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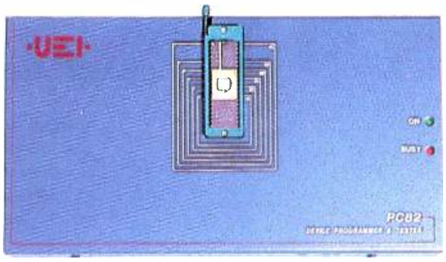

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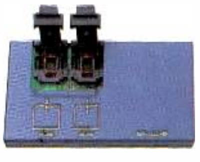

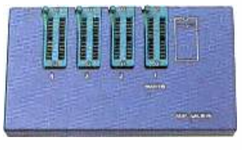

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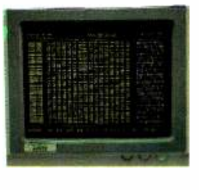

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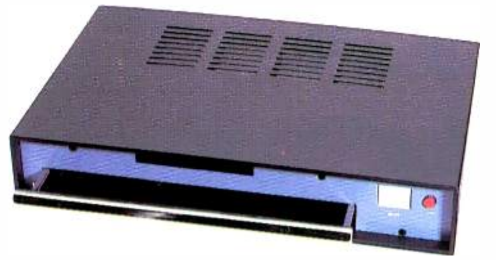

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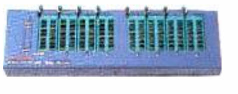

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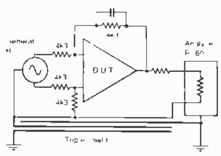

REGULARS

In next month's issue: Low power radio links
Since the DTI removed the licencing requirement for low power radio equipment working on certain spot frequencies, some interesting type-approved modules have emerged. Ian Hickman provides a number of practical application examples using an RX/TX pair operating at 418MHz. You also have the opportunity to purchase modules for development purposes at half price. See next month's issue for full details.

FEBRUARY ISSUE IS ON SALE JANUARY 28

Unacceptable standards

Why should the UK have just passed a law which states that, from January 1, 1994, the selling and renting of a widescreen television without a D2MAC decoder is an offence?

And why should anyone want to insist on the fitment of decoders for which there are presently no programmes and virtually no prospect of there ever being any?

The answer to the first question is that the European Commission told the UK Government to pass the legislation in accordance with a Brussels directive. The answer to the second is simply that euro-electronics giants Philips and Thomson think it would be good for business if EC consumers stumped up for their microchips regardless of the fact that there is presently no call for them. Nor is there likely to be.

This odd and costly state of affairs has its origins in the early Eighties when broadcasters researched replacements for the ageing PAL and SECAM transmission standards. The result was MAC, an ingenious mix of analogue and digital electronics which matched available technology. Rightly anxious to maintain a European initiative in the development of a new television standard, the European Commission launched the Eureka 95 HDTV development programme with millions of ECUs steered towards European broadcasters and setmakers.

There was of course no public demand for a new TV standard so Brussels, bureaucratic as ever, issued one of its Directives to create one. The resulting edict, now UK law, will force setmakers to include a MAC decoder in all widescreen sets, and new satellite operators to broadcast future services using the MAC standard. Fine, except that broadcasters think further and faster than Brussels: they realise that MAC, nearly ten years in evolving, is already past its sell-by date, eclipsed by digital HDTV.

Almost as one, existing satellite service broadcasters applied for – and gained – exemptions from the MAC broadcast rule. They can continue to transmit PAL/SECAM indefinitely to European audiences not conspicuously unhappy with current picture quality. And should there be a few viewers who would like a better picture, existing broadcasters will be able to transmit in backwards compatible PAL Plus.

To be fair to all parties involved in HDTV decisions, development of a viable digital system occurred quickly: it went from scepticism to reality in just two years. It has already reached the point where HDTV can be broadcast in the same sized frequency slot as standard 625 line PAL. Even more remarkable, it appears immune to the ghosting and co-channel interference which bedevil existing terrestrial TV services. A report about the latest BBC trial appears on page six of this issue.

Meanwhile, the European Commission is insisting that we use and pay for a system which is obsolescent before it even starts. It intends to hand out £750 million from the European budget despite objections from the UK Government about this expenditure.

It goes without saying that the main beneficiaries will be the MAC component suppliers, Philips and Thomson. These companies are spending millions of guilders and francs pushing the Commission for a place at the ECU money trough.

It will demonstrate a corrupt system if the MAC lobby is allowed to succeed.

Frank Ogden

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REED BUSINESS PUBLISHING

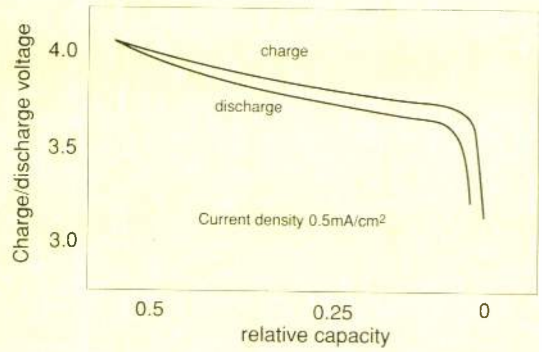
Lithium rechargeables to replace NiCds?

A new rechargeable battery system developed by the Sony subsidiary company Energytec outperforms conventional NiCd battery systems in virtually all aspects.

The battery, known as *Lithium Ion*, delivers 3.6 volts per cell compared to 1.2V of a NiCd while being able to support the heavy peak discharge currents of motorised applications such as camcorders and portable phones. Energy storage density, the stored electrical power compared to weight and volume of the battery, is said to be at least three times that of a NiCd.

Sony claims that its new batteries can be charged and discharged 1200 times, with only 12% self-discharge after one month of storage compared to 25% for NiCds. After six months NiCds have lost 60% of their charge but the lithium battery has lost only 30%. It may be used with similar charging rates. Unlike NiCd systems, Sony's lithium rechargeables do not exhibit a memory effect, a factor of great significance to the electronics industry.

Sony's lithium rechargeable battery system behaves more like lead acid than NiCd in its characteristics. In particular, it doesn't exhibit the memory effect which reduces effective cell capacity.

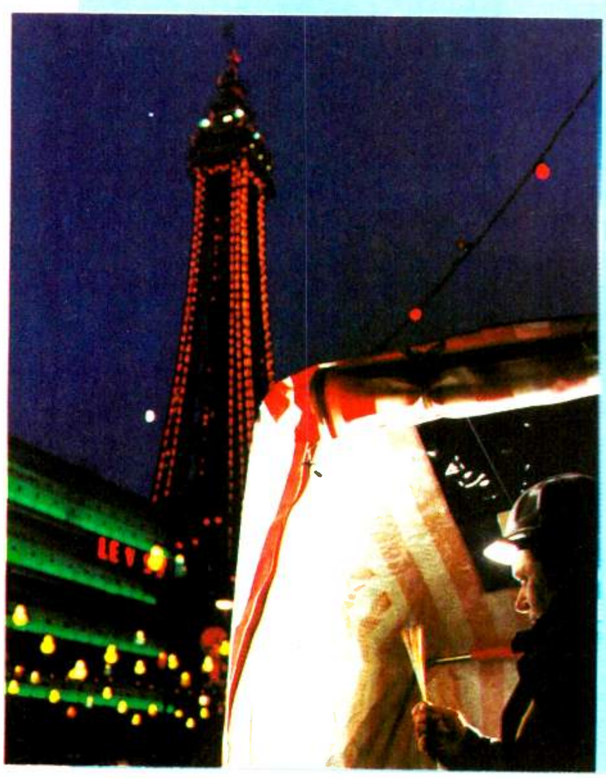


The news about the lithium rechargeable system was buried in the text of a Sony camcorder announcement. The *CCD-TR8* machine uses a rechargeable lithium battery, the first to be offered with a consumer product. The company says that it is cheap enough to incorporate in consumer equipment, and safe enough for consumers to use, abuse and mishandle.

Until now portable electronics has relied on nickel-cadmium batteries. These work well when new, able to deliver bursts of heavy current while maintaining a relatively flat discharge curve. Although NiCds are sold with the promise that they may be recharged up to a thousand times, this is only true if used in a completely unnatural way. A NiCd must be drained of all charge before recharging. If it is topped up with charge, it develops a "memory", delivering back little more than the amount of charge used to top it up. Sony's lithium rechargeable system exhibits no such drawbacks.

The memory effect leaves users with a flat battery when they most need one. Even if warned of the danger, few can resist the temptation to top up their battery before leaving home to make video movies or work on a journey with a portable computer. Owners need to do this because a NiCd loses half its charge if left unused for a few months. A battery that should hold an hour's charge may only drive a camcorder for ten minutes. Lead-acid batteries can be topped up but are too heavy and only survive a few hundred charge cycles.

Sony's rechargeable lithium cell uses a



***Blackpool optical illuminations:* communication systems for the Radio Wave radio station operating from the Blackpool Tower use optical fibre cable. The installation by British Telecom involved running a 2000ft continuous length of cable under the street, down into the basement, and up through a 90 degree angle to the top of the Tower, 500 feet up.**

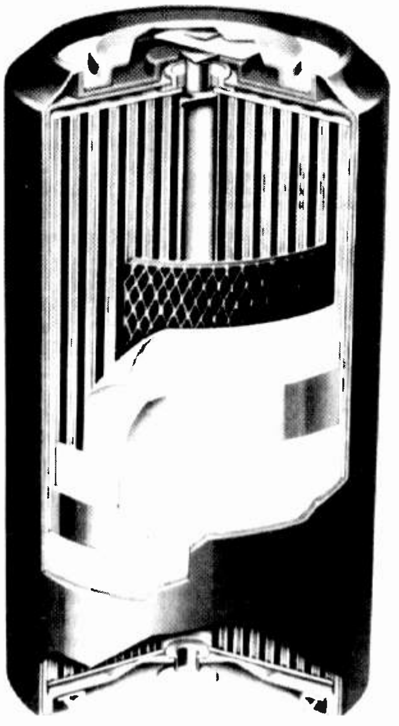
Because Blackpool Tower is a listed building with restrictions on drilling and structural interference, the cable was installed using a technique which "blows" it through a tube on a cushion of compressed air. The outer tube, carrying the cable, was lashed to the structure.

BT has now installed more than two million kilometres of optical fibre in Britain.

carbon anode electrochemically doped with lithium. The cathode is made from a compound of lithium, cobalt, nickel and manganese. The anode carbon is doped with lithium during a first charging cycle. Some of the lithium moves from the cathode to the anode through non-aqueous electrolyte, a mix of propylene carbonate and diethyl carbonate. The anode carbon is derived from polyfurfuryl alcohol, with heat treatment at 1100°C after adding phosphorus-bearing compounds, such as phosphoric acid and phosphorus pentoxide.

Sony is committing heavily to the technology. Energytec's factory in Japan has installed a production line which can make 100,000 lithium ion batteries a month. The initial price will be 30% higher than a NiCd, but is expected to fall as production expands.

Sony believes that Lithium Ion batteries may finally make electric vehicles viable. **Barry Fox.**



Development work carried out by the University of St. Andrews and Dowty Batteries has also resulted in a rechargeable lithium manganese dioxide battery system which is claimed to have around 50 per cent more storage capacity of equivalent nickel cadmium cells. The St Andrews cell is said to have achieved almost 50W-hours/kg although commercial cells are still more than a year away.

Britain vetoes MAC subsidy

Europe's policy on high definition television (HDTV) has fallen apart after the UK blocked attempts to give a subsidy of up to 850m ECUs (£650m) to build a broadcasting infrastructure around MAC technology. At an EC telecomms ministers' meeting, both Britain and Denmark vetoed a French compromise proposal to agree on the principle of a MAC subsidy, leaving the amount to be decided later.

Britain stood alone in rejecting a further meeting to discuss HDTV funding. A DTI spokesman said the government believed that if public money is to be spent on HDTV development it should go on technology with a long-term future, such as digital HDTV, rather than on a stop-gap analogue system like MAC.

LCD with integrated drivers

Researchers at GEC's Hirst Research Centre have developed an active matrix liquid crystal display (LCD) with driver circuits integrated on the glass substrate. The display is built using a low cost, low-temperature glass-compatible polysilicon process. Similar technologies use quartz substrates for the integrated driver circuitry. In a paper to be delivered at *Japan Display 1992*, researchers describe a 2.8in. diagonal display comprising 200 x 200 pixels, each driven by a dual gate nmos transistor. In the architecture used, the row and column drivers are on opposite plates of the display. This maximises the yield of the matrix because there are no row/column crossovers.

The transistors' polysilicon active layer was built using a Rytrak low-power chemical vapour deposition machine to deposit silicon on to a glass substrate. GEC has already built 6.2in. 240 x 960 pixel LCDs using external drivers and claims that displays of this size are possible using the integrated driver technology.

Student aid for industry

USITT (the University of Southampton Institute of Transducer Technology) is co-ordinating a Student Project Scheme offering member companies the chance to submit science and engineering based projects for investigation by University undergraduates.

As part of their degree course, all science and engineering students are currently required to devote one day per week (for two terms) to a project. The Institute, having recognised that the students' time, labour and expertise could be successfully used to address latent ideas and problems faced by industry, recently invited member companies to submit project ideas.

A number of assignments of interest have already been received. Companies who have not already suggested projects are welcome to do so. The projects may cover a variety of topics, ranging from the completely theoretical to the largely practical.

Although projects may be submitted at any time of the year, they will only be distributed to the relevant University Departments during the Autumn or Spring terms.

Britain loses to EC lobbyists

French electronics companies are increasing their influence with the European Commission, causing their British counterparts to lose out. Companies such as Bull, Alcatel, Thomson and France Telecom now have permanent lobbyists in Brussels.

For instance, French promoters of MAC-format TV for European HDTV were instrumental in persuading the Commission to spend 850m ECUs of European tax payers' money on subsidies to promote D2MAC (see story "Britain vetoes MAC subsidy).

The number of offices that represent French interests has risen from eight in 1988 to 72. Lobbying is left to trade associations by British companies. French companies, whose day to day operations involve support for European standards, are also quick to inform the Commission of possible transgression by Far Eastern companies, whether this involves local content or tariff infringements.

GPS breakthrough

The cost of GPS (global positioning by satellite) RF circuitry is likely to fall as a result of GEC Plessey Semiconductors' launch of a highly integrated RF circuit.

Current GPS front-ends use a mixture of discrete components and GaAs ICs. The *GP1010* is fabricated on a bipolar process with an f_T of15GHz, the result of co-operation with the Canadian Marconi Company for application in their GPS avionics. It requires few external passive components to convert GPS Ll-band spread-spectrum signal at 1575.42MHz to digital data.

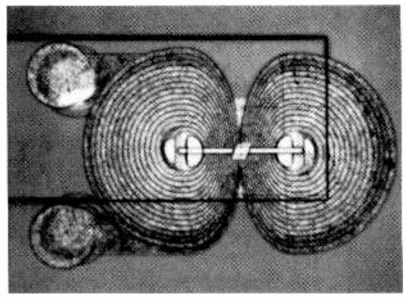
The availability of low cost circuitry will help drive Global Positioning products into the commercial domain, so GPS receiver prices are certain to fall. This will encourage new markets to develop, such as vehicle Location, and handheld personal navigators for hikers and bikers.

EMC test threat

Thousands of small electronics design houses could be driven out of business by the cost of getting prototype equipment and one-off units tested to conform with the EC's EMC Directive when it becomes effective after 1st January, 1996.

Small companies will be expected to conform to the Directive, even for single pieces of equipment. Testing costs could be as high as £10,000 for a troublesome piece of equipment.

Disk drive head integrated directly onto silicon. This revolutionary design bythe French company Silmag also puts the head slider mechanics onto the same piece of silicon rather than a conventional thick film hybrid ceramic substrate. Using silicon as the substrate material also allows the use of photolithography rather than micromachining in manufacture. The new heads will be made at Lincoln, England by GEC-Plessey Semiconductors.



Bulletin on electronics

Up-to-date information on microcontrollers and development tools can be had through a bulletin board set up by in-circuit emulation equipment manufacturer, Hitex.

The board is host to a wide range of information on popular microcontrollers and programming languages. It also houses the *C51* and *C166* user groups for the *8051* and Siemens *SAB80C166* devices.

It includes real applications and demonstration software can be downloaded free-of-charge. The board is open from 8.00am to 9.30pm Monday to Saturday. Membership is free and there is no charge for time spent on the system. Call 0203-690026

Skills initiative welcomed

The Engineering Council has praised the Government's intention to produce greater numbers of young people with technical skills. Denis Filer, Director General of The Engineering Council, said that the Council welcomes the announcement by Education Secretary John Patten that he wishes "to see young people emerging from education with skills that are relevant to the needs of industry, skills which will make them the Brunels and the Stephensons of the future".

Mr Filer said: "Mr Patten's pledge to provide funds to improve the quality and quantity of engineers and technicians is good news for the engineering profession, for industry and for the nation. It will not only help to rebuild our manufacturing sector but it will also help the United Kingdom to be more competitive in the international field".

Digital HDTV success spells end for MAC

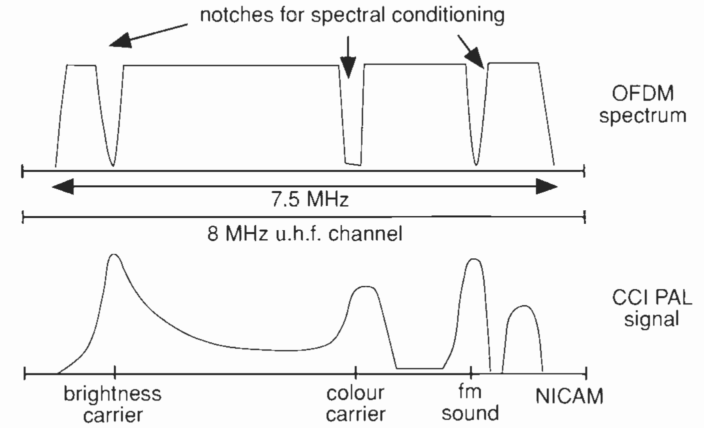
Just a year or so ago experts in digital signal processing, image compression and broadcasting were still arguing about the possibility of squeezing a digital high-definition television signal into a standard European terrestrial TV channel, just 8MHz wide.

Last week the BBC and French electronics giant Thomson CSF announced successful tests of a system which can squash not one but two full HDTV programmes into such a channel. Asked to fit a quart into a pint pot, Thomson and the BBC have managed to pour in a full gallon without spilling a drop.

The secret is a modulation technique that can achieve a data rate of 7.5bit/s for each Hertz of bandwidth. This enables the system to deliver a total data rate of 60Mbit/s in an 8MHz channel, more than enough to carry two HDTV signals after they have been squashed using the latest digital image compression techniques.

The system, developed by Thomson CSF and Laboratories Electroniques de Rennes (LER), is similar to the digital HDTV setup produced by the Swedish HD-Divine consortium which stole the show at the International Broadcasting Convention in Amsterdam in July. But by using a different modulation technique Thomson and LER have increased the channel capacity.

How OFDM reduces cross-channel interference



To send a raw digital HDTV signal (1250 lines, 25 frames per second), with the brightness information sampled at 72MHz and the colour at 36MHz, requires a data rate of around 1Gbit/s. But by using a combination of compression techniques this can be cut by a factor of 40 or so to 25Mbit/s.

First, a discrete cosine transform (DCT) is applied to blocks of 8x8 pixels in each frame. This converts the blocks into two-dimensional spatial frequency arrays, where each value in the array represents the magnitude of a particular spatial frequency component in the original block. The fact that the human eye finds it harder to detect

Continued over page

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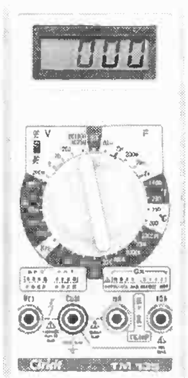

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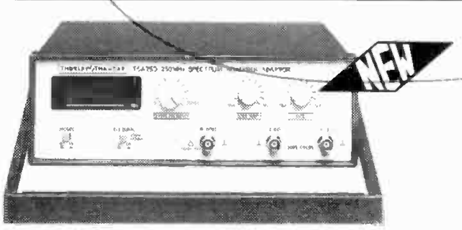

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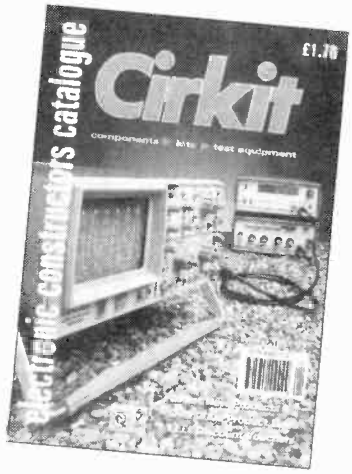

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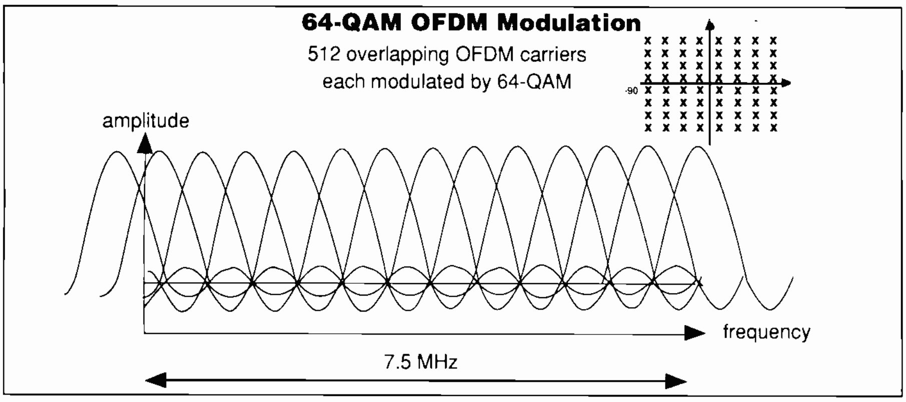
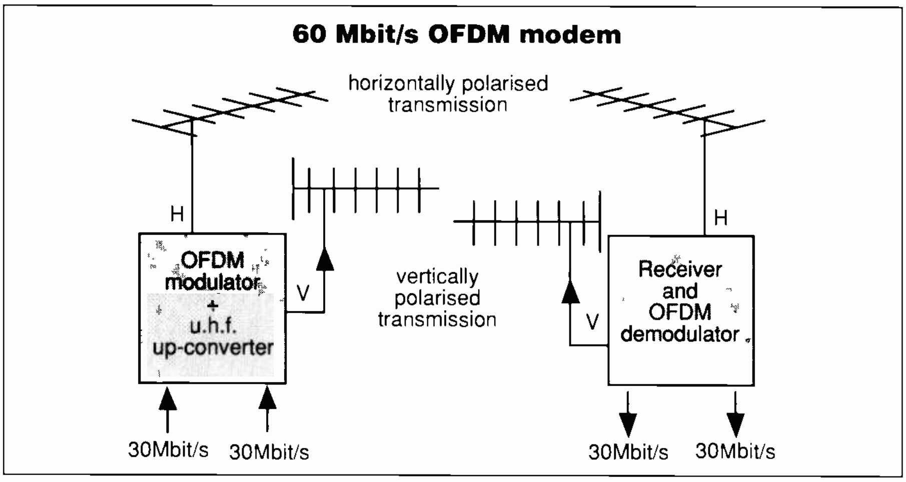




errors in images with higher spatial frequencies means that it is possible to get away with using fewer bits to code the high frequency components.

Other techniques take advantage of the relationships between adjoining pixels, either next to each other in the same frame or in the same place in adjacent frames. Coding the differences between these pixels rather than the raw colour and brightness information produces a big data rate saving.

Finally, a variable length coding system is used to minimise the number of bits required to code the compressed difference signal. The codings are chosen so that the most common values (ie the values of difference information most likely to occur) are given the shortest code numbers. The same technique is used in Morse Code, where the most common letters are given codes with just one or two pulses.

Impressive though these compression techniques are, they are now well established. The really clever part is the way Thomson and LER have managed to achieve a spectral data density of 7.5 bits per second per Hertz.

At the heart of the modulation system is a technique known as orthogonal frequency division multiplexing (OFDM). This spreads the signal across the entire 8MHz bandwidth in a very efficient way, at the same time allowing some freedom to avoid parts of the channel where interference is most likely to occur.

Instead of a single carrier frequency, the picture information is divided between a large number (in this case 512) of closely-spaced but independent carriers (see diagram). Although the frequency bands associated with these carriers overlap the signal do not interfere because the carriers are orthogonal to each other.

The technique is analogous to replacing a single signal cable carrying data at a high rate with lots of smaller wires in parallel, each carrying a much lower data rate.

Because the data rate on each carrier is relatively low, around 60kbit/s, problems such as ghosting caused by reflections from buildings can be virtually eliminated. The delay between the original and reflected signal is generally much shorter than the gap between pulses in the original signal, so the two cannot be confused.

Another advantage of OFDM is that the orthogonal carriers can be spaced across the frequency band in a way that minimises interference with PAL TV signals from nearby transmitters. Carriers close to the peaks of the PAL luminance, colour and sound carrier frequencies are not used, leaving notches in the OFDM spectrum. In the BBC tests the system was able to tolerate a PAL signal in the same channel with a power 12dB lower than the digital signal, ie about a sixteenth as strong, without any affect on the picture quality. A PAL signal would show considerable picture interference with a second signal this strong. The compressed TV signal is coded using 64-state quadrature amplitude modulation (64-QAM) on each of the 512 orthogonal carriers . Each cycle is coded into one of 64 states, created by combining eight possible values of amplitude and eight values of phase angle. This means each cycle on each carrier can send a 6-bit number (since $2^6 = 64$).

Using these modulation methods an 8MHz channel can carry up to 30Mbit/s. The final trick is to double this by sending two separate signals in the same channel, one polarised vertically and the other horizontally.

In successful tests last month the BBC showed that the modulation technique can work at the data rates proposed, using standard definition TV signals. The next step is to repeat the tests for full HDTV pictures, using prototype compression hardware.

Karl Schneider, *Electronics Weekly*

Superconductors wired for action

US government researchers profess to be close to developing a way of forming high temperature superconductor materials into usable wires. This has previously posed a problem: the materials are very fragile; wires formed from them are brittle and cannot handle large current loads.

Ames researchers have teamed up with researchers from Babcock and Wilcox, a firm that specialises in manufacturing furnace insulation fibres. They have adapted a B&W technique that produces superconductor wires five times more flexible than types previously manufactured.

The wires are formed by forcing molten superconductor material through a supersonic nozzle. The material cools quickly when it emerges and solidifies into long fibres.

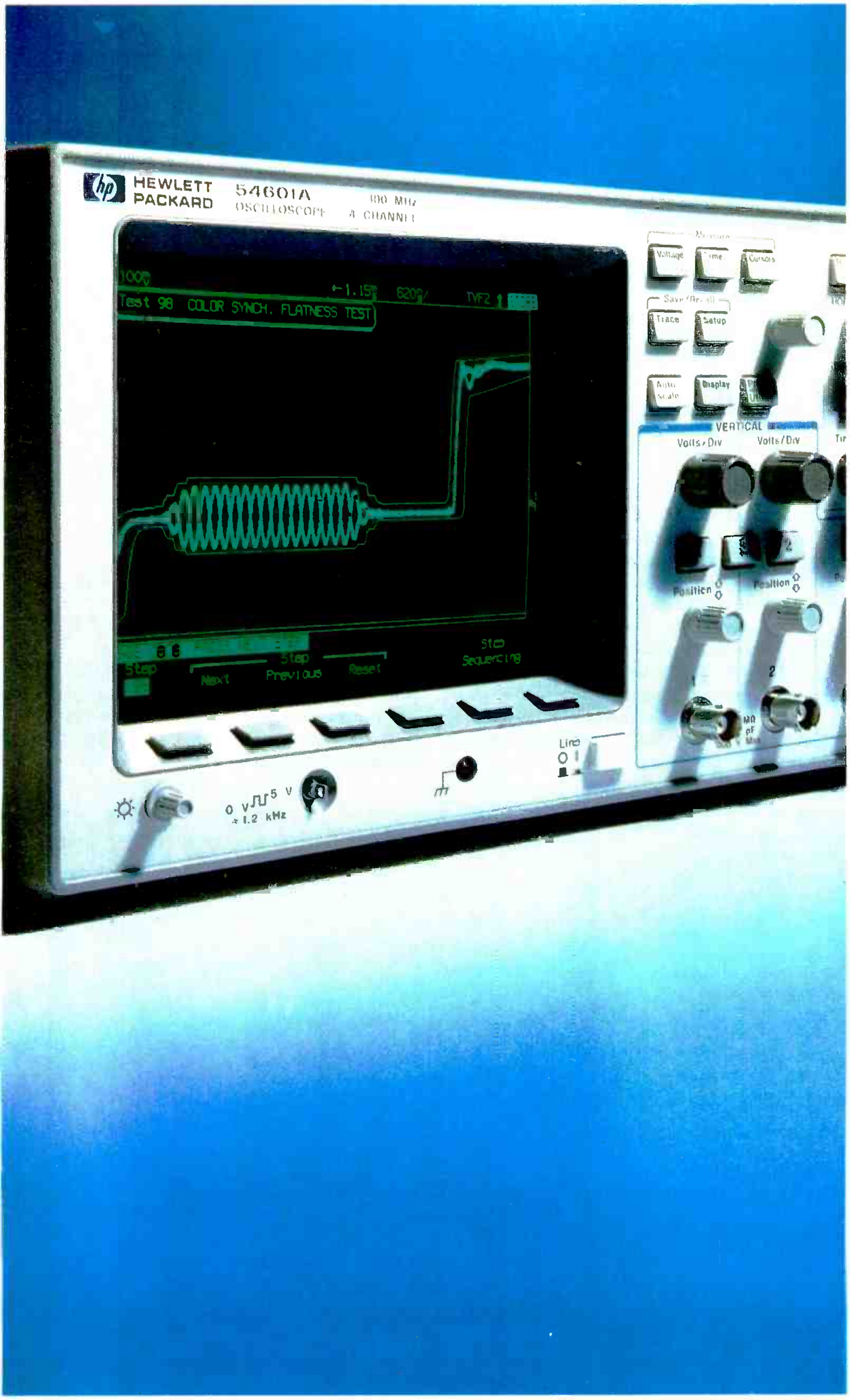
Physics degree "best in Europe"

The Institute of Physics has singled out Sheffield Hallam University's Engineering Physics degree course as "inspirational and unique in Europe".

The lead editorial in the institute's journal *Physics World* claimed that the erstwhile polytechnic had put together a course which could deliver skills not only in physics, but in applied engineering. Graduates will be able to take a direct route to either chartered engineering or physics.

Course content includes instrumentation and measurement, modern materials and maths/computer modelling of physical systems. The course includes a year long industrial placement.

Within budget. Without compromise.



The value of this 100 MHz digital scope is easy to see.

Take a close look at the HP 54600 oscilloscope, and you can't help but notice certain things.

It looks and feels like an analogue scope, with dedicated knobs and a display that responds instantly to your control changes. Yet it has all the digital power that analogue can't give you – high accuracy, automatic measurements, hard copy output and programmability. And superior viewing of virtually any waveform, even at low rep rates and slow sweep speeds.

But what really stands out is that the HP 54600 gives you this performance for just £1860 (4-channel) and £2160 (2-channel version).* And that's a value worth looking into.

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HP 54600 Digital Oscilloscope	
No. of channels	2 or 4
Bandwidth	100 MHz
Timebase accuracy	±0.01%
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There is a better way.

HEWLETT PACKARD

*As at 1 September 1992

CIRCLE NO. 107 ON REPLY CARD

RESEARCH NOTES

Diamonds make chips sparkle

A breakthrough in the manufacture of synthetic diamonds is set to open up the possibility of faster, denser and more powerful chips. Researchers at AT &T Bell Labs in Murray Hill, NJ, and Crystallume Inc of California, have come up with what looks like a practical answer to the seemingly insurmountable problem of heat dissipation.

Removing heat from a semiconductor junction depends on creating good thermal contact between the junction and the outside world. But good conductors of heat are also usually good conductors of electricity, so they cannot be used as the first link in the "heat chain". Instead, insulators – often silica or ceramics – are used to transport heat for those critical first few microns of its journey to a metal heat-sink. In general electrical insulators are many times poorer than metals at conducting heat. Unless they happen to be one of the world's most remarkable... diamond.

John Graebner who heads the team from Bell Labs says that diamond is about five times better as a heat conductor than copper, largely because of the very stiff carbon-carbon bonds in diamond and also because of the highly rigid nature of the crystal lattice.

Unfortunately, slicing up the *Koh-i-Noor* to make substrates for chips is not altogether practical, which is why Graebner and others have been eagerly pursuing the idea of making synthetic diamonds. For nearly 15 years, a whole variety of approaches has been adopted in the search for a practical method of making useful synthetic diamond. Chemical vapour deposition (CVD) is now one of the most successful methods, producing diamond deposits of up to a millimetre in thickness and covering a substrate 100mm in diameter. These are films of tiny interlocked crystals rather than single crystal diamonds. But Graebner says that they have improved so much in recent years that their thermal properties are becoming as good as single crystal gemstones. "As little as a year ago the films were generally black because of impurities and defects, but now you can buy films that are optically transparent."

Latest research (*Nature*, Vol 359 No 6394), has led to diamond films that conduct heat as well as copper on their bottom surface, but five to six times *better* near the top.

This remarkable result shows that some synthetic diamond is as good as the best natural gemstones. Graebner says that the achievement will open the way to the widespread use of diamond films as heat spreaders under many high-powered electronic devices.

Glimpsing the limits (and potential) of magnetic storage

For the first time the quantum effect that may limit the amount of information storable on magnetic media such as tapes and disks has been observed. The effect – macroscopic quantum tunnelling – becomes significant at sizes about an order of magnitude smaller than those used by present magnetic recording technology and represents an ultimate barrier to how much data can be stored. Its occurrence in magnetic materials has been predicted for about 20 years, though until now it has not been seen.

But macroscopic quantum tunnelling, causing groups of magnetic atoms to act as single magnetic units, has now been reported by scientists from the University of California Santa Barbara and the IBM Thomas J Watson Research Center, Yorktown Heights, in a naturally-occurring iron-containing protein called ferritin (*Science*, Vol. 258 p.414).

Ultra-sensitive magnetic measurement techniques allowed researchers to record the behaviour of magnets one hundredth the size of those previously studied. They found that when the 4500 iron atoms making up a ferritin core were cooled to liquid helium tem-

Chaotic communications: Research into the behaviour of laser systems at the US's Georgia Tech has revealed a new aspect of chaos and highlighted possible problems for fibre optics communications.

Work at the institute, sponsored by the National Science Foundation and Department of Energy's Chemical Science's Division – has shown that at certain placements, light travelling through the lasing crystal moves at slightly different velocities depending on its polarisation. When these velocities interact with similar variations in the laser's frequency doubling crystal, chaotic fluctuations result in the amplitude of the light beam.

Georgia scientists are using the effect to investigate chaos, applying their results to such areas as data encryption and even the mathematics of Josephson junctions which behave as a similar type of oscillator to the laser system.

But the research also reveals a potential problem for fibre-optic telecomms. In the drive to carry more information, fibre-optic engineers are adding more signals on ever closer frequencies. At the same time, signals are being increased in intensity to reduce the number of repeater stations needed to convey telecomms over long distances.

Georgia's programme is showing that if the frequency separation between pulses is small and the intensity of the pulses is high then they start interacting with each other through the non-linear susceptibility of the fibre optic media. Under certain conditions the effect could be to make cable systems unstable because signals will become chaotic, leading to scrambling of communications.

Picture: Joe Schwartz

CIRCLE NO. 108 ON REPLY CARD

peratures, their magnetic fields began to fluctuate in unison rather than freezing in place. Such behaviour is not consistent with classical theories of magnetic behaviour, but fits that predicted by quantum mechanics.

This form of quantum tunnelling appears at a scale of about 7nm, a size which therefore defines the upper limit for the amount of information that can be packed in a given space on a magnetic surface.

Commenting on the discovery, Dr Philip Stamp of the University of British Columbia says (*Nature*, Vol 359 No 6394) that the advance is important for a number of reasons, not least because it offers the opportunity to test the predictions of quantum mechanics. He goes further and suggests that the use of biological molecules (like ferritin) to do fundamental physics may lead to a new kind of applied physics using biologically-manufactured components. Stamp recollects the late Richard Feynman's advocacy of "quantum computers" – devices that could store 10^{20}bit/s in a volume of $1cm^3$. He adds that such a computer would dissipate less than 1mW at 0.25K and would easily exceed the total of all existing world computing power.

X-rays to throw light on black holes

Scientists at Columbia University, along with those at the Lawrence Livermore National Laboratory, the Marshall Spaceflight Center and two Italian research institutes are preparing an X-ray astrophysical experiment to fly on a Russian satellite which will explore some of the most exotic objects in the Universe.

The experiment, made possible under an agreement announced by Nasa and the Russian Institute for Space Research, is based on an instrument designed to measure the polarisation of X-rays from space. The X-ray polarimeter, it is claimed, will be the most sensitive instrument of its type ever flown in space.

Polarisation measurements will help astrophysicists understand whether an X-ray emitting region of space contains a black hole – X-ray emission is thought to be one of the signs of the highly energetic processes going on in the vicinity of black holes. The polarimeter is also expected to reveal how rotating neutron stars – pulsars – emit X-rays.

The polarimeter package is designed to be placed on the Spectrum-Roentgen-Gamma satellite, scheduled to be launched by the Russians in mid-1995 with a 2t scientific payload. Scientists will participate in experiments to be placed on the satellite, which will carry the most powerful X-ray and gamma ray detectors ever orbited.

The satellite's orbit will be highly elliptical, with an apogee of about 200,000km, ie about half the distance to the moon. A single orbit will last four days, creating the opportunity for unusually long uninterrupted observation of specific X-ray sources.

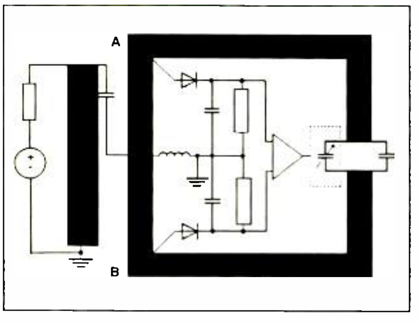
The polarimeters will be placed at the focus of one of two 8m focal length grazing incidence X-ray telescopes built for the project by Russia and Denmark. Unlike light or radio waves, X-rays cannot easily be focused to a point by means of lenses or conventional mirrors. Reflection is only possible when the beam hits a plate at a very small angle – like a stone skimming a pond.

The X-ray polarisers will be developed and calibrated at the Lawrence Livermore Laboratory, and the detectors will be supplied and tested by Columbia University and by research institutes in Frascati and Palermo, Italy. Marshall Spaceflight Center will be responsible for developing software to simulate the experiment and verify the overall design.

Antennas that tune themselves

A novel self-tuning loop antenna has been developed by two researchers from the University of Stellenbosch in South Africa (*Electronics Letters*, Vol 28, No. 22). It is a miniature etched loop, designed for portable VHF and UHF equipment and is capable of remaining resonant at the transmit frequency, even when the latter is altered, for example by changing channels.

Loop antennas have extremely desirable directional properties and are also compact. Their major disadvantage when used with hand-held equipment is that their bandwidth is extremely narrow and the resonant frequency is affected by changes of temperature and by the proximity of conductive objects. So even if the tuning could be switched along with the transmitter's channel selector, it would not necessarily hold for long.



Self-tuning loop antenna, developed in South Africa, designed for portable VHF and UHF equipment.

Step forward the self-adjusting antenna.

The South African loop is designed to monitor the transmit frequency, compare it with the actual resonant frequency of the antenna itself and then use a feedback loop to bring the antenna into tune. The loop is fed from the signal source in two distinct ways: magnetically using a short line and electrically through a capacitor.

The electrically coupled signal, injected at the centre of the loop, excites voltages that are symmetrical with respect to earth, so magnitude and phase of the voltages are therefore identical at points A and B, regardless of whether the antenna is resonant. The magnetically coupled signal, which forms the radiating component, induces voltages at A and B which are 180° out of phase with respect to each other and 90° out of phase with the electrically coupled signal. But only when the antenna is resonant. If the antenna drifts off resonance or the transmitter frequency is changed, this 90° relationship no longer holds. Phase difference between the electric and magnetic components of the signal at points A and B changes in an opposite and equal way; one goes greater than 90°, the other less. The circuit shown in Fig.1 is designed to rectify and subtract the signals, producing an error voltage proportional to the difference between the signal frequency and the antenna resonant frequency. Bringing the antenna into tune automatically is then a simple matter of using this error voltage to drive a voltage-controlled tuning capacitor.

The authors say that for a small milliwatt-sized transmitter a variable capacitance diode will do the job admirably. Their experiments with a system working at 402-406 MHz shows that it is possible in practice to increase the effective bandwidth of a small loop antenna by an order of magnitude compared with that of an uncompensated loop.

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

PUTTING PUNCH INTO VOICE COMMUNICATIONS

ILLUSTRATION JAMEL AKIB

Conventional speech processing enhances talk power at the expense of intelligibility.* Louis Thomas *presents an alternative system which operates without waveform clipping.

Simitar is a mean-loudness controller and waveshape compressor that in essence controls loudness and raises the level of the lower amplitude elements of a complex waveform. But unlike traditional compressors and clippers, Simitar preserves audio fidelity and controls loudness without degrading the signal-to-background noise ratio.

In communications and broadcast systems from PA to FM radio, Simitar improves speech intelligibility – whether conditions are ideal or poor. Equally importantly, by enhancing voice and audio-frequency signals in general, the system offers the potential to make much better use of RF spectrum. Preliminary tests with speech enhancement for the hard of hearing are encouraging too.

Loudness is controlled over a 30dB range while obtaining, typically, an increase of 6dB on mean power level and 15dB on consonant speech levels. Frequency modulated radio systems benefit from an overall increase of 6dB in audio power while in AM systems, an increase of 6dB in mean carrier power leads to lower background noise. On AM, a similar improvement in intelligibility without Simitar would mean an increase of 10 to 15dB in mean power.

Whether at the sending or receiving end of a communications system, Simitar can restore signal levels and improve noise performance. Modulation levels are improved since maximum output can be precisely controlled under all conditions.

Simultaneous processing

Simitar (SIMultaneous near Instantaneous and Time Averaged Response) represents the first major step forward in crest factor reduction since Licklider's work in 1948. Two elements make up the system – a waveshape compressor and a loudness controller – with the compressor operating almost instantaneously on the waveform and the loudness controller affecting the mean level. Either mode can be used independently but the greatest benefits are obtained with the two combined.

Being implemented digitally, Simitar allows programmable levels of waveshape compression to be added to any mean loudness control characteristic. Distortion caused by gain switching is minimal since gain only changes at the waveform's zero-crossings.

In effect, the system reduces the ratio of peak-to-mean power through sampling the peak of the input's half cycles, detecting zero-crossing points and adjusting gain accordingly. Gain applied by the mean loudness controller and/or waveshape compressor is determined by the amplitude transfer characteristic needed. Numerous functions can be carried out including, limiting with background noise reduction, compression, expansion and noise gating.

In the mean loudness controller, the rate of gain change is determined by a digital filter whose response is synchronised to the waveform's frequency content. Gain is allowed to decrease by the maximum amount over a single half cycle, or increase over 256 half cycles, thereby retaining envelope characteristics. The waveshape compressor has an instantaneous response and may increase or decrease gain every half cycle.

While all work on the process has been aimed at improving the performance of professional communications systems, there is no reason why the principles cannot be applied to studio and broadcast applications.

Waveshape compression

Three parameters define the operation of any loudness controller:

- the quantity in the waveform that represents its loudness;
- level of gain or loss needed to bring the measured level to that required, and
- way in which gain is applied.

In this method, the quantity representing loudness is the peak amplitude of individual waveform half cycles, **Fig. 1**, between consecutive zero crossings. Only magnitude information is needed, the sign being ignored.

The peak determines the amount of gain applied to the waveform to adjust its level according to the amplitude transfer characteristic.

Gain changes are applied as a step function only at the zero-crossing instants, **Fig. 2**, and so the input and output half cycles have the same shape, differing only in their peak amplitudes.

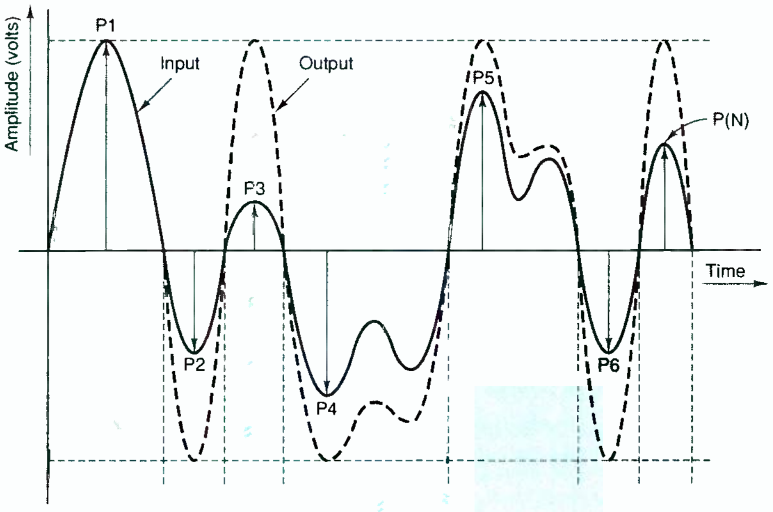
Several reasons justify choosing the waveform's peak value, which is relatively easy to measure. Controlling peak amplitude avoids peak limitation distortion, and the peaks are least affected by noise. Measurement and control of peak amplitude provides better speech quality than options based on the mean or RMS area under the half cycle. Finally, since peaks occur twice per cycle, amplitude information is available at a higher rate than with methods based on averaging.

To allow gain to be applied at the zero crossing point of the start of the waveform just measured, delay is needed. This delay must be at least as long as the longest zero-crossing period expected. For a lower limit of 300Hz, a delay of 8ms is suitable – assuming that a sixth-order Butterworth high-pass filter is applied to the input. Higher-order filters reduce the delay to 4ms or less. The process is shown in **Fig. 3**.

The amplitude transfer characteristic defines the relationship between input peak amplitude and the output peak. The gain needed to implement the characteristic is defined by the gain function, **Fig. 4**. Any function that can be drawn within the axes of the amplitude transfer diagram can be implemented, but in practice, incremental slope must be limited to avoid subjective degradations. Peak voltages referred to on the diagram are of individual waveform half cycles.

For a given mean input level, an increase in the limiting range causes an increase in the compression level as more and more of the smaller half cycles are raised. As a result, the

Fig. 1. Input and output waveforms show how the compressor lifts lower amplitude half cycles while maintaining waveform characteristics.



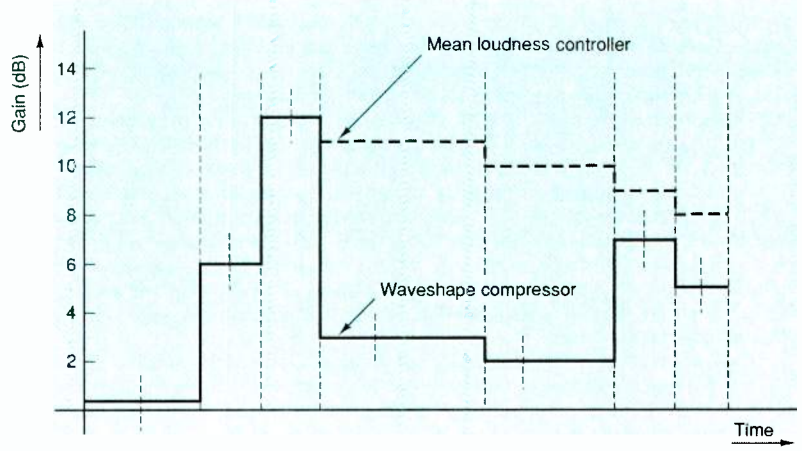


Fig. 2. Gain variation for the waveshape compressor is near instantaneous while that of the loudness controller varies over a longer period.

Fig. 3. The author's system combines a mean loudness controller and waveshape compressor have been available in a single digital process.

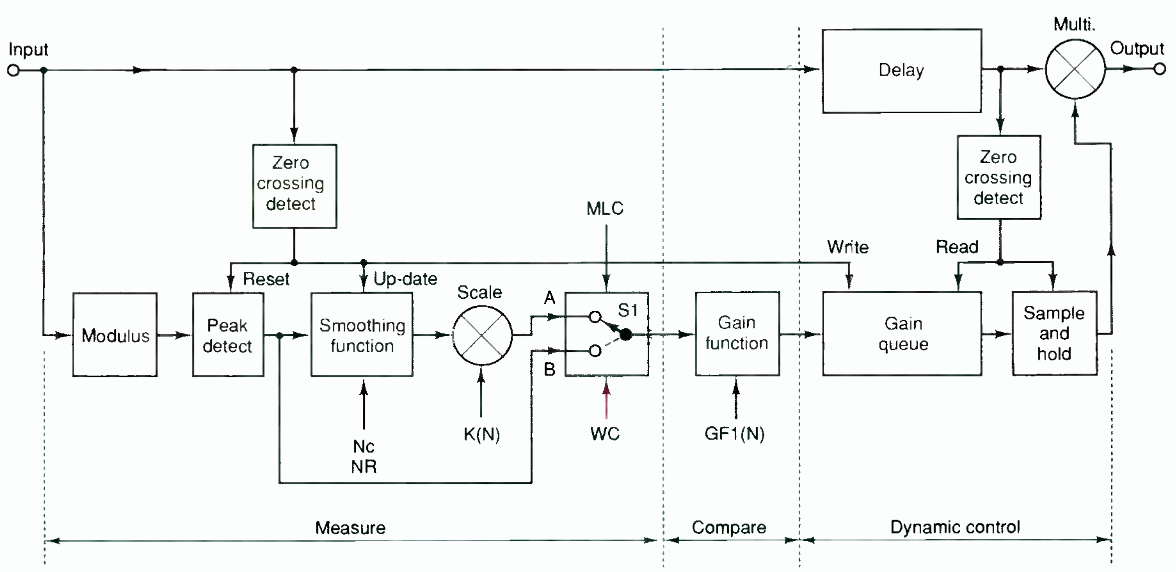


Table 1. Filter time constant as a function of filter constant (m).

Filter constant (m)	2^m	Capture or recovery No (N)	Time constant ($t=CR=2^mT_z$)ms Sinewave (T_z=0.5ms)	Speech (T_z=1.5ms)	Noise (T_z=0.27ms)
0	1	0.5	0.5	1.5	0.27
1	2	1	1	3	0.54
2	4	2	2	6	1.08
3	8	4	4	12	2.2
4	16	8	8	24	4.3
5	32	16	16	48	8.6
6	64	32	32	96	17
7	128	64	64	192	34
8	256	128	128	384	69
9	512	256	256	768	138
10	1024	512	512	1536	276
11	2048	1024	1024	3072	552
12	4096	2048	2048	6144	1105

Clipping drawbacks

An ideal waveshape compressor would raise the level of quieter speech sounds without the listener being aware of a change in quality. This requires a process that amplifies spectral components of the waveform without introducing *any* unnecessary distortion.

Clipping is simple and effective at making speech more intelligible under poor listening conditions. But since it relies on simply chopping off the tops of large-amplitude waves until they are at a similar level to the quieter sounds, intelligibility is severely degraded when listening conditions are good. Clipping introduces harmonic and intermodulation distortion, increases background noise, attenuates low-level tones and produces a non-zero waveform mean value. All these factors degrade speech intelligibility when listening conditions are good. As a result, clipping has few applications outside HF SSB radio.

Attempts have been made to stabilise the clipping ratio by preceding the clipper with a mean loudness controller. Background noise becomes a greater problem and impulse noise can effectively switch off the channel for the duration of the compressor's recovery time.

Attempting to reduce background noise by adding a supersonic tone has also been tried. But as the tone needs to be typically 10dB greater than the noise and 10dB less than the signal, the allowable signal-to-noise ratio is constrained, as is the absolute level of the source.

Companding is sometimes used to enhance radio communication but it only serves to limit channel noise. Fast-acting compressors are more complex and more effective but since they continually adjust gain, they may also introduce significant amounts of subjective distortion and increase background noise.

Current loudness compressors are always a compromise between the need to respond rapidly to transients and the need to vary gain slowly enough to minimise intermodulation distortion.The main difficulty in providing a practical alternative to clipping is in raising the quieter portions of a waveform with a minimum of spectral impurity. To do this, gain adjustment should preferably be carried out at the waveform's zero-crossing instant to minimise distortion. Secondly, for the method to be generally useful in communications, any compressor should introduce a minimum of delay.

limiting range determines the compression level. Conversely, for a given range of limiting, a variation in mean talker loudness causes a change in compression level.

The limiting characteristic of Fig. 4 is shaped so that low level background noise is not increased. Section *AB* of the curve needs to incorporate an incremental expansion to reduce the gain at low levels. To avoid subjective effects, the expansion rate is set no greater than 1.4dB/dB.

A look-up table holds the transfer characteristic but it may be calculated in real time if need be. Other useful curves are "soft knee" and noise-gate functions and a constant expansion rate to allow a further reduction in background noise. The soft knee, that is point B on the diagram if rounded, provides slightly better speech quality at higher compression ratios but reduces the maximum modulation range. With a noise gate function, background noise is reduced without affecting speech levels. Implemented in this way, a noise gate is instantaneous, involving neither capture nor recovery time constants.

Mean loudness controller

Apart from the addition of a digital filter, the loudness controller configuration is the same as that of the waveshape compressor, Fig. 3. The measured quantity of interest, namely the half-cycle peak, is also the same.

To provide a slower variation in gain, the digital filter smooths the series of peak values, Fig. 2.

Gain applied to the waveform depends on a gain function whose input is derived from the digital filter output. The gain function defines the gain inserted to implement the relationship between input peak amplitude and output peak amplitude – the amplitude transfer function.

Functions implemented include:

- a conventional limiter with thresholds of –9,

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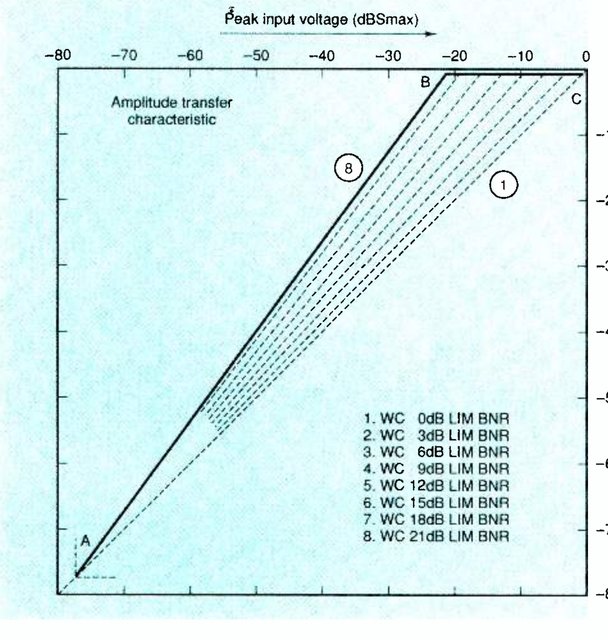
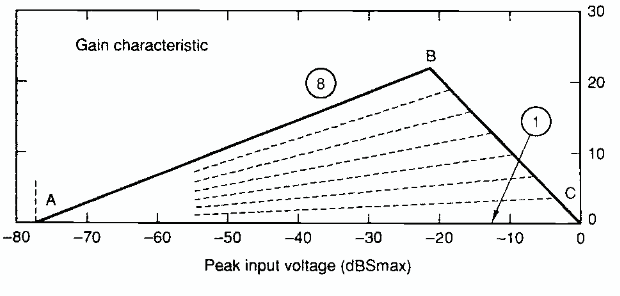


Fig. 4. Waveshape compressor limiter characteristics, shaped to avoid increasing low-level background noise.



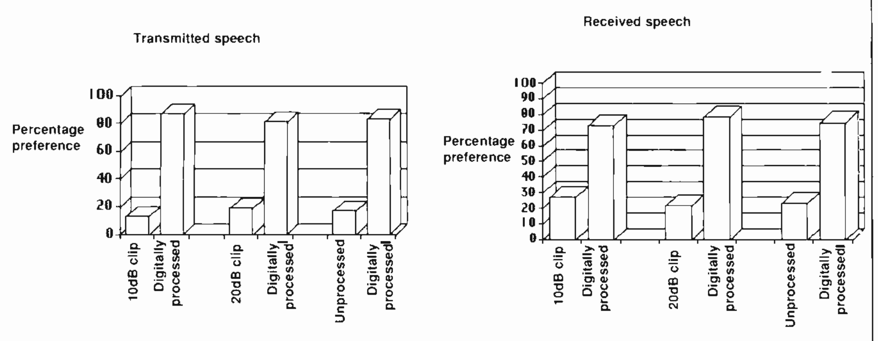


Table 2. Left: speech processed prior to transmission. Right: speech processed after reception. Thirty listeners provided these overall performance scores.

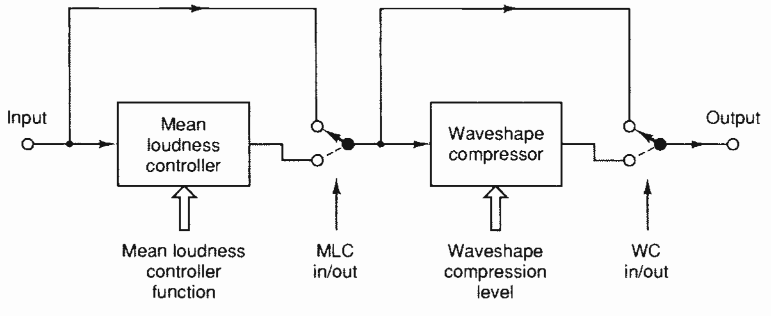


Fig. 5. Mean loudness controller with waveshape compressor in tandem causes no degradation or alteration to either process.

–15, –21 or –27dB relative to peak system voltage, S_{max};

- background noise limiter with the same selectable thresholds;
- 2:1 compressor with and without background noise reduction from –40dB and 1:2 expander, and
- noise gate with –30, –40 or –50dB thresholds relative to S_{max}.

Within certain subjective limits, any function that can be drawn within the input/output space is permissible, including the soft knee type.

The digital filter responsible for smoothing the peak values approximates to a first-order linear low-pass filter with unequal rise and fall times. Apart from when it is updated at zero-crossing instants its output remains constant. In this way, the compressor responds in a given number of half cycles of the input waveform and is synchronous with the input frequency content. Filter constant m_1 defines the filter rise time or capture time and m_2 the fall, or recovery time. Filter constants are set at m_1=0 and m_2=8, and with m_1=0, the compressor has no transient overshoot and the filter is effectively a so-called leaky-peak detector.

In the digital filter, it is better to use the terms capture and recovery cycles instead of half cycles since the defined time constant depends on the particular input waveform.

For telephony-bandwidth speech and noise, the mean zero-crossing periods are typically 1.5 and 0.27ms respectively. Filter performance as a function of filter constant m for speech, noise and a 1kHz sinewave is given in Table 1. The term cycles is used rather than hertz since it is an integer rather than a unit of frequency.

With the recovery number set at 256 cycles the fall times for speech and noise are on average 384 and 69ms respectively. Transient undershoot is thus short and determined in amplitude by the gain function. A smaller recovery number modifies the short-term spectrum of speech and begins to cause waveshape compression and intermodulation distortion.

Filter constants in the waveshape compressor are m_1=m_2=0. Setting fall time at zero provides satisfactory waveshape compression while setting it between unity and about seven does not. This is because these intermediate values cause rapid gain changes that are applied to parts of the waveform with low correlation with the measured peak. With fall time ≥8, gain changes slowly enough to avoid intermodulation distortion.

Applying the theory

The Simitar process is formed from configuration of the mean loudness controller and waveshape compressor in tandem, **Fig. 5.** Because the two processes are almost identical, they can be superimposed, **Fig. 6**, with the bonus of avoiding duplication of signal path delay.

The process is implemented using a *TMS320C17* digital signal processor with 356byte of ram and 4K rom built in. Most of the processor's instructions only require one 200ns cycle. It interfaces to a pair of *A*- or µ-law PCM codecs each operating at 8kHz to provide the 16kHz sampling needed for peaks measurement.

Additional stop-band filtering is needed since that built into the codecs is insufficient. A sixth-order Butterworth filter with 300Hz cut-off provides high-pass filtering while a third-order Butterworth circuit with 3kHz cut-off performs low-pass.

Output samples from the codecs are added and, having noise components of the same amplitude but uncorrelated, provide a theoretical 3dB noise reduction from signal averaging. The whole process is interrupt driven at 16kHz.

Waveshape compressor performance

Performance curves for a limiter with a 21dB limiting range and background noise reduction are shown in **Fig.** 7 (see also Fig. 4). For the DC output, peak amplitude is controlled within ±1dB which represents the combined errors from quantisation and peak measurement.

For AC, the non-zero mean-value resulting from the compression process appears as gain-induced ripple in the overall peak output. At high compression levels, amplitude ripple is ±2dB about peak system voltage, S_{max}.

During gross compressor overload, from inputs of +20dB S_{max} or above, ripple is ±4.4dB maximum. Gross overload is normally accommodated in transmission system specifications and the compressor is no exception.

The peak-to-RMS voltage ratio for speech is reduced from 15dB to a minimum of 6dB at maximum compression. Except under gross overload, the crest factor cannot be reduced

How Simitar differs

In 1948, Licklider showed that speech intelligibility could be increased by raising its mean power level using clipping. Since then various methods of clipping have been applied, differing only in the way in which they minimise the distortion produced.

Dr Thomas's digital waveshape compressor is the first step forward in crest-factor reduction since Licklider's work. It was specifically designed for radio systems with the objective of providing better speech quality than was obtainable from clipping alternatives.

With the technology available at the time they were invented, 1976, the waveshape compressor and mean loudness controller described here were very difficult to implement, requiring an inordinate number of discrete components and ICs. This ruled out the processes for practical applications on both cost and physical size grounds.

But digital signal processors have now allowed the two processes to be combined cost-effectively on a PCB measuring only 110 by 116mm. This is probably the first time ever that a waveshape compressor and mean loudness controller have been combined into a single digital process.

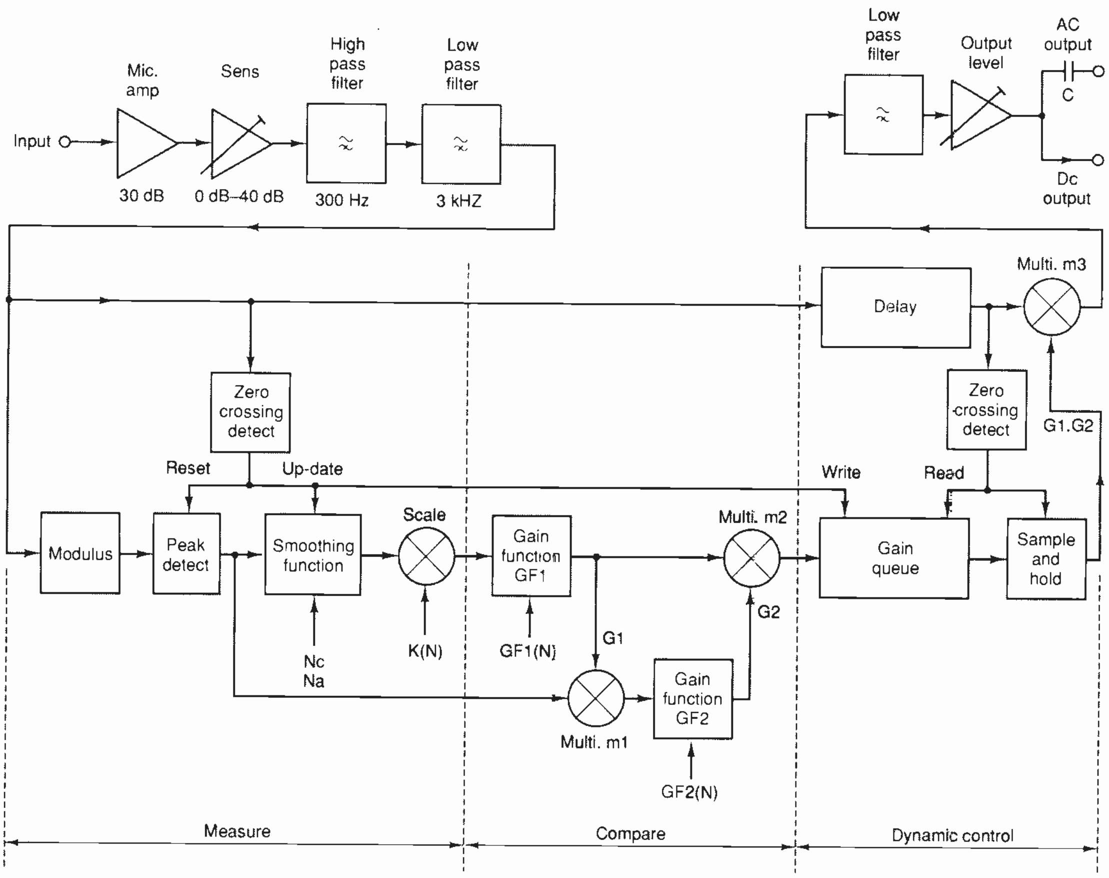


Fig. 6. Being almost identical, the loudness control and waveshape compression processes can be superimposed, avoiding duplication of the signal path delay.

below 6dB so the maximum attainable compression is 9dB. Compression for the average talker is 5dB and gain applied to the overall peak amplitude is 10dB.

Background noise reducing limiter-characteristics generally increase signal levels to a greater extent than noise levels. If input SBNR is low and talker loudness high, the SBNR will be degraded.

Subjective view

Subjective listening tests involve comparing compressed speech with speech that was either uncompressed or subjected to a sideband clipper.

The first test involved speech transmitted over a 7.5MHz SSB carrier connection to a receiver 160km away. Peak envelope power was 200W resulting in a mostly 100% readable signal at *S9*. Compressed speech in this case had noise-reducing limiter characteristics with a –20dB threshold and was compared with:

- uncompressed speech;
- SSB speech clipper with 10dB clipping ratio, and
- SSB clipper with 20dB clipping ratio.

Three levels of speech were tried, namely 0, –6 and –12dBS_{max}, each conditioned with background acoustic noise from a Gaussian noise source to provide SBNRs which were either noise free or with 10 or 15dB of SBNR.

Source material was a five-second sentence in male voice. Listeners comprised 13 "normal" males, 13 males familiar with degraded SSB radio speech and seven females.

Each test was statistically designed using three-by-three superimposed orthogonal squares to eliminate bias due to material presentation order.

Results (**Table 2**) indicate an unconditional selection of the digital speech processor and a clear preference for processed speech over unprocessed, under optimum conditions.

Mean loudness controller

Performance curves for a 21dB limiter with background noise reduction show that overall peak amplitude is controlled to within ±1dB, the sum of quantisation and peak measurement error. RMS output for speech follows the peak output curve but is 15dB below it since expansion and compression effects are absent. Intermodulation and harmonic distortions, contributed by the codecs not the process, are shown as RMS levels relative to system peak voltage S_{max}.

Crest factors of the single and dual-tone test signals are 3 and 6dB respectively.

Overshoot is non existent but undershoot is present in an amount depending on the magnitude of the step decrease in amplitude. Due to the shaping of the gain function, undershoot is reduced or eliminated if the step decrease is placed symmetrically about the knee

In this situation, gain before and after the step remains unchanged. But a temporal effect as the gain tracks back through the knee causes transitory gain elevation though, subjectively, this is not detectable since it occurs

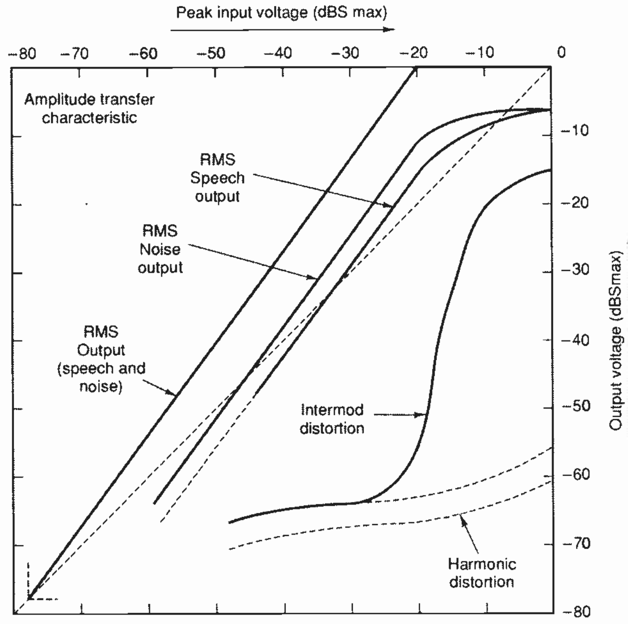
Good news for the deaf

Tests were carried out on fifteen people with impaired hearing, made up of three groups of five people. One group contained people who normally use a hearing aid; one group, people who were not profoundly deaf; the third group of people had been diagnosed as profoundly deaf.

All the groups listened to compressed speech through headphones both from recorded material and from live speech via a microphone. Listeners with moderately impaired hearing preferred compressed speech from the new system to that from their hearing aid – without exception. Those who were more severely impaired, and could not benefit from a hearing aid, all said that they benefited from compressed speech while of the five profoundly deaf people, two said that they were "able to hear sound for the first time".

Under laboratory conditions, improvements could have been expected, so the results can only be taken as an indication of possible usefulness. But the technique offers freedom from the fatigue of background noise enhancement and has higher immunity to the onset of feedback instability due to reduced gain at low levels.

(The above results refer to the waveshape compressor acting alone.)



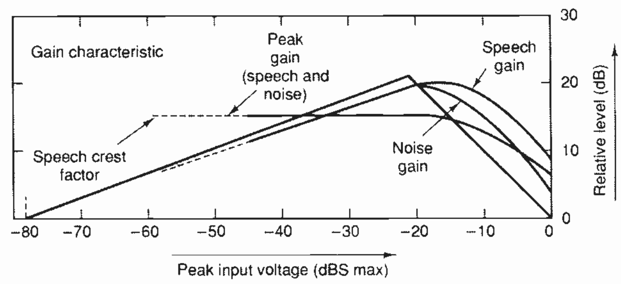
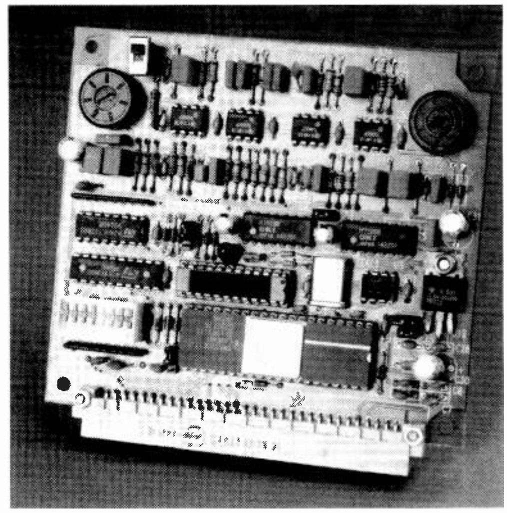


Fig. 7. Performance curves for the waveshape compressor with a limiter having a 21dB limiting range and background noise reduction.



Simitar UdB100 board

within the tail end of sounds. The background noise pumping effect common to analogue controllers is also absent.

The amplitude transfer characteristic provides an improvement in SBNR for all but the loudest talkers at low input SBNR. An SBNR of 30dB or less represents an unusually high level of background noise and is probably fatiguing for the talker. If it is unavoidable then the controller loading may be set at, say, $-15dB_{Smax}$ for the medium talker. The lower quartile of talkers may not provide a limited output but the improved SBNR may give an overall advantage. It is probable that talkers would in any case raise their voice in the presence of high background noise.

The SBNR improvement should be compared with that from a more familiar limiter characteristic.

For these characteristics, background noise is always raised by an amount equal to the limiting range selected.

Any effects due to the response rate being synchronised with the waveform zero crossings are inaudible.

Simitar performance

Configuring the mean loudness controller with the waveshape compressor following in tandem – Simitar – causes no degradation or alteration to either process.

Acting with a limiting amplitude transfer characteristic, the mean loudness controller stabilises the waveshape compression level. There is no detectable change in speech quality for input loudness variations of ±10dB about the medium talker with a 21dB limiter characteristic.

Speech clarity and "brightness" are increased with the addition of quite small amounts of compression with a limiting range of 3 to 6dB. Clarity and loudness are greater with the Simitar process than can be obtained with either the loudness controller or compressor acting independently.

Applications

Performance improvements allow a greater range of applications for level controllers. Other methods which raise background noise are usually sited at the sending end where the benefits are greatest and background noise low.

In a communications environment, an increase in background noise is at least a distraction, and at worst a source of misinformation. The digital compressor may used at the sending end, the receiving end or in between to re-establish levels while usually providing an SBNR improvement.

Modulation levels are improved since maximum output is controlled and without overshoot, but a modulation overhead is still required to accommodate gross overloads.

Compared with simple analogue controllers and clippers, the digital compressor is more complex and more expensive, making it mainly applicable to high-performance systems such as base stations and common apparatus locations. Incorporating it into an existing digital system however brings the cost down significantly.

There is little benefit in adding the digital compressor to a system with an existing analogue compressor since overall performance is determined by the weakest link.

The method shows that speech levels may be maintained and enhanced with improved clarity and lower background noise without compromising naturalness. An option now exists for enhancing transmission performance without incurring the familiar sound of compressed speech.

Fabrication of the process as a monolithic device is possible as the implementation uses readily available technology and devices. ■

Organisations wishing to use the technology should contact DR Easson, The British Technology Group, 101 Newington Causeway, London SE1 6BU (Tel: 071 403 6666 or Fax: 071 403 7586).

Acknowledgements

Thanks to current and former staff of University College Swansea for generous help in developing the waveshape compressor, in particular Dr VJ Phillips, Prof JV Oldfield, Dr M Barton and Dr M Towers. Also to Racal Communications Systems, AB Electronic Products Group and Serc.

For patent provision and development funding for Simitar, thanks are given to The British Technology Group and to Aculab for provision of a demonstrator.

Further reading

Licklider, JCR and Pollack, T. Effects of Differentiation, Integration and Infinite Peak Clipping upon the Intelligibility of Speech, *J Acoust. Soc Am*, 1948 vol 20, No 1, pp. 42-51.

Richards, DL, Statistical Properties of Speech Signals, *Proc IEE*, 1964, Vol 11, No 5, pp. 941-949.

Richards, DL, Telecommunication by speech, Butterworth, London, ISBN 0 408 70344X.

Thomas, LD, MSc thesis, An Investigation into a Non-linear Method of Automatic Level Control for Speech Channel Signals, October 1976, University College of Swansea, Singleton Park, SA2 8PP.Thomas, LD, British Patent Spec No 1599401, Input Signal Level Control, 22 Jan 1977.

Thomas, LD, British Patent Spec No GB2201310B, Signal Level Control, 18 Jan 1988 (continuation patent USA 4947133, FRG application P3801849).

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HP Signal Generator type 8640B – Opt001 + 003 – .5-512Mc/s AM/FM – £1000.
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Racal/Dana 9301A-9302 RF Millivoltmeter – 1.5-2GHz – £250-£400.
Racal/Dana Counters 9915M – 9916 – 9917 – 9921 – £150 to £450. Fitted FX standards.
Racal/Dana Modulation Meter type 9009 – 8Mc/s – 1.5GHz – £250.
Racal – SG Brown Comprehensive Headset Tester (with artificial head) Z1A200/1 – £350.
Marconi AF Power Meter type 893B – £200.
Marconi RCL Bridge type TF2700 – £150.
Marconi/Saunders Signal Sources type – 6058B – 6070A – 6055B – 6059A – 6057B – 6056 – £250-£350. 400Mc/s to 18GHz.
Marconi TF1245 Circuit magnification meter + 1246 & 1247 Oscillators – £100-£300.
Marconi microwave 6600A sweep osc., mainframe with 6650 PI – 18-26.5GHz or 6651 PI – 26.5-40GHz – £1000 or PI only £600.
Marconi distortion meter type TF2331 – £150. TF2331A – £200.
Microwave Systems MOS/3600 Microwave frequency stabilizer – 1GHz to 40GHz £1k.
Tektronix Plug-ins 7A13 – 7A14 – 7A18 – 7A24 – 7A26 – 7A11 – 7M11 – 7S11 – 7D10 – 7S12 – S1 – S2 – S6 – S52 – PG506 – SC504 – SG502 – SG503 – SG504 – DC503 – DC508 – DD501 – WR501 – DM501A – FG501A – TG501 – PG502 – DC505A – FG504 – P.O.R.
Alltech Stoddart receiver type 17/27A – .01–32Mc/s – £2500.
Alltech Stoddart receiver type 37/57 – 30-1000Mc/s – £2500.
Alltech Stoddart receiver type NM65T – 1 to 10GHz – £1500.
Gould J3B Test oscillator + manual – £200.
Infra-red Binoculars in fibre-glass carrying case – tested – £100. Infra-red AFV sights £100.
ACL Field Intensity meter receiver type SR – 209 – 6. Plugs-ins from 5Mc/s to 4GHz – P.O.R.
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Tektronix Mainframes – 7603 – 7623A – 7633 – 7704A – 7844 – 7904 – TM501 – TM503 – TM506 – 7904 – 7834 – 7104.
Knott Polyskanner WM1001 + WM5001 + WM3002 + WM4001 – £500.
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SE Lab Eight Four – FM 4 Channel recorder – £200.
Alltech 757 Spectrum Analyser – 001 22GHz – Digital Storage + Readout – £3000.
Dranetz 606 Power line disturbance analyser – £250.
Precision Aneroid barometers– 900-1050Mb – mechanical digit readout with electronic indicator – battery powered. Housed in polished wood carrying box – tested – £100-£200-£250. 1, 2 or 3.

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HP 3580A LF-spectrum analyser – 5kHz to 50kHz – LED readout – digital storage – £1600 with instruction manual – internal rechargeable battery.
Tektronix 7D20 plug-in 2-channel programmable digitizer – 70 Mc/s – for 7000 mainframes – £500 – manual – £50.
Datron 1065 Auto Cal digital multimeter with instruction manual – £500.
Racal MA 259 FX standard. Output 100kc/s–1Mc/s–5Mc/s – internal NiCad battery – £150.
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EIP 451 microwave pulse counter 18GHz – £1000.
Marconi RF Power Amplifier TF2175 – 1.5Mc/s to 520Mc/s with book – £100.
Marconi 6155A Signal Source – 1 to 2 GHz – LED readout – £600.
Schlumberger 2741 Programmable Microwave Counter – 10Hz to 7.1GHz – £750.
Schlumberger 2720 Programmable Universal Counter 0 to 1250Mc/s – £600.
HP 2225CR Thinkjet Printer – £100.
TEK 576 Calibration Fixture – 067-0597-99 – £250.
HP 8006A Word Generator – £150.
HP 1645A Data Error Analyser – £150.
Texscan Rotary Attenuators – BNC/SMA 0-10-60-100DBS – £50-£150.
HP 809C Slotted Line Carriages – various frequencies to 18GHZ – £100 to £300.
HP 532-536-537 Frequency Meters – various frequencies – £150-£250.
Barr & Stroud variable filter EF3 0.1Hz–100kc/s + high pass + low pass – £150.
S.E. Lab SM215 Mk11 transfer standard voltmeter – 1000 volts.
Alltech Stoddart P7 programmer – £200.
H.P. 6941B multiprogrammer extender. £100.
Fluke Y2000 RTD selector + Fluke 1120A IEEE-488-translator + Fluke 2180 RTD digital thermometer + 9 probes. £350 all three items.
H.P. 6181 DC current source. £150.
H.P. 59501A – HP-IB isolated D/A/power supply programmer.
H.P. 3438A digital multimeter.
H.P. 6177C DC current source. £150.
H.P. 6207B DC power supply.
H.P. 741B AC/DC differential voltmeter standard (old colour) £100.
H.P. 6209B DC power unit.
Fluke 80 high voltage divider.
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Tektronix M2 gated delay calibration fixture. 067-0712-00.
Tektronix precision DC divider calibration fixture. 067-0503-00.
Tektronix overdrive recovery calibration fixture. 067-0608-00.
Avo VCM163 valve tester+book £300.
H.P. 5011T logic trouble shooting kit. £150.
Marconi TF2163S attenuator – 1GHz. £200.
PPM 8000 programmable scanner.
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B&K 4815 calibrator head.
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Farnell power unit H60/50 – £400 tested.
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Racal/Dana 9300 RMS voltmeter – £250.
H.P. sweeper plug-ins – 86240A – 2–8.4GHz – 86260A – 12.4–18GHz – 86260AH03 – 10–15GHz – 86290B – 2–18.6GHz. 86245A 5.9–12.4GHz.
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H.P. 461A amplifier – 1kc–150Mc/s – old colour – £100.
H.P. 8750A storage normalizer.
Tektronix oscilloscopes type 2215A – 60Mc/s – c/w book & probe – £400.
Tektronix monitor type 604 – £100.
Marconi TF2330 or TF2330A wave analysers – £100-£150.
HP5006A Signature Analyser £250 + book.
HP10783A numeric display. £150.
HP 3763A error detector. £250.
Racal/Dana signal generator 9082 – 1.5–520Mc/s – £800.
Racal/Dana signal generator 9082H – 1.5–520Mc/s – £900.
Claude Lyons Compuline – line condition monitor – in case – LMP1+LCM1 £500.
Efratom Atomic FX standard FRT – FRK – 1–1–5–10Mc/s. £3K tested.
Racal 4D recorder – £350 – £450 in carrying bag as new.
HP8350A sweep oscillator mainframe + HP11869A RF PI adaptor – £1500.
Alltech – precision automatic noise figure indicator type 75 – £250.
Adret FX synthesizer 2230A – 1Mc/s. £250.
Tektronix – 7S12–7S14–7T11–7S11–S1–S52–S53.
Rotek 610 AC/DC calibrator. £2K + book.
Marconi TF2512 RF power meter – 10 or 30 watts – 50 ohms – £80.
Marconi multiplex tester type 2830.
Marconi digital simulator type 2828A.
Marconi channel access switch type 2831.
Marconi automatic distortion meter type TF2337A – £150.
Marconi mod meters type TF2304 – £250.
HP 5240A counter – 10Hz to 12.4GHz – £400.
HP 3763A error detector.
HP 8016A word generator.
HP 489A micro-wave amp – 1–2GHz.
HP 8565A spectrum analyser – .01–22GHz – £4k.
HP 5065A rubidium vapour FX standard – £5k.
Fluke 893A differential meters – £100 ea.
Systron Donner counter type 6054B – 20Mc/s–24GHz – LED readout – £1k.
Takeda Riken TR4120 tracking scope + TR1604P digital memory.
EG&G Parc model 4001 indicator + 4203 signal averager PI.
Systron Donner 6120 counter/timer A+B+C inputs – 18GHz – £1k.
Racal/Dana 9083 signal source – two tone – £250.
Systron Donner signal generator 1702 – synthesized to 1GHz – AM/FM.
Systron Donner microwave counter 6057 – 18GHz – Nixey tube – £600.
Racal/Dana synthesized signal generator 9081 – 520Mc/s – AM–FM. £600.
Farnell SSG520 synthesized signal generator – 520Mc/s – £500.
Farnell TTS520 test set – £500 – both £900.
Tektronix plug-ins – AM503 – PG501 – PG508 – PS503A.
Tektronix TM515 mainframe + TM5006 mainframe.
Cole power line monitor T1085 – £250.
Claude Lyons LCM1P line condition monitor – £250.
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Rhodes & Schwarz vector analyser – ZPV+E1+E3 tuners – .3–2000Mc/s.
Bell & Howell TMA3000 tape motion analyser – £250.
Ball Efratom PTB-100 rubidium standard mounted in Tek PI.
Ball Efratom rubidium standard PT2568-FRKL.
Trend Data tester type 100 – £150.
Farnell electronic load type RB1030-35.
Fairchild interference analyser model EMC-25 – 14kc/s–1GHz.
Fluke 1720A instrument controller+keyboard.
Marconi 2442 – microwave counter – 26.5GHz – £1500.
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B&K 7003 tape recorder – £300.
B&K 2425 voltmeter – £150.
B&K 4921+4149 outdoor microphone.
Wiltron sweeper mainframe 610D – £500.
HP3200B VHF oscillator – 10–500Mc/s – £200.
HP3747A selective level measuring set.
HP3586A selective level meter.
HP5345A electronic counter.
HP4815A RF vector impedance meter c/w probe. £500–£600.
Marconi TF2092 noise receiver. A, B or C plus filters.
Marconi TF2091 noise generator. A, B or C plus filters.
Tektronix oscilloscope 485 – 350Mc/s – £500.
HP180TR, HP182T mainframes £300–£500.
Bell & Howell CSM2000B recorders
HP5345A automatic frequency convertor – .015–4GHz.
Fluke 8506A thermal RMS digital multimeter.
HP3581A wave analyser.
Philips panoramic receiver type PM7800 – 1 to 20GHz.
Marconi 6700A sweep oscillator+6730A – 1 to 2GHz.
Wiltron scaler network analyser 560+3 heads. £1k.
R&S signal generator SMS – 0.4–1040Mc/s – £1500.
HP8558B spectrum ANZ PI – .1–1500Mc/s – o/c – £1000. N/C – £1500 – To fit HP180 series mainframe available – £100 to £500.
HP8505A network ANZ + 8503A S parameter test set + 8501A normalizer – £4k.
HP8505A network ANZ + 8502A test set – £3k.
Racal/Dana 9087 signal generator – 1300Mc/s – £2k.
Racal/Dana VLF frequency standard equipment. Tracor receiver type 900A + difference meter type 527E+rubidium standard type 9475 – £2750.
Marconi 6960–6960A power meters with 6910 heads – 10Mc/s – 20GHz or 6912 – 30kHz–4.2GHz – £800–£1000.
HP8444A–HP8444A opt 59 tracking generator £1k–£2k.
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HP8755A scaler ANZ with heads £1k.
Tektronix 475 – 200Mc/s oscilloscopes – £350 less attachments to £500 c/w manual, probes etc.
HP signal generators type 626 – 628 – frequency 10GHz–21GHz.
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HP3730B down convertor – £200.
Bradley oscilloscope calibrator type 192 – £600.
Spectrascope SD330A LF realtime ANZ – 20Hz–50kHz – LED readout – tested – £500.
HP8620A or 8620C sweep generators – £250 to £1k with IEEE.
Barr & Stroud variable filter EF3 0.1Hz–100kc/s+high pass+low pass – £150.
Tektronix 7L12 analyser – .1Mc/s–1.8GHz – £1500 – **7L14 ANZ** – £2k.
Marconi TF2370 spectrum ANZ – 110Mc/s – £1200–£2k.
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Racal receivers – RA17L–RA1217–RA1218–RA1772–RA1792 – P.O.R.
Systron Donner microwave counter 6057 – 18GHz – nixey tube – £600.
HP8614A signal gen 800Mc/s–2.4GHz old colour £200, new colour £400.
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CIRCLE NO. 111 ON REPLY CARD

PC ENGINEERING

Protel's new product makes window shopping worthwhile

After *Easytrax* and *Autotrax* comes Protel's new PCB layout, placement and autorouting package for *Windows*. Reluctant *Windows*–user John Anderson likes what he sees.

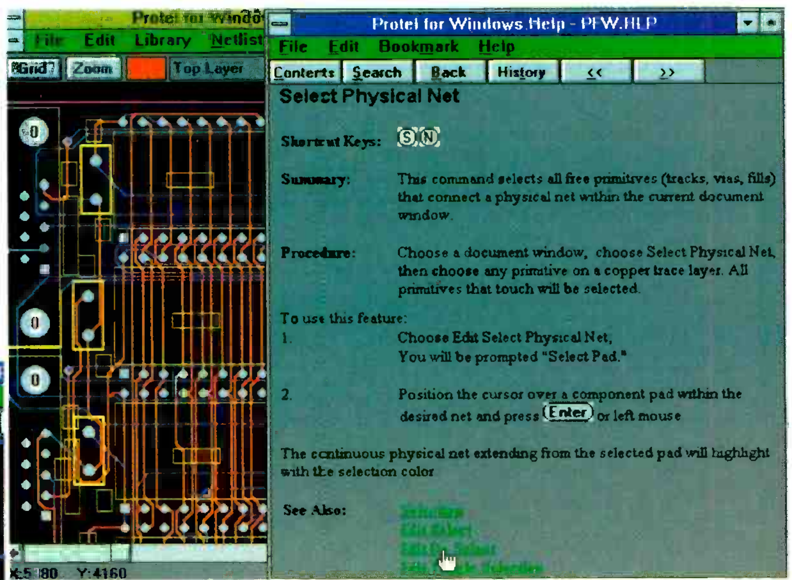
Anyone who has used the earlier non-*Windows* products, will be immediately at home with *Protel for Windows*. The look and feel of the software is similar, and the short-cut key strokes needed to obtain particular results are unchanged. Strangely the package has been launched without schematic capture and requires nets to be input from elsewhere, eg *Protel Schematic*, or typed in directly (supplier JAV says a schematic capture package for *Windows* is scheduled for imminent release).

But the basic graphical user interface offered by *Windows* is ideally suited to a PCB design product, and the common user interface between various *Windows* applications means new users should barely need the reference manual – well that's the theory.

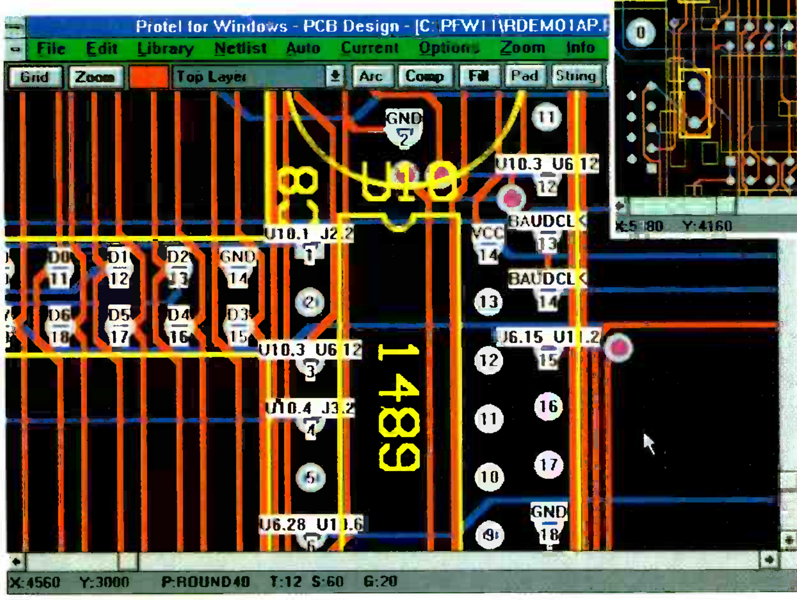
34 layers

PCBs up to 32 x 32in can be handled, with a position resolution of 1mil (1/1000in). In total, 34 layers are supported: up to 16 signal layers, four internal power planes, four mechanical assembly layers, two silkscreen overlays, two resist masks, two paste masks, drill guide, drill drawing, multilayer and DRC errors layers.

The working screen looks similar to any other *Windows* application, with an additional tool bar which allows short-cuts to some of the more commonly used features.

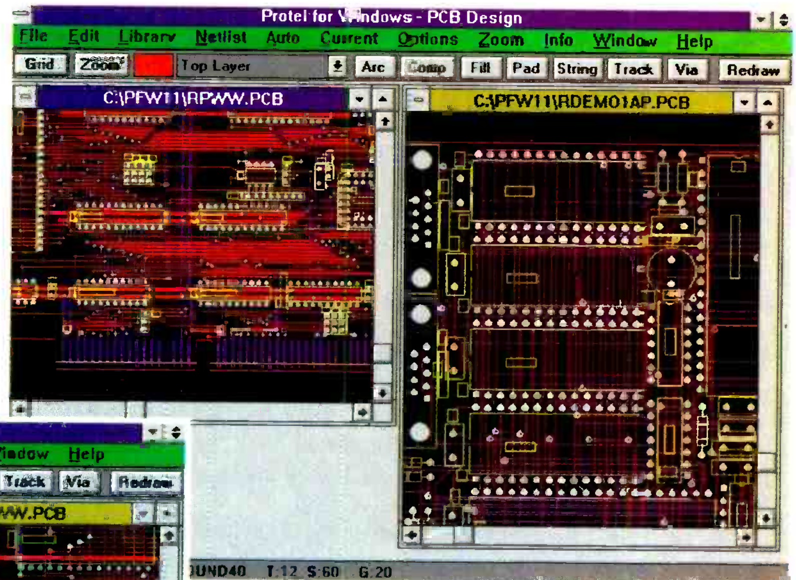


On-line help screens are available for what is already an easy to use package.



Net identifiers shown on a zoomed screen – very effective.

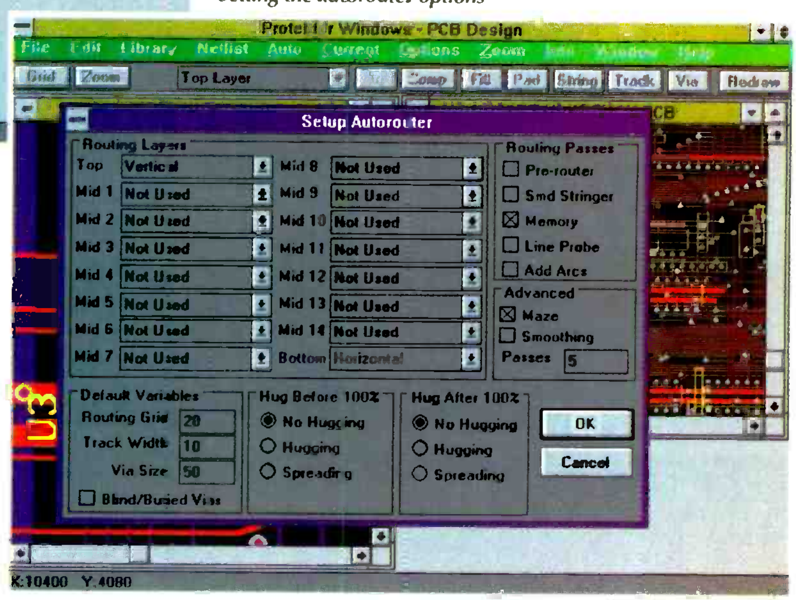
System Requirements
Windows V3.0 or later
Windows standard mode: 80286 processor with 1Mbyte ram
Windows 386 enhanced mode: provides access to Windows virtual memory and multi tasking.
Output to a *Windows* supported pen plotter or printer.
Gerber format photoplotter.
Review machine was a 33MHz 486 machine with 8Mbytes ram



Working on two or more different windows with two different designs is possible.

Showing two different views of the same PCB. There need be no particular linkage between displayed windows.

Setting the autorouter options



Mouse, cursor keys or scroll bars can all be used for moving around the PCB, and zoom level is controlled by page up/page down keys, the zoom toolbar (which is a window zoom function), or the zoom menu from which any zoom can be set.

Unlike most *Windows* applications, *Protel* clears the keyboard buffer after each command, preventing queuing of the short cut key strokes. As a result the user has to wait for the program rather than typing in commands while others are being processed. Frustration is heightened by the fact that keyboard macros – a very useful feature – have been dropped for *Protel for Windows*.

Ground planes can be generated in two ways: the simplistic area-fill scheme, where a ground plane can be built up from a series of rectangles. Or the much better polygon-fill command, where a complex polygon is defined, optionally assigned to a net, and *Protel for Windows* then fills the area with a cross hatch avoiding tracks and pads. The result is neat. But it does take a few tens of seconds to run and produce a large number of tracks. These in turn slow down the redraw time, so it is sensible to use this function only when the main layout is complete.

Manipulating components

Data on pad size and placement for over 300 component footprints are stored in the library, and users can also generate custom components, or indeed modify those provided. Following creation or modification with the program library editor, a "component" may then be loaded to the PCB in the same way as any other.

Prior to routing, components must be placed, either manually, or automatically using an interesting example of topological optimisation. In automatic mode, after the settings for the auto-placement have been arranged, the program moves components around to exclude them from the keep-out regions, and to minimise total net length. The method requires the testing of many possible arrangements, so that, in about an hour, 56000 arrangements were tried on a board of 10 ICs. But the final result seemed to be very good.

Import of the net list allows display of the rat's nest of connectivity data, a simple point-to-point net, and these nets can then be routed manually or automatically.

There is a choice of autorouting options, and of particular interest are hugging (placing tracks close to existing tracks to preserve bard space) and spreading to equalise the distance between tracks to improve manufacturing yield.

Starting up and trading up

Protel for Windows comes as a professionally published spiral bound manual, disk files totalling about 3Mbytes and a parallel port dongle.

The manual is well produced and includes a command reference and a general design topics section covering all aspects of the program. The package is so straightforward to pick up and use that the manual is only of importance when you get stuck. There is also a lot of useful information in the reference supplement that adds up to a manual in its own right. A nice touch is that the exact format of the PCB files (when stored in ascii) is described, allowing alteration with any ascii word processor.

Executing SETUP from the *Windows* file manager installs the package and all unpacking and transfers are accomplished without user intervention. Being *Windows* there is no systems installation to worry about and no tampering with dos system files. The package does however fiddle with the *Windows* installation, introducing its own icon bitmap into the applications group.

For those using the previous Protel products, *Protel for Windows* can import and export *Autotrax* files, although, by default, files are saved in a compressed form. So investment in files and user training in the dos product is not lost with the *Windows* package.

Other options include "SMD stringer" (a scheme to route SMD pads to power planes), a heuristic memory router, and a so called "line probe" which produces "L, C or Z" 1/2 via tracks.

On the subject of surface mount components, these can be handled perfectly well – even in a mix with pin lead devices on both sides of the board. All necessary reverses of the surface mount component patterns are effected and outline data are transferred to the appropriate silk screen.

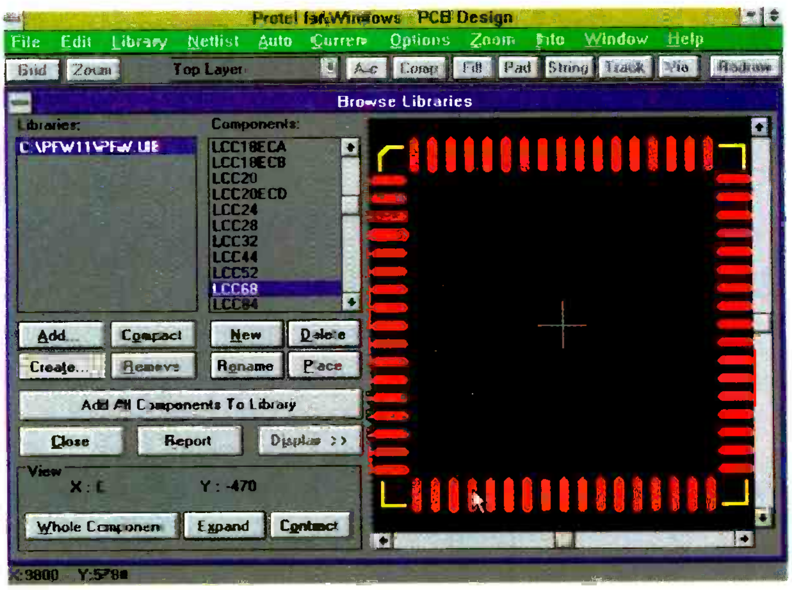
The advanced router offers a maze router with rip-up and retry capability and smoothing – a review scheme to minimise the number of track segments and vias.

Text strings can be placed on any layer and can be set to be one of a variety of sizes and stroke widths, and one of three vector fonts. Special strings such as date, time and layer can be used to identify the PCB automatically.

After manual routing, the program will check that the PCB is routed to the set of design rules, such as track spacing and via-to-track clearance. Although again rather slow this is useful to help keep errors to a minimum.

The dos product had a serious problem in that there was no direct way of highlighting unrouted nets for manual intervention. (For our *Autotrax* system we had to write a utility to map the DRC file to an unused internal layer). Now Protel has copied this approach and allotted a specific layer to DRC error highlights.

Data on pad size and placement for over 300 component footprints are stored in the library.



Hard copy and reporting

Plotting is handled by the *Windows* resource interface. The result is a simpler program, but means there is a tedious rigmarole if you wish to do anything other than output to the default printer.

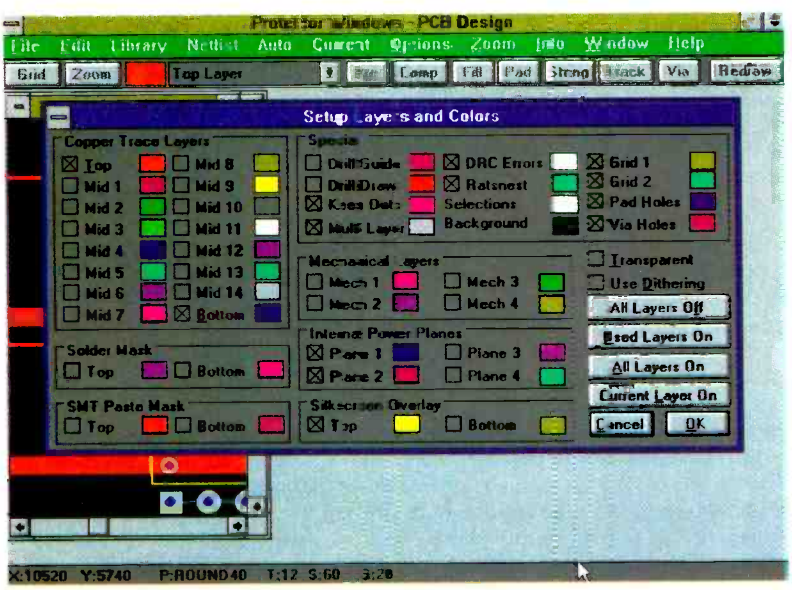
For example, plotting to a file requires switching to Program Manager then Control Panel then Driver Menu, defining the default port to be a file – then finally plot.

Similarly, outputting to a Gerber photoplotter requires a suitable Windows driver.

For reporting, the file pull-down menu has options to produce reports on bills of materials, board specifications and netlists. Bills of materials – and incidentally back annotation – should be handled through the link with schematic capture. But as this was not available at review time, this more logical route could not be followed.

DXF format files can be exported to 2-D drafting packages so that detailed manufacturing drawings can be generated.

Layer set-up and colours

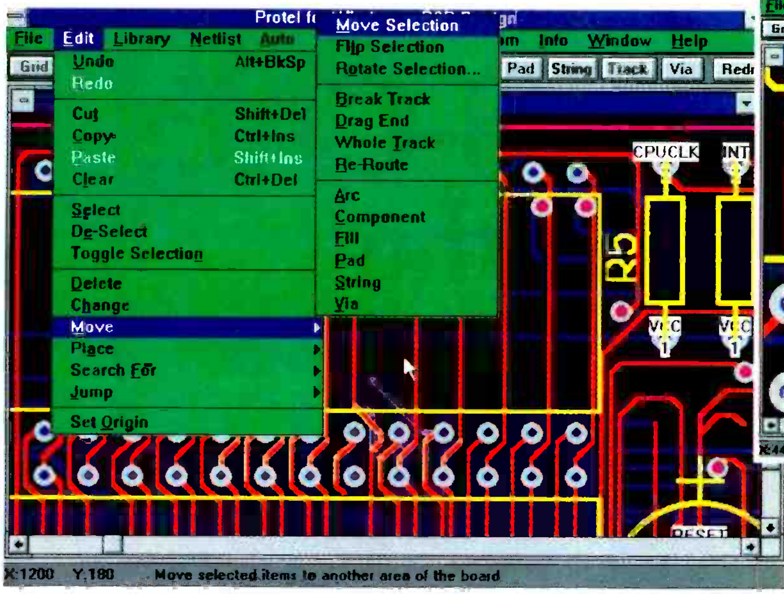


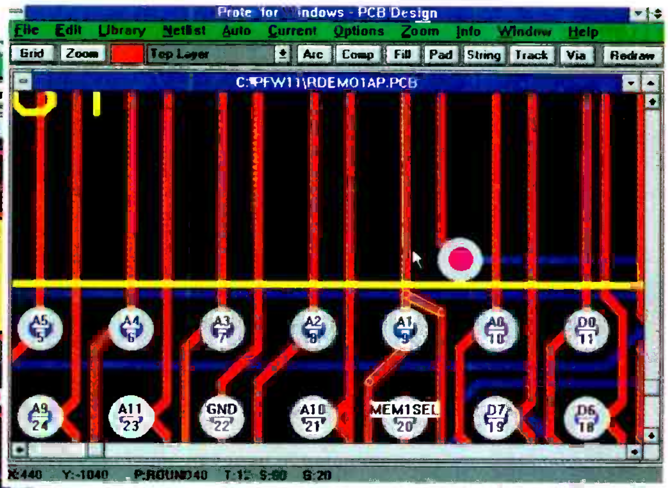
Quality package

Protel for Windows is similar in use to *Autotrax* and remains simple to learn and operate. Unfortunately the lack of a companion schematic capture program to generate the circuit diagram and netlist is a real drawback. A schematic capture package is promised, but until it is the system is seriously flawed. As a result integration between the capture and layout processes could not be assessed, and gate swap or back annotation issues have not been reviewed.

But all the basic facilities are there, though some need to be used with care, particularly the auto placement and autorouting. However once the ground rules have been set for these facilities, they perform well, and the autorouting is significantly better than the earlier *Autotrax* offering.

On-line help is extensive and using *Windows* is easy. It is a simple matter to access the report files (eg the DRC report)





using *Windows* notepad. So the user manual is little needed except for help with formats and the tricks and tips not covered by the help system.

I am no fan of *Windows* and its terrible slowness and hunger for memory. But as packages of this quality are produced, the *Windows* band wagon will become unstoppable. A truly excellent product. ■

Protel for Windows' design rule check requires the testing of thousands of possible arrangements.

Highlighting DRC violations.

Supplier details

Protel professional manual router £650; Protel advanced auto route/place £1300; Protel rip up/retry router £650; Protel auto component placement £650 Protel professional schematic £650. Digital simulation £1300. 15% discount for cash with order.

JAV Electronics Ltd, Gallery House, 677 Manchester Road, Denton, Manchester M34

WINDOWS-CAE: more than just the easy option?

Future promise of true multi-tasking, plus growing numbers of suppliers moving into Windows products means engineers are taking a new look at Windows-cae. Shaz Horner reports on packages and facilities.

Microsoft *Windows*, introduced to make the PC easier to use, is set to become a familiar sight to engineers as PC cae suppliers rush to redevelop their products to run on *Windows*.

For engineers new to cae, working with the *Windows* graphical user interface is a painless route to rapid learning. But even seasoned cae users can gain something from *Windows* – hopping around between applications for example, or moving files without having to exit the cae package.

True multi-tasking is not possible with *Windows*. Several windows can open at the same time, but this is different to Unix-based cae systems which allow several applications to be run and viewed within *X-Windows*: However, release of Microsoft's *Windows NT* promises to provide a multi-tasking operating system in the same league as Unix. *NT* should run dos/*Windows* applications as well, so current *Windows* users will be able to upgrade their dos to *NT* and run existing applications on the new operating system.

Easy integration of software

As well as ease of use, there are other more fundamental benefits that can be gained by running cae applications under an object oriented user interface like *Windows*. One of the major advantages is that applications can be more tightly integrated through the *Windows* dynamic data exchange (DDE) – a mechanism allowing related or linked data existing in separate files to be updated simultaneously. This would allow, for example, back annotation from PCB layout to schematic to be carried out in real time.

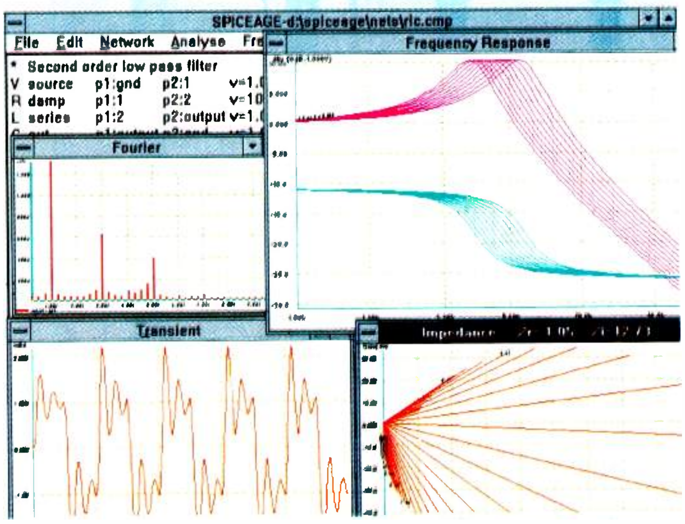
DDE also makes possible close integration of applications from separate vendors, probably benefitting vendors as much as users. This allows cae to be linked with ease into DTP or wordprocessing for writing documentation and technical manuals. In addition, object linking and embedding enables such links to be established over a network of PCs, with obvious advantages for teams of engineers working in a networked PC environment.

Cae suppliers moving their products onto *Windows* have taken the opportunity to tweak their products.

Charles Clarke of Those Engineers, points out that Windows memory management makes it impossible to run out of memory. *SpiceAge for Windows* simulation has been built from 25,000 lines of object oriented code. But being able to use larger programs has, for example, allowed more

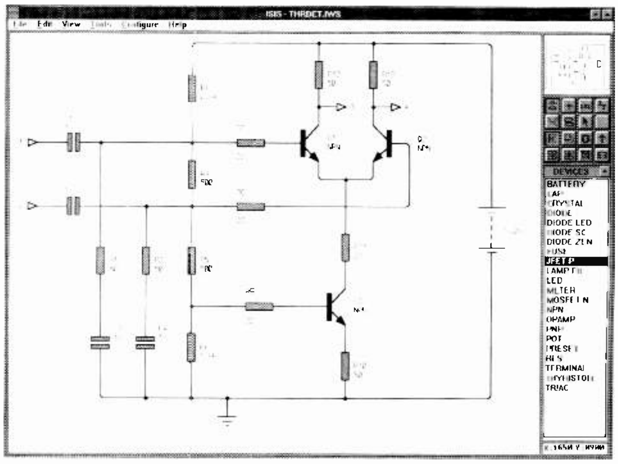
Spice Age for Windows (simulation) includes enhanced AC, non linear DC, non linear transient and Fourier analysis with tolerancing, component value sweeping, temp. distribution, noise generation and a scope mode for viewing results as they are being calculated. Output functions include, gain, phase, group delay, complex plane, real and imaginary impedances, DC quiescent voltages, time dependent voltages. Auto-scaleable and selectable units, reference levels. Frequency zooming in Fourier analysis with a Hanning window option. Models can be created or changed and op-amps, transistors, fets, thyristors, triacs, diodes, etc are supplied. An intelligent circuit editor is provided for tidying syntax including that generated by other programmes.

Requires *Windows* 3.0 or 3.1 running in standard or 386 mode. Maths coprocessor is optional. £395 for single user.
Those Engineers Ltd 31 Birkbeck Road, Mill Hill, London NW7 4BP. Tel: 081 906 0155.



Intelligent Schematic Input System (Isis) is a schematic drawing package giving full control of line width, fill style and colour of all elements of the schematic, curved or angular wire corners, round, square or diamond junction dots, access to all *Windows* fonts, automatic wire routing and dot placement. 2-D drawing capability with symbol library. A comprehensive device library is standard and new devices can be created directly on the drawing. Display uses high resolution *Windows* display drivers. Output is to any bitmapped *Windows* printer device or to the clipboard. The package loads Isis Supersketch and designer files directly

Requires compatible PC with 286 upwards. 2Mbyte ram, 2Mbyte hard disk space, mouse with *Windows* driver, and *Windows* 3.X £179.
Labcenter Electronics, 14 Marriner Drive, Bradford, BD9 4JT, Tel: 0274 542868.



flexible plotting to be provided. So users can now plot in the complex plane and graph real and imaginary components of impedance. Phasor diagrams can be plotted with some of the dependent variables swapped around to the X axis to produce a graph of residue figures. Bigger libraries can be developed too. *SpiceAge for Windows* includes new facilities such as bipolar junction transistor models, jfet models and diodes.

Falling resistance

Despite all the benefits, some users are still reluctant to use *Windows*, assuming *Windows*-based cae is only for trainees. But ease of use, and improved features - closer to those offered by the more powerful Unix-based cae systems at maybe half the price - are making the interface harder to resist.

Suppliers believe the main reason stopping users upgrading to *Windows* based cae is that many are still using 286 PCs. 286s can run *Windows* but performance suffers. For example screen refresh can be frustratingly slow.

But cae suppliers say demand for *Windows* products is growing, to such an extent that *Windows* packages are becoming their best sellers. Once the graphical interface has been tasted it is difficult to return to the not-so-friendly dos environment, and the lowering cost of 386 machines means more users can be expected to upgrade their hardware and move to *Windows*-based cae. What is more, as *NT* becomes available with its true multitasking - depending on how competitively it is priced - the dividing gap between Unix-based cae and the *Windows* cae look set to disappear. ■

Design Cad 2D for schematic has a curve capability and tangent/perpendicular functions. *BasicCad* programming language with its *Basic*-type syntax for writing commands or complete programmes is included. Icons menus can be customised.

Requires: Compatible PC, *Windows* 3.0. £275.
PMS Instruments Ltd, Waldeck House, Reform Road, Maidenhead, Berks SL6 8BR. Tel: 0628 38036.

Protel for Windows for PCB design is modular in construction so that it can be purchased either as a full package (Advance Pack) or by different functions: maze routing, and global placement can be purchased separately. Features include 32x32 inch workspace, 20 layer capacity with blind/buried vias, complete Gerber tool set. Imil placement resolution. Netlists can be loaded from popular schematic capture formats.

Requires *Windows* 3.0 or 3.1 running in standard- or 386-enhanced mode. Min 2Mbyte ram. 3Mbyte of hard disk is required for installation. £2,200 for (top of the range) Advance Pack.
JAV, Electronics Ltd, The Gallery House, 677 Manchester Road, Denton, Manchester M34 2NA Tel: 061 320 7210.

Design Centre 3 for integrated simulation and schematic capture. Schematic capture provides graphical circuit editing, complete symbol library, and a fully integrated means of simulating and graphically analysing circuit behaviour.

Mixed mode circuit simulation with *Pspice* includes analogue behaviourial modelling (non-linear and frequency-domain transfer functions require no programming); Monte Carlo and sensitivity/worst-case statistical analysis; fully integrated event-driven digital simulation supporting state, strength, and timing modelling, as well worst timing simulation; extensive analogue and digital libraries. Device equations can be customised to tailor a device's behaviour by supplying partial code which can be modified. Graphical waveform analysis allows simultaneous viewing of analogue waveforms and digital signals produced by *Pspice* simulations.

Requires 386/486 or compatibles, 640K memory, 3Mbyte of extended memory, 387 floating point processor, *Windows 3.0*, 386 enhanced mode or standard. £59.95.
ARS Microsystems, Herriard Business Centre, Alton Road, Basingstoke Hants RG25 2PN. Tel: 0256 381400.

Quickroute 2.0 for Windows is a schematic design package that allows point and pick editing, curved tracks and filled polygons. Schematic capture has Spice-compatible net-list generation. Imperial or metric units – resolution 1mil. Seven preset and one custom zoom. Turbo draw speeds up redraw times by drawing objects as outlines. Up to eight PCB schematic layers plus two silk screen layers, a solder mask and rats-nest layer handled. Five different types of object can be placed: pads, tracks, ICs, symbols and graphic primitives.

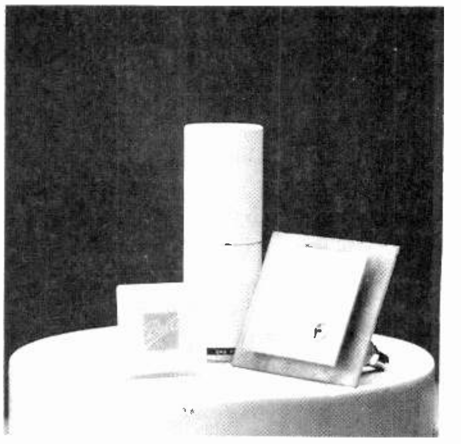
Requires *Windows 3/3.1*. £59.
Powerware, 14 Ley Lane, Marple Bridge, Stockport SK6 5DD. Tel: (6-9pm only) 0860 602486.

LabWindows data acquisition and processing package is a powerful aid for the development engineer producing code to load up a microprocessor. The product has been designed to generate code for processing data derived from instruments via the IEEE488, RS232 or PC expansion cards. *QuickBasic* and *C* are both available as programming languages. When creating code in *LabWindows* the designer, in effect, assembles a list of library subroutines to perform the required task. Users can single-step through code while watching the effect on variables in another window.

Requires 386/486 PC and 4Mbyte extended ram. An *80387* coprocessor takes advantage of routines contained in the *Advanced Analysis* version.
LabWindows 2.2 £783, *Advanced Analysis* £1685.
National Instruments UK Ltd, 21 Kingfisher Court, Hambridge Road, Newbury, Berkshire RG14 5SL. Tel: 0635 523545.

GPS

2: receiver architecture



Aerials for GPS: The signal modulation represents less than 1% of carrier frequency but the element bandwidths must be made much wider to allow for manufacturing tolerances.

GPS signals are a curious mix of narrowband information imposed on wideband signals showing varying degrees of doppler shift. Reception requires a radical approach to receiver design. By Philip Mattos*

The signal available to the commercial GPS receiver is −160dBW, or −130dBm total, spread over about 2MHz bandwidth by the spread spectrum coding, though most of the power can be found in the central 1MHz section. The thermal noise power in 1MHz, derived from Boltzman's constant KTB, is −114dBm using a perfect receiver, or −112dBm if the radio front end achieves a 2dB noise figure.

Thus the receiver starts with a theoretical signal to noise ratio of about −18dB. In practice the antenna may have a few dB of gain, but the required hemispherical response limits aerial gain. Losses in connectors and coax before the LNA may actually nullify aerial gain. The satellites also transmit a few dB above spec, but the equipment must be designed to operate at minimum signal strength.

To give an idea of received power, −160dBW into 50Ω is equivalent, as a single CW carrier, to about 70nV. A good VHF communications receiver expects almost a microvolt. But the GPS receiver must take in 1MHz of bandwidth – compared with 25kHz for the VHF comms receiver – so gets 40 times the noise power.

Thus the GPS receiver has to separate a 70nV signal equivalent from under about 600nV of equivalent noise – an interesting challenge.

Note that these voltage equivalents barely illustrate the problem – they relate to single sinewave power. In real GPS both the noise and the signal are spread over a wide bandwidth.

GPS antennae

The GPS antenna must receive from the satellites wherever they are in the sky, and because it must accommodate several satellites at the same time, it must cover the entire sky. (The exception is certain military arrays, which can have multiple lobes electronically steered to satellites and nulls to interferers.)

GPS systems use two classic aerial types, the quadrifilar helix and the patch. The first resembles a vertical cylinder, and inside has four helical wires with a phased feed network. (**Fig. 1**)

The second is simply an area of copper, of electrical halfwave tuned size, on a high quality dielectric such as PTFE or ceramic. The phased feeds are often printed onto the same substrate. (**Fig. 2**)

The purpose of the feed network is to optimise the response to right hand circularly polarised signals from the satellite, while minimising the response to reflections which tend to be left hand polarised from metallic surfaces.

The quad helix can be designed to give a true hemispherical response, but for sailing vessels is usually designed to cover an even greater solid angle, so that when the vessel is heeling by say 30°, the windward horizon is still covered. (**Fig. 3**)

For surveying applications, it is usually optimised for 25° and higher as using satellites below this angle increases the position error due to ionospheric diffraction.

The patch antenna naturally has a very good response overhead, getting progressively worse nearer the horizon. Many alternative

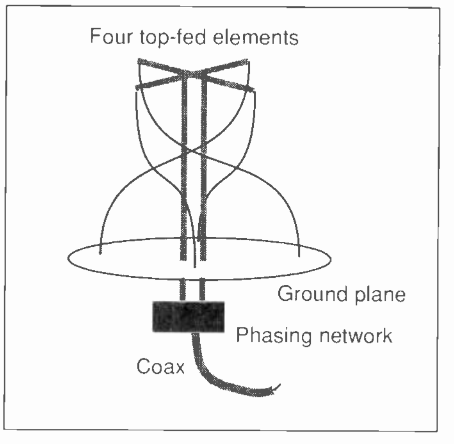


Fig. 1. The Quad-helix antenna achieves hemispherical coverage. Four correctly phased elements provide horizontal coverage, while the twist in the helix allows some endfire response. The proportion between the two can be adjusted by altering the spacing between the active elements and the integral groundplane. It maintains a good ellipticity performance at all angles.

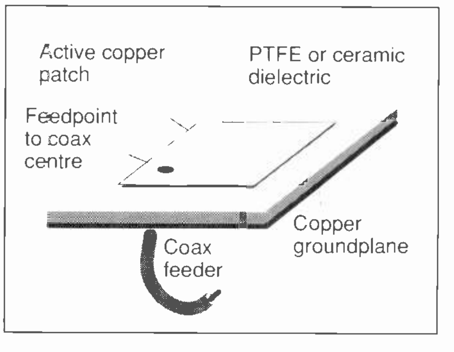


Fig. 2. The patch antenna, an electrical halfwave per side set against a reflector backplane, costs far less to produce, and is more easily faired into streamlined structures such as cars and aeroplanes. It has better gain at the zenith, dropping off towards the horizon, and also losing circular polarisation and becoming linear near the horizon.

*Consultant engineer, Inmos Ltd

designs exist to improve this, and the best offer about 0dBiC on the horizon and about +7dBiC overhead. Clearly the average over the entire hemisphere cannot exceed +3dBiC, and the 0 to +7dB figures are compatible because the solid angle around the zenith is so much smaller than that around the horizon. (**Fig. 4**)

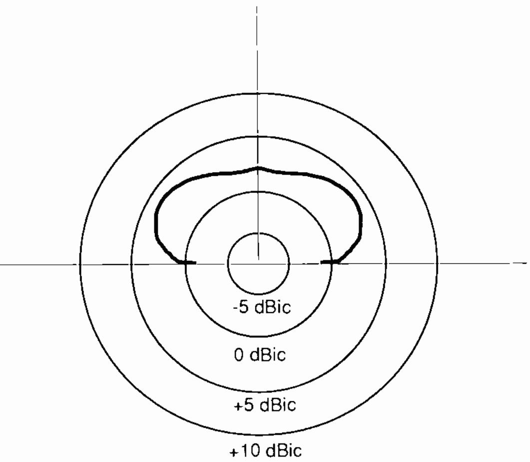


Fig. 3. The quad-helix comes close to the ideal gain of 3dBiC over the entire hemisphere.

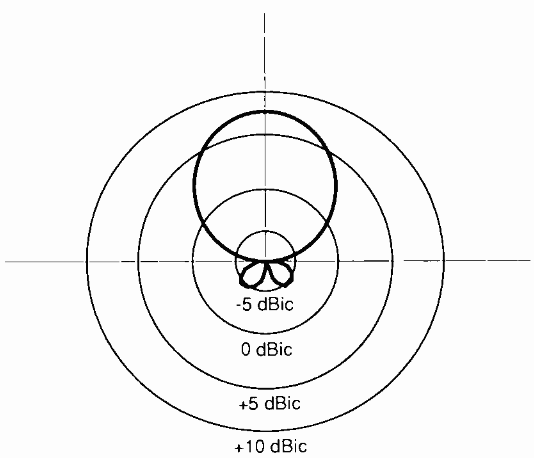


Fig. 4. With up to +7dBiC overhead, the patch antenna is very useful on land, where low angle satellites are usually blocked by trees/buildings anyway.

For land mobile use the patch is very suitable, as at low angles, trees mountains and buildings obstruct the satellites anyway. The military GPS specification assumes that the antenna can receive from satellites down to 5° above the horizon. This is only practical when the physical horizon allows, such as at sea or in the air.

Because the limiting signal strength is greater for acquisition than tracking, most receivers can track satellites right down to the physical horizon, but do not attempt to acquire them below a predefined mask angle.

Radio requirements

The very low power levels of the received signal described above put considerable restrictions on the radio design, and translating a signal from 1.5GHz to baseband requires minimal phase noise and great stability on the local oscillators. On the other hand, being a wideband signal, tight IF filters are not required, but their phase response is critical.

The intrinsic bandwidth of the signal, before spreading in the satellite, and after despreading in the receiver, is only 100Hz, twice the data rate of 50 bits per second. While it is passing through the analogue sections of the radio, it is 1MHz wide. Any distortion that changes the relative phases of the components of the signal, or spreads a spectral line, will prevent the despreading operation from bringing the energy back together to the 100Hz wide original, and thus reduce the peak energy that can be recovered.

If the reference oscillator for the system is a 50ppm crystal over temperature, the carrier will appear to drift by 75kHz. This dwarfs the errors caused by the Doppler shift of the satellite, and that of the user. When the final processed signal is fed through a 100Hz wide filter, it can be seen that even 0.1 ppm (150 Hz) would lose the signal. Thus measures must be taken to either control, measure or accommodate drift in the reference oscillator.

Traditional radio architectures use double down conversion superhet circuitry with a TCXO reference, and I and Q outputs. (For abbreviations see glossary). They also tend to be designed traditionally, by skilled designers experienced with narrowband radios. A clean sheet approach can yield a much lower cost design of equivalent performance specifically for spread spectrum use, and will be covered in a future article.

The choice of intermediate frequencies is up to the designer, but with two policy decisions made first: is the second IF, the handover to the signal processing, done at baseband, or at some higher frequency, and are the frequencies to be code synchronous or not? If the handover frequency is baseband, amplifiers must pass DC, if not in the radio, certainly after despreading in hardware. However DC does remove the need for image filters, removes a spurious response, and using I+Q the phase/frequency of the signal can be recovered. The classic double conversion architecture is shown in **Fig. 5**.

Early radio architectures from the 1970s and early 80s always used a code-synchronous frequency plan. All the frequencies in the satellite are generated from a single source, at 1.023 MHz, known as f_0. The C/A code chipping rate is this and the P-code rate (military) ten times it, and the carrier at 1575.42MHz is 1540 times it. Thus 10.23MHz is a very useful reference frequency, as is any multiple thereof.

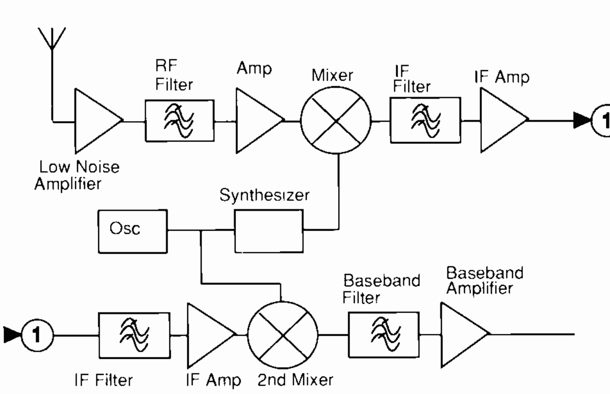
While excellent for hardware code tracking (discussed later) synchronous designs suffer from the fact that 20.46, 10.23 and any sub multiples thereof generate harmonics at the GPS carrier frequency, and due to the extreme sensitivity of the radio, these are picked up by the antenna and swamp the set. This was not a major problem on naval/aircraft sets, as the antenna was well away from the main radio, but once portable sets were developed, it became serious. Modern sets use a 40.92MHz reference if they wish to operate synchronously, as the 38th and 39th harmonics span the GPS carrier cleanly. Few synthesisers will operate with such a high reference, so it is divided down defeating the object. As a result, most sets operate asynchronously to the satellites.

Note the difference between synchronous and coherent. Synchronous means at a multiple of the satellite's code-chipping rate. Coherent means that all the frequencies in the set are derived from the same source, eg first LO, second LO, code generator reference, sampler clock, CPU clock, etc. Coherence is a definite advantage, as it minimises the unknowns in the equations, and also the intermodulation products in the receiver.

The first IF must be chosen to allow the front end filters to remove the spurious response adequately at the image frequency, twice the IF away from the carrier. Thus anything below about 40MHz is unreasonable, as the image response at 80MHz away, or 5% away, would demand excessive Q from the RF filters. The RF filters are usually ceramic resonators, and can be bought off the shelf for a reasonable cost, such is the demand for GPS. They are available with two to four poles, and skirts 30-40dB down, so at least two are usually used.

The two major interfering sources for GPS are terrestrial microwave links, and radar,

Fig. 5. A full traditional dual conversion radio resembles communication radios except in the nature of the IF filters, which do not need to select a channel, rather just reject an image. Only the baseband filter deliberately controls the channel bandwidth, as the phase response is critical.



while a co-located mobile satcomms uplink at 1620MHz can burnout the low noise amplifier if badly sited!

If the final handover IF is not DC, the percentage bandwidth of the filters before the second mixer must be considered. If a 5MHz final output was required, the image would be 10MHz away. On a first IF of 500MHz this would not be appropriate, but at 70MHz, a satcomms standard, there would be no problem. However if the final frequency were to be 2MHz, then even 70MHz would be too high: 71-73MHz is required, without phase distortion. But the image at 67-69MHz must be rejected, albeit only for noise purposes. There will not be any transmitters at 1571MHz, its equivalent at RF, as it is in the reserved GPS band. Remembering that the passband must be 1-2MHz wide, phase characteristics dictate an IF considerably higher than the bandwidth.

If the final handover is to be DC, the second image problem disappears, but a different problem arises. The second local oscillator frequency will be the same as the first IF, and if there is any leakage into the signal path, it can saturate and thus block the IF strip.

Thus the conventional double downconversion design has some problems, and must make the best compromise between them.

Signal path

The first stage of the radio is a low noise amplifier, though it is often sited at the antenna, rather than in the radio itself. While semiconductor technology permits the construction of 1dB noise figure amplifiers, 2 or 3dB is more usual, for cost, reproducibility and protection reasons.

For protection reasons, it is convenient to put a filter before the first transistor, and this necessarily adds the loss of that filter, about 2dB, to the noise figure of the amplifier. This rather defeats the benefit of using a Gallium Arsenide first stage. A solution is to use a robust silicon bipolar first stage with the filter after it.

The LNA must have enough gain such that the signal is boosted to a level where the noise from the second and subsequent stages is negligible by comparison. 15dB is the usual minimum, though up to 45dB is available commercially. Clearly if a long downlead is used, the gain must also cover the coax losses.

At the front end of the radio proper there is normally further gain, and another filter before the first mixer, then a low pass filter to remove RF and LO components before the IF strip. This is necessary because inductors and capacitors used in the IF strip do not retain their values at gigahertz frequencies, so the effect of such frequencies is very ill-defined, both in input impedance and in spurious responses from the IF amplifiers.

The IF strip is usually conventional except for the need to pass a wideband signal with linear phase dictating a fairly low Q for the filters. This is the cheapest place to provide gain, and as some 120dB is needed overall, more than a third of it is provided here.

Equipment with I and Q outputs requires

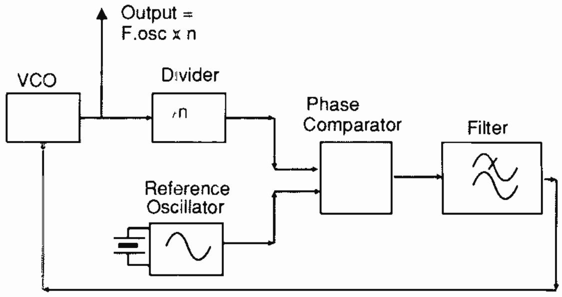


Fig. 6. While most synthesisers are complex designs making use of chips designed for multichannel communication radios, once an asic is included in the radio, a more simple fixed frequency version can be used, benefiting in phase noise from a higher comparison frequency.

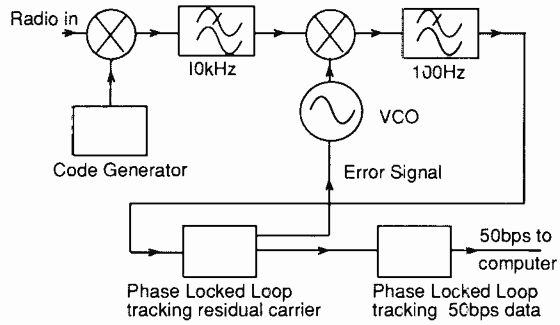


Fig. 7. Tracking loops implemented entirely in hardware were the norm before the day of the micro. They consist of two interlocked loops, one tracking the spreading code, the other the carrier frequency. Initial acquisition is difficult as if either loop is misadjusted, no signal is seen.

two second mixers, fed with local oscillator signals 90° apart in phase. These are followed by the baseband gain, be it DC or a few MHz, and filtering, producing the signal to be fed to the sampling stage.

The sampling stage consists of a primitive A-D converter, normally 2-bits, sampling at video rates. This is fairly recent. Earlier systems would have fed a local satellite code into the radio stage, which would have been narrow band filtered with further gain before digitisation.

The local oscillators of the radio are normally derived from a reference crystal using a synthesiser. Thus the critical first LO is actually a VCO running at either L-band or half that with a passive doubler, the output of the VCO being divided down to a few hundred kilohertz and phase locked to the reference crystal, itself divided down. (**Fig. 6**)

While such synthesisers are readily available, derived from multichannel communications products, it is inevitably a compromise. Since GPS is fixed frequency, a programmable synthesizer is overkill. Specifically, the available chips expect 25 or 50kHz channel spacing, so the reference frequency is designed to be around there. This is much too low for acceptable phase noise at L-band, even if the synthesiser chips are pushed to their limits of a few hundred kilohertz.

If an asic or custom silicon is allowed in the radio, a simple fixed frequency synthesiser with a comparison frequency of a few megahertz is sensible. The cheapest and most effective solution, however, is a high frequency crystal and harmonic multiplier stages.

Hardware tracking loops

The original receivers processed the signal in analogue hardware right through the tracking loops, and only digitised the final output to assess the signal strength and polarity, both to control the loops and to extract data. (**Fig. 7**)

By the late 80s, the phase locked loop that removed the last of the carrier had migrated into the CPU through software together with narrowband filters (**Fig. 8**). This meant that a much faster sampling rate was needed, to handle the frequency errors and Doppler even if the handover frequency was set at DC.

By 1990, with asic hardware progressing, some carrier handling went back to the hardware as it allowed coherent receivers to guide their code trackers from the carrier tracker.

My own design continued to move more and more to software, putting the entire code tracker and carrier tracker in software. The original 1970s designs most clearly illustrate the work that must be done on the signal. (**Fig. 7**)

A pair of fed back shift registers generates a mimic of the satellite spreading code. (**Fig. 9**) Initially this code generator is run a few parts per million fast or slow, to ensure that it drifts past the correct synchronisation point. The drifting must be slow, so that the integrators have time to build the signal.

When this correct sync point is reached, the output from the code mixer becomes a narrow band carrier at a low frequency, that being simply oscillator error and Doppler shift. The filter following the mixer then realises the correlation gain, as it reduces the bandwidth to a few hundred hertz, and here lies the problem. The oscillator error and Doppler error may be a few tens of kilohertz, so the signal misses the filter. Thus in addition to sweeping the code generator in the time domain, one of the downconverting local oscillators must be swept in frequency.

A magnitude detector is used, with a wider filter, about a kilohertz, during acquisition.

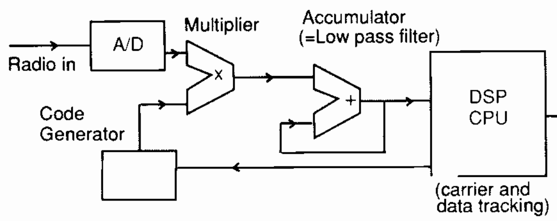
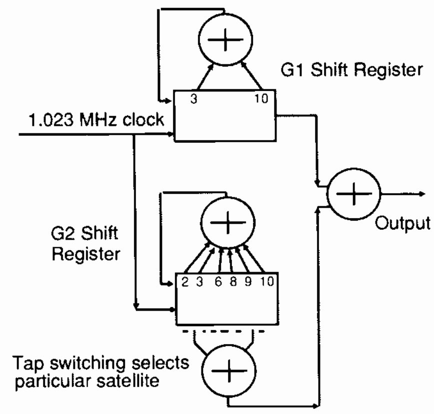


Fig. 8. Most modern tracking loops are digital, be they hardware or software. Few processors can handle the civil code in software, and none the military code, so a hardware code generator, multiplier and accumulator /integrator are used, but their control algorithms, and the carrier loop, are all in software.

Fig. 9. A hardware code generator is very simple, being two fed-back shift registers generating maximal length sequences that are XORed together. Tap selection generates different codes for each satellite. This circuit is used in both satellite and receiver, but the receiver is complicated by the need to search in the time domain, so the 1.023MHz clock must be adjustable first to search, and then to track, the timing of the incoming signal.



Once a likely cell is found, the code tracker is run at the correct, not the offset, rate, and the frequency sweep locked. This allows the phase locked loop to lock to the residual frequency. A final mixing operation can then yield the data waveform to be sampled by the processor.

It is the slow drift rate of the code through its 1023 possibilities, times say 10,000 frequency bins to search, plus the PLL lockup time on each likely candidate, that makes signal acquisition by the fully hardware route such a time consuming process, sometimes as much as five to 10 minutes. Modern sets with greater processor involvement can do a much faster job.

The next generation of sets still used a hardware code generator and correlator, but allowed the processor to sample directly the resultant output, and pin down the frequency of the carrier with an FFT. This is equivalent to using many simultaneous filters.

Note that in a coherent set, once the exact carrier frequency is pinned down, the rate at which the code must be run is known as all oscillator error and Doppler effects have the same percentage effect. Doppler affects both carrier and code.

Carrier tracking and data extraction

Once the code tracker is locked, its output is a narrow band carrier, ie CW modulated with 50 baud data BPSK. However, in the shared part of the radio, the frequency cannot be brought to zero exactly, as the satellite frequencies differ due to Doppler shift.

Thus the final residual frequency error must be removed in the processing channel. This is done using either a phase locked loop, originally in hardware, but more recently in software, or alternatively a direct measurement of phase using an arctangent calculation on the I and Q components of the signal.

Either method yields two outputs: a very tightly filtered copy of the residual carrier signal and a very oversampled datastream.

The cycles of the carrier are counted, and the phase noted. The cycle count can be used both for velocity calculations and for millimetre accuracy in post processed surveying applications.

The main output is the datastream. Extracting the data requires further processing in a filter matched to the databit length of 20ms, and the data sampled once per bit at the peak. This is achieved by a very simple loop that rectifies the signal and monitors it early, punctually and late. By adjusting the sample point until the early and late levels match, the loop rapidly stabilises at the correct point. While this is an exact mimic of the loop in the code tracker, it runs at just 50 bits per second against over a million for raw code. It requires almost no CPU time to implement in software.

Multiple satellites

The story so far relates how a signal from a single satellite is demodulated. A real GPS receiver must receive from four satellites simultaneously to generate a position fix; the received data must be interpreted to derive the orbit parameters which allow the position of the satellite to be calculated.

The method used in dealing with multiple satellites used to have a serious effect on the cost of the receiver. The design became a cost/performance trade off. With reducing cost of asics, receivers now provide five copies of the processing functions in hardware, with only control and data extraction done in software. But there are many variations available to the designer.

The lowest cost solution provides a single hardware channel performing the satellite tracking which it shares over the four satellites. If this timesharing is done slowly, the system is known as a sequencing receiver. A multiplex receiver downloads data from all satellites using short time slots.

Both these types behave adequately in marine and land environments, where the acceleration environment and dynamics are low. However they are not suitable for missiles or fighter aircraft. The sequencing receiver also has the problem that it must stop sharing while downloading the orbit data. This means that for half a minute or so, positioning is not available. It must also download the data from all four satellites at startup sequentially which takes several minutes.

An effective compromise between single channel and the fully parallel set is to provide two channels. At startup, both may be used for downloads. After acquisition, one is used for positioning while the other becomes the housekeeping channel: it acquires new satellites and downloads from them without interrupting positioning. When all in the sky have been downloaded, the second channel can also be timeshared, thus tracking all in view, so that if one or more are obstructed, there are still sufficient for a fix.

The fully parallel set usually provides five channels using four to track the optimum satellites while the fifth downloads data from the rest. There are some sets that provide eight or twelve channels so that all in view can be tracked continuously, rather than multiplexing on the spare satellites.

GLOSSARY

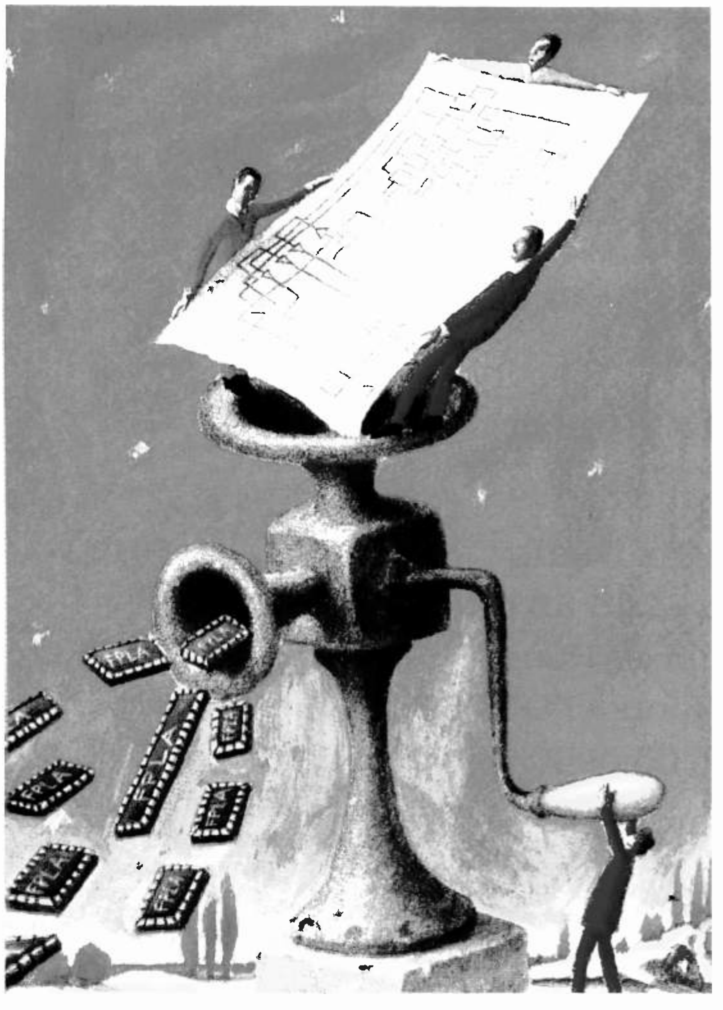
BPSK Binary Phase Shift Keying, inverting the signal for '1' data
I + Q In phase and quadrature components of a signal
LNA Low noise amplifier
PLL Phase locked loop
PTFE Polytetrafluoroethylene, a strong, inert plastic
ppm parts per million, usually a tolerance, here on frequency
Q Quality factor of a tuned circuit
TCXO Temperature compensated crystal oscillator
VCO Voltage controlled oscillator

Software Correlation

My own contribution to the GPS field, developed in 1988 and reported in *Wireless World* in February 1989, is real-time processing of the GPS signals entirely in software. My approach has been not multiple harware channels, but feeding the signals through an A-D converter directly into a computer for processing in software.

The method has two major benefits. It can be implemented without resorting to special asics and, perhaps more importantly, software may be remodelled as the market changes without added chip development costs.

The software can emulate a single channel or parallel set, using code tracking, squaring or composite codes, and mixing these as appropriate through the phases of handling the satellite. The operation of the receiver in software will be covered later.



DIY LSI

3:tools of the trade

In the third and final part of their series on programmable logic families, Dave Nicklin and Nick Sawyer* examine the role played by software in designing programmable devices. In particular, it considers the shift toward VHDL and logic synthesis together with the concept of high level graphical design.*

Most FPGA designs are currently completed using normal gate and macro level schematic entry tools. The FPGA vendor supplies a library of macros and generic symbols which are connected together using either proprietary or standard schematic entry tools. The netlist produced by this software is then translated into a vendor specific format for partitioning, placement and routing within the target device.

While schematic entry is the normal route to FPGA design, the majority of EPLD designers employ text based forms of entry. These tools vary in complexity, and there are a number of products and standards currently available which allow the engineer to design using complex Boolean equation and state-machine description syntax. This form of entry naturally allies itself with the types of design that fit most suitably into an EPLD style architecture. That is not to say that schematic entry isn't used for EPLD design, nor equation entry for FPGAs. They are both used to some degree. However, both have adopted the design practices of their roots: EPLDs use the entry tools familiar to PAL designers, while FPGAs follow asic design practices.

Deadline timing

In contrast to EPLD design, it is hard to predict the performance of an FPGA circuit before it has been placed and routed. The FPGA architecture trades predictable timing against flexibility and efficient silicon usage. However, as a result of automatic placement and routing software development, it becomes possible to incorporate some predictability in FPGA design.

Deadline timing is the ability to control the partitioning and routing of a design in terms of critical paths and overall circuit performance. Xilinx calls its version of the software *XACT-Performance*. With this system a designer can assign (in his schematic) various timing attributes, termed TSPEC attributes. Each attribute can define a path in terms that the designer is familiar with, such as clock-to-setup, pad-to-pad, clock-to-pad and pad-to-setup. Placing the attributes on the design schematic eases update of timing parameters.

While the use of deadline timing does not guarantee the design will work at the specified rate, the software does detect impossible timing constraints, i.e. those due to block delays which will cause the path timing to fall outside specification. It allows the designer to ensure that logic paths are prioritised by the software in terms of true system requirements. Even so, EPLD implementation will always remain the only way to achieve 100% predictable timing.

Towards logic synthesis

The last couple of years has seen efforts to use hardware description languages, such as

Continued over page

*Xilinx UK

DEADLINE TIMING - AN EXAMPLE

Figure 1 shows an example of an LCA design with *XACT-Performance* specifications. Figure 2 is the same schematic highlighted to illustrate the effects of deadline timing. The XACT-Performance specifications in this design result in the following (numbered items in this list refer to the appropriate numbers in the schematic).

Terminology

DP2P = Default Pad-to-Pad specification, i.e. the default delay for any path in a design sourced by an input pin leading to an output pin.

DP2S = Default Pad-to-Setup specification, i.e. the default delay for any path in a design which starts at an input pin and finishes at the input of a flip-flop.

DC2S = Default Clock-to-Setup specification, i.e. the default delay for any path sourced from a flip-flop output which finishes at the input of another flip-flop.

DC2P = Default Clock-to-Pad specification, i.e. the default delay for any path sourced from a flip-flop which finishes at an output pin.

P2P = Specific Pad-to-Pad timing constraint. i.e. a specific path delay value which overrides the default pad-to-pad timing. This delay can be either shorter or longer than the default value.

P2S = Specific Pad-to-Setup timing constraint. i.e. a specific path delay value which overrides the default pad-to-setup timing. The value may be either shorter or longer than the default value.

C2S = Specific Clock-to-Setup timing constraint. i.e. a specific path delay which overrides the default clock-to-setup value. The value may be either shorter or longer than the default value.

C2P = Specific Clock-to-Pad timing constraint. i.e. a specific path delay value which overrides the default clock-to-pad value. The value may be either shorter or longer than the default value.

TSnn = Time Spec attribute (*nn* can be any integer value). An attribute can be attached to a net which is part of the path to be constrained, thereby assigning the timing information to that specific path.

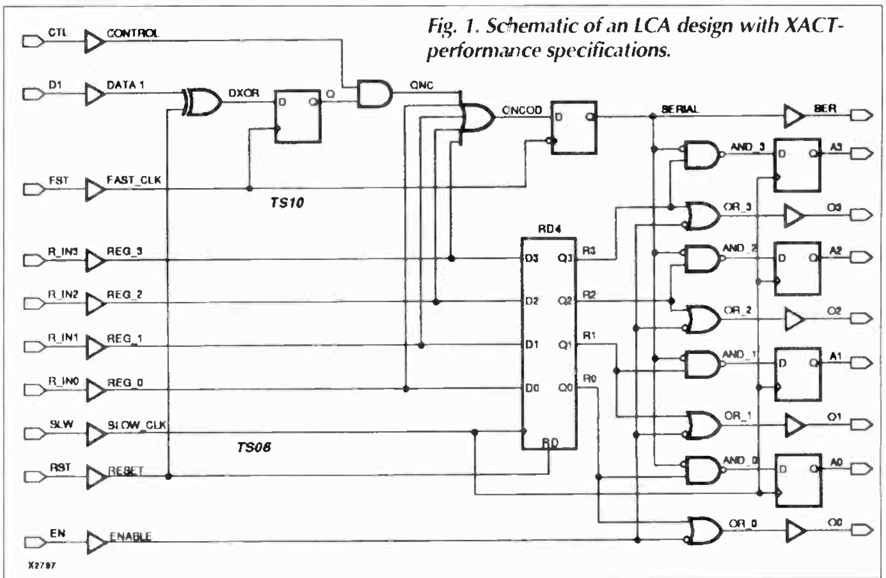


Fig. 1. Schematic of an LCA design with XACT-performance specifications.

Timespec	TS06=LINK:TS07:TS0
TS01=DC2S:40:25	8:TS09
TS02=DP2S:30	TS07:C2P:100
TSO3=DC2P:25	TS08=C2S:200
TS04=DP2P:100	TS09=P2S:110
TS05=P2P:60:*:Q0	TS10=P2S:45:CTL

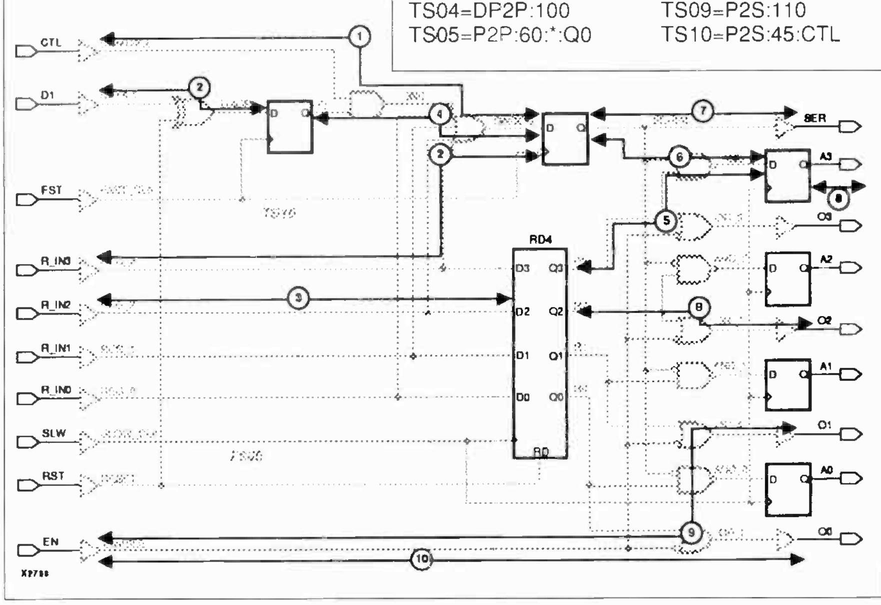


Fig. 2. Annotated version of Fig. 1. schematic.

Referring to the paths of Fig. 2:

(**1**). DP2S of TS02 is overridden by P2S of TS10, resulting in a pad-to-setup specification on this path of 45ns, compared to the 30ns default.

(**2**). DP2S of TS02 is not overridden by P2S of TS10, as TS10 specifically refers to the input CTL in the timespec table. Therefore the pad-to-setup for this path is the default 30ns.

(**3**). DP2S of TS02 is overridden by P2S of TS09 for all flip-flops clocked by the net SLOW_CLK, resulting in a pad-to-setup of 110ns.

(**4**). DC2S of TS01 specifies a high time of 25ns. As the timing for this path is from the rising-to-falling clock edge, it takes this specification as the clock-to-setup. If no high time had been specified then the software would calculate the parameter based on a 50-50 duty cycle clock.

(**5**). DC2S of TS01 is overridden by C2S of TS08 for all flip-flops clocked by the net SLOW_CLK, resulting a clock-to-setup specification of 200ns.

(**6**). This path is a clock-to-setup type. However, the source and destination flip-flops are already governed by different timing specifications. The source flip-flop is governed by the default clock-to-setup of 40ns given by TS01, whereas the destination flip-flop has a much looser specification, of 200ns given by TS08. In this case the tightest specification always takes precedence, and this results in a clock-to-setup of 40ns.

(**7**). DC2P of TS03 is 25ns, therefore the clock-to-pad of this path is 25ns.

(**8**). DC2P of TS03 is overridden by C2P of TS07, resulting in a clock-to-pad specification of 100ns.

(**9**). DP2P of TS04 results in a pad-to-pad specification of 100ns.

(**10**). DP2P of TS04 is overridden by P2P of TS05 for the path to O0. This results in a pad-to-pad specification of 60ns.

VHDL, and logic synthesis tools more effectively for both EPLD and FPGA design. This has led to design entry developments which include high-level schematic based synthesis tools.

As FPGAs increase in size and performance, designs increase in size and complexity. By using a synthesis tool rather than designing at the gate level, the designer enjoys a higher level of abstraction. This enhances creativity and should provide faster design cycles. Being largely architecture and manufacturer independent, it avoids becoming too tied to proprietary library or design tools.

The growth of VHDL and logic synthesis is analogous to the development of high-level language compilers for software development. The writing of assembly language programs and early generations of interpreted software languages were time consuming. Although the end result was often a highly efficient technical solution, it took little account of development time.

Today the use of assembly code has been

SCHEMATICS AND SYNTHESIS

To illustrate the advantages which high level synthesis can bring to the designer, we will compare a simple circuit implemented in both a classical schematic method, and using the new high level libraries and *x-blox* software.

The design (**Fig. 3**) is for a pulse generator which produces a variable frequency with programmable mark/space ratio. The circuit is simply two loadable counters that trigger each other when they reach terminal count. The TC outputs also control a toggle flip-flop that then produces the desired output waveform. The concept of the design is extensible, more counter modules would be added so that more complex waveforms could be generated.

In this first figure we can see that the data to be loaded into the counters (DA or DB) is latched into the registers contained in the macro block INREG8 whose schematic is shown in **Fig. 4**. It is important to note that we had to decide the length of each counter before we started on the design; four bits were chosen for clarity. We had to decide, because we needed to draw an i/o pad for each data bit to be input.

Having latched in the data, it is then presented to the count macros C16BCRD. These are loadable up-counter macros with clock enable, reset direct and parallel enable, and are part of the standard Xilinx library for the *XC4000* devices. The schematic of the macro is shown in **Fig. 5**. The length of count of the top counter (COUNT1) determines the low time (TL) of the output waveform, and the bottom counter (COUNT2) determines the high time (TH).

Consider now the circuit shown in **Fig. 6**, which performs an identical function, but has been drawn using the parameterisable *x-blox* modules. Each of the circuit inputs is drawn using the generic INPUTS symbol which can represent either a single input or an *n*-bit bus. The width of the input bus (where applicable) is specified by means of the attached BOUNDS attribute. Thus the two four bit buses are given by the BOUNDS=3:0 expression. The other nets are simply bit inputs, so no BOUNDS are required. The inputs are fed to a generic DATA_REG symbol. This represents *n* registers, and the symbol has inputs for all possible register scenarios, i.e. clock, clock enable, async reset,

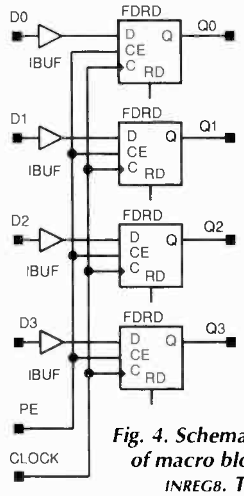


Fig. 4. Schematic of macro block INREG8. This contain the registers into which counter data is latched.

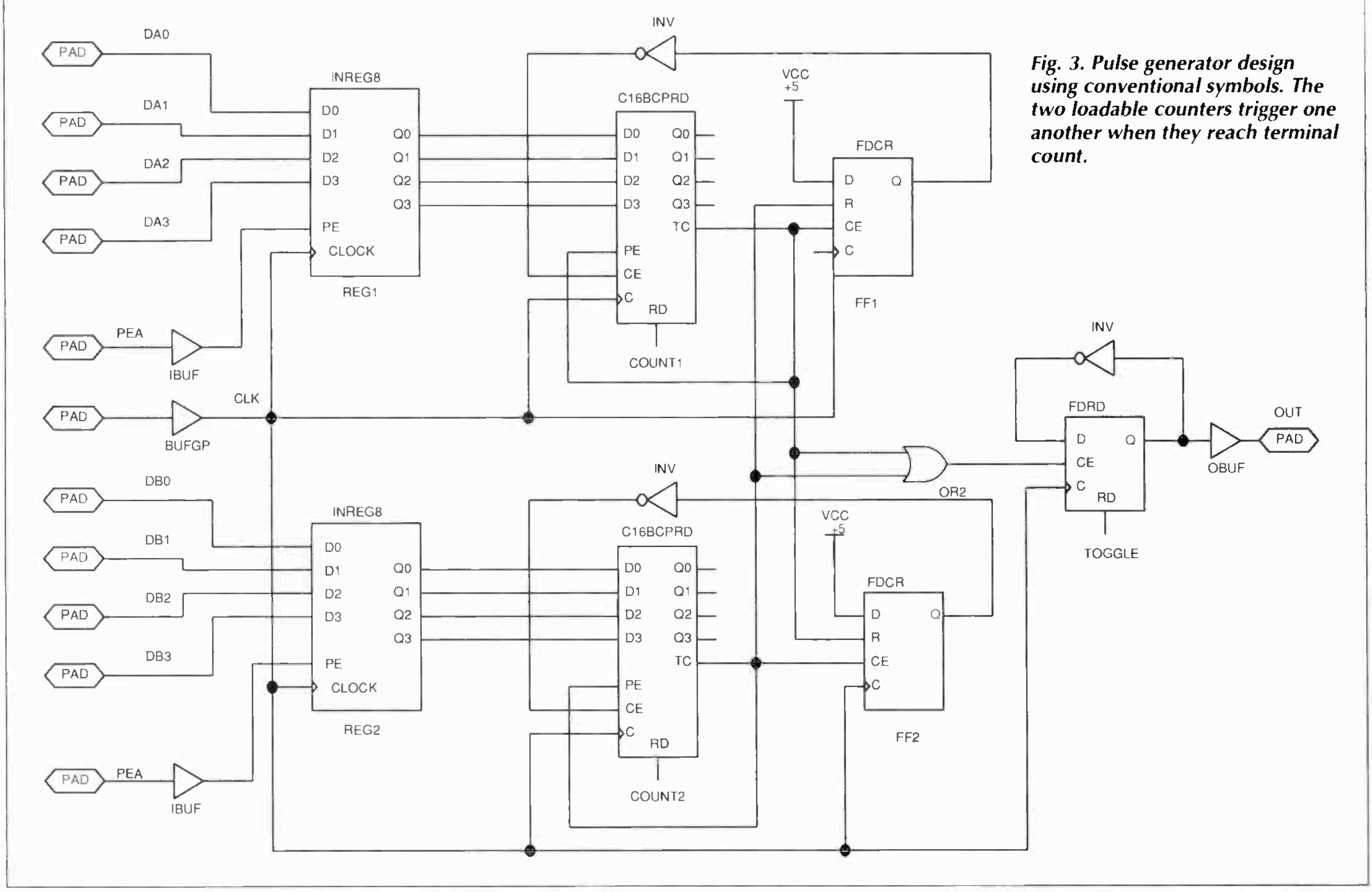


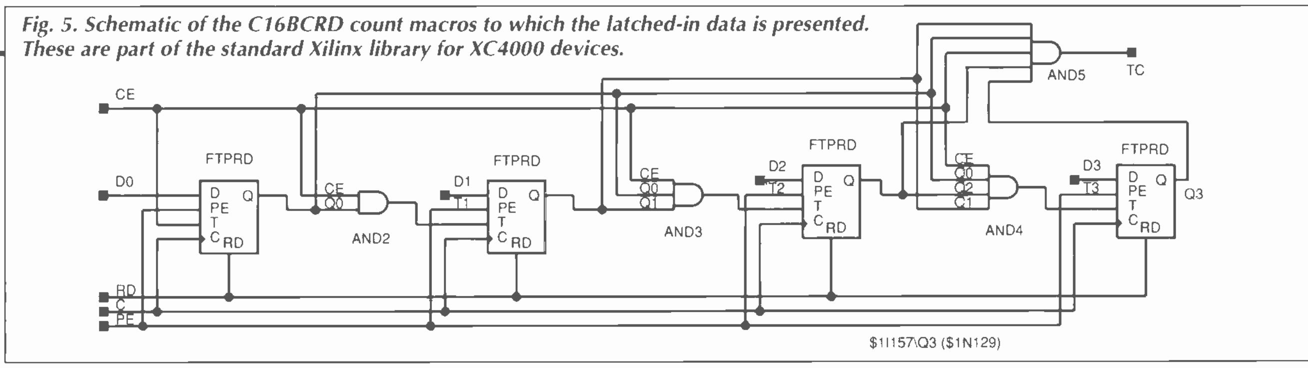
Fig. 3. Pulse generator design using conventional symbols. The two loadable counters trigger one another when they reach terminal count.

reduced to those parts of a program, such as interrupt handlers, which must be as fast as possible. In fact as compiler technology has progressed, the code generated by compilers has very nearly approached the efficiency of hand coding.

Why are synthesis tools not used more widely? An obvious reason is cost. The software is not cheap and it requires a long learning-curve. It is also intended primarily for the gate-array market and that granular architecture. As a consequence it has always been less efficient at synthesising logic for devices with complex logic block structures such as EPLDs and FPGAs.

A typical asic vendor library consists of anything from a few dozen to a few hundred gate level elements. The synthesis tool will map the most efficient available element for a given function. The Xilinx FPGA architecture consists of look-up tables rather than gates, and

Fig. 5. Schematic of the C16BCRD count macros to which the latched-in data is presented. These are part of the standard Xilinx library for XC4000 devices.



etc. Notice that we do not specify a size (in bits wide) for the symbol. The register outputs feed the generic COUNTER module, which again has no size information. Once more the symbol contains all the inputs and outputs that could potentially be needed for a counter (though obviously not all have to be used). The type of counter, e.g. binary, Johnson, can be specified via attribute, but the optimisation software should make this decision for the designer.

The *x-blox* based circuit therefore is very similar to the classic implementation, but has been drawn in the same manner as a designer thinks, i.e. an input then a register, then a counter, rather than having to worry about the exact sizes of registers and counters.

The first task the software undertakes is to size the design. It sees that there are four inputs to the data registers, therefore it will synthesise four registers. As there are four registers the counter must be a four-bit counter, and that is therefore synthesised.

Having synthesised the logic, *x-blox* will then optimise this logic for the requested Xilinx device. For example global buffers, and hard macros will be created. The output of the program is a netlist in standard format. It also generates a report shown here for interest as it demonstrates each phase of the process.

Continued over page

```
  X-BLOX(tm) - Blocks of Logic Optimized for Xilinx (Version: 1.04)
   supports the Xilinx XC4000 Family of Field Programmable Gate
Arrays
  (c) 1990, 1991, 1992 by Xilinx, Inc. All Rights Reserved. U.S. and
  Foreign Patents Pending.

  READING 'art3_1.xnf' as the design file ...
  CREATING THE X-BLOX internal database ...

  DEDUCING THE BUS WIDTHS ...
  => Signal DA given Data Type unsigned binary indexed from 3 to 0
(Bus of
  => BIT indexed from 3 to 0).
  => Signal QA given Data Type unsigned binary indexed from 3 to 0
(Bus of
  => BIT indexed from 3 to 0).
  => Signal FBQ2 given Data Type BIT.
  => Signal $1N34 given Data Type BIT.
  -> Signal DB given Data Type unsigned binary indexed from 3 to 0
(Bus of
  => BIT indexed from 3 to 0).
  => Signal QB given Data Type unsigned binary indexed from 3 to 0
(Bus of
  => BIT indexed from 3 to 0).
  => Signal FBQ1 given Data Type BIT.
  => Signal $1N23 given Data Type BIT.
  => Signal QOUT given Data Type BIT.
  => Signal $1N51 given Data Type BIT.
  => Signal OUT given Data Type BIT.
  => Signal TC2 given Data Type BIT.
  => Signal TC1 given Data Type BIT.
  => Signal TG given Data Type BIT.
  => Signal CLK given Data Type BIT.
  => Signal FF1/A1 given Data Type BIT.
  => Signal FF1/A0 given Data Type BIT.
  => Signal FF1/DI given Data Type BIT.=> Signal FF2/A1 given Data
Type BIT.
  => Signal FF2/A0 given Data Type BIT.
  => Signal FF2/DI given Data Type BIT.
  => Signal PEA given Data Type BIT.
  => Signal PEB given Data Type BIT.

  IMPROVING YOUR DESIGN by using the special features of the Xilinx
device
  ==============================================
  CHECKING IF GLOBAL BUFFERS can be used ...
  * INFO: Signal 'CLK' now connected to a 'GLOBAL PRIMARY BUFFER
(BUFGP)'
  * INFO: symbol.

  EXPANDING THE X-BLOX modules into optimised logic functions
  ==============================================
  CHECKING IF FAST CARRY LOGIC can be used ...
  * INFO: The symbol called 'COUNT2' has been mapped into an
arithmetic hard
```

```
  * INFO: macro AND HAS BEEN RENAMED COUNT2_1 to utilise the fast-
carry
  * INFO: logic.
  * INFO: The symbol called 'COUNT1' has been mapped into an
arithmetic hard
  * INFO: macro AND HAS BEEN RENAMED COUNT1_1 to utilise the fast-
carry
  * INFO: logic.
  => The COUNTER 'COUNT1_1' has been expanded into a hard macro.The
hard
  => macro has been written to the file xblox5.hm
  => The COUNTER 'COUNT2_1' has been expanded into a hard macro.The
hard
  => macro has been written to the file xblox6.hm
  => Register 'REG2' driven by net 'QB' expanded into 4 elements,
which
  => includes flip-flops and logic symbols ...
  => Register 'REG1' driven by net 'QA' expanded into 4 elements,
which
  => includes flip-flops and logic symbols ...
  => INPUTS '$1I1' driving net 'DA' expanded into 4 groups of IPAD
and IBUF
  => symbols ...
  => INPUTS '$1I33' driving net 'PEB' expanded into 1 group of IPAD
and IBUF
  => symbols ...
  => INPUTS '$1I41' driving net 'DB' expanded into 4 groups of IPAD
and IBUF
  => symbols ...
  => INPUTS '$1I70' driving net 'PEA' expanded into 1 group of IPAD
and IBUF
  => symbols ...
  => INPUTS '$1I71' driving net 'CLK' expanded into 1 group of IPAD
and IBUF
  => symbols ...
  ==============================================

  MODULE EXPANSION IS COMPLETE, merging modules into a single design.
This
  process can take a few minutes on a large design ...
  ==============================================

  WRITING THE IMPROVED AND EXPANDED DESIGN to a Xilinx Netlist Format
file
  ==============================================
  * INFO: WRITING NEW X-BLOX Netlist File to 'art3_1.xg' ...

  X-BLOX EXPANSION AND OPTIMISATION IS COMPLETE
  ==============================================
  Your expanded and optimised design can now be Partitioned, Placed
and
  routed using the Xilinx PPR program. See the file
'd:/nick 4k/art3_1.blx'
  for a log of X-BLOX outputs.
```

this makes these potential libraries much too large. As an example, a four input function generator has the potential of generating 65,536 different logic functions. This number can be minimised using input pin-swapping, and 'inverter absorption' to a library of only 223 separate elements. However, a five input function generator can be configured to produce any one of 4,294,967,296 logic functions and can be minimised down to a figure of roughly one million separate functions. This library is obviously impractical and therefore, a subset of the most commonly used functions is used instead. This means that functions that could be performed in one function generator and are not in this library, need to be constructed out of the available library elements by the synthesis tool. As a consequence, a function which could be produced in only one function generator, may have to be implemented in two or three function generators by the synthesis software. Very inefficient.

As the number of FPGA and EPLD designs increases, so the demand for synthesis support rises. Some manufacturers are looking to synthesise directly from look-up table function generators, thereby greatly increasing the efficiency of the logic implementation. This will require comprehensive software development to meet FPGA/EPLD market requirements.

Graphical design and synthesis

Until recently, virtually all high-level logic design was text based, with VHDL and Verilog-HDL being the two main standards. However, there have recently been attempts to produce graphical synthesis products. These are still very new, and will be formalised under a new standard called the "Library of Parameterised Modules" (LPM).

The first implementation of graphically based high-level design and synthesis for FPGAs, was *x-blox* from Xilinx. The product partitions into two separate entities. The front end of the package comprises a graphical library, which may be used with the schematic tools familiar to the designer. These modules are non-specific drawing blocks, i.e. adders, counters, multiplexers etc., but none of these specify gates as with older libraries. For example, the same counter module may be used to specify a 32-bit binary counter or a 128-bit Johnson Counter. Size, and design style, is defined on the schematic by giving the module specific attributes. Very large and complex circuits can be designed by connecting these modules together, as well as by combining them with regular schematic macros to perform small and compact functions – analogous to software development where assembly code routines are compiled into *C*.

Once completed the schematic is processed to a high level Xilinx Netlist Format (XNF). This is a text description of the schematic design and is read by the second part of the package, the *x-blox* program. This software generates the optimum design for each module, depending on the size, style and target device. As the program was written in the artificial intelligence language, prolog, it uses rule based decision trees and models of the Xilinx FPGA look-up table structure to produce the design. It can also use the system features on each type of device.

For example, in the *XC4000* family it can implement logic to make use of the dedicated carry-logic circuitry, and also the wide decoder logic. It can also ensure the best performance and device usage for the design by choosing the correct global low-skew resources for clocks and high fan-out nets. It will even move flip-flops from the internal logic blocks into the i/o blocks wherever possible to improve system speed and design efficiency. Overall, the intention is to reduce development time for complex circuits.

As the modules include programmable parameters it is relatively easy to scale a design. For example, if a chip were originally programmed for an 8-bit wide address bus, it only requires one or two amendments to the attributes of a module to create a 16-bit wide address bus.

The future

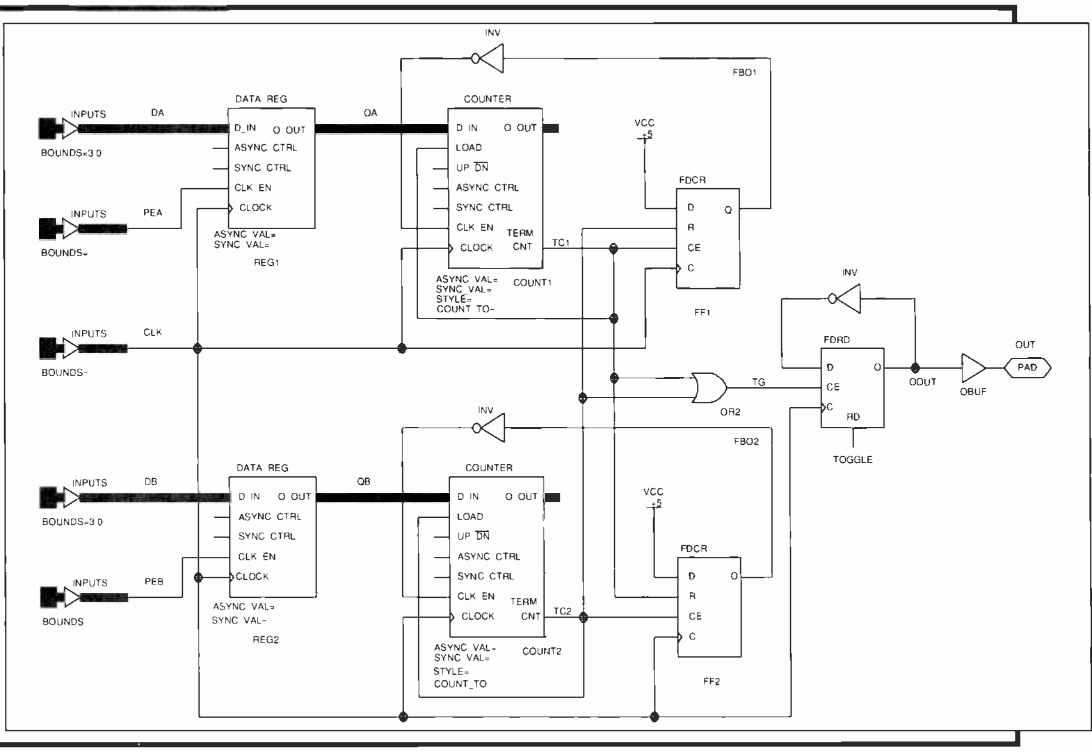
While the market value of asics is still larger than the FPGA and EPLD markets, the latter are growing at a much greater rate than that of asics. In fact, many asic manufacturers have pulled out of the 1k-10k gate area, because of customer preference for programmable logic. Many of the larger semiconductor companies view the complex programmable logic markets with an eye to developing their own products. But they face many barriers to entry, not least of which are the formidable array of patents taken out by the American pioneers.

Both the FPGA and EPLD markets are arguably the most dynamic semiconductor markets around today.

Hardware development will be directed to speed, density, cost reduction and system features. For example, maximum clock rates have increased from 33MHz to 230MHz in just six years, while device density has increased from around 1000 to 10,000 usable gates. This has been followed by consistent price reductions and increasing numbers of system enhancing features, such as boundary scan test logic.

It will not be long before FPGAs provide up to 20,000 usable gates combined with 100MHz system level performance. This will cause second thoughts over vast majority of new asic design starts. It remains to be seen what the consequences will be. ■

Entering the design in this way has the major advantage of being able to change it quickly. Consider the case where the designer realises that he actually should have used a higher definition, for example the requirement is now for five bit counters. In the classic schematic, he will have to edit the top-level drawing to add more input pads, edit the register macro to add another register, and change the clock macro to a five-bit counter. This will also mean adding a lot more nets (and naming them). With the *x-blox* approach, all that is required is to change the BOUNDS attributes to indicate a 5-bit data path (i.e. BOUNDS=4:0), and then re-process. The software runs in about two minutes on a 486 PC for this design.



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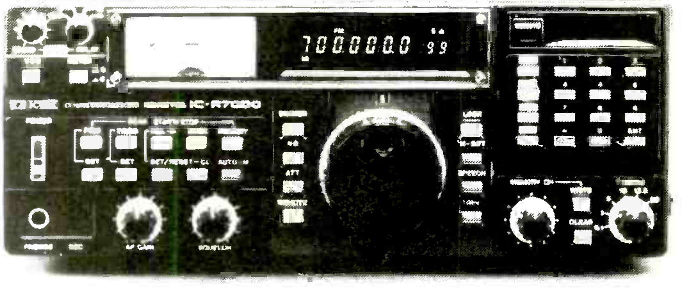

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THE 180 SERIES

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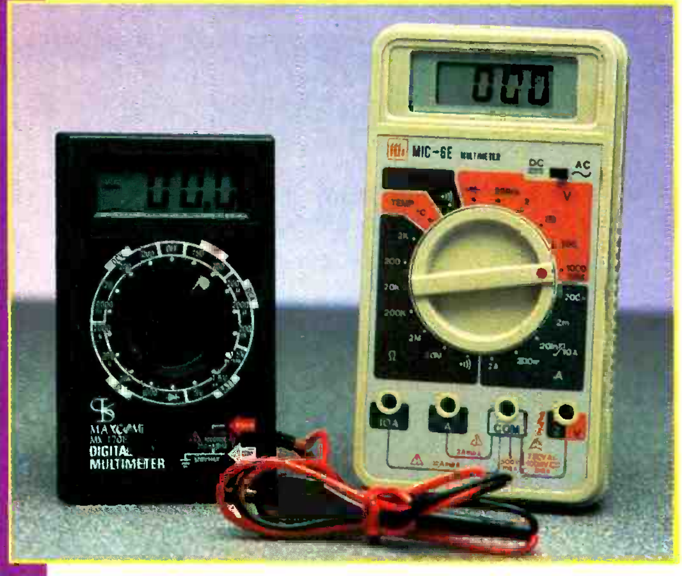
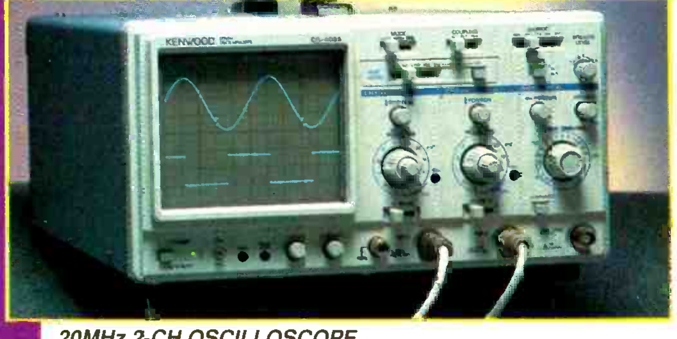


MX170B MIC-6E

20MHz 2-CH OSCILLOSCOPE

The CS4025 20MHz dual trace oscilloscope offers a comprehensive range of facilities including a high sensitivity vertical amplifier providing from 1mV to 5V/div in CH1, ALT, CHOP, ADD, CH2 modes with inverse polarity on CH2. The horizontal timebase offers a sweep range of 0.5s/div to 0.5µs/div plus x10 sweep expansion and X-Y mode. Triggering can be auto or normal from vert, CH1, CH2, line or external sources with coupling provided for AC, TV-F and TV-L. The CS4025 is supplied complete with matching probes for £295.00 plus VAT (£346.62).



20MHz 2-CH OSCILLOSCOPE

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The PPS series of GPIB programmable DC power supplies offer high performance yet are extremely competitively priced using a 16 x 2 backlit LCD and 14 button keypad. All functions and conditions are easily selected and displayed. Overvoltage and overcurrent are selectable as is output enable/disable. Terminals for output and sense are provided on the front and rear to allow easy rack mounting.
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FUNCTION GENERATOR

The **MX2020** 0.02Hz – 2MHz sweep function generator with LED digital display offers a broad range of features. Output waveforms include sine, square, triangle, skewed sine, pulse and TTL. Lin and log sweeps are standard as is symmetry, DC offset and switchable output impedance from 50Ω to 600Ω. The digital display provides readout of the generators' frequency or can operate as separate 10MHz frequency counter. £199.00 plus VAT (£233.83).

LCR METER

The **MIC-4070D** LCD digital LCR meter provides capacitance, inductance, resistance and dissipation measurement. Capacitance ranges are from 0.1pF to 20,000μF plus dissipation. Inductance ranges from 0.1μH to 200H plus a digital readout of dissipation. Resistance ranges from 1mΩ to 20MΩ. Housed in a rugged ABS case with integral stand it is supplied complete with battery and probes at £85.00 plus VAT (£99.88)

FOUR INSTRUMENTS IN ONE

The **MX9000** combines four instruments to suit a broad range of applications in both education and industrial markets including development work stations where space is at a premium.

The instruments include:

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2. An 8-digit LED display 1Hz - 100MHz frequency counter with gating rates of 0.1Hz, 1Hz, 10Hz and 100Hz providing resolution to 0.1Hz plus attenuation inputs and data hold;
3. A 0.02Hz to 2MHz full featured sweep/function generator producing sine, square, triangle, skewed sine, pulse and a TTL output and linear or logarithmic sweep. Outputs of 50Ω and 600Ω impedance are standard features;
4. An auto/manual $3^1/_2$ digit LCD multimeter reading DCV, DCA, ACV, ACA, resistance, and relative measurement with data hold functions.

The MX9000 represents exceptionally good value at only £399.00 plus VAT (£468.83).

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The **FG500 series** sweep/function generators provide two powerful instruments in one package, a 6MHz or 13MHz sweep/function generator and an intelligent 100MHz frequency counter. The microprocessor based instruments offer sophisticated facilities yet remain extremely competitively priced. A menu driven display allows easy set up and operation. A 16 character by 2-line LCD display provides clear and unambiguous readout of generator output and frequency measurement.

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EWWO/93

HOW CLEAN is your audio op-amp?

Most op amps – particularly when configured to employ high feedback ratios – produce small to vanishingly small amounts of harmonic distortion at mid frequencies in the audio band. Manufacturers often claim reductions in %THD in modern devices. But makers are not audio engineers so data provided is sparse, incompatible, and measurement technique is rarely documented.

Collective distortion data covering common and esoteric op-amps was originally published by Walt Jung[1] 15 years ago (which I recently revisited[2]).

The golden ears brigade has always claimed that some distortion artefacts remain audible at levels approaching those of system thermal noise. **Ben Duncan's** *objective op-amp distortion test programme quantifies harmonics down to 140dB below the signal.*

But a great deal of op-amp "distortion" – whether harmonic or intermod – is really noise, and while this overrides the test equipment's reading, it is not necessarily masked to the ear.

In the past decade, the capability of test sets like Audio Precision's *System One* to plot %THD + N against frequency and level has aided designers' understanding of how to analyse mechanisms, to reduce or tweak THD residues. But they have not directly indicated how to make a circuit sound acceptable to critical listeners.

In fact, many listeners hold a fear of op-amps, knowing that most rely nearly completely on high (almost brute force) feedback for linearity.

As high order products are hard to eradicate, they perhaps have a point. The ear is acutely – and increasingly – sensitive to the higher harmonics, particularly odd orders, produced by high NFB ratios. To use an old analogy, it's like beating a large dent out of a car's wing: feedback replaces the dent with many smaller ones. No amount of beating will completely eradicate the damage – if you look closely enough.

An op-amp's harmonic structure can help steer audio quality, but conventional wave analysers are beset with the same noise masking problem as %THD measurements.

Improved testing

Following earlier testing (see box) the Audio Precision *System One* test set's DSP facility has been employed to record harmonics below the noise floor of conventional wave

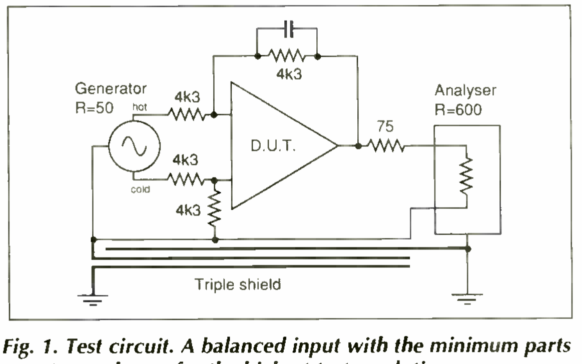
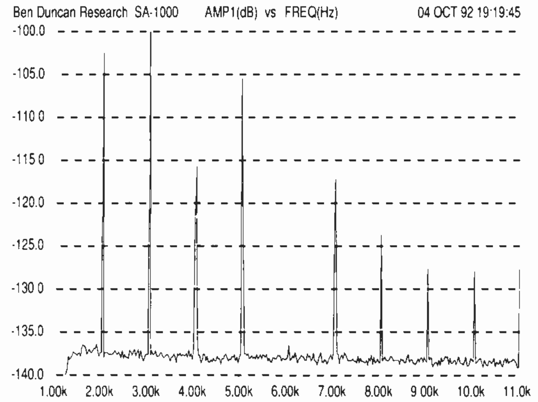
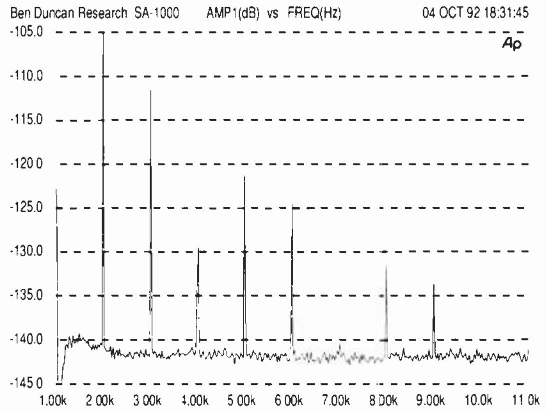


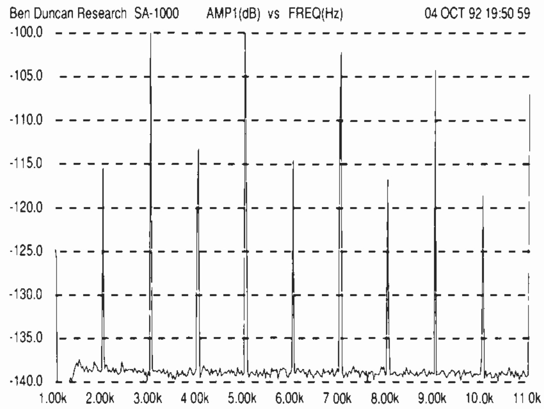
Fig. 1. Test circuit. A balanced input with the minimum parts count was chosen for the highest test resolution.



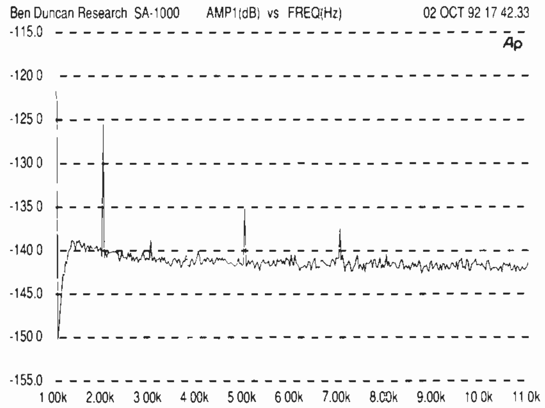
Texas TL071



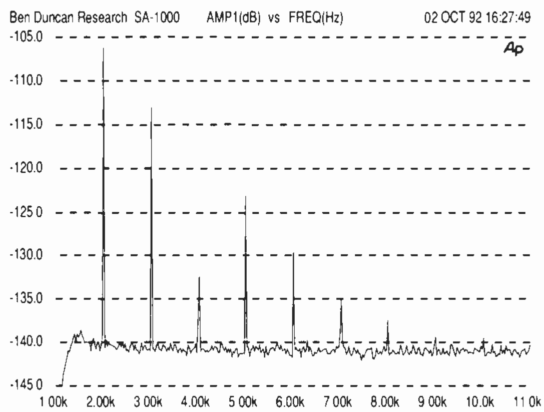
PMI (AD) OP42



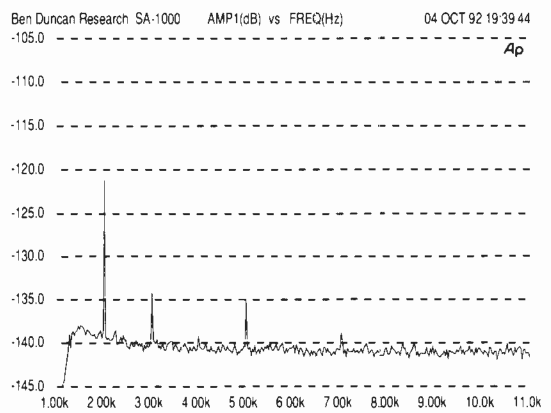
Motorola L351



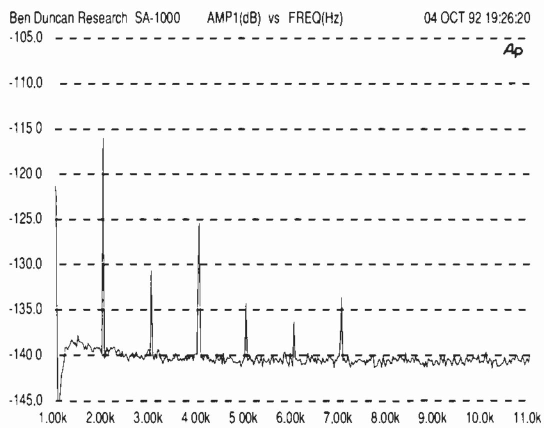
PMI (AD) AD845



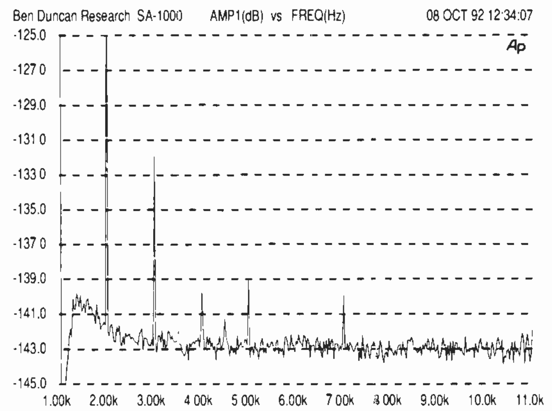
Linear Tech LT1056



Linear Tech LT1122

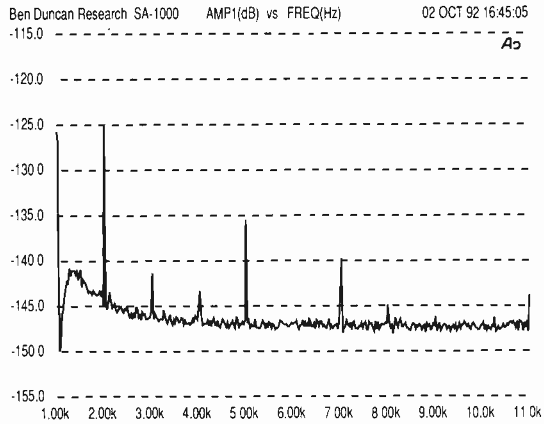


Analog Devices AD744

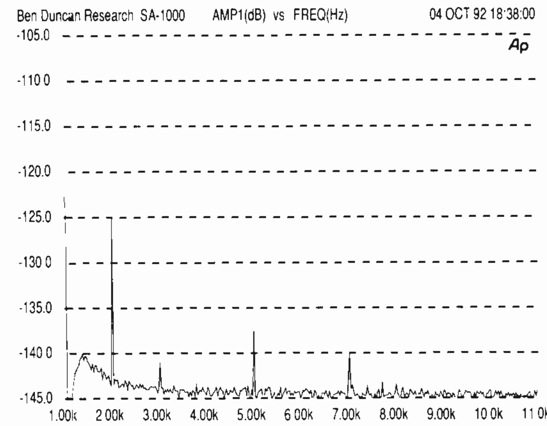


Burr Brown OPA604

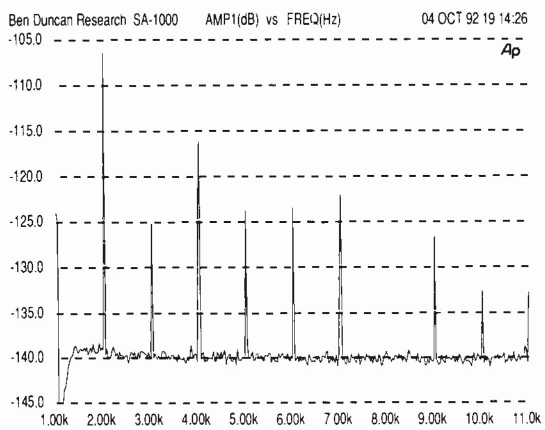
Figs. 2-16: Graph scaling maintains a 20dB window, but the base-line has been adjusted according to each op-amp's noise floor. Note how the test also yields smooth spectral data on the op-amp's noise. Amplitudes of harmonics with off-scale peaks are enumerated at the graph tops.



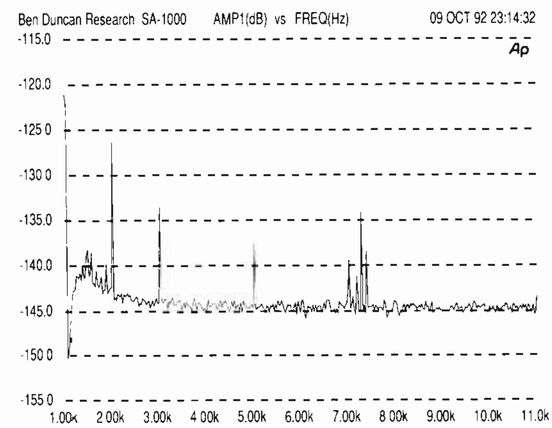
Signetics NE5534



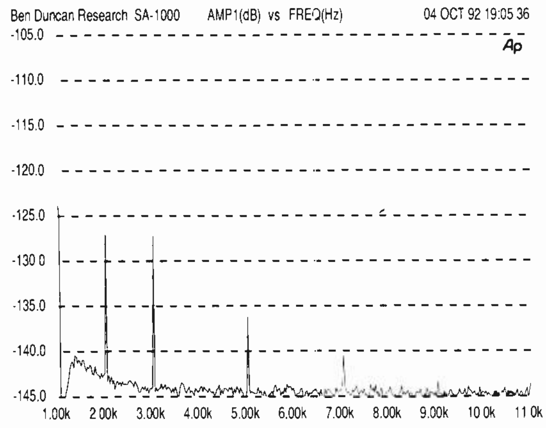
Harris HA5221



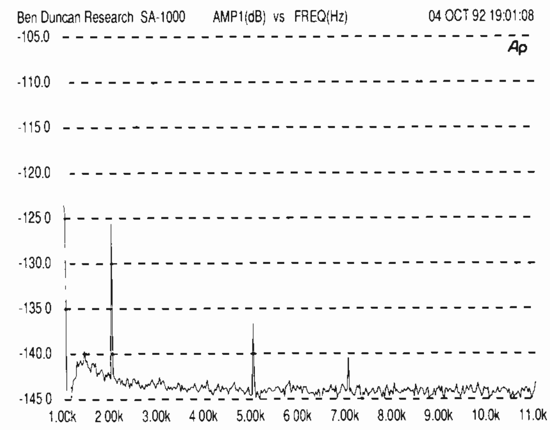
Harris HA2525



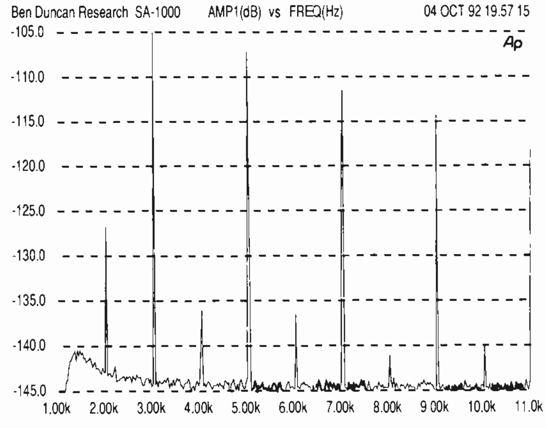
Texas TLE2027



PMI (AD) OP27



Analog Devices AD797



Linear Tech LT1007

Strong signatures
Devices listed in ascending order of magnitude, ie. most polluted first, are:
Op-amps with highest sum of even harmonics (2, 4, 6th): *LF351, TL071, HA2525, LT1056, OP42.*
Op-amps with highest sum of odd harmonics (3, 5, 7th): *LF351, TL071, LT1007, OP42, HA2525.*

Harmonic Ranking

Numeric ranking of spectra in figs 2-16, dBs referred to Audio Precision test set residue. Grouped by ascending even, then odd harmonics. The lowest, cleanest op-amps appear at bottom of table. The most euphonic sounding will emphasise low and even order over high and odd order. But see text for caveats regarding interpretation.

Even					Odd				
2nd	**4th**	**6th**	**8th**	**10th**	**3rd**	**5th**	**7th**	**9th**	**11th**
>30dB above system residue:									
					351	351	351	351	351
					071	071		1007	
					1007 1007				
+21 to 30dB:									
OP42	351	351	351	351	OP42		1007	2525	1007
071	071	2525	071				071	2525	OP42
	744								
+11 to 20dB:									
2525	1056	1056	OP42	071 1056	OP42	2525	071	071	
1056				2525	-	1056		OP42 2525	
604					-	2525			
+6 to 10dB:									
351	1007	1007	1056	1056	797	1122		1056	
744	1122	071	1007	1007	2525	744			
		744			5534	845			
					5221	OP27			
					OP27	5534			
< +6dB – not ranked:									
1007	-	-	2525	-		-	1056	-	1056
-	-	-	744	744	744	-	744	744	744
-	OP42	-	-	OP42	-	-	OP42	-	OP42
845	845	845	845	845	845	-	845	845	845
1122		1122	1122	1122	1122	-	1122	1122	1122
797	797	797	797	797	-	797	797	797	797
-	604	604	604	604	604	604	604	604	604
5534	5534	5534	5534	5534	-	-	5534	5534	5534
OP27	OP27	OP27	OP27	OP27	-	-	-	OP27	OP27
5221	5221	5221	5221	5221	-	5221	5221	5221	5221
2027	2027	2027	2027	2027	2027	2027	2027	2027	2027

analysers. Clarity and repeatability of testing has been greatly improved by tightening the resolution through taking the average of 64 samples – though time needed for testing has risen to about 3h for 14 devices. But the noise floor is lowered and a clearer plot is produced, less "spiked" by noise. A 48kHz sample rate allows the AP's ultra-narrow 3Hz measurement bandwidth to be applied.

Repeatability is evaluated by making random retests of spectra, after several hours &/or days have elapsed, sometimes after repowering all the equipment. A peak uncertainty of ±5dB appears on the third harmonic only, with a peak divergence of ±2dB on the remaining harmonics. Variables may include the test set's relay contact resistance, $1/f$ noise components, and the period allowed for op-amp warm-up.

To make any harmonics align perfectly with gridlines, a 1kHz "prodding" tone, subject to a tolerance of a few Hz, is used. Non-integer frequencies place harmonics between grid lines – a situation the AP's presently limited cursor facility finds difficult to handle. Instead, grid lines are locally erased, taking care to reduce, below the test set's residue, pick up of AC mains harmonics in the test band.

Resolving the cleaner devices demands the highest resolution, and the test set has to drive the device under test in balanced mode at 1V RMS. This requirement (**Fig. 1**) subjects the device to a common mode voltage. So distortion arising is included, and usefully, as common mode distortion is prevalent and the dominant distortion mechanism in many bipolar input types.

The device under test's in/out gain must be unity, as noise gain is the last requirement. Fortunately – and despite working with maximum loop gain – op-amps are no more subjectively-perfect at unity gain than at any other setting, though their ranking may change.

Surrounding resistor values were originally set to load the output stage lightly; 4k3 is about the lower working limit for "weedy" devices such as the *TL071*. But at the 1V test level, a 600Ω load helps most harmonics of the cleaner devices stand clear of the test set's residue. At 1V RMS, 600Ω draws under 3mA peak current which is easily within the current sourcing capabilities of all the op-amps tested. The 600Ω simultaneously simulates a noise-free gain condition in the low tens of dB, by causing a droop in open loop gain. Amount of loop gain "loading" depends on output stage gain, and will be higher for op-amps not rated to drive 600Ω. The 75Ω output resistor causes a small yet constant attenuation that is insignificant on a scale of 145dB.

Test results

Figures 2 to **16** show large differences in harmonic amplitudes and, more importantly, structures. But this is only the starting point. Any designer seeking corroboration with subjective results faces a exhausting obstacle course:

- Some op-amps' noise floors may be hiding harmonics. This is significant with many of the jfet input types having a noise floor that is well above the test set's residue.
- Phase of each harmonic residue is indeterminate. For devices having harmonics close to the residue, at any frequencies where the test sets' harmonic is similar or larger, cancellation (down to $-\infty$ dB) or addition (up to +6dB) results. In these instances, the true magnitude of each harmonic is uncertain. For harmonic magnitudes less than 6dB above the AP's residue, the 4th, 6th and 8th to 11th harmonics – where the test set has no visible residue – are the most certain.
- Analog Devices has noted that in a similar test set-up a steel shielding plate located a few mm under the device under test (as in this test) increases 3rd harmonic content in its

Earlier testing

In a recent article[5] the Audio Precision test set's DSP-based enhanced harmonic analysis was used in a lesser mode to examine the spectrum of the residue when an amplifier was driven with a 1kHz tone, before and after changing the mode of DC blocking capacitor in the feedback's lower arm. While differences were clear-cut, more than one reader commented that the harmonics are multiples of the UK's 50Hz mains frequency. The extra parasitic inductance (including a length of exposed conductor) arising from the back-to-back capacitor configuration may well have increased 50Hz harmonic pickup, as the amplifier has an unshielded (albeit toroidal) mains transformer close to both channel's circuit boards.

But any attempt to shield the transformer or move the driver PCB to "improve" the test would detract from the validity of the measurement, ie the patient was a working sample with only one explicit variable.

Whether or not audiophiles prefer back-to-back electrolytics only because 50Hz AC mains pollution is added or changed, the measurable difference stands. Cause of the difference is interesting, but relatively unimportant.

Responding to further points made in the *EW + WW* Letters column, the harmonic magnitudes are quite distinct from grid lines on screen, and measures have been taken to clarify them on the page.

V-I non-linearity in the elcap should not be an issue, as the DC voltage across the capacitor is well below 500mV, and capacitative impedance is low enough to keep even the worst case AC swing (at 20Hz) below 10mV in the audio band. It is also worth noting that the sonic aberrations introduced by tone controls are of a vastly greater nature than the harmonic changes discussed here.

AD797. Other types with similarly low harmonic distortion may be similarly affected. Several op-amps operate in this condition, but many others are spaced well away from high-μ metalwork. An ideal test must embrace both conditions.

- Loop gain loss caused by the 600Ω loading will be more severe in some op-amps than others, notably the *TL071*, *LF351* and *OP27*. Also harmonic structure may differ from high gain, low load condition. But a higher load will reduce the amount of certain data above the noise floor that can be captured for the op-amps with low residues. The residues apply solely to the simple circuit shown, at the 1V level. Harmonic patterns may be very different at lower and particularly higher levels, and if common-mode distortion is dominant, results will also differ with a purely inverting topology.
- Finally, the residues have to be related to psychoacoustics. High harmonics require strong weighting, but the formula may depend on whether least audibility, or least objectionableness is the target. The magnitudes also require relating to sound pressure levels in the room, to provide an overlay for aural sensitivity at the harmonics' frequencies[3, 4].

The device list and harmonic ranking Table (see boxes) skip these reservations and make fair use of the available data, ranking the devices in different ways for each even, then odd harmonic. In the harmonics Table, devices occupying a high position on the left, descending to the lowest boxes on the right, should sound euphonic. Transparent devices can be expected to occupy only the lower boxes, not more than 20dB above the test residue, which broadly indicates a sound pressure level below the threshold of perception.

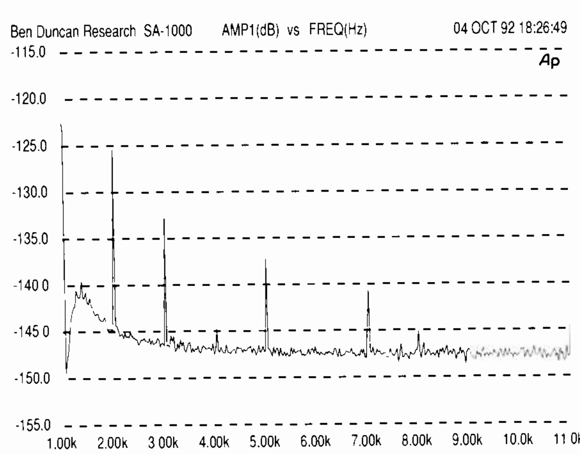


Fig.17. Typical residue of the AP test set. Despite averaging 256 samples, it varied slightly with every "take", particularly the 3rd harmonic.

Path through the chaos

Recognising the view-in-isolation mistake made by many equipment designers and academics clears up the chaos-like complexity of audio design based on pure objectivity. Even after data have been extended and fully processed (as dictated above) and we feel we have reached a full and fair judgement as to what is audible, harmonic preferences will then depend on the product of the harmonics of all the cascaded stages.

In effect, an op-amp with a nominally poor harmonic structure may cancel or harmonise with products created by a subsequent mosfet output stage.

So far, little or no data has been published showing the harmonics generated by resistors (some interim tests with the Audio Precision failed to resolve any at 1V and 230μA RMS). Time taken to model these factors and the increasing matrix of possibilities, plus the decreasing certainty, explains why the bottom line – listening – remains the central tool of development by companies who make their living not by theorising, but by designing and making equipment to satisfy critical listeners.

Measurement is never more than a map, ever finite in its bounds, and is often no more than a pencil beam, rather limited in audio's dark spaces. ■

Acknowledgements The author is indebted to Walt Jung, Analog Devices, Audio Synthesis, Burr-Brown, Linear Technology, Harris Semiconductor UK, SSE Marketing and TI.

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2. Ben Duncan, "Evaluating Audio Op- Amps", Pts.1-3, *Studio Sound*, July-Oct '90.
3. Ben Duncan, "Harmonic Convergence", *Stereophile (USA)*, Oct '92.
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Test group

How were the fifteen op-amps tested arrived at? They had to be types previously or presently used for audio – or recently introduced with audio in mind – and they had to work in an existing test circuit, without adding compensation and/or adjusting the test circuit values. In general, excluded devices are those not rated for use at unity gain, and current-mode types.

Grouped in two categories and approximately listed in historic order, oldest first:

Model & grade	Type	Company or manufacturer
(all jfet inputs)		
TL071	CP	Texas
LF351	-	Motorola
LT1056	CN8	Linear Tech
AD744	AQ	Analog Devices
OP42	FZ	PMI (AD)
AD845	JN	PMI (AD)
LT1122	CCN8	Linear Tech
OPA604		Burr Brown
(all bipolar low noise)		
NE5534	AN	Signetics
HA2525	–5	Harris
OP27	GP	PMI (AD)
LT1007	ACN8	Linear Tech
HA5221	–5	Harris
TLE2027	P	Texas
AD797	XN	Analog Devices

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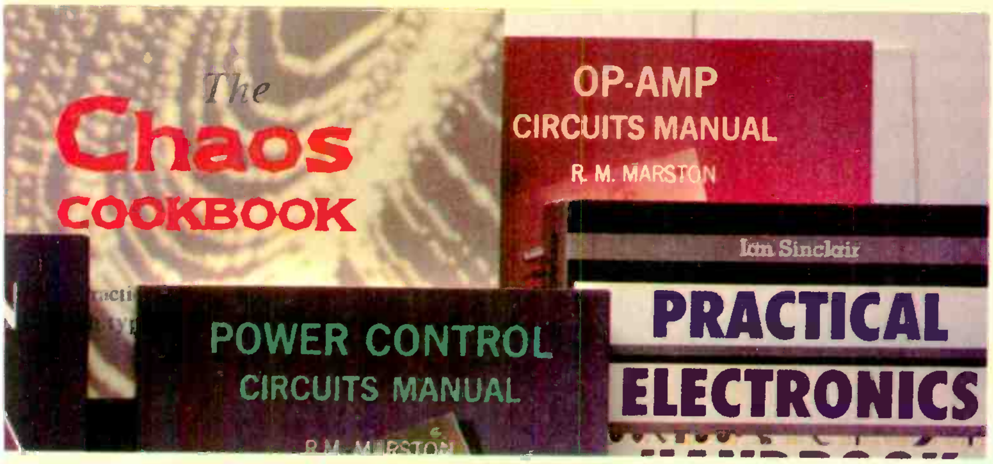

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A guide to microprocessor architectures explaining the differences, advantages and disadvantages, and design implications. With most semiconductor manufacturers supplying only one type of architecture, processors becoming more complex than mainframe computers, and design cycles shortening from two years to six months, it is vital for designers to choose the right processor. Steve Heath is processor technology specialist for Motorola.Hardback 300pages.

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Sony's NT system stores 2h of stereo on a tiny cassette. But who wants it?

PALMED OFF with a new format

Where will Sony's NT cassette format fit into the market? George Cole wonders if Sony is asking the same question.

Just when you thought you had seen enough new audio formats to last a lifetime, along comes Sony with its NT (non-tracking) tape system, storing up to 2h of audio stereo sound on a cassette tape the size of a postage stamp.

The NT player dispenses with advanced tracking systems. But it is size that is its most striking factor. Sony's player is powered by one small alkaline battery and the cassette is around 25 times smaller than a compact cassette – one quarter the size of a microcassette. Cassette weight is only 2.8g and a small lid, rather like those found on video and dat cassettes, protects the tape from dust and dirt.

NT is bi-directional, so the tape is turned over after the completion of one side, with two write-protect slots – one for each side.

Recording is by helical scan. But unlike VCR and dat, the tape is not drawn from the cassette and threaded around the head drum. NT cassettes have built-in tape transport handled by two pinch rollers and two moulded tape guides located just inside the cassette. The head drum slots inside for recording and playback. As a result loading is simpler and cassettes can be manually inserted.

The tape itself is metal evaporated (ME), a mixture of cobalt and nickel where the metals are boiled in a vacuum chamber and condense on to a chilled base film to form ultra-thin tape. Total tape thickness is around 5µm, with the magnetic layer just 0.2µm thick (compared to 3µm in metal powder tape) and width is 2.5mm. A 2h NT cassette holds around 20m of tape and has a data capacity of 690Mbytes.

ME tape has many advantages. Its thinness means the recording magnetic field can easily penetrate down to the bottom and over-write any existing data, so no erase head is required. It also has a lower coercivity than MP tape (around 1050 oersted compared to 1500 oersted) so that a lower recording current can be used. Remanence is higher than that of MP tape too (around 4000 gauss compared to 2600 gauss) and there is a higher output level and carrier-to-noise ratio.

Main disadvantage of ME tape is its cost, because the tapes are made by a batch rather than continuous process.

Recording and playback

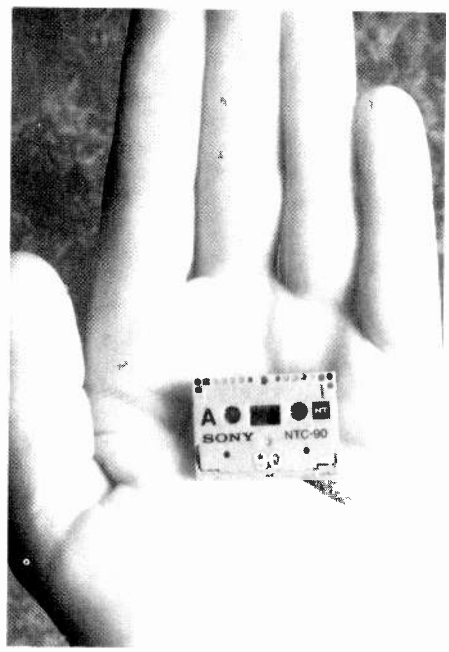
The NT player's head drum is 14.8mm in diameter and rotates at 3000/min. Tape wrap angle is 100°, speed is 6.35mm/s, and each track is around 9.8µm wide at an angle of 4.4°. Two-channel stereo sound is recorded and the sampling frequency is 32kHz, with 12-bit non-linear quantisation – Sony claims this is equivalent to 17 bits. Error detection and correction is by cyclic redundancy check code and cross interleave Reed Solomon code. Recording of the tracks is almost haphazard: no attempt is made to minimise the risk of tracks over-lapping or them being recorded at the wrong angle. This is because there is a repeat area at the top and bottom of each track and the non-tracking system deals with any misaligned tracks.

Non-tracking

Video 8 and dat formats use complex pilot tone tracking systems to ensure that the tape and heads are properly aligned: VHS uses a control track. But NT has no tracking system. Instead, it uses a double density scan.

One way of achieving double density would be to double the drum rotation speed, but NT uses an alternative system. The head drum has two head units composed of three heads: A, A' and B. Head A is used only for recording and its azimuth is +27°. Head A' is for playback only (+27°), and head B for recording and playback (–27°). During recording, only one head unit is in operation, but during playback, both head units work to double the amount of data read off-tape.

A track contains 116 data blocks, including eight unused data blocks – three at one end and five at the other – occurring at the head switch-over points. So 108 blocks are actually usable. Two sets of repeat blocks, copies of the first blocks at the edge of the tracks, act as a back up system in case there is edge damage to the tracks or track over-lap during recording. The remaining 92 blocks are divided into two blocks of 40 separated by four control blocks, one inter block gap (IBG) two auxiliary blocks, another IBG and another four



NT format is 25 times smaller than a compact cassette. But can it compete with disc-based systems?

A single AA-sized battery provides 7h recording or 6h playback.

Cassette tape systems

Cassette systems for open reel tapes were developed in the 1950s and 60s by Grundig and others, and in 1963 Philips introduced the compact cassette – originally designed as a dictation medium. Six years later Olympus launched the microcassette. Olympus' system was promoted as a rival to the compact cassette, but its slow tape speed and limited frequency response (plus a lack of software support) resulted in the format being pitched at the dictation recording market (the reverse of Philips' experience with compact cassette).

The early 1980s saw the rise of audio VHS recorders, offering audio-only recording, and in 1985, the Video 8 format was launched. Some Video 8 machines have an optional PCM audio-only recording system so that a 90 minute pal cassette played at half speed can store up to 18h of digital stereo sound.

Digital audio tape or dat, reached the European consumer market in 1990 and stores several hours of high quality PCM sound on a credit card-sized cassette. Philip's digital compact cassette (DCC) holds around two hours of high quality, digitally compressed audio on a standard-sized cassette.

Now we have NT. In fact NT development work began at the end of 1980 – two years before CD was launched in Japan and around six years before dat reached the market. The design goals were to produce a very small cassette and player which used little power and gave 90min of stereo music.

Various events have since over taken these original goals, not least the launch of DCC and Sony's Mini Disc. Interestingly, Sony now says that future audio formats will be disc-based and that tape technology is old technology. For this reason, NT is being promoted as a high-tech replacement for the microcassette.

Comparison of tape formats.

	Micro cassette	Compact cassette	Video 8 (PCM)	Dat	DCC	NT
Cassette size	73x52x 20mm	102x64x12mm	95x62.5x15mm	73x54x10.5mm	100x64x9.6mm	30x21.5x5mm
Tape width	3.81mm	3.81mm	8mm	3.81mm	3.78mm	2.5mm
Tape speed	24.0mm/s	47.6mm/s	20.1mm/s	8.15mm/s	47.6mm/s	6.35mm/s
Sampling freq	–	–	31.25kHz	48kHz	48kHz	32kHz
Quantisation	–	–	8bit	16bit	16bit	12bit
Frequency Resp	400-4kHz	40-18kHz	20-15kHz	4-22kHz	5-22kHz	10-14.5kHz
Max rec time (SP)	60min	120min	540min*	120min	120min	120min
Standard tape type	Ferric oxide	Ferric oxide	Metal powder	Metal powder	Chrome	Metal evaporated

* With 90 minute cassette. Note that dat and DCC have variable sampling rates.

control blocks. The control block contains time code, play back level and emphasis information and the serial copy management system (SCMS) prevents multiple digital dubbing. IBGs keep the auxiliary and control blocks separate and make it possible to re-record the auxiliary block. The auxiliary block is currently unused (Japanese designers always like to keep their options open!). Sony says they could be used for recording sub-codes for a track-search system.

Each block contains 288 bits, the first 11 of which are used for synchronisation. The address area is composed of 13 bits: six are for track address and seven for block address. The next part contains four parity words (odd and even), each composed of 12 bits. These are followed by 16 12-bit words which contain left and right channel audio data. At the end of the block are 24 bits for the cyclic redundancy check system and its back-up overwrite protect coding.

During playback, the head skims across a series of tracks. For any particular track during the first sweep, data blocks 1 to 5 are read, during the second sweep, blocks 4 to 13 and so on. Only data which is "safe" (ie. error free) is read into a buffer memory. Each block has an unique address and the memory buffer has an area reserved for it. Sony says that the memory buffer is like a jigsaw with many slots and that the data blocks are like pieces with their number written on the back.

Incidentally, the data blocks are not built up in sequence, but the end result is a complete set of blocks for each track.

No servo control is required for the non-tracking system. But a servo is needed to correct tape speed fluctuations, because there is a delicate balance between the rates at which the data is read off-tape, fed into the buffer memory and read out of memory. In the NT's servo system, the address is separated from the playback data and compared to a reference address produced by the quartz generator. Any difference between the two addresses means the speed at which the data is being fed into the memory buffer is faster or slower than the rate at which it is being read out. The offset value is subtracted to produce the phase error.

Phase error is averaged out by a digital low pass filter and the gain adjusted and mixed with the speed error component derived from the motor tacho generator. A pulse width modulator (PWM) converts the servo signal into a PWM signal and at the second low pass filter the signal is demodulated to produce the motor drive voltage. A micro-controller LSI performs the servo arithmetic processing.

The NT-1 Player

The chocolate-bar-sized NT-1 player uses a single AA-sized battery to give 7h recording time or 6h playback. Of the ultra-compact drum, just 14.8mm wide. the section that slots into the cassette is less than 4mm in length. The drum is mounted horizontally and only the middle section rotates – a fine technical achievement. But the price is that the motor cannot be serviced. If any fault occurs, replacement is the only option.

The NT-1's circuit board, tightly packed and folding up like a concertina, contains nine LSI and IC components, six of which have been specially developed for the NT system:

- DSP LSI: tasks include A-to-D and D-to-A conversion, non-tracking processing, error correction and detection. The non-tracking buffer and servo buffer are loaded into 1Mbit dram.
- ADA LSI: contains the A-to-D/D-to-A converter, audio circuits, cmos op-amp and digital circuits.
- DET LSI: includes the RF equaliser and PLL.
- Micro-CTL LSI: one-chip servo microcomputer includes servo control software and LCD drive.
- DRV LSI: uses DC-DC converters and regulators to extract voltage from the AA battery.
- R/P IC: recording and playback amplifier.

Connection of the LSIs by a high speed simple serial bus allows large data exchange and real time control between the microcomputer and the LSIs. It also reduces the number of LSI pins and wires. A non-volatile ram serves as a back-up for the control data.

Scooping the market?

Tapes for the £545 NT-1 or Scoopman (because the non-tracking system "scoops" up the data) cost around £10 each. The player itself measures 115 x 50 x 21mm and weighs around 138g, including battery and tape. Power consumption is around 27mW.

Scoopman is easy to use, great for carrying around in a pocket or bag and sound quality is also good – Sony says the dynamic range and signal-to-noise ratio are both over 80dB, and total distortion is below 0.05%.

But here lies the problem. The NT-1 is too good – and expensive – to be used simply as dictation system and yet it is hard to see it succeeding as an audio product, especially as Sony is busily promoting Mini Disc. So who will buy the system?

The Non-Tracking system shows that Sony undoubtedly has the technology, but whether it has a market remains to be seen.

Acknowledgment

Thanks to Sony's Eric Kingdon and Chris Baker for their help during the preparation of the article.

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TYPE 9177 4 watts output 20–200MHz 26dB gain £314
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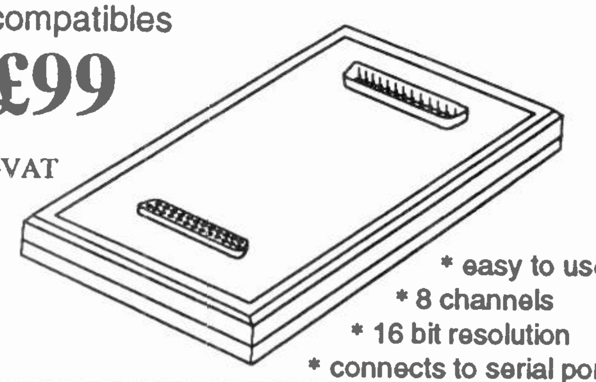

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Making switched DC/DC conversion easier – and cheaper

The newest switched mode power supply chips are so undemanding to work with, they make design of a linear supply look difficult. Scott Brown shows how.

Linear regulators are a simple way to convert from the relatively high bus voltage down to the logic levels of 5V and 3.3V required for distributed power supplies. But large voltages across the regulator cause high power dissipation, requiring costly heatsinks and possibly a cooling fan.

Switching regulators can drop large voltages with efficiencies in excess of 80%. Silicon costs are higher than with linear regulators, but less heatsinking is required and power losses are reduced so a smaller power bus and a lower-power transformer can be used. The main drawback is that lots of components are needed along with a good understanding of switching techniques. DC/DC modules offer high efficiencies and guaranteed performance but they also come with a high price tag.

The switched mode supply combines high efficiency and off-the-shelf performance of modules at the cost of a switching regulator. In addition, National Semiconductor's *Simple Switcher* family has been designed to need fewer components than conventional switching regulators so that less board space is needed and building is easier.

Simple Switcher™ is a trademark of National Semiconductor Corporation.

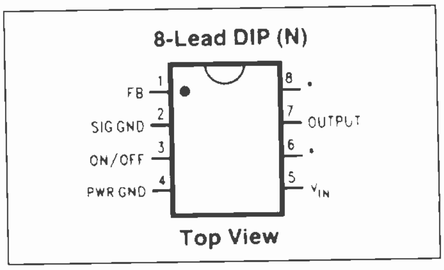
Two examples demonstrating how SMPSs can be used for stepping down and stepping up are the *LM2574* and *LM2577*.

DC/DC conversion

A typical application for a DC/DC converter using the *LM2574* is a 24V bussed power supply feeding a board packed with 3.3V logic (the same design procedure would apply for a 5V system).

There are four external components to select and essentially three design steps: determine $V_{in(max)}$ and $I_{out(max)}$ and select inductor from chart; choose output capacitor; and select catch diode according to $V_{in(max)}$. The *LM2574* datasheet takes the designer though the selection process step-by-step, providing values for the four components along with recommended manufacturers and part numbers of the products.

But design software *Switchers Made Simple* considerably simplifies the task. In this example the user starts by selecting a step-down design (referred to as BUCK in the software), then enters the input parameters, maximum and minimum input voltage range, followed by the operating temperature range, the output voltage and current. Calculations are executed once all the parameters have been entered by hitting the END key.



LM2574 PINOUT – 8-PIN DIP

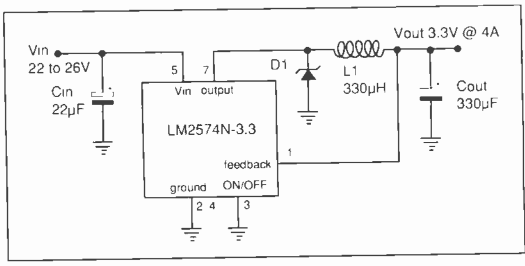


Fig. 2. LM2574 schematic diagram is generated by the software.

Component limits and suggested components values are displayed (**Fig. 1**) – values can be altered and performance re-calculated – and once component values have been selected, the cross-over frequency and phase margin are calculated and displayed. Thermal analysis can now be performed for a particular package type with the software calculating junction temperature and determining the need for a heatsink.

Finally the design can be saved, and the schematic diagram (**Fig. 2**) displayed and printed along with a complete parts list.

A printout of the schematic, a complete parts list and a design – guaranteed to work

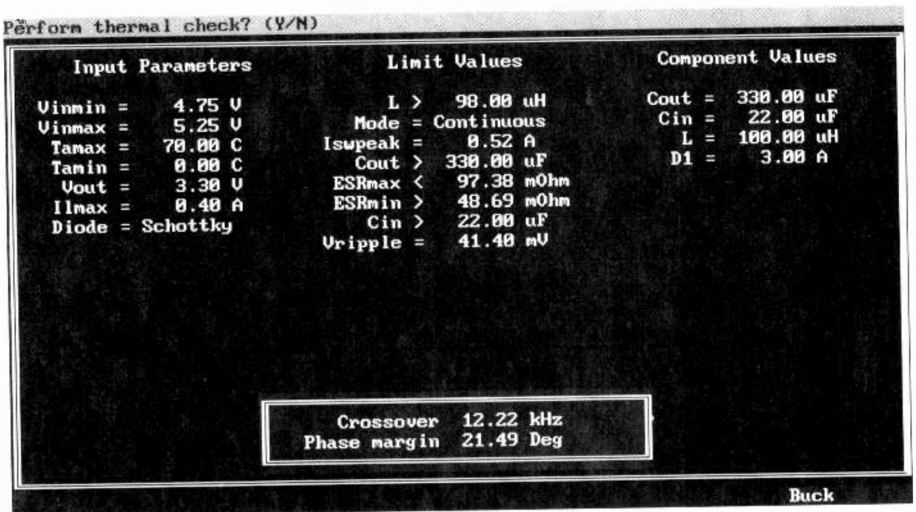


Fig. 1. Software considerably simplifies the design process.

$$a).\ I_{LOAD}(max) < \frac{2.1A \times V_{IN}(min)}{V_{OUT}} \Rightarrow 1.17\ A$$

$$b).\ D(max) = \frac{V_{OUT} + V_F - V_{IN}(min)}{V_{OUT} + V_F - 0.6V} \Rightarrow 0.506$$

$$c).\ E{*}T = \frac{D(max)\ (V_{IN}(min) - 0.6V)10^6}{52kHz} \Rightarrow 40.38\ V.\mu s$$

$$d).\ I_{IND,DC} = \frac{1.05 \times I_{LOAD}(max)}{1 - D(max)} \Rightarrow 2.13A$$

$$e).\ C_{OUT} > \frac{0.19 \times L \times R_C \times I_{LOAD}(max)}{V_{IN}(min) \times V_{OUT}} \Rightarrow 1200\mu F$$

$$f).\ C_{OUT} > \frac{V_{IN}(min) \times R_C \times (V_{IN}(min) + (3.74 \times 10^5 \times L))}{487{,}800 \times V_{OUT}^3}$$

$$g).\ I_{RIPPLE}(P\text{-}P) = \frac{1.15 \times I_{LOAD}(max)}{1 - D(max)} \Rightarrow 2.33A$$

$$h).\ ESR < \frac{0.01 \times 8.5V}{I_{RIPPLE}(P\text{-}P)} \Rightarrow 25m\ Ohms$$

$$i).\ ESR < \frac{8.7 \times 10^{-3} \times V_{IN}}{I_{LOAD}(max)}$$

$$j).\ R_C < \frac{750 \times I_{LOAD}(max) \times V_{OUT}^2}{V_{IN}(min)^2} \Rightarrow 2.4K\ Ohms$$

$$k).\ C_C > \frac{58.5 \times V_{OUT}^2 \times C_{OUT}}{R_C^2 \times V_{IN}} \Rightarrow 220nF$$

$$l).\ V_{OUT} = 1.23V(1 + R1/R2) \Rightarrow R1 = 11.8KOhms,\ R2 = 2KOhms$$

Fig. 3. Datasheet calculations for the LM2577

V_{IN} = Input Voltage
$V_{IN}(min)$ = 4.75V
V_{OUT} = Output Voltage = 8.5V
V_F = Diode Forward Voltage
= 0.5V (Schottky Diode)
D(max) = Duty Cycle
$I_{IND,DC}$ = Average Inductor Current
C_{OUT} = Output Capacitance
I_{RIPPLE} = Ripple Current
C_C = Compensation Capacitance
R_C = Compensation Capacitance

first time – are all produced within minutes.

Generating new voltages

Linear regulators can only take an input voltage and produce an output voltage of the same polarity with smaller magnitude. But switching regulators are equally well suited to stepping-up or inverting.

For example the *LM2577* can be used to step-up, or generate multiple outputs (flyback mode). As before, software can be used to do the hard work. But datasheets and calculation will still reach the same solution.

To demonstrate the datasheet route take the example of a DC/DC converter with an output

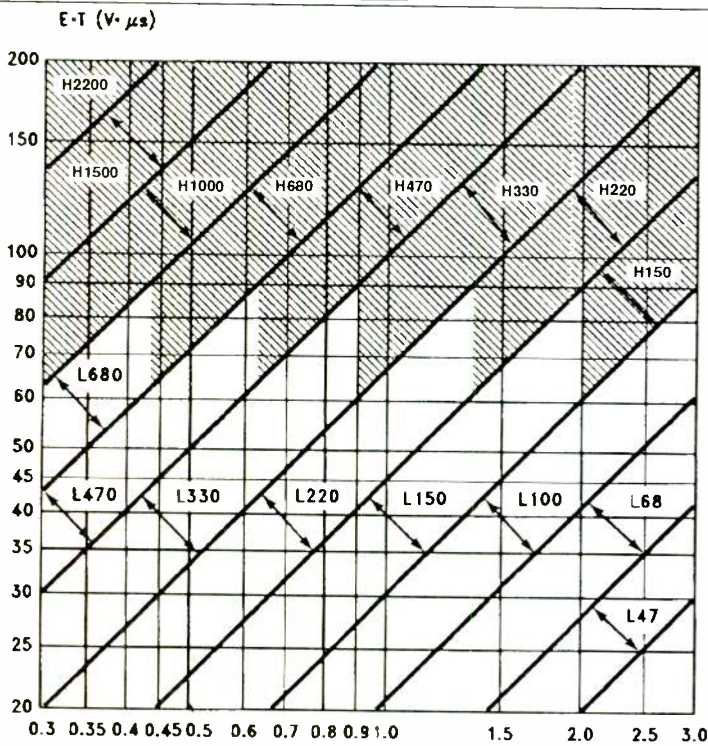


Fig. 4. LM2577 inductor selection graph.

Distributed Power Supplies

Distributed power supplies are not a new idea but they are becoming increasingly popular because they offer better regulation at the load, isolation between boards and greater design flexibility.

Traditional centralised power supplies gave a range of supply busses at the various voltages required and each board within the system would take its supply from one or more of the buses. Unfortunately loads at the bus extremities receive a degraded voltage, and the faster switching speeds of today's logic devices make current demands that corrupt the bus voltage. Centralised buses also offer no isolation between loads.

In a distributed power supply, one intermediate DC voltage generated in the main supply is bussed around the system. The chosen bus voltage varies from system to system but is typically in the region of 20 to 40V. In theory, higher bus voltages mean lower bus currents and so cheaper, smaller buses. This of course assumes ideal power conversion at the load – a critical factor! But ideally, the bus is efficiently converted at the load to the required voltage. Regulation is improved and isolation between loads can be achieved.

Perhaps the largest distributed power supply is found in the telecommunications industry. British Telecom (like other telecommunications operators) buses out an intermediate voltage – in this case -48V – converted at the load, the telephone.

At the other extreme, the "bus voltage" in a modern hand-portable phone unit comes from five nicad cells, equating to a nominal supply of 6V. Every phone will have three or more regulators situated at the load, and a distributed approach is taken to achieve isolation of the loads.

Mobile phones are perhaps an exception as in most cases distributed power supplies rely on DC/DC converters that are capable of dropping large voltages both efficiently and cost effectively.

Table 1. Standardised inductors and manufacturer's part numbers

Inductor Code	Manufacturer's Part Number		
	AIE	Pulse	Renco
L47	415 - 0932	PE - 53112	RL2442
L68	415 - 0931	PE - 92114	RL2443
L100	415 - 0930	PE - 92108	RL2444
L150	415 - 0953	PE - 53113	RL1954
L220	415 - 0922	PE - 52626	RL1953
L330	415 - 0926	PE - 52627	RL1952
L470	415 - 0927	PE - 53114	RL1951
L680	415 - 0928	PE - 52629	RL1950
H150	415 - 0936	PE - 53115	RL2445
H220	430 - 0636	PE - 53116	RL2446
H330	430 - 0635	PE - 53117	RL2447
H470	430 - 0634	PE - 53118	RL1961
H680	415 - 0935	PE - 53119	RL1960
H1000	415 - 0934	PE - 53120	RL1959
H1500	415 - 0933	PE - 53121	RL1958
H2200	415 - 0945	PE - 53122	RL2448

Table 2. Diode selection chart.

V_{OUT} (max)	Schottky		Fast Recovery	
	1A	3A	1A	3A
20V	1N5817 MBR120P	1N5820 MBR320P		
30V	1N5818 MBR130P 11DQ03	1N5821 MBR330P 31DQ03		
40V	1N5819 MBR140P 11DQ04	1N5822 MBR340P 31DQ04		
50V	MBR150 11DQ05	MBR350 31DQ05	1N4933 MUR105	
100V			1N4934 HER102 MUR110 10DL1	MR851 30DL1 MR831 HER302

of 1.0A at 8.5V from an input of 5V (±5%).

Input range of the *LM2577* is 3.5V to 40V, so a 5V input is no problem, and the output voltage of 8.5V is also well within the specified 65V maximum. Care must be taken with the load current. The *LM2577* has an output switch current maximum of 3.0A, but maximum load current has to be calculated (Eq a). In this case maximum load current is 1.1A so again the target of 1.0A is within specification.

Inductor selection is based on two parameters, *ExT* (the product of voltage and inductor charge time) and $I_{IND,DC}$ (the average inductor current under full load). To obtain the value of $E*T$, first check the duty cycle (Eq b) is less than 90%. Then $I_{IND,DC}$ and $E*T$ are calculated (Eq c and Eq d) and used with inductor selection graph in **Fig. 4** to obtain an inductor value of *L68*.

The last step is to select an inductor (**Table 1**). *L68* corresponds to an inductor value of 68μH and the table provides part numbers for three manufacturers. AIE inductors use ferrite, pot-core construction, and benefit from low electro-magnetic interference (EMI), small physical size and low power dissipation. The inductors from Pulse use powdered iron toroidal cores and also offer low EMI. They are also capable of withstanding ET and peak currents above their rated value. Renco's inductors use ferrite bobbin-cores and are low cost though they tend to generate more EMI.

The next stage is to select the values of the output capacitor and compensation network. C_{out} is calculated using two equations (Eqs e and f) with the larger value being the minimum that ensures stability, in this case 1200μF. Equivalent series resistance (ESR) is the primary cause of output ripple voltage so the ESR of the output capacitor is critical. Calculate the ripple current (Eq g) and from this two more equations (Eq h and i) give an ESR of 25mΩ. It is important to ensure that the ESR is less than 25mΩ at the switching frequency of 52kHz.

The compensation network (values calculated using Eqs j and k) also serves as a part of the soft start circuitry. On power-up it causes the switch duty cycle to rise gently. Without it the duty cycle would immediately rise to 90% drawing huge currents from the input power supply.

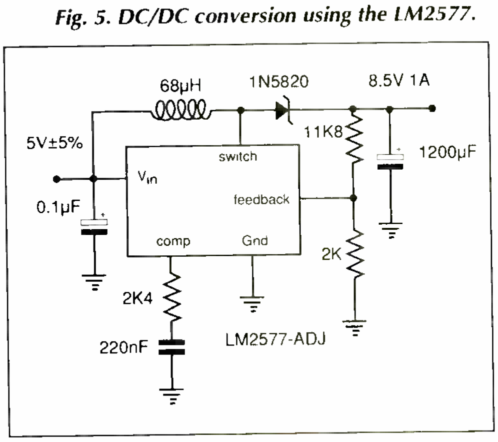
Output voltage is set using two resistors in much the same way that an adjustable linear regulator is set (Eq l). For an output of 12V or 15V these resistors would not be necessary as there are fixed voltage versions available.

An input capacitor with low ESR prevents the triangular switching current corrupting the supply current – a 0.1μF capacitor is normally sufficient.

The diode must be able to withstand a reverse voltage equal to the output voltage and handle the average and peak currents to the load. Schottky diodes are usually favoured as their low forward voltage results in higher efficiency (**Table 2**).

Figure 5 shows the completed circuit incorporating all the component values. ■

Fig. 5. DC/DC conversion using the LM2577.



Switching regulator basics

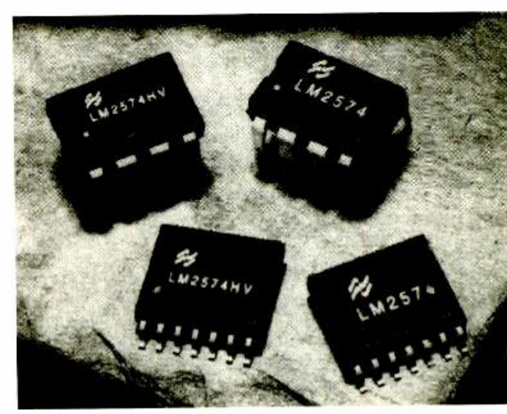
Switching regulators employ a switching transistor to achieve DC/DC conversion, with duty cycle of the switch proportional to the input-output voltage differential.

For step-down applications, output voltage is fed-back to a voltage reference and the error determines the duty cycle of the switching transistor. The output filter, comprising the inductor and capacitor smooth the waveform to DC.

In an ideal system, output voltage is equal to the input voltage multiplied by the duty cycle. But there are losses associated with the diode and the saturation voltage of the transistor. To maximise efficiency Schottky diodes are commonly chosen in preference to fast recovery diodes for their low forward voltage drop.

Output ripple voltage is inevitable when using switching regulators and the magnitude of the ripple is set by the output capacitor. Surprisingly, ripple voltage is primarily a function of equivalent series resistance (ESR) of the capacitor, and is equal to the product of the amplitude of the inductor current and the ESR.

The step-up converter operates in much the same way. The inductor stores energy during the switch on-time and releases it to the load when the switch turns off. The *LM2577* actually operates in current mode so that a voltage is fed-back from the output and compared with the voltage reference. The error signal is then compared with a voltage proportional to the switch current, to determine the switching of the output transistor. Current mode operation provides excellent transient response, good loop stability and is inherently self limiting.



Switched mode power supplies combine high efficiency and off-the-shelf performance with switching regulator pricing.

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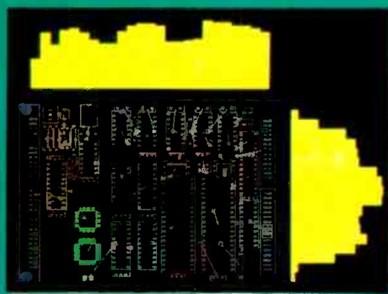

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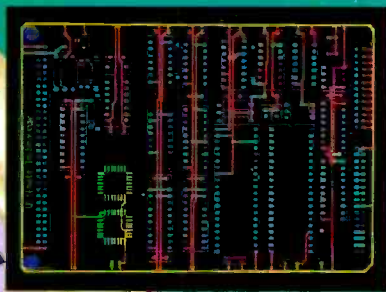

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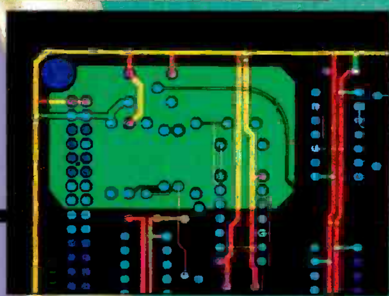

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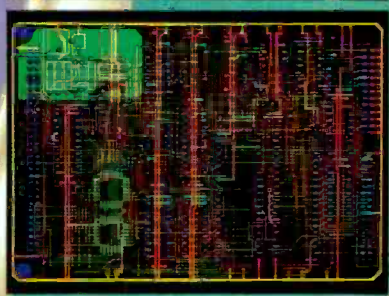

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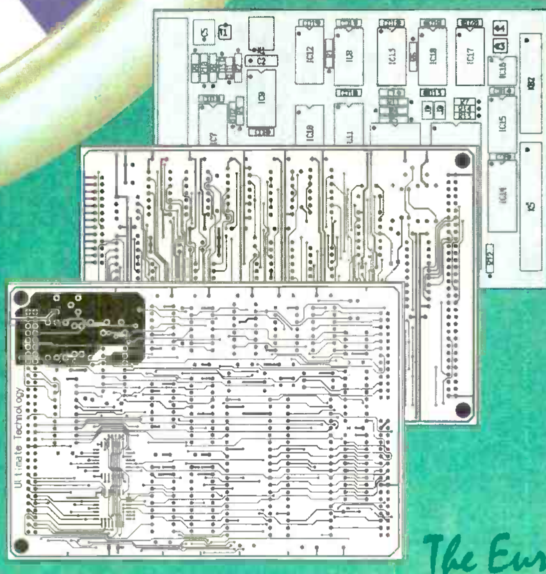

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APPLICATIONS

Switch-mode PSU technology runs at 1MHz

Working in conjunction with IBM, Harris Semiconductor has produced a series of switch-mode PSU chips needing very few external components to form compact power modules suitable for both centralised and distributed power architectures. The new *HIP506x* ICs are capable of up to 100W output with just 6W switching loss – an efficiency of 94% – and are said to permit the design of DC-to-DC converters with power densities greater than $50W/in^3$.

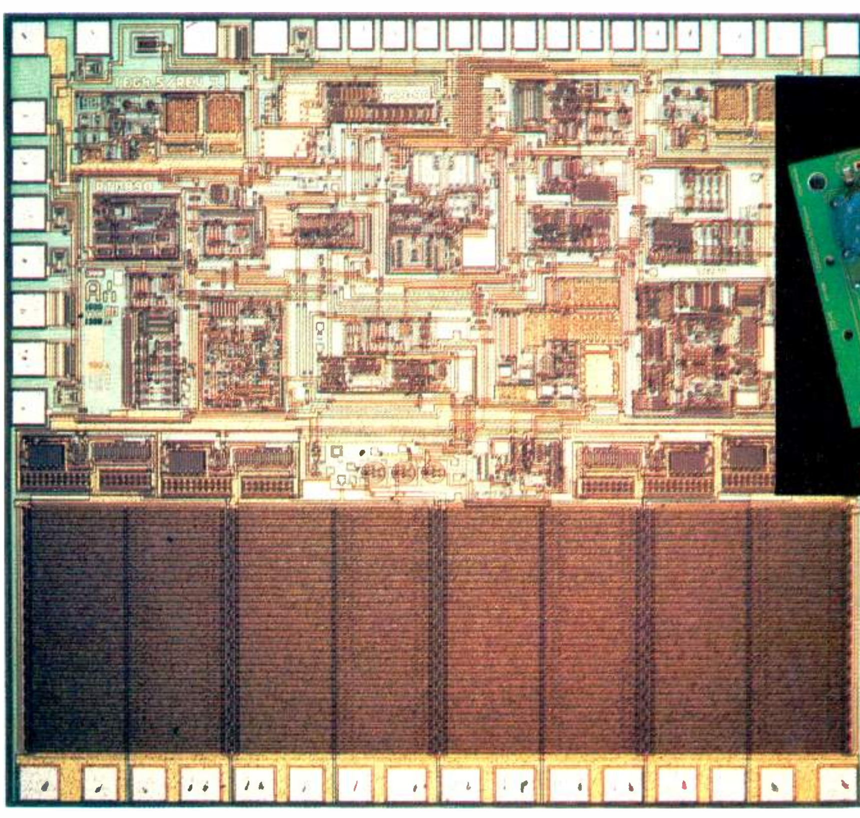
Two important features characterise the new devices. First, they permit operation in current-mode PWM at the relatively high switching frequency of 1MHz. Compact and efficient DC-to-DC converter modules capable of similar switching speeds have been in existence for some time, from Vicor for example, but they mainly operate in resonant or quasi-resonant mode. They provide softer switching than PWM alternatives, but resonant-mode converters often vary widely in operating frequency and require components capable of handling high peak voltages. Conduction losses are higher too. Fixed-frequency PWM architectures are preferable since they are easier to filter and more reliable but until now, resonant-mode converters have had the edge due to their low switching losses.

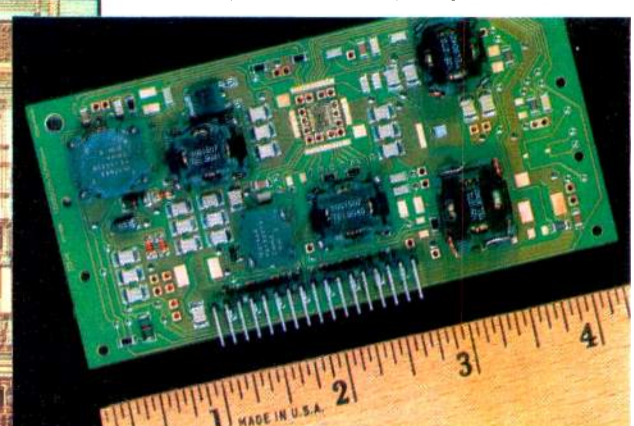
High integration is the second key feature, so that the main power device – a mosfet capable of switching up to 10A at 60V in less than 3ns – forms part of the IC. The only additional silicon needed to form a complete power supply is a diode or two. Special non-dissipative current-sensing circuits for the power-transistor are incorporated to minimise losses, while integral detectors shut down the power section if output voltage chip temperature becomes excessive, protecting both the IC and its load.

Operating at a frequency claimed to be twice that of their nearest rival – the *SGSL4970* – the *HIP506x* series ICs are configurable for almost all topologies, including single-ended primary inductance converters (sepics) and can offer performance better than discrete components, says Harris.

In linear supplies, PCB tracks and power-device wiring form simple resistors. But with nanosecond timings, ampere switching and 60V levels, these same tracks and wires become significant and complex impedances, stopping practical discrete PWM circuits operating above a few hundred kilohertz.

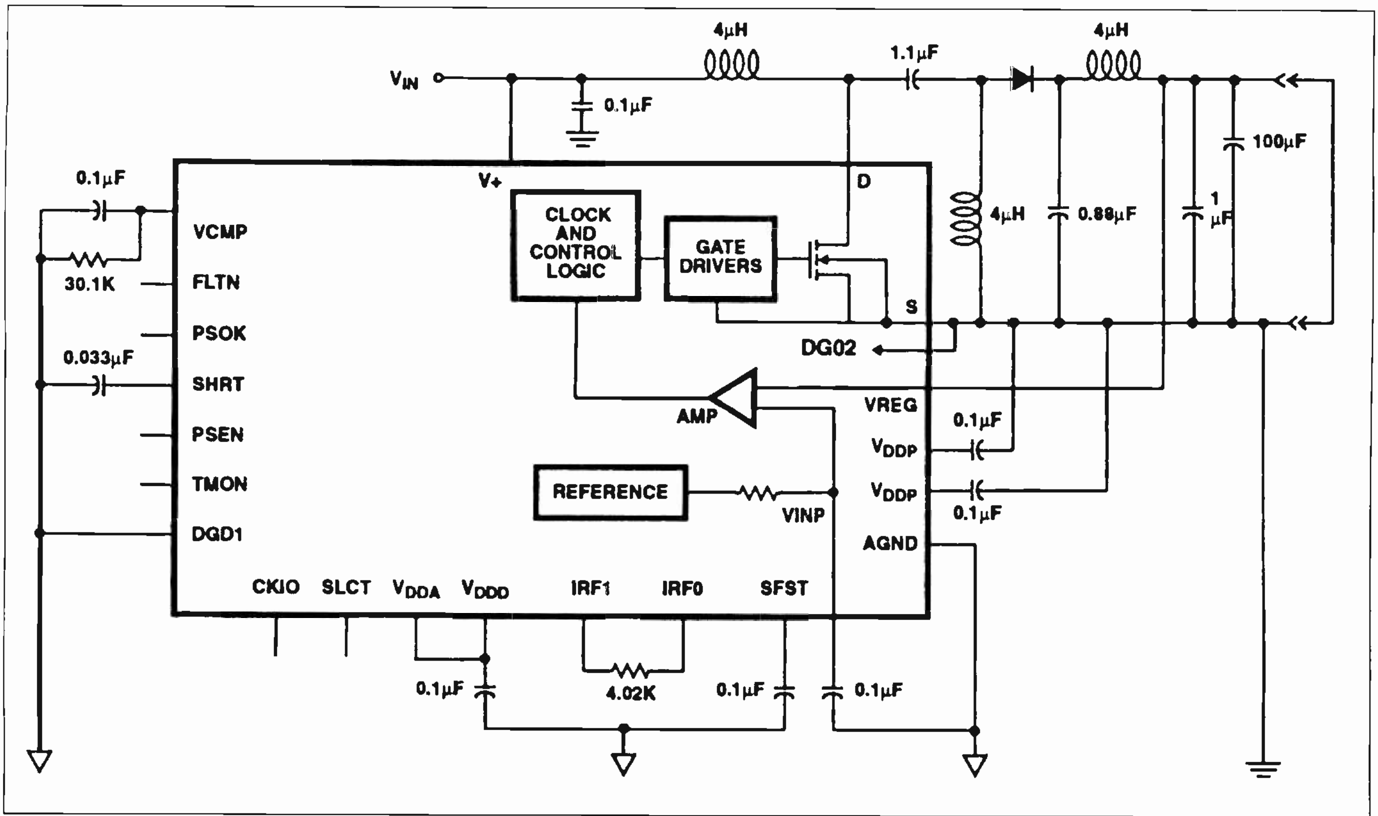
Increased efficiency leads to lower dissipation which in turn leads to higher reliability, as evidenced by a triple-rail DC-



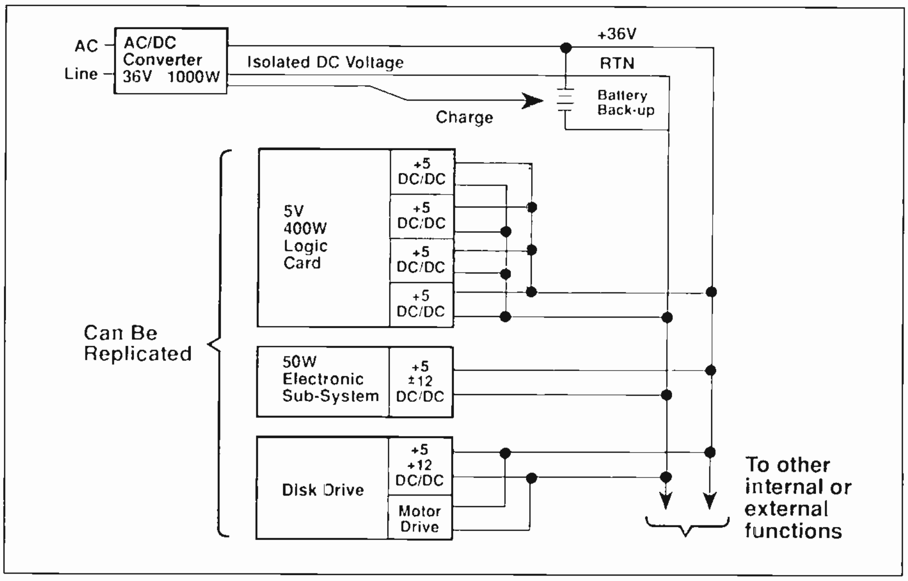


DC-to-DC converter incorporating the HIP506x power asic from Harris is capable of 88W power output, exhibits an efficiency of 82% yet measures only 110 by 55 by 9mm. Operating at a switching frequency of 1MHz, the IBM-manufactured module is particularly suited to distributed power architectures.

Apart from a diode or two, this is all the silicon needed to make a switchmode power supply capable of up to 100W. A significant part of the chip is a mosfet capable of switching up to 10A at 60V in less than 3ns.



The single-ended primary inductance converter, or sepic, is not widely known. But since it seems the best configuration for use with what are possibly the world's best commercial switchmode power ICs, all that could change.



Distributed power systems offer significant advantages in reliability and cost and are now increasing in popularity in the computing field.

to-DC converter module that allows a claimed 5 million hour system MTBF figure - nearly 600 years. This module, produced by IBM, has an efficiency of 82% and its power output is 88W yet it measures only 110 by 55 by 9mm.

Distributed power increases reliability

All the attributes needed for distributed power architectures are incorporated in the *HIP506x*. Distributed power still involves a central AC-to-DC converter but this converter needs no regulation and is easily supported by battery backup. Uninterruptible power supplies are inherently unreliable and, ironically, unsuitable for critical applications. This is due to the need for the additional DC-to-AC converter following the battery for stepping battery voltage back up to AC mains level. Since the final electronic circuitry to be powered by the uninterruptible supply needs DC, this extra circuitry is a waste of money and significantly decreases reliability.

Distributing power via, say, a 36V unregulated rail as opposed to a 5V line also means less cable loss for a given copper cross section. In a distributed power architecture, regulation is carried out remotely by numerous DC-to-DC converters situated at the points where power is needed.

In telecomms, each PCB in a racking system might have its own local card-mounting DC-to-DC converter. This distributes any heat dissipation due to regulator losses and minimises the effects of cable voltage drop. Having a number of converters also means that a failure can only cause a partial system loss, which need not be catastrophic, and the cost of replacement of a smaller distributed regulator is much less that the cost of a large central supply.

Originally, the new switchmode ICs were designed for IBM applications using Harris's Power asic technology, launched mid 1990, which was also jointly co-developed with IBM. Following successful implementations of the chips in distributed-power applications, IBM has granted Harris rights to offer them as standard parts. Dice are available now for hybrid applications and surface-mount packaged devices are expected to be available during the first half of 1993.

Three chips are currently on offer. Two are single-output types and one a dual-rail option incorporating two 60V 5A output devices instead of the 60V 10A transistor featured in the other two. One of the single-output devices has facilities for an external clock and loop-closure amplifier to allow extra design flexibility – a flexibility extending to potential applications of the device to wideband power amplifiers for motor control. All three devices operate up to 1MHz and exhibit 3ns typical switching times.

Applications

Although it contains no fully detailed practical circuits, application note 9208 from Harris is a comprehensive guide to designing high-frequency power converters. It summarises the various switchmode topologies, including Cuk, buck, boost, buck-boost, flyback and forward. But the sepic configuration is discussed in considerable detail. Implicit in the note is the fact that the IBM power module is implemented using this topology, and that sepic is the best solution for high-frequency switchmode converters using the new ICs.

The note points out that Cuk, flyback and forward converters are well known to power supply designers but sepic circuits have not had the airing they deserve.

Now over 20 years old, sepic was devised by AT&T Bell Laboratories who were looking for a configuration with the ability to buck or boost the input voltage without reversing voltage polarity.

Some 17 references are quoted in the note, which spans nine A4 pages.

***Harris Semiconductor,** Riverside Way, Camberley, Surrey GU15 3YQ. Tel: 0276 686886.*

Decoding RDS

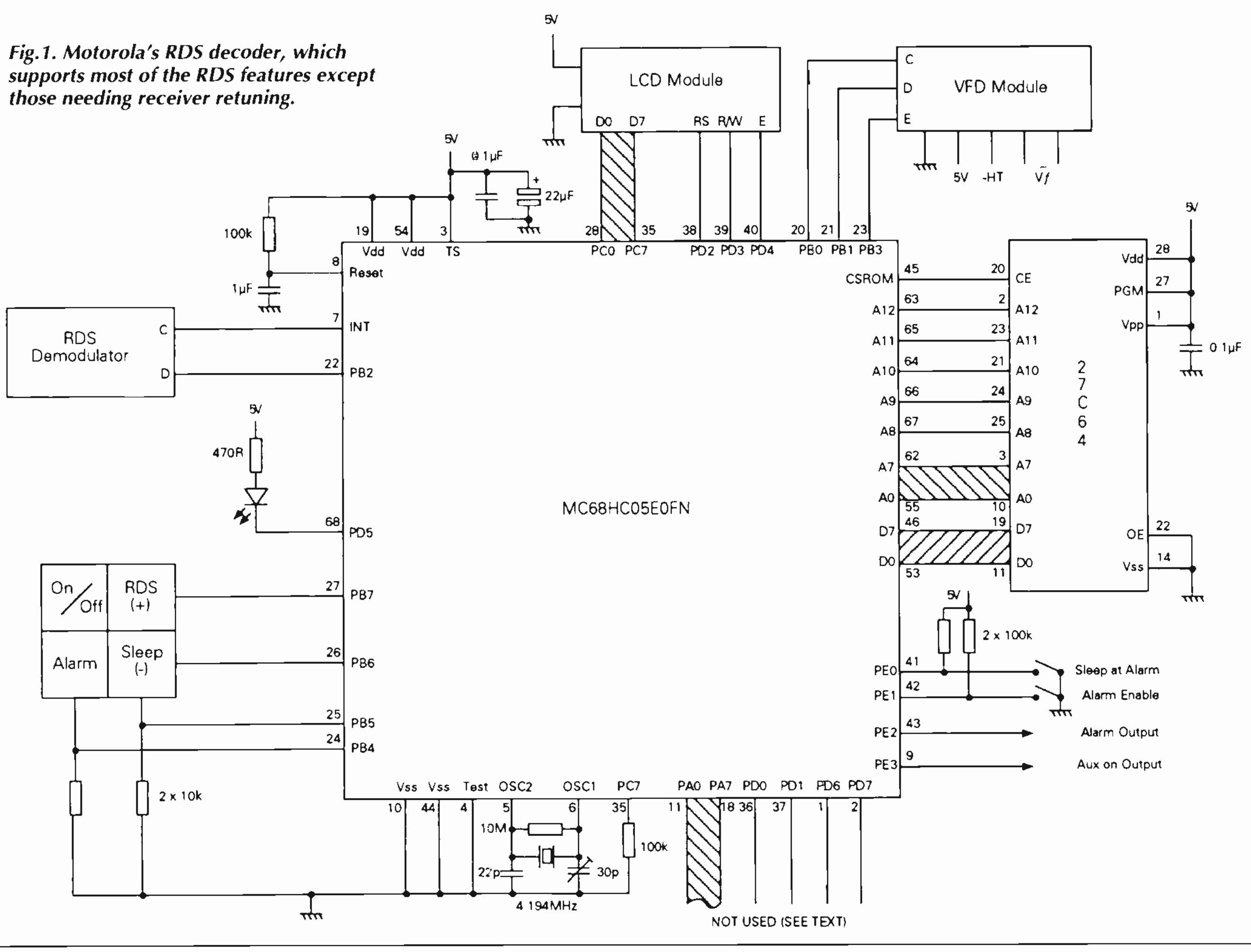
Application note AN460 from Motorola describes the use of an *MC68HC05E0* microprocessor to decode the demodulated radio data system signal and to display the result on either LC or vacuum fluorescent displays. Suggested demodulators are *SAA7579T, TDA7330, LA2231* or RDS hybrids. An alarm-clock function is included which, assuming the unit is permanently powered, could be used to switch on the RDS receiver at the appointed time. RDS data can either be updated when the audio is muted or simply be switched on and off manually. Most of the RDS features can be used, but alternative frequencies (AF) and enhanced other networks (EON) would need some way of retuning the radio. Since the radio forms no part of the design, these features are not supported. Radiotext is displayed on request by scrolling the 16-digit dot-matrix displays.

Figure 1 shows the hardware to be simple — the microprocessor, a *27C64* 8Kword/8bit rom and one or two externals, including a four-button keypad.

AN460 provides a full software listing. EBU Technical Document 3244 gives specifications of RDS.

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

Fig. 1. Motorola's RDS decoder, which supports most of the RDS features except those needing receiver retuning.



Digital audio

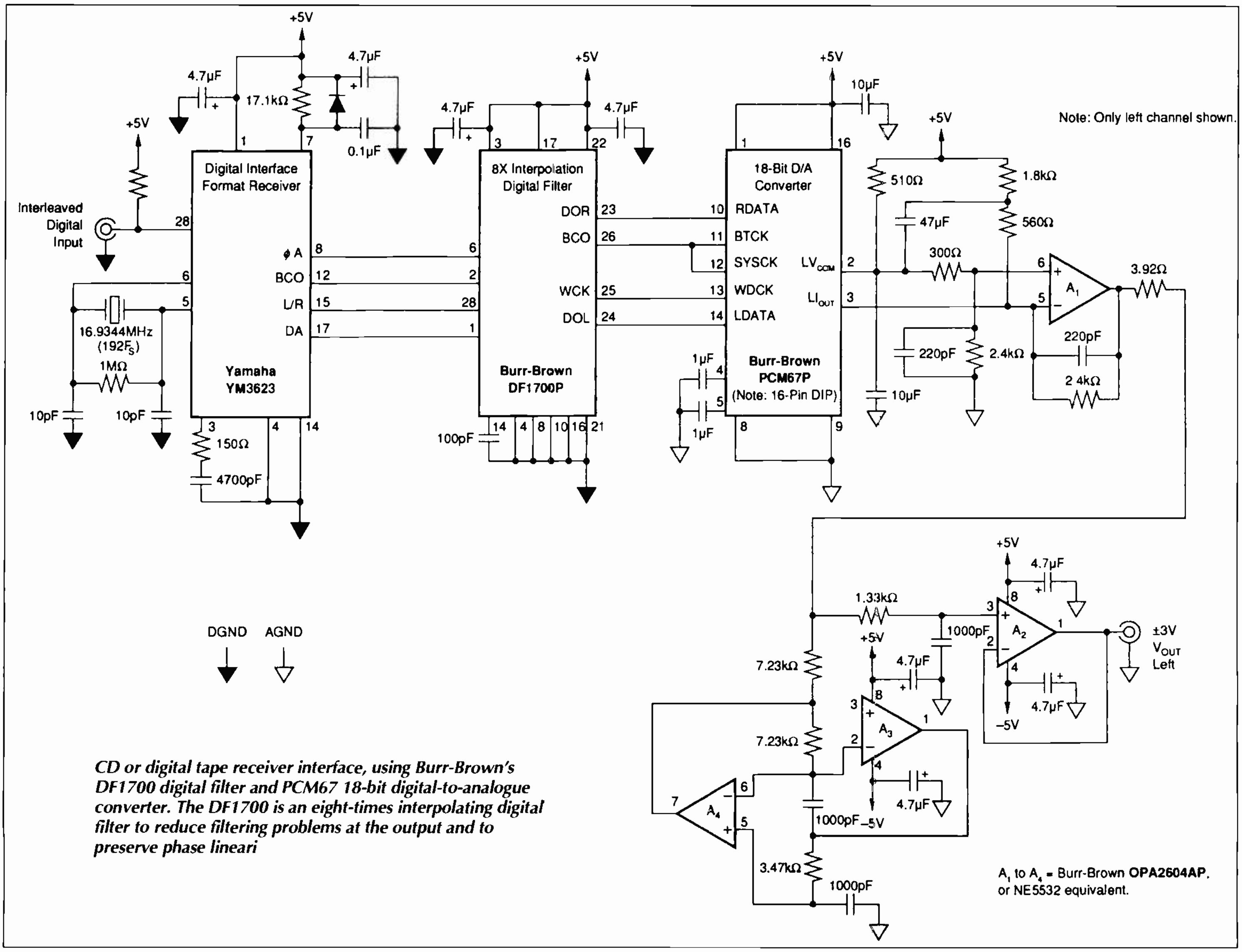
Two chips from Burr-Brown, the *DF1700* digital filter and the *PCM67* D-to-A converter are characterised for use in consumer digital audio playback systems, such as compact disc and tape equipment.

DF1700 is an eight-times oversampling cmos filter, accepting 16-bit input and selectable 16, 18 or 20-bit output. Since the oversampling multiplies the input frequency by eight, the following D-to-A converter accepts a lower-order filter at its output to give a better phase response and simpler design. The application note gives connection details from receivers by Yamaha, Matsushita, Sony, Mitsubishi and Toshiba to the DF1700.

PCM67, the D-to-A converter, is a dual 18-bit bicmos type, which combines a thin-film R-2R ladder D-to-A, a digital offset technique with analogue correction and a one-bit D-to-A to achieve good low-level performance by combining both one-bit bit-stream and conventional 10-bit D-to-A, which is trimmed to 18-bit linearity. This approach minimises the problems inherent in both bitstream and conventional circuits, mainly by reducing the size of major carry error and confining it to higher-amplitude signals.

The diagram is of a circuit combining both the chips in a stereo (one channel shown) audio playback chain, in this case using the Yamaha *YM3623* digital interface format receiver chip. The circuitry at the output is a generalised immittance converter (GIC) filter to filter out the shaped noise of the one-bit D-to-A converter.

***Burr-Brown International Ltd**, 1 Millfield House, Woodshots Meadow, Watford, Hertfordshire WD1 8YX. Telephone 0923 33837.*



CD or digital tape receiver interface, using Burr-Brown's DF1700 digital filter and PCM67 18-bit digital-to-analogue converter. The DF1700 is an eight-times interpolating digital filter to reduce filtering problems at the output and to preserve phase lineari

Gunn diodes

Although not in our usual style of application notes, the introduction to Gunn diodes in GEC Plessey's 1992 Microwave Products handbook deserves mention, since it is a more lucid explanation of device operation than is seen in most introductory texts.

A brief look at basic Gunn action is followed by a description of both transit-time and delayed domain modes of operation; other modes are mentioned but not described. In the second part of the piece, the operating parameters of Gunn diodes are described in detail and the note is rounded off by a section on graded-gap diodes, which give much greater high-frequency power, reduced turn-on voltage and improved temperature stability.

The handbook is obtainable from:
***GEC Plessey Semiconductors Ltd**, Cheney Manor, Swindon, Wiltshire SN2 2QW. Telephone 0793 518000.*

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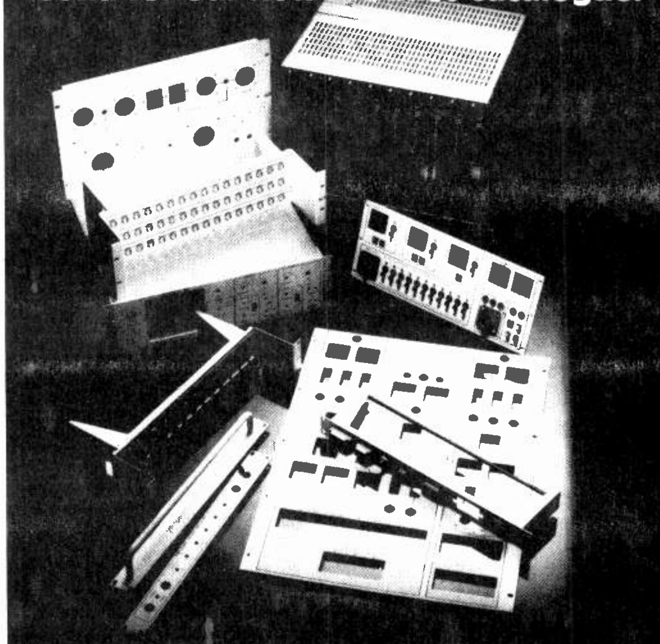

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A-to-D & D-to-A converters

Digital video encoder. The *TRC22090* converts computer graphics into television signals in a digital process, accepting a wide range of formats between 10 and 15 Mpixels/s and producing outputs with separate luminance and chrominance and RGB in NTSC or pal. It will also take a composite digital video stream and soft-key it into the encoded video. Ambar Components Ltd, 0844 261144.

4ns comparator. Maxim's *MAX910/1* are high-speed voltage comparators that accept a digital setting of threshold voltage via a built-in D-to-A converter to a 10mV resolution. Propagation delay is 8ns for the TTL-compatible *910* and down to 4ns in the ECL-compatible *911*. Maxim Integrated Products UK, 0734 845255.

24-bit A-to-D. Since they dissipate only 100mW maximum, the *CS53222/3* 24-bit variable-bandwidth A-to-D converter combination from Sequoia are suited to low-power logging applications, being principally meant for seismic and low-frequency passive sonar. *CS5322* is a digital finite impulse-response filter with programmable decimation and *CS5323* is the A-to-D chip for high resolution between zero and 2kHz. At a clock frequency of 512kHz, dynamic range is 123dB and distortion is –120dB. Sequoia Technology Ltd, 0734 311822.

Discrete active devices

SM active mixer. A version of H-P's double-balanced active mixers is now available in an SO-8 surface-mounting package. *IAM-820008* covers the cellular, cordless, GPS, microwave distribution spread-spectrum at 900MHz and 2.4GHz and optical-fibre system requirements. Conversion gain is up to 15dB up to 5GHz. The devices contain mixer, IF amplifier and local-oscillator buffer in the one package. Hewlett-Packard Ltd, 0344 362277.

Transistors for portables. Low-current, low-voltage transistors intended for use in battery-powered equipment, the *AT60111* and *AT60211* work up to 2GHz and to 4GHz as oscillators. The 111 has a maximum stable gain of 11.5dB biased to 3V, 400µA at 1GHz and the 211 14.5dB at 3V, 800µA. Noise figure is 1.4 (111) and 1.5dB(211). Hewlett-Packard Ltd, 0344 362277.

Schottky diodes. Two new series of schottky diodes from Siemens are usable up to 15GHz without large matching losses. *BAT 14-XXX* are medium barrier types and *BAT 15-XXX* low barrier devices, both offering a diode capacitance of 0.35pF throughout the cut-off region and a 5.5Ω forward resistance. Siemens plc, 0932 752631.

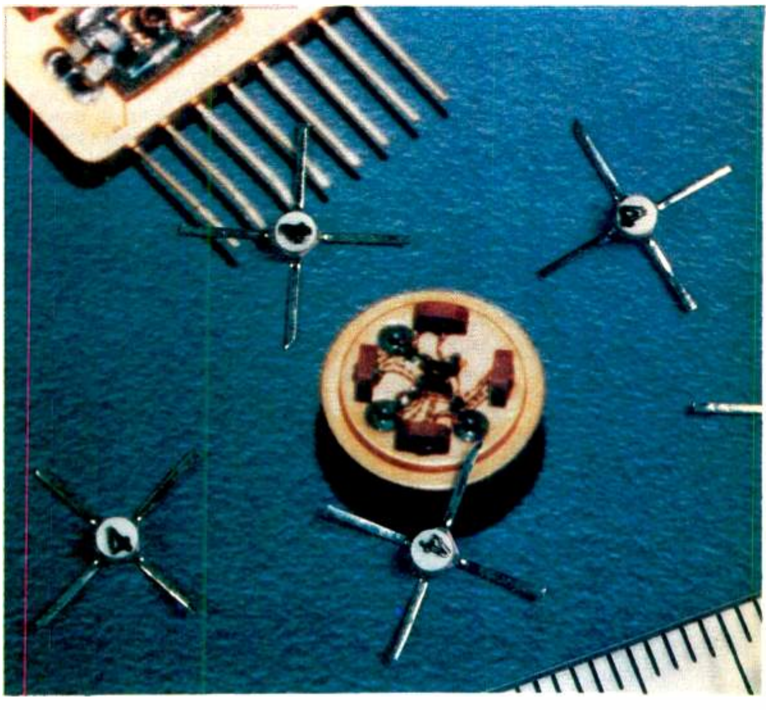
Little mosfets. Two new members of Siliconix's Little Foot family of mosfets are rated at 200V and 30V.7A. *Si9420DY*. The 200V type, is meant for low-power DC-to-DC converters, and the 30V *Si9410DY* is for disk drives and portable computers, in which it can be driven by 5V logic supplies. Siliconix Ltd, 0635 30905.

Linear integrated circuits

Tunable Butterworth filter. A 5th-order switched-capacitor Butterworth filter from Linear, the *LTC1063* allows full clock tunability with no degradation of the filter's 1mV offset and provides 12-bit DC and AC gain linearity with no clock feedthrough. Low-frequency gain accuracy is ±0.01dB. Cut-off frequency is 50Hz, THD 0.01% and wide-band noise 95µV RMS. Linear Technology (UK) Ltd, 0276 677676.

SM op-amp. *LT1128* is an ultra-low-noise op-amp giving 0.85nV per root hertz at 1kHz and is stable at a gain of +1. Maximum offset voltage is 40µV and minimum voltage gain 7,000,000. To achieve the unity-gain stability, gain/bandwidth product is reduced from the *LT1028*'s figure to 13MHz and slew rate is 5V/µs. Linear Technology (UK) Ltd, 0276 677676.

Fast comparators. *LM6685* and *LM6687* are 2.6ns propagation-delay comparators in single and dual form respectively. Analogue input is differential and the outpts are ECL-compatible, with enough output current to drive 50Ω lines. Input offset voltage is typically 0.6mV. The *LM6687* dual type has separate latches so that each comparator can be used independently in sample-and-hold mode. National Semiconductor, 0793 614141.



Low-noise transistor. *B12V114* from Bipolarics has a noise figure of less than 1.4dB at 900MHz, a transition frequency of 10GHz at 25mA and power gain of 16dB. Current range is 1mA - 60mA. Tekelec UK, 0753 548585.

Memory chips

10ns ram. Three new versions of Hitachi's 256Kbit TTL srams offer a 10ns address access time. *HM6708SH* and *HM6709SH* are 64K by 4-bit devices and the *HM67832SH* is a 32K by 8-bit type, all having minimum write-pulse widths of 8-9ns, the *6708* and *67832* offering output enable access times of 5-6ns. The 6709 has an extra output enable pin to avoid data competition when i/o pins are taken to a data bus. Hitachi Europe Ltd, 0628 585000.

Microprocessors and controllers

Risc microcontroller. AMD's *Am29205* is a lower-cost version of its Am29200 risc microcontroller for 16-bit embedded application. The device incorporates a complete set of common system peripherals and glueless interfacing to external ram and rom. It has a 29k core processor, memory controller for rom, ram or sram, a two-channel DMA controller and a video interface. Advanced Micro Devices (UK), 0483 740440.

Microcontroller. From Microchip, the *PIC16C71* is a low-cost, 8-bit controller with an on-board A-to-D converter. It is intended for use as an embedded unit and has an enhanced CPU with internal and external interrupt, 1024 by 14 eprom and 38 by 8 sram, and 14-bit-wide instructions. Four analogue inputs use one sample-and-hold, the A-to-D having a conversion time of 20µs/channel. Operating speed is 20MHz and current consumption is 1515µA at 32kHz from a 3V supply.

Please quote "Electronics World + Wireless World" when seeking further information

Arizona Microchip Technology 0628 850303.

4-bit microcontrollers. With 4K and 2Kword by 10bit of rom respectively, Hitachi's *HD404814S* and *HD404812S* microcontrollers cost less than but retain the performance of 8-bit types. They have LCD driving capability for up to 16 digits and possess 30 i/o pins, 10 of them high-current outputs. Versions with a 4MHz clock give cycle times of 1μs and lower power 800kHz types cycle at 5μs. Hitachi Europe Ltd, 0628 585000.

PC speech recognition. Two chipsets from Sierra, when built into add-on sound boards, confer speech recognition powers on a PC. Two Aria Listener chipsets, *ST8003* and *ST8004*, include the Aria synthesiser, a sound library, ADPCM, a joystick port, digital audio and playback and a Midi interface. *T8003* has a 512Kbyte sound library, while that of the *8004* is a 1Mbyte one. There is the facility to make the device recognise either one speaker's voice or any voice, the system "training" itself to cope with accents or impediments. Sierra Semiconductor Ltd, 0793 618492.

Mixed-signal ICs.
Telephone synthesisers. New PLL frequency synthesisers by Fujitsu are intended for the cordless and cellular telephone sector. For CT2, there is the *MB1509*, with two PLLs on the chip for transmit and receive and two dual-modulus prescalers. *MB1505/7* are for cellular use, at 600MHz and 2GHz respectively. Hawke Components Ltd, 0256 880800.

Switch array for video. *MT88V32AP* is video crosspoint switching up to 32 bidirectional cmos T switches in an 8 by 4 non-blocking array. Mitel's device is digitally programmable to route signals between 0 and 50MHz at 3dB. Feedthrough and crosstalk are less than -60dB at 15MHz, handling 12Vpk-pk bipolar signals with a 75Ω on resistance. Mitel Semiconductor, 0291 430000.

Motor controller. Philips's *TDA5145* is an integrated driver for three-phase brushless DC motors, giving 1.8A drive and needing no Hall-effect sensor, since rotor position is determined from the motor's back EMF. It generates full-wave output for forward and reverse drive and has motor braking, output protection and start-up circuitry. No-load current is 6.8mA. A square-wave output at half the commutation frequency is usable by a microprocessor for speed control. Philips Semiconductors Ltd, 071 436 4144.

Optical devices

Higher CCD sensitivity.
ICX039BNA-6 is a 0.5in interline transfer CCD image scanner by Sony for use in pal cameras and has twice the sensitivity of the earlier *ICX039AKA-6*. It offers built-in yellow, cyan, magenta and green complementary mosaic filters, a variable-speed shutter and the hole-accumulation diode for low smear and blooming. Sony Components Ltd, 0784 466660.

Fast IGBTs. Six new high-voltage insulated-gate bipolar transistors from Harris have built-in ultra-fast recovery diodes to improve switching. Breakdown voltages range from 400V to 600V and collector currents from 6A to 24A. Reverse recovery time at the device's rated current is 60ns. A sample and data pack is available for evaluation Harris Semiconductor UK, 0276 686886.

Red laser diodes. Three types of red-light visible laser diodes by Toshiba are meant for use in optical recording, bar-code readers and measuring sensors. *TOLD9150(S)* puts out 30mW at 690nm, *TOLD9412(S)* operates at 650nm at up to 50°C and *TOLD9111(S)* at up to 60°C at 680nm. Toshiba Electronics (UK) Ltd, 0276 694600.

Power semiconductors

Power opto-isolator. New from Motorola is the *MOC2A40-10* opto-isolator, which is a 2A 140V RMS, zero-crossing triac that is drivable by TTL logic levels and needs no heat sink. Input/output isolation is 3750V RMS and it will withstand a 60A single-cycle surge. Motorola Semiconductor Ltd, 0296 395252.

12A mosfet. Zetex has the *ZVN4310A*, a 12A buffer to interface logic levels to loads such as lamps or inductive components, its on resistance being 0.36Ω on a 10V gate drive. Breakdown is 100V, maximum dissipation 850mW and continuously conducts 0.9A (1.13W and 1A on a 1sq.in heat sink. It turns on and off after 8 and 30ns delays at 25V and 3A with rise and fall times of 25 and 16ns. Zetex plc, 061 627 4963.

Low-noise HEMT. 2SK1977 is a new member of Hitachi's range of high-electron-mobility transistor low-noise amplifiers for use in broadcast satellite converters. Noise level is 0.55dB and power gain 11.5dB at 12GHz. Hitachi Europe Ltd, 0628 585000.

PASSIVE

Passive components

SM inductors. Cambion's 0.1-1000μH range of surface-mountng inductors now has *BS9752-F0001* approval. These are fixed RF inductors in the E12 series to BS2488 1966 (IEC63). Dimensions are 3.2mm by 4mm and either 2.4mm or 3.4mm in height. Interconnection Products Ltd, 0433 21555.

Surge protection. Murata's *DVZ07/10* surge protectors handle 140V or 275V AC and 180V or 350V DC, depending on the model. Maximum clamp voltage is 710V at 25A and maximum withstand current 1250A. Capacitance values lie between 200pF and 700pF. Murata Electronics (UK) Ltd, 0252 811666.

SM electrolytics. This is a range of surface-mounted aluminium electrolytic capacitors capable of working at temperatures between –55°C and 105°C. The WT capacitors have a range of 0.1μF to 100μF ±20%, at 4V to 50V and with 3μA leakage current. Nichicon (Europe) Ltd, 0276 685393.

SM trimmer capacitors. Philips's SMD811 surface-mounting dielectric-film trimmer is intended for impedance matching and tuning, having low energy loss and a resonant frequency between 400MHz and 1GHz. It comes in four versions with maximum C of 5, 10, 15 and 18pF, with a tan of 0.001 at maximum capacitance and 1MHz. Philips Semiconductors, 071 436 4144.

Displays

Bright led. Belling Lee Dialight has the Series *557* panel lamp which is said to combine the brightness of an incandescent lamp with the reliability of a led. It uses a multiple led construction and a lens to give a luminance of up to 1000 foot-lambert and a lifetime greater than 100,000h. It comes in red, green or yellow and in a red/green bicolour. BLP Components Ltd, 0638 665161.

Hardware

SM prototyping. Prototype boards that allow the use of both surface-mounted and through-hole components are available from BICC-Vero. SMD adaptors cater for 28, 44, 68 and 84-pin gull-wing or J-lead quad packages on the upper surface. Or, the VSM pins press fit into 1mm dia holes in the PCB, suitable for both gull-wing and J-lead packages. BICC-Vero Electronics Ltd, 0489 780078.

Instrumentation

Radio test set. Marconi's new radio test set includes a full-span, high dynamic range spectrum analyser, an accurate power meter, a storage oscilloscope, a fast -switching RF generator, an audio analyser and three independent audio generators. There is also a built-in multimeter. A typical test of of transmitter frequency, power and deviation and a measurement of receiver audio level and *SINAD* takes less than three seconds. Control is by IEEE 488.2 and RS-232 and the instrument's own Basic interpreter. Marconi Instruments Ltd, 0727 59292

VSWR power meter. For use with GSM phones, the *NAS-Z6* insertion unit is coupled to the NAS directional power meter, made by Rohde & Schwarz. Since the signal bursts encountered on GSM systems render conventional methods of power measurement up to 100% inaccurate, the NAS-Z6 has a signal-controlled circuit to remove transients so that forward and reflected power and VSWR can be correctly measured. PEP is measured, rather than one of a number of average measurements, as is usually the case. Five models are produced, for 1-1000MHz working and handling up to 1200W. Rohde & Schwarz UK Ltd, 0252 811377.

RS-232 instrument control. ARC is a system enabling a number of measuring instruments to be controlled via RS-232 from a PC. It needs no special cabling, software or hardware — not even an extension card. The system allocates an unique address to each of up to 32 instruments. Thurlby-Thandar already produces several instruments incorporating the ARC interface, which is effectively a lower-cost alternative to GPIB control at a lower data rate. Thurlby-Thandar Ltd, 0480 412451.

Interfaces

Data logger. Orion's Tinytalk is a little self-contained data logger which is to be made for a number of functions, the first one, Tinytalk-Temp, being for temperature logging. Several ranges are available, the widest being -39°C to 123°C, with a resolution of 0.35°C. Non-volatile memory stores 1800 readings at intervals from 0.5s to 4.8h. Setup and down-loading is by PC, the output being a plot or spread-sheet data in Excel and Lotus 1-2-3. Orion Components Ltd, 0243 778088.

Digital oscilloscope. LeCroy's Model *9304* digital storage oscilloscope has four flash A-to-D converters, one for each channel, which digitise one-shot signals at 100Msample/s or repetitive waveforms at 4Gsample/s. It also has 10k of memory. Each channel is also provided with its own FFT display. LeCroy Ltd, 0235 533114.

Literature

DSP handbook. In a 286-page book, GEC Plessey describes its large range of digital signal-processing ICs, including arithmetic and logic units, algorithm-specific ICs, image processing, filtering and freuency-domain processing. Gothic Crellon Ltd, 0734 788878.

Maplin 1993. Maplin's enormous 1993 catalogue is now with us, containing in over 700 pages many new arrivals, including 10 pages of computer equipment and accessories and expanded tools, protection and test gear sections. £50 worth of vouchers are included. The catalogue is available from Maplin shops, W H Smith or by mail-order at £3.45. Maplin Electronics plc, 0702 554161.

Microwave components. Krytar microwave components cover the 0-50GHz range and are described in the new catalogue, freely available. Included are directional detectors and couplers, 3dB 90° hybrids, power dividers, zero bias Schottky detectors and coax. terminations. Tony Chapman Electronics Ltd, 0992 578231.

Power supplies

Switched-mode power supply. An open-frame power supply from BICC-Vero exceeds the requirements of IEC601, with regard to ground leakage current in medical and dental equipment, maximum being 100μA at 233V AC, 50Hz. Modules in the *BVS/BVM40* range also meet RFI requirements and give a number of output voltage combinations from 5V to 24V. BICC-Vero Electronics Ltd, 0489 780078.

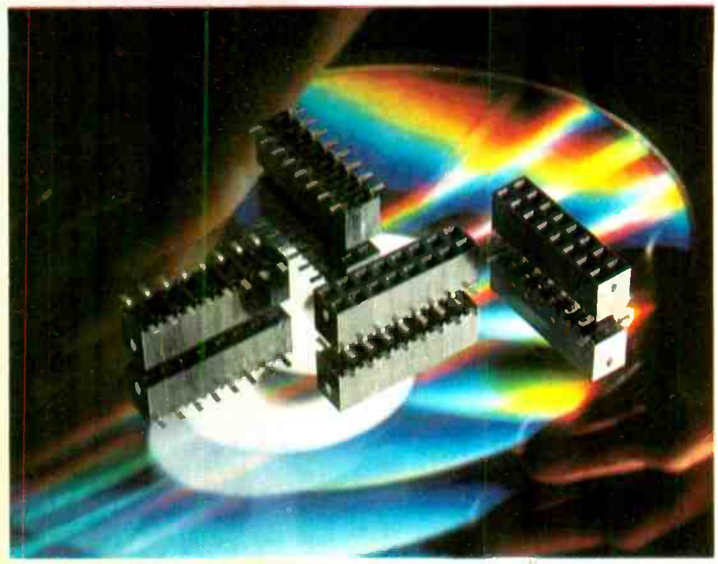
High-current power supply. Trio-Kenwood's *PD18-20* bench power supply provides 0-18V at 0-20A, with a ten-turn pot. for control. A phase control circuit with a built-in preregulator gives a fast response, a choke-input filter providing good regulation and low ripple and noise. Remote sensing is used and the output can be set remotely. The output is protected. Trio-Kenwood UK Ltd, 0923 816444.

Radio communications products

RF mixer. Atlantic claims its *MT45* double-balanced RF mixer to be the smallest eleven-octave, discrete device available. RF and local-oscillator range is 0.5MHz to 1000MHz, IF being from zero to 1000MHz. Local-oscillator drive range is 3-13dB, conversion loss 6dB at mid range and oscillator/input isolation 35dB. It is packaged in a TO-5 can and can handle temperatures from -54deg. C to 100deg. C. Atlantic Microwave Ltd, 0376 550220.

Switches and relays

High-frequency relay. *RG* relays by Matsushita offer an isolation of 65dB and an insertion loss of 1dB maximum at 900MHz. They are available in monostable or latching form with a characteristic impedance of 50Ω or 75Ω and the gold-plated contacts are rated at 24W at up to 24V, with an initial contact resistance of 100mΩ. Matsushita Automation Ltd, 0908 231555.



PCB connector. A dual-row inverse connector from Methode with an off-board height of 7.1mm is designed for surface mounting and is available with vertical holes or horizontal holes, both types having between 2 and 40 positions spaced at 0.1in. Contacts are either tin/lead or selectively gold plated and handle up to 3A. Methode Electronics Europe, 0535 603282.

Transducers and sensors

Displacement sensors. Fast linear displacement transducers from Control Transducers now come with ranges from 304mm to 609mm, with maximum errors of 0.15% or 0.1% as an option. Temperature range is

Magnetic sensor. Intended for position sensing in printers, motors, RPM measurement and other devices, the *SS400* series by Honeywell are temperature-compensated, Hall-effect devices measuring 4 by 3 by 1.6mm. Operating conditions are 3.8-24DC supply voltage and 10mA and bipolar, unipolar and latching operation is available. Honeywell Ltd, 0344 424555.

improved to -50°C to 125°C. Signal processing is carried out by the SP50 board, which allows measurement to 4kHz, the SP100 going up to 15kHz. Control Transducers, 0234 217704.

Hall-effect sensors. Optek Hallogic range of sensors include on-chip logic functions and are chiefly intended for use in encoding and speed monitoring in tachometers and DC motor control. Temperature range is -55°C to 150°C. Highland Electronics Ltd, 0444 236000.

COMPUTER

Computer board level products

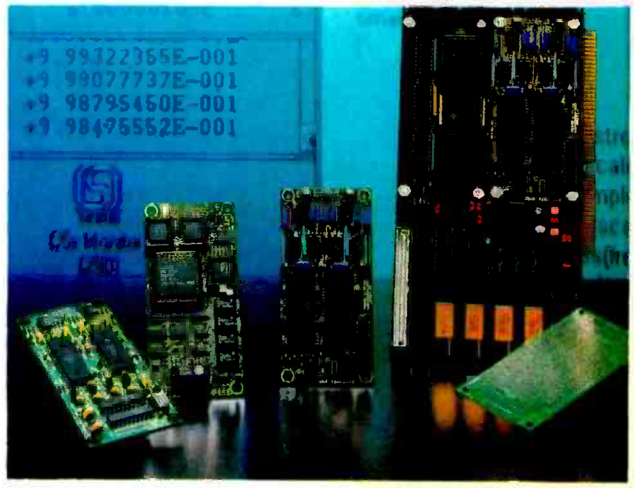
Thermocouple board. Amplicon's *PC73A* board amplifies and digitises thermocouple signals for the logging and display of eight temperatures per board. Boards are linked by cable so that thermocouples can be easily changed. Gain is set to 1, 100, 200 or 500. Supplied software requires only basic knowledge of the board's operation, C source code being provided. Amplicon Liveline Ltd, (Free)0800 525 335.

Development and evaluation

Neural net development. *SigNet*, claimed to be the world's first development system for real-time neural networks, is introduced by Data Beta. *SigNet* is a combination of neural software and DSP boards for PC and VME buses and will carry out processor-intensive tasks not feasible for traditional algorithmic processes, the user not needing any knowledge of computer programming. Expansion cards for the PC have a processing power of up to 200Mflops and include 16-bit A-to-D-to-A for capture and output of analogue data. Data Beta, 0734 758222.

Software

PC frequency management. Harris's *RF-7790* is a program to select optimum frequencies for HF communications. It analyses HF propagation conditions using an ionospheric prediction model, a ground-wave prediction model, real-time oblique sounding and real-time channel occupancy measurement. It can be adapted for use in fixed and mobile land-based systems or in marine work. Harris Corporation, 0101 (716) 244 5830.



Applications board: *PC/C31* combines the 32bit real-time processing power of TI's TMS320C31 with flexible i/o options and high development support. The $^{2}/_{3}$ length PC AT card is accompanied by a package of development and integrations software. Loughborough Sound Images Ltd. 0509 231843.

Art attack

It is true that other countries are making products that we in the UK seem to have forgotten how to make, and making money, which we also seem to have forgotten how to do ("Paying for the Arts", November *EW + WW*). But to blame "unproductive" disciplines such as art, English and history is short-sighted.

Science and Engineering came about because people learned to look at the world around them, separate and categorise the phenomena, construct theories, experiment, and build useful things. The fact that these useful things are often weapons of destruction, either directly or indirectly, just shows the minute difference in evolutionary terms between the human mind and that of a shark. Art, story, history all have a role in forming an engineer's ability to *see* the world, both in form and function. These are skills which no design package I have come across has dared to claim: it is called being creative.

Education is for training minds to interpret the world. It should not be a sizing up for a set of blinkers or rosy spectacles.

Malcolm Bloor
Stafford

Your comment that government under-investment in education and scientific research is a contributor to the present state of the UK economy is certainly valid, and is of concern to all educationalists and industrialists alike.

But your suggestion that the present malaise might be cured by the reallocation of research and education funds away from the arts and into scientific studies is nothing short of scandalous.

Certainly it is true that German and Japanese economies are powerful and have enjoyed more or less continuous expansion since the end of World War II. It is also true that the manufacturing industries of these two nations produce desirable consumer products. But reasons for this are historic and certainly not as a result of German or Japanese students being forced to study engineering subjects in their universities at the expense of the arts.

Following the war, Germany and Japan were forbidden to manufacture military hardware in mass quantities or to engage in any kind of military expansionism. The long term result has been that their economies have flourished and now produce useful well-designed consumer products, unlike the economies of the UK and US where some 40% of all graduate physicists, mathematicians, computer scientists and engineers still work exclusively in the weapons industry.

At a time when 85% of German teenagers go into further education to study the subject of their choice in well funded FE colleges and universities, the government in this country is continuing to decrease investment in industry and education to such an extent that a large proportion of our top graduates are being forced to emigrate to find work.

A decline in the manufacturing requirement of the defence sector coupled with a government which believes that market forces are all-powerful has directly contributed to the depth and duration of the current recession.

An alternative strategy to that suggested in your comment, to aid recovery, is that as a nation we must attach more value to education and insist on government investment in education – especially at FE and university level – while allowing our students to choose whatever subjects they wish to study.

The wealth of the nation cannot be measured simply in economic terms but is embodied by a balanced education in the humanities as well as the sciences.

Mathew Steinberg
New Cross
London

Two questions dominate the debate over the importance of manufacturing to the UK: why don't people believe that the ability to make widgets is the ability to generate wealth? ; and, how are we to convert Britain's lethargic and moribund manufacturing organisations into the pro-active and empowered organisations of our Continental Europeans and Far Eastern competitors?

I am not suggesting that all UK companies are moribund. Far from it. The UK has some of the best factories in the world. A Japanese manufacturing system professor told me just a few months ago that he was sending some of his students to the UK to learn what "best practice manufacture" means! It seems that Sony's most productive factory is in fact in the United Kingdom! So it can be done. Buy why don't more of us achieve it?"

I too would like to see copies of "Paying for the Arts" sent to MPs.

But I would also like to see our professional institutions taking a look sideways. Not only must engineers have multidisciplinary skills, we must also foster interdisciplinary skills.

Other *EW + WW* readers who have views on this matter should make their opinions known. Do not feel as though "I can do nothing about it" because we can. Every little helps. Manufacturing is the cornerstone of our economy and we must work to ensure its survival.

Andrew Ainger
Human Centred Systems
Hertfordshire

You are correct in assuming that this country needs to invest in its infrastructure: ie more investment in science and technology in education, industry and business. But you are incorrect in assuming that because the sciences are beleaguered, the arts receive great piles of cash and an unfair leg-up concerning job prospects compared with the training and research grants available to science subjects at universities. What is your evidence for this view?

It is an outdated notion that art and science should be forever consigned to opposite sides of the debate on issues concerning economic growth, culture and the future of the global village. To provide for ourselves and create stability in Europe in the next century, shouldn't we really be thinking in terms of the world rather than just Britain in this context?

At a time when new developments in IT and spin-offs from the more esoteric sciences (eg "anti-chaos theory") appear to present fantastic new options and opportunities for learning, it is hard to understand why anyone should advocate such a narrow view of our collective future. We are amazed that in your editorial you resort to art-bashing and advocate some sort of "Thatcherite style assessment" of the situation (your words). Can you really be suggesting that the corridors of power are patrolled exclusively by bands of ex art history students?

Yes, it is vital that the UK government develops a longer term strategy for British industry, but surely we must also work as hard as we can to bring about a longer term strategy or greater understanding between everyone working in both the arts and sciences. Production, exploration, conservation, education, development and cultural awareness – all these qualities have their place in both art and science and we should be working towards their equal balance and endeavouring to identify the real enemy of our present difficulties. In this context, would it not be more helpful if arts and sciences were encouraged to interact *more* rather than *less* at every level?

A great number of inventions would be unmarketable, were it not for their artistic design content – and quite rightly so. The main reason we are surrounded by such ugliness is because too many people don't care how things look. A more balanced arts and sciences education might also assist those readers tempted to send your letter to their MPs to write their own.

Please spare us any more "Comments" similar to the one we have just read. Next time, couldn't we have something a little more constructive and more worthy of your record of interesting debate and open-mindedness?

Clifford Williams *(for the Sciences)*
Linda Williams *(for the Arts)*
- happily coexisting despite a mixed marriage.
Lancashire

The UK grossly undervalues science and its industrial application. The Whitehall villagers and their visitors prize a classics education more highly than a science based one. Until the UK changes its priorities and perceptions, we shall remain well and truly dans la merde. I make no apologies for my comment.

Frank Ogden
Editor.

Speakers corner

PC Meunier (Letters, EW + WW, November) appears to be unaware that there are several types of DC motor which differ according to how the magnetic field, in which the armature revolves, is produced.

These fall into three main categories:

- Permanent magnet motor: the field is produced by a permanent magnet and is the type to which the normal permanent magnet loudspeaker is closely related.
- Series motor: the field is produced by an electromagnet whose winding is connected in series with the armature winding.
- Shunt motor: the field is produced by an electromagnet whose winding is placed in shunt with the armature winding.

There are also combinations of these, such as the compound motor where the field is produced by an electromagnet with two windings, one connected in series and one in shunt with the armature winding.

If we consider just the three main types, we note that for the permanent magnet motor, the magnetic flux is quite independent of either the current flowing through the motor or the voltage applied across it. For the series motor, the magnetic flux depends on the current through the motor. If it were not for the non-linear characteristics of iron, the flux would be proportional to the current, and quite independent of the voltage across the armature winding. In the shunt motor, the flux depends on the voltage applied across the motor. Again, if it were not for the non-linear characteristics of iron, the flux in this case, would be proportional to the applied voltage and be independent of the current through the armature winding.

Connecting two motors of the same type in series gives two different results according to the type of motor. Note that when two motors (or two loudspeakers) are connected in series, the currents through each are identical, whereas the voltages across the two may differ.

In the permanent magnet motor, or permanent magnet loudspeaker, since the magnetic fields are produced by permanent magnets and the same current flows through each device, the torques of the two motors and the forces on the voice coils of the two loudspeakers will be very nearly equal – assuming the corresponding devices are nominally of the same design. The torques and forces will differ only due to manufacturing tolerances so that in the case of the two loudspeakers, their acoustical outputs will be very nearly equal. There will be no untoward effects, as suggested by Meunier. Assuming that the two motors or two loudspeakers feed into equal linear loads, half the applied voltage will be developed across each device and the current will be half of what would flow if the same voltage were applied to but one device. Each motor would revolve at approximately half the speed that would result if all the supply voltage were applied across it.

The case of the series motor is somewhat similar. Since the same current flows through each motor and also flows through each field winding, the magnetic fluxes in each motor must be the same, within the limits of manufacturing tolerances, and so the torques must be very nearly equal. With equal mechanical loads, the speeds will then also be equal, which, in turn makes the back EMF in each motor the same, and each motor will have half the supply voltage across it. Magnetic flux depends on the current through the motor, so there is, in this case, a non-linear relationship between the speed of the motor and the applied voltage and the speeds are found to be faster than half those obtained if all the voltage were applied to one motor only.

The case of the shunt motor is very different, and is that discussed by Meunier. Suppose we have managed to start the two motors running at equal speeds with half the supply voltage dropped across each. If we increase the mechanical load on one, as Meunier suggests, it will slow down. The back EMF produced in this motor, being proportional to its speed and magnetic flux, will fall and so the voltage across the machine likewise falls. But this voltage is also that across the field winding, so that, as the voltage falls, so must the current through the winding. Magnetic flux depends on this current, so it, too, must fall, in turn reducing the back EMF still further – and so on. The result is a positive feedback system, and stability will only be reached when the motor with the additional load stops rotating, with essentially no voltage across it.

At the same time, the voltage across the other motor will increase, since it must be the difference between the supply voltage and that across the first motor. This will both make it revolve faster and also increase its magnetic flux. Stability will be reached when nearly all the supply voltage is across this second motor, and we have in effect an electro mechanical flip-flop, as Mr. Meunier describes.

The important point is that loudspeakers are not equivalent to shunt motors and do not behave like them. As explained above, for loudspeakers in series, the voltages across them will be equal, within manufacturing tolerances, if the loudspeakers are of the same design.

But the general question which Meunier has raised, regarding the desirability of placing loudspeakers in series, requires further consideration of a rather different nature.

If the two loudspeakers are exactly equal in all respects, then the applied voltage would divide equally between them and their acoustic outputs would be the same. They would be equivalent to a single loudspeaker of the same type wound with more turns on the voice coil, so as to double the impedance. (There might be some small difference due to the acoustic loads on the two loudspeakers mutually affecting each other. If the two loudspeakers were placed very close together then they would be equivalent to a larger loudspeaker with twice the radiating area, which would have a different value of acoustic load, particularly at low frequencies.) In practice, the two loudspeakers are bound to differ slightly and we need to consider what happens under these circumstances.

It is well known that to obtain the best reproduction through loudspeakers, they must be fed from amplifiers of very low output impedance so as to damp, as far as possible, any of the resonances in the loudspeaker system. The main one of these is the bass resonance of the loudspeaker, which may be split into two or more other resonances if the loudspeaker is mounted in a bass reflex cabinet, or similar acoustic device. At low frequencies, away from the bass resonance, the loudspeaker impedance is very nearly equal to the resistance of its voice coil. Thus, when two loudspeakers with different bass resonant frequencies are connected in series, at the bass resonance of one, it will be fed effectively from a low output resistance amplifier in series with the resistance of the other loudspeaker. It will no longer be damped.

As the bass resonant frequencies of the two loudspeakers get closer together, their (complex) impedances will react with each other so as gradually to merge into a single resonance. The two loudspeakers will then appear as a single one with twice the impedance of either, and good damping will occur when fed from a low output impedance amplifier. So, if the two loudspeakers are closely matched, the damping will be very nearly the same as for a single one. But if they are badly matched, the damping will be considerably less, with noticeable deterioration in the acoustic reproduction. We see then, as Meunier asserts, that it is desirable to connect loudspeakers in parallel, rather than in series, so that, whether they are matched or not, the low output impedance of the driving amplifier is always effectively in parallel with each, giving good damping.

But a word of caution: if transformers are used to adjust the combined load presented by the speakers to that required by the amplifier, there is little point in

Be-low frequency

ELF and VLF radio bands have been largely neglected by amateur radio and electronic enthusiasts. This is surprising in view of the fascinating properties of this portion of the electromagnetic spectrum.

Practical uses of frequencies below 150kHz include broadcasting standard frequency and standard time services, hyperbolic navigation, and communication through sea water or solid rock to submarines, miners and pot-holers.

With negotiations between the RSGB and the licensing authorities for an amateur VLF allocation well underway, the neglect of the bottom end of the radio spectrum could be coming to an end.

One British group, the Cave Radio and Electronics Group of the British Cave Research Association, is intent on improving techniques of VLF radio communication with a particular emphasis on its use in cave surveying and cave rescue.

Anyone wanting more information can send for a copy of our journal that contains a broad mix of practical and theoretical articles. Recent articles have covered the principles of inductive communication, antenna design, modulation methods, a design for an ultra-sensitive flash trigger, cave-proofing equipment, NiCd battery charging, surveying software and cave detection using geophysical techniques.

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using this type of connection if the transformer introduces more degradation in performance than the reduction of damping due to a series connection. If the performance is not to be materially degraded, the transformer must have a very good frequency response; must not introduce non-linearity distortion, and it must be very efficient.

Transformers meeting these requirements are inevitably bulky and expensive. If cheap, poor quality transformers are used, the results would probably be much worse than if the loudspeakers were connected in series. It should be noted that since the transformer is connected to the output of the amplifier, it will not be within the negative feedback loop of the latter, and so will not have its performance significantly improved by the feedback.

Roger Chapman.
Bexley
Kent

P C Meunier (Letters, *EW + WW*, November) correctly warns that errors can arise from wiring two or more loudspeakers in series. But his proposed experiment does not prove his point.

He attempts to extrapolate from the observed behaviour of two unloaded motors wired in series to the speculative behaviour of two or more loaded loudspeakers wired in series. A speaker without a load might exist in a vacuum, but in practical speaker designs where speakers are wired in series, or in complex series and parallel networks, the loading on each speaker driver includes the air pressure on the surfaces of the cone, and the mechanical resistances of the cone's suspension. These loads are virtually identical for each speaker in the system, and are not analogous to unloaded motors.

A speaker designer will attempt to obtain equal efficiency from each driver in the network, and in so doing will be distributing the air-load equally amongst the drivers by coupling each driver to a similar load on both the front and rear surfaces of its cone. The internal surface includes the path to the enclosure's port, so the designer will try to position the port equidistantly from each cone's effective centre.

If Meunier repeats his experiment with identical loads on each of the two motors wired in series, or even just approximately similar loads, he will find that both motors turn. The electrical impedance of each device (whether it be a motor or speaker) is much more than the unloaded resistances and inertia of the driven coil or magnet: it is the electrical impedance presented to that driven element, which in turn is the mechanical impedance presented to the moving parts. The air resistance "seen" by both speaker drivers, wired in series in an enclosure, will tend to ensure that one cone moving further will do so against the greatest mechanical impedance, and therefore the greatest electrical impedance, consequently enabling the other driver to move further. The upshot is a stable mechanism in which speaker drivers are self regulating in their ability to share the load. Even ill-matched speakers and enclosures will share their load to a considerable extent.

There are other problems in wiring speakers in series. Quite simply, the amplifier, for all its negative feedback, is only able to ensure that the voltage output is as accurate as its makers demanded, and as long as that output is connected directly to a single cone driver, or parallel cone drivers, then we can reasonably expect that the driver will perform as designed. But if any additional impedance is placed in series with the amplifier's output, whether it be a passive crossover network, excessive cable resistance, the light bulb as found in Bose 802 speakers, or simply another speaker placed in series, then the amplifier can no longer be expected to be in control of the voltage and current reaching the speaker coil.

This is where Meunier's concern for unequal damping is valid.

Any resonance produced by the combination of the amplifier's Zobel network (placed outside the feedback loop); the highly reactive components in the passive crossover; all of the speaker cable's impedance and currents induced from adjacent wiring; and the natural resonance of the speaker's cone, will allow the cone to resonate at a high impedance frequency. The electrical "error" signal thus generated by that resonating cone will not be detected by the amplifier's feedback input, but will be presented to the other driver (wired in series) as part of its input signal.

It is possible in asymmetrical designs that one driver will resonate in one phase, and another driver in the opposite phase. But the air impedance would have to be quite dissimilar for each driver, and I guess that this is only going to arise if the speakers are placed in separate enclosures, possibly at opposing ends of a room facing each other!

Bigger problems occur in such speakers as the Bose unit mentioned, not because of the fact that its eight drivers are connected in a parallel/series arrangement, but because of the light bulb wired in series to the whole lot. This deliberately increases the resistance between the output from the amplifier's feedback loop, and the chain of drivers so that any resonance can, and will, develop within the cabinet and remain poorly corrected by the amplifier. The error is not a result of the series wiring of the drivers. Similar errors of course arise with passive crossover networks. Poor component selection is common, with disregard for the often horrendously wide tolerances admitted by manufacturers of high voltage non-polarised capacitors, such that an amplifier may drive impedance peaks of double the nominal value at certain frequencies. The crossover point at the lower end of a driver's designed frequency range is close to the frequency where its moving mass is most resonant.

Meunier advocates the use of transformers to distribute the loads across the network. Problems occur in all non-resistive loads where the voltage at the amplifier's output is not in direct proportion to the current being delivered. The amplifier's negative feedback input, its inverting input, is unable to compensate for any errors in the output current. This is particularly true of that popular method of connecting many speakers using "100V line" transformers. The method is often employed in announcement systems in large buildings, in which the speakers are sharing dissimilar loads, sometimes in series parallel networks, and driven by amplifiers which have to provide maximum current when the output voltage is near zero. While such set-ups have surely disappeared from most music systems of any power and quality, it is astonishing that they behave as well as they do, thanks to the virtually constant air pressure on each cone.

Finally, it may be worth examining the reason for connecting speakers in series. If they are in the same enclosure as discussed, then it will be to create an approximation of a speaker with a pistonic diameter much greater than that of the drivers available, with the high frequency performance of small low-mass cones. Such systems overcome the need for different drivers for different parts of the frequency range, and therefore should not need passive crossovers in series with the drivers. The theory works well in practice, even with the addition of a light bulb between the amplifier and the drivers.

Some of the problems I have mentioned are eliminated by complex systems, in which the actual movement of the driver is monitored by the control electronics, or at least the actual output of the amplifier is monitored, notably systems by Meyer Sound of California and Nexo of Paris. Otherwise, it is quite acceptable to connect identical speakers in series as long as they are in the same or adjacent enclosure, and there is no significant impedance between amplifier and series driver.

It remains an anachronism of the electronic industry that audio reproduction involves the use of such crude devices as paper loudspeaker cones glued onto bent coil formers which are suspended by cloth in a large magnet – all of which are outside any servo control or other feedback loop. In any other device today, we would expect position detection to control the amplifier's output (as once produced in a Philips active loudspeaker), if not a transducer actually monitoring the sound near the surface of the transducer.

Dave Cross
Newcastle

Read shift

I was delighted to read the letter from AJ Quinton of Australia (Letters, August) which raised the incompatibility of a constant speed of light and the Doppler effect. This is something I have questioned myself and though I have asked some quite bright people to explain how the two ideas can be made compatible, none has been able to answer me.

Quinton goes on to say that he still thinks in terms of particles rather than waves and I sympathise with him. No one seems to be able (or even willing) to try to answer the question -Waves of what?

After all, waves must be composed of something - a "wave" merely describes the motion of the particles which make up the body which exhibits wave motion.

Perhaps some of your more learned readers will enlighten us. After all, the expanding universe is so described because of the red shift, which is the Doppler effect; is this really inconsistent with a constant speed of light ?

Martin W Berner
Trinidad
The West Indies

Circuits, Systems & Standards

First published in the US magazine EDN and edited here by Ian Hickman

Two-way amplifier uses few parts

Two-way amplifier for two-wire line
A conventional amplifier for a two-wire line uses two-wire to four-wire conversions using hybrids, with a separate amplifier for each path direction. In applications where the return loss of the ends is large, a simpler arrangement is possible, shown here implemented for unbalanced two-wire lines. It can be extended to balanced lines if required, by duplicating the whole circuit for the other wire.
IH

A conventional two-way amplifier used with two-wire telephone lines (**Fig. 1**) requires separate amplifiers for each direction, hybrid two- to four-line transformers, and carefully designed complex-impedance terminations to match the capacitive reactance normally associated with telephone lines.

For two-wire signal lines whose impedance is mostly resistive (telecommunications test equipment, for example), the simpler, resistively terminated circuit of **Fig. 2** can be used. The boxes labelled "termination" are two telecomm devices (such as modems, modem test equipment, short-haul two-wire communication lines, or minimum-loss impedance-matching pads) that are communicating with each other.

Like all two-way amplifiers, both of these circuits may oscillate unless terminated with the proper impedance. Moreover, higher audio gain imposes greater accuracy constraints on the terminating impedance values. Because resistors R_A, R_B, R_C, and R_D have a fixed relationship with one another, the circuit's "resistance level" can be set by choosing a value for one of them – R_D – for example. Given the desired termination resistance R_L and amplifier gain A, values can then be calculated for the remaining resistors R_A, R_B and R_C.

First, note that R_A contributes the only significant resistance between each amplifier and its load. Source impedance should equal the load impedance, so

$$R_A=R_L \qquad (1)$$

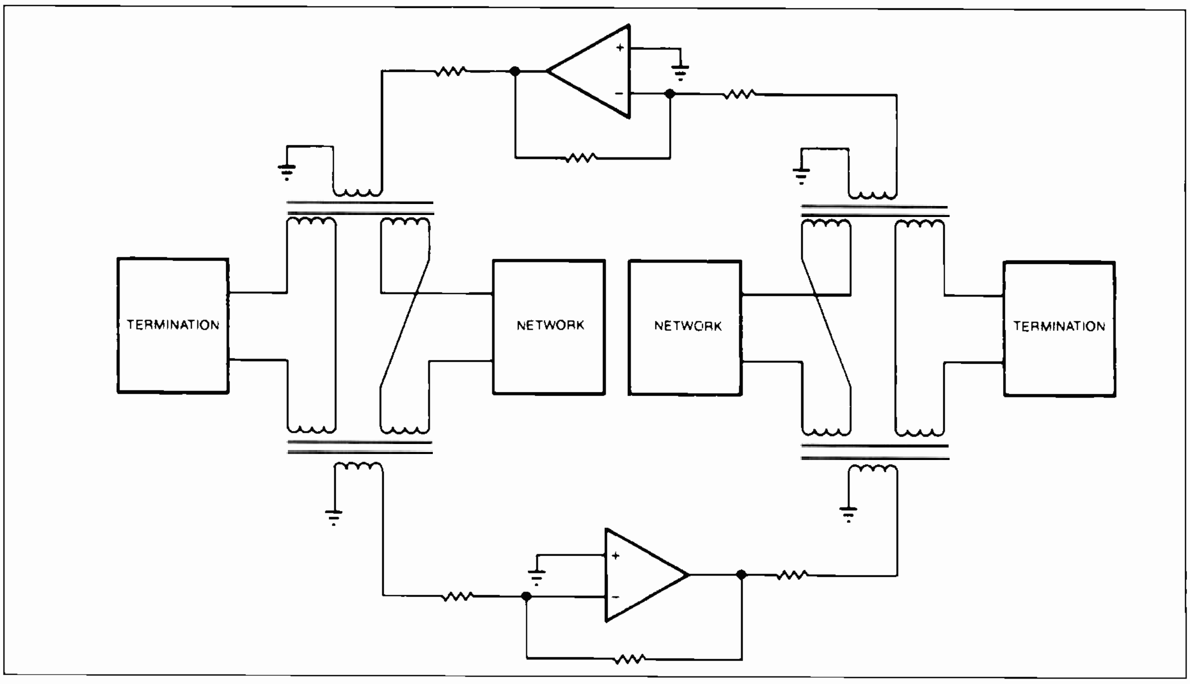
is the first of three design equations. Because IC_2 must not contribute a signal while IC_1 is operating (and vice versa), IC_2's output must remain at virtual ground during that time. Therefore, R_A is also the input impedance for each amplifier.

Next, the gain from IC_1's non-inverting input to its output is

$$\frac{R_D + \dfrac{R_C R_B}{R_C + R_B}}{\dfrac{R_C R_B}{R_C + R_B}}$$

and the gain from IC_1's output to R_L is $^1/_2$ because R_A=R_L

Fig. 1. Schematic showing a typical two-way amplifier used with telephone lines.



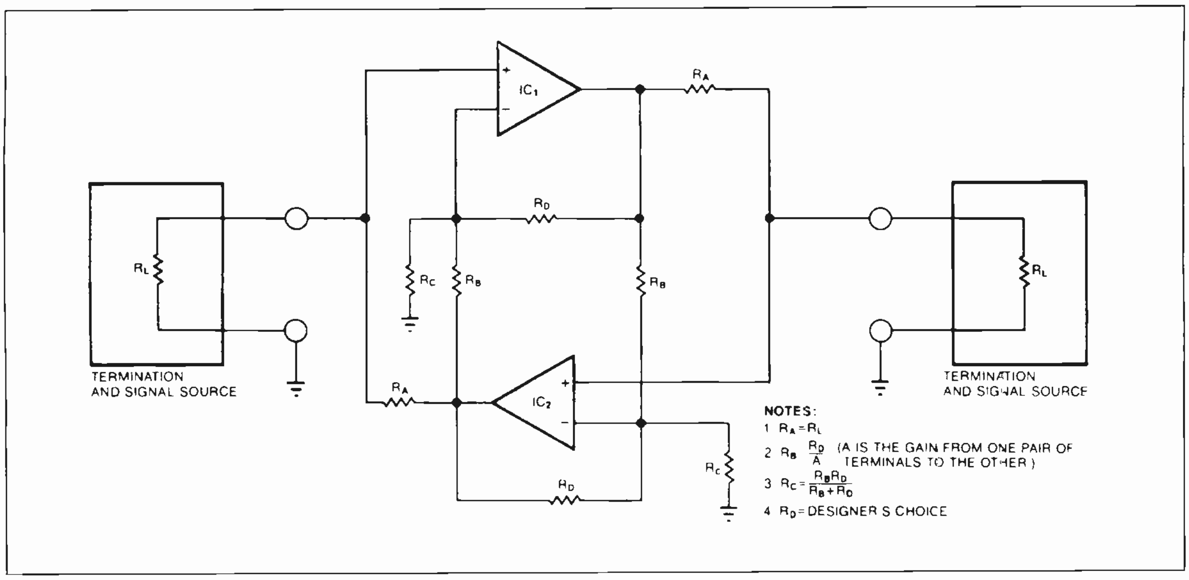


Fig. 2. Simpler two-way amplifier suitable for lines whose impedance is primarily resistive.

(Eq 1). The product of these gains must equal the given gain A:

$$A = \frac{1}{2}\left[\frac{R_D + \dfrac{R_C R_B}{R_C + R_B}}{\dfrac{R_C R_B}{R_C + R_B}}\right] \qquad (2)$$

To ensure that no signal is emitted from IC_2 when only the left hand signal source is active, IC_2's differential input must be zero. This is so if IC_2's inputs are equal, which implies equal voltage dividers:

$$\frac{\dfrac{R_C R_D}{R_C + R_D}}{R_B + \dfrac{R_C R_D}{R_C + R_D}} = \frac{R_L}{R_A + R_L} = \frac{1}{2}$$

Simplifying this equation:

$$\frac{R_C R_D}{R_B R_C + R_B R_D + R_C R_D} = \frac{1}{2} \qquad (3)$$

Eqs 2 and 3 can be solved simultaneously for the remaining unknowns R_B and R_C. First, simplify Eq 2:

$$A = \frac{1}{2}\left(\frac{R_D R_C + R_D R_B + R_C R_B}{R_C R_B}\right) \qquad (4)$$

Next rearrange Eq 3 as

$$R_B R_C + R_B R_D = R_C R_D \qquad (5)$$

and substitute in Eq 4:

$$A = \frac{1}{2}\left(\frac{R_D R_C + R_C R_D}{R_C R_B}\right)$$

Therefore

$$R_B = \frac{R_D}{A}$$

is the second equation. For the last equation solve Eq 5 for R_C

$$R_C = \frac{R_B R_D}{R_D - R_B}$$

Figure 3 shows an example of a circuit using 600Ω resistive loads and a gain of 2. The signal at IC_{1B}'s output measures only 2 to 3% of the signal-generator output.

Rudy Stefenel, *Luma Telecom, Santa Clara, CA*

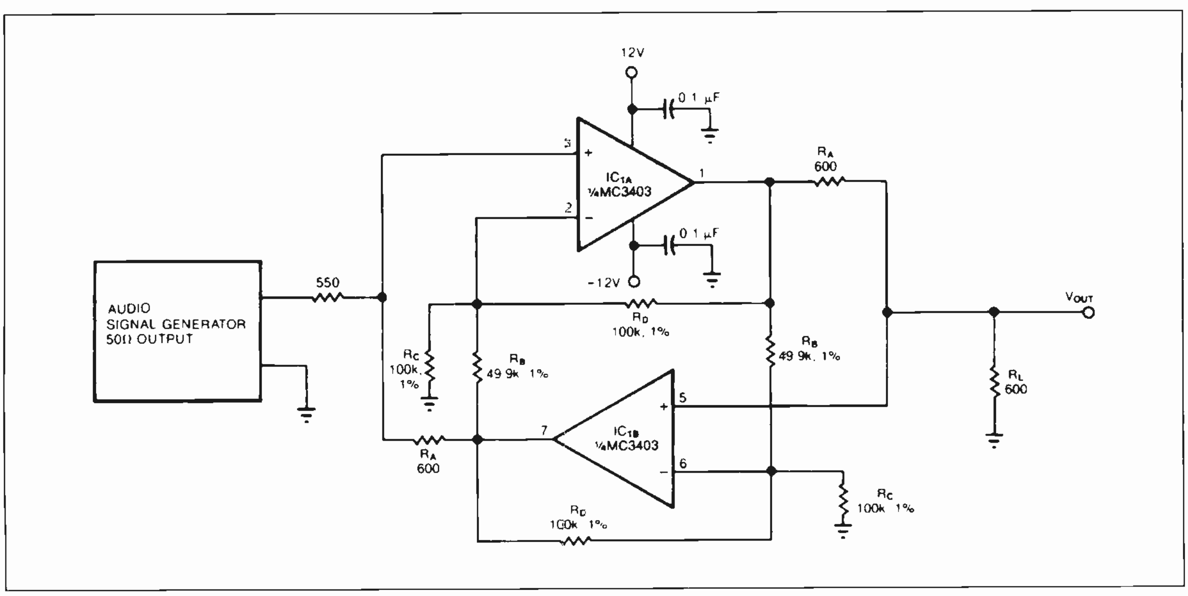


Fig. 3. Based on the circuit of Fig. 2, this two-way amplifier provides a gain of 2 between 600Ω terminations.

Is it Gaussian?

A spectrum analyser will measure the bandwidth of a noise source, but different equipment is needed to determine whether or not the noise amplitudes have a Gaussian probability distribution. One approach is to use a simple circuit (**Fig. 1**) and a DC voltmeter to measure the circuit's output.

Going back to basics, Gaussian noise voltage (or current) with RMS value S and instantaneous value V has a probability-density p(V) distribution defined by the equation

$$p(V) = \frac{e^{\frac{-V^2}{2S^2}}}{S\sqrt{2\pi}}$$

For given values of S, the equation produces curves such as those in **Fig. 2**. The probability that the instantaneous value of V will occur between any particular two levels is the area under the curve between those levels. And the area under each curve from –infinity to +infinity is unity because the chance that V will occur between those limits is 100%.

The probability that V will occur between the arbitrary levels V_1 and V_2 can be calculated by integrating p(V) between those levels. Or the cumulative probability that V will occur between a level V_1 and +infinity can be calculated: integrate p(V) from V_1 to +infinity for V_1 values between 0V and 10 times the RMS value S (**Fig. 3**). The cumulative probability falls off sharply as V_1 increases, so V_1 values to 10x the RMS level only need to be considered. The probability that noise will exceed 10S is less than 1ppm. Note also that the noise voltage near 0V has a near-equal chance of being positive or negative. So the cumulative probability for Gaussian noise near 0V is 0.5, regardless of the RMS level.

The cumulative-probability circuit of Fig. 1 can be used to generate the curves of **Fig. 3**. First connect the noise source and DC voltmeter as shown and then adjust R_1 for the desired V_1 value at the comparator's inverting input. The And-gate output V_{IN} will be high when the input noise is above V_1 and low otherwise, producing a duty cycle corresponding to the cumulative probability that V will exceed V_1.

The DC voltmeter's output divided by 5V equals the cumulative probability if the voltmeter input swings between 0 and 5V. The cmos gate IC_2 provides such an output swing when lightly loaded, provided the power supply is adjusted to obtain an accurate 5V output level.

Then, with the noise source connected and V_1=0V, the effect of comparator-offset voltage can be removed by adjusting R_2 for a voltmeter reading of 2 5V. Under these conditions, V_{IN} will spend 50% of the time at 0V and 60% at 5V, yielding a cumulative probability of 2.6V/5V=0.5.

Test whether a noise source is Gaussian
The frequency distribution of a noise source's output may (or may not) be white, up to a given frequency. This can easily be checked by a spectrum analyser. Likewise, its output amplitude distribution may not have the normal distribution of true random noise. Founded upon basic theory, this article describes how to assess whether a noise source's amplitude distribution is, in practical terms, Gaussian, following the normal distribution.
IH

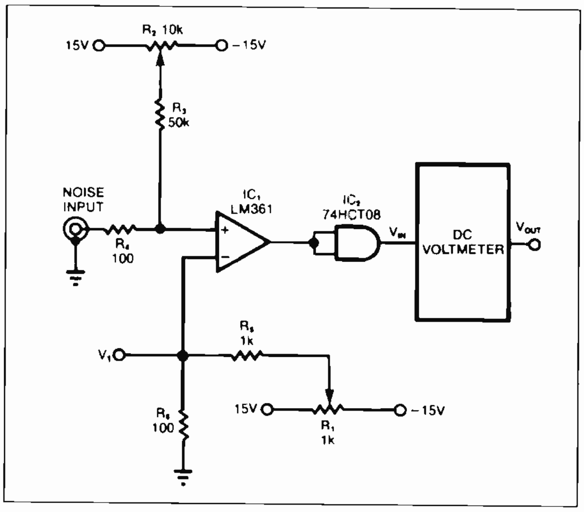


Fig. 1. Circuit to measure the cumulative probability (V0/5V) of the input noise-voltage excursions exceeding the V1 threshold.

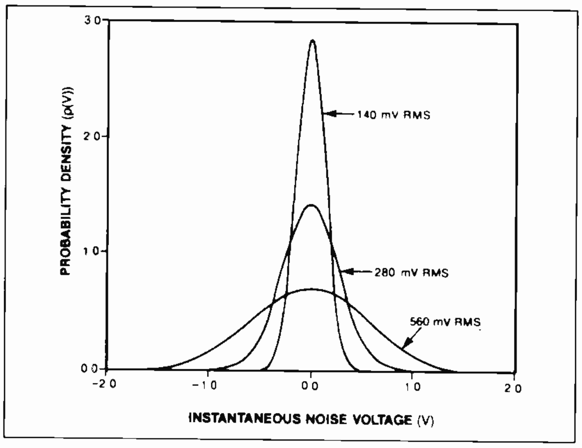


Fig. 2. Classical Gaussian distribution for noise change shape with the RMS level, but the area under all such curves is unity.

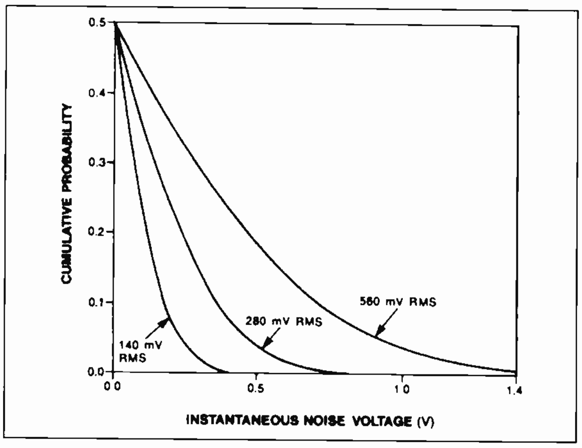


Fig. 3. Circuit in Fig. 1 yields these cumulative-probability curves, which correspond to the Gaussian distributions in Fig. 2.

A non-zero, positive V_I will cause V_{IN} to spend less time at 6V, resulting in a lower value of cumulative probability.

For accurate measurements, the comparator's bandwidth should exceed that of the input noise by a minimum factor of 10. Otherwise, it will cause error in the comparison of measured and theoretical cumulative probabilities – that is, by reducing the measured RMS voltage. To avoid this problem, certain factors should be taken into account. For $V_I=S$, the theoretical cumulative probability for Gaussian noise is 0.1589. If V_I is adjusted to V_0=794.3mV (which corresponds to a cumulative probability of 0.1589), V_I will equal the RMS noise voltage S. Further, the noise source is Gaussian if the same value of V_0 is obtained for various multiples and sub-multiples of V_I.

Figure 4 illustrates the result obtained when using the circuit in Fig. 1 to test a Gaussian source that is part of a commercial noise tester. The congruence of the curves for measured and theoretical data indicates the source is "very" Gaussian.

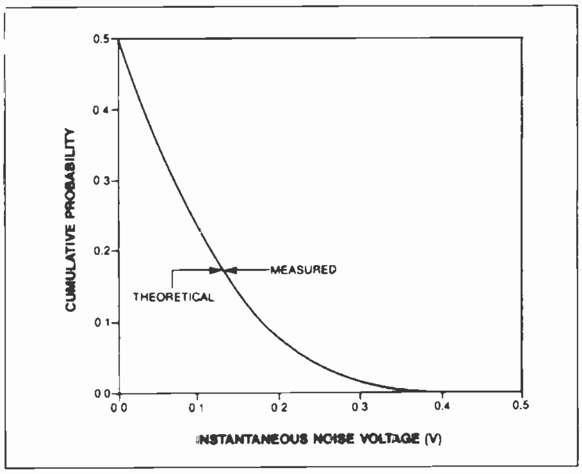
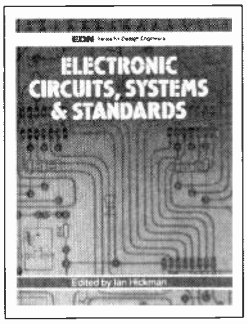


Fig. 4. Curve showing the result of using the test circuit of Fig. 1 and illustrates near-ideal Gaussian characteristics for the noise source in a commercial noise-tester.

Stuart R Michaels, *ILC Data Device Corp, Bohemia, NY*



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VHF frequency divider

Working on an analogue principle, this divider covers the 140-150MHz range, dividing by 16 and presenting a 50Ω input impedance. It offers two main advantages over the usual logic solution: low power consumption and an input sensitivity of around 20mV. Although this circuit operates at a fairly low frequency, an adapted stripline version fitted with high f_t devices could operate as a low power divider to several gigahertz.

Division by 16 is obtained here by four divide-by two stages, **Fig. 1** being the block diagram and the practical circuit. In essence, the circuit is that of an oscillator which does not start. To operate as a divider, feedback from the collector produces a voltage of the same frequency and phase as the input, the 5k trimmer in Fig. 2 bringing the circuit near to oscillation and *LC* being chosen so that their resonant frequency is at the centre of the sync. range. Further cells use the input circuit shown in **Fig. 2**.

Taking the divide-by-sixteen circuit as a whole, the minimum sync. range was 138-152MHz, output fundamental signal was about 50mV and, for inputs from 25mV to 75mV, current consumption was 30mA.

Mihai Sanduleanu
Suceava District
Romania

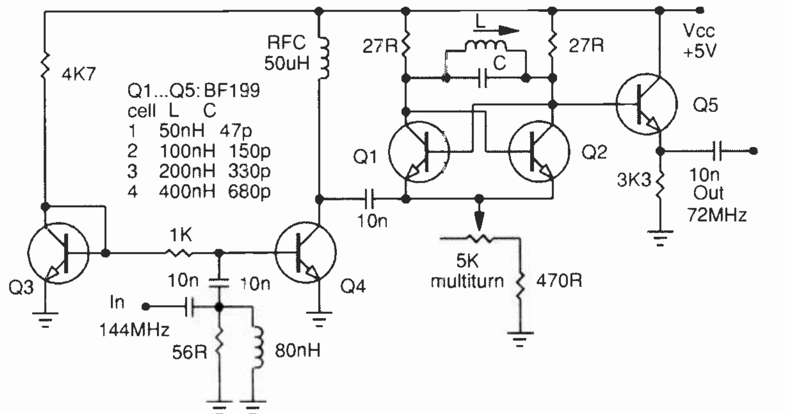


Fig. 1. Block diagram of analogue frequency divider – basically, an oscillator set just short of oscillation, its practical form being the emitter-coupled "oscillator" shown. Values of C and R for the four cells are: 50nH/47p; 100nH/150p; 200nH/330p; 400nH/680p.

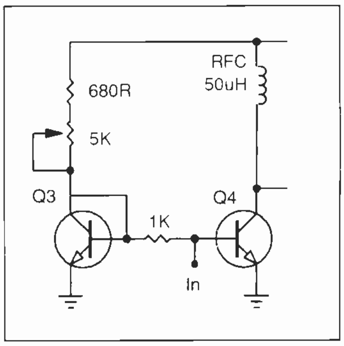


Fig. 2. Input circuit of succeeding division cells, in which the potentiometer provides control of Tr4 bias current.

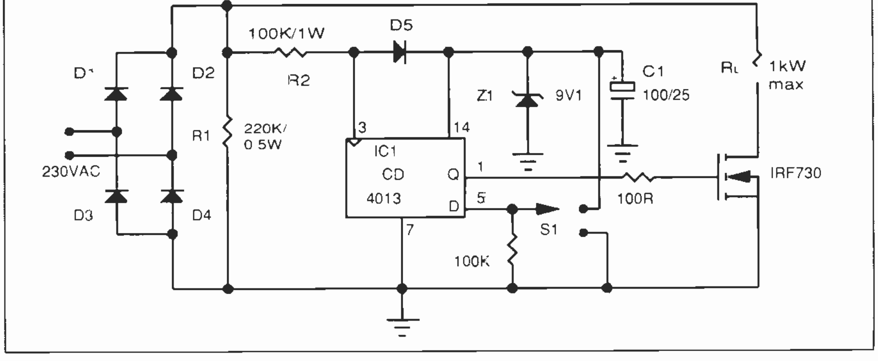
Zero-voltage switch

Switching resistive loads such as lamps and heaters at zero voltage lengthens their life, but usually requires transformers and relays. This circuit uses a flip-flop and some diodes, with a mosfet.

Diode bridge D_{1-4} rectifies the mains supply, the resulting 9.1V across C_1 energising the *CD4013B* D-type flip-flop, its clock input being clipped by D_5 to V_{DD} plus a diode drop. Resistor R_1 pulls it to ground when line voltage is zero. Connecting the *D* input to ground or V_{DD} via the switch sets the flip-flop and the information is latched by the clock. The *Q* output now controls the mosfet, switching it on or off, depending on the position of the switch.

Edge triggering immunises the circuit against bounce and the zero-point switching protects the switch contacts.

M S Nagaraj
ISRO Satellite Centre
Bangalore, India



Simplified on-off switching for resistive loads taking high inrush currents.

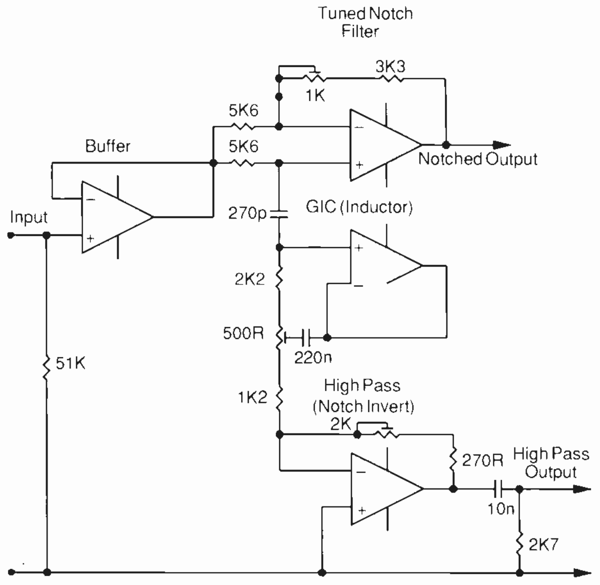
Notch and high-pass filter

This somewhat unusual filter was intended to extract the 11kHz pilot tone from the Russian Gorizont (Horizon) satellite television and sound channels with a view to using it to control an *NE571* compander in its expanding mode.

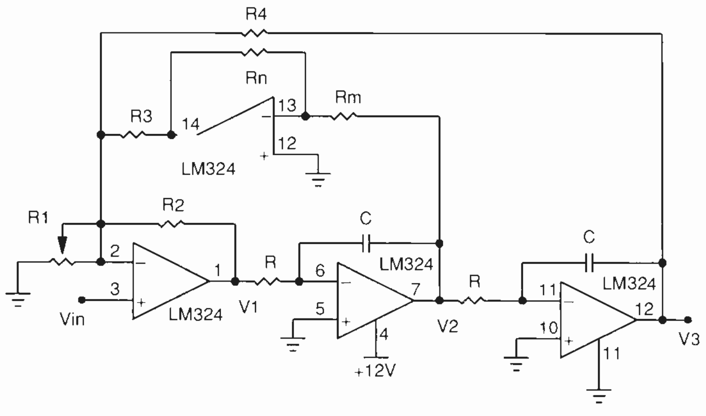
At the top is a balanced bridge, with one arm taken to signal ground of a series-tuned circuit using the impedance converter, which is tunable by the 500kΩ pot to allow for circuit tolerances; for the same reason, the bridge balance is also adjustable by the 1kΩ pot. This arrangement gives a notch with a very high Q – about 40dB rejection.

The earthy end of all this goes to the virtual earth of the bottom op-amp; a high-pass filter with its maximum output at 11kHz and a steep slope. This is also adjustable because of uncertainties about the level of the 11kHz pilot tone.

Reg Williamson
Kidsgrove
Staffordshire



Combined notch and high-pass filter, designed to extract the 11kHz pilot signal from the Russian Horizon satellite transmissions



Three op-amps make up this state-variable filter, which gives all three configurations simultaneously; the fourth one simply allows for Q adjustment.

Active filter

Compared to other filter designs of this broad type, this one offers the advantages of higher input impedance, adjustable gain and adjustable Q.

High-pass, low-pass and band-pass configurations are available simultaneously, the outputs coming from V_1, V_2 and V_3 respectively.

Since Q = R_n/R_m and $\omega_0 = 1/RC$, Q is variable simply by changing R_n and $_0$ by either R or C. Similarly, as the gain A of the filter is $A = (3 + R_2/R_1)$, it is also adjustable by R_1 only.

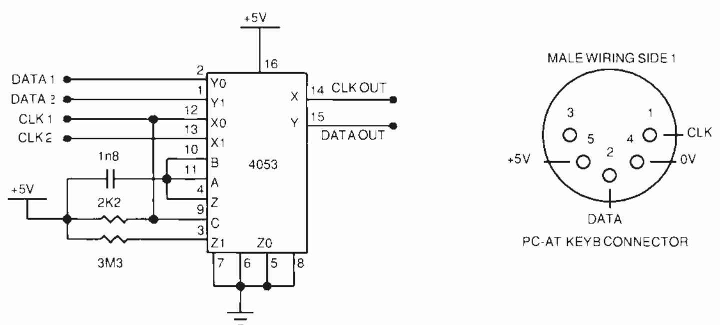
Kamil Kraus
Rokycany
Czechoslovakia

Remote keyboard for your PC

With one IC, you can add another, remote keyboard to a computer or control remote peripherals.

The *4053B* is a triple cmos single-pole, double-throw switch with very low on impedance and off leakage current. In the connection shown, the keyboard giving CLK2 and DATA2 is the default, a 0 from the remote unit's clock on CLK1 making the switch and connecting the remote keyboard for a few milliseconds. Zero and 5V keyboard supplies are common.

Ron Weinstein
Centralab Tel Aviv
Israel



4053 cmos triple switch allows connection of a remote keyboard.

DESIGN BRIEF

CFBOs: delivering speed at any gain?

How hard can you push the LM6181 CFBO? Ian Hickman reports.

Current feedback op-amps have now achieved a ready acceptance by semiconductor manufacturers and design engineers. At first, it was the specialist analogue IC manufacturers such as Comlinear and Elantec that pioneered their introduction. But since then, mainstream semiconductor manufacturers – Analog Devices and Harris Semiconductor – have all come forward with products.

National Semiconductor also produces the *LM6181* CFBO, mounted in an evaluation board with accompanying data sheet, applications leaflet and *Spice* model on floppy disk. The device itself is packaged in an 8-pin DIP outline, with the usual pin-out. Specified typical –3dB bandwidth at a closed loop gain A_v of +2 and a 1kΩ load, is 100MHz. It is still well over 50MHz when driving a 100Ω load such as a back-terminated 50Ω cable.

Testing the LM6181

The evaluation board circuit diagram (**Fig. 3**) shows positions of the links and optional components exactly as found on the board as supplied.

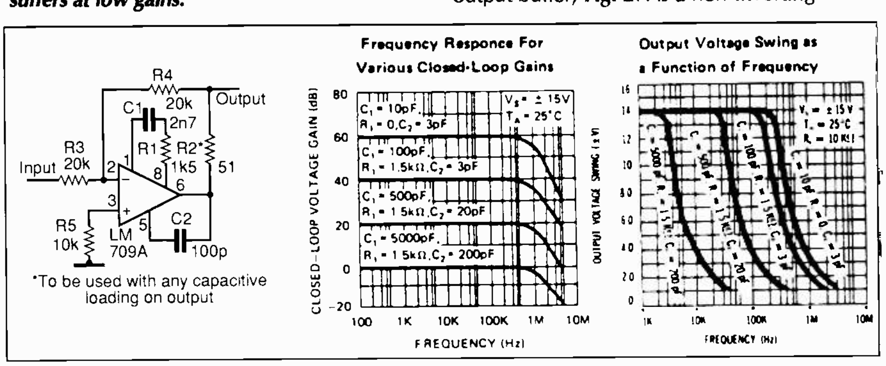
Absence of a spectrum analyser but having an oscilloscope with a 300MHz bandwidth dictated time domain testing had to be my approach. Rise and fall times of the 10MHz square-wave output of my 10Hz–10MHz video oscillator are a leisurely 20ns – scarcely enough to exercise the *LM6181*'s capabilities. So I built the test circuit of **Fig. 4**, using ground-plane construction and a strip-line connection between pin 11 of the Harris *CD74AC00* quad two-input Nand gate and the BNC output socket. Output was connected to a 50Ω three-way resistive splitter, with one output applied to channel 1 of the oscilloscope by a coaxial lead with a 10dB pad at the split-

High speed CFBOs

CFBOs (current feedback op-amps) are noted for their high speed and wide bandwidth – qualities that do not deteriorate too markedly as the demanded gain rises from unity to times ten or even more. Their performance is in marked contrast to conventional VFBOs (voltage feedback op-amps) with internal compensation for gains down to unity, such as the *741*. Even using a VFBO, the band-width reduction with increasing gain is largely avoided when an op-amp with external compensation is used.

Fig. 1 shows the closed loop bandwidth achieved by the earliest popular monolithic op-amp, the *709*, when the compensation is correctly selected, according to the set gain. But though bandwidth is remarkably well maintained with increasing gain, it still amounts to only a few percent of the bandwidth of a CFBO. Further, like a VFBO internally compensated for gains down to unity, a CFBO does not need any external compensation components selected according to the required gain (except perhaps when seeking to wring the last ounce of performance from the circuit). There is only one internal node at which there is any significant voltage swing – the point immediately preceding the unity gain output buffer, **Fig. 2**. As a non-inverting amplifier, the *I* terminal is connected to ground through resistor R_s and to the output through resistor R_f. The input is applied to the high impedance *NI* terminal and the gain is $(R_f + R_s)/R_s$. Due to the unity gain of the input buffer and the very low value of the amplifier's I-terminal internal resistance R_{in} (tens of ohms), input voltage at the *NI* terminal is closely reproduced at the *I* terminal, the small difference being equal to I_{in} times R_{in}. Current mirrors cause I_{in} to flow in R_t, the input impedance (typically several megohms) of the output buffer, giving the device a low-frequency open loop voltage gain of R_t/R_{in}. The open loop current gain is simply equal to the current gain of the output buffer. I_{in} is the difference between the current in R_s and that in R_f, a CFBO working as a transimpedance amplifier.

Fig. 1. The venerable 709 externally compensated op-amp provides a wider (small signal) bandwidth at high gains than even a modern voltage feedback op-amp internally compensated for gains down to unity, although the large signal bandwidth suffers at low gains.



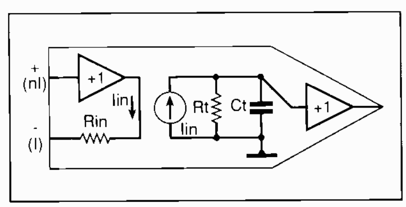
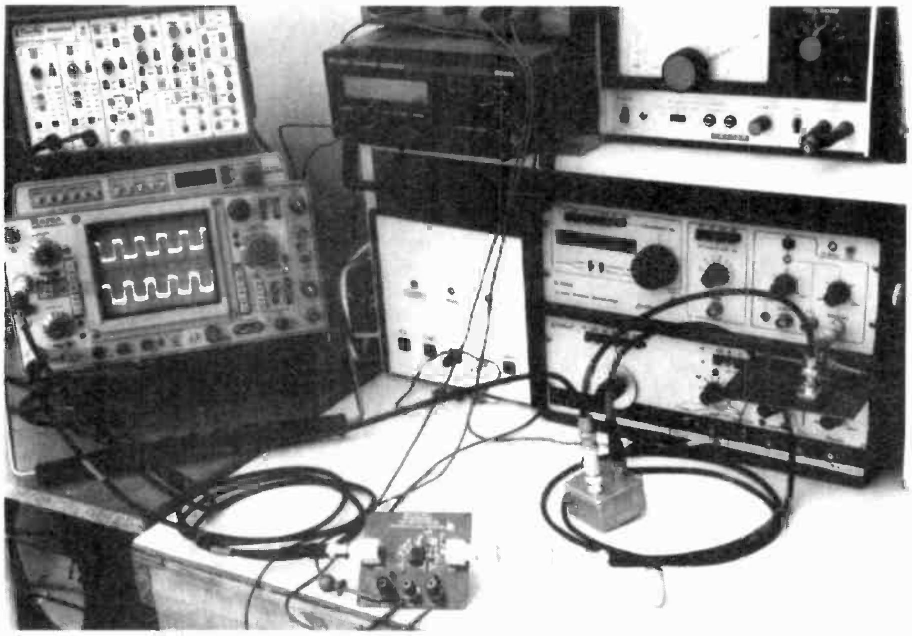
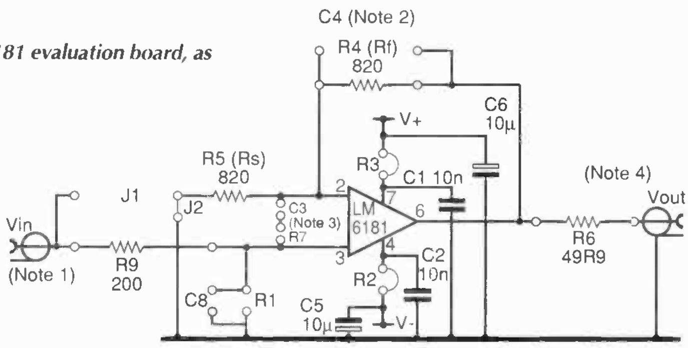


Fig. 2. In a CFBO, all gain is raised as current gain, except for the node immediately preceding the unity gain output buffer. So there is only one significant lag RtCt, rather than two or more as in a VFBO. (Courtesy Analog Devices.)



Equipment set-up that produced the test results .

Fig. 3. LM6181 evaluation board, as supplied.



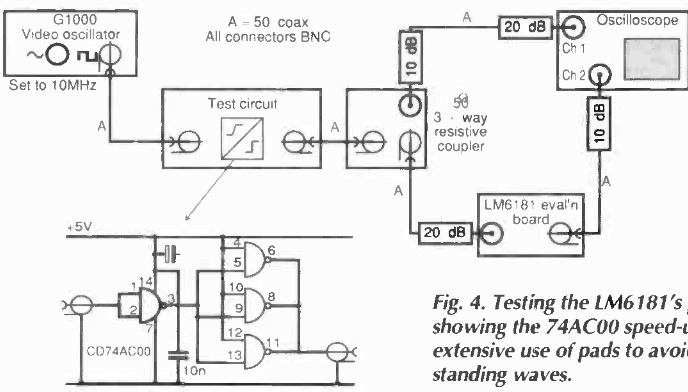
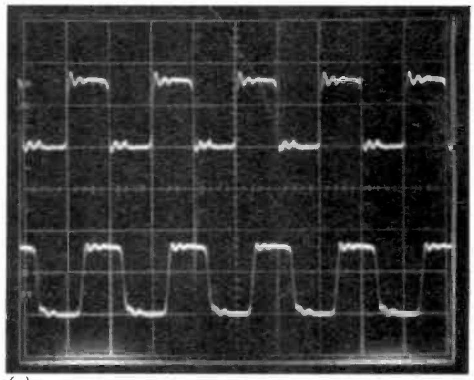


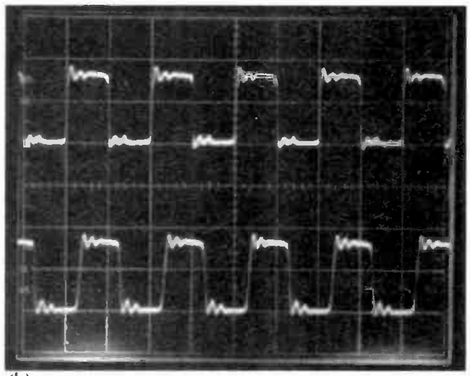
Fig. 4. Testing the LM6181's performance, showing the 74AC00 speed-up circuit and extensive use of pads to avoid reflections and standing waves.

ter end and a 20dB pad at the oscilloscope end. With 30dB of pads and 6dB loss in the splitter, the (almost) 5Vpk-pk amplitude out of the Nand gate board might be expected to be 75mV pk-pk. But the scope presents a high input impedance, leaving the 20dB pad unterminated. So the channel 1 input is in fact 0.15V pk-pk (**Fig. 5a** upper trace), 100mV/division.

As a test signal, it is not as clean as is ideal. But at a little over 2ns, the rise and fall times are good and fast. Decoupling includes a ceramic chip capacitor connected directly from pin 14 of the Nand gate to ground-plane, but the device is socketed, resulting in longer



(a)



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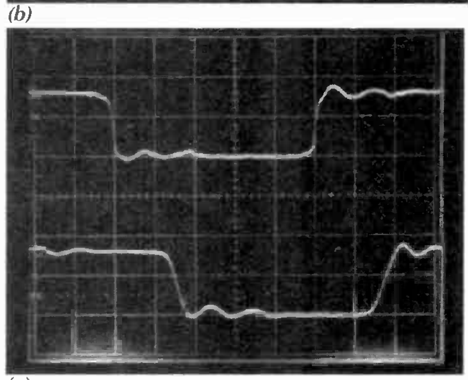

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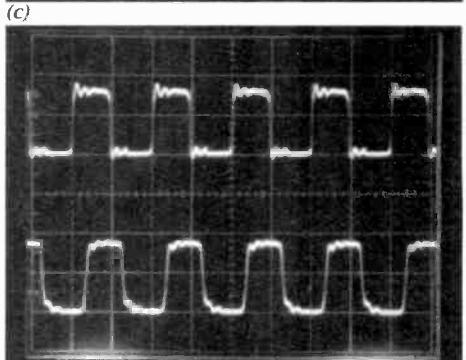
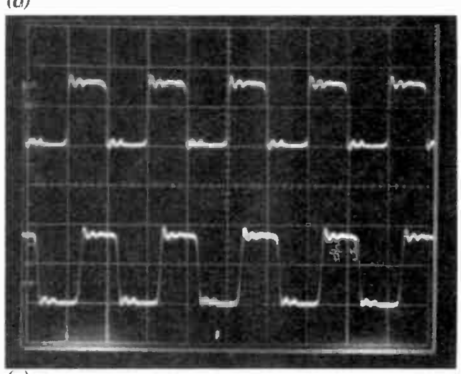
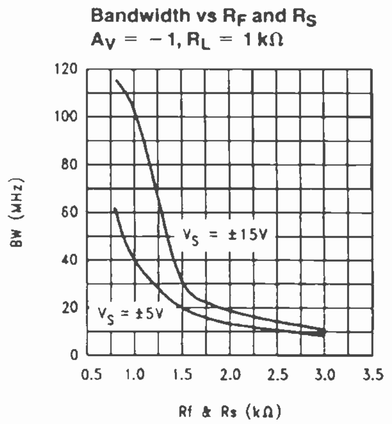


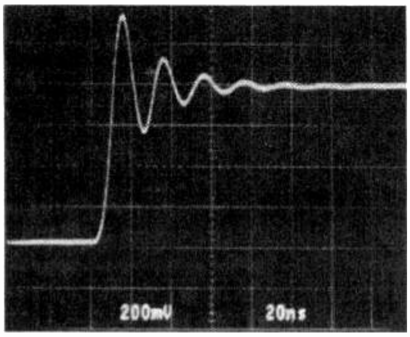
(d)



(e)

Fig. 5a) Performance of the LM6181 on its evaluation board, at unity gain; upper trace, input waveform (channel 1, 100mV/div); lower trace, output waveform (channel 2, 200mV/div); b) with R9 shorted; c) timebase speed increased from 50ns/div to 10ns/div d) demanded-gain increased by 10dB; and e) a peaking network of 10pF in series with 180Ω connected in parallel with R5 (Rs).





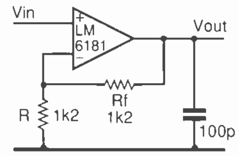


Fig. 6a) With suitable values of Rf and Rs, the LM6181 is capable of driving a load capacitance as high as 100pF. b) Normally, if Rf = Rs = 820Ω, the LM6181 would oscillate with 100pF of capacitive load. In this example the feedback Rf and Rs values are scaled to 1.2kΩ so that the closed loop gain is Av = +2, but the open-loop band width decreases, maintaining adequate phase margin. c) By scaling both RF and Rs the closed-loop gain stays constant but the bandwidth changes.

leg lengths and the consequent ringing shown.

Channel 2 at 200mV/div (lower trace of Fig. 5a) shows the *LM6181* output. As in channel 1 there is a total of 30dB of pads in circuit, the scope end again being unterminated. The *LM6181* drives the output coax through a 49.9Ω source resistor and in conjunction with the x2 voltage gain set by R_4 and R_5 this should result in unity overall gain. Apparent gain of x2 is due to the fact that the 20dB pad at the amplifier's input port is also unterminated, and the amplifier presents a high input impedance due to operating in the non-inverting mode (unterminated pads do not result in any significant mismatches, since the input of an unterminated 10dB pad presents a 20dB return loss to the source.) The output waveform is a fair copy of the original – though the rise time is clearly degraded because of the pole created by R_9 and the board and device capacitance at the op-amps's *NI* input, pin three. **Figure 5b** lower trace shows the result of shorting R_9; the output rise-time is improved, while the ringing is slightly increased compared with the input (upper trace) - a value for R_9 somewhere between zero and 200Ω would seem to be optimal. **Figure 5c** is the same as 5b except that the timebase speed has been increased from 50ns/div to 10ns/div.

Output rise and fall times (lower trace) are about 5ns. Given that the amplifier's bandwidth when driving a 100Ω load is about 55MHz, this is a very creditable performance bearing in mind the rule of thumb for a single pole response which says that risetime (ns) is approximately equal to 350/bandwidth (MHz). Incidentally, the 16ns mid-swing delay evident between the two traces indicates the length of the extra coaxial cable in the path via the amplifier. Assuming the wave velocity in the cable is 2×10^8m/s (2/3 of that in free space), the length is indicated as 3.2m, whereas it actually measured only an extra 2.2m (11ns). The remainder is due to the tracking on the evaluation board and the 5ns propagation delay in the device itself.

Demanding more gain

Obtaining a higher gain with a CFBO means increasing ratio R_f to R_s, just as with a VFBO. But this is achieved by *lowering* R_s, not *increasing* R_f (which would have a deleterious effect upon bandwidth). The increased gain is thus obtained by increasing the drive current to the input buffer portion of the CFBO (Fig. 2). **Figure 5d** shows the result of increasing the gain by 10dB (from x2 to x6 approx), by connecting a 180Ω resistor in parallel with R_s; an additional 10dB pad has been inserted between the splitter and the evaluation board so channel 2 deflection factor is unchanged.

Some deterioration in performance is evident, but not nearly as much as would be the case with a VFBO internally compensated for gains down to unity. The question is: will the sort of compensation techniques used for decades by designers of video amplifiers and oscilloscope deflection stages work with the *LM6181*? The answer, obtained after only a very little experimentation is yes, as **Fig. 5e** shows. Apart from the slower rise and fall times, the output trace is a faithful copy of the upper input trace. Indeed it is a more faithful copy than either Fig. 5a or 5b. As indicated above, when changing the gain of the stage, normally R_f is left at 820Ω and R_s adjusted as required. But sometimes changing the value of R_f can be beneficial – such as when driving capacitive loads. Load capacitance in conjunction with the CFBO's output resistance adds another pole in the loop, reducing the phase margin and, at the extreme, causing oscillation. At a gain of +2 and with a 100pF load capacitance (**Fig. 6a**) the *LM6181* would oscillate if $R_f = R_s$ = 820Ω. Scaling both to 1200Ω keeps the closed loop gain at +2 but decreases the open loop bandwidth, restoring an adequate phase margin (**Fig. 6b**). The way that bandwidth changes with R_f scaling, for the case where $A_v = -1$, is illustrated in **Fig. 6c**. As so often in electronics there is more than one way to achieve the desired result. Thus with $R_f = R_s$ = 820Ω, even a 48pF load will result in excessive ringing (**Fig. 7a**) but buffering it from the *LM6181*'s output with a 47Ω resistor (**Fig. 7b**) gives a much more satisfactory result (**Fig. 7c**).

References

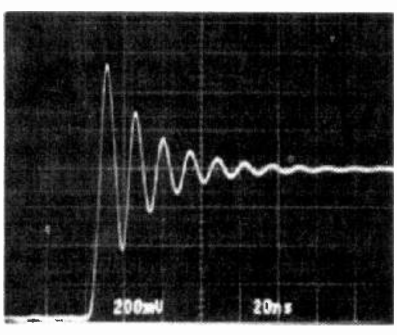
1. "The current alternative to operational amplifiers", F Ogden, *EW + WW*, August 1992 p. 643.

Acknowledgments

Figures 1, 3, 4, 6 and 7 courtesy National Semiconductors.

Inverting illustrations

Illustrations all relate to the non-inverting circuit. But the *LM6181* and other CFBOs can also be used in the inverting connection. Indeed some CFBOs from one manufacturer are stated to be designed primarily for use as inverting amplifiers, although they are "also suitable for use in many non-inverting applications".



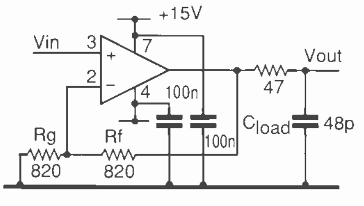
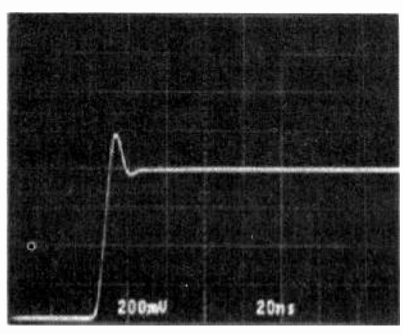


Fig. 7a) With Rf = Rs = 820Ω, a 48pF load will cause excessive ringing b). This can be avoided by buffering the capacitance from the op-amp with a 47Ω resistor, giving a healthier result c).



BBC puts digital TV into practice

A lot of European effort is being expended on digital television and HDTV. Tom Ivall went along to the BBC to see how the UK's contribution was progressing.

As the man crossed the road he left a bit of his leg behind – at least for a split second. But this was no circus trick, rather one of the characteristic problems of digital television.

The picture was a busy street scene, shown on a 16:9 monitor, and we were viewing results of the latest work on digital compression for bit-rate reduction at the BBC's research department, one of the many collaborators in the major international effort on digital TV.

One demo showed an HDTV picture after being compressed to give a bit-rate of 25Mbit/s. The system, using a new algorithm, was based on a coding method currently being considered by the Motion Picture Experts Group (MPEG) of the International Standards Organization. But the MPEG project is for compressing ordinary definition TV signals to give bit-rates up to about 10Mbit/s.

25Mbit/s may still sound high but it is in fact a considerable compression. To retain sufficient resolution in a digitised HDTV picture, the signal must be sampled at a rate above 70MHz in the A-to-D converter and the resulting raw data rate is over 1000Mbit/s. A compressed rate of 25Mbit/s for digital HDTV is about four times the mid-range rate which the MPEG algorithm is aiming at for conventional television (where the raw video data rate is 216Mbit/s).

Compression factors of about 30 to 40 are now being seen – though the MPEG has conducted tests on algorithms from a variety of organisations with compression factors ranging from about 20 up to 54.

International collaboration

The current international work, which includes a substantial European input from the Eureka 625 project, exploits the psycho-physical limitations of human vision, the spatial and temporal redundancy of TV pictures and the statistical characteristics of the picture information. To do so, it uses a combination of techniques such as discrete cosine transform coding, motion-compensated inter-frame prediction and variable-length coding. (Aubrey Harris outlined the main benefits of digital signalling in *EW + WW*, June 1992)

Ultimately the digital signals have to be transmitted in television channels of limited bandwidth, and for HDTV particularly the spectral efficiency of the modulation system is all-important. At the demonstration, along with its own work, the BBC showed a recording of digital TV pictures broadcast in the UK with the extremely high spectral efficiency of 7.5bits/s/Hz.

In fact this was an experimental transmission of 60Mbit/s in a standard 8MHz television channel The modulation technique giving this remarkable performance, developed by France's Thomson-CSF/Laboratoires Electroniques de Rennes, uses orthogonal frequency division multiplexing (OFDM) and quadrature amplitude modulation (qam).

One contribution to the spectrum space efficiency comes from use of both horizontal and vertical polarisation. A 30Mbit/s signal is transmitted on each, made up of an OFDM ensemble of about 500 closely spaced carriers, all digitally modulated with 64 qam.

UK tests

The BBC's involvement was to provide the conditions for a realistic broadcasting test of the French system in the UK. The digital signal was radiated in UHF channel 28 (a standard 8MHz TV channel) from a low-power transmitter at the Crystal Palace station in London. Signals were received about 10 miles away at the research department in Kingswood, Surrey, and at other places in South London and Surrey.

To an outsider, the system looks complex and vulnerable. But engineers involved claim that it is rugged in performance. It certainly seems to offer the prospect of transmitting at least one digital HDTV service in a standard terrestrial broadcasting channel – without digital compression being applied.

For contribution links (eg from outside broadcasts back to studios) the BBC has developed, in a collaborative project, a codec for sending HDTV signals at high quality over a standard 140Mbit/s telecommunications circuit.

The BBC has now demonstrated how this codec can also be operated at the lower bit-rate of 70Mbit/s, to save cost, when the programme material is less demanding. ■

Camera with adaptive approach

Digital processing is bringing benefits in a new design of radio-camera for TV outside broadcasts. Instead of trailing cables behind, these portable cameras carry a small microwave transmitter and antenna, allowing the camera operator considerable freedom and flexibility. A nearby receiving dish picks up the signals.

Currently the BBC is using circularly polarised omni-directional antennas to combat multipath propagation problems. But the omni-directional radiation pattern still causes some reflections and the signal between camera and receiving dish can be blocked by obstructions.

To overcome these difficulties the research engineers have developed a radio-camera with a cluster of six directional horn antennas, all pointing in different directions. These are automatically selected by a switching system so that at all times during an OB the best signal path between camera and receiving dish is brought into play.

The system works by repeatedly transmitting video test lines through the horns in turn. Decisions on which horn gives the highest signal strength at any instant are made in the receiving equipment, which sends back corresponding data signals to the camera. The test lines and switching decisions occur during field blanking intervals, and because the whole system is updated at frame rate 25 times/s it adapts immediately to all camera movements.

BBC engineers have developed a radio-camera with a cluster of six directional horn antennas selected by a switching system to ensure the best signal path between camera and receiving dish.

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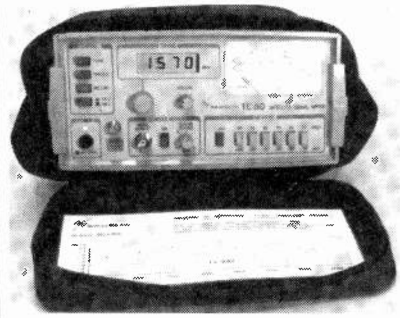

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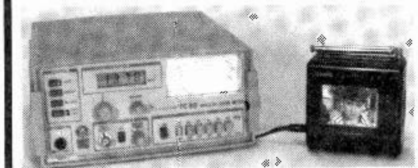

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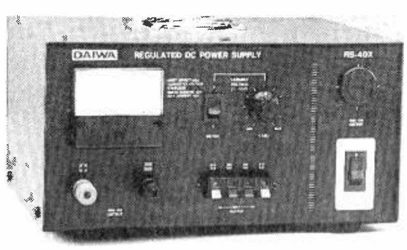

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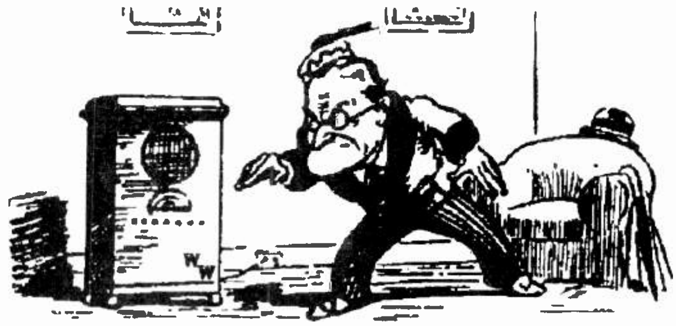

WHITE NOISE *by Hot Carrier*

Return of the filament?

In the days of the valve, hot carriers were just that – electrons boiled off from a red hot cathode or glowing filament. Now students at the University of California have developed "micro-filaments" using IC fabrication techniques. The 200µm long devices, grown on a silicon chip with a tungsten filament 5µm wide and 0.7µm thick, emit almost half a lumen at less than 100mW of input power, and operate as a micro-sized triode. Performance is comparable with the 35mW drawn by the 1.4V 25mA filaments of the *DK96* range of B7G battery valves. Presumably if the filament could be coated with the usual mixture of rare earths to reduce the work function, the same electron emission could be obtained at a much lower filament temperature, drawing less heater power. This opens up the prospect of a renaissance of thermionic devices. After all mobility of electrons *in vacuo* is much greater than in silicon, gallium arsenide or even diamond, currently being researched as the next super solid state technology. Nor should the presence of a filament be regarded as necessarily a source of fragile unreliability. Special purpose valves were successfully used in artillery fuses where the peak shock due to acceleration reached 50000g. The side-slap forces in an end-of-life barrel were estimated to reach a quarter of a million g.

Words of complaint

Recycling as an aid to economy in the use of the world's limited resources is a laudable aim. Not so attractive is the re-use of technical terms with an entirely different meaning. For example a "phantom" used to be an additional telephone circuit carried over a pair, effected by using the centre taps of the line-side windings of two other pairs or "side circuits". The latter were physical circuits whereas the additional circuit operated (apparently) without its own pair: hence it was known as a phantom.

But in the modern world of communications, where we worry about the effects on our brains and eyes of a handheld phone unit transmitting maybe a couple of watts of RF at 900MHz or more, the word has an altogether different meaning.

The NRPB (National Radiological Protection Board) has laid down limits on the permissible absorption of radiated power in human tissues. But are the limits being met? Volunteers ready to have thermocouples poked into their brains are unsurprisingly hard to find. So we must produce jellies with the same density, thermal conductivity and RF absorptivity as a real brain – the necessary target data presumably having been measured on the organs of a (recent?) cadaver. Sadly, the packet of goo representing a brain or whatever is now referred to as a "phantom", a new and grisly twist to an old term.

Cynical manipulation

Is it better to be pessimistic and have the occasional pleasant surprise or optimistic and be doomed to repeated disappointment?

Pessimistic optimism: Only blinkered isolationists believe the UK can thrive without active trading links with the rest of the world – and Europe in particular. This especially applies to electronics goods and manufacturers. So the move by the seven member EFTA group and the EC to form a 19 nation free trade common market of 380 million people with effect from January 1 1993 must be hailed as good news: *if it goes ahead.*

Optimistic pessimism: We all puffed our chests out with pride in November 1991 when scientists on the Joint European Torus added tritium to the reactor fuel and produced a significant amount of power from the resulting controlled fusion. The achievement thrust the project ahead of others in the USA, Japan and Russia, and perhaps laid the foundations for a golden age of abundant polutionless virtually free energy for all. *Or then again perhaps it did not.* Maybe the latter would be the best outcome for mankind in the long run. Who knows what the effect of a vastly increased release of energy (which mostly all finishes up as heat) will have on the environment?

Logic not language leads to clarity

British engineers are inaccurate and slipshod in their use of language, and unable to express themselves clearly and unambiguously. Not my words, but a summary of a tetchier than usual *IEE News* editorial.

It seems that disconnected relevant clauses, verbalised nouns and split infinitives are solecisms indicating a slipshod and untrained mind.

But sometimes only a split infinitive will do: "Add water to just cover the crystals" is perfectly clear while neither "– just to cover –" nor "– to cover just –" expresses quite the same thing. Of course, we could recast the sentence: "Add water until the crystals are just covered". But to be forced so to do is the kind of pedantry up with which a famous British statesman said he would not put.

Some educationalists still advocate the teaching of Latin as a way of inculcating a logical turn of mind. What a farce! The French, whose electronics industry and economy in general, has arguably outperformed our own in recent years, have a much more logical approach.

To help their schoolchildren to think and express themselves logically, they teach them a subject which never features in a UK curriculum. They teach not Latin but *Logic*.

Put an engineer on top

At an "innovation" seminar organised by the DTI and attended by many of Britain's manufacturing bosses, Akio Morito (founder and chairman of Sony) lectured on the disastrous results for industry of the low esteem in which engineering is held in the UK. He wanted to see more engineers in top management – as is common in Japan (and in Germany too!). According to Morito, too many people running British industry are simply illiterate in technical matters, no matter how many business and economics qualifications they might have. Amen to that.

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