on a Smith chart and/or given in tabular form.

Test station for rf power transistors used by Motorola.

Z_{in} and Z_{out} apply only for a specified set of operating conditions, – power output, voltage and frequency – and are determined in a similar way: place the DUT in a tunable circuit and tune both input and output circuit elements to achieve maximum gain for the desired set of operating conditions. At maximum gain, DUT impedances will be the conjugate of the input and output network impedances. So terminate the input and output ports of the test circuit, remove the device and measure Z looking from the device – first, toward the input to obtain the conjugate of Z_{in} and, second, toward the output to obtain Z_{OL}, the output load required to achieve maximum P_O.

A network analyser is used to determine the complex reflection coefficient of the circuit using, typically, the edge of the package as a plane of reference .

Once Z_{in} and Z_{OL} of the transistor are known as a function of frequency, cad can be used to design L and C matching networks for a particular application.

The complete impedance measuring process is somewhat time-consuming since it must be repeated for each frequency of interest: note that the frequency range permitted for characterisation is that over which the circuit will tune. For other frequencies, additional test circuits must be designed and constructed, explaining why it is sometimes difficult to get a semiconductor manufacturer to supply impedance data for special conditions of operation such as different frequencies, different power levels or different operating voltages. ■

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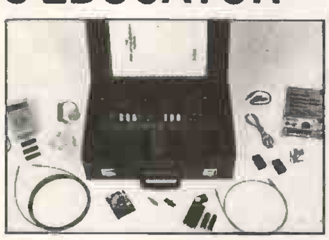

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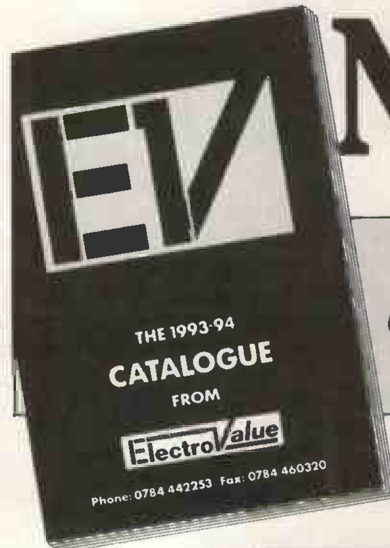

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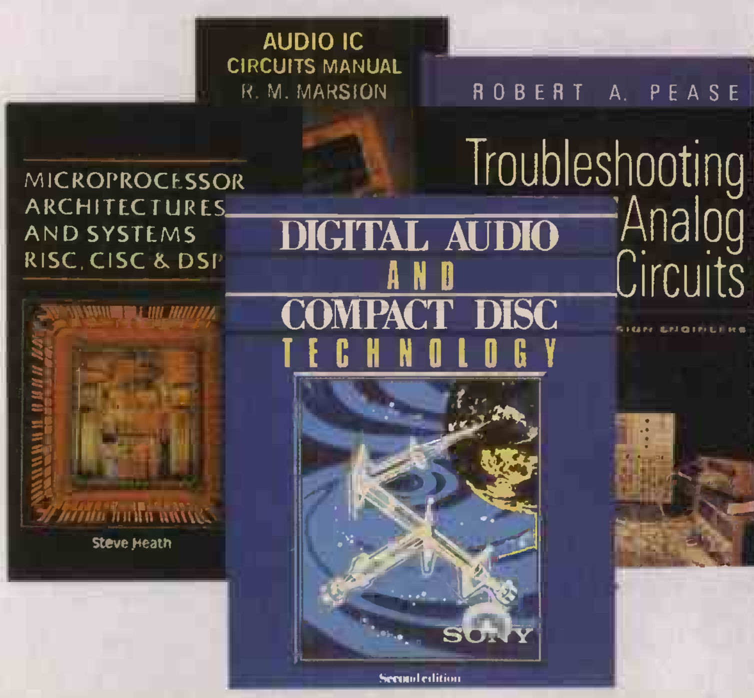

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Rewriting the rules with current mode amplifiers

Current mode amplifiers open up new designs for wide bandwidth circuits while retaining similarities with conventional operational amplifiers. Bashir Al-Hashimi uses commercially available devices in design examples of high performance, wideband amplifiers.

Circuits working at video frequencies used to depend largely on discrete or hybrid designs. But now, better fabrication techniques coupled with novel circuit design have given rise to a new family of integrated circuit amplifiers based on the current feedback approach[1] (CFA).

Performance of these monolithic amplifiers matches or surpasses their hybrid counterparts at a fraction of the cost.

To minimise deviation from the standard approach, CFA manufacturers are designing their products to be used directly in text-book op-amp configurations – in spite of their significantly different internal design from conventional op-amps. So obtaining optimum performance with the various commercially available CFAs[2, 3], (**Table 1**), has meant a new set of design rules has been formulated. It is these rules for design of wideband amplifiers that will be considered, using the *EL2030* 8-pin device (**Fig. 1**, Table 1) from Elantec.

As an illustration, we will look at the requirement for a HDTV system amplifier (**Table. 2**).

HDTV amplifier design

The HDTV amplifier (**Fig.** 2) needs a gain of 2 to provide matching to its termination. and uses the standard non-inverting op-amp configuration where the gain is $(1+R_F/R_I)$. Voltage feedback amplifiers (VFAs) allow an almost arbitrary choice of feedback – provided the ratio of R_F to R_I is correct.

This is the first, and possibly most important, difference between CFAs and VFAs.

With CFAs, the feedback resistor sets the amplifier bandwidth and the frequency response shape, as well as defining the gain of the circuit. R_F and R_I must not be arbitrary.

Table 1. Typical list of commercially available CFAs.

Company	Part no	3dB bw (MHz)	SR (V/µs)	ST (ns)	Op current (mA)	Quiescent current (mA)	Diff gain and phase
Elantic	EL2030	120	2000	40 to 0.25%	65	15	0.01%, 0.01
Harris	HA-5020	100	800	45 to 1%	60	7.5	0.02%, 0.03
Linear Tech	LT1223	100	1000	75 to 0.1%	50	6	0.02%, 0.12
Comlinear	CLC410	200	2500	12 to 0.05%	70	16	0.01%, 0.01
Analogue Dev	AD811	140	2500	50 to 0.1%	100	15	0.01%, 0.01
Burr-Brown	OPA603	100	1000	50 to 0.1%	150	21	0.03%, 0.02
Nat Semi	LM 6181	100	2000	50 to 0.1%	100	8	0.5%, 0.4

Table 2 Typical specification of a HDTV amplifier

Design parameter specification	Target
Bandwidth	DC to 30MHz
Gain flatness	<0.1dB
Group delay ripple	<2ns
I/O signal level output	1V (pk-pk) into 75Ω

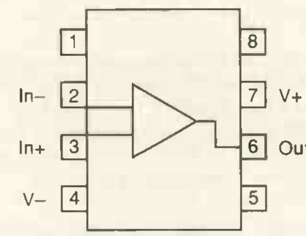

Fig. 1. The EL2030 8-pin device from Elantec used to test out the new design rules.

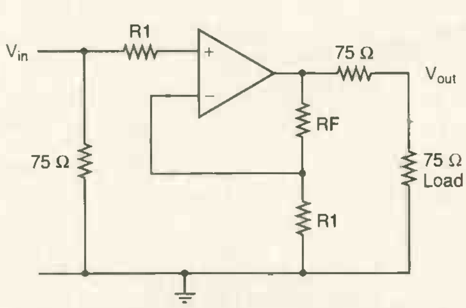

Fig. 2. The HDTV amplifier has a gain of 2 to match its termination and uses the standard non-inverting op-amp configuration where the gain is $(1+R_F/R_1)$.

Design rules to tackle disadvantages

Compared to VFAs, CFAs have several advantages and disadvantages. Advantages are:
Bandwidth largely independent of the closed loop gain;
Superior AC performance, including high slew rate and low settling time;
Linear phase response.

Disadvantages include:
Feedback resistor must be optimised
Unstable with capacitive feedback
Supply voltage affects performance.

But by following certain design rules, CFAs can be used much more effectively, with disadvantages largely overcome:

Choose the feedback resistor to set the amplifier bandwidth and the shape of the frequency response;
- Choose the other resistor (R_I) to set the amplifier gain;
- Do not use capacitive networks in the feedback path, since this drives the amplifier into oscillation;
- Use maximum allowable supply voltage since this allows CFAs to achieve optimum performance;
- Always terminate the amplifier input and output;
- Use RF layout techniques (see box).

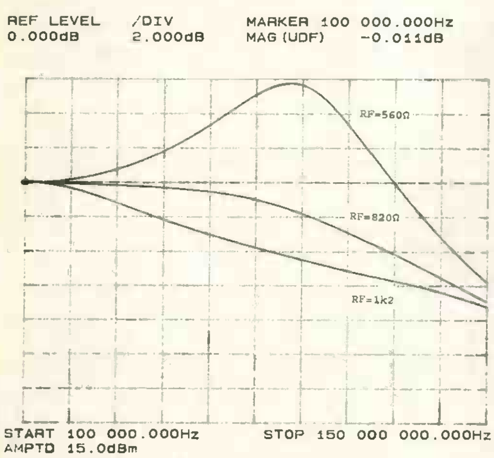

Fig. 3. HDTV amplifier frequency response for different values of RF. Adjusting R_1 for correct gain shows that RF affects the amplifier bandwidth and the frequency response peaks. R1 has no effect on bandwidth and frequency response, only affecting the gain.

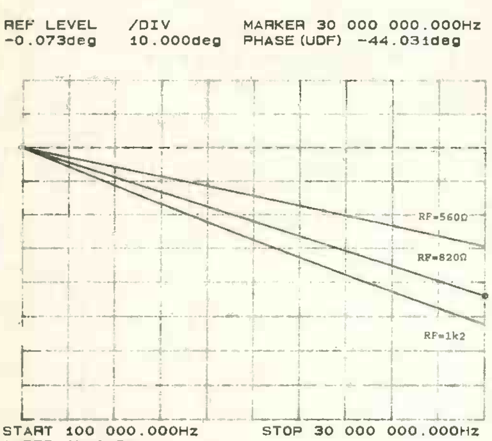

Fig. 4. Examining the phase response of the amplifier shows good phase linearity at high frequencies and the phase shift increases as the bandwidth decreases.

Frequency response of the HDTV amplifier at three different values of R_F (560Ω, 820Ω and 1k2) – adjusting R_I for correct gain – shows that R_F (**Fig. 3**) affects the amplifier bandwidth and the frequency response peaks. R_I has no effect on bandwidth and frequency response, only affecting the gain[5]. But it becomes clear that optimum bandwidth and maximally flat frequency response at a gain of +2 is obtained when R_F = 820Ω, the measured –3dB bandwidth being approximately 110MHz. Higher values of R_F decrease the bandwidth: lower values increase bandwidth at the expense of peaking in the amplifier frequency response. The minimum value for a given CFA – below which the CFA becomes unstable – is usually given in its data sheet. For example, lowering R_F in the HDTV amplifier below 400Ω causes the amplifier to oscillate. So, to change the amplifier gain, R_I should be varied not R_F.

Good signal fidelity will depend on a video amplifier exhibiting linear phase shift or a flat group-delay response. Examining the phase response of the amplifier for the previously used three different values of R_F (**Fig. 4**), we can see good phase linearity at high frequencies and the phase shift increases as the bandwidth decreases (or R_F increases).

Another difference between CFAs and VFAs is the effect of power supply. For example, lowering the supply voltage of the HDTV amplifier from ±15V to ±7V reduces amplifier bandwidth from approximately 110MHz to 96MHz (**Fig. 5**).

Returning to the target specification (Table 2), for the required gain flatness and group delay ripple, R_F must be 820Ω. Both requirements for the gain flatness and group delay are met (**Figs. 6** and **7**).

As with VFAs, the ratio of the feedback resistor to the input resistor ($-R_F/R_I$) determines the CFA voltage gain (**Fig. 8**). R_F sets

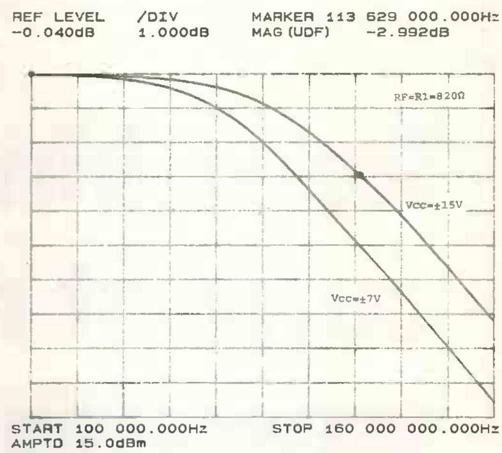

Fig. 5. Lowering the supply voltage of the HDTV amplifier from ±15 to ±7V reduces amplifier bandwidth from approximately 110MHz to 96MHz.

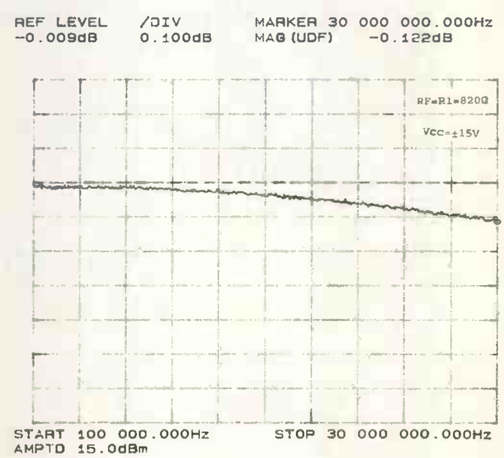

Fig. 6. Requirements for gain flatness...

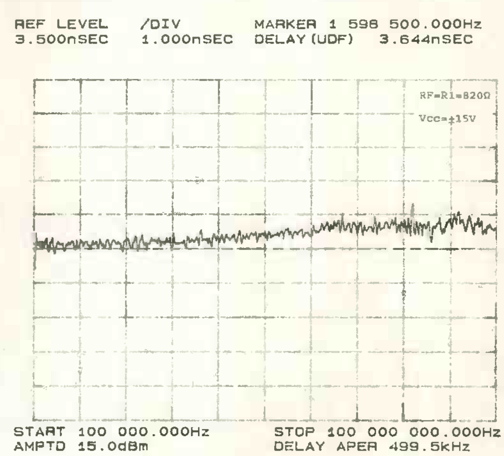

Fig. 7...and group delay are both met.

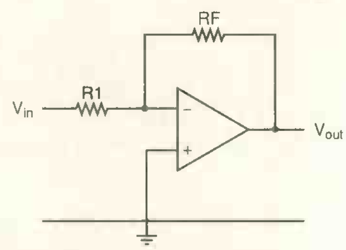

Fig. 8. Ratio of the feedback resistor to input resistor determines the CFA voltage gain.

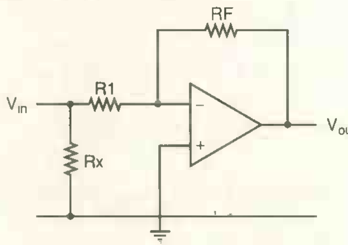

Fig. 9. There is no problem in simultaneously achieving the amplifier input resistance and the gain because a shunt resistor can be added.

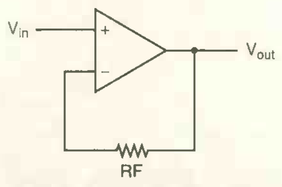

Fig. 10. In CFA-based buffers the output must be connected to the inverting amplifier input through the recommended feedback resistor.

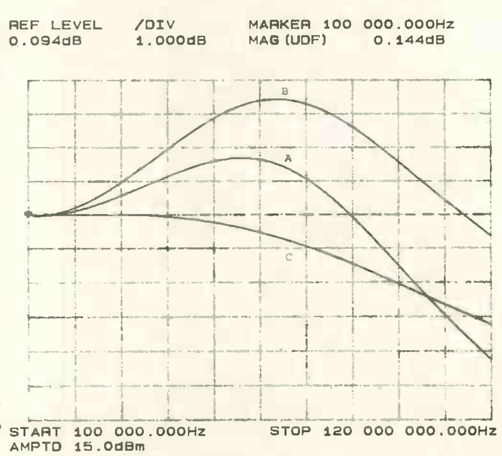

Fig. 11a. Modifying the circuit to a gain of +1 causes frequency response peaking due to stray capacitance to ground from the inverting input. 11b. Adding 2pF results in a further 2dB peaking, implying that 2pF of equivalent strays originally existed. 11c. Changing RF to 960Ω provides a flat response.

the bandwidth and frequency-response peaking, while R_I sets the gain. R_I also determines the input impedance of the amplifier, so a maximum input impedance is possible for a given amplifier gain. For example, take a 75Ω input impedance amplifier with a gain of –4. We know that the *EL2030* requires $R_F = 820\Omega$ to achieve maximally flat frequency response and optimum bandwidth. This gives $R_I = 205\Omega$.

Clearly there is a no problem in simultaneously achieving the required amplifier input resistance and the gain because a shunt resistor (R_x) can be added (**Fig. 9**). Amplifier gain is still ($-R_F/R_I$), and the value of R_x is given by:

$$R_x = (R_I R_{in})/(R_I - R_{in})$$

where R_{in} is the required amplifier input resistance.

Referring back to the example, if $R_{in} = 75\Omega$, $R_F = 820\Omega$ and $R_I = 205\Omega$ (to achieve a gain of –4), this gives $R_x = 118\Omega$. Care needs to be taken when using CFAs in applications requiring low impedances and relatively high gains. If in the previous case the gain required was –30, R_I would be 27Ω which would provide too low an impedance, and a buffer would be necessary.

Common practice in voltage feedback inverting amplifiers is to connect the non inverting input to ground through a resistor of value equal to R_F/R_I, giving bias current cancellation. Unlike a voltage feedback amplifier, a current feedback amplifier does not have two high impedance inputs[1].

The non-inverting input is a high impedance, of the order of 1MΩ, while the inverting input is a low impedance, around 30Ω. It means that the two CFA bias currents are unrelated and no attempt need be made to minimise them through the impedance matching of the inverting and non inverting inputs. In CFA-based inverting amplifier circuits, the non-inverting input should be connected directly to ground or through a small resistor (<30Ω) to ensure stability[6].

Buffer amplifier design

A common application of CFAs is in buffer amplifiers driving high speed flash A-to-D converters. These amplifiers must have wide bandwidth; fast settling time; low output impedance and the ability to drive large and variable capacitive loads.

Current feedback amplifiers meet all these requirement (Table 1), but to realise the full capabilities of an A-to-D converter, bandwidth of the driver amplifier should be at least three times the Nyquist frequency (half the sampling frequency), minimising gain and phase aberrations[7]. Also, the driver amplifier should be able to settle within one-half LSB of the correct value within the sampling period. For example, the driver amplifier must settle to 0.2% within 10ns, for use with an 8-bit 100MHz flash A-to-D converter.

In voltage feedback buffer amplifiers, the output may be connected directly to the invert-

Board layout and passive components

As with any high frequency device, care must be taken in board layout to maximise performance of current feedback amplifiers. Key points include:

- Use a large ground plane to asure that low impedance ground is available throughout the layout.
- Do not extend the ground plane under nodes which are sensitive to stray capacitance, in particular the inverting amplifier input. Make short and wide connection tracks to minimise losses.
- Bypass power supplies very close to the amplifier pins. For best results, bypass the power supplies with 1 to 10µF tantalum capacitors in parallel with 100nF ceramic capacitors. The power supplies should be well stabilised.
- Use surface mount passive components since they have the lowest inductance and capacitance.
- Avoid use of IC sockets.

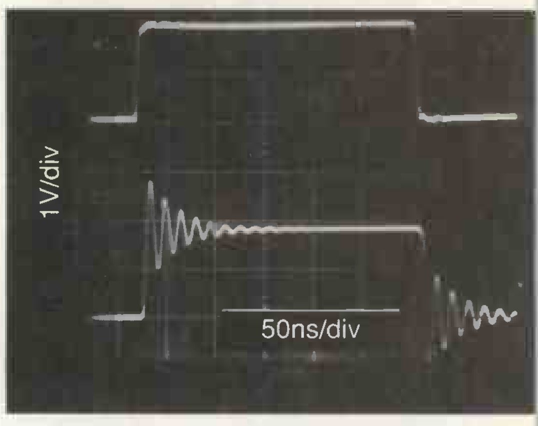

Fig. 12. Pulse response of the buffer amplifier shows excessive overshoot. (Top trace input, bottom trace output).

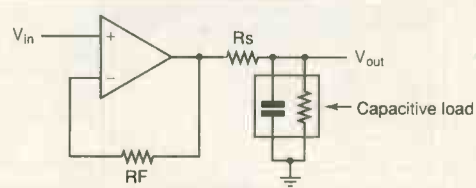

Fig. 13. To reduce overshoot, insert a series resistor between the amplifier output and the load.

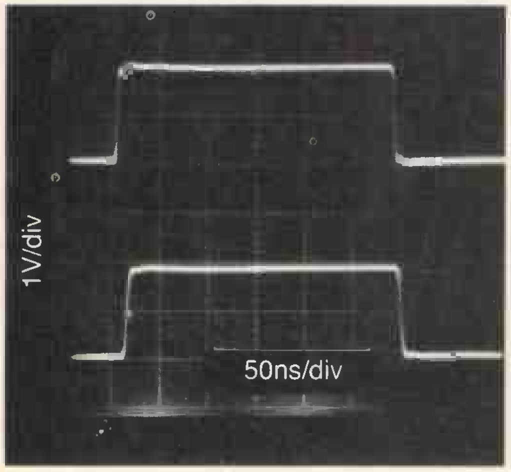

Fig. 14. Much improved performance is the result of adding the series resistance. (Top trace input, bottom trace output).

ing input, but in CFA-based buffers, the output must be connected to the inverting amplifier input through the recommended feedback resistor **(Fig. 10)**, ensuring amplifier stability. Current feedback amplifiers are sensitive to stray capacitances, in particular stray capacitance to ground at the inverting input. Effect of this capacitance is shown in the form of peaks in the frequency response **(Fig. 11)**.

From the previous results outlined for a gain of +2, the optimum feedback resistor R_F is 820Ω. Modifying the circuit to a gain of +1 by removing R_I, causes frequency response peaking to take place **(Fig. 11a)**, due to stray capacitance to ground from the inverting input.

When gain is employed, R_I shunts the effects of strays, and adding a capacitor to simulate larger strays gives the results in **Fig. 11b**. Here adding 2pF results in a further 2dB peaking, implying that 2pF of equivalent strays originally existed. Changing R_F to 960Ω provides a flat response **(Fig 11c)**.

An important feature of CFAs is that once R_F is chosen correctly, the –3dB bandwidth of the amplifier is approximately 116MHz with a gain of +1, and 110MHz with a gain of +2. Gain-bandwidth limitation of VFAs is not followed by CFA's.

Input impedance of flash A-to-D converters is highly capacitive, so that a typical 8-bit, 20MHz video flash converter will have an impedance of 25pF//100kΩ.

Pulse response of the buffer amplifier driving directly a 24pF//100kΩ load **(Fig. 12)** shows excessive overshoot, and increasing the load capacitor further results in the amplifier becoming more unstable. One way of reducing the overshoot and boosting the buffer-capability to drive higher capacitive loads, is to insert a small series (R_s) resistor between the amplifier output and the load **(Fig. 13)**. The result is a much improved performance **(Fig. 14)**, when R_s = 47Ω. Value of resistor (R_s) depends on the value of the load capacitor and should be determined from manufacturer data.

The drawback to this method is a reduction in amplifier bandwidth. ■

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Breaking Windows

With the profusion of *Windows* software being marketed and reviewed in the computer press, I'm struck by the paradox that so few people that actually like and use it.

Everyone seems to have their tale of woe regarding installation followed by the realisation that the cost in computer memory and disk space appears to be too great in comparison to any benefits gained.

Although computer magazines offer plenty of technical criticism of *Windows*, I don't ever recall seeing the need for its very existence being called into question. The truth of the matter I believe is that *Window*'s hungry demand for high performance machines is the driving force behind the computer sales market today. And the personal computer magazines that depend on the industry's advertising budgets are hardly likely to bite the hand that feeds them.If the trend is for software producers to concentrate their efforts only on *Windows* compatible versions, I suppose all of us will eventually be forced into buying expensive upgrades to maintain compatibility.

It seems a pity that the steady improvements in PC price and performance should be soaked up by such a cumbersome and gimmicky operating system rather than more powerful applications.

In the meantime is there anyone else out there prepared to stand up and say "I'm sorry I don't do *Windows*".

John Carrey
Malvern

Clunky versus cost

Having spent a large proportion of the past six years investigating the perfect audio switch for use in high quality music recording consoles, I would like to add a few comments to Mike Meechan's article on the subject (*EW + WW*, July).

First, I'm afraid that you cannot dismiss the mechanical switch or relay too lightly; it still provides the closest we have to an ideal automatable switch. In both laboratory and listening tests I have found no problem with either speed or wear with modern relays. Power consumption, too, can be overcome by using either reed or latching relays, although the latter can cause a problem if the system forgets how they are set.

The problem is a trade-off between clunkiness and cost. A clunky relay and its driver cost about the same as a decent semiconductor switch, a reed relay rather more. The Focusright console, regarded as having extremely high audio purity, uses as far as I know only reed relays for signal routing.

Secondly, the humble jfets. Meechan passes over these rather quickly, yet they have been the standard for top-of-the range mixing consoles for many years. They have the advantage over cmos switches of a lower on impedance, higher voltage range, and infinitely lower propensity to blow up when you look at them. Simply switching the gate via a 10MΩ resistor gives good results. However, if the gate drive is bootstrapped, as Meechan shows, then high frequency distortion will improve due to the negation of the gate-drain capacitance.

Bootstrapping the gate can also avoid the other annoying problem of jfet switches, that the gate current appears in the channel, requiring it always to be fed from a source of low dc impedance. It was as a result of our development at Solid State Logic of the bootstrapped jfet arrangement that Analog Devices developed the SSM2402, and the derivative 2412 mentioned in the article.

Last, but definitely not least, a J112 jfet costs about a quarter the price of a VN88 mosfet.

As a final thought, having finally escaped from the audio industry I am now able to start using cmos switches seriously. Can anyone explain why they seem to be so much more prone to static damage than other modern cmos devices?

Andy Millar
Whitstone, Devon

Sounding out the critics

I was pleased to see Drs Blake-Coleman and Yorke report on their investigation into fancy speaker cables (*EW + WW*, May). Their results seem to be pretty much in accord with what Davis[1], Greiner[2] and others have been trying to tell us for years. In fact, I was personally involved in a similar test recently that produced pretty much the same result. I am pleased that the good doctors have publicly put their weight behind a more informed approach to hi-fi interconnections.

I also read with keen interest Ben Duncan's article on op-amp distortion "How clean is your audio op-amp?" (*EW + WW*, January)[3]. It was a thought provoking piece, more for the explanation of the exercise than the actual data. I was a little amused, however, that Duncan has been unable to uncover any harmonics generated by resistors.

Resistors are about the most linear circuit components around. I think his time might be more productively spent investigating the non-linear behaviour of electrolytic capacitors! There are also many other factors, such as circuit layout, more important to distortion than resistor non-linearity.

I disagree with Duncan's idea that the "Cause of the difference is... relatively unimportant". Before we can accept his hypothesis as anything more than conjecture, it is essential that we establish any genuinely audible effects and to ascribe them accurately to differences in the circuit. For example, it is pointless to measure differences in distortion if listeners respond only to a change in mains noise.

As far as Chris Daly's remarks go (*EW + WW*, April), I think he has the wrong end of the stick. One does not need a hot soldering iron nor any practical application to recognise a poor argument.

My criticism of Duncan's is that he proves nothing much. Since there is no scientific evidence that the effects are audible Duncan has furnished us with no more than hypothesis. Unsubstantiated anecdotes are not proof. It is often the lack of theoretical and experimental rigour that renders claims from the golden eared or subjectivist club invalid.

Reproducible experimental results and a firm basis in theory are needed before we can take them seriously.

For the record, I consider it to be very bad practice to trust critical signals to the quite dubious performance of a reversed biased electrolytic capacitor. In fact it would be better if electrolytic capacitors were not used in the signal path at all. The only good features of electros are their relatively small physical size and their cheapness, and in all other respects their performance is pretty lousy.

Also, I have no doubt that some people will think that using this circuit makes their amp sound more euphonious, but that does not mean that they can actually hear any differences, whatever the cause. The levels of distortion that we are discussing are probably inaudible to the vast majority of people[4]. The fact is that average enthusiasts cannot perform properly controlled listening tests in their own living rooms. They very rarely have the facilities or the expertise.

For example, it is necessary to control the effects of room acoustics, extraneous noise and the position of the listeners. It is essential that the sound level and the frequency response be maintained to within 0.2dB[5] (that is about 2% across the audio band, which is outside the capabilities of most multimeters and oscilloscopes).

Also from my own experience I know the sound will depend upon the age and experience of the

Power to protect

On reading R Gough's letter (*EW + WW*, August), I was struck by several points. I study with a group of people that doesn't know that the human body is tuned to a cosmic keyboard, but there you are; there's just no telling some people.

However, I was most perturbed to read that certain music sounds can stimulate the pancreas, pineal glands and so on. Clearly there are unsuspected dangers in listening to music or indeed any other sound; unlooked-for stimulation of the pancreas is likely to send us into insulin coma while nudging the pineal is going to cause sudden uncontrollable attacks of jet-lag, or possibly mass hibernation.

The hazards are obvious, and I think that Mr Gough has a clear duty to specify which frequencies have these hitherto unsuspected effects, so that we audio engineers can guard the public against them with suitable banks of filters.

Seriously though, what is this kind of wretched nonsense doing in *EW +WW*? Without wishing to be catty, surely the "I've out-thought Einstein" contingent are enough of a cross to bear.

Douglas Self
London

As editor, I have a duty to allow heretical discussion and presentation of controversial views. This is not just a result of some woolly liberalism, but the shameless desire to give entertainment to other readers - FO.

listener, as well as their state of health and fatigue, and whether they have had any stimulants or depressants (such as coffee, tea, alcohol, cigarettes,and so on) within hours before any auditioning. Lipshitz and Vanderkooy even suggest that the source material often has greater effect on the sound than the performance of the reproduction equipment[6].

Based on the evidence that Duncan has published it is unlikely that the new harmonic structure is due to harmonics from the mains. If it were caused by mains we would see many more spectral lines at multiples of 50Hz, not just the 1kHz spacing that we do see. I think Duncan would probably agree with me there.

It is interesting then that Daly claims "the near total absence of mains noise (hum)" when Duncan's results show no effect on mains noise. Duncan shows that the reverse biased capacitor causes an increase in even harmonic distortion. Daly on the other hand says only that he "noticed no sonic degradation". One cannot say that Daly's claims support Duncan's results, since Daly's experience with this modification is quite at odds with what Duncan seems to be saying. If anything Daly's claim that he "noticed no sonic degradation" tends to suggest that these levels of distortion are inaudible.

However, I am perplexed by a philosophy that seems to be telling me that some distortion is acceptable because it makes the music more euphonious (to some people), but at the same time tone controls should not be allowed because they introduce distortion, even though they reduce sonic aberrations and make the music more euphonious when properly used. Such an illogical and inconsistent philosophy is not likely to maintain much credibility.

Phil Denniss
University of Sydney, Australia

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Authors wanted

There are a number of subjects and ideas that I have not seen in print that could provide the base for some interesting articles. For example information on surface acoustic wave resonators; they come in one and two port configurations. One manufacturer of these, Siemens, is fully booked through 1994. I would like to see a list of other possible suppliers.

Incidently did you know saw resonator oscillators hold the record for noise floor/spectral purity, they give a better figure than the best crystal oscillators. It is not only *Q* that matters, saws can operate at much higher power.

A review of university departments would be interesting for prospective students and people with research money. Who is an expert at what and where?

Good technical information means finding out what is not on the manufacturer's data sheet. This is possible if its made in the UK and almost impossible from elsewhere. Few distributors are interested in providing technical information on products from their suppliers that are outside the range they stock.

What MPs have a technical background? To whom are we to write if we wish to complain about the sale of the UK's last truly world class electronics capability – STC Submarine Cables.

Submarine fibre optic cables are preferred by telecommunications companies worldwide to satellite for commercial and user convenience. The competitors ATT and Alcatel are regarded in their own countries as critically strategic. They are directly supported and control would never pass outside their national boundary.

Inertial sensors could only be afforded by the military. An inertial platform for a submarine could set one back £0.5m. Accelerometers are available to the automotive industry for a few pounds. Angular rate sensors (the equivalent of a rate gyro) are required to sense car turning for computer control of steering geometry, suspension rotation, steering wheel spin (the Magimix effect), and to interpolate between GPS fixes.

Who makes these low cost sensors, what specification parameters are important, and what other applications could ride on low cost automotive devices?

The department of the government chemist is really useful in letting interested parties communicate in the field of chemical sensors; this should be a model for the rest of the DTI which in most things is truly appalling.

Douglas Dwyer
Okehampton, Devon

If any readers are interested in taking up some of the above ideas for articles, may I remind them of the writers' award scheme that we run. *Turn to page 813 to find out more* *- Editor.*

A question of theory

Surely messrs Catt, Elmendorf, and Goldberg (*EW + WW*, August) have missed the point. It is rare indeed for a scientific theory or formula to be proved correct or even incorrect, as the continuing big-bang versus steady-state controversy illustrates.

All one can generally say is that the predictions of one tend to fit the evidence better than those of others. The other reason for preferring a theory is simplicity. Generally a very large number of possible theories will fit the observations; to pick the simplest one has itself proved to be a useful principle.

The Michelson-Morley experiment indicated the aether to be stationary with respect to the Earth, a circumstance untenable in the prevailing philosophy. Another possibility not considered by Goldberg or anyone at the time is that the aether is viscous, and stationary wherever it is measured – rather like air, which tends to attach itself to moving bodies. Of course this makes nonsense of the constancy of the velocity of light, and can be dismissed lightly; its only value as an explanation would be if it fitted reality better than current theories.

We sometimes forget that our theories – and indeed basic concepts like particle and wave – are only convenient fictions to describe complex behaviour.

I could go on, but since challenges seem to be in fashion, I challenge Elmendorf to indicate first with respect to what is the Earth's rotation to be measured, and secondly, if he finds our weather systems unconvincing, what conceivable type of evidence he would accept as absolute proof that it does?

AM New
Bristol

Drawing error

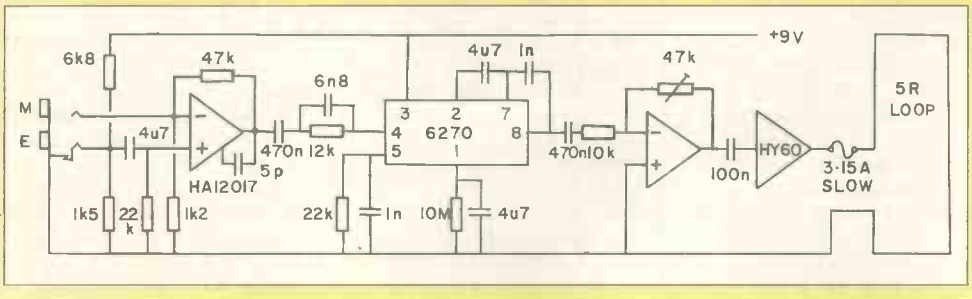
Unfortunately there is an error in the redrawing of the circuit diagram in my article "Audio induction technology for the deaf" (*EW + WW*, September).

The universal preamp should have shown separate jack sockets to different points in the circuit for the dynamic (M) and electret (E) microphone inputs, the former being open circuit and the latter being short circuit when unused.

This provides the different configurations needed for the two types of microphone and results in a similar overall sensitivity.

Specifically, for M, the E input is grounded giving a low noise, low impedance input and a higher gain for lower impedance microphones. For E, M is open circuit giving medium gain, medium impedance input and providing dc to power the fet in the microphone.

JP Wilson
Keele University



CIRCUIT IDEAS

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THIS MONTH'S £100 CIRCUIT IDEA

FSK receiver has auto decision-threshold control

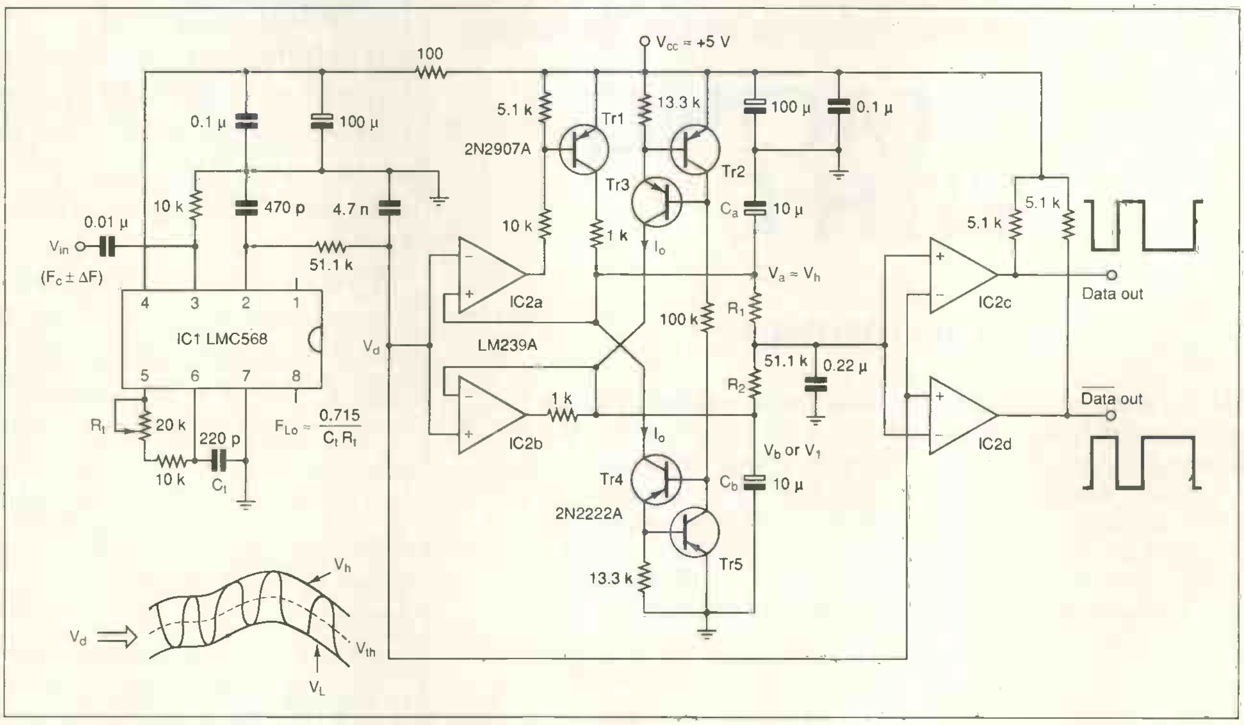
This low-cost, low-power PLL receiver for FSK binary-coded data automatically sets the best decision-threshold level for recovering the data.

Ideally, V_{th} should be midway between V_h and V_l, the upper and lower peak voltages of the demodulated signal, which are not usually stable due to the effects of drifts and gain inequalities. Signal from the *LM568* demodulator passes to comparators $IC_{2a,2b}$, which work in conjunction with current source $Tr_{2,3}$ and sink $Tr_{4,5}$ as positive and negative peak detectors, tracking variations of the peak voltages. Their outputs are combined and averaged by $R_{1,2}$ to put V_{th} half way, this voltage going to the output comparators, which also accept the input from the demodulator. Current generators improve the dynamic range of the peak detectors.

With a 100mV carrier at 110kHz, the local oscillator at $2F_c$ and frequency deviation between 500Hz and 5kHz, drift was simulated by adding a 0.1Hz signal to the 300b/s digital modulation. When $V_h - V_l$ is between 100mV and 600mV and V_{th} 1-3V, data is always recovered with negligible jitter.

G Stochino
Ericsson Fatme Spa
Rome
Italy

Simple method of adapting data-recovery decision-level threshold to take account of noise, drift and gain variations in the amplifiers and demodulator.



Cheap mosfet audio power

To provide very good performance using cheap mosfet output devices, this circuit uses an op-amp driving a conventional source follower output stage.

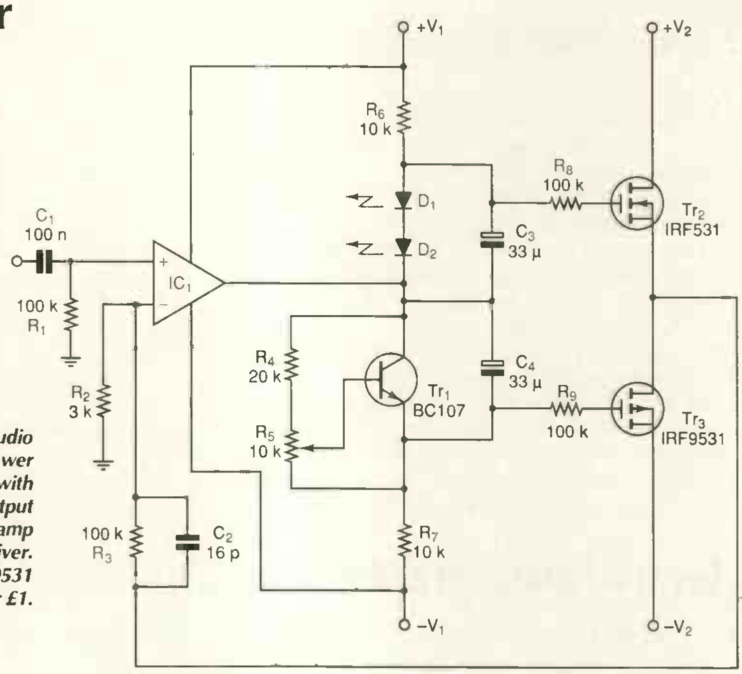
Transistor Tr_1 sets the quiescent current, the two leds dropping a fixed 3.2V; a single zener would be just as suitable. There is a small drift, but since quiescent current is not critical, there is no correction.

Since the output stage voltage gain is less than unity, the op-amp should be chosen to have enough internal frequency compensation to avoid instability; it should also have a high slewing rate and reasonable current drive. Capacitor C_2 reduces gain at higher frequencies.

Power output depends on supply voltage; at ±22V, power is 15W into 8Ω.

Being a low-cost circuit, it may be feasible to use several with active crossovers, rather than one higher-quality amplifier.

Andrew Southgate
Sutton, Cambridgeshire



Low-cost audio power amplifier with mosfet output and an op-amp driver. IRF531/9531 cost under £1.

No-loss tuned circuit

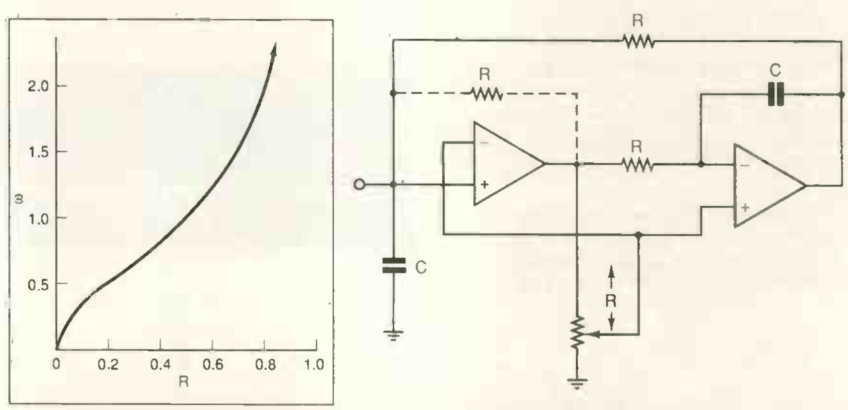
The circuit shown is (ideally) a loss-free parallel tuned circuit, for use as part of a band-pass filter or as a reasonably accurate sinewave oscillator if the dotted resistor is in place, although a diode limiter across one of the capcitors may be needed in the oscillator form. With component values normalised to unity, the dotted resistor would be typically 50.

Frequency control is also unity, *R* being the top half to give

$$\omega = \sqrt{\frac{R}{1-R}}.$$

Varying the control gives a fairly straight line over around two octaves, accurate enough for use in a simple spectrum analyser.

McKenny W Egerton jr
Owneys Mills, Maryland, USA.



Loss-free parallel tuned circuit with one frequency control to give a nearly straight resistance/frequency response over two octaves.

Transconductance squarer

An op-amp and a dual fet combine to give output kv_i^2 when v_i>0.

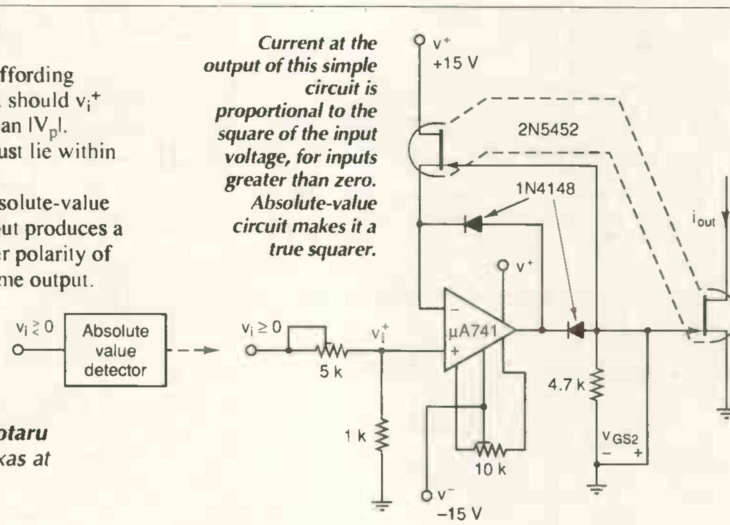
National Semiconductor's *2N5452* n-channel dual fet has good matching between the two devices and low output conductance. Normally, the voltage between the fet gates and the non-inverting op-amp input is constant at V_p, the pinch-off voltage of the two transistors. V_{GS2} is $(v_i^+ + V_p)$ and i_{out} is proportional to v_i^2.

The coefficient k is adjustable by means of the 5kΩ input variable, the two diodes and the 4.7kΩ resistor ensuring that i_{out} does not exceed I_{DSS} and affording negative feedback should v_i^+ become greater than $|V_p|$. Output voltage must lie within the 5-15V range.

Including an absolute-value detector at the input produces a true squarer: either polarity of input gives the same output. With a *µA741*, the circuit works at several kilohertz.

Alexandru Ciubotaru
University of Texas at Arlington
Texas,
USA

Current at the output of this simple circuit is proportional to the square of the input voltage, for inputs greater than zero. Absolute-value circuit makes it a true squarer.



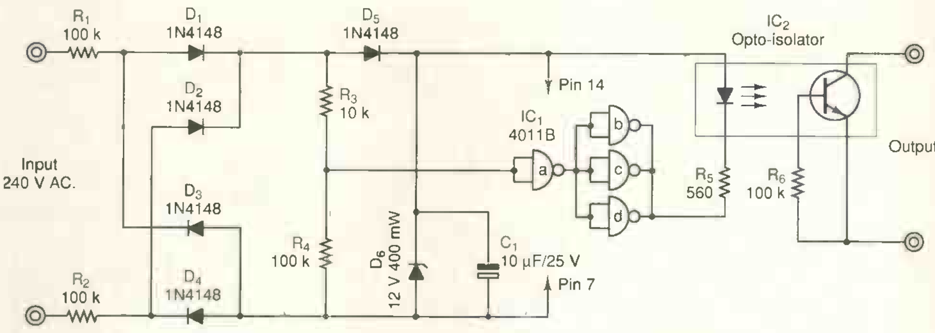
Zero-crossing detector

At every zero crossing, this circuit produces a 150μs pulse, centred on the crossing, the output being optically isolated from mains input.

Diodes D_{1-4} rectify the input current, which is limited by $R_{1,2}$ to around 1.3mA; power dissipation is minimal, so virtually any kind of resistor will suffice. Output from the rectifier powers the quad Nand IC_1, zener D_6 limiting the voltage to about 12V.

Since input to IC_1 is normally high, output from the three parallel gates is high, except for the period around each zero crossing, when input to IC_1 is low and so is the output, which turns the led and output transistor on. Resistor R_6 avoids trouble caused by stray coupling at the transistor base. Capacitor C_1 takes about 100ms to charge at switch-on.

AJ Flind
Taunton, Somerset



Simple zero-crossing detector produces an isolated pulse, used by the author to synchronise a sawtooth generator.

Electrocardiograph simulator

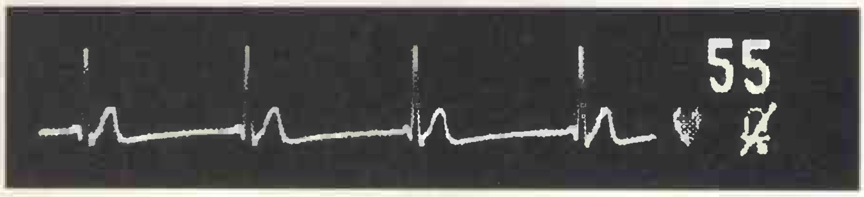
This simple ECG simulator has been useful for some years in teaching and equipment test and development.

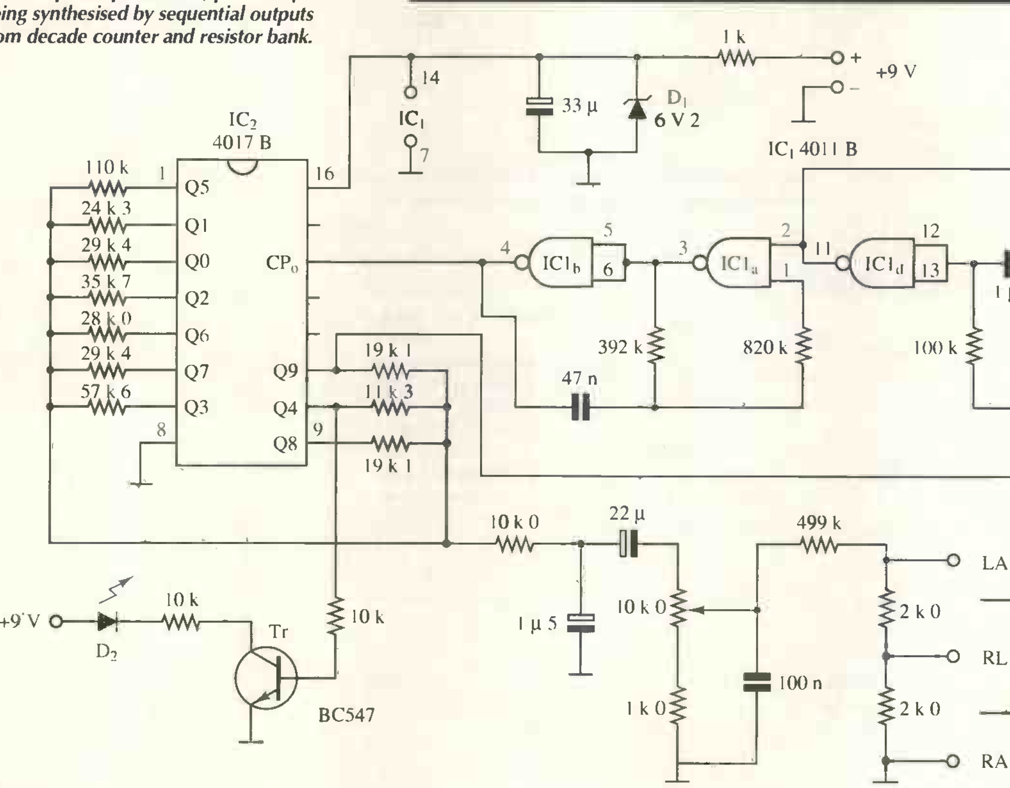
$IC_{1b,1a}$ form an astable flip-flop at 20Hz to drive the clock input of the decade counter IC_2, which produces sequential highs on its Q_{0-9} outputs. The voltage at the common point of the output resistors is determined by the potentiometer action of one on resistor and nine off, the values being chosen to form the required pulse shape.

When Q_9 goes high, it inhibits the $IC_{1c,d}$ monostable for a time set by the 2.2MΩ variable resistor, so controlling the pulse rate from 30pulses/min to 160p/m. At each pulse, led D_2 blinks.

The 10kΩ resistor and 1.5μF capacitor smooth the stepped output from the resistor bank and the 10kΩ pot. sets the output to 0.2mV-2mV pk-pk. Over 4kΩ output impedance could trigger the "lost electrode" alarm on some cardiac monitors. ■

Alberto R Marino
Madrid
Spain





Electrocardiograph simulator, variable from 30 to 160 pulses per minute, pulse shape being synthesised by sequential outputs from decade counter and resistor bank.

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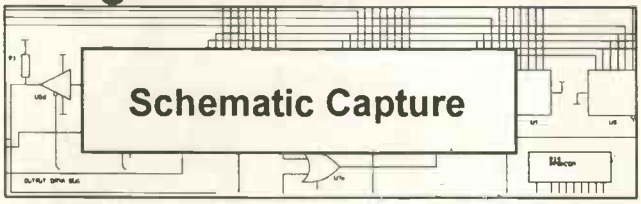
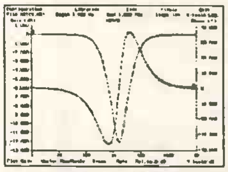
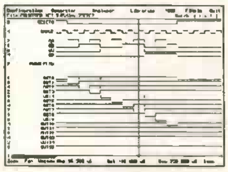
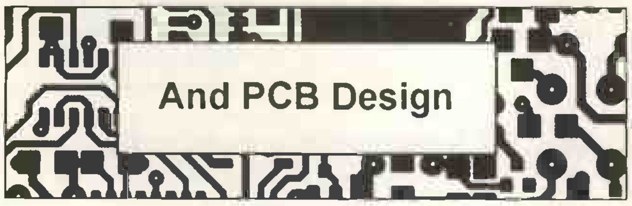




Analogue & Digital Simulation





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12-bit, 2MHz ADC. Datel's *ADS-132/883* analogue-to-digital converter, qualified to MIL-STD-883, is a high-speed A-to-D combined with a fast sample-and-hold amplifier in a small 32-pin TDIP package. It has its own reference, three-state outputs, digital correction circuitry and internal clock. Total harmonic distortion is –80dB. Datel (UK) Ltd, 0256 880444.

Discrete active devices

13.5GHz transistor. The industry's highest gain/bandwidth product of 13.5GHz is achieved by Hitachi's 2SC5080, which also offers a noise factor of 1.1dB at 900MHz, 5V and 5mA. At a bandwidth of 2GHz, forward transfer gain at around 900MHz is 11.7dB. The device is in a miniature SM resin moulded MPAK-4 package. Hitachi Europe Ltd, 0628 585000.

SM rectifier diodes. 1W surface-mounting rectifier diodes by Temic-Telefunken are available in the 50V-1000V range in standard, fast and ultrafast versions. ITT MULTIcomponents, 0753 824212.

UHF mesfet. *SGM2014M* is a GaAs dual-gate mesfet by Sony, intended for UHF low-noise amplifiers, mixers and oscillators. Noise figure is 1.5dB at 900MHz and gain is 18dB; cross modulation is low and there is a gate-protection diode built in. The device is packaged in a standard four-pad SOT-143. Sony Semiconductor Europe, 0784 466660.

Linear integrated circuits

VHF linear amplifier. *RF2103* is a medium-power linear amplifier for digital or spread-spectrum transmitters operating in the 800-1000MHz region, or as an exciter for higher powers. The IC is self-contained, with a power-down facility, and produces 800mW CW or 400mW average for two-tone input, at 6.3V DC or 3V (135mW CW). Gain is 25-30dB. Anglia Microwaves Ltd, 0277 630000.

Current-feedback op-amp. Burr-Brown *OPA623* has a large-signal bandwidth of 350MHz at 2.8V pk-pk and $\pm$70mA output. Quiescent current is 4mA. An external feedback resistor sets open-loop gain for best frequency response and constant bandwidth with varying gain. Slew rate is 2100V/μs. Burr-Brown International Ltd, 0923 233837.

Power-factor controllers. *MC34262/33262* are power-factor controller ICs meant for use as preconverters in electronic ballasts and in off-line power converters. They include start-up timer, single-quadrant multiplier for near-unity power factor, zero-current detector, transconductance error amplifier, trimmed bandgap reference and output for power mosfets. There is a full range of protective measures. Motorola Ltd, 0908 614614.

Logic building blocks

Modems. Hitachi has released three new single-chip modem ICs, *HD819003/04/10*, based on the earlier *81900*. They conform to CCITT V.29, V.27ter and V.21ch2 standards and carry out G-III fax mod./demod. operations. Functions now on-chip include touch-tone telephone control, transmission message record and playback and reception data error correction. Hitachi Europe Ltd, 0628 585000.

Bus interface. IDT's *IDT74 FC1625 11T* is said to be the fastest buffered parity generation and checking device available, and can be configured as a register, latch or transceiver, supporting 16 bits with byte-level parity generation and checking. The 56-pin IC replaces up to eight chips. Integrated Device Technology, 0372 377375.

RS232/562 transceiver. Designed for low-power application, the Linear *LT1331* is a three-driver, five-receiver interface IC that keeps only one receiver alive during shut-down. It operates in 3V and 3V/5V systems and includes $\pm$10kV electrostatic discharge protection, which is claimed to be the industry's highest. Using 0.1μF charge-pump capacitors, the device supports data rates to 120kbaud into 3Ω and 2500pF in either mixed 5V/3V or 3V only modes. The active receiver wakes the system up on receipt of data. Linear Technolgy (UK) Ltd, 0276 677676.

CD interface/audio processor. *MCD220* by Motorola interfaces CD drives with the CD-I system bus, performing all CD-I base case audio functions; it is compatible with CD-DA, CD-ROM, CD-karaoke and Photo-CD standards. The device runs at normal, double or quadruple speed and supports JTAG boundary scan. Motorola Ltd, 0908 614614.

Home automation modem. *ST7537* by SGS-Thomson is a modem IC intended for the control of domestic heating, security and other functions in the home. The use of frequency-shift keying makes for reliability, data rate being 1200b/s at a carrier frequency of 132.45kHz. Data is conveyed by AC power cables in compliance with Cenelec EN50065-1 and the US FCC standards. Multiple devices operate on the same power line. SGS-Thomson Microelectronics, (Italy) 39-6035-597.

Fifos. TI's new first-in-first-out families are expressly designed for communications and digital signal processing, using multi-stage clocked architecture to handle data at up to 67MHz clock speed. One-bit fifos are available for telecomms, in which users have been forced to specify up to 9-bit devices, wasting memory. For DSP, there is a 36-bit design, the *SN74ACT36XX*, running at 67MHz. Texas Instruments, 0234 223252.

Memory chips

64Kb eeprom. AMT's *24C65* is claimed to be the world's first 64Kb device and is the first in a new series of smart serial eeproms. Bus rate is 400kHz, there is a 64Kbyte data-input cache and up to eight devices can be on the same bus to give 512Kb. A 4K block of one million cycle erase/write memory is provided, as is a 60K block 10,000-cycle memory. Arizona Microchip Technology, 0628 850303.

Microprocessors and controllers

Floating-point accelerator. An arithmetic co-processor board using the risc machine's floating-point accelerator chip is meant for use with the Archimedes *A540* and *A5000* and the *R260* Unix station. The *FPA10* interfaces with the ARM family of CPUs,implementing often-used instructions on-chip and a software/chip combination for the rest. The chip fits a socket on the *A5000* motherboard and the *A540* card to increase floating-point performance by about 50 times; double and extended precision calculations are accurate to 14 and 18 decimal places respectively. Acorn Computers Ltd, 0223 245200.

Microcontroller. An eight-bit risc microcontroller by Arizona, the *PIC16C84* contains 1K by 14b of eeprom program memory, 64Kbyte of eeprom data memory and 38 by 8b of

Telecomms controllers. *HD404638/9* are low-power telecommunication system controllers operating from voltages down to 2.7V, several power-saving stand-by and stop modes being provided, as well as a 32kHz "sub-active" clock. The two devices have 8K and 16K by 10 bits of rom respectively and 1152 by 4 bits of ram. The *HD4074639* has 16K of one-time-programmable memory, and all three have DTMF generator, timers, serial interfaces, a voltage comparator and input capture. Hitachi Europe Ltd, 0628 585000.

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general-purpose ram. Supply is 2-6V and the sleep mode current is 400nA. There are four hardware interrupts, including wake-up on keystroke and the instruction set has 35 single-word instructions to give a single-cycle execution rate of 400ns. Arizona Microchip Technology, 0628 850303.

Frugal micro. *TMP68003F* by Toshiba is a microprocessor based on the 68HC000 microprocessor core, but with two types of power-down and a clock generator. In power-down, either the clock alone is left running to support external circuits or the whole thing is turned off, in which case current drain is 0.1mA. The 80-pin PQFP package has all the address bus pins on one side and all the data pins on the other. Toshiba Electronics (UK) Ltd, 0276 694600.

Mixed-signal ICs.

PC sound synthesiser. Yamaha's *YMF278* sound synthesiser generates both PCM and FM sound through a single device; PCM comes from a 16Mb rom. With the addition of a D-to-A converter, an effects processor and a wave rom, it forms a MIDI level 1 system. The FM generator is compatible with the *YMF278* sound source LSI. Twenty-four tones may be generated simultaneously at 44.1kHz sampling rate, the data word being 8, 12 or 16bit. Output is in stereo. Polar plc, 0525 377093.

Voice for modems. Rockwell's *RC96V24ACW* cmos modem chipset allows a PC to act as a comms station for data, fax and voice storage and playback from one telephone line. It conforms to the TR29.2 standard and Rockwell has a development program (VAPI) for Windows and dos which provides a driver. VAPI stores data on disk and has intelligent call discrimination to sort out the use of the single line. RCS Microsystems Ltd, 081-979 2204.

Optical devices

Bright led. Tosbright *TLOA156P* by Toshiba is an exceptionally bright led, intended in the main for use as high-mounted car stop lights, where it offers size and placing advantage over the more usual incandescent bulb, measuring only 5mm diameter. Intensity is 800mcd at 620nm over 30deg angle, with a forward current of 20mA. Toshiba Electronics (UK) Ltd, 0276 694600.

Programmable logic arrays

5ns eecmos pal. Claimed by AMD to be the fastest electrically erasable 20-pin programmable logic device currently being sold. Maximum clock speed is 142.8MHz and the device uses less power than bipolar 5ns ICs, also offering the facility of frequent and fast reprogramming. *PALCE16V8H-5* has pull-up resistors on inputs and i/o and is a plug-in replacement for the *PAL16R8* and most of the *PAL10H8* series. Advanced Micro Devices UK Ltd, 0483 740440.

600MHz DSO. With a sampling rate of up to 5Gsample/s, LeCroy's *9360* digital oscilloscope is able to digitise waveforms completely in one shot, bandwidth being 600MHz. Smart triggering offers edge, window, glitch, pulse width, interval width, state and edge qualified, dropout and television triggers. Basic waveform processing is included, with enhanced processing a firmware extra. A built-in printer produces screen dumps in 10s. LeCroy Ltd, 0235 533114.

PCMCIA PLA. *EPM7032* is Altera's first entry to the personal computer memory card market and is one of the multiple-array matrix 7000 family of high-speed EPLDs, having 32 macrocells and 36 i/o pins. Pin-to-pin delay is 7.5ns. The devices are supported by *MAX+PLUS II* software for PCs and workstations. Altera UK Ltd, 0628 488811.

Power semiconductors

95% efficiency regulator. Linear's *LTC1147* is an 8-pin step-down switcher exhibiting 95% efficiency at an output current of 20mA-2A. with a stand-by current of 160µA (quiescent 20µA). It comes in fixed 3.3V and 5V versions, accepting 4V-16V at the inputs. An external mosfet is driven at up to 250kHz. Burst mode stand-by power is 2mW with a 10V input. Linear Technology (UK) Ltd, 0932 765688.

2A, 5V switcher. *MAX726* is a step-down switching regulator, accepting 8V-40V (60V in the 726HV) and producing 2.5V-40V. It is a buck regulator and can be arranged as an inverter, a negative boost converter or a flyback converter with input down to 5V. Power switch, 100kHz oscillator and control circuitry are built in. Maxim Integrated Products Ltd, 0734 845255.

Transient suppressor. Semtech's tranient voltage suppression diode array handles 300W peak pulse power per line, has data, signal and DC supply line protection, and is obtainable in unidirectional and bidirectional forms. Reverse stand-off voltage is 5-24V, leakage current 1µA or 100µA and capacitance 63-550pF. Semtech Ltd, 0592 773520.

Motor control. Siliconix's *Si9901* provides low-to-high-voltage interfacing to the Si9911 and Si9914 high-side drivers, with built-in protection for high-voltage motor control. Si9901 low-side driver includes the bootstrap diode, protection circuitry, fault indication, cmos-compatible input and a pump oscillator. A 1µA quiescent current reduces drain on the boostrap capacitor. Siliconix, 0344 485757.

Power mos arrays. Under the name *POWER+ ARRAYS*, Texas offers up to seven power transistors on one chip. In various packages, *TPIC5201/2202/2301/2701* contain from two to seven devices. Those with up to three transistors are 7.5A, 60V devices, while the seven-device package (TPIC2701)contains 0.5A types. Texas Instruments, 0234 223252.

PASSIVE

Passive components

Encapsulated transformers. For switched-mode power supplies, MicroSpire has introduced the MPMP series of transformers and MPLP inductors, both of which offer high power in a small size; 40W from 3400 cubic millimetres, for example. Height is as little as 5mm above the board and, since the core material touches the board, it acts as a heat sink, so that the units withstand 150°C. Microspire UK Ltd, 0227 740368.

Tantalum chips. Ultra-miniature tantalum-chip capacitors in NEC's SVS range measure 2 by 1.2 by 1.25mm, but come in values from 0.33µF to 2.2µF at voltages from 2.5V to 16V DC. Leakage current is 0.5µA maximum with a dissipation factor of 0.1. The existing R series covers the 0.047-330µF range at up to 50V DC. NEC Electronics (UK) Ltd, 0908 691133.

Electrolytics. Nichicon's Muse electrolytic capacitors come in nine varieties, all of them intended for audio use. Range of values is 0.1-22,000µF at 4-100V DC. The ranges have various levels of use: KZ, for example, is meant for high-quality audio, ES is non-polarised, FA is for mid-to-top quality work and KV is a 6mm long chip type. Nichicon (Europe) Ltd, 0276 685393.

Lithium batteries. High-performance manganese dioxide lithium backup batteries in button form for 3V operation are available from P Caro. Six sizes from 25mAh to 500mAh provide continuous current of 2mA to 20mA. P Caro & Associates Ltd, 021 742 1328.

Current limiters. Surge-Gard current-limiting devices from Rhopoint handle steady-state current from 0.3A to 30A and are meant for use in bridge power supplies to limit the inrush current for up to two seconds, whereupon their resistance decreases so that voltage drop is negligible. Several may be placed in series. Rhopoint Components, 0883 717988.

Chip capacitors. Solid tantalum chip capacitors in the Matsuo Tanchip *267* range are resin moulded and designed for surface mounting, being as small as 3.2 by 1.6mm and precisely sized. Capacitance range is 0.1µF-100µF at 4-35V, or up to 220µF in a larger case. Operating temperature is –55°C to 125°C. Surtech Interconnection Ltd, 0256 51221.

Connectors and cabling

Edge connector. Methode has an edge connector with a 0.05in pitch and 0.3in off-board height. Between 20 and 132 beryllium-copper pins are mounted on a 0.05 by 0.1in grid in a heat-resistant polyphenylene sulphide body. ITT MULTIcomponents, 0753 824212.

Filtered BNCs. Noise problems in coax. transmission lines are reduced by Oxley's BNC connector with built-in filtering. The cable outer is not connected to chassis, avoiding ground loops, integral multilayer ceramic capacitors providing AC coupling, so that noise currents on the screen can be decoupled to ground. VSWR is better than 1.3:1 over 0-4GHz and, with 10nF between cable screen and equipment shield, attenuation is over 30dB. Oxley Developments Co. Ltd, 0229 52621.

Euro V35 connector. St Cross now stocks the Belgian CDM range of products, including the CDM-V.35 connector designed for the V.35 CCITT interface, which incorporates strain relief and RFI/EMI shielding. St Cross Electronics Ltd, 0703 227636.

Displays

Panel meters. Digitron's new *440* series panel indicators work with a variety of sensors, for which the indicators provide a 12-24V supply. Model *445* is a 3.5 digit led type, while the *448* also has a single set-point changeover relay. The units are approved to BS 5750. Digitron Instrumentation Ltd, 0992 587441.

Slim displays. Epson's *EG8501* graphics displays have 320 by 240 pixel screens, the pixels being 0.36mm by 0.36mm. Total size of the units is 154mm by 110mm by 18.1mm, active area taking up 115mm by 86mm. Displays come in blue and black-on-silver with backlights and a reflective silver. Hawke Components Ltd, 0256 880800.

Filters

UHF filters. Toko UHF dielectric filters are now available in surface-mount packages. They are meant chiefly for RF front-end filtering in cellular and cordless telephones and GPS, with centre frequencies in the 800-1700MHz range and ±12.5MHz bandwidth. Insertion loss of the two and three-pole types is 2dB, pass-band ripple less than 1dB and VSWR 2dB. Cirkit Distribution Ltd, 0992 444111.

Satellite filters. Four new satellite filters from Matthey are meant for use in SNG video upconverter equipment and professional satellite receivers. Centre frequency is 70MHz and the filters conform to Eutelsat amplitude and group delay shaping characteristics for both full and half transponder use. Matthey Electronics, 0782 577588.

Hardware

Air packing. *Air Box* is a double-walled plastic bag for transporting delicate components. When the package is inflated, the inner bag contracts and is surrounded by air under pressure, the whole then being put into a carton. The contents can be seen before opening. The Post Office recognises the *Air Box* for sending blood products. The bag can be inflated by mouth or by air line. Air Bag Packaging Ltd, 0789 297000.

RF shielding material. Available in a number of forms, EMI shielding material from RS is made from copper-coated non-woven nylon fabric, in self-adhesive and non-adhesive varieties. There is a 25mm door-seal tape in gasket form, laminated to a PVC foam strip. All types attenuate by 60dB in the 1MHz-10GHz range and have a typical surface conductivity of 0.04Ω per square. RS Components Ltd, 0536 201234.

Instrumentation

Subminiature DVM. Datel's *DMS-20PC* is a 3.5-digit digital voltmeter that occupies around 0.5 cubic inch. Thirty-six models cover the range ±0.2-200V, with 1000MΩ inputs, autozero and autopolarity. Display is by 0.37in leds and the units are surface-mounted or panel-mounted. Datel (UK) Ltd, 0256 880444.

20MHz oscilloscope. Hitachi Denshi's *V-212* 20MHz oscillscope costs only £365, yet offers all the usual facilities: Alternate and chop sweep at all speeds, Ch1 output, triggering vert mode, which allows examination of signals at different repetition freequencies on Ch1 and Ch2 simultaneously. Vertical sensitivity is 1mV/div. with magnification; rise time is 17.5ns and sweep time is 0.2µs/div. to 0.2s/div., or 0.1µs/div. uncalibrated. Hitachi Denshi (UK) Ltd, 081-202 4311.

DAT instrument recorder. Sony's *PC116* instrument recorder uses digital audio tape techniques for increased accuracy. Analogue input is converted to PCM digital form, which is then recorded without the corruption that sometimes accompanies analogue recording and with a recording time of 120min. Sixteen channels have a dynamic range of over 80dB. Livingston Hire Ltd.

Vector wattmeter. Simultaneously performing 144 quantities, Infratek's *305A* vector wattmeter works in the 0-800kHz range up to 16kV and 800A, in single. two and three phase versions. It provides direct indication of current, voltage, power, apparent power, reactive power, impedance, energy and frequency, determining harmonic current and voltage and showing power in vectorial form for harmonics to the 59th. Versions for transformer, motor and burden test are available. Tandem Technology Ltd, 0243 576121.

Interfaces

Networks and PCs. For networked keyboard users who need to use PC facilities such as Windows and MS-DOS, Cherry has the *G80-2550* IBM-compatible keyboard, which combines the functions of an MFII keyboard for PCs with those provided by a data-entry device for IBM hosts. Cherry Electrical Products, 0582 763100.

Literature

Test gear. New instruments in the Alpha product guide include power analysers, power supplies and oscilloscopes from Tektronix and Hameg, together with equipment by Fluke, Avo, Edgcumbe, Robin, Kewtechnic, Seaward and Thurlby Thandar. Alpha Electronics plc, 0942 873434.

Reflection coefficient. Marconi now has the *6210*, an add-on to the *6200* series microwave test set family of instruments, which is designed to provide precision reflection analysis on microwave and RF systems and components in the 250MHz-26.5GHz range. Time-domain analysis is also offered by way of a simplifying user interface. Display formats include phase, polar, real, imaginary and Smith chart, in addition to normal Cartesian return loss and VSWR. Accuracy is such that return loss measurement of 20dB is accurate to within 0.5dB. Marconi Instruments Ltd, 0727 59292.

Cirkit catalogue. Cirkit's Summer 1993 catalogue includes details of Kenwood oscilloscopes, new low-cost alarms, scanning receivers and new kits by Velleman. The catalogue costs £1.90 direct from Cirkit or newsagents. Cirkit Distribution Ltd, 0992 444111.

PSU handbook. Computer Products' 1993 power-supply handbook covers AC/DC supplies and DC/DC converters. It also has a technical section on recent advances in design, with a multilingual glossary of terms. There is an offer of free technical handbooks on safety and EMC and

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thermal management. Computer Products, 0494 883113.

Audio amplifiers. Philips Semiconductors offers the Audio Amplifier Cook Book, which lists audio power devices from 150mW to 50W and contains over 100 pages of application notes, including PCB design. Gothic Crellon Ltd, 0734 788878.

DSP. The DSP Catalogue gives details of what LSI claims to be the world's most comprehensive range of digital signal processing and data acquisition boards and software support, including products for VMEbus, PCbus and Sbus. There are also specifications of all the leading DSP chips currently available, including the TI *TMS320C40* and Motorola's *DSP96002*. Loughborough Sound Images Ltd, 0509 231843.

Radio communications products

Microwave link. Designed for secure, point-to-point Ethernet and voice, Digilink 60 from Wadsworth can be installed in less than a day and offers communication between sites up to 1km apart. Beam width is 2deg for invulnerability to movement. Two versions operate at 2Mb/s and 10Mb/s and frequency is 60GHz, at which the equipment is said to be completely safe; an operating licence is supplied with the units. Wadsworth Electronics Ltd, 081-941 4716.

Transducers and sensors

Gas sensors. The Capteur G series of semiconductor selective toxic gas sensors measure ammonia and hydrogen sulphide concentration in the 0-100ppm and 0-1000ppm ranges. Unusually, the sensors exhibit a resistance increase in the presence of the target gases, with negligible interference from methane and other toxic gases. Contained in one cubic centimetere, they can be PCB-mounted or in flame-proof enclosures. Circuit diagrams or ready-built interface boards are offered. Capteur Sensors & Analysers, 0235 821323.

Gas sensor modules. Only a 5V reference is required by Nippon Ceramic gas sensor modules to convert gas concentration into voltage. An output indicator completes the imstrument, which measures air cleanliness of 100-3000ppm hydrogen or tabacco smoke, 0.1-10ppm nitrous oxide, 10-500ppm alcohol or 10-100ppm organic solvents. Chartland Electronics, 0372 363666.

Optical encoders. Codechamp optical encoders are now distributed by Pandect. Specifications include a maximum resolution of 22bits/revolution with an accuracy better than 2 seconds of arc. Units are discrete or in slip-ring assemblies. Pandect Precision Components, 0494 526303.

Winding temperature

Sensors. Sensor elements designed to measure the temperature of motor and generator windings are fitted in place of a stator slot separator and varnished in. The sensors use a bifilar winding on a glass former, the whole encapsulated in a silicon glass board. Twisted PTFE leads are provided. Class H insulation withstands 2kV RMS for a minute. Range of sizes is up to 700mm long, 50mm wide and anything over 2mm thick. Thermocouple Instruments Ltd, 0222 734121

COMPUTER

Development and evaluation

Embedded computer development. Arcom's *Fastcycle-88* is a multi-tasking STEbus processor board with the SourceVIEW PC-based debug environment. It works with Borland C or Pascal compilers, the combination optimising Borland's Turbo Debug to produce code for embedded systems. Arcom Control Systems Ltd, 0223 411200.

Fuzzy logic toolset. The Togai Infralogic Fuzzy Logic toolset enables system development for a range of microcontrollers such as the 8051, 68HC11 and 80C166. Resulting fuzzy rule base can then be integrated with C code, which can also form part of the rule base, whose output is a defuzzified variable for use in the application. A simulator proves the system, which is then compiled to C code for the processor. A free demo version is available. Hitex (UK) Ltd, 0203 692066.

CAN development. Hitex has a range of development tools for the Controller Area Network protocol, in software and hardware. The range includes the Keil C51 Ansi C compiler, Keil *RTX51/CAN* operating system for the 8051, a Phytec microcomputer card, Hitop51 source-level debugger and the Hitex Teletest 51 emulator. Hitex (UK) Ltd, 0203 692066.

PLD development. *PDS-1016* software from Micro Call is an entry-level system to enable full implementation of designs with Lattice *pLSI* and *ispLSI 1016* 110MHz programmable logic devices. It runs on 386/486 PCs under Windows 3.0 or later, using Boolean logic equations and TTL-like logic macro entry. Micro Call Ltd, 0844 261939.

Data acquisition. New features of the latest Windmill data acqusition and control software include alarm indication on all inputs, up to eight channels on each window showing the chart recorder and up to 100 channels for each copy of the data logger. A graphics program allows custom displays to be designed and another enables RS232 interfacing. The package can be linked to *Windows* spreadsheets, word processors or database software. Windmill Software Ltd, 061 883 2782.

Computer peripherals

Shielded VME bus. Elco's VME bus system provides shielding and data protection and is available in up to eight-layer form for use in every known bus application. Press-fit, multilayer back panels make for optimum capacity and cross talk to adjacent pins is 350mVSS. Ceramic capacitors between slots reduce LF disturbances. Elco Europe Ltd, 0638 664514.

Software

Antenna calculation. A set of programs called RF Tools assists in the production of critical base antenna calculations and tailoring to special requirements. DXPLOT allows precise calculation of beamtilt coverage; PATPLOT displays digitised base antenna patterns; and ANTPLOT develops patterns for side-mounted base antennas. The programs come on three 5.25in disks, but can be down-loaded free from the company's bulletin board, of which the modem format is 300/1200/2400/9600-N81; telephone (US) (216)349-8698. The Antenna Specialists Co., (US) 0101 216 349-8400.

PCB design. *TangoPRO* PCB board design software now runs under *Windows*, a cut/copy/paste capability allowing selected items to be moved to and from the *Windows* clipboard. Design errors are shown on screen and there are enhanced report formats and improved attribute editing. Redraw, file loading and move/delete are now faster. *TangoPro Route*, also running under *Windows*, allows simultaneous routeing on all layers, true 45° routeing, manufacture improvement passes and copper sharing. Pentica Systems Ltd, 0734 792101.

RF amplifier plotting program. Motorola's Plotting Utility Program disk enables the plotting of RF amplifier characteristics on Smith chart or polar plots. It will plot input and output stability circles, transition frequency against log frequency and maximum gain in dB against log frequency. Needs are an IBM-type PC with a 5.25in, 1.2M drive, a VGA display and dos. The program disk is free. Motorola Inc., (USA) 602 994 6561.

CIRCLE NO. 135 ON REPLY CARD

APPLICATIONS

Are 4Mbit drams a drop-in upgrade?

	4M		1M	
	Normal Power	Low Power	Normal Power	Low Power
Number of rows	1024	1024	512	512
Number of bits per row	4096	4096	2048	2048
Refresh period (t_{RFSH})	16ms	128 ms	8 ms	64ms
Distributed refresh period	15.6µs	124.8µs	15.6µs	124.8µs
Burst refresh period	16ms	128ms	8ms	64ms
Time to refresh 1 row (t_{RC})	130 ns	130 ns	130 ns	130 ns
Cumulative time to refresh entire array	133.1µs	133.1µs	66.6µs	66.6µs
Refresh time/operating time	0.833%	0.104%	0.833%	0.104%

In most respects, industry-standard 4Mbit dynamic RAMs are a drop-in replacement for their 1Mbit predecessors. Minor refresh and power-up incompatibilities can cause problems however. This table compares standard 70ns parts.

One megabit drams with pin designations to jedec and EIAJ standards were designed with upgradability in mind. This applies to both small-outline J-lead and zip package formats. In theory, any PCB layout for a standard 1Mbit chip will work equally well with a 4Mbit upgrade without needing any modification, assuming only 1Mbit needs to be addressed of course.

Standardisation fulfils two needs. Firstly, upgradability is simplified. The only circuit modification needed to address a 4Mbit chip as opposed to a 1Mbit device is to connect an extra multiplexed address line (for the additional two bits) to a previously unused pad. Secondly, standardisation ensures that equipment designs performing adequately with 1Mbit chips will not be forced into redesign or obsolescence when the unit price of a 4Mbit ram chip falls below that of its 1Mbit predecessor.

Electrically however, compatibility is not 100% in all applications. Motorola application note *AN1124* discusses possible incompatibilities with refresh and power-up sequences. Both incompatibilities exist because the 4Mbit dram has a different test-mode entry sequence from its predecessor.

Because 4Mbit drams have 1024 rows as opposed to 512 for a 1Mbit device, refreshing takes twice as long. Refresh versus operating time is however the same, as the table shows. In common with many contemporary dynamic rams, both 1Mbit and 4Mbit industry-standard parts can be refreshed in a number of ways.

A potential incompatibility between the two device sizes can occur when CAS-before-RAS refresh is used. For a 1Mbit dynamic ram, the state of the write input is irrelevant during a CAS-before-RAS refresh. With 4Mbit devices however, the negated write input must be disabled, ie high, to prevent the device from entering the jedec standard test mode. Test mode on a 1Mbit memory is entered in an entirely different manner.

During power up, both 1 and 4Mbit drams need a 200ms pause followed by eight RAS cycles to guarantee proper operation. To prevent a 4Mbit ram from erroneously entering test mode however, these RAS cycles should be either RAS-only or CAS-before-RAS types.

Motorola Ltd, *European Literature Centre, 88 Tanners Drive, Blakelands, Milton Keynes MK14 5BP. Telephone 0628 585000.*

Battery life meter from Chinatown?

In the film Chinatown, Jack Nicholson placed a cheap analogue watch under a parked car's tyre so that he could return at his leisure to find out precisely what time the car departed.

This trick allegedly provided the inspiration for the circuit shown here. The design shown is intended to provide an accurate reading of battery life. Unlike its ancestor however, this new idea provides a read out without irreversibly stopping the time piece.

Essentially the clock shown in **Fig. 1** – a cheap analogue quartz movement - is powered until cell or battery voltage falls below a programmable threshold. Provided that the initial state of the clock is known, it records very accurately how long the battery has been in use.

Battery life is hardly affected by the metering circuit. Most quartz analogue clocks consume very little power, running for many months on one AA cell. The monitoring circuitry needs less than 20µA of supply current and gradually less as the battery voltage falls due to discharge.

To set the "battery-expired" point, a voltage equalling the minimum needed to power the circuit relying on the battery or cell is connected to the input terminals marked on the diagram. Next, the ten-turn potentiometer is adjusted to the threshold where the clock starts operating.

To commission the circuit, the note advises, "remove the power supply, set the clock to 12:00, connect the test circuit and go home". Presumably, connect the test circuit means connect the equipment that the battery normally supplies to the battery's positive and negative terminals.

There is no mention of connecting the same battery or cell to the input terminals shown on the diagram but it is difficult to see how the circuit would operate otherwise.

Among other circuits discussed in Maxim's Engineering Journal Volume 11 are a third-order high-pass filter incorporating a synthetic inductor and two circuits for converting 3V supplies to 5V. There is also a pulse stretcher for capturing pulses down to 15ns that, unlike most flip-flop alternatives, will respond to amplitudes down to 100mV.

The bulletin's in-depth article discusses new transconductance amplifiers, culminating in the op-amp circuit.

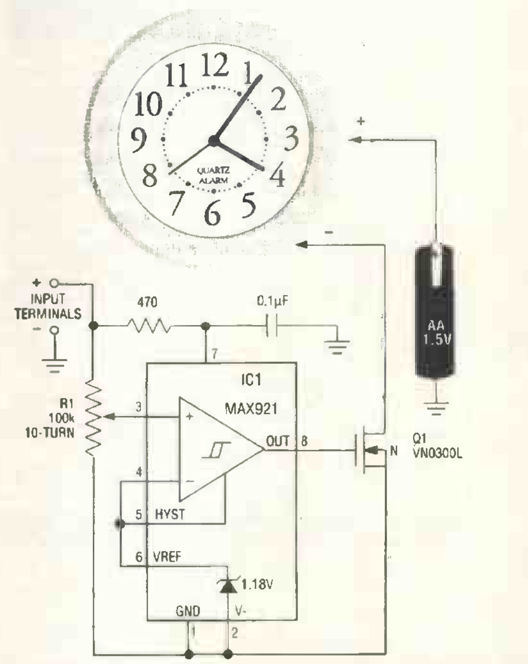


Fig. 1. Battery life can be indicated very accurately but cheaply using an analogue quartz clock. When cell or battery voltage falls below a threshold programmed via the ten-turn potentiometer, the clock stops. Provided that the initial time on the clock is known and battery life is less than 12 hours, this simple circuit provides a very accurate readout of battery life.

Op-amp transmits baseband video over 1500m

Although very easy to implement, the high-speed link shown in **Fig. 2** achieves exceptional performance. Using new high-performance transconductance op-amps, it provides an interface for transmitting baseband video via a twisted pair over distances up to 1.5km.

Twisted-pair cable is cheaper and more compact than its coaxial counterpart. On the other hand, twisted pair links involve transformers and have limited bandwidth. Without sacrificing the noise immunity of transformer-based twisted-pair transmission systems, this circuit signals down to DC without coupling transformers.

Operating with around 150m of 22 gauge (American) twisted pair intended for burglar-alarm wiring, the circuit attenuates a baseband video 3.58MHz colour burst by about 6dB, **Fig. 3**. In many cases this will not matter. If it does, overall brightness resulting from ohmic losses is adjusted via R_1 and bandwidth can be extended by trimming C_1. **Figure 4** compares a waveform resulting from R_1 and C_1 being in their nominal settings.

According to the note, stranded and unstranded twisted pairs exhibit similar bandwidths. Unshielded pairs with low dielectric constant polyethylene or polypropylene insulation offer the highest bandwidth. Gauge of the twisted pair becomes increasingly important as distance increases due to ohmic losses in the cable.

At the driver end, each cable is terminated by 50Ω to ground. Mismatches can degrade video quality but amplifier stability is unaffected since the *MAX435* has no feedback.

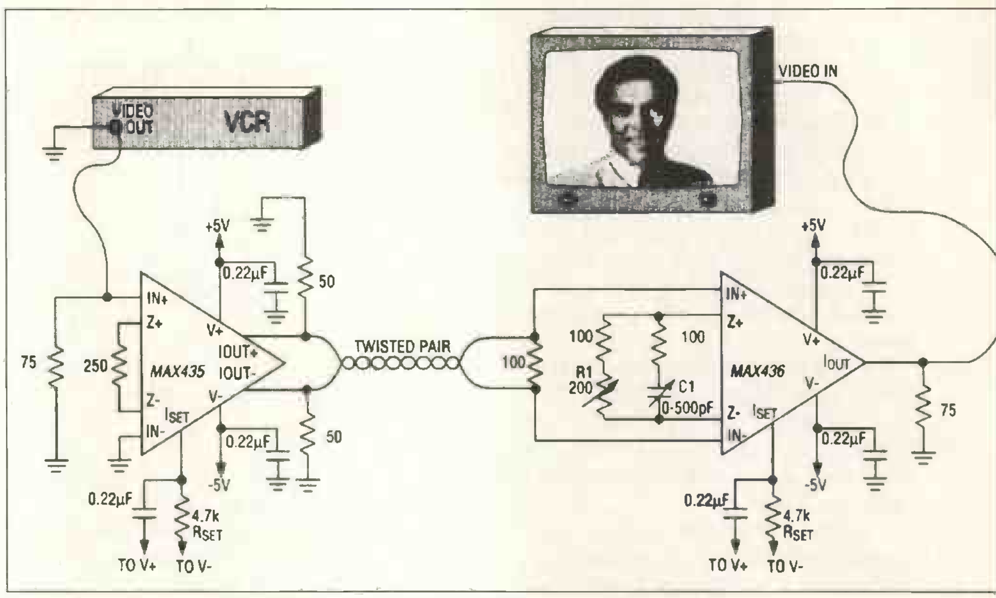


Fig. 2. Video and data transmission via coaxial cabling is expensive relative to twisted-pair alternatives. This simple transmitter and receiver communicates signals over a twisted-pair link up to 1.5km without the bandwidth restrictions of transformer coupling.

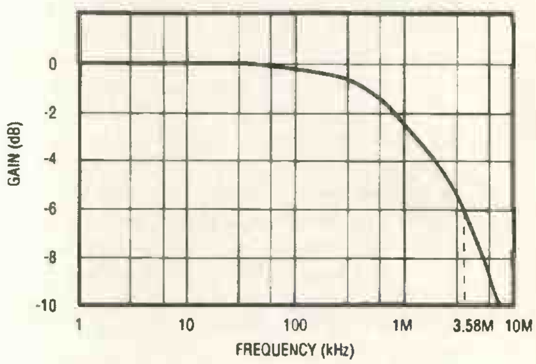


Fig. 3. Operating with around 150m of twisted pair intended for burglar-alarm wiring, the data transmission circuit attenuates a baseband video 3.58MHz colour burst by about 6dB.

Maxim Integrated Products, *21C Horseshoe Park, Pangbourne, Reading RG8 7JW. Telephone* 0734 845255.

Versatile fast-charger chip has only three pins

A single three-terminal IC, whose functional diagram is shown here, is capable of fast charging lithium, NiCd, NiMH or lead-acid batteries without any additional components. Called the *DS1300*, this IC holds its charging parameters digitally in user programmable non-volatile memory. As a result, it can be set up to produce charging characteristics for a wide variety of cell types and sizes.

Primarily designed with portable equipment in mind, the *DS1300* handles standard or trickle charging of batteries up to 3.7V, given a 5V input. With a 6V input, maximum battery voltage capability rises to 4.7V.

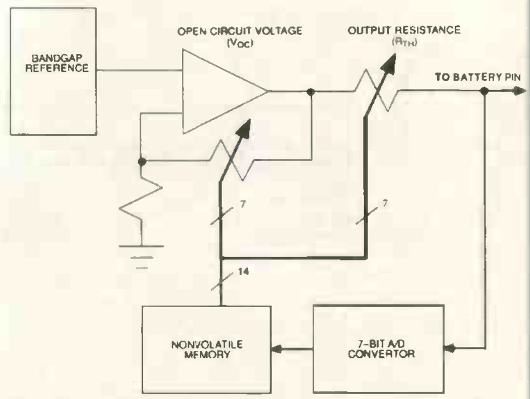
Throughout the charging cycle, output current can be directly related to battery voltage. With simple chargers, output current falls as battery voltage rises. Most batteries however can handle the same current at all stages in their charge cycle so the gradual decrease in current wastes time.

With the *DS1300*, output current versus battery voltage is programmable. The device can be programmed to provide constant-current charging until the battery voltage reaches a certain threshold. Alternatively it can be programmed to provide any charge current at any of 32 discrete battery voltages in roughly 37mV steps.

Being TO220 packaged, the device has only three pins – one connecting supply input, one feeding the battery and one providing a ground connection. For programming, the pins are fed with a special access sequence which initiates single-wire communications.

Once initiated for programming, the device can be fed with parameters including power supply range, charge-current load line and trickle-charge rate. An optionally selectable timer provides charge termination after a programmed period. There are also five fully preprogrammed options catering for triple NiCd packs charging at 20 to 100mA.

Dallas Semiconductor, *Unit 26, West Midlands Freeport, Birmingham B26 3QD. Telephone* 021 782 2959.



Housed in TO220 power-device packaging, this battery charger IC has only three terminals yet can be programmed to fast charge a variety of rechargeable technologies up to 4.7V. Five programmable parameters are charge on, pulse width, Thevenin resistance field, open-circuit voltage and breakpoint voltage. All are programmable in 32 steps. There is an additional register for limiting charge time from 2 to 32 hours.

Using PCB as a current shunt

To keep power losses to a minimum, current sensing shunts for mosfets passing tens of amps need to drop as little voltage as possible. Purpose designed power mosfet current sensors like the *LTC1154/5/6* can detect voltage drops of less than 100mV.

For a mosfet passing ten amps, the resistance needed to develop 100mV is just 0.01Ω. In this example, dissipation is 1W. Resistors with such a low value and capable of dissipating more than a watt can be difficult to obtain, particularly in surface-mount format.

For applications where the accuracy of a special current shunt in unnecessary, *Application Note 53* from Linear Technology suggests using a printed circuit board track to provide the necessary voltage drop.

Copper used for PCBs is often expressed in ounces of copper per square foot. The thickness of 1oz copper-clad board is approximately 0.00343cm. Resistivity of pure copper is 1.822μΩ-cm so the 'sheet' resistance of a 0.00343cm thick layer of copper is around 530μΩ/square. This means that a section of 1oz copper-clad board one unit long by one unit wide has a resistance of 530μΩ, regardless of the unit size.

Using 1oz copper board, a 0.01Ω resistor for sensing 10A will be approximately 20 squares long, i.e. a strip one unit wide by twenty units long. For 1oz board, a current density of 50A per 25mm width of copper is considered conservative. This means that a shunt 5mm wide by 100mm long should suffice.

From **Fig. 1**, you can see that four connections are made, as is common with power shunts. Two large terminals connect the supply to the power mosfet drain and two sensing terminals provide the voltage signal for the low-power circuit that senses current. As far as the voltage drop seen by the sensing circuit is concerned, only the distance between the two sensing terminals is important.

If a long stretch of PCB is unavailable, the resistor can be folded. In this case, the corner resistors should be counted as 0.6 squares since current concentrates at the inside corner, **Fig. 2**.

Copper has a rather high positive temperature coefficient of around 0.39%/ C. As a result, shunt resistance rises with increasing temperature. In some applications this may be a disadvantage but in others, it will be desirable to reduce the current limit as circuit board temperature rises. ■

Linear Technology, *111 Windmill Road, Sunbury-on-Thames, Middlesex, TW16 7EF. Telephone 0932 765688.*

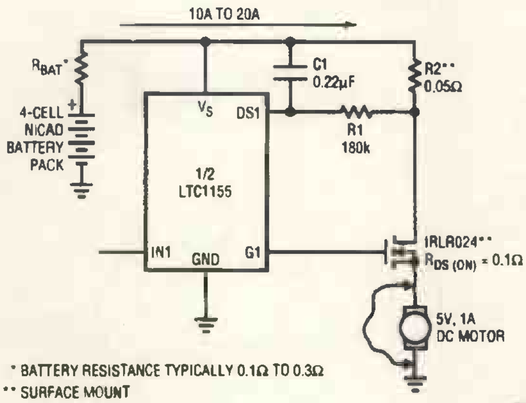
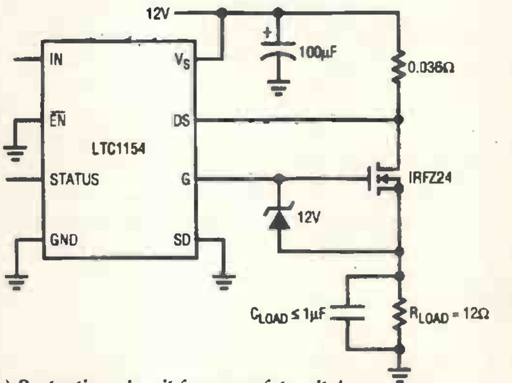
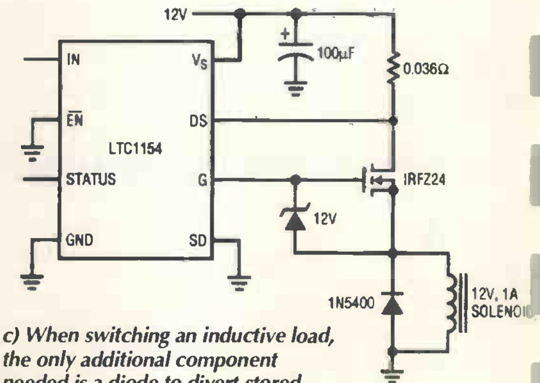


Fig. 3. While providing ideas for mosfet switching, these circuits illustrate designs where low-value resistors formed from copper on a PCB could be useful.

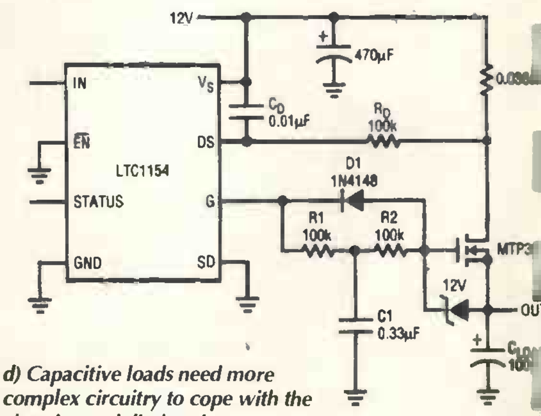
a) Here, a low-value resistor senses current to make sure that the mosfet is kept within its safe operating area.



b) Protection circuit for a mosfet switch handling a resistive load. A 10μs delay built into the drain sense circuitry eliminates false triggering due to power supply or load transients.



c) When switching an inductive load, the only additional component needed is a diode to divert stored energy to ground.



d) Capacitive loads need more complex circuitry to cope with the charging and discharging.

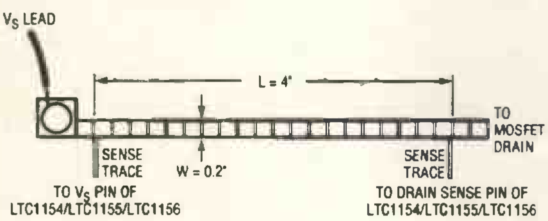


Fig. 1. Resistors with values below 0.1Ω are difficult to find, particularly in surface mount form with power ratings above a few hundred milliwatts. Circuit-board tracks form ideal power shunt resistors where accuracy and power dissipation are not too critical.

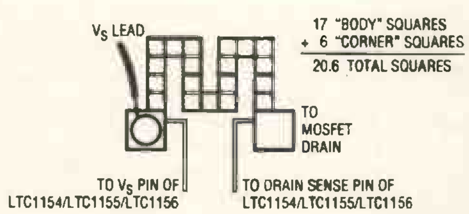
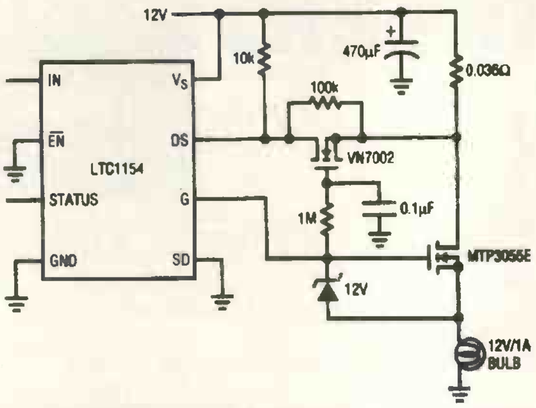
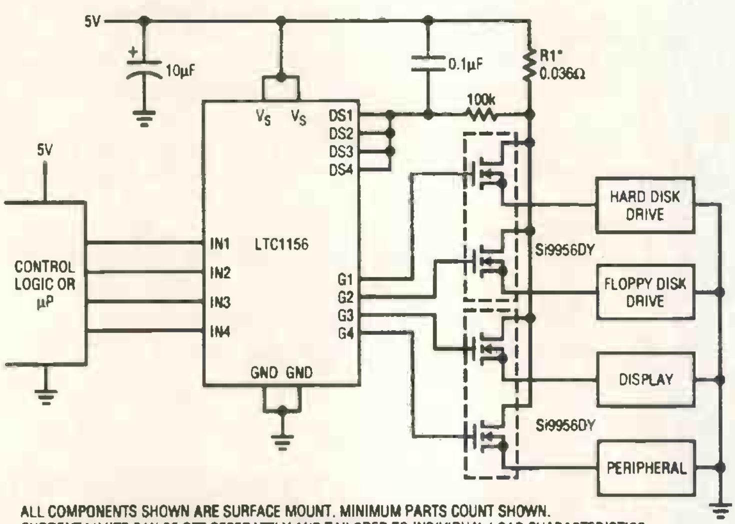


Fig. 2. Where board area available for a PCB track low-value resistor is limited, a serpentine layout may offer an alternative. Compared with the straight-line resistor layout, overall length needs to be increased since adjacent corner squares exhibit 0.6 the resistance of their linearly spread counterparts.

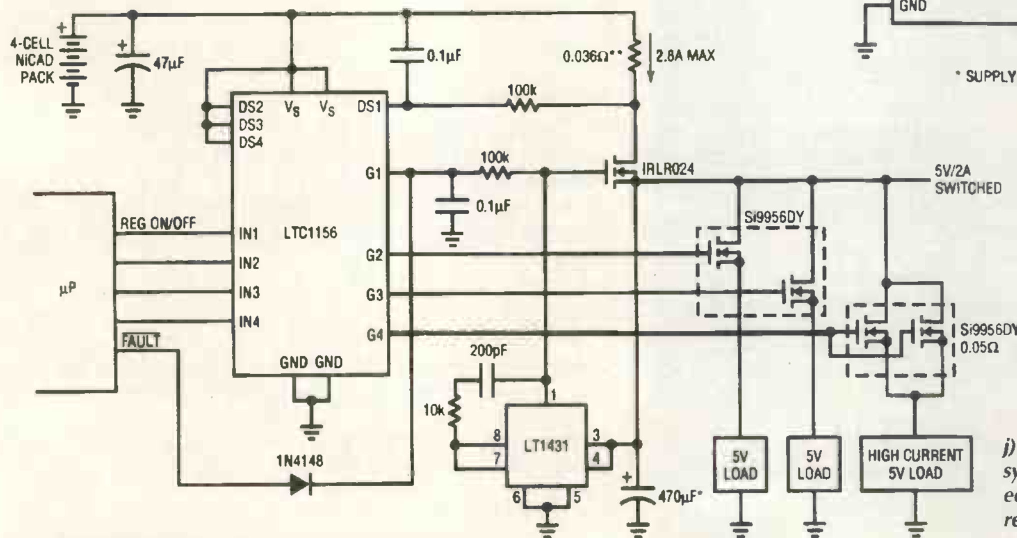


e) Inrush current when a lamp turns on can be up to 20 times the rated operating current. In this circuit, the current limit is raised at turn on.



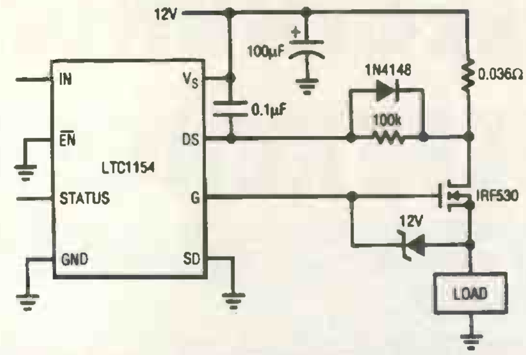
ALL COMPONENTS SHOWN ARE SURFACE MOUNT. MINIMUM PARTS COUNT SHOWN.
CURRENT LIMITS CAN BE SET SEPERATELY AND TAILORED TO INDIVIDUAL LOAD CHARACTERISTICS.
* IMS026 INTERNATIONAL MANUFACTURING SERVICES, INC. (401) 683-9700

i) Quad switch designed for a laptop computer. One sensing circuit controls all four outputs, although each may be switched independently.

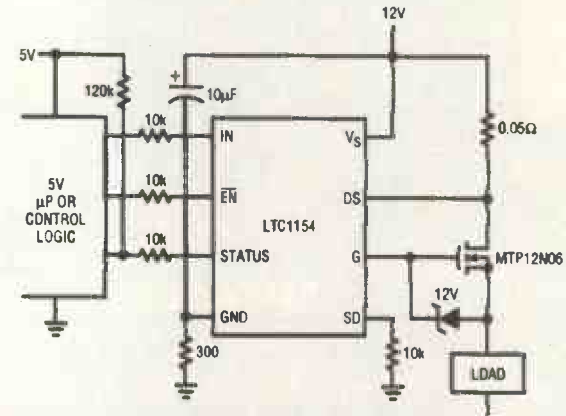


* CAPACITOR ESR < 0.5Ω
** IMS026 INTERNATIONAL MANUFACTURING SERVICES, INC. (401) 683-9700

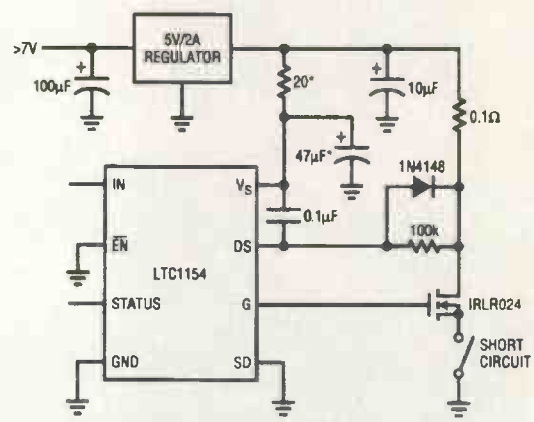
j) Complete power management system for battery operated equipment. By using an LT1431 reference as the heart of a regulator, a costly switching converter is avoided.



f) To reduce the amount of time that the power mosfet is in short-circuit condition, the delay resistor can be bypassed with a speed-up diode.



g) Simply connecting a resistor protects against reverse battery connection by limiting current.



* SUPPLY FILTER COMPONENTS

h) Adding power supply filtering to the IC feed ensures that the 3.5V needed for correct operation remains in the event of a momentary regulator overload.

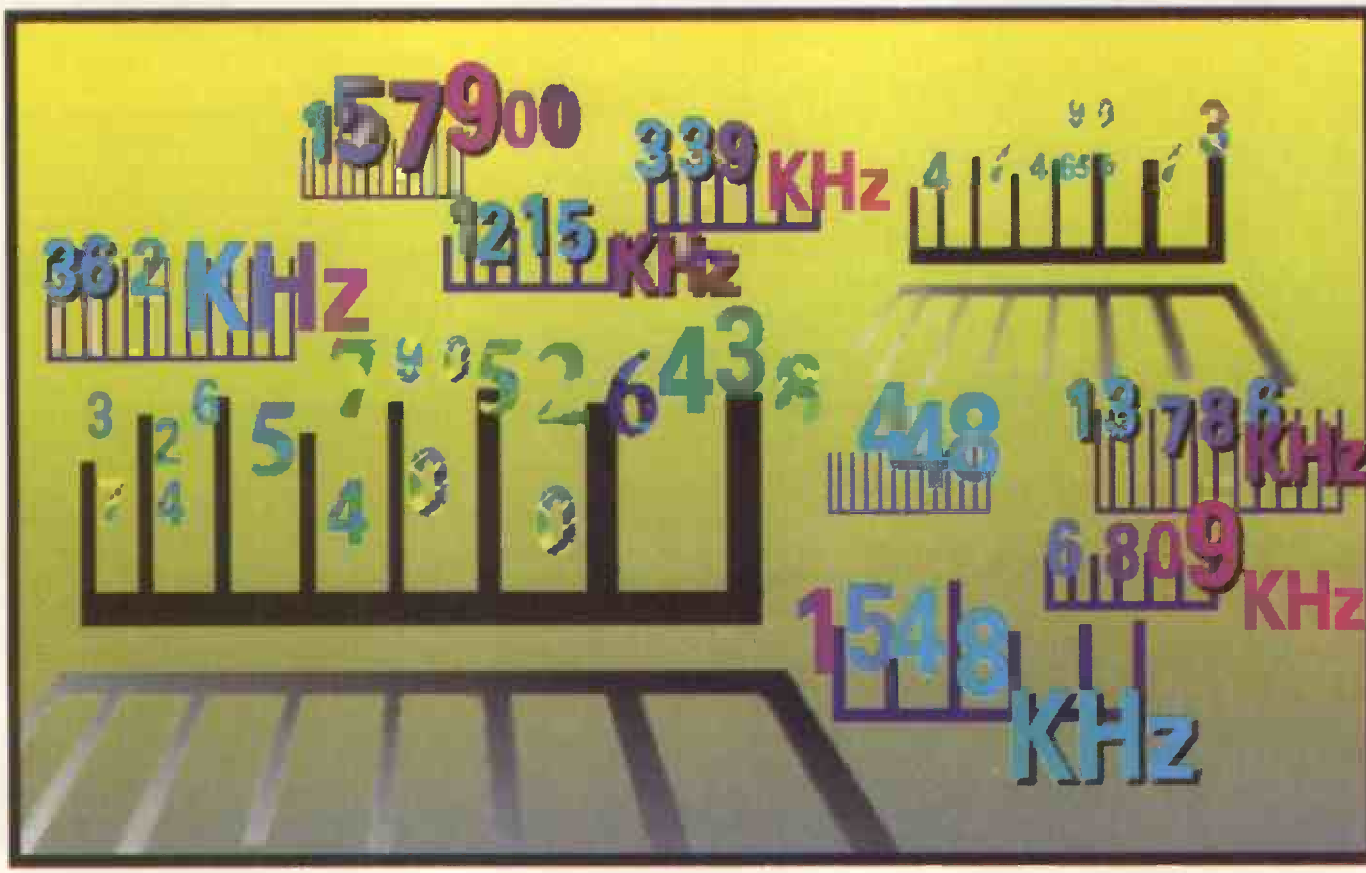


Illustration Tracy Martin

Implementing a band pass filter on a microprocessor

A hybrid comb FIR filter/second order IIR filter has useful benefits. Allen Brown shows how to implement the hybrid on a general purpose microprocessor.

Digital filters have become regular building blocks in the design of digital based systems. Remarkable roll-off and ease of design and implementation has led to their widespread use. Like their analogue counterparts, they are available in various forms and can suffer from the same problems – pass band ripple and ringing.

Fixed gain digital filters broadly fall into two camps, the finite impulse response (FIR) and infinite impulse response (IIR) filters.

An FIR filter is dependent on the current and the previous input samples: the IIR filter is not only dependent on the current and previous input samples but also on previous output samples. So it is recursive and can suffer from instability problems due to the feedback of output to input. FIR filters on the other hand are unconditionally stable (no feedback) but require more computation (more taps) to achieve the same performance as equivalent IIR filters.

The ideal would be to have the benefits of an IIR filter without the problem of potential instability, achievable by constructing a hybrid of a comb FIR filter and a second order IIR filter.

Under certain conditions this hybrid digital filter can be implemented on a general purpose microprocessor. Normally that is not an option since digital filters require intensive multiplications that cannot be performed efficiently on microprocessors. But the problem can be overcome by eliminating multiplications through ensuring that the digital filter only requires the numerical additions at which microprocessors excel.

A comb filter **(Fig. 1)** has an input sample

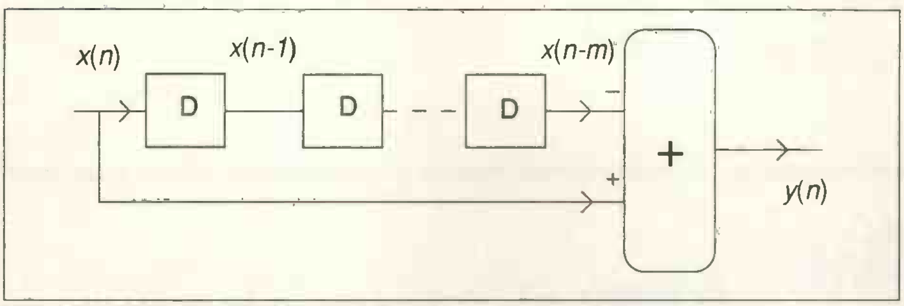


Fig. 1. A comb filter has an input sample sequence {x(n)} and an output sequence {y(n)} with m intermediate storage locations for the m previous input samples.

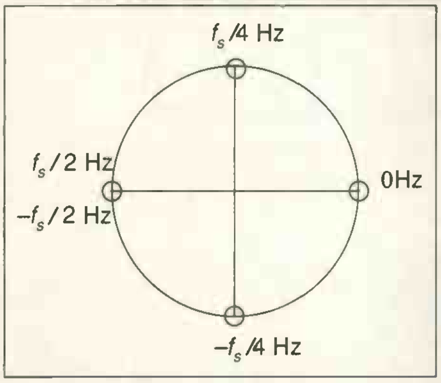


Fig. 2. Unit circle showing the positions of the four zeros and their respective frequencies.

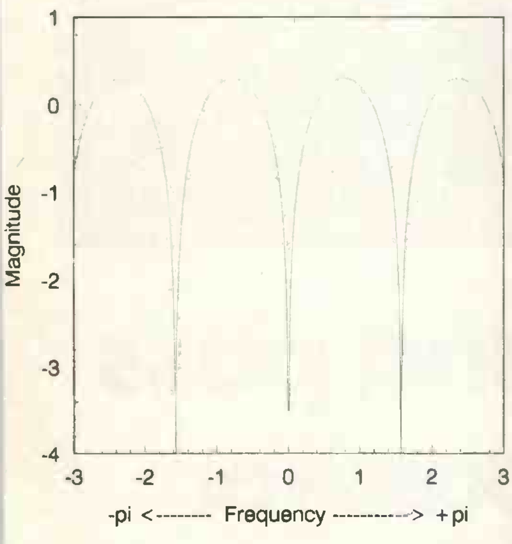


Fig. 3. Transfer function of comb filter with four zeros. Every point where the value of |H(q)| is zero corresponds to a zero in the unit circle.

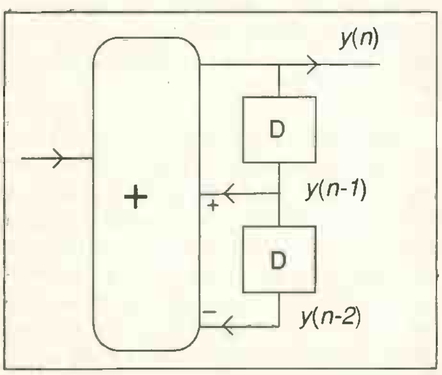


Fig. 4. Schematic representation of the second order recursive filter.

sequence $\{x(n)\}$ and an output sequence $\{y(n)\}$ with m intermediate storage locations (D memory shift registers) for the m previous input samples. An attraction of the comb filter is its non-recursive nature, in that there are no feedback components which can lead to instability in a filter's operation.

Actual output of the comb filter is the difference in the newest input $x(n)$ and the oldest $x(n-m)$ expressed as,

$$y(n) = x(n) - x(n-m) \qquad (1)$$

Now the behaviour of the filter must be considered in the Z domain, and for this the Z-transform needs to be invoked. Two properties of the Z-transform will be used,

$$Z\{x_1(n) + x_2(z)\} = Z\{x_1(n)\} + Z\{x_2(n)\} \qquad (2)$$

and,

$$Z\{x(n-m)\} = z^{-m}Z\{x(n)\} \qquad (3)$$

When the Z-transform is applied to Eq. 1, then

$$Z\{y(n)\} = Y(z) \text{ and } Z\{x(n-m)\} = z^{-m}X(z)$$

which leaves,

$$Y(z) = X(z) - z^{-m}X(z) \qquad (4)$$

where X(z) is the Z-transform of $x(n)$ and Y(z) is the Z-transform of $y(n)$. Eq. 4 can be written as

$$Y(z) = X(z)[1 - z^{-m}] \qquad (5)$$

The transfer function H(z) for the comb filter can be defined

$$H(z) = \frac{Y(z)}{X(z)} = 1 - z^{-m} \qquad (6)$$

Dividing the right hand side by z^{-m} this equation becomes

$$H(z) = \frac{z^m - 1}{z^m} \qquad (7)$$

But what are the conditions needed to ensure that H(z) = 0?: when the right hand side is zero. For this to happen

$$z^m = 1 \qquad (8)$$

When H(z) is zero it means that when a signal, whose frequency matches the value of z is fed into the filter, there is no output from the filter. In the Z plane, all the activity revolves around the unit circle, a circle with a radius of one unit (**Fig. 2**).

The east point of the circle represents 0Hz and the west point represents the maximum frequency which can be realised in a digital system, and that is half the sampling frequency.

Moving anti-clockwise along the perimeter from east to west is the positive frequency direction: moving east to west clockwise is the negative frequency direction. (In DSP, negative frequencies are just as real as positive ones.) The points at which the transfer function is zero are displayed as small zeros on the unit circle. These are frequencies where there is no output from the filter.

Taking a closer look at Eq. 8 shows that every time Eq. 8 is true then a zero will appear in the comb filter (a zero in the unit circle). From this equation there will be m zeros in the unit circle (each delay element in Fig. 1 will give rise to a zero in the unit circle). Remembering that z is a complex number with real and imaginary components, then z may be expressed as,

$$z^{m\theta} = e^{jm\theta} = \cos(m\theta) + j\sin(m\theta) \qquad (9)$$

For z^m to be a real, then the imaginary component in this equation must be zero. For this to occur then $m\theta = 2\pi$, since $\sin(2\pi) = 0$.

Our comb filter will therefore have m zeros, equally distributed on the perimeter of the unit circle. So for example if there were four zeros ($m = 4$), then

$$z^4 - 1 = (z^2 + 1)(z^2 - 1)$$
$$= (z + j)(z - j)(z + 1)(z - 1) \qquad (10)$$

The first zero at the south point ($z = j$, $-f_s$/4Hz), the second zero at the north point ($z = j$, $+f_s$/4 Hz), the third zero at the west point ($z = 1$, $\pm f_s$/2Hz) and the forth point at east point ($z = 1$, 0Hz). These are shown on the unit circle in Fig. 2. As the number of delay elements increases in the comb filter, more zeros appear on the unit circle, all equally spaced and closer together. To see what the transfer function of the comb filter looks like in real terms, use the complex conjugate:

$$H(z)\,H^*(z) = (z^4 - 1)(z^{-4} - 1)$$
$$= 2 - (z^4 + z^{-4}) \qquad (11)$$

Since $z^4 = e^{j4\pi}$

$$\cos(4\theta) = \frac{e^{j4q} + e^{-j4q}}{2}$$

then

$$|H(\theta)|^2 = 2\{1 - \cos(4\theta)\}$$

A plot of H(θ) is shown in **Fig. 3**, showing

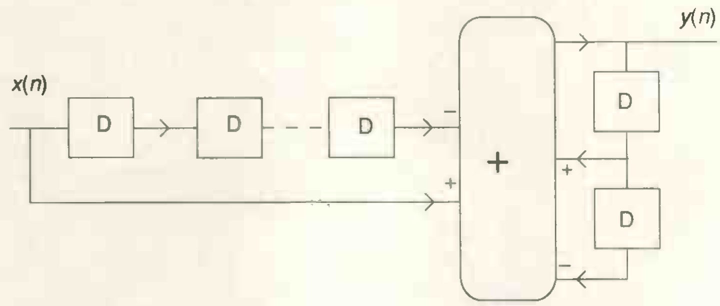


Fig. 5. Combining the comb filter with a second order recursive filter.

why the comb filter is so called. Every point where the value of $|H(\theta)|$ is zero corresponds to a zero in the unit circle.

We are now going to use a comb filter for $m = 12$ in combination with a second order recursive filter which has conjugate poles. Remember the poles in a filter gives rise to its resonance behaviour.

Consider a second order filter whose transfer function is,

$$H(z) = \frac{z}{z - e^{j\alpha}} \frac{z}{z - e^{-j\alpha}}$$

$$= \frac{z^2}{z^2 - 2\cos(\alpha) - 1} \qquad (13)$$

The filter will have a second order zero at the origin, but we are now interested in the behaviour of the poles which are on the perimeter of the unit circle. The condition for the filter to remain stable is that its poles must remain within the unit circle. This filter is on the margin of stability, if it is subjected to an impulse it will be excited into an oscillating condition. Now look at the case when $\alpha = 60°$ then $2\cos(\alpha) = 1$, the above equation will therefore become,

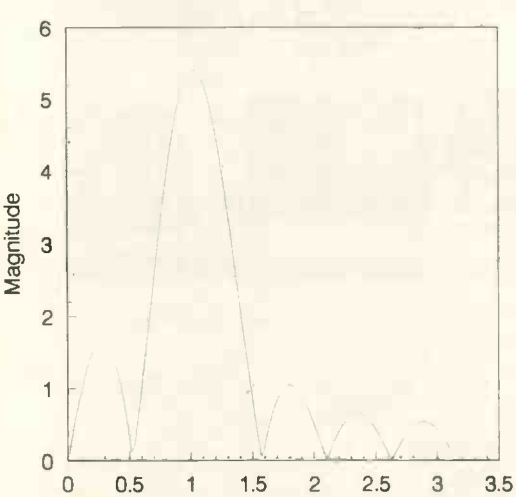


Fig. 6. Transfer function of band pass filter (order 12).

$$H(z) = \frac{z}{\left(z^2 - z + 1\right)}$$

$$= \frac{z^2}{(z - 0.5 + 0.866j)(z - 0.5 - 0.866j)} \qquad (14)$$

with pole values, $z_1 = 0.5 - 0.866j$ and $z_2 = 0.5 + 0.866j$, where $|z_1| = |z_2| = 1$ which shows that the conjugate poles are on the perimeter of the unit circle. Performing an inverse Z-transform on this equation, we can rewrite the equation as,

$$H(z) = \frac{Y(z)}{X(z)} = \frac{1}{1 - z^{-1} + z^{-2}} \qquad (15)$$

Rearranging the terms, we get,

$$Y(z)[1 - z^{-1} + z^{-2}] = X(z) \qquad (16)$$

noting that the inverse Z transform will give,

$$Z - 1\{Y(z)\} = y(n),$$
$$Z - 1\{z^{-1}\,Y(z)\} = y(n-1)$$
$$Z - 1\{z^{-2}\,Y(z)\} = y(n-2)$$

then the difference equation for the recursive filter becomes,

$$y(n) = y(n-1) - y(n-2) + x(n) \qquad (17)$$

which can be represented schematically as shown in **Fig. 4**. Now what has to be done is to combine this second order recursive filter with a comb filter and ensure that the two poles reside on top of two of the zeros of the comb filter. The operation will cancel out the effects of the poles and the two zeros under them. To do this we can choose a comb filter of order 12 since two of its zeros will reside at an angle of 60°, the same angle as the recursive filter. The resulting system will be non recursive, having the advantage of being unconditionally stable. Schematically the combination is shown in **Fig. 5**. The difference equation for this hybrid filter is,

$$y(n) = y(n-1) - y(n-2) + x(n) - x(n-12) \qquad (18)$$

A program in *C* to implement this filter is given in **Listing 1**.

The transfer function for this expression is

```
/* Program listing for the coding of
Eq.18 in C

In this coding the arrays are
defined, y[0] = y(n), y[1] = y(n-1),
y[2] = y(n-2) and x[0] = x(n), x[1]
= x(n-1), .... x[12] = x(n-12). The
input and output instructions must
be added to the code depending on
the source and destination of the
data. If used with a data
acquisition card there will probably
be call instructions to perform the
I/O operation.*/


#define M 12             /* Size of
the Comb Filter */

int x[M+1], y[3], n, i ;

main()
{
        /* Zero the shift registers
*/
        y[1] = y[2] = 0 ;
        for( i = 0 ; i <= M ; i++)
        {
                x[i] = 0 ;
        }
        do
        {
                /* instruction to
input data sample and equate to x[0]
*/
                y[0] = y[1] - y[2]
+ x[0] - x[M] ;
                /* instruction to
output data sample y[0]
                update data array
*/
                for( i = M ; i > 0
; i-)
                {
                        x[i] = x[i
- 1];
                }
                y[2] = y[1] ;
                y[1] = y[0] ;
         } while( abs(y[0] < 32788
);  /* If an overflow occurs the
program stops */ }
```

shown in **Fig. 6** which clearly demonstrates the band pass nature of the filter. The greater the order of the comb filter the narrower the band-pass filter. Some problems may arise when trying to implement this filter due to the truncation errors which occur in fixed point processors. The poles may not exactly reside over the zeros which can give rise to some oscillation in the output.

Using a microprocessor

When implementing the band pass filter structure it is possible to choose an appropriate sampling frequency which avoids the need to perform multiplication as seen in Eq. 18. If this is the case the filter can effectively be performed by a general purpose microprocessor, very efficient at moving data and performing arithmetic additions of fixed point numbers. The option is a very cheap one and is ideal for low frequency (< few kHz) applications ■

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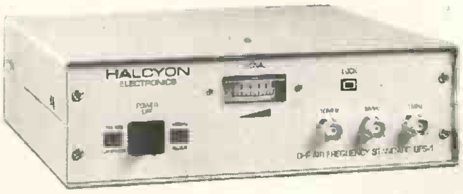

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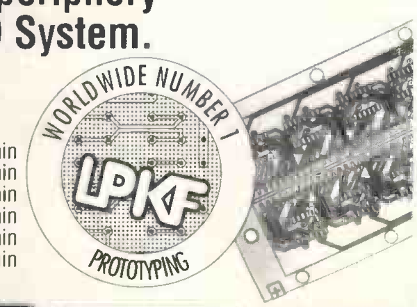

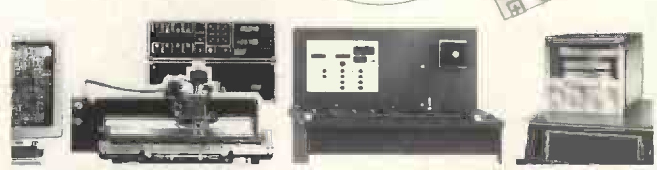

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CIRCLE NO. 123 ON REPLY CARD

Is there life in cold fusion after Fleischmann and Pons? Andy Wright reports on a growing scientific community that says yes.

Clawing back respectability for cold fusion?

Cold fusion looked dead. Fleischmann and Pons were discredited, the public had lost interest, and within a year work published on the subject had been cut to a trickle. Science magazines were jostling to write obituaries.

But now the technology seems to have come back to life, with scientists from Europe, the US and Japan revealing evidence showing that perhaps the two professors were on the right track.

Following the Fleischmann and Pons debacle (see box) cold fusion was starved of funds, with work occasionally carried on in researchers' spare time, and sometimes lacking the scientific rigour demanded by such a controversial topic. But at the *Third International Conference on Cold Fusion*[3]. heretics with substantial academic credentials were presenting credible results indicating both the generation of excess heat and nuclear products.

Excess heat

M C H McKubre's group at the Stanford Research Institute in California is funded by the Electric Power Research Institute. Using sealed and pressurised electrochemical cells, they observed excess power between 2 and 50% of the power they put in – with occasional bursts of 350 to 500% of power-in. The excess only occurred when the palladium cathode was highly loaded with deuterium – specifically when x is greater than approximately 0.9 in the beta phase of the palladium-deuterium system PdD_x. Other factors are also implicated, but these have not been characterised in detail.

At face value, there seems little room for error. Some 17 electrochemical variables were carefully controlled and the reaction parameters recorded on-line. Great care went into the cell design and electrochemical aspects of the loading process gained close attention. In particular, deuterium loading was carefully monitored *in situ* for the duration of the test so the team could be sure that this was indeed a key condition for excess heat.

McKubre's team is not alone in reporting excess heat. In similar experiments, a group from Osaka University in Japan has recorded 70% excess heat from a cell with a palladium plate cathode. Other experiments using the same method (by E Celani of INFN Laboratory, Frascati, Italy) have also generated 25% excess power.

So, from the Los Alamos National Laboratory in New Mexico, through IMRA in Japan to IMRA SA, France, scientists are independently gaining tantalising glimpses of the effects first noted by Pons and Fleischmann.

Apparently, high deuterium/palladium loading ratios are necessary to create excess enthalpy. But it is also clear that this is not the only significant variable – and still the results are not repeatable on demand.

Fusion physics

Several nuclear fusion reactions have been proposed as cold fusion candidates.

Deuteron-deuteron fusion

$d+d \rightarrow$ 3He(0.82MeV)+n(2,45MeV)

$\rightarrow$ t(1.01MeV)+p(3.02MeV)

is one possibility, and some authors have claimed to have observed characteristic 2.5MeV neutrons in cold electrolytic experiments.

More likely at low temperatures is the proton-deuteron reaction

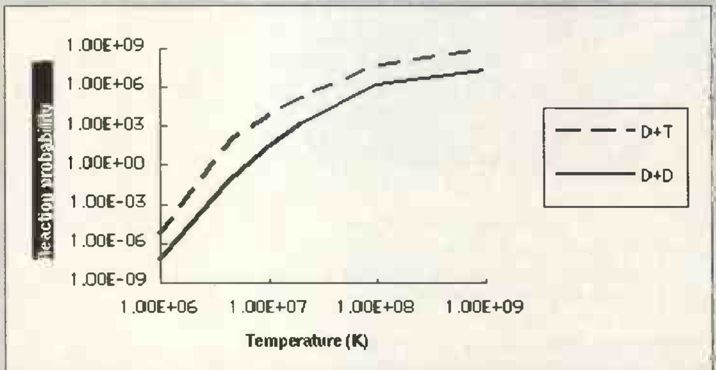
$p+d \rightarrow {}^3He+g$ (5.49MeV)

calculated to proceed at rates more than eight orders of magnitude faster than d+d. Observation of 5.49MeV gamma rays should therefore provide a stricter test of cold fusion than neutron measurements

Deuteron-triton reactions may account for the emission of ^4He, through the mechanism

$d+t \rightarrow n+{}^4He$.

As illustrated in the graph, such fusion has a higher cross-section than d+d.



Text-book nuclear fusion at extremely high temperatures. Reaction probability per second is shown for density one gram-atom per cm³ for each reacting nucleus. For example two gram atoms D/cm3 for D+D, one gram atom T + one gram atom D for D+T.

Furthermore, an explanation involving nuclear reactions requires other corroborative evidence.

Nuclear products

Direct observation of nuclear products at ordinary temperatures is more convincing evidence for fusion. Soon after the Pons-Fleischmann bombshell, fusion-generated neutrons were indeed observed in electrolytic cells[1], though the validity of these observations remains in doubt.

Nevertheless the work has continued. A collaboration between Provo Canyon Laboratory, Utah and the University of Tokyo has recorded small steady emissions of neutrons in an underground laboratory. In Osaka and Rome teams have reported neutron emissions in electrochemical experiments, coinciding with apparent excess heat production. In the Italian experiments, neutrons came in bursts.

Another group, working at NTT Basic Research Laboratories, Tokyo, claims to have identified He-4 coming from deuterated palladium. Early-on in the investigation, large neutron bursts were seen simultaneously with explosive release of gas, with plastic bending of the sample and production of excess heat. Later, ultra-high resolution mass spectrometry indicated the presence of He-4 during evolution of excess heat.

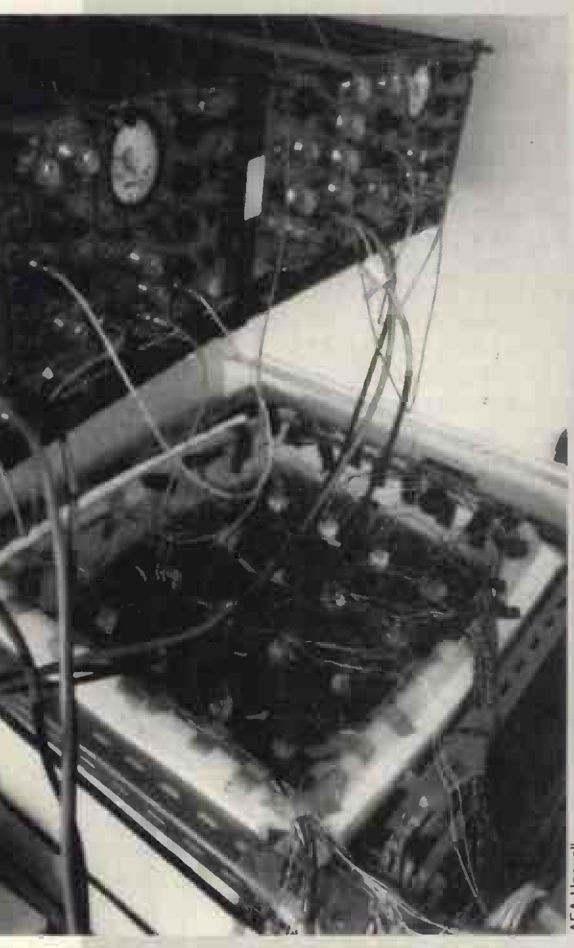
Similarly M F Miles and B F Bush of China Lake, California, have repeatedly correlated He-4 signals with excess heat in electrochemical reactions, though they have not so far been able to make the excess heat production repeatable.

J O'M Bockris of Texas A & M University has reported production of "massive quantities" of tritium at a palladium electrode, accompanied by He-4 production. Thermal expulsion and mass spectrometry by N Hoffmann of Rockwell International Corporation confirmed that He-4 was present in nine out of ten specimens – in quantities corresponding to two or three hundred times the background. No helium was present prior to tritium production, implying that the element was synthesised by fusion

Pro-cold fusion data suppressed

Several explanations have been put forward to explain cold fusion. For example, substituting a charged radioactive particle called a muon,

Cold fusion explosion



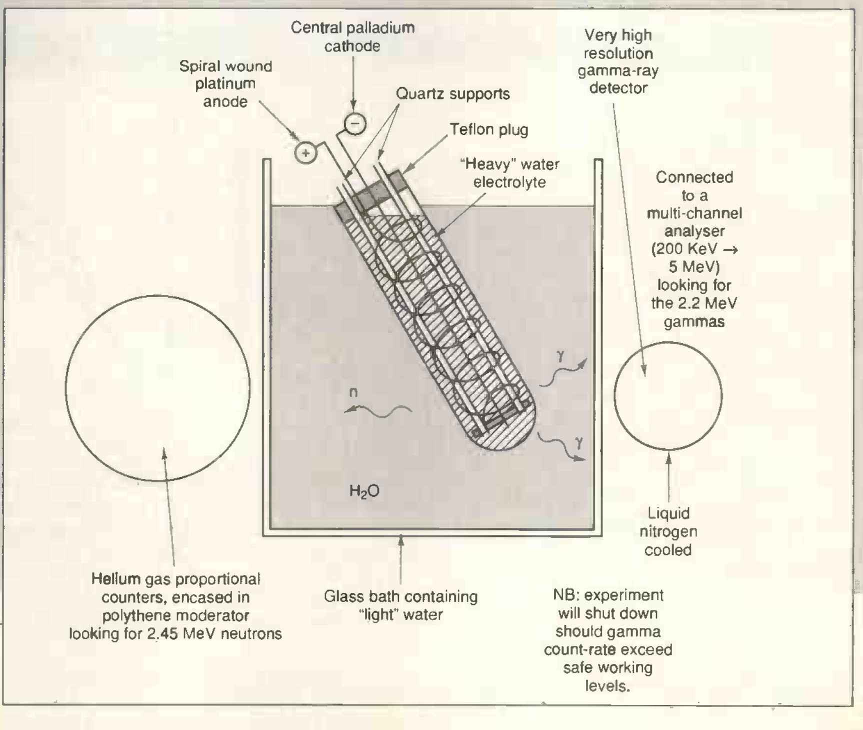
The scientific community was shocked in March 1989, when Professors Pons and Fleischmann announced[2] they had discovered nuclear fusion at ordinary temperatures. Their evidence was production of excess heat during heavy water electrolysis experiments, using a palladium cathode and platinum anode in a simple calorimeter. The only way excess heat could have been produced, it was claimed, was via nuclear reactions between deuterons (heavy hydrogen nuclei).

The reactions just shouldn't happen. After all conventional wisdom is that fusion can only happen at very high temperatures – millions of K – such as the nuclear reactions that fuel the stars. Such energies are needed to bring the nuclei close enough together for fusion. For it to occur in condensed matter, at room temperature, some novel mechanisms must be at work, perhaps as a result of the non-equilibrium conditions.

Spurred on by the possibility of low-cost inexhaustible energy, scientists tried to repeat the experiments, concentrating on verifying the energy-balance calculations and attempting to find nuclear isotopes or particles that could only come from fusion reactions. Most failed, or at least produced cryptic effects.

Some of the most painstaking research was carried out at AEA Harwell under Professor Williams[3]. In three months, of seven-days-a-week investigation at a cost some £320,000, the Harwell team found that observations attributed to cold fusion could be artefacts arising from several possible errors. Among defects identified were inadequate controls; imprecise material characterisation; insufficiently thorough calorimeter calibration; problems in distinguishing neutron counts from the background; and well-known spurious effects with $^{10}BF_3$ and ^3He proportional counters.

The AEA work was perhaps the turning-point for cold fusion research. While producing an essentially negative result, it also set a tough standard for researchers to follow, concluding: "Claims of observations of cold fusion ought now to meet similar standards of data analysis and materials characterisation, so that a proper assessment can be made."



Top: AEA Harwell's sixteen 50cc 'Fleischman' cells undergoing excess heat studies. Some of these are control cells. The experiment would normally be covered with an insulated lid.

Right: Experminetal cold fusion cell structure co-located with neutron and gamma detectors.

200 times heavier than the electron in a hydrogen molecular ion, increases the probability of fusion by around eighty orders of magnitude.

A similar distortion of the wave function between nuclei could arise when deuterons are loaded into a metal lattice, particularly under the special conditions that might occur during electrolysis.

But this amounts to little more than speculation until it is backed by real scientific results.

The lack of reproducibility is not in itself surprising as electrochemical reactions are by their very nature highly sensitive to the state of the surfaces involved. In addition, cold fusion is a notoriously difficult field involving several highly specialised scientific disciplines spanning nuclear physics, materials science, electro-chemistry and instrumentation technology. Nevertheless, it is surprising that none of the pro-cold-fusion data gathered has earned publication in a respectable, refereed scientific journal – arguably publishing data at conferences of fellow enthusiasts does not add to the body of scientific evidence.

So can the latest research pass the scrutiny of the orthodox scientific community? Increased funding, especially from Japan, may produce results. But the fact that so much effort has yielded so few conclusions, gives little room for optimism. The case for cold nuclear fusion remains unproved through lack of reliable witnesses. Unfortunately, on the current evidence it seems unlikely to even reach court. ■

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Inconclusive search: Dr Derek Craston examines data logging equipment used in the calorimetric measurements on 24 cold fusion cells. Despite closely matching the original conditions which were said to have produced excess heat, AEA Harwell's researchers were unable to duplicate the results.

DESIGN BRIEF

RF REFLECTIONS

Ian Hickman delves into the true effects of matched and unmatched loads, to map the movement of energy in a transmission line.

The maximum power theorem says that a practical power source will deliver the maximum power of which it is capable to a load, when that load is matched – that is, has the same value of resistance as the source's internal resistance. Its corollary is that the internal dissipation in the source is then equal to that in the load R_L, giving an overall system efficiency of 50%. This applies at dc, or at ac if reactive components are absent – and even if they are present given certain conditions.

To keep the sums simple, a 1Ω source delivering 1W to a matched load is assumed, except where otherwise stated. But for other value loads, the internal dissipation in the source can be anything from zero to four times that delivered to a matched load.

In practical situations where high system efficiency is desired, eg a flash-lamp or a 500MW turbo-alternator, the load resistance will be much higher than the internal resistance of the source, so that most of the energy finishes up where it is wanted, in the load.

In **Fig. 1b** the 1Ω source is connected to a transmission line with a 1Ω characteristic impedance, terminated with either a short- or an open-circuit. Either way, 100% of the incident energy is reflected. But the internal dissipation in the source is anything but 2W (1W as with a matched load plus 1W supplied to the line but entirely reflected back by the load).

I had come across this conundrum years before and concluded that what was reflected was basically voltage (and current), and that the question of power had to be deferred until we saw how the reflected voltage combined with the incident voltage at the output terminals of the source.

But this is a wrong conclusion. Looking back into the output terminals of the source you see a matched source, yes: but because of the presence of the ideal voltage generator, a matched load, no.

This difference becomes clearer if we consider what happens in more detail **(Fig. 2a)**, when the source is connected to a loss-free coaxial transmission line 200,000km long.

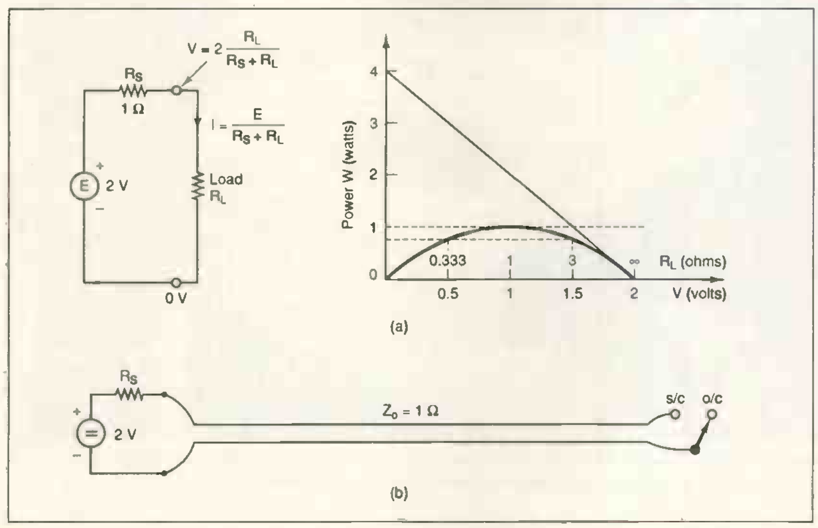
Velocity of propagation in the line is two thirds that of the speed of light – commonly the case with coax – and on closing switch S_1 at time $t = 0$ seconds, the source sees a matched load and delivers 1W to it. After 1s the source has supplied 1J (1watt second) to the line, and its internal dissipation is 1W. At t = 1s, **(Fig. 2b)**, the 1V wavefront reaches the load, which we will assume to be open circuit, but cannot continue any further. Therefore the 1A current in the line must somehow fall to zero at this point, the current being reflected back towards the source.

To cause the current to flow back again against the incident wavefront, the voltage at the open circuit end of the line must rise to 2V – the current is reflected in antiphase and the voltage in phase. The 2V step propagates back along the line, arriving at the source at t = 2s. Line voltage now equals the source emf so there is zero voltage across the internal source resistance R_S and no current flows in it. The source has effectively disconnected itself at t = 2s exactly, so we can open S_1 at t = 2s or any subsequent time without making the slightest difference to the 2J of energy stored in the line.

But if instead of opening S_1 at t = 2s, we had instead replaced the 2V generator by a short circuit, the source would now look like a matched load and would dissipate 1W during the next 2s, until at t = 4s, all the energy reflected from the load would be dissipated in the "source".

Radar transmitter designers uses a line charged up to kilovolts in just this way to to feed a short square high power dc pulse to a magnetron.

Fig. 1a. 1Ω source which delivers 1W to a 1Ω load dissipates 1W internally, but for other loads the source's internal dissipation may be anything between 0W and 4W. Power in load (curved line) and source's internal dissipation (sloping line) plotted against load resistance and terminal voltage. Power in load exceeds 75% of maximum for 0.33 < RL < 3Ω[1].
1b. 1Ω source tries to deliver 1W to a matched line terminated with a 100% reflection. Why isn't the internal dissipation in the source 2W?



Avoiding contradictions

It is tempting to think that at t = 2s, and thereafter, there is no current in the line and all the energy is stored in the line's capacitance. But this leads to a contradiction, only avoided by an apparently slightly absurd hypothesis. Namely that there really is a current of 1A flowing left to right and another flowing, on the same line at the same time, from right to left, like a train going round the long thin loop of London's Circle Line - a train so long that its head is hitched up to its tail. This is just a case of a pulse of 1V amplitude and 2s duration circulating in the line and there is no reason why it should not. Consider the case of a 1V pulse of 1s duration, **Fig. 3a**. Here, the switch connects the line to source for just one second from $t = 0$ to $t = 1$. **Figure 3b** shows the state of affairs at various subsequent times.

After the source is disconnected, the pulse bounces back and forth along the loss-free line for ever, the voltage being 1V at $t = 1, 2, 3...$ seconds, but 2V on half of the line and zero on the other half at 1.5, 2.5, 3.5s.

If total capacitance of the line is C Farads, then at 1, 2 seconds etc, if the energy were all stored in the line's capacitance it would total $1/2\ C\ 1^2$ or (numerically) $C/2$ joules: at 1.5, 2.5 seconds when the energy is all stored in one half of the line, it would total $1/2\ (C/2)\ 2^2$ or C joules – twice as much! The conclusion can only be that the energy is stored jointly in the line's capacitance and *inductance*, which presupposes that current is flowing in the line.

Sinewave signal

If the signal supplied to the 1Ω open circuit line were not a dc pulse of whatever duration, but a sinewave, the situation at the open circuit is just the same. Voltage is reflected in-phase and the current in antiphase (**Fig. 5**). Vectors further down the line (ie to the right in the case of the incident wave and to the left in the case of the reflected) are shown lagging – displaced clockwise – since they will not catch up until a little later. The incident v and i vectors are unit vectors, 1V and 1A. It can be seen that at $\lambda/8$ from the open circuit, the resultant voltage and current is 1V and 1A in quadrature, with the current leading. So at a distance of $\lambda/8$, the impedance seen looking towards the open circuit is a capacitive reactance equal to Z_0, or $-j1\Omega$ in this case. Drawing the diagram for the short circuit case gives $+j1\Omega$, illustrating the well known fact that

$$Z_0 = \sqrt{(Z_{oc} \cdot Z_{sc})}$$

The voltage and current both vary from zero to twice the incident value, at different points along the line, giving a standing wave ratio of infinity to 1. This would apply only on the right half of the 200,000km line at $t = 1.5$s, if an rf source were connected to the line for just 1s, the rest of the line being devoid of energy travelling in either direction.

As in the dc case, a 2s pulse would fill the line twice over, the standing waves demonstrating that energy was indeed travelling both ways all the time. Hanging a very high input impedance buffer amplifier on one end of the line, would make available – almost indefinitely – a sample of the rf , giving a signal which was greatly "stretched" in time.

Receiving a radar pulse and retransmitting it at low level in such a stretched form is the basis of one type of counter-measure used in electronic warfare.

The fact that a line of (about) 200,000km provides a delay of 1s (ie the signal travels at about two thirds the speed of light) comes about as follows:

$$Z_0 = \sqrt{(L/C)}$$

(see box) where L and C are the inductance and capacitance of any length of line.

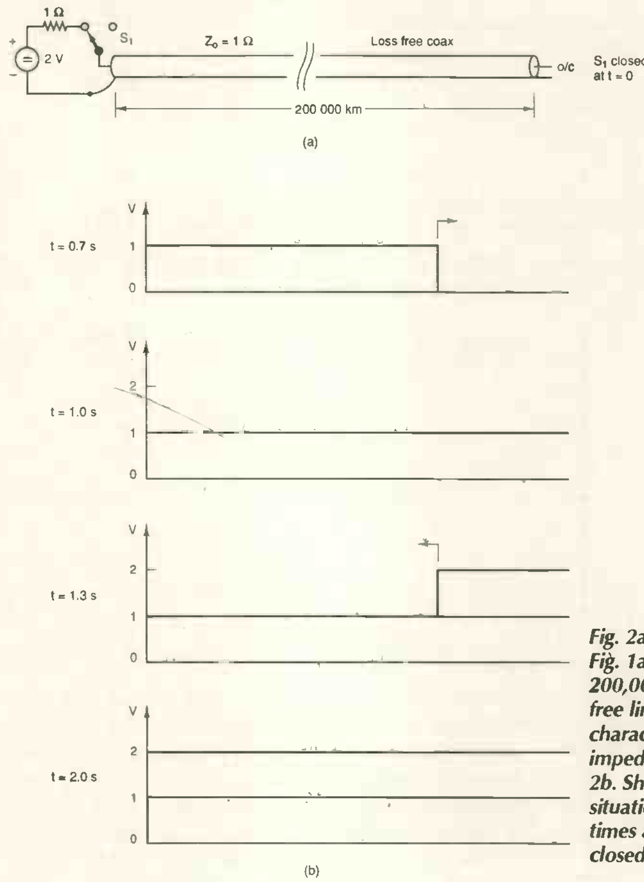


Fig. 2a. The source of Fig. 1a driving a 200,000km long loss-free line with a characteristic impedance Z0 of 1Ω. 2b. Showing the situation at various times after switch S1 is closed.

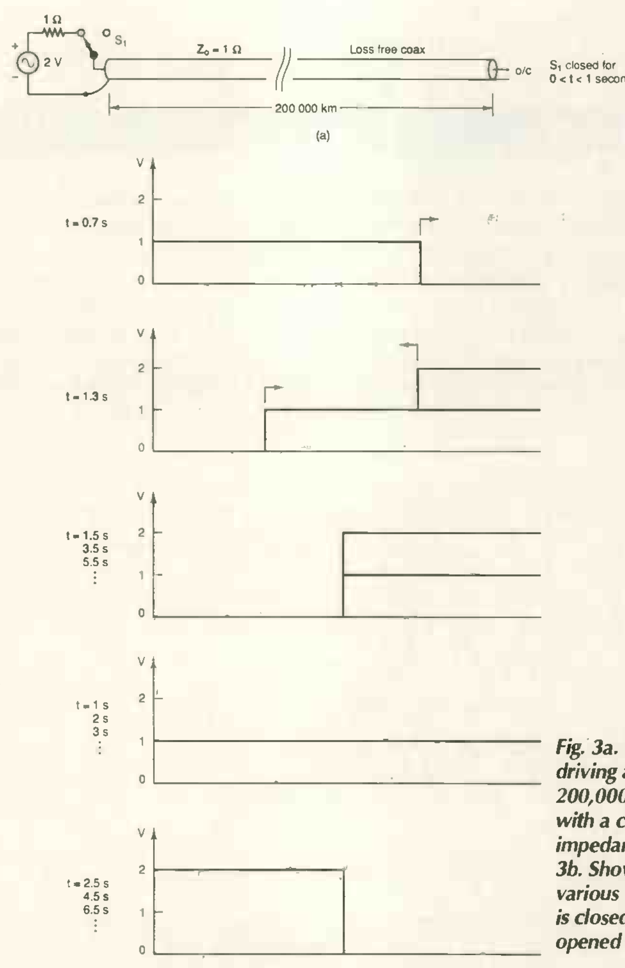


Fig. 3a. The source of Fig. 1a driving a 1s long pulse into 200,000km long loss-free line with a characteristic impedance Z0 of 1Ω. 3b. Showing the situation at various times after switch S1 is closed at t = 0 (S1 is opened again at t = 1s).

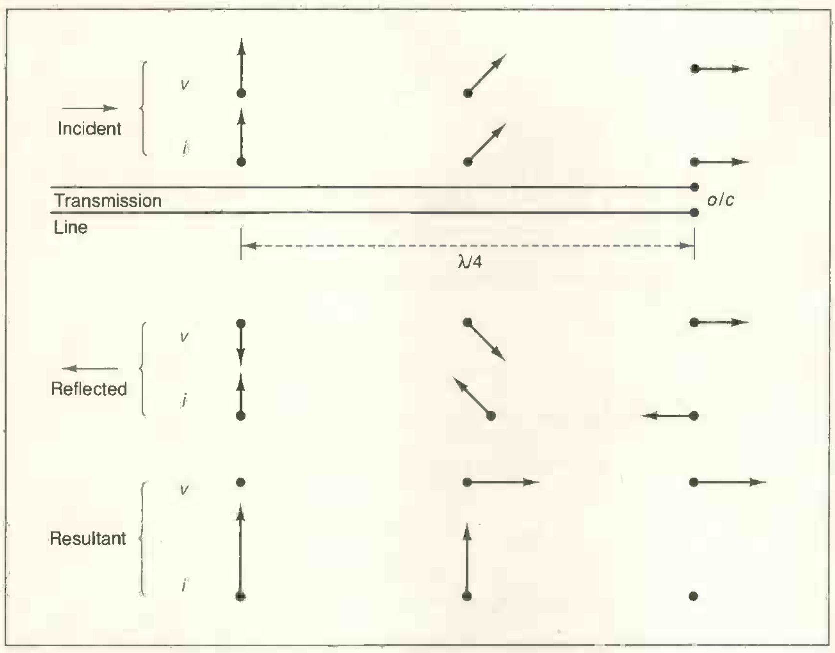


Fig. 5. Radio frequency signal applied to an open circuit transmission line[2].

Actual length is immaterial as it cancels out. For a coaxial cable, C is typically about 100pF/m, so L works out at 250nH/m. The delay experienced by a signal in travelling along the line is $\sqrt{(LC)}$ns/m (see box). Substituting the given values of L anc C gives the delay as 5ns/m where the velocity v is 200,000km/s. Reducing L without increasing C, or vice versa, would raise the velocity. But in practice any substantial increase proves to be impossible.

Increasing the diameter of the inner until it nearly fills the outer would reduce the inductance: the length of the magnetic flux paths would be increased and the cross sectional area available to carry the flux reduced – both effects increasing the reluctance of the magnetic circuit, reducing the total flux caused by the current. But capacitance would be increased, offsetting any gain.

Conversely, moves such as increasing the diameter of the outer could only reduce the capacitance at the expense of increasing the inductance. A fortune awaits the inventor who can half both L and C simultaneously, as the velocity in such a cable would exceed that in free space. Values of L and C per metre in free space are the lowest that can ever be achieved, giving $v = c$ (the velocity of light) and $Z_0 = 377\Omega$.

Given that $Z_0 = 377\Omega$ and $\beta = 1/c = 3.33$ns/m, we can find out just what they are, since:

$Z_0 \times \beta = \sqrt{(L/C)} \times \sqrt{(LC)} = L = 377 \times 3.33\ 10^{-9} =$ 1256nH/m

and

$\beta\ /\ Z_0 = \sqrt{(LC)}\ /\ \sqrt{(L/C)} = C = 3.33\ 10^{-9}\ /\ 377 =$ 8.85pF/m.

The value of 8.85pF/m for C is easy to visualise: it is the capacitance of two metal plates each of area $A = 1\text{m}^2$ separated by distance $d = 1$m (ignoring fringing). Since $C = (\eta_0\ \eta_r A)/d$ – and the relative permittivity η_r for air is unity – then 8.85pF/m is numerically equal to η_0. 1256nH/m is more difficult to visualise, but is in fact equal to $4\pi 10^{-7}$, the value of the permeability of free space μ_0.

So $\beta = \sqrt{(\eta_0\mu_0)}$ and

$c = 1/[\sqrt{(\eta_0\mu_0)}]$.

That the internal dissipation in the source is apparently

Unit discussion

The characteristic impedance Z_0 and phase constant β of a line are given by approximately $\sqrt{(L/C)}$ and $\sqrt{(LC)}$ respectively where L and C are the inductance and capacitance per unit line length (or per section in the case of a delay line composed of LC sections) – assuming the losses are low. At least, these are the commonly quoted expressions. A low loss line is one such as the balanced line in Fig. 4 - which shows a very short section and its equivalent circuit. G and R are both near zero. Actually it is not the values of inductance and capacitance which determine the line parameters, but their reactances; the product in the case of Z_0 and the quotient in the case of β . But as $j\omega L$ and $1/(j\omega C)$ both have units of ohms, their product has units of ohms squared. Z_0 has units of ohms independent of frequency, as the $j\omega$ terms cancel out.

With β, it is the ohms that cancel out, leaving units of... what? To find out, it is best to go to the full expressions for Z_0 and the line propagation constant.

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}}$$

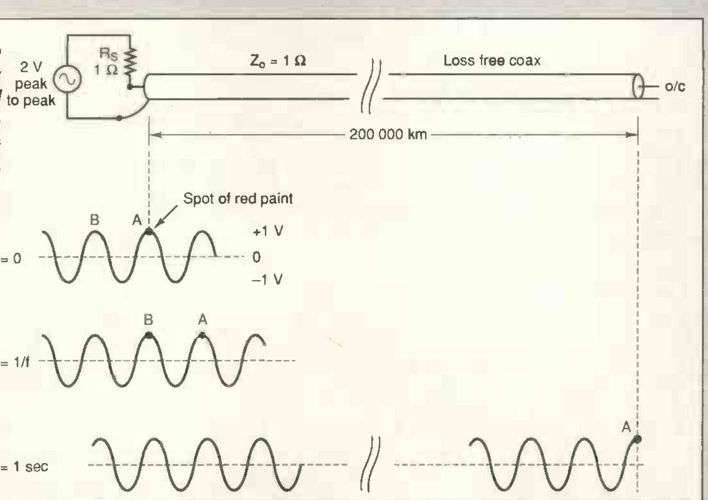
$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)}$$

$$= \sqrt{(R + jX)(G + jB)}$$

In the case of Z_0, if $R = G = 0$ then the expression simplifies, as described, above to $\sqrt{(L/C)}$. The propagation constant is in general a complex quantity γ which is made up of $\alpha + j\beta$, where α and β are the attenuation and phase constants respectively, per unit line length (or per section in the case of an LC delay line). As the wave propagates along the line, its amplitude and phase relative to the input can be expressed succinctly as:

$e^{-\gamma} = e^{-(\alpha + j\beta)}$.

Radio frequency signal applied to an open circuit transmission line.



not 2W, when the load reflects 100% of the incident energy, is true of both a Thevenin generator (Fig. 1a) and a Norton generator (**Fig. 6a**) – at dc. But it is not necessarily so at ac.

When a 1Ω Thevenin generator tries to deliver 1W to a purely reactive load which, at the frequency in question, has a reactance of 1Ω, the internal dissipation is indeed 2W, **Fig. 6b**. The same is true for the Norton generator. But it would be elegant to find a generator which always dissipated 2W internally. The clue to such an arrangement is the fact that on short-circuit a Thevenin generator dissipates 4W internally while a Norton dissipates zero, and vice versa on open-circuit. Take two 2Ω loads, one receiving 0.5W from a matched Thevenin generator and the other receiving 0.5W from a matched Norton generator, connect them in parallel and the result will be 1W delivered to a 1Ω load from a source, dissipating 2W internally on either short- or open-circuit (**Fig. 7a**).

2W is also dissipated internally when trying to deliver 1W into any purely reactive load (**Fig. 7b**). The situation can be a little confusing to envisage without resorting to formal mesh analysis. But the vector diagram is easily derived if the Thevenin and Norton sections are provided with separate j2Ω loads, which are subsequently paralleled. Furthermore, for any impedance load with any phase angle, the sources in the two generators always supply a total of 2W between them – any power not finishing up in the load being dissipated internally. I leave algebra to prove this to you.

Compared to a 1Ω 1W Norton generator, the Norton section of the "Hickman type I" generator (Fig. 7a) has a 1A constant current generator instead of a 2A one, and a 2Ω shunt resistor in place of a 1Ω one. By contrast, in

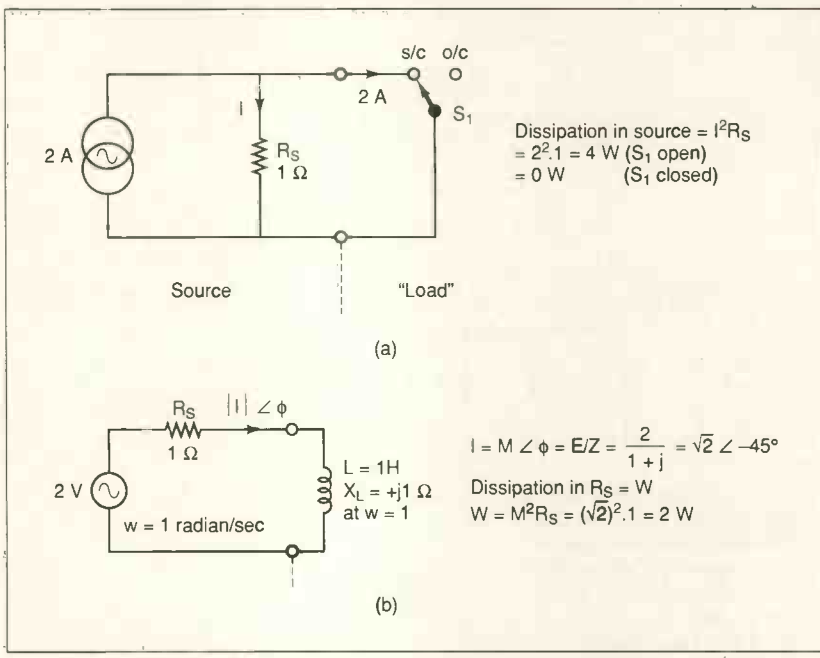


Fig. 6a. When a Norton generator representation of the source is used, its internal dissipation is zero with a short circuit load and 4W on open circuit, just the opposite from a Thevenin generator.
6b. A 1Ω Thevenin generator supplies a purely reactive load of 1Ω. Its internal dissipation is exactly 2W. (M is the modulus or magnitude of the current; I = M/ f where M = | I |)

For a loss free line where $R = G = 0$, γ simplifies to β, the j in the exponential indicating that it refers to the sine wave's phase, not its exponentially decreasing amplitude. So, if $R = G = 0$, then:

$$\beta = \sqrt{j\omega L.j\omega}$$

$$= \sqrt{-\omega^2 LC}$$

and $|\beta| =$

$$= j\omega\sqrt{LC}$$

at 1 radian/s.

The phase constant is proportional to frequency and has units of radians per unit distance along the line – radians per metre if L and C are the inductance and capacitance per metre of the line.

The difference beween θ and β must be borne in mind. θ refers to the continually changing phase of the signal at a fixed point on the line - $\theta = \omega t$ where the instantaneous voltage $V = V_{max}\sin(\omega t)$ – the latter to the phase at one point on the line relative to another at the same instant. The Figure shows a sinewave of frequency fHz (f cycles per second) entering a transmission line, the voltage at the input being $+V_{max}$ at time $t = 0$, which makes it in fact a cosine wave. To see

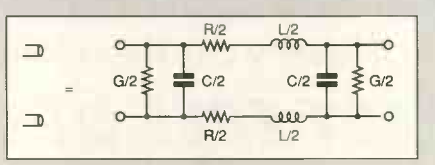


Fig. 4. A short section of a balanced line and its equivalent circuit. For a coaxial line the circuit is similar except that the upper leg consists of R and L in series while the lower leg is a straight through connection2.

what happens, put a spot of red paint on the waveform at the input at $t = 0$, point A in the figure. Then at $t = 1/f$s later, the next peak B is just entering the line while the spot of red paint is distance d down the line. Distance d between successive peaks on the line is the wavelength λ on the line, in the case of coax typically two thirds of the free space wavelength. The phase of the signal at the point on the line where the red spot is now (at $t = 1/f$) is lagging that at the input by 2π radians - it won't have caught up until $t = 2/f$, by which time the red spot will be a further distance $d = \lambda$ down the line. After 1s, the red spot will have travelled a distance $f\lambda$ down the line, so the velocity v of the wave is $f\lambda$m/s.

β is the phase lag per metre. The lag at the point distance d down the line at $t = 1/f$ is one cycle or 2π radians. So $d = 2\pi/\beta = \beta$ and $v = f\beta = (\omega/2\pi)(2\pi/\beta) = \omega/\beta$.

The result can be arrived at more directly by noting that:

$$\beta = J\omega\sqrt{LC}$$

(L and C are values for 1 metre of line) has units of radians/metre. Dividing by ω gives the result independent of frequency and has units of (rad/m)/(rad/s) = s/m.

$\beta/\omega = \sqrt{(LC)}$ has units of seconds delay per metre for the particular cable in question, and the reciprocal of this has units of metres per second, giving $v = \omega\beta$ directly.

That excellent if ancient textbook, *The Handbook of Line Communications*, calls v the phase or wave velocity, adding darkly, without further explanation, that: "the group velocity (ie, the velocity at which energy is transferred along the line) is $d\omega/d\beta$".

In the present case of a transmission line, where β is directly proportional to ω their ratio is constant so the wave and group velocities are the same and are independent of frequency.

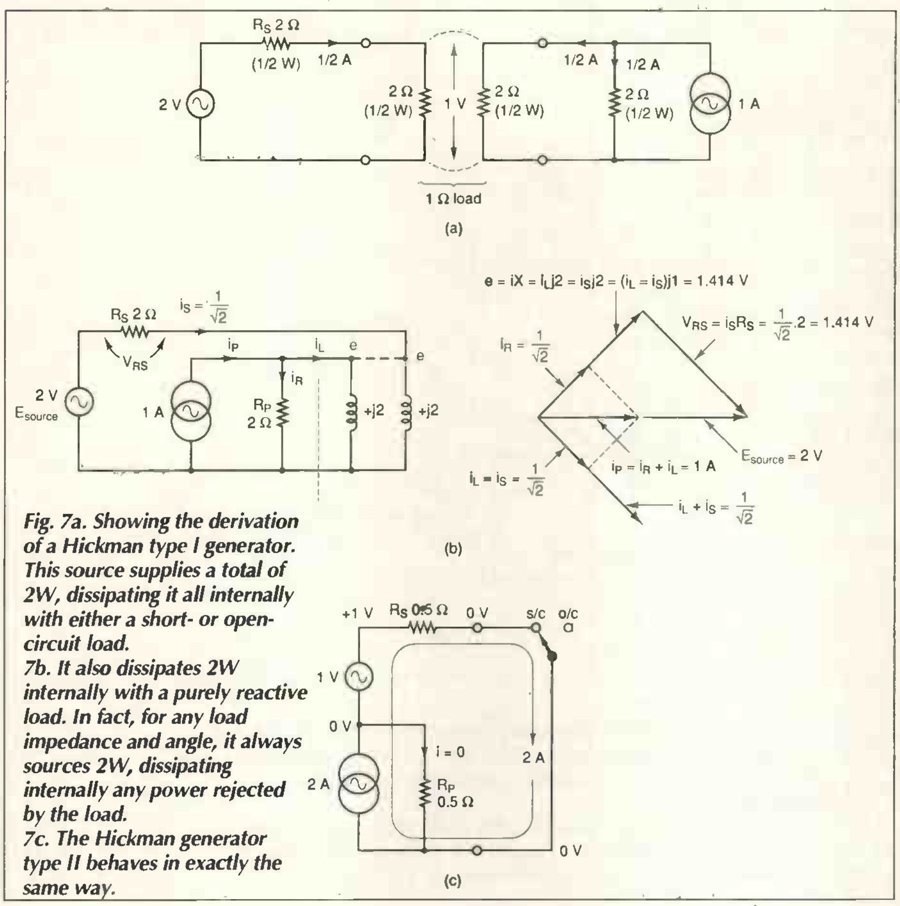


Fig. 7a. Showing the derivation of a Hickman type I generator. This source supplies a total of 2W, dissipating it all internally with either a short- or open-circuit load.
7b. It also dissipates 2W internally with a purely reactive load. In fact, for any load impedance and angle, it always sources 2W, dissipating internally any power rejected by the load.
7c. The Hickman generator type II behaves in exactly the same way.

the Thevenin section, only the series source resistor changes, from 1Ω to 2Ω. From a philosophical point of view there seems an odd lack of symmetry. But this is due to the fact that – in the short circuit case – currents in parallel add, whereas – in the open circuit case – voltages in parallel do not.

Symmetry is restored by taking into account the dual, the Hickman type II generator (**Fig. 7c**). Here it is a case of voltages in series add, but currents in series do not.

I have not seen this type of generator mentioned in any text book, but almost certainly some-one else has invented it already, in which case doubtless a well-informed reader will kindly write in to the Editor, quoting chapter and verse, and the name by which it should rightly be known. ■

References

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2. The Newnes Practical RF Handbook Ian Hickman Butterworth-Heinemann 1993 ISBN 0 7506 0871 4

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