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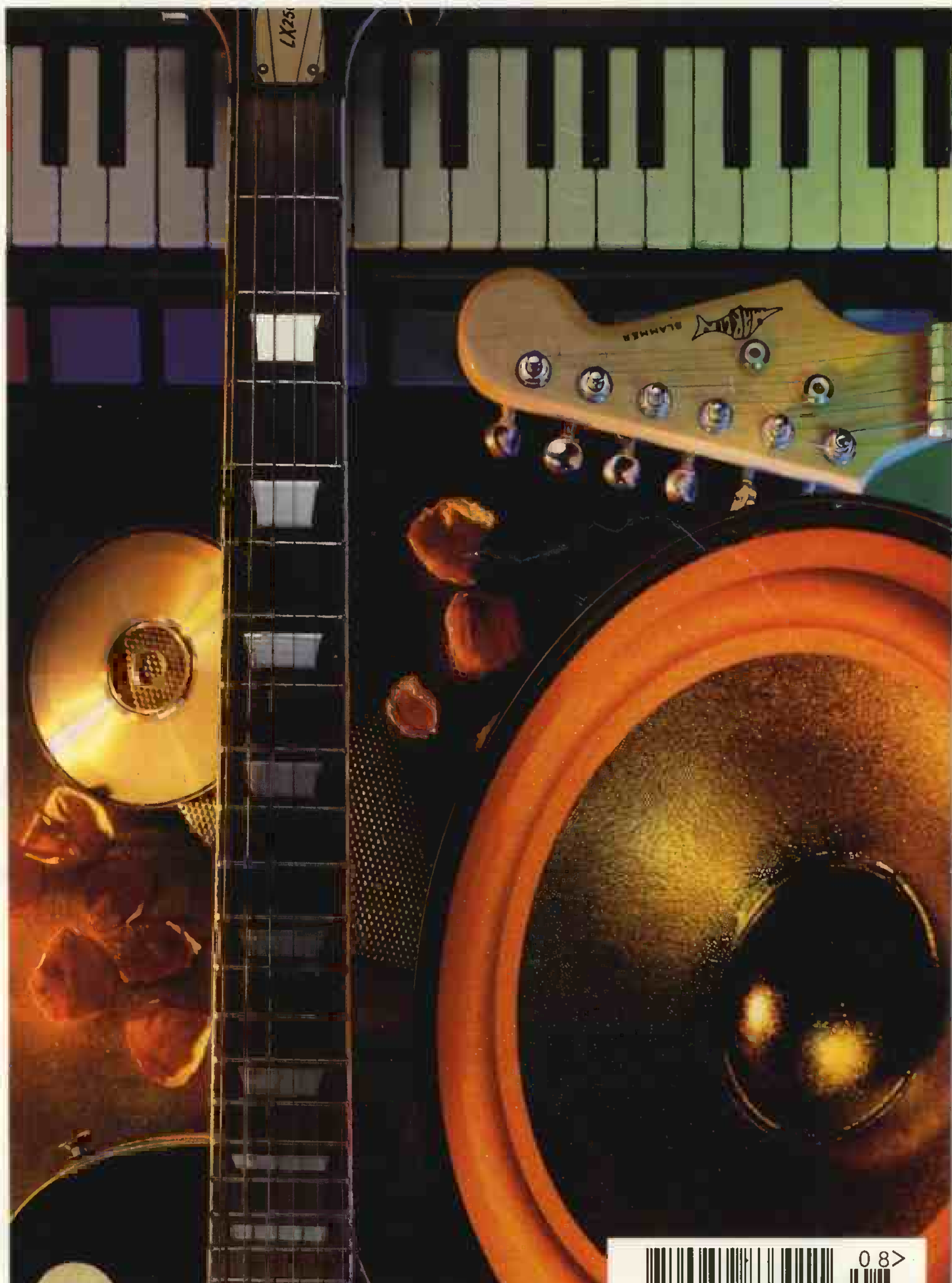

20Hz active subwoofer

Getting more from RS232

PC engineering: signal analysis

24cm antenna

DSP demystified

Analogue signal processing

Linsley-Hood comments on mosfets

RF oscilloscope probe using sm

Exclusive - GPS designer's kit

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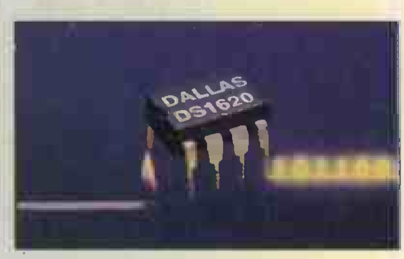
A special offer for John Linsley Hood's book 'The Art of Linear Electronics' appears on page 685.

REGULARS

Cover – Trevor Bailey

A GPS designers kit – that can form the basis of anything from a pc based GPS data analyser to a portable navigation aid – is one of this month's special reader offers, page 664.



Heat-in, bits out – a combined thermometer/thermostat chip operating from –55 to 125°C. A designer's kit based on this chip is also the subject of a unique reader offer.

Next month:

Top designer Walt Jung discusses working with single-rail op-amps, Ian Hegglun presents a high-performance audio power amplifier based on square-law techniques and Douglas Clarkson looks at isolated RS232 links.

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Jug wine or Chambertin?

It was early June in Montreux. The Organisers of the 19th International Television Symposium and Technical Exhibition were holding a lakeside party to welcome the press. Hungry hacks were grabbing anything edible from passing trays of nibbles. At an exchange rate of less than 2 francs to the pound – and no provision made for helping visiting journalists to eat – free food was everyone's first priority.

Exhibition organiser Joe Flaherty stood up. Flaherty is Senior Vice President in charge of Technology at CBS Inc, and for ten years he has been the driving force in the USA behind HDTV and the digital tv Grand Alliance.

"Europe's DVB will be a digital version of MAC. It will be a disaster", Flaherty pronounced, "When, as is sure to happen, perhaps at the Olympics, broadcasters start transmitting in high definition, existing sets will go dark. You can't ever let sets go dark".

All round the area the press dropped food and grabbed notebooks.

Flaherty then explained his point. Europe's Digital Video Broadcasting standard, which will be frozen this September to allow a service launch in 1997, does not provide the same upgrade path from Standard Defintion to High Defintion that the Grand Aliance system offers. The SD digital receivers sold from Day One will not be able to receive an HD broadcast, and decode it as an SD signal. This, Flaherty repeated, will make DVD "a disaster".

"In the future Europe won't be able to receive high-definition pictures from the same transmission channels. And it won't be long before someone puts out HD programming. You can go on drinking jug wine until you taste the Chambertin...".

As the word spread round Montreux of Flaherty's outburst, members of the DVB project, which is now backed by over 160 companies, were asked to comment, often in open seminars. The diversity of comment told its own story. As Henry Price, BBC Head of Engineering Information put it back in the UK.

"There is a huge amount of confusion over all this", NTL already provides the analogue transmitters used by the UK's commercial tv companies, and has offered to build the BBC's netowrk. Tony Gee, NTL's Marketing Manager for digital broadcasting, admitted that the issue of whether DVB receivers sold on Day One will later be able to decode HD programmes for display in SD is "a key question". But he was unsure of the answer. "My understanding is that the likelihood is no" says Gee, suggesting that broadcasters could simulcast the same programme in both HD and SD formats.

Richard Wiley, Chairman of the Federal Communication Commission's Advisory Committee on Advanced Television Service, boasted "In the US no set will ever be dark. We are making sure of that. The Grand Alliance handles both HD and SD transmissions perfectly. The cost of adding HD compatibility to an SD set is only marginal".

George Waters, Director of the Technical Division of the European Broadcasting Union, agreed with Wiley's low cost estimate and added that "Joe was right to stimulate discussion. The US has approached this from the top down; Europe has approached this from the other way round because the project began with cable and satellite. There's a new awareness in Europe. We have to look at the situation again."

So why, if there is now doubt over the standard, is Europe rushing towards finalisation in September 1995? Why not wait another six months or even a year to sort out the issue of HD/SD downwards compatibility? Who is driving the rush?

"The BBC", said George Waters, turning to look at Michael Starks, BBC Controller of Digital Feasibility. Starks looked sheepish, and remained silent.

Does the BBC really believe there will be rioting in the streets of London if it delays the start of digital tv broadcasting by six months? Does anyone in the electronics trade or general public really care about a six month delay, especially if it means Day One sets are futureproofed?

Starks listened to the questions and remained even more sheepishly silent. With the help of engineers from NTL and the BBC it has now been possible to piece the scenario together.

The DVB originally planned to specify a hierarchical system, but it was dropped for three reasons.

It puts up the cost of even the basic receiver, because it needs more memory. Also, if the broadcaster tries to guarantee good reception for portable tv sets, by adding a very robust signal with a low data rate, even high quality SD receivers with roof aerials will lock onto the low resolution picture. And hierarchical coding is very inefficient. It wastes so much spectrum space, that the broadcaster might as well do as NTL's Tony Gee suggested at Montreux, and transmit the same programme twice over, once in standard definition and once in high definition.

But the Grand Alliance thinks it has licked all these problems.

Meanwhile the BBC favours a third option. Programmes will be shot in 1250 line definition, down-converted to 625 line signals and transmitted at high bit rate, probably 9Mbit/s, which is half the capacity of a uhf channel.

Adanced tv sets will then artifically double the number of picture lines to simulate HDTV. However good the up-conversion this option still cuts Europe off from true 1250 line HDTV.

While everyone scurries around, giving different reasons for not doing what the Americans think they can do, Jim Norton, Chief Executive of the Radiocommunications Agency, warns that simulcasting SD and HD will be a no-no. But Norton also offers a neat and simple solution.

"The tv manufacturers should borrow an idea from the computer world, and provide every digital tv with an upgrade slot. That way owners can bolt on whatever upgrade circuits they need in the future".

Is it too late for the DVB group to write such a sensible approach option into its final standard? ■

Barry Fox

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REED BUSINESS PUBLISHING

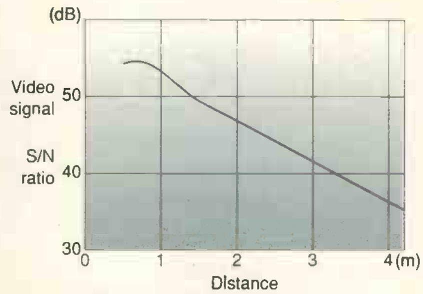
Optical link makes camcorder cordless

Hitachi has launched a camcorder incorporating an optical link which is intended to make playback more convenient. The system comprises a camera with integral transmitter and a remote receiver with connections to a tv or vcr.

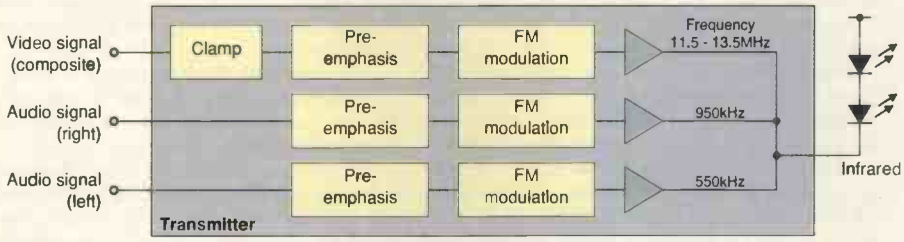
A composite video signal is clamped, pre-emphasised then frequency modulated onto a 11.5-13.5MHz carrier; both stereo audio audio channels also undergo pre-emphasis and frequency modulation, onto carriers at 950kHz and 550kHz.

The signals are combined and emitted by two infrared leds. In the receiver, the signal is sensed by a photo-diode and the process reversed after signal amplification.

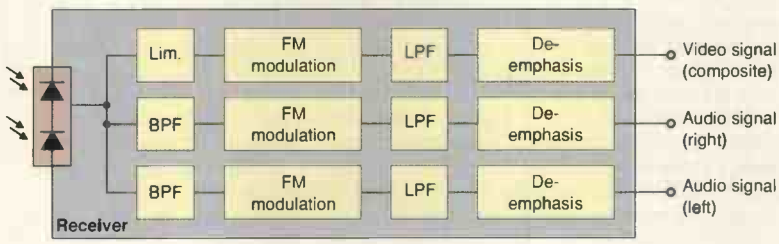
Usable range of the link is 3m, at which point the video signal-to-noise ratio is at a threshold 42dB. The maximum capture angle is about 60°.



As distance between the optical transmitter increases, so does noise. Past 3m, the s:n ratio falls below 42dB and starts to become a problem.



Transmitting and receiving systems of the camcorder one-way optical link. Both audio and video signals are transmitted.



● Coincidentally, JVC Professional has launched a portable video presenter, also capable of optical linking. Its built-in ccd camera is designed to provide the electronic equivalent of an overhead projector, so banishing cables is particularly apt. The principle is the same as the Hitachi optical link but with greater quality and range, and no audio is involved.

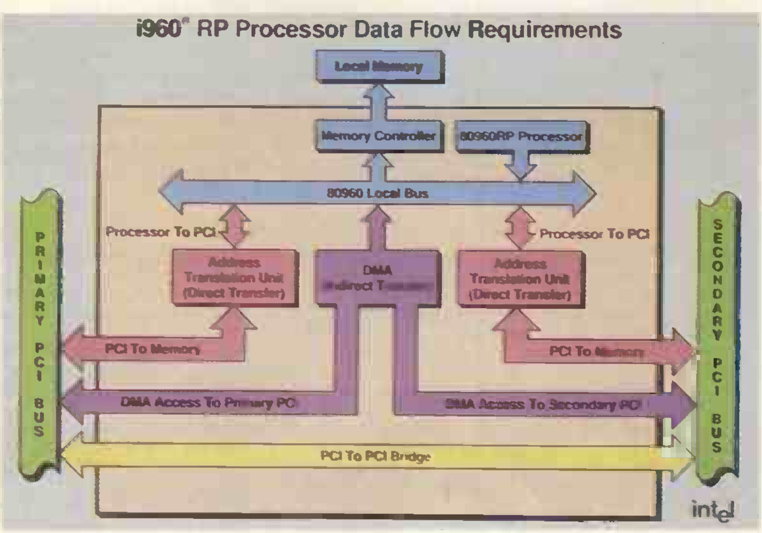
New pc co-processor helps i/o keep up

An intelligent i/o processor unveiled by Intel addresses more exacting applications demanded by present pc technologies such as x86 microprocessors and PCI local bus.

According to Elliot Garbus, Intel's technical marketing manager, Embedded Systems division, the best example of this is the evolution of the pc architecture for use in application server systems: "This move to client/server architectures has put great strain on the i/o."

Moreover, advancements in i/o technologies for net working and storage has also increased i/o demands. These include 100Mbit/s Ethernet and ATM; and for storage the Ultra-SCSI and 100Mbyte/s Fibre Channel standards.

To help i/o subsystems keep up with advances in x86 microprocessor performance, Intel is launching an intelligent i/o processor with dedicated PCI bus facilities.



Called the *i960RP*, the new device, device is based on the *i960Jx* core, which is rated at 31Mips (VAX) given a 33MHz clock. It incorporates a 4Kbyte instruction cache, and 2Kbyte data cache.

The i/o processor is designed to off-load the demands placed on the host processor, while offering i/o expansion via a secondary PCI local bus. In addition to the core and PCI-to-PCI bridge, the *i960RP* features two dma controllers, address translation units, pci bus arbitration logic, a memory controller and a I^2C interface.

Intel plans to announce further *i960* core upgrades and produce other, more tailored product variants.

Roy Rubenstein
Electronics Weekly

Cash transactions over Internet

UK electronic purse company Mondex is believed to be squaring up to its European rival Europay to provide secure, chip-based financial transactions over the Internet.

Last week Europay joined Forces with technology giant IBM to develop a smart card based payment system for commercial transactions over the Internet.

This mirrors the plans of Mondex, led by NatWest Bank, Midland Bank and BT, which according to one industry source wants to provide its own electronic funds transfer services over the Internet.

Europay, which is backed by credit card firms Mastercard and Visa, is making the running and IBM has already devised an open payment protocol, named Internet Keyed Payment Protocol – iKP – that is fully compatible with leading operating systems such as Windows, OS/2, Macintosh and others.

It has proposed this protocol to the World Wide Web organisation and the Financial Services Technology Consortium in an effort to make it an industry standard.

Another contending protocol is Visa and Microsoft STT's, which needs a proprietary system to run.

In order to allow electronic commerce over the Internet, specially designed card-readers fitting to PCs or multimedia boxes, are already in development. "A range of devices are under development: multi-functional tvs, fax pcs and television set combinations and not just one device," said a Europay spokesperson.

According to Adrian Cannon, general manager of smart card and reader provider CP8 Transac UK "We are likely to join in an alliance as the banks are our customers."

Svetlana Josifovska,
Electronics Weekly

Widescreen plus plus

The obvious advantage of a tv with a 16:9 *Widescreen* aspect ratio is that 4:3 ratio programmes – i.e. standard tv pictures – can be expanded to fill the screen. Initially, *PALplus* condenses pictures to a letterbox across the middle of the screen. It does this by extracting information and encoding it in the black area to enable the full image to be reconstituted. But this as a result of this, a 4:3 aspect-ratio picture is only expandable by adding blank lines.

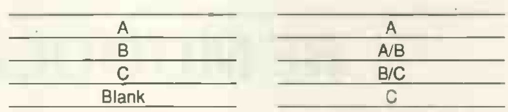
Now Philips is planning to overcome this with *WideScreen Plus* tvs. A 16:9 letterbox image contains only 432 active lines by comparison with the 576 of a full picture. Hence expansion requires one extra line for every three received.

With *WideScreen Plus* this is achieved by interpolation. The incoming picture signal is digitised and stored, then processed to create lines composed of elements from existing lines. For example, A, B, and C lines are written to screen as A, $^2/_3$ of A+$^1/_3$ of B, $^2/_3$ of B+$^1/_3$ of C, and C. This returns the vertical resolution to near normal.

Unfortunately, the system does not compensate for the loss of horizontal resolution – as *PALplus* does – but perhaps that is something for the future.

In the meantime *WideScreen Plus* will coexist with *PALplus* and enable standard letterbox programmes to be viewed with improved quality.

To partly compensate for the loss of one line in four when viewing standard images on Widescreen, an interpolator is used to spread useful information over all four lines.



Standard letterbox **Widescreen plus**

Neural net speeds up fingerprint matching

To catch a thief – or any other re-offending criminal – can take a matter of minutes with a little help from the UK-based DSP firm, Cambridge Neurodynamics Limited (CNL).

The company has developed a neural network Integrated Automatic Fingerprint Recognition System (IAFRS), that can match a ten fingerprint (tenprint) set to one from a total of a million stored in a central database.

The system comprises an encoder, image analyser and image matcher which are linked via a LAN. Once a tenprint has been scanned in, it is stored as a greyscale image in a 40:1 compressed format using C-Cube's *CL-550 JPEG* device. In order to receive/transmit noise-free images from/to remote locations, CNL also executes its proprietary compression algorithms on the images. Techniques such as the wavelet-based modified fast lapped transform encoding, and error resilient entropy coding are used. These avoid blocking artifacts that result from the compression, smoothing away image irregularities and spreading errors throughout the image.

The encoder/analyser is implemented as a board containing five 80MHz Texas Instruments *TMS320C50 DSP* chips. The processor are linked in parallel, each executing a different part of CNL's algorithm. For a basic unit these processors can deliver peak processing speeds of. 600Mips.

The unit extracts and encodes 16 of the most identifiable physical characteristics of the fingerprint. The unit uses a neural network to extract the statistics from a large scale image. Neural networks are used because they are seen as one of the best methods of data analysis for poor quality inputs, which can be unclear, smudged or superimposed.

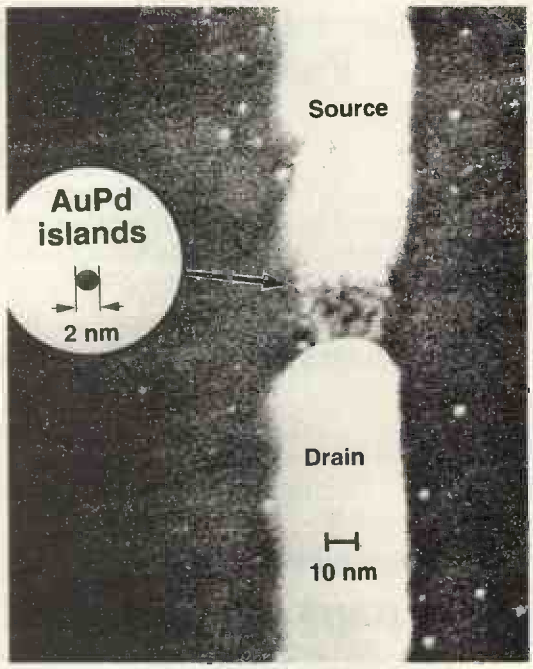
The matcher is based on the same processing board as the encoder, but runs a different algorithm to identify a kilobyte of information with one from a million-image library of fingerprint information. The matching of fingerprints can take 15 minutes on average, whilst the encoding takes approximately one minute per set.

The system, being scalable, can be expanded as required. Currently the IAFRS is being used by the South Yorkshire police, but its applications go beyond policing. It can be used in ballistics and DNA. matching (for which there is currently no extensive database), immigration and passport control, banking and social security applications.

Svetlana Josifovska
Electronics Weekly

One-electron transistor gets warmer

Hitachi's quest to build electronic circuits from structures that, use a single electron has taken a significant step forward. Researchers at its Cambridge lab,with Cambridge University, demonstrated a single-electron transistor device working at 77K rather than the 4.2K. Professor Haroon Ahmed of Cambridge University, said: "When we next talk about developments it will be when we can demonstrate room temperature operation. We don't want these things to remain novelties but have exciting, real applications." The 77K device is a thin metal wire on a silicon-dioxide substrate. The gold-palladium islands are the so-called Coulomb blockades letting single electrons cross the gate. The islands have a "world record" diameter of 2nm. But to get room temperature operation the dots would have to be five times smaller, said Ahmed. "The important thing is to reduce the capacitance of the Coulomb blockade by using different materials rather than making them much smaller." The group has also demonstrated an inverter at 4.2K and is striving to build more complex logic primitives.



Researchers have demonstrated a single-electron transistor operating at 77K rather than 4.2K and work is underway towards a room-temperature version.

VHDL model standard suffers further delay

Work in the IEEE group striving to specify analogue system modelling extensions to the VHDL digital design language has slowed to a crawl for the second time in 12 months. Finalised specifications of the 1076.1 extensions is now unlikely before the early next year, pushing ratification to at least end 1996.

Onlookers from the EDA industry and Asic user community are deeply frustrated with the latest delays. It was hoped, in September last year, that a language specification could have been released at the time of last week's Design Automation Conference.

Stan Krolikowski, senior member of the IEEE VHDL community, is disappointed. "It's really frustrating: we have all the concepts mapped out but there seems little progress in getting them implemented," he said. "The mood of the working group seems to be changing from democracy to mob rule with senior members seemingly unable to work together."

Krolikowski says the danger is that analogue extensions to the Verilog language – about to become an IEEE standard – will be available first, confusing potential users. "At this rate we'll have analogue Verilog out a long time before 1076.1, and that will really change the picture."

But Ernst Christen, vice chairman of the 1076.1 committee, defends progress saying the process is slowing down because it depends on volunteer work. "We have to let the volunteers catch up with the two of us being paid to develop the language. That's what makes the schedule unpredictable. But we absolutely have to have consensus because we are bound by the IEEE process and because the language has to survive without being continually modified."

Christen says the current bottleneck is validation of the fledgling language and not design of the language itself. There are quite a lot of open issues such as frequency domain ac analysis – we haven't seen anything substantial there yet – but 75% of our investigations have reached their first drafts and 20% have produced results and been provisionally approved."

But in some areas arguments still rage about language semantics and syntax. "Language design is an art, not a science," said Christen. There are various ways of doing the same thing and it is the task of the language design committee to weigh the benefits of one approach over others. We have to consider if it is consistent with the main body of the existing digital language, whether it provides sufficient functionality and usability. It's not a black and white process."

TDM optical net promises multi-Gbit/s performance

A fibre-optic network transmitting multi-gigabit per second data using optical time division multiplexing has been demonstrated by researchers at BT laboratories.

The BT network avoids the bottleneck associated with converting optical signals to an electrical form at switching centres. Instead, long distance transit traffic passes through switching centres in optical pulse form. By performing channel multiplexing and extraction processes at high speed using optical devices, much higher aggregate data rates can be attained, potentially as high as 100Gbit/s.

Andrew Ellis, senior professional in networks research said: "The key advantage of this experiment is not the long distance capability or even the high capacity that can be achieved, but the ability to route traffic flexibly on demand without needing a lot of costly electronic switching." ■

RESEARCH NOTES

Jonathan Campbell

Simple electronics takes a leap

Application of relatively simple electronics is being credited with helping discover previously unknown populations of frogs. Ecologist Michael Dorcas, working with his electrical-engineer father Eugene have built a 'frog-logger' using a tape-recorder that automatically turns on and off at particular times and uses a voice clock to audibly time stamp the start of each sampling interval.

In what is described as a dramatic "success" for the device, a breeding population of barking tree-frogs has been discovered where the species was previously represented by a single road-killed specimen — despite 19 years of field work. Other extensive monitoring work is being carried out with colleagues at the Ecology Laboratory, operated by University of Georgia.

One problem remains with the 'frog-logger': you have to sort through a lot of tapes to find the right croak. So Dorcas is now working with Ontario Hydro Technologies to perfect computer software that can identify animal vocalisations, making it much easier for scientists to analyse their data.

Fortunately, frogs are easier to analyse than birds, say, because their are fewer species and their calls are less variable. That is fortunate, because scientists believe that monitoring the diversity and relative abundance of frogs and toads could serve as an overall indicator of the entire environment.

Frog dying to be logged.

Morphing robot with the amorphous role

The concept of a robot that can change its form in response to its environment has been the basis for everything from a hit children's cartoon series to the blockbuster film Terminator II. But a robot system built up from interlocking body blocks that can slide over each other could be the start of a shape changing reality.

The interlocking-cube idea has been proposed by Joe Michael of London. He has just patented a design of cube with a mechanism that push out and retract wedges from all its faces. The wedges lock into place in other cubes and a complete structure can be built up like a series of Lego bricks.

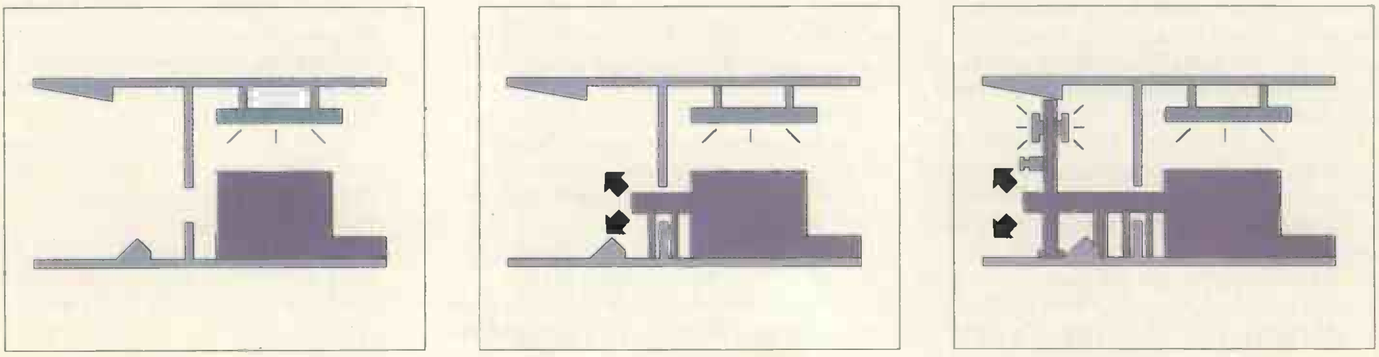
But the clever part has been to design a mechanism to allow the cubes to move over each other. Michael uses a gear wheel to engage with the serrated edge of a wedge when it is place, enabling the wedge – and hence the block – to be moved.

Obviously movement is only possible when locking wedges perpendicular to the direction of intended movement are disengaged. But Michael says that the design will enable the blocks to move while holding tools etc.

The first working prototype has yet to be built and so most of the practical problems remain to be faced. But Michael is supremely optimistic at present, proposing his design as suitable for use in microsurgery, using nanorobots, up to massive machines to be used for bridge building.

Whatever the actual potential, the attraction of a robot able to flow itself through a keyhole and then reform itself in a room has obvious attraction – to film makers if no-one else.

Sliding cube structure could allow a robot to slip in and out of almost anywhere.



Induction loopy

Odd electromagnetic effects surface every so often, prompting scientists to review induction theory and re-examine Faraday's law. Now apparent generation of a positive emf as well as back emf in experiments on a motor operated by an *LCR* circuit are causing physicists to scratch their heads once more.

The effect has come to light as a result of work by Osamu Ide to develop a motor operated by discharge of a capacitor in an *LCR* circuit ("Increased voltage phenomenon in a resonance circuit of unconventional magnetic configuration, *J App Phys*, Vol 11, pp. 6015-6020).

Ide's motor is designed to make use of the force of attraction between a current-carrying coil and a movable magnetic core. That attraction, resulting from a capacitor discharge, is converted to a rotary motion, and unconsumed magnetic energy is recycled as electrical energy by recharging the capacitor.

But what Ide has noted, is that the capacitor recharge voltage depends on the precise configuration of the system, so that under a certain magnetic configuration, it is bigger than normal theory expects.

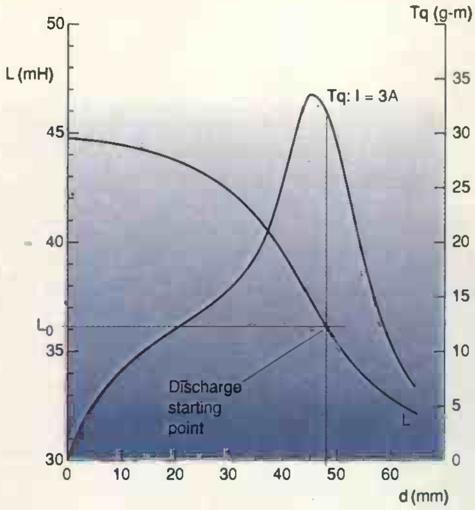
The construction of Ide's motor is unusual, but relatively straightforward. He starts with a basic *LCR* circuit, with a silicon controlled rectifier (scr) used as a switch so that instead of the usual voltage and current oscillations, a negative charge is retained in the capacitor after discharge. The scr automatically turns off after the half-cycle current, recharging the capacitor to a recharge voltage which is smaller than the initial voltage due to resistance loss in the circuit.

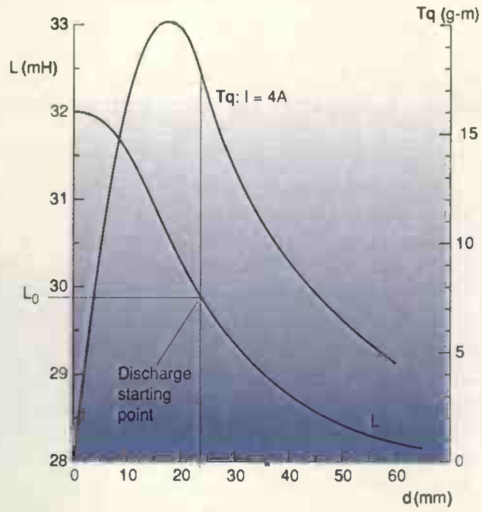
Ide has then replaced the inductance with two pairs of electromagnets, which form the stator around a rotor having two ferromagnetic cores. The magnetic fields attract the cores, producing motion, but as the core approaches the magnets, combined inductance increases and the magnetic flux moves, affecting the discharge current and recharge voltage.

The total recharge voltage should decrease because the system produces mechanical output as the core moves, but Ide found an increase in the average current and recharge voltage when the magnetic field in the electromagnets was opposing, ie N-S:S-N).

His explanation is that his motor demonstrates that two types of emf (actually suggested by other workers 50 years ago) – motional emf caused by cutting of the magnetic flux and induced emf caused by Faraday's laws – are independently at work.

In Ide's motor the two emfs have a contradicting effect, with the motional emf having a positive effect on the recharge voltage.





How inductance of the electromagnet relates to displacement of the core from the elecrtromagnet. Rate of change of the inductance depends on the magnetic fields and is greater in the attracting mode.

Silicon technology will catch sun-spot flares

Silicon processing techniques developed from microchip manufacture are being exploited by Sandia National Laboratories to make telescope imaging grids with features up to 100 times smaller than possible using traditional manufacture.

Liga – a German acronym for lithography, plating and molding – is being used at Sandia to create moulds for electroplating deep structures made of metal, or for casting ceramics, plastics or other polymers.

The telescope will be used on a Nasa mission to image the sun's solar flares, tracking particles that burst in a sudden release of power from areas where sun spots arise. Neither gamma nor neutron rays produced by these solar outbursts have been imaged before.

Using liga, the telescope will be less than half the length initially anticipated, measuring overall less than 2m long. The smaller size allows it to be launched on the cheaper spacecraft Pegasus. Orbiting almost 400 miles above Earth, the telescope will capture images through a series of 12 mesh grids and detectors. As the telescope rotates through space, it will provide a three-dimensional view of radiation from solar flares, in a manner similar to a pinhole camera. Proof-of-concept grids have already been electroplated at Sandia, and the next step is to produce perfect, large-area structures, and assemble and characterise flight instruments.

Other Sandia liga projects include a disk drive suspension arm for IBM and a high-performance stepping motor that involves creating a large nickel-iron alloy part with fine resolution teeth.

Space camera ...that doesn't take up any

A device claimed to be much smaller and cheaper than current state-of-the-art electronic imaging systems but comparable in performance is being developed by Nasa as part of its space program.

The camera on a chip is a considerable leap beyond current charge-coupled devices, according to Eric Fossum, leader of the team who developed the active pixel sensor at Nasa's Jet Propulsion Laboratory (JPL), Pasadena.

JPL has so far signed a technology cooperation agreement with AT&T Bell Laboratories, while several other companies are said to be seeking licensing agreements to commercialise the technology.

Up to now, ccds have been used mostly in video camcorders and space-craft, and have led to relatively low-cost, compact imaging systems compared to Vidicons and other tube technology. CCD technology has also advanced as the microelectronics industry has improved its quality and fabrication techniques. But ccd devices with a million pixels, or picture elements, are still expensive to make, costing about $1000 per million pixels in low-volume applications.

Active pixel sensors, by contrast, are made with main-stream cmos microelectronics technology which can reduce the cost to under $200 per million pixels.

The active pixel sensor developed at JPL for space applications is designed to use less power and has lower susceptibility to radiation damage in space.

Solid progress to portable audio

Are you concerned that the mechanical component of current audio technology is limiting your in-sport audio activity? Do you find you are frustrated with the limited choice (?) of tape and cd formats? Then a prototype solid state audio system being developed by researchers at NEC could be just for you.

The basis of the system, which the Japanese team is calling silicon audio, is a semiconductor memory card equipped to store compressed audio data ("The Silicon audio: an audio-data compression and storage system with a semiconductor memory card," *IEEE Transactions on Consumer Electronics*, Vol 41, No 1, pp.186-194).

Advantages seem to be that there is no need for any mechanics; unlike tape, random access is possible, while unlike cd, users can re-record.

A general purpose dsp and specially-designed gate array chip, is used for decoding data on the card and the system will both play-back and record.

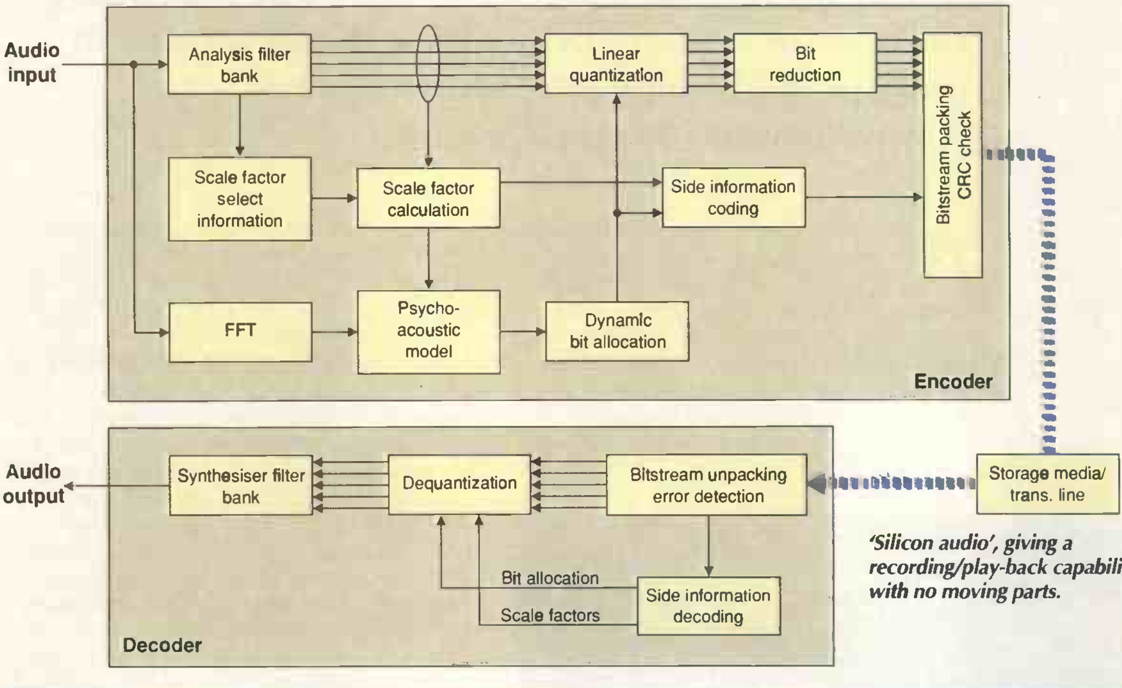
For audio compression, the designers propose using the mpeg/audio layer II algorithm, handling a 20kHz bandwidth.

Length of recording time is obviously a function of the size of memory card available and the employed bit-rate. In the prototype work, encoding is being carried out at 96kbps/ch, giving a reasonable compromise between recording capacity and sound quality.

So far the developers have built a play-back machine able to store 12min of music on a 16Mbyte card. But their eventual aim is 60min recordings, which they say is well within the capability of fast evolving card memory technology.

Silicon audio is being proposed as a complete portable multimedia system spanning teaching, guides, news and books, and integrating sound, speech, text, still pictures and even images.

However, the current close links between technology progress and fashion have not been missed by the researchers either. In parallel with detailed electronic development, the team has been trying out a variety of packaging colours and graphics to attract the eye of potential consumers. Only when that is finalised, will we be able strap on our new toy and go "jogging, skiing, biking and even roller-blading", while we listen to something other than our own laboured breathing.



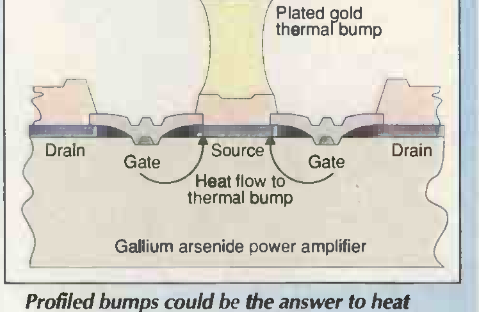
'Silicon audio', giving a recording/play-back capability with no moving parts.

Hot GaAs technology gets a cooler

Widespread application of gallium arsenide devices in portable telephones phones is testament to their low noise amplification characteristics – advantages that are expected to see GaAs use spread to high speed switching networks and digital logic. But difficulties in removing the heat generated in GaAs power amplifier chips has so far led to compromises in design which have impacted on yield.

But work at Motorola looks to have gone some way to solving that problem, by shrinking the heat transfer path from the hot zones in the active area of the device to the heat sink, and bypassing the bulk of the GaAs itself. ("A novel active area bumped flip chip technology for convergent heat transfer from GaAs power devices, Debabrata Gupta, *IEEE Transactions on Components, Packaging and Manufacturing technology – Part A*, Vol 18, No 1, pp.82-86).

The technique relies on plating bumps of gold onto the active area of a flip chip device. Bumped amplifier flip chips bonded to aluminium nitride substrates can dissipate heat fluxes up to 300W/cm^2 with an acceptable rise in junction temperature. ■



Profiled bumps could be the answer to heat dissipation problems in GaAs power amplifiers.

Feel the bass

With this active 100W sub-woofer design – featuring a –3dB point of 20Hz – Jeff Macaulay once more demonstrates how inadequacies in loudspeaker cabinet systems can be compensated for electronically.

Although the quality of audio electronics has now advanced to a state of near perfection, loudspeaker systems have not kept pace. The main problem is still the reproduction of bass – especially in domestic surroundings. Large speaker systems are not usually practical, but bass reproduction requires large enclosures, when conventional design techniques are applied.

Historically, most of the effort in speaker design has been expended on understanding how drivers and enclosures interact to produce predictable results. Generations of engineers have contributed to this work, culminating in the Theile-Small model, which now forms the basis of low frequency speaker design.

Briefly, the model compares the response of the speaker system to that of a high-pass filter. This enables the response to be accurately predicted. In fact the response curve of an unmounted speaker or one mounted in a sealed box is that of a second order high-pass filter in the bass. This response shape is dictated by the fundamental bass resonance, possessed by all speakers, between the cone mass and speaker surround compliance.

Although a driver may have a desirably low resonant frequency when measured in free air, mounting it in a sealed enclosure raises the resonant frequency because of the added stiffness of the enclosed air. Unfortunately some kind of enclosure is necessary because the radiation from the front of the cone is in antiphase with that from the rear. Consequently, because of the long wavelengths of bass sounds the two outputs diffract around the driver and cancel each other out, a kind of acoustic short circuit occurs.

Below the resonant frequency, response rolls off rapidly, reaching an ultimate slope of 12dB/octave. An obvious thought is to make the resonant frequency in free air lower to reduce the cut-off frequency in the enclosure. Unfortunately the efficiency of a speaker is proportional to the cube of its resonant frequency. Although you can trade off efficiency and bandwidth, the benefits are limited. The result is that most domestic speaker systems are limited to a –3dB point of 70Hz or more. At least an octave of bass is lost.

Driving amplifier

There is a parallel train of thought which seeks to modify the behaviour of a speaker system by altering the response of the driving amplifier. Three main areas of research have been pursued.

Motional feedback. Probably the most well known is motional feedback. Here a transducer is fitted to the speaker cone and the information obtained is used to control the cone in a classic feedback circuit. This technique has many advantages. The distortion and lack of bandwidth of a small box can be overcome. On the down-side, the system can be easily overdriven by bass transients and requires special transducers and electronics.

Damping. Another way of controlling the cone is based on the idea of damping. Power amps are designed with a zero-impedance output to control the speaker cone. The idea is that speaker resonance can be damped out by connecting a zero impedance source of signals. Unfortunately the voice coil resistance is always in series with the resonant sound radiation circuit.

Figure 1 shows the equivalent electrical circuit of a speaker mounted in a sealed box. Here the various reactive elements combine to produce a parallel tuned circuit. Resonance of these elements produce the well known rise in impedance at the speaker's fundamental resonance.

If these elements were to be fed directly from a zero-impedance source, the bass resonance would be completely suppressed. Although voice coil resistance cannot be physically removed from the circuit, an amplifier's output impedance can be made negative to cancel it. When this is done the speaker resonance is completely suppressed and the cone motion becomes independent of frequency.

The result is that the sound pressure level, spl, measured in front of the driver, increases at a steady 6dB/octave for a constant drive signal amplitude. In other words, the speaker acts as a differentiator. This is exactly how a perfect massless piston would behave – if such a device could be built. Furthermore, near perfect transient response is obtained and because the non linear compliance of the speaker surround is controlled this results in a considerable reduction of distortion.

All that is required to obtain a flat response from the speaker is to integrate the input signal. It is a sobering thought that if perfect speakers were available, our power amplifiers would have to become power integrators to drive them. In fact the only reason that loudspeakers have a flat response is that the bass resonance tilts the response downward by 6dB/octave above the resonant frequency.

As with motional feedback systems, special amplifiers with positive current feedback are necessary. To be effective, the speaker inductance also has to be cancelled if new unwanted resonant peaks are to be avoided. One advantage over the motional feedback system is that the speakers are stock items. No special features are required.

Equalisation. Although it is true to say that equalisation is used as an adjunct to both motional feedback and damping systems, it can also be used on its own. The main difficulty is that although Theile-Small software can predict the roll-off curve of any speaker system, suitable equalisers are rare.

Sixth-order alignment

Although equalised speakers are not in general use, the sixth-order alignment has gained acceptance. This speaker system is a reflex type using a vent to provide bass boost. The system is adjusted for an over-damped response curve which is then boosted back to flatness by an auxiliary under-damped second order high-pass filter as **Figure. 2** shows. Not only can the system extend lower into the bass region, but the subaudio noise limiting the system's excursion ability is rejected.

A new filter configuration for audio

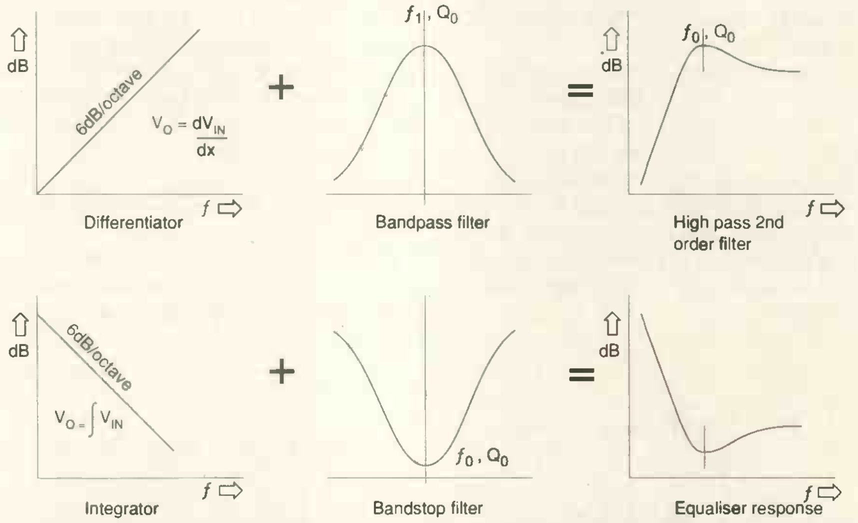
A standard second-order filter response can be simulated with a second-order bandpass filter and an integrator.

The bandpass filter needs to have the same centre frequency and Q as the second-order filter, or speaker response to be simulated. The flat response audio signal is fed first through the bandpass filter then an integrator. A second-order high-pass response is obtained.

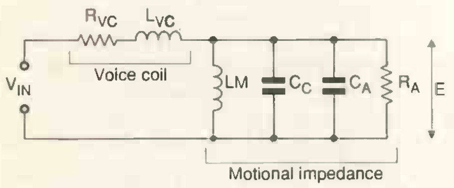
If the integrator is replaced with a differentiator, a second-order low-pass response is obtained. To simulate a reverse speaker response, all that is required is to replace the bandpass filter with a band-reject type of the same turnover frequency and Q.

A practical difficulty with such a circuit is insertion loss, but it avoids the use of high-gain circuits in the equaliser. As the gain of a high-pass filter at dc is zero, an infinite-gain amplifier would be needed to obtain full equalisation. In practice, this is not necessary since each halving of input frequency brings a fourfold increase in cone excursion to maintain the same sound pressure levels.

Obviously this soon gets out of hand and at sub audio frequencies, excursion requirements will eventually exceed the displacement limits of the speaker.



Synthesising a second-order high-pass filter from a differentiator and bandpass filter. For equalisation purposes, an inverse response can be generated by combining a bandstop filter and integrator.



Equivalent circuit to speaker in sealed enclosure.

$LM = B^2L^2 \cdot \frac{C_S C_B}{C_S + C_B}$

Where C_S = cone suspension compliance
C_B = compliance of enclosed air in box

$C_C = \frac{M_a}{B^2L^2}$ Where Ma = mass of air load on speaker

$R_A = \frac{B^2L^2}{\pi A}$ Where πA = radiation resistance applied to speaker

E = BLU where
B = magnetic flux density
L = voice coil length
V = velocity of voice coil

Fig. 1. Parallel tuned circuit – the equivalent electrical circuit of a speaker mounted in a sealed box.

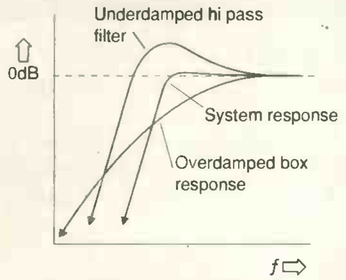


Fig. 2. Sixth-order alignment. System response is extended by an external filter.

Equalised speakers have most of the advantages of the other systems described. Transient response is improved because this is simply a function of the rate of roll-off imposed by the system. Provided that sensible limits are observed, the extension in bass response can be as great.

Unfortunately, the total harmonic distortion of the system is not improved. However this is no great advantage since it has been shown that 40% of second harmonic distortion below 80Hz is inaudible on programme material. Modern drivers tend to produce far less odd harmonic distortion than older models. These facts, taken together, indicate that sub-resonant operation works well – a fact borne out by experience.

Circuit details

Electronics textbooks are full of circuitry to produce the classical filter configurations. But a reversed second-order response is not amongst them. The only filter that I know of that can simulate this response is the Linkwitz filter.

Unfortunately this filter tends to be component critical, and as often as not, components have to be selected by experimentation – a time consuming process. I find this unsatisfactory, so I have developed a new type of filter. The key to understanding this new circuit is that the standard second-order filter response can be simulated with a second-order bandpass filter and an integrator, see panel.

A practical filter of this type is shown in **Fig. 3**. Since band-reject filters tend to have awkward component values, I have used a multiple-feedback bandpass filter, or mfb. Band-reject response is produced by subtracting the bandpass response from the input signal in A_2. Since the mfb filter is inverting this can be done with a shunt feedback amplifier without fear of unwanted signal interaction. The integrator function is performed by shunting the feedback resistor. Although sealed box systems can be very effective when operated in sub-resonant mode, greater efficiency would be a distinct advantage.

The Microreflex system described in an earlier article[1] is efficient and has excellent bass performance for its size. It makes use of the reflex principle but in a grossly undersized box. Such a system exhibits a roll-off characteristic closely resembling those of a closed box. This means that it is easy to equalise.

To understand the Microreflex enclosure requires some understanding of standard reflex types. In a reflex enclosure, rear radiation from the cone is used to reinforce the output from the front. As discussed earlier, radiation from the front of a speaker cone is in antiphase with that from the rear.

For rear energy to be useful, some method has to be found to phase invert the rear sound. In a reflex enclosure this is done by making the enclosed air resonate at a fixed frequency by means of a vent. The mass of the air in the vent resonates with the compliance of the air within the enclosure. It forms a mechanical tuned circuit which is excited by the rear of the cone.

Above the enclosure resonance, the vent's radiation is in phase with the front of the cone. Below the resonance the vent and cone radiation move out of phase producing a rapid bass roll-off. The system operates as a fourth-order high-pass filter.

At very low frequencies, the cone is unloaded by the air in the enclosure and thus large sub audio cone fluctuations can occur, unless suitable filtering is applied. The advantages of a reflex system over a sealed box are also applicable in the Microreflex system. Primarily the speaker's resonant frequency is virtually unchanged by reflex loading, leading to lower distortion at bass frequencies. Secondly the excited enclosure resonance acts as a heavy acoustic load reducing both cone excursion and distortion.

Most importantly however is the increased efficiency it provides relative to a sealed enclosure. Unfortunately it is a waste of time trying to extend the bass response of a properly designed reflex enclosure below its cut-off frequency. All that happens is that the cone moves further but the antiphase vent output cancels any improvement.

To take advantage of the reflex system, the enclosure has to be redesigned. It so happens that when the enclosure is made small enough the natural fourth order filter roll-off characteristic changes to a second order roll-off, just like a sealed enclosure.

Active equalisation is used in the sixth-order speaker system discussed earlier. Enclosure resonance is lowered and the overall response is then boosted flat again. Before equalisation is applied, the response of the system resembles an over damped fourth-order filter. To obtain a flat response the enclosure has to be carefully matched to the driver. Although sixth-order systems are usually smaller and produce deeper bass, the design still requires set enclosure volumes and driver parameters.

When the enclosure size is too small, as in the Microreflex, the response becomes that of a second-order filter. This yields response curves that are far from flat. However this is not a limitation. The equalisation applies a drive signal that corrects both the amplitude and phase of the natural response, rendering

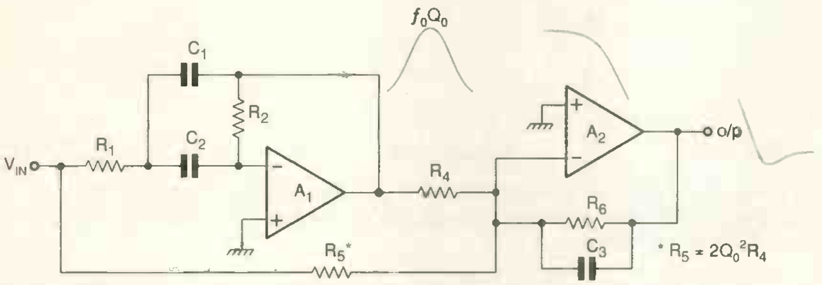
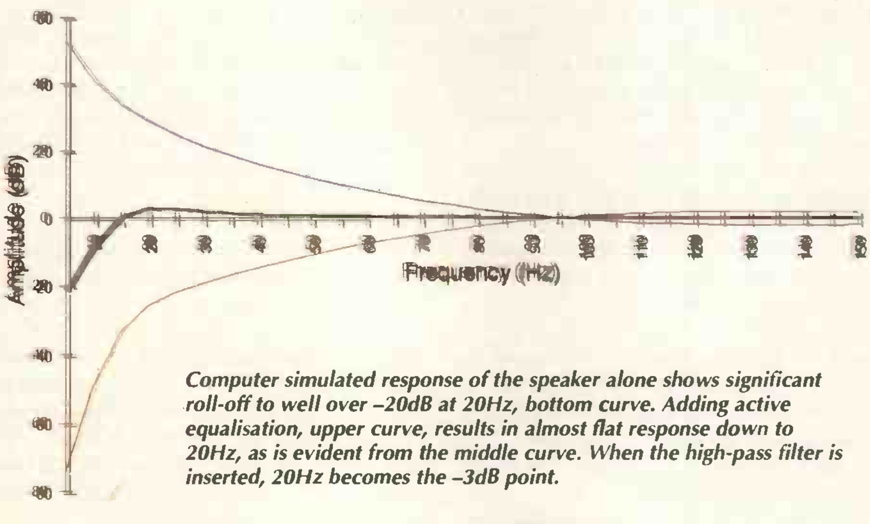


Fig. 3. Amplifier A_1, with R_1, R_2, C_1 and C_2, forms a bandpass filter with the same turnover frequency and Q as the speaker system. The required bandstop response is obtained by subtracting the bandpass from the input by A_2. Capacitor C_3 and R_6 produce an integrator response to produce equalisation.



Computer simulated response of the speaker alone shows significant roll-off to well over –20dB at 20Hz, bottom curve. Adding active equalisation, upper curve, results in almost flat response down to 20Hz, as is evident from the middle curve. When the high-pass filter is inserted, 20Hz becomes the –3dB point.

the system response flat.

Small reflex enclosures have a useful characteristic. That is, that the second-order roll-off extends to below the enclosure resonance. Investigation reveals that the enclosure resonance's Q has been reduced, leading to lower phase shift, and the vent output supplements the driver down to and below the enclosure resonance. This is important because it allows bass to be extended towards the sub-audio region. Provided that the excursion limit of the cone is observed, the performance of the system can be predicted with the standard Theile-Small model.

This article describes a small-sized active subwoofer with a –3dB point extending down to 20Hz. Computer simulation, and testing show that the system is thermally limited within its working range.

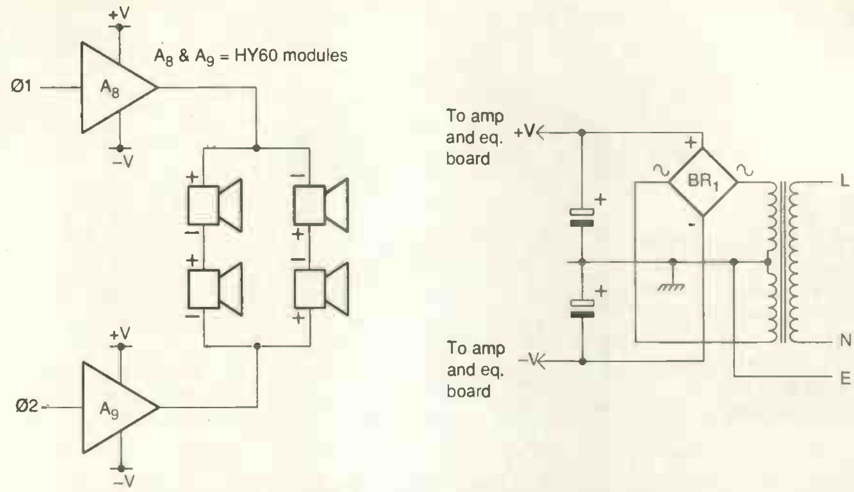


Fig. 4. Wiring the drivers in series/parallel results in an overall impedance of 8Ω from four 8Ω drivers. Pairs of drivers are driven in antiphase and mounted face-to-face.

Subwoofer systems

To be effective, a subwoofer needs to shift a large volume of air at low frequencies. Two approaches are possible. Either a large diameter driver can be used, or several smaller units operating in phase. High efficiency is also helpful to offset the amount of equalisation required. Large speakers with the right qualities are hard to find so four 8in *PF81HR* types are used.

Speaker size is dictated by the need to reflex the enclosure to a suitably low frequency. A 1.5ft^3 enclosure was chosen, tuned to 40Hz. By wiring the drivers in series/parallel, Fig.4, an 8Ω impedance was obtained with a reference efficiency of 97dB/W. The resulting raw frequency response of the system was predictably dire closely resembling a closed box system with a Q of 1.3 and a resonant frequency of 124Hz.

Examination of the alignment shows that the vent and driver are still operating in phase down to below 20Hz. The vent provides several decibels more in the deep bass than the same system operated as a sealed box. Moreover the system is capable of providing 95dB spl at 1m at 30Hz. Further examination of the large signal characteristics showed that the system was thermally limited, rather than displacement limited, to 20Hz.

With correct equalisation, the bass response can be rendered flat. Auxiliary bass filtering below 20Hz ensures that the displacement limits imposed by the drivers are not exceeded. Although the vent resonance reduces distortion from the speakers at low frequencies they are used in push-pull pairs. This effectively eliminates even-order distortion from the system.

Integration into an existing system

The other interfacing problem posed by subwoofer systems is integrating it into an existing set-up. Having solved this problem several times, I have found some useful practical guidelines.

Imposing a low-frequency crossover between the subwoofer and existing speakers is counter productive. To prevent any problems with the stereo image the subwoofer's range should not exceed 120Hz. At these frequencies the existing speakers will be already beginning to roll off with unpredictable amplitude and phase shifts. Add to this equally unpredictable resonant modes in the listening room and you have a problem that virtually defies analysis.

The best solution that I have found is to leave the existing speakers alone and make the subwoofer roll-off variable. A low Q second order low pass filter is the best. With a little adjustment seamless integration of the system is possible.

The power rating of the system is 100W continuous. With four drivers there is a choice as to the impedance chosen. By wiring these in series/parallel I kept this at 8Ω to provide standard loading. However it is not strictly

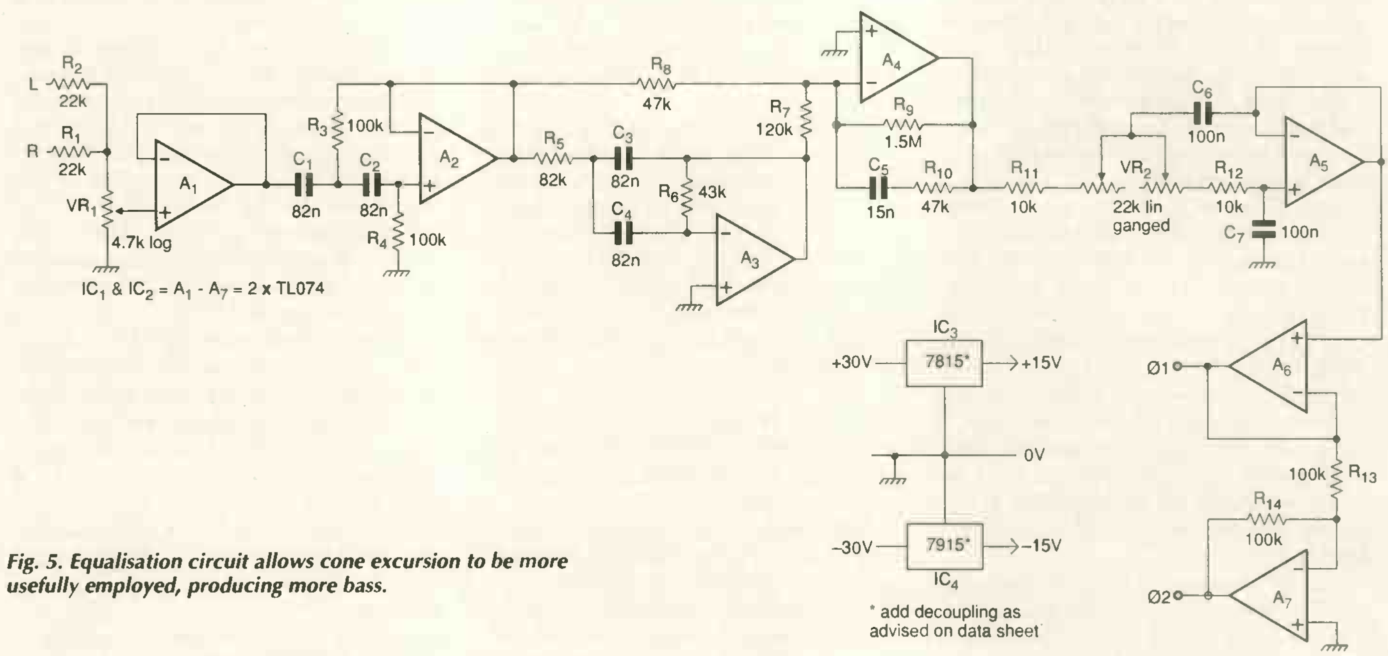


Fig. 5. Equalisation circuit allows cone excursion to be more usefully employed, producing more bass.

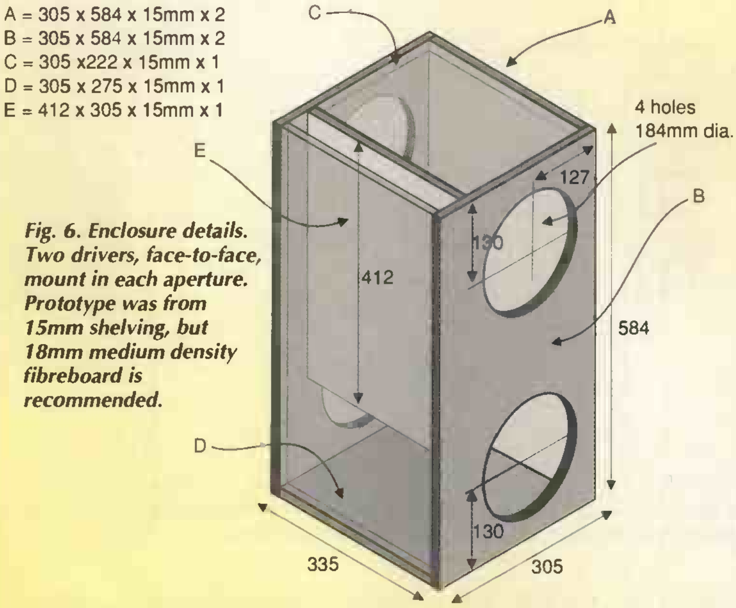


Fig. 6. Enclosure details. Two drivers, face-to-face, mount in each aperture. Prototype was from 15mm shelving, but 18mm medium density fibreboard is recommended.

Frequency (Hz)	Relative response (dB)	Max power input (W)	Max infinite baffle response (dB)
5	–74.59	49.37	38.95
10	–49.16	39.66	63.43
15	–33.90	35.64	78.22
20	–25.80	71.03	89.32
25	–22.00	100	94.60
30	–19.31	100	97.29
35	–16.97	100	99.63
40	–14.84	100	101.76
45	–12.86	100	103.74
50	–11.02	100	105.58
55	–9.31	100	107.29
60	–7.72	100	108.88
65	–6.24	100	110.36
70	–4.87	100	111.73
75	–3.62	100	112.98
80	–2.50	100	114.10
90	–0.65	100	115.96
100	0.66	100	117.26
120	1.85	100	118.45
150	1.95	100	118.52
200	1.31	100	117.91

Theile-Small analysis of the subwoofer shows high acoustic output above 35Hz. Additional roll-off imposed by 20Hz high-pass filter ensures 100W rating throughout the range. The PF81HR has a VAS of 1.462ft³, a QTS of 0.45 and an f_s of 45Hz.

Bass units – special offer

Semiconductor Supplies International is offering sets of four PF81HR bass units needed for the subwoofer at the special price of £42.77 – exclusively to *EW+WW* readers mentioning this article. This price is fully inclusive of vat and UK next-day delivery. Normally, the set would retail at £55.27 fully inclusive.

Send postal order or cheque to Semiconductor Supplies International at Dawson House, 128 Carshalton Road, Sutton, Surrey SM1 4TW, or ring or fax with credit card details, tel 0181 643 1127, fax 0181 643 3937. Note that SSI's part number for these speakers is LO31.

necessary to use 100W drive. In practice any amplifier which can deliver 30W or more is suitable. In a domestic environment, using more than a few watts input will not impress the neighbours.

The complete circuit

The whole circuit, Fig. 5 is built around two quad *TL074* op-amps. Right and left channels are passively mixed via resistors $R_{1,2}$ across the level potentiometer VR_1. It makes sense to obtain the input from the right and left speaker outlets of your main amplifier. This way the sub woofer correctly tracks the system output.

The signal is first buffered by A_1 before being fed to the high-pass filter built around A_2. Component values chosen for this circuit establish the -3dB point at 20Hz. This heavily attenuates out-of-band sub-audio signals. Without the filter these may have sufficient amplitude to push the drivers beyond their excursion limits. Not having the cones aimlessly pumping in and out allows the cone excursion to be more usefully employed, producing bass.

From A_2's output, the signal feeds the response equalisation circuitry built around A_3 and A_4. Op-amp A_3 forms a band-pass filter with the same turnover frequency and Q as the system. Output of the bandpass filter is subtracted from the input signal to produce a bandstop filter.

Because A_3 inverts, A_4 acts as a differential amplifier. Resistors $R_{8,9}$, together with C_5, integrate the signal. This produces accurate equalisation, flattening the system's response.

Final equalisation is provided by a variable low-pass filter, built around A_5. This strips off out of band signals . Potentiometer VR_2 – a dual ganged type – makes this stage's cut-off frequency continuously variable between 40 and 120Hz. The Q of the filter is set at 0.5 for the best transient response within the subwoofer's passband.

Although the subwoofer gives a good account of itself when driven from a low-power amp a reasonably substantial power input is required to drive it to its limit. A 100W amp was required and this was produced by using a pair of *ILP HY60* modules in bridge mode.

To use the amps in this way requires a phase splitting circuit, A_6 and A_7. Input signals are fed into A_6's non inverting input. As this is wired as a buffer, the output signal is applied across R_{13}. Op-amp A_7 is wired as a shunt-feedback amp. With R_{13} and R_{14} having equal values, A_7 output is equal to A_6 but shifted through 180°. The circuit is powered by a standard ±15V power supply using *7815* and *7915* regulators.

Enclosure details

An ideal material for the enclosure is 18mm medium density fibreboard, mdf, Fig. 6. The material is not too critical, but ensure that the internal volume is kept constant if you use material of a different thickness. The enclosure must be completely airtight.

I implemented the tuning vent by forming a partition within the cabinet and an undersized top panel. Although the vent is rectangular, it operates as a circular type of equivalent area.

The speakers are mounted in two pairs. One pair with the magnets external to the case, the other pair conventionally. Its also important to ensure that the drivers make an airtight seal to the case. With the rear mounted pair they can be simply screwed up against their gaskets.

Both the front pair and rear pair are wired in series. Both pairs are then connected in parallel for an 8Ω load. The rear pair of speakers are wired out of phase with the front pair to give push pull operation.

A neat and airtight way of taking the connections through the case is to use M5 screws fitted with solder tags. The electronics can be mounted inside the enclosure if need be. A rather neater result is obtained if the circuitry is mounted in an ABS case which is then screwed onto the rear panel.

This last method gives easy access to the electronics without having to rip apart the case. It also makes the task of keeping the enclosure airtight easier.

Finally I would commend this design to those of you who yearn for a decent bass response, but who are unwilling to give house room to large enclosures. Everyone who has heard the sub-woofer has been favourably impressed. ■

Reference

1. Macaulay, J., Bigger bass – smaller box, *EW+WW*, June 1995, pp469-475.

BULL CLEAROUT SALE

WIRELESS VIDEO BUG KIT Transmits video and audio signals from a minature CCTV camera (included) to any standard television! All the components including a PP3 battery will fit into a cigarette packet with the lens requiring a hole about 3mm diameter. Supplied with telescopic aerial but a piece of wire about 4" long will still give a range of up to 100 metres. A single PP3 will probably give less than 1 hours operating time. £99 REF EP79. (probably not licensable!)

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Discrete-time key to dsp

Real-time digital signal processing calls for an understanding of the principles of discrete-time linear systems. Howard Hutchings explains the theory, and then offers a practical way to try out the results.

Interfacing to a pc and real-time programming undoubtedly call for engineering ingenuity.

But successful real-time digital signal processing requires something more – an understanding of the principles of discrete-time linear systems.

The fundamental mechanism of sampled data manipulation is time-domain convolution, or z-domain multiplication. These particular topics are well-understood and have been extensively documented, typically with the assistance of mathematics.

But with only a small investment of time it is possible to demonstrate the principles of dsp in action in a dramatic way.

The approach is to use a pc to examine the time- and spectral-performance of a sampled sinusoid detailed as a rational z-function, to be subsequently developed as a particular case of a general discrete-time linear system.

Convolution is not needed as any real-time numerical manipulation uses recursive methods, while engineering design and performance of the discrete-time linear system is introduced using the z-transform of a sampled sinusoid.

As a result there is no requirement for synchronised data transfer between the a-to-d and numerical processor, and engineering principles are not compromised.

Equi-spaced samples generated by the sinusoidal digital generator are written to the real world via an 8-bit d-to-a, while the resonant frequency can be adjusted by selective scaling.

Advantage of the method is that it demonstrates several concepts in a visual way, using a worked example. Sampled time-domain performance is visualised on an oscilloscope, and the frequency-domain behaviour of poles is visualised on the z-plane and displayed using computer graphics. Other topics discussed include discrete Fourier transforms, the audible and visible effects of aliasing in time and frequency, unit-pulse response of an IIR system, and the real-time realisation of a pc-based digital filter programmed using *C*.

Digital oscillator

First step is to examine how to use a numerical system to generate sine waves, exploiting a real-time pc-based digital oscillator, interfaced to the real world through an 8-bit d-to-a converter.

Typically, such a system would compute the required output samples ahead of time and store them in a wave-form look-up table to speed up the process of real-time wave-form generation. But rather than follow such a well-trodden path, we are going to look at a recursive real-time realisation that is more relevant to the characteristics of a discrete-time linear system.

It is developed from the z-transform of a sampled sinusoid:

$$H(z)=\frac{z\sin\Omega_0}{z^2-2z\cos\Omega_0+1}$$

To understand the behaviour of this transform detailed as a linear system, we should first undertake a brief review of the properties of linear systems, and then follow this with a discussion of discrete summation to assist the notion of open-and closed-forms. Then, the effects of the sampling process can be considered, and real-time performance demonstrated visually and audibly using computer-managed instruction.

Frequency preservation

Many signal processors do not possess memory. Frequently the purpose is simply that of signal amplification ($A > 1$) or attenuation ($A < 1$). The output signal should be a faithful reproduction of the input, represented by the scalar product: $y(t) = Ax(t)$. Impulse response $h(t)$ of such linear systems would ideally consist of a single weighted impulse $h(t) = A\delta(t)$ **(Fig. 1a)**.

Instantaneous output of this system is an

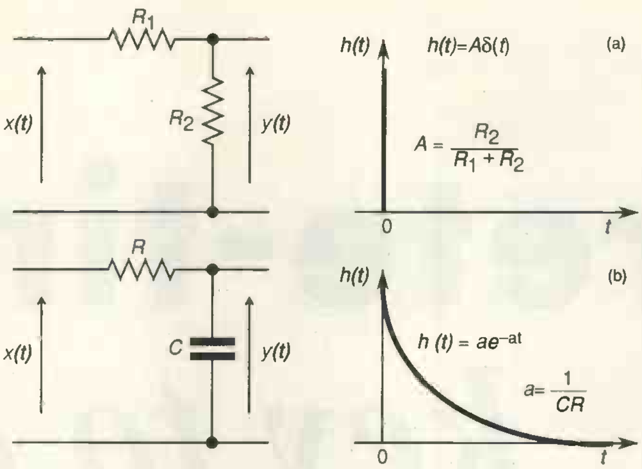

Fig. 1. Consideration of the impulse response of the spectral performance, from a time-domain perspective, improves understanding of the effect of memory on frequency-domain filtering.
1a. Memory less system characterised by a single scaled impulse.
1b. System with memory, characterised by a time-extended impulse response.

attenuated version of the input, and does not depend on past or future inputs.

Practical analogue systems seldom meet this requirement. Typically, capacitance, inductance and resistance modify the intended response at certain frequencies. A testament to the frequency-domain performance is inferred by examination of the characteristic of the time-domain impulse response. System parameters such as cut-off frequency, Q-factor and undamped natural frequency can all be determined directly from the historical record contained in the impulse response. A first-order response is detailed in **Fig. 1b.**

All linear signal processors exhibit frequency preservation so that no new frequencies are generated within the system. Any signal represented in terms of its component frequencies will be processed in a very simple way, with only the amplitude and phase of individual components being modified.

This is a key factor in relation to amplitude scaling and time delay – the parameters of a discrete-time linear system.

Filtering – a time domain perspective

Selectively manipulating input data, or filtering the signal in such a way that certain frequencies are especially favoured, requires memory. Numerous practical examples of analogue systems exhibiting this effect can be found in the technological world around us. For example, in a simple first-order system, a mercury-in-glass thermometer retains a history of the input step-function, long after the stimulus has been removed. More evocatively, striking a cathedral bell – a highly resonant second-order system with a Q-factor typically of the order of 1000s – will give rise to a prolonged response, audible long after the impulse has been removed.

Unwanted second-order effects in the passband of video amplifiers also give rise to ringing, an undesirable, but clearly visible effect.

Time-domain convolution and frequency-domain filtering can be thought of as complementary perspectives of the same processing operation. Here we will see how it can be desirable to present the characteristics of a linear system as a frequency-selective filter, specified either by the frequency response, transfer function, or by the impulse response.

Although convolution is a significant concept in many diverse scientific fields, the convolution integral plays a mainly theoretical role in analogue linear system theory. Discrete convolution is inherently practical. The convolution sum is frequently a direct realisation of a linear discrete-time system. It is the natural mechanism for describing the performance of processing devices operating on sampled signals, whether such systems are purpose-built electronic circuits or real-time programs implemented on a digital computer.

Open and closed discrete summation

The two principal types of digital filter algorithm are non-recursive, described by a finite impulse response (FIR), or recursive, characterised by an infinite impulse response (IIR).

A non-recursive filter calculates the current value of the output by combining the present value of the input, with scaled past values of the input.

But in a recursive filter, the current output depends on the present input together with combinations of scaled values of previous outputs and inputs.

Computational economy in real-time signal processing frequently involves minimising the number of calculations per sample, often by expressing the algorithm in closed form.

The sum of the infinite geometric series can illustrate the technique. In the equation $S = \sum 1 + r + r^2 + r^3 + \ldots$ we could simply summate all the terms on the right-hand side of the equal sign, and infer the sum as a limiting process. But, providing that $|r| < 1$ – a condition that applies to all the closed forms we are going to meet – the calculation can be simplified by taking advantage of the enumeration:

$$\frac{1}{1-r} = \sum 1 + r + r^2 + r^3 \ldots$$

This is an example of a discrete summation formula, where the left-hand side of the expression is said to be a closed form of the sum. Expressions of this form are so common in dealing with sampled-data sequences that we introduce a new symbol z^{-1} to represent unit time delays (for reasons that will become apparent in a later section).

Consider the sampled-data sequence $u(n) = 1, 1, 1, 1 \ldots$

To model delays algebraically, $U(z)$ can be expressed as a power series in z^{-1} as $U(z) = 1 + z^{-1} + z^{-2} + z^{-3} + \ldots$

Each sample in the sequence can be clearly identified and its appropriate position in time identified. In this case $r = z^{-1}$ so the series may be summed as:

$$U(z) = \frac{1}{1-z^{-1}} = \sum 1 + z^{-1} + z^{-2} + z^{-3} \ldots$$

Multiplying numerator and denominator by z gives:

$$U(z) = \frac{z}{z-1}$$

Physically, this may be clarified as the closed form of the sampled unit-step, the first sample occurring at $n = 0$.

How to use delay operator z^{-1} may be be made clearer by analysing the sequence $u_1(n)$, describing the sampled unit step delayed by one sampling interval, so that $u_1(n)$=0, 1, 1, 1, 1 ... The sampled data sequence may be modelled as $U_1(z)=z^{-1}+z^{-2}+z^{-3}+z^{-4}\ldots$ and summed as:

$$U_1(z) = \frac{z^{-1}}{1-z^{-1}} = z^{-1}U(z)$$

If the sequence is delayed by one sampling interval, then its closed form (or transform) is multiplied by z^{-1}. The delay property of z^{-1} can be expressed as a formal operation which fulfils the following requirements.

- Each numerical sample must be identified in some way, by reference to its particular time frame;
- Multiplication by z^{-1} should represent a time

Practical advice

Aim of this article is to develop a time-domain perspective on frequency-domain filtering, with the practical result being a computer-based account of the real-time performance of a digital resonator.

Engineering design and performance of the discrete-time linear system is introduced using the z-transform of a sampled sinusoid. This avoids the need for synchronised data transfer between the a-to-d and numerical processor – but not at the expense of engineering principles.

Equi-spaced samples generated by the sinusoidal digital generator are written to the real world via an 8-bit d-to-a and the resonant frequency is adjusted by selective scaling.

Principles can be developed into applications using computer-managed instruction.

There is no need to attempt any Fourier analysis or pole-zero plotting. The software will do it all for you.

delay of one sampling interval – moving a sample value from one time frame to the next. In general the z-transform of the sample $x(n-k)$ is $X(z)z^{-k}$.

Sampling complex exponentials

The kth sample value of a sampled sinusoid can be written as $f(k) = A\sin(\omega k + \phi)$, where k is an integer. The signal is characterised by three parameters: amplitude of the sine wave A, angular frequency ω, and angular starting point of the cycle in relation to the origin ϕ.

Euler's formula is a convenient way to describe $F(k)$ as the real part of a complex exponential $F(k) = Real\{Ae^{(j\omega k + \phi)}\}$.

The continuous phasor Acan be interpreted as a point that moves around the circumference of a circle of radius A in the complex plane. Sine wave $F(k)$ is basically one-dimensional, and can be modelled as the projection of the point on the Y-axis, rather than its position in two dimensional space.

A digital phasor can be visualised as the process of sampling a continuously varying phasor, represented by $\exp(j\omega t)$, at equally-spaced time intervals, represented by $t = kT$. Behaviour of a complex digital phasor sampling various signals is shown in **Fig. 2**.

Programs listings

The three *C* listings of programs referred to in the article, and written by Howard Hutchings, show dsp theory in action. For a hard copy of these listings, send an sae marked dsp to *EW + WW's* editorial offices (address on Comment page).

A disk containing the *C* program code is also available – for £10 to cover the cost of disc, copying, postage and packaging – from the same address.

Or we can include the three listings, free, on the *Interfacing with C* listings disk, obtainable from *EW + WW* for £15.

Listing 1: encourages the user to enter the parameter f_{cyclic}/f_{sample}. The associated complex conjugate pole-pair is plotted on the circumference of the unit circle with the assistance of the computer graphics. The parameter input from the keyboard controls the frequency of the digital oscillator, the output of which is made available via the Blue Chip data acquisition card ACM-44, using an 8-bit d-to-a (*AD7226*).

Base address is identified in the program.

Listing 2: provides a graphical representation of the amplitude and phase response of the rational function $H(z)$. To avoid exceeding the dynamic range of the monitor graphics the maximum value is normalised to unity. For best visual results, it is recommended that a sampling frequency of 640 be selected. This maximises the horizontal, relative frequency axis.

Listing 3: requests the user to specify a sampling frequency F_s, and an increment Δf in Hz. The program then prints the amplitude ratio and phase angle at frequency intervals of Δf over the range 0 to $F_s/2$.

All the programs are written in Microsoft C, version 5.1.

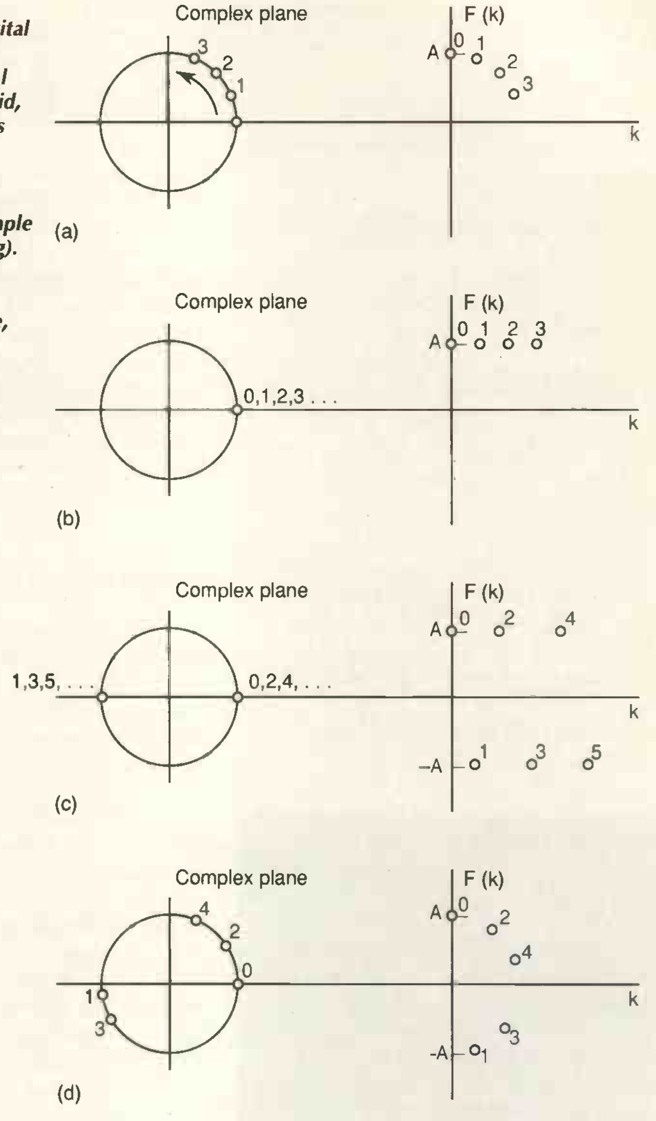

Fig. 2a. Complex digital phasor, and the projection of the real part, a digital sinusoid, onto the vertical axis labelled F(k)
2b. Complex digital phasor of frequency zero radians per sample interval (dc sampling).
2c. Complex digital phasor of frequency π radians per sample, sampling at Nyquist frequency.
2d. Complex digital phasor of frequency (π+δ) radians persample interval; aliasing due to undersampling.

Sampling is analogous to the stroboscopic effect noticeable in motion picture films of rotating objects, such as stage-coach wheels. Cine film consists of a series of still pictures captured at 24 frames a second. When the film is projected we interpret any displacement from one frame to the next as motion.

If you take a picture of a wheel with a spoke, in the 12 o'clock position and rotating clockwise, and then another a 24th of a second later, the wheel will have rotated and the spoke may well be in the 11 o'clock position. Another 24th of a second the spoke will point to 10 o'clock and so on. Our brain wrongly concludes that the wheel is travelling anti-clockwise.

A negative frequency means that the phasor is rotating in the clockwise rather than the counter-clockwise direction. Beware, by interpreting the sine wave as a projection on the Y-axis, the image will rise and fall sinusoidally, exactly the same whether the wheel is turning clockwise or anti-clockwise.

To visualise the performance, and limitations, of the sampling process, imagine a white rotating disc with a black dot painted near the rim, illuminated by a stroboscope.

The revolving dot represents the fixed frequency phasor, while the sequence of illuminated dot positions represents the discrete-time sampled data. If the dot moves slightly more than 180° between flashes, it will appear to be moving backwards at a rate slightly less than one-half a revolution per flash; what is more, the dot will never appear to be rotating more than π radians per flash.

Our conclusion must be that to represent the speed of revolution uniquely, we must sample at least twice per period of rotation.

Sampled sinusoid

In the z-plane, the frequency-selective performance is characterised by the complex conjugate pole-pair, on the circumference of the unit circle, together with a single zero at the origin.

Location of the poles of this digital resonator, given by the characteristic equation $z^2 - 2z\cos\Omega_0 + 1 = 0$, will determine the frequency of oscillation.

Clearly, by varying Ω_0, the linear system can be made to resonate in the range dc to half the sampling frequency.

Frequency of oscillation is related to the sampling frequency by,

$$\frac{\Omega_0}{2\pi}=\frac{f_{cyclic}}{f_{sample}}$$

The consequence of varying this parameter, and the effect on spectral and time-domain performance is detailed in **Fig. 3**, which shows time and spectral response at f_0=0.125f_s (45°).

First, notice how in the time-domain, Ω_0 corresponds to a sinusoid with eight samples per period. It is interpreted as the required frequency in fractions of the sampling frequency, so that the angular displacement Ω_0 is f_{cyclic}/f_{sample} multiplied by 2π.

Secondly, the frequency-domain model would ideally consist of a single ordinate, the spectral response of a sinusoid.

Computer-based software written in *C* provides a useful opportunity to investigate the performance of the variable frequency, discrete-time, linear system. On converting from transforms to sequences, the relationship between the input and output samples, **(Fig. 4)** may be expressed by the recurrence relationship

$$y(n) = 2\cos\Omega_0 y(n-1) - y(n-2) + \sin\Omega_0 x(n-1)$$

The unit-pulse response of this resonator is eternal, and once stimulated, the processed output will continue indefinitely with no further inputs. Of course, the position of the oscillator's complex poles, precisely located on the circumference of the unit circle will characterise a conditionally-stable system. For this reason, the multiplier coefficients are carefully declared in floating point format, with six digits of precision.

Practical implications

Practical implications of time-domain sampling and the effect on frequency-domain performance are important considerations in real-time numerical signal processing. Mathematical details are well-understood and have been extensively documented by many authors. But a more contemporary approach is to employ a practical pc-based investigation of the sampling theorem and to explore the relationship with sampled-data models or *z*-transforms.

To demonstrate the stroboscopic effects of sampling, an interesting computer-based study is to vary the location of the system poles, and to monitor the sampled-data output of the d-to-a, using an oscilloscope. An audio power amplifier and loudspeaker connected following d-to-a conversion, produces an audible output. With the amplifier connected to the output, List 1 (see note at end of article) is run.

The designed output frequency is varied by changing the location of the system poles. On passing through the Nyquist frequency two tones, sounding the same, are obtained. Clearly, the Nyquist frequency "mirrors" the unique wanted signal and unwanted alias.

Performance of a sampled-data system, which visualises the dynamic position of the system poles plotted on the *z*-plane, together with the audible, variable frequency output, detailed in the panel entitled 'Poles and zeros in action'.

When the software is run the computer-based text will initially outline a statement of the problem, before interactively encouraging the user to enter parameter Ω_0 to control the frequency of oscillation.

Output from the digital oscillator is available through an 8-bit d-to-a.

Computer graphics display the unit circle drawn in the *z*-plane and indicate the position of the selected poles.

Principles into applications

To develop a more realistic understanding of the properties of sampled data, we should formulate the *z*-transform more rigorously.

Expressed simply, the *z*-transform formulates a set of sequences or a difference equation into an algebraic structure, allowing much of the manipulation to be carried out according to a few very simple rules. To design and analyse numerical signal processing systems successfully we must understand the rules.

Stated formally the *z*-transform of a sequence $x(n)$ is given by:

$$X(z)=\sum_{n=0}^{\infty}x(n)z^{-n}$$

Conceptually, the transform may be visualised as a power series in z^{-1}, with coefficients equal to successive samples of the signal $x(n)$:

$$X(z)=\frac{X(0)}{z^0}+\frac{X(1)}{z^1}+\frac{X(2)}{z^2}+\ldots$$

This expansion is a refined perspective of the more primitive embodiment of z^{-1} as the unit-delay operator. Interestingly, any digital transfer function can be expressed as a rational function (a ratio of polynomials) in z^{-1}:

$$H(z)=\frac{Y(z)}{X(z)}=\frac{a_0+a_1z^{-1}+a_2z^{-2}+\ldots+a_Mz^{-M}}{b_0+b_1z^{-1}+b_2z^{-2}+\ldots+b_Nz^{-N}}$$

Rational *z*-functions of this form figure prominently in the analytical description of sampled-data systems – consult any table of *z*-transforms for details.

Previous use of the operator z^{-1} to model time delays placed no restriction on the shape of the time shifted waveform. Evidently the complex variable $s = a + j\omega$ in the expression $\exp(sT)$ took care of this.

We will concentrate on how a digital filter will process sine waves. Relaxing the original

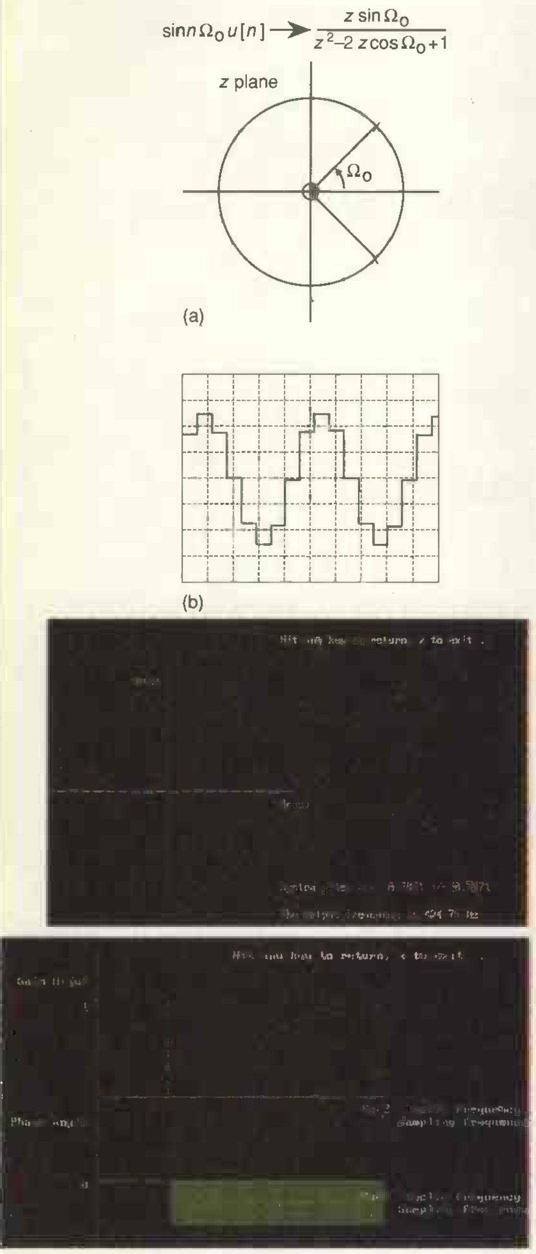

Fig. 3a. The pole-zero diagram of the sinusoid, visualised with the assistance of Listing 1. The frequency of the processed output is 434.78Hz using a pc clocked at 25MHz.
3b. Oscillogram of the d-to-a output. Ω_0=π/4 radians. Sensitivity 1V per division, time-base 0.5ms/division.
3c. The pole-zero diagram of the sampled sinusoid, visualised with Listing 1. The frequency of the processed output is 434.78Hz.
3d. Spectral response of the sinusoidal output, obtained with the assistance of Listing 2, demonstrates the unique, single-frequency performance of a second-order system characterised by an infinite Q-factor.

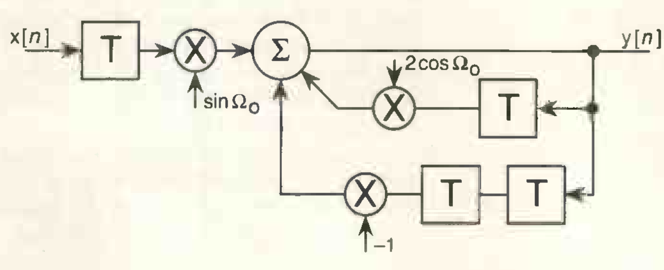

Fig. 4. System diagram of a digital resonator. The frequency of oscillation is controlled by the parameter Ω_0.

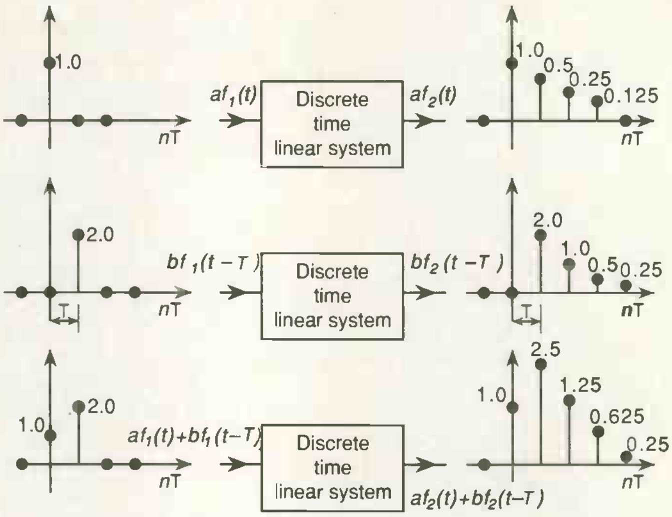

Fig. 5. Testing for linearity. An input signal formed by the weighted components $af_1(t)+bf_1(t-T)$ will generate the proportional output $af_2(t)+bf_2(t-T)$, without need to reanalyse the response.

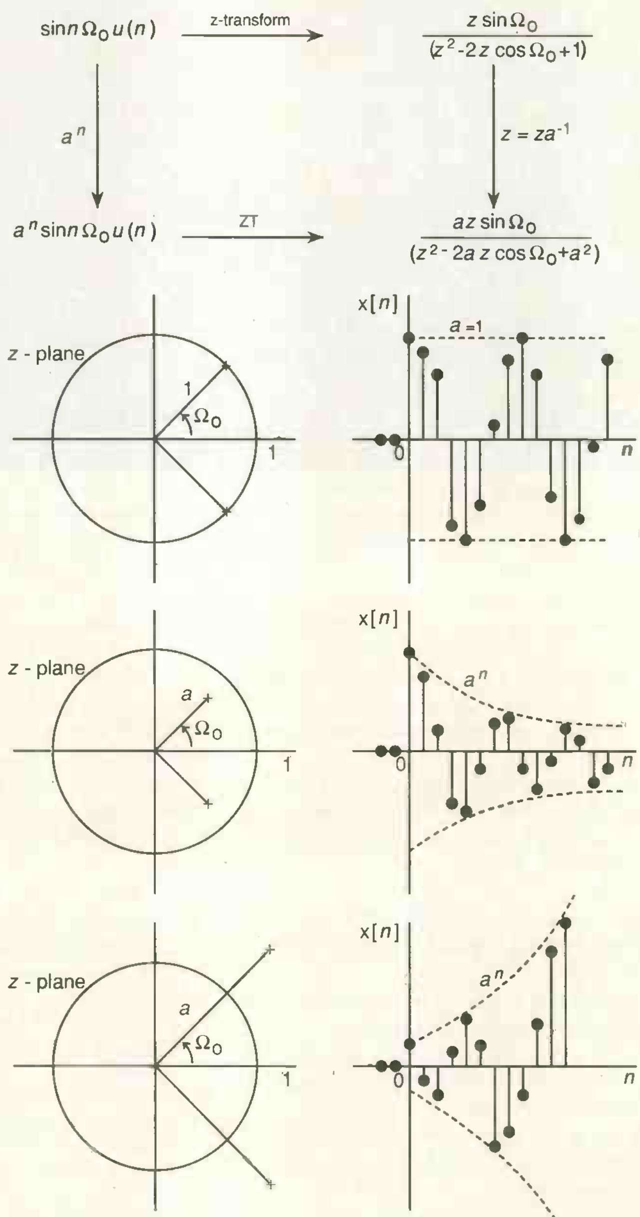

Fig. 6. Scaling in the x-domain. To evaluate the z-transform of $a^n f(n)$, trace the top right-hand path giving F(z) followed by substitution $z=za^{-1}$. Or follow the left-had path to obtain $a^n f(n)$ and then perform the transformation. Both routes are equivalent.

Poles and zeros in action

To generate a variable-frequency sampled-output, the digital oscillator is designed using the z-transform of a sampled sinewave

$$H(z) = \frac{\sin\Omega_0 z^{-1}}{1-2\cos\Omega_0 z^{-1} + z^{-2}}$$

Behaviour of this transform detailed as a linear system is easier to understand if we consider how the general digital function $H(z)$ can be expressed as the ratio of two polynomials, of the general form

$$H(z) = \frac{Y(z)}{X(z)} = \frac{a_0 + a_1 z^{-1} + a_2 z^{-2} + \ldots + a_M z^{-M}}{b_0 + b_1 z^{-1} + b_2 z^{-2} + \ldots + b_N z^{-N}}$$

By inspection, the coefficients of the sinusoidal generator are given by $a_1 = \sin\Omega_0$, $b_0 = 1$, $b_1 = -2\cos\Omega_0$ and $b_2 = 1$.

The numerator and denominator polynomials can always be factorised to give the pole-zero plot. Characteristics of a particular data sequence are specified by the roots of the two polynomials – numerator $N(z)$ and denominator $D(z)$ typically plotted on an Argand diagram. A unit circle is generally drawn to indicate all values of z for which $|z| = 1$.

Poles and zeros can be plotted graphically, and for a sampled-data system the unit circle is a reference from which the frequency response of the filter can be measured geometrically.

Coordinates of the upper, unit semi-circle, traced out in an anti-clockwise fashion represent frequencies from dc to half the sampling frequency; the operating range. To represent frequencies from half the sampling frequency to the sampling frequency, the 'walk' around the circumference of the unit circle is continued. The amplitude ratio is determined by the position of the poles and zeros, relative to the circumference of the unit circle.

A pole close to the circle will cause a peak, a zero a dip in the spectral response; the closer they are to the unit circle the more extreme is their effect. Thus, the complex poles of the sinusoidal generator detailed previously will model a second-order system characterised by an infinite Q-factor.

The term zero can be misleading. Located at the origin of the z-plane (0,0) a zero represents a time advance of one sampling interval, this will modify the phase response only. Any radial displacement towards the unit circle progressively increases the attenuation, until, upon reaching the circumference the attenuation is a maximum and the amplitude ratio is zero.

The frequency of oscillation is controlled by the parameter Ω_0.

definition of s and discarding the exponent a, (by making $a = 0$) means that z may be viewed as a phasor defined by $z = \exp(j\omega T)$. Of course a complex signal or system will be composed of several frequencies and may be represented as the sum, or superposition, of many such phasors. This model fits naturally with the signal processing characteristics of linear systems.

Any signal represented in terms of its component frequencies will be processed by the linear system in a very simple way. Only the amplitude and phase will be modified.

But how can we be certain a system is linear? To test for linearity we must apply the principle of superposition (**Fig. 5**). An input signal formed by the weighted components $af_1(t) + bf_1(t - T)$ will generate the proportional output $af_2(t) + bf_2(t - T)$ without need to re-analyse the response.

Z-transforms in time and frequency

Magnitude and phase of the frequency response of the general sampled-data system $H(z)$ can be determined by replacing z by phasor $\exp(j\omega T)$. Employing this substitution, $H(\exp(j\omega T))$ may be regarded as a complex number, which characterises the frequency-selective properties of the rational transform in terms of the amplitude ratio and phase angle over the range of interest.

Viewed as a sinusoidal signal processor the transfer function of the transform may be expressed as:

$$H\left(e^{j\omega T}\right) = \frac{a_0 + a_1 e^{-j\omega T} + a_2 e^{-j2\omega T} + \ldots + a_M e^{-jM\omega T}}{b_0 + b_1 e^{-j\omega T} + b_2 e^{-j2\omega T} + \ldots + b_N e^{-jN\omega T}}$$

Rational z-transforms play a central role in the development of sampled-data systems. A relatively modest table of transforms will now be selectively explored using pencil-and-paper analysis assisted by computer managed instruction. Program Lists 2 and 3 are available to demonstrate this.

There is considerable merit in developing the z-transform of the decaying exponential signal from first principles, as a foundation from which to obtain other commonly used transforms and to develop a real comprehension of pole-zero models – a delicate topic that must be thoroughly mastered and understood.

Example 1. $x(n) = \cos(n\Omega_o)$ *for* $n \geq 0$: Consider how the sinusoidal response normally associated with complex poles can be usefully developed from the decaying exponential $x(t) = e^{-at}$.

Using Euler's formula, it follows that the periodic function $x(t) = A\cos(\omega t)$ can be expressed in complex exponential form as:

$$A\cos n\Omega_o = A\frac{\left(e^{jn\Omega_o} + e^{-jn\Omega_o}\right)}{2}$$

Converting to z - transforms:

$$X(z) = \frac{A}{2}\left(\frac{z}{z - e^{j\Omega_o}} + \frac{z}{z - e^{-j\Omega_o}}\right)$$

$$= A\frac{1 - \cos\Omega_o z^{-1}}{1 - 2\cos\Omega_o z^{-1} + z^{-2}}$$

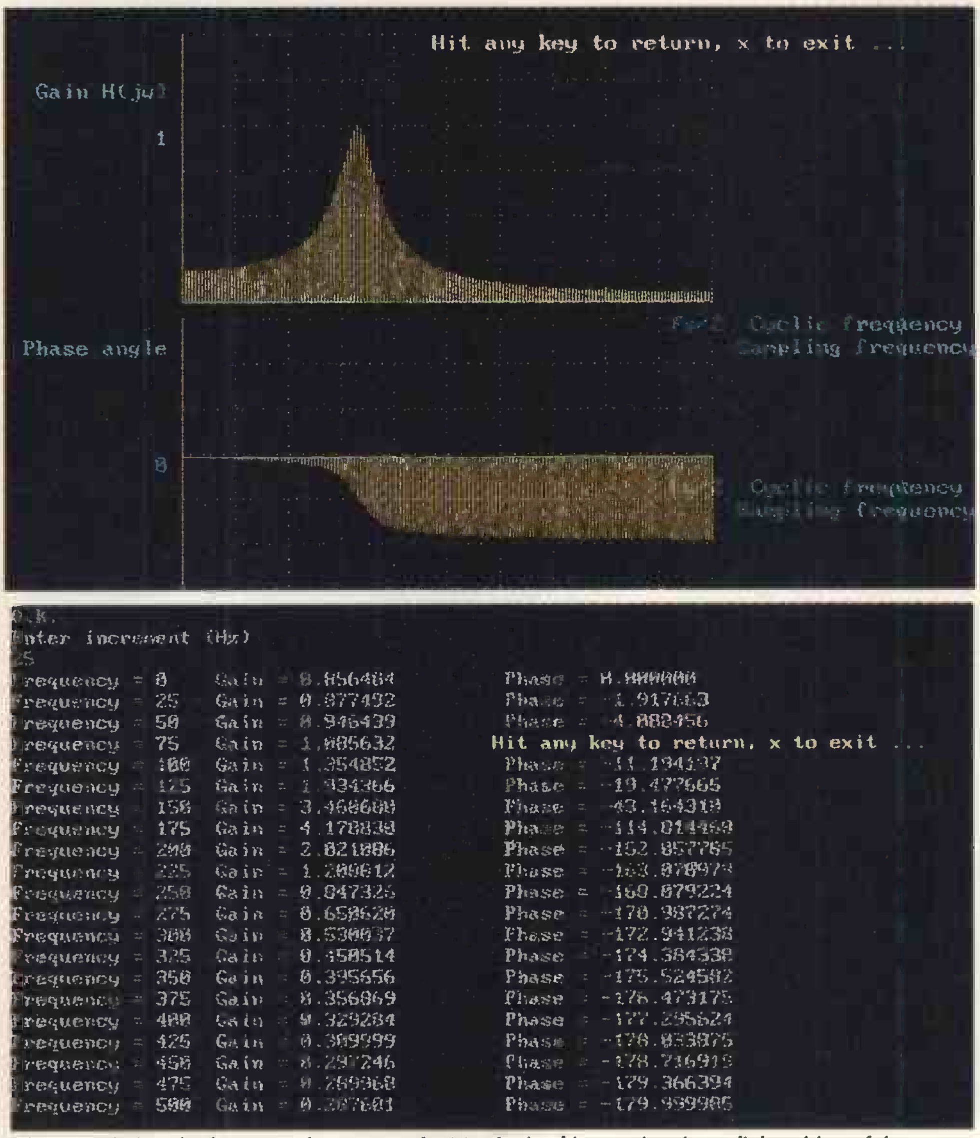

Fig. 7. Variations in the spectral response of H(z), obtained by varying the radial position of the poles, graphically demonstrate how the Q-factor or sharpness of the peak in the amplitude ratio is related to the positions of poles relative to the unit circle. Values are $\Omega_0 = \pi/3$, a = 0.9. The numerical response demonstrates performance for F_s=1000Hz, δf=25Hz.

Evidently, the concept of complex poles has arisen as a natural consequence of the z-transform of the cosine function. Not surprisingly the characteristics of under-damped second-order systems and frequency selective filters can be effectively developed from complex pole-zero models.

To investigate the frequency domain properties of this transform, the software can be used (Listings 2 and 3) to vary parameter Ω_0 over the range $-\pi \leq \Omega_o \leq \pi$.

Example 2. $h(n) = a^n\sin(n\Omega_0)$: Let us investigate the damped sinusoidal sequence $h(n) = a^n\sin(n\Omega_o)$.

Clearly the z-transform has been developed from the time-domain multiplication of the sequences $f(n) = a^n$ and $g(n) = \sin(n\Omega_o)$, or equivalently the frequency-domain convolution of $F(z)$ and $G(z)$.

The tedious algebra associated with this operation is avoided by taking advantage of the multiplication by a^n property of z-transforms, to demonstrate scaling in the z-domain (**Fig. 6**). The effect on the poles will be a radial displacement along the same frequency vector (Fig. 6b).

(To examine the effects of varying the parameter a, for a particular value of Ω_0, (say $\Omega_0 = \pi/3$), Listing 2 could be be run, using for example, 0.6, 0.9 as values of a. The spectral response is shown in **Fig. 7**.

For a numerical response of the amplitude ratio and phase angle then Listing 3 can be run. ■

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Best rf article '95

Following the success of 1994's Writers Award, ***Electronics World*** and **Hewlett-Packard** are launching a new scheme to run from January to December 1995.

Only articles which have an element of rf design will be eligible for consideration by the judging panel. It is hoped that this year's award will focus writer interest on rf engineering in line with the growing importance of radio frequency systems to an increasingly cordless world.

The aim of the award scheme is to locate freelance authors who can bring applied electronics design alive for other people.

Qualifying topics might include direct digital synthesis, microstrip design, application engineering for commercially available rf ICs and modules, receiver design, PLL, frequency generation and rf measurement, wideband circuit design, spread spectrum systems, microstrip and planer aerials... The list will hopefully be endless.

All articles accepted for publication will be paid for – in the region of several hundred pounds for a typical design feature.

The prize for the coming year's award is a £4000 Hewlett-Packard HP8647A 1GHz programmable signal generator. It features HPIB interface, solid state programmable attenuator and built in AM-FM modulation capability.

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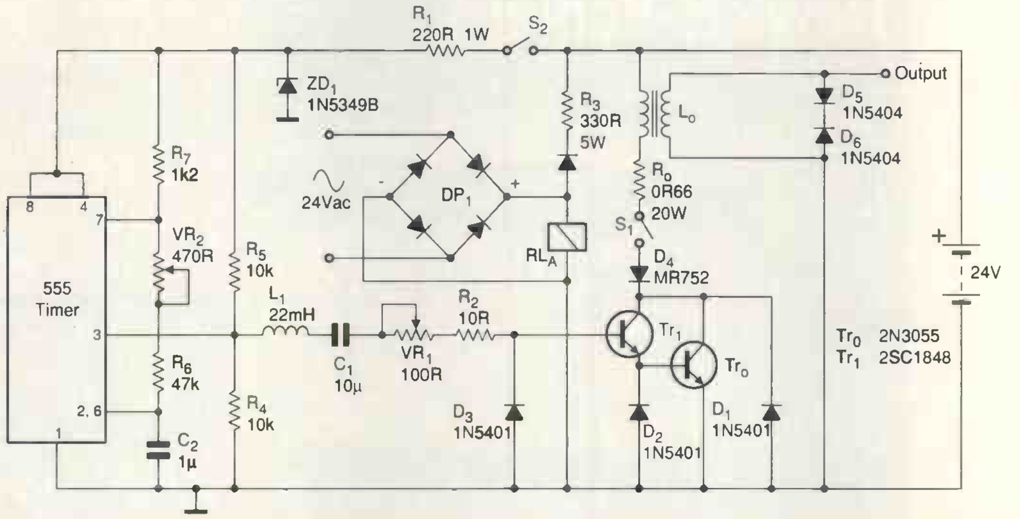
Backup source using resonant inverter

For equipment that can tolerate non-sinusoidal ac supplies, this source of backup power is simple and efficient. Its efficiency arises from its use of an *RLC* resonator driving output transistors from saturation to cut-off, a further benefit of this arrangement being that the circuit is less sensitive to transformer performance than the more common Class-A push-pull inverter. Capacitive loads do introduce losses.

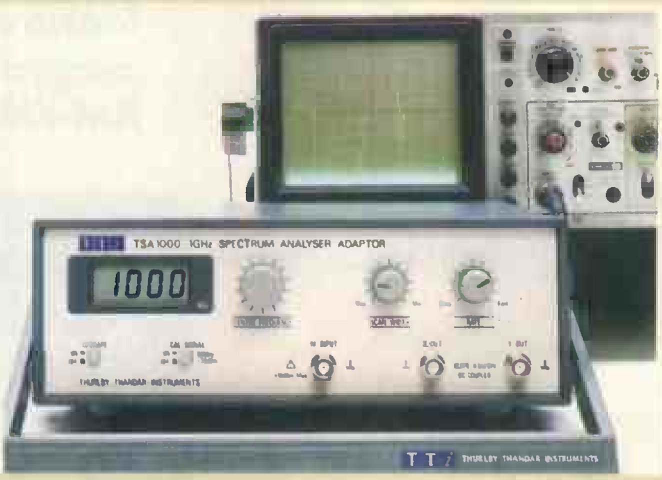
A *555* oscillates at 1kHz and feeds the resonator circuit $R_{4,5}$, L_1, C_1, VR_1, R_2, in combination with the input loading of the transistors, which are protected against reverse breakdown by diodes. If mains power fails, the normally open contacts of RL_A, supplied by a step-down transformer and bridge rectifier diode pack, close and enable the inverter. Voltage from the bridge rectifier also maintains a charge on the battery pack through the diode and current limiter R_3.

Diodes $D_{5,6}$ remove spikes at the output; VR_1 varies resonator frequency to match the output load; and VR_2 determines maximum power for a given load.

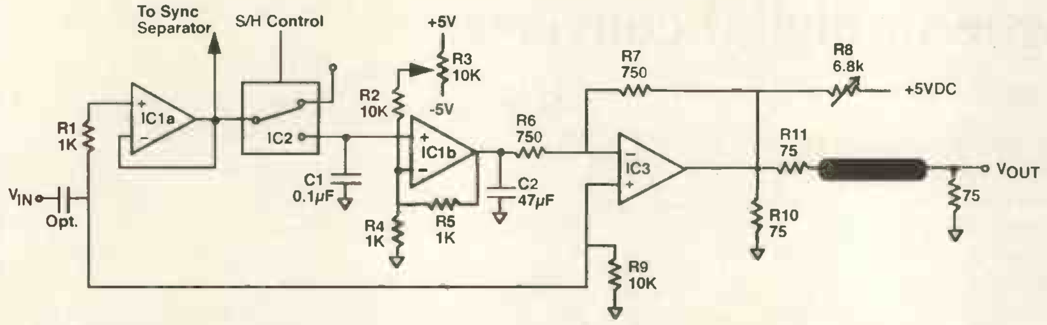
Sode-shinni N Rumala
Federal University of Technology
Minna
Niger State
Nigeria



Simple and efficient inverter provides a backup source for waveform-tolerant equipment and maintains a charge on the battery pack.



Video amplifier with sync stripper and dc restore



One of the benfits of this circuit is that by stripping off the sync from a video signal and digitising only the active information, better use is made of the a-to-d converter's resolution.

This circuit transmits 220MHz, –3dB bandwidth, video signals while stripping off the sync pulse and performing dc restoration. It is configured for a typical video cable driver application feeding a double-terminated 75Ω load. The *HFA1103*, IC_3, is configured for a gain of +2 to ensure unity gain overall.

Stripping sync

Sync is often combined with one or more of the red, green, and blue video signals in component video distribution amplifiers. This is done to decrease the number of input and output channels required. Also, as the video signals exit the switching network, the sync pulse must often be removed.

Other applications benefiting from sync removal are video digitising circuits and hdtv systems. Consider a typical 1V peak-to-peak rgb video signal with a –300mV sync pulse and +700mV video data. By stripping off the unwanted sync pulse and digitising only the active video, designers can use the full dynamic range of the a-to-d converter for the +700mV video data. This results in a 30% increase in resolution using the same a-to-d converter.

The *HFA1103* video op amp is specially designed to perform sync stripping. Its open emitter n-p-n output forms an emitter-follower with the load resistor, and passes the active video signal while virtually eliminating the negative sync pulse.

Residual sync of the *HFA1103*, defined as the remainder of the original –300mV sync pulse, referenced to ground, is only 8mV at the cable output.

A particular advantage of sync stripping with the *HFA1103* is the resultant larger – by 0.7V – output voltage swing, compared to simply using a wideband video op-amp with an external emitter follower.

Because the *HFA1103* contains no active pull-down, output linearity degrades as the signal approaches zero volts. To deal with this a 6.8kΩ pull up resistor, R_8, and a 75Ω pull-down resistor, R_{10}, on the output ensure a fixed positive voltage offset, in this case +50mV. This offset was arbitrarily chosen as a good compromise between linearity near the dc level and minimum residual sync. Increasing R_8 decreases residual sync, at the expense of linearity. Conversely, decreasing R_8 decreases linearity error, but increases residual sync.

Dc restoration

Another common video function is dc restoration, used when ac coupled signals have lost their dc reference and must have it reset line-by-line in order to retain brightness information.

The circuit accomplishes dc restoration using a *CA5260* dual op amp ($IC_{1a,b}$) coupled with a sample-and-hold circuit based on the *74HC4053* switch IC_2). Vin, consisting of the input video signal and a dc offset (V_{dc}), is connected to the non-inverting input of the *HFA1103* (IC_3). The *HFA1103* is configured in a gain of +2, which would result in an output of $2\times V_{in}=(2\times video+2\times V_{dc})$, were it not for the dc-restore circuit.

V_{in} also travels through half of the dual *CA5260* amplifier to the sample-and-hold circuit, where the 0.1µF capacitor (C_1) is the hold capacitor. (The sample-and-hold control is triggered by a back-porch pulse from a sync separator or by horizontal video blanking) This dc signal is then amplified by a gain of +2 by the second op amp (IC_{1b}). The gain of +2 is required because the dc offset is input to the *HFA1103*'s inverting input, which provides only a gain of –1. Thus 2×Vdc is summed into the inverting input of the *HFA1103* and is subtracted from the output signal.

Because the output impedance of IC_{1b} is high, and would affect the gain at the non-inverting input of the *HFA1103*, a 47µF capacitor (C_2) is used to provide an ac ground and to maintain good high-frequency gain accuracy.

A potentiometer (R_3) is used prior to IC_{1b} to null out any offset voltage contributed by the dc-restore circuitry.

The resulting output is a 220MHz, dc restored video signal in which the sync pulse has been stripped to a residual level of no more than 8mV.

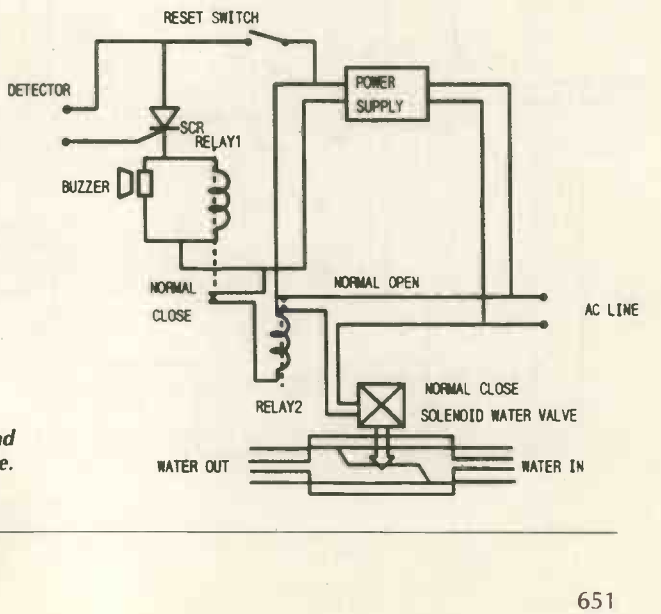
Jeff Lies, Chris Henningsen and Mike Press
Harris Semiconductor
Melbourne, USA

Water leak detector/stop valve

As a means of avoiding total disaster in the presence of unlooked-for quantities of flood water, this very simple circuit gives early warning and is extendable to turn the water off, given that an electrically operated valve is installed.

If water bridges the two SCR electrodes the device conducts and continues to do so until reset, sounding the buzzer. The SCR drives relay 1, which in turn drives another relay capable of operating the solenoid water valve.

Shin'ichiro Asai
Research Centre of Denkikagaku Kogyo KK
Tokyo



If the washing machine leaks, this circuit will buzz and operate an electrical stop valve.

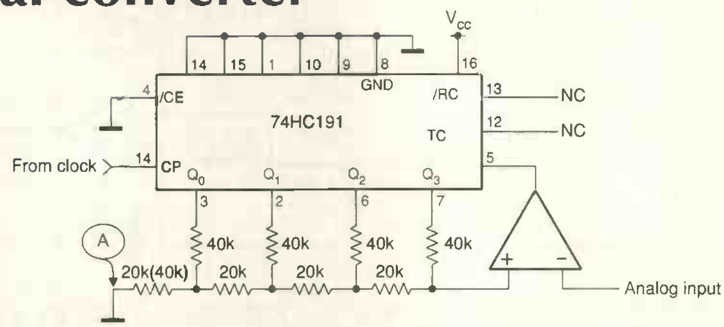
R/2*R* analogue-to-digital converter

This 4-bit a-to-d converter tracks the input and was originally meant to drive the mouse input of an Amiga. It is easily extended to eight bits or more.

Under the control of the op-amp, which drives the 'direction' pin, the *191* up/down counter counts until the *R*/2*R* ladder output on the non-inverting op-amp input equals the analogue input, at which point the counter state represents the digital equivalent of the input.

If another *191* is to be added, connect pin 13 of the new one to pin 4 of the existing counter, point A on the ladder going to Q_3 of the new one.

John Henningsen
Copenhagen
Denmark



Up/down counter tracks analogue input to give a 4-bit a-to-d converter, convertible to give more bits.

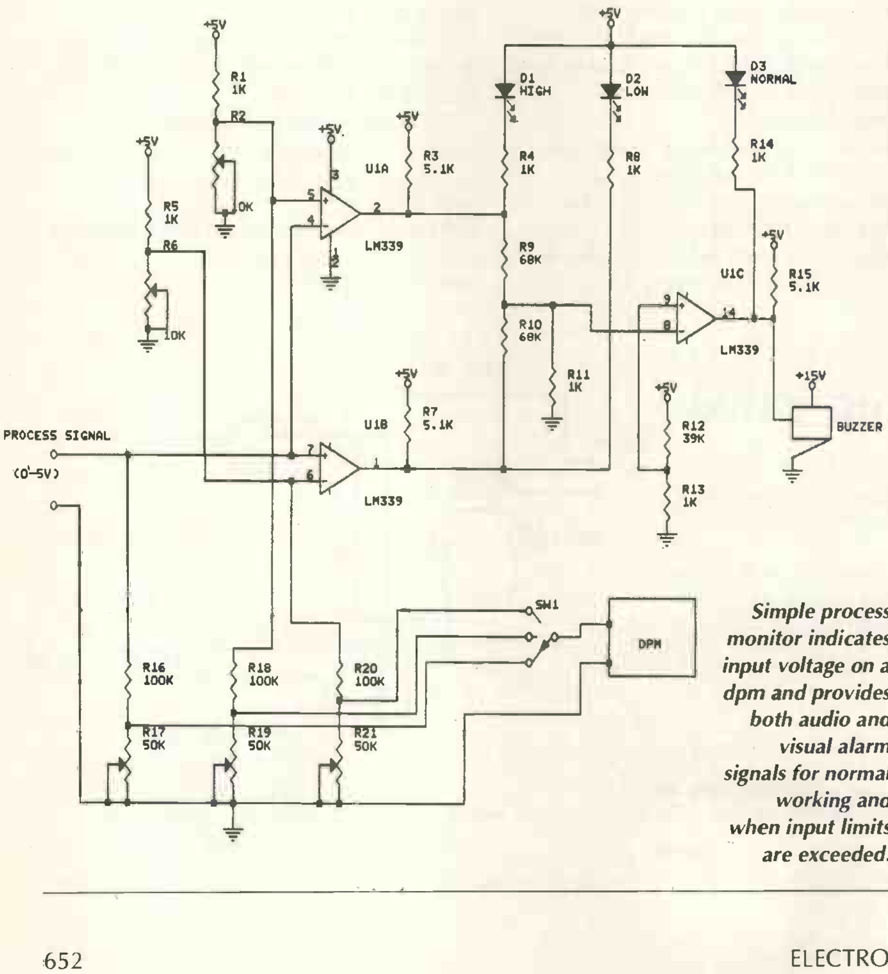
Process monitor

Accepting an input of 0-5Vdc, this circuit monitors an industrial process and provides out-of-limits alarms.

Input goes to the two comparators, which have their references set high and low respectively and adjusted by R_2 and R_6. If the input reaches either of the two reference voltages, one of the comparator outputs goes low and a led is illuminated, the low at either comparator output causing the buzzer driver to trigger the audio alarm. In normal conditions, while the input stays between the two input reference voltages, the buzzer driver output is low and the "normal" led illuminated.

Input voltage is displayed by the digital panel meter, which is switchable to show the set limits, the calibration method being to set the inputs at half-scale and to adjust $R_{17,19,21}$ to make the dpm indicate half-scale.

P Bhanu Prasad and R S Mahajan
Central Electronics Engineering Research Institute
Pilani



Simple process monitor indicates input voltage on a dpm and provides both audio and visual alarm signals for normal working and when input limits are exceeded.

Hf buffer with zero-offset

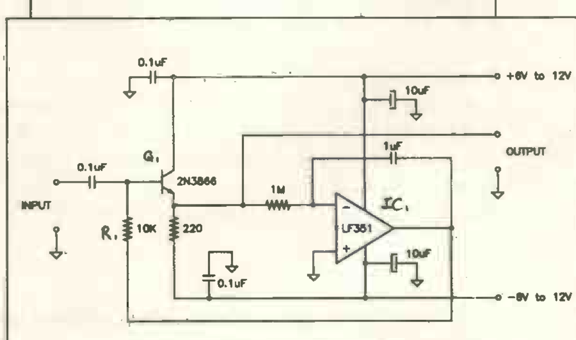
A high-frequency, low-impedance drive to a fet switch uses an emitter follower, the dc offset being removed by feedback.

At switch-on, the transistor is off and the integrator output ramps in a positive direction, but slowly, given the 1μF and 1MΩ timing components. Feedback to the transistor base turns it on, its emitter rising towards the 0V on pin 3 of the integrator. When settled, the integrator output is enough to maintain conduction in the transistor, its emitter being at 0V, except for a possible slight offset caused by the input offset voltage of the op-amp. The *LF351* has an adjustment to remove this offset.

With the components shown, the circuit buffers frequencies up to 40MHz.

G W Väth
University of Natal
Durban

With no offset to cause a pedestal in a following fet switch and no unity-gain op-amp buffer, this clamped emitter follower buffers 40MHz signals.



Cross-point switch controller stores input

Cross-point switch *AD75019* from Analog Devices keeps switching data in a shift register formed from dynamic memory and requires input to be at a minimum 20kb/s. This circuit idea allows the device to be used in the presence of interrupts.

A main processor uses a *PIC 16C84* as a local controller, providing 36 bytes for 256 bits to control the 256 switches in the array and 4byte extra for other purposes. Either a serial or parallel interface to the main processor is possible; in this case, an 8-bit parallel port was used for speed. An on-board 64-byte eeprom will hold two sets of switch data for frequently used configurations.

Data to the *AD75019* is clocked serially into a 256-bit register, as S_{in} goes high, and control starts with the Y15/X15 cross point, going down to Y15/X0, proceeding to Y14 and finishing with Y0/X0. Data is latched into the device as PCK is pulsed low, where it remains while power is applied.

From the main processor, data is four bits for the X address, four for Y, one to indicate required switch status and four for the command.

We implemented six commands:

RESET to clear the ram image and update the cross-point switch to all open;

SETSW to set the ram image for switch X,Y open or closed and then to update the cross-point;

SAVE-A saves the current ram image to eeprom area A;

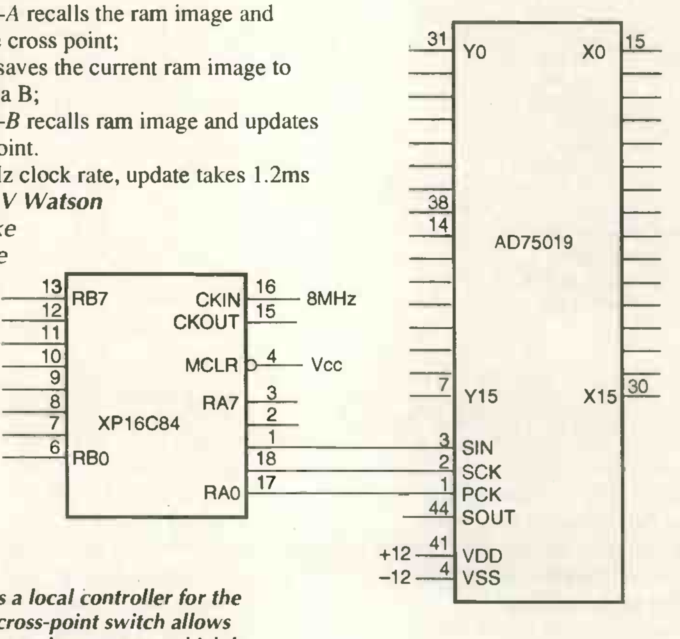
RECALL-A recalls the ram image and updates the cross point;

SAVE-B saves the current ram image to eeprom area B;

RECALL-B recalls ram image and updates the cross point.

At an 8MHz clock rate, update takes 1.2ms

Michael J V Watson
Basingstoke
Hampshire



PIC used as a local controller for the AD75019 cross-point switch allows control by a main processor which is unable to comply with the minimum 20kb/s transfer rate.

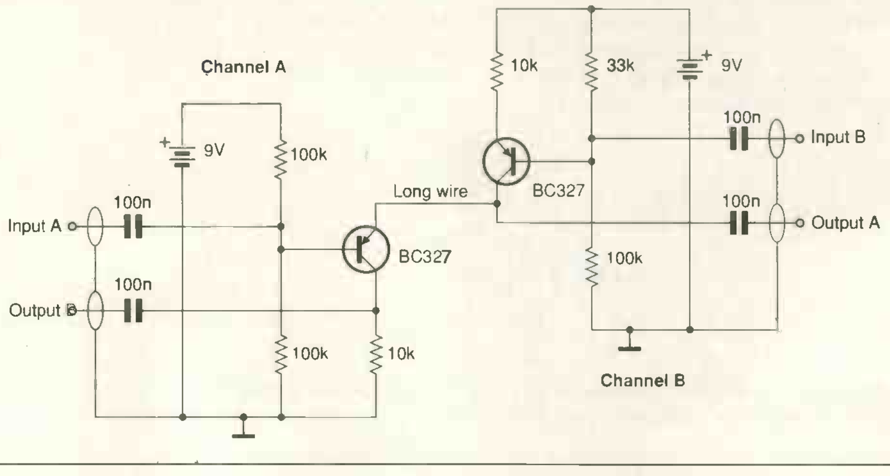
Single-wire, duplex communications link

This two-way link could hardly be simpler. It carries bi-directional signal simultaneously, as in a telephone link, the household earth being used for the return path.

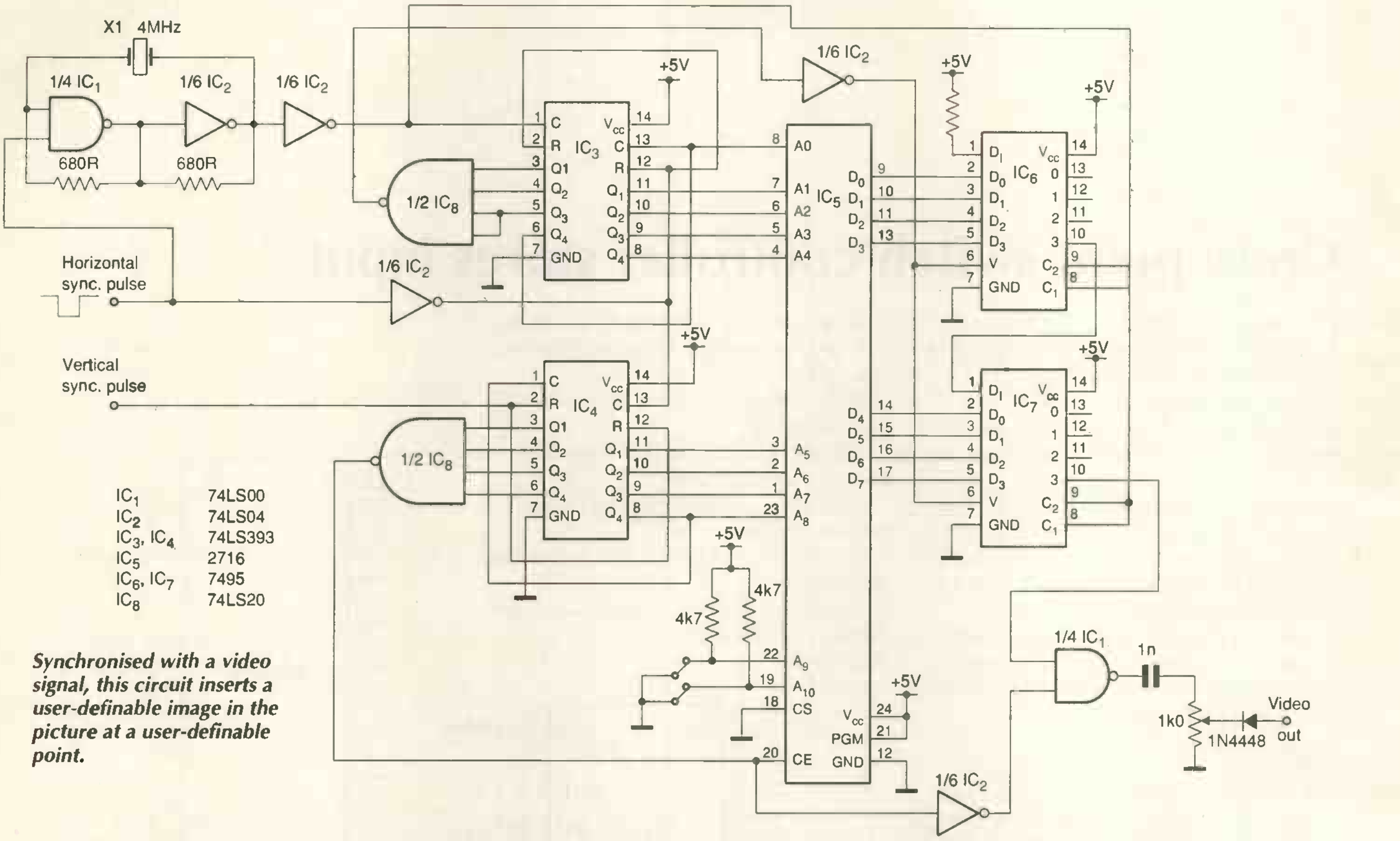
Conflict is avoided by making channel A drive the voltage source and channel B the current source.

Andrew M Wilkes
Wokingham
Berkshire

Very simple circuit provides simultaneous two-way communication, using one wire. If noise is a problem, coaxial cable will improve performance.



Video image inserter



Synchronised with a video signal, this circuit inserts a user-definable image in the picture at a user-definable point.

Using an eprom, this circuit inserts a message at a user-definable point on the screen. Oscillator $IC_{1,2}$ provides 4MHz pulses outside the horizontal sync pulse. These pulses clock counter IC_3 and registers $IC_{6,7}$. Write and shift signals for the registers are provided by parts of $IC_{2,3,8}$.

Adderss-selection for the eprom is carried out by $IC_{3,4}$. Via address lines A_{0-4}, the rom can store 32×8 horizontal points and 16 lines vertically via A_{5-8}. Address bit $A_{9,10}$ can be used to select one of four pictures.

Pins 3-6 of counter IC_4, driving part of IC_8, are used to determine the position of the picture on the screen. Inserting inverters between these ICs exchanges the position of the picture.

Output of the circuit connects to the video input. The horizontal and vertical sync pulses must be at ttl levels.

Vasily Fyodorov
Leo Tolstoy
Russia

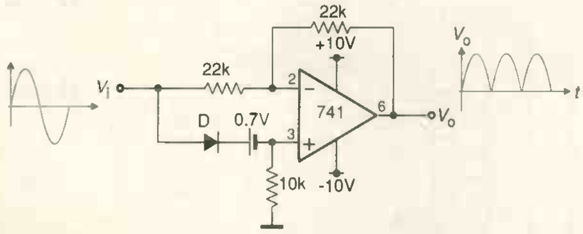
Single op-amp full-wave rectifier

Most full-wave rectifiers using op-amps need two amplifiers; this one uses only one.

On positive half-cycles, the diode conducts and the circuit becomes a voltage follower, and on negative half-cycles a unity-gain inverting amplifier, so that the output is about the same in both halves.

Since, however, the diode is not in the feedback loop, its forward voltage drop is not reduced and must be cancelled by the 0.7V.

Uday S Tandi
Raichur
Karnataka
India



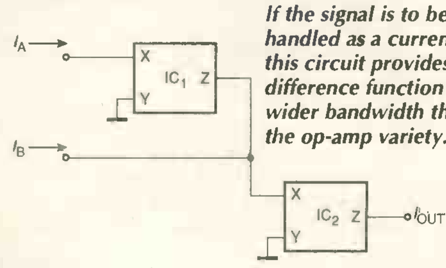
This full-wave rectifier is unusual in its use of a single op-amp, paying the price for simplicity by its need for a bias voltage.

Subtracting currents

When it is simpler to regard signals as currents, rather than voltages, in signal processing, two current conveyors can be used to produce the difference function.

Input I_A is inverted in IC_1 and added to I_B at the X input of IC_2 to give $I_B–I_A$, the output then being $–(I_B–I_A)$ or $I_A–I_B$. A bonus is the comparatively wide bandwidth of current conveyors.

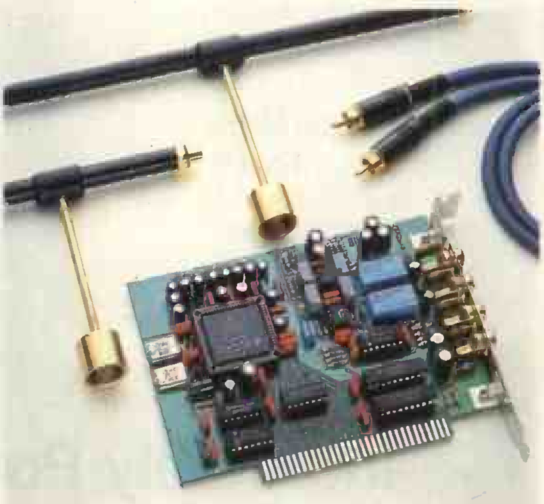
L Szymanski
Stamford Lincolnshire



If the signal is to be handled as a current, this circuit provides the difference function at a wider bandwidth than the op-amp variety.

AF analysis *via the pc*

A relatively low-cost virtual instrument originally intended for loudspeaker analysis has now expanded into a comprehensive lf measurement suite. Richard Lee has been looking at how this analyser – Clio – performs.



With virtual instruments – i.e. instrumentation simulated on a pc screen – signal generation and analysis are carried out by hardware, under the control of user-friendly software. So far, virtual instrumentation has been expensive, but low-cost systems are starting to appear.

Clio – from Italian recording studio equipment manufacturer Audiomatica – offers several virtual audio instruments in one, at a basic price of £799. It was originally designed for loudspeaker system engineers, but has expanded into a suite of instruments with a wide number of other uses involving audio frequencies.

Although Clio's software will not run under Windows, it does have a mouse controlled graphical user interface. Combined with a laptop, it forms a portable measuring system needing only a power amplifier for on site loudspeaker or room acoustics measurements. No additional amplifier is required for impedance measurements, or testing electrical or electronic circuits.

Virtual instruments included are a swept-sine frequency response analyser, 10Hz–20kHz storage oscilloscope with 20kHz bandwidth and a fast Fourier transform, FFT, analyser. Spectrum analysis to 20kHz is possible, and the system can measure total harmonic distortion, ac millivolts, Vrms, dBV, dBm and dB relative, fast and slow reading. Sensitivity is up to +30dBV and down to –40dBV. An *LC* meter is also built in.

The signal generator is programmable and comes with a library of signals including sine waves, with a claimed thd of 0.015%, pink noise, white noise, square waves and pulsed sine waves, **Fig. 1**.

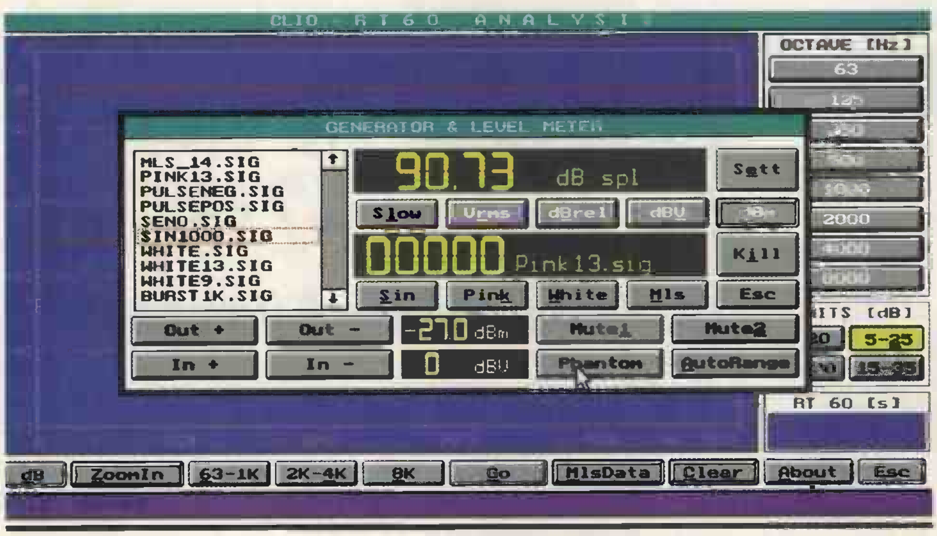


Fig. 1. The generator/voltmeter front panel window allows input/output conditions to be set and then dropped down over most instrument displays. The Clio board has two inputs and two outputs, but current software commons both outputs but uses only one input. Being able to use both sets independently would simplify stereo testing.

For acoustic measurements, the Clio board includes a preamp with switchable dc phantom powering and microphone sensitivity calibration. The optional microphone plugs straight into the board's input. Instruments dedicated to sound measurements include a third-octave real-time analyser, rta, with internal pink noise source and A-weighted sound level meter, **Fig. 2**. Also for acoustics engineers is the *RT 60* analyser, which carries out room reverberation time measurements.

Maximum length sequence, mls, time-domain measurement and FFT analysis allows loudspeaker frequency/phase responses to be measured in-situ. This avoids the need for a

Requirements
PC compatible with an 8/16bit ISA slot.
286 processor or higher
Ega display
640KB of ram or more.

two-channel FFT analyser because the mls is a recognised signal. Reflections from surfaces can be removed by post-measurement truncation of the time-domain signal, **Fig. 3**.

Additionally, mls allows 3D sound amplitude/frequency versus time 'waterfall' plots to be generated, and energy time curves which show how sound decays over time, **Fig. 4**.

Clio uses two direct memory access channels, dma, one for signal generation and the other for data acquisition. The board also makes use of three i/o lines. This might be restrictive if your pc already has a few occupied slots. Clio uses the same dma and i/o as *DRALABS* maximum length sequence signal analyser, so if your pc will run mlssa it should run Clio.

The mls measuring system samples at 51.2, 12.8 and 3.2kHz. Set to 51.2kHz for a 20kHz bandwidth, it is very fast and displays do not autoscale irritatingly. If set to 3.2kHz sampling frequency, it will measure down to 0.125Hz.

Filing data

Clio will export frequency responses to hard disk for use with cad loudspeaker design packages such as *Leap* and *Calsod*. Ascii time-domain files cannot be exported, but Audiomatica can supply a Pascal structure describing the file format needed to do this.

Swept sine-wave measurements comprise tone-bursts at discrete frequencies, with interpolation to create smooth curves, **Fig. 5**. Connecting input to output produced an overall response flatness within 0.5dB. Frequency response files may be loaded as a reference and new measurements displayed as a difference curve. This is ideal for comparing production items to standards. Files can also be imported from outside the sine wave environment, displayed and post processed. Curves for instance, can be averaged.

Another facility takes the measured in-box sound pressure level, spl, and simulates an outside-of-the box spl. Text files can be edited to splice two separate sets of readings, for example mls mid-to-high response and near-field low-frequency response, to give a complete anechoic frequency response.

Theile/Small loudspeaker characterisation is automatic, very quick and simple. The analyser hunts for the impedance curve resonance peak and half points to an accuracy of $^{1}/_{24}$ octave, a worst case frequency error of 2.9%. Accuracy of Theile/Small parameters was found to be about as good as you can get without measuring the *BL* factor directly.

Hardware in the Clio package includes measuring cables and a half-length pc card holding

***Fig. 2.** Clio rta plot. The internal pseudo-random pink noise is very accurately distributed in frequency. Because its energy content is known, the analyser doesn't have to wait as long to average at low frequencies – thus making it faster than a normal real time analyser using ordinary random pink noise*

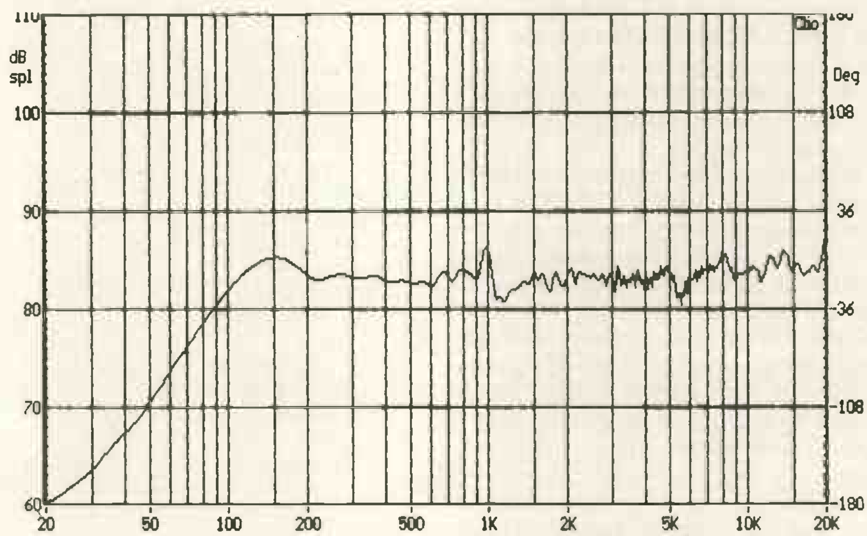


***Fig. 3.** Clio mls loudspeaker measurement taken in an anechoic room but truncated to 100ms. As with any truncated measurement, this results in losing low-frequency information, (the longer the sample time, Ts, the better the low-frequency performance), but it does allow the room to be taken out of the measurement frame completely. Indeed, a truncated response excluding reflections is more anechoic down to the 1/Ts frequency (100Hz in this case) than the usual swept measurement in even the best chambers. In room one could expect anechoic accuracy down to 250Hz or so. Setting the time 'window' is particularly easy and the display shows the 1/T_s frequency.*

***Fig. 4.** This waterfall plot was taken in room. These plots can help pinpoint sources of colourations in loudspeakers caused by energy storage and reflections.*

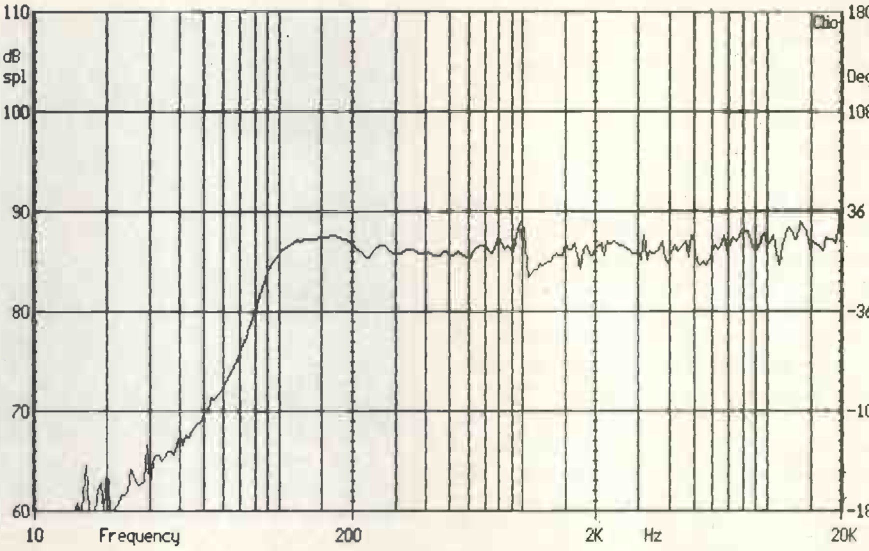


Fig. 5. Loudspeaker sine-wave response measured with Clio in an anechoic room (compare to Fig. 3 mls plot taken with same loudspeaker). The previous ten curves can be stored and shown as overlays, the oldest being erased for each new acquisition.

signal processing circuits and 16-bit sigma-delta a-to-d and d-to-a converters for performing analogue i/o.

Listening in

At an extra £150, the optional 8mm diameter microphone is an electret capsule. It comes with no curve but it does have a specification window. Tested in Celestion's anechoic room with a known loudspeaker against a B&K microphone, the frequency response was found to be flat except for a slight rise at high frequencies of about +1dB from 5kHz–15kHz. Sensitivity as measured via a B&K calibrator was probably within 0.5dB. This is typical of a good, electret insert.

The storage oscilloscope worked well, with excellent triggering. At some settings, where the screen is frequently updated, mouse operation became jerky due to the fact that the pointer cannot move while the screen is updating. This is easily overcome by using hot-keys to operate the buttons.

Testing the FFT spectrum analyser with the internal sine oscillator at 1kHz, Clio's thd meter measured down to a minimum reading of 0.01% – slightly better than the claimed 0.015% thd of the generator alone, **Fig. 6**. All one can be certain of from these tests is that the thd and FFT analyser easily meet their respective specifications. With intelligent use of input sensitivity and FFT scaling, Clio provides a handy check on distortion spectra and noise in electronic circuits.

Overview

Clio is an easy to operate, fast, user-friendly package, combining several instruments geared to acoustic and electronic testing at audio frequencies. From a speaker engineer's viewpoint, it is ideal for use as a professional design and development tool, and could even find its way from the laboratory to the factory floor as an inexpensive and easy to operate on-line tester.

Acoustics and speaker installation engineers will find the real-time spectrum and *RT60* analysers invaluable.

Electronic engineers should find the electronics measurements equally useful, especially considering that here is a complete work bench all in one package – and well suited to companies on a tight budget. ■

CLIO Specifications

Generator	
Type:	Two channels 16 bit sigma-delta d-to-a converter
Frequency range:	1Hz-22kHz (+0/–1dB)
Frequency resolution:	0.01Hz
Output Impedance:	100Ω
Max. output level (sine):	12dBm (3.1V rms)
Attenuation:	64x1.5dB steps plus mute
Thd+noise (sine)	0.015%
Analyser	
Type:	2-channel 16bit sigma-delta a-to-d converter
Input range:	+30 to –40dBV
Input impedance:	64kΩ (2.7kΩ mic. input)
Phantom power:	8.2V
Miscellaneous	
Sample frequency:	51.2kHz, 12.8kHz, 3.2kHz
Card type:	14cm 8bit pc slot card
Card connections:	four RCA plugs
Adaptor cables:	for speaker, *LC* tester
Microphone Mic–01 (Mic–02)	
Type:	Condenser electret
Accuracy:	±1dB 20Hz-10kHz, ±2dB 10Hz-20kHz (±2dB 10Hz-20kHz direct field)
Max. level	120dB spl
Dimensions:	
MIC-01:	8mm dia. 25cm long
MIC-02:	As MIC-01 but 12cm long
Accessories:	3m cable, stand adaptor
Preamp 3381/A	
Type:	Rechargeable battery,individually calibrated for MIC-01 or MIC-02
A-weighting filter:	IEC651 – Type 1
Level range:	70-120dB spl
Attenuator:	10dB step, 0.1dB accuracy
Accessories:	AC mains adaptor

Fig. 6. Distortion spectra of Clio generator looped into input and set at –3dBm (–5.25dBV), showing second and third harmonic both at 83dB down on the fundamental. The thd meter reads 0.01%. Drop the input lower and the thd reading rises. Level can rise to +1.5dBm (–0.77dBV) with change to spectra but thd reading stays at 0.01%. Distortion spectra could be bettered by 10dB or so by using an external low-distortion oscillator carefully set in level and frequency, with windowing, indicating that most distortion is from the internal sine generator. Though clearly meeting its specification, with 16 bits available, distortion could possibly be bettered. Muting the generator (input set to –40dBV) produced a noise floor of around –105dBV — not at all bad

Switchers *for the masses*

Highly-integrated chips and easy-to-use software bring designing a switch-mode psu within the reach of all electronics engineers – as Al Kelsch and Wanda Garrett show.

*Al Kelsch Wanda Garrett are with National Semiconductor Corp. in Santa Clara, CA

In many applications, on-board dc-dc conversion – sometimes called point-of-load conversion – is desirable and practical. National Semiconductor's Simple Switcher power converter family concept was created to occupy a position between complex, design-intensive circuitry of a fully-custom switching converter, and the simplicity of plug-in power conversion modules.

Advantages are lower cost, compared to modules, and reduced design expense relative to discrete solutions.

This article reviews the principles and viewpoint which led to the the Simple Switcher converter family, and highlights ease-of-use aspects of the concept.

Henry David Thoreau, the iconoclastic American writer of the nineteenth century, stated the case for simplicity in 1846 while searching for truth at Walden Pond:

"Simplicity, simplicity, simplicity!, I say, let your affairs be as two or three, and not a hundred or a thousand."

William of Occam, the Medieval English philosopher, dealt with the subject of simplicity in the metaphysical realm before expiring in 1349. He is best remembered for what has become known as Occam's Razor. William's famous razor says that if more than one theory is advanced to explain the same set of facts, the theory that is overall the simplest and needs the fewest assumptions is the true one.

Occam's Razor applies to integrated circuits, too. If two integrated circuits do basically the same job, the one that is overall the simpler and requires the least design time and external aids is the superior product.

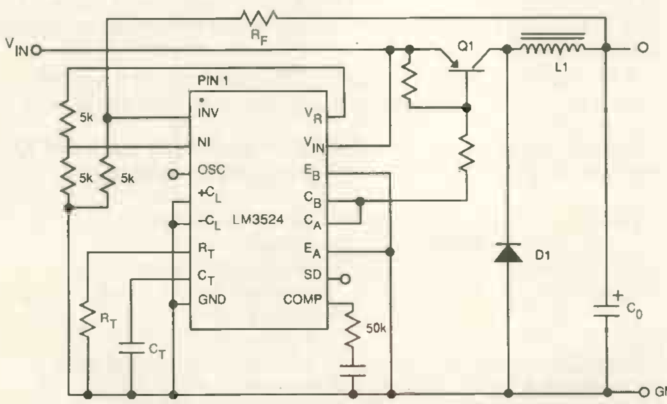
In the not-too-distant past, it was common to design a voltage regulator as a part of designing the end equipment. Whether it was a simple zener diode regulator or a real series voltage regulator, on-board voltage stability was considered to be a part of equipment design. This all changed with the advent of the first three-terminal linear regulator – the *LM109*.

From the time these devices were available, equipment designers everywhere ceased designing the regulator part of their circuits from the transistor level. They found themselves free to spend more of their engineering energy on the performance of the equipment or chassis itself.

Prior to 1978, the task of efficiently converting a dc voltage to another dc voltage using switching techniques was done with discrete transistors, diodes and passive components. At that time, there were two possibilities open to electronic equipment designers that had made the decision to enjoy the benefits of switchers — namely smaller size and increased efficiency. They either made use of an in-house power supply design capability, or opted for off-the-shelf dc/dc converters.

The first integrated circuit for switching power conversion – the *SG3524* – was introduced by Silicon General in 1978. This type of design was very complex, compared to a linear voltage-conversion solution, **Fig. 1**. However, it did represent an advance compared to bottom-up discrete switching designs.

At the same time, an industry for solving the complexity problem associated with switchers was emerging. The on-board conversion of one voltage to another, such as +24V to +5V, could be accom-



Design equations

$$RF = 5k\left(\frac{V_o}{2.5} - 1\right)$$

$$R_{cl} = \frac{200mV}{I_o\,(max)}$$

$$F_{osc} = \frac{1}{R_t C_t}$$

$$L_1 = \frac{2.5\,V_o(V_{in} - V_o)}{I_o V_{in} F_{osc}}$$

$$C_o = \frac{(V_{in} - V_o)V_o T^2}{8\,V_o V_{in} L_1}$$

Fig. 1. Even though early switch-mode ICs integrated a fair number of control functions, power supplies incorporating them were still complex in design and component hungry.

Table I. DC converters compared.*

Parameter	LM352-based 1A, 5V converter	Module 1A, 5V converter	Simple switcher design
# of components	16	1	5
Design time	months	minutes	<l hour
Parts inventory	high	lowest	low
Parts price	$2.10	$9.00-$13.00	$3.50
Labour	high	lowest	medium
Efficiency	80%	65%	80%
Offerings	one device, many designs	broad line	broad line

*source National Semiconductor.

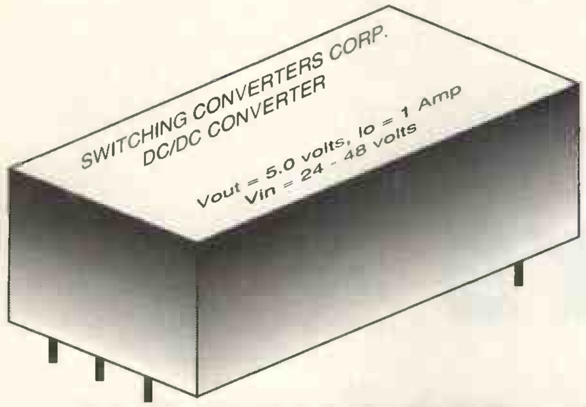


Fig. 2. Allowing designers to reap the benefits of switch-mode power supplies without incurring all of the associated design time and cost penalties, standard modules have become a viable alternative to discrete circuitry.

plished in minutes by the decision to purchase a self-contained dc-to-dc converter in the form of a potted module or hybrid assembly, **Fig. 2**.

Opposite ends of the on-board dc/dc converter spectrum in the early 1980s were represented by the self-contained power component on the one hand, and the *LM3524*-based IC design on the other. These are compared in the left-hand columns of **Table 1**.

Thoreau's simplicity of viewpoint, William's logical efficiency, and the need for a counterpart to the three-terminal linear regulator came together in 1985 to spark an idea for a new class of on-board dc/dc converters. This five-terminal integrated switching converter became known at its introduction as the Simple Switcher. This new class of converter took up a position between the one component $10.00 solution and the 16-component $2.25 solution, with a five-component $3.50 solution. This intermediate market approach is compared to the competing solutions in the fourth column of Table 1.

There were four elements in the creation of an intermediate position of switching converters that appealed to non-expert digital and microprocessor equipment designer; the sum of these became the 'simple' family approach. First, the 16 component dc/dc converter had to be reduced to the absolute minimum number of components possible in the context of an IC solution. Second, the procuring of the non-IC components had to be straightforward, foolproof, and well-supported. Thirdly, any design that was required had to be ultra-simple and had to be supported with friendly software. Finally, the overall convener performance had to be guaranteed with system-level specifications.

Reduced component count

Table 2 summarises the system decisions that led to the paucity of external components seen in the Simple Switcher type converter.

Bringing the 1A transistor on board eliminated three external components. On-board trimming to standard voltages such as 3.3V, 5V, and 12V eliminates two components. Fixing the frequency at a standard value of 52kHz eliminates a capacitor. Fixing the current limit internally and offering a range of current values, such as 0.5A, 1A, and 3A eliminated a current-sense resistor. Eliminating a Bode compensating network in favour of on-board compensation reduces the count by two more. Removing a resistive level setting network in the feedback path eliminated two more components. In this manner, a sixteen-component design becomes a five-component converter, **Fig. 3**.

The five-lead minimum design converter uses two electrolytic capacitors, one inductor, one Schottky diode and one integrated circuit. The design value of the on-off switch – something not actually offered in the 16-component design – dictated its inclusion as the fifth active pin.

Readily available components

The four non-IC external components that remained also had to be relatively free of the fear factor when viewed by non-expert designers, with regard to the design and the actual procuring of parts.

Aluminum electrolytic capacitors were specified for the input and output capacitors. These are readily available and popular. The Schottky is slightly more challenging, but these devices are likewise readily available and easily specified.

By far the most important challenge in guaranteeing ease of use was designing and procuring the power inductor or transformer that is at the heart of switching converters.

Here the fear factor plays a dominant role. Gauss, B-H curves, magnetising inductance, core materials, custom magnetics – all of these words and the practices associated with them have the potential to affect design time and difficulty level.

To remedy this, we took several actions. First, the design procedure had to be simple and unambiguous; second, there had to be pre-specified, industry standard inductors, suitable to the converter application. Third, the standard inductors had to be supported by magnetics vendors that could offer stocked values and instant support.

Calculating an inductance value is just the start of the problem. Given simple assumptions, the calculation is straightforward. The challenge was to reduce the solution space for all combinations of input/output voltages and load currents, to a small number of well-defined inductors.

Figure 4 illustrates the type of mapping needed to fully define converter solution space in terms that are meaningful to inductor manufacturers. The vertical axis on the graph turns out to be calibratable in E-T, or the volt-second product needed for full specification of the inductor in magnetic terms.

Knowing maximum load current, and E-T values as dictated by input and output voltages, the solution space is divided up into standard inductance regions. These are then specified in inductor terms. This lead to a well-specified series of inductors that covered the converter solution space.

With this set of basic inductor specifications, we then approached several magnetics manufacturers that had the capability and interest to

Table 2. Comparison of external components required for typical 1A step-down switching regulator design.

Function	LM3524-based design	Simple Switcher design
1A o/p current	Transistor, 2 Rs	On-board >1.3A transistor
V_{out} setting	2 Rs	Internally-set, fixed output voltage options
f_{osc}	1 C	Internal, 52kHz
Current limit	1 R	Internal, fixed (several options)
Compensation	1 R, 1 C	internal
V_{ref} level	2 Rs	Internal
Net component reduction	0	11

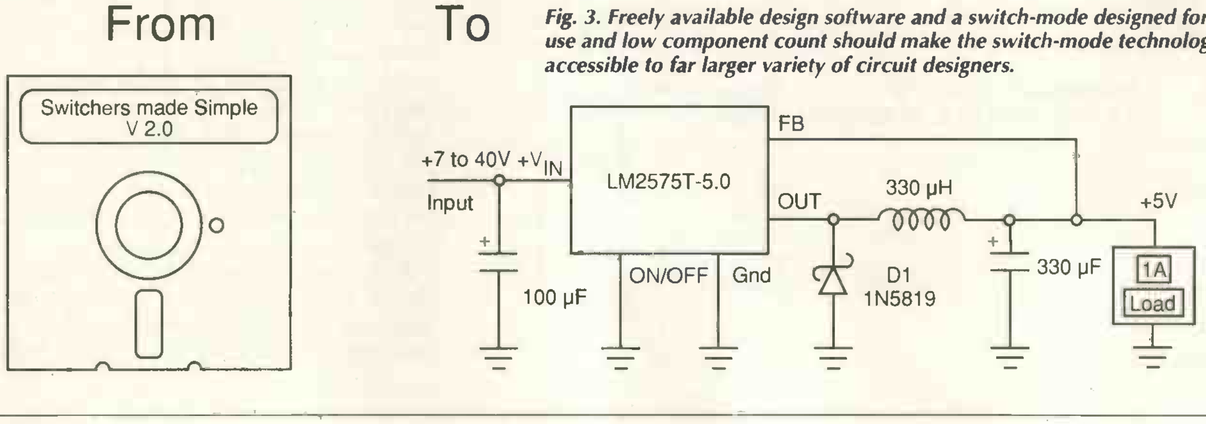


Fig. 3. Freely available design software and a switch-mode designed for ease of use and low component count should make the switch-mode technology accessible to far larger variety of circuit designers.

Customising the design

Although a successful power supply design can be generated by using the default or standard options in the program, there are opportunities for customisation. Alterations can be made to a design that has been previously saved, or during the design of a new regulator.

It is a good idea to have a 'baseline' design to start from, using the default choices for components. Modifying the inductor selection For the buck and boost converters, the standard inductor selection is based on an assumption that the inductor ripple current will be 30% of the average inductor current at fill load. This gives continuous operation, striking a balance between inductor size and output power capability.

Output power is for a given regulator is limited by the average inductor current; inductor ripple current is determined by the inductor value, the input and output voltages, and the switching frequency; the sum of the average and ripple inductor currents is limited by the regulator IC current limit threshold, as the IC must conduct this peak current when its switch is on.

For low output power applications, the inductor can afford to have a greater ripple current since its average current is low. This allows its value to be lower than the normal standard value that would be chosen automatically. As the inductor selection is made immediately after the input parameters are entered, you can take the option to not take the standard default inductor. The user can then customise the inductor selection by either entering the desired inductor ripple current or a desired inductor value, and the remaining components will be chosen accordingly.

If the choice of inductor value or current causes the peak current to exceed the limit available from the regulator IC, an error message instructs the user to make another choice between lower current or greater inductance.

When the choice is accepted, the user should check the peak switch current, shown in the middle column on-screen, to see how close it is to the current limit of the regulator IC chosen for the final circuit.

The output capacitor value and ESR can be modified after the preliminary calculations and component choices have been made. Recommended ranges for those values are given in the 'limit values' column. These ranges assure the stability of the regulator, and a reasonable limit for output ripple voltages. Modifying databases used or component selection 'Switchers Made Simple' selects the recommended components from a set of databases associated with the program. These are Ascii files, containing component values, characteristics and ratings, and vendors' part numbers for components including the output capacitors, inductors, and diodes. The user can extend the databases to include other components, if they are specified for use in switching regulators.

The database format is shown in a header for each file. In addition, the header also contains the strategy for component selection, for example which parameter is looked for first.

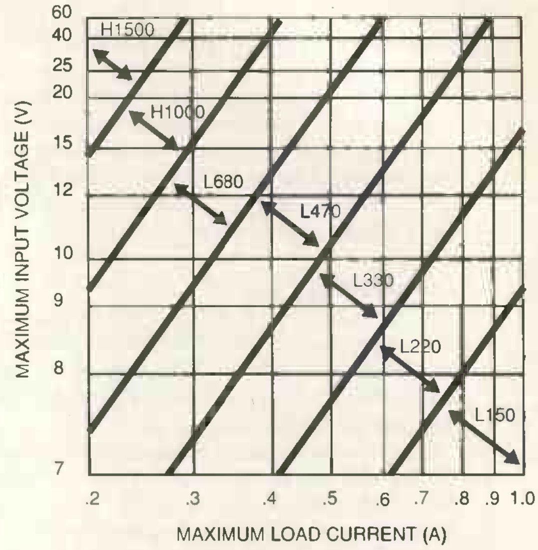


Fig. 4. With the Simple Switcher concept, choosing an inductor is simplified by focussing on a range of readily available standard parts.

Standard inductors and sources.

Inductor code	Inductor value	AIE	Pulse Eng.	Renco
L150	150μH	415-0953	PE53113	RL1954
L220	220μH	415-0922	PE52626	RL1953
L330	330μH	415-0926	PE52627	RL1952

support this effort at standardisation. This level of interdependence had not been attempted before between the semiconductor and the magnetics industry.

When choosing inductor suppliers, we attempted to offer a range of styles and prices from the stick inductor, to the popular toroids, to low-EMI pot cores. Once on-board, the magnetics industry has done a fine job of supporting first-time converter designers. This included no custom charges – or any talk of charges – stocking of all values, overnight response and attractive prices.

Create new or recall old design? (N/O)

This program supports four types of power supplies.

Boost:
Used to step up the input voltage, e.g. Vin = 5V, Vout = 12V

Flyback:
Used for multiple output voltages, positive or negative, with the possibility of isolation. Both step up and step down are possible. High output voltages are achievable. A transformer is needed instead of an inductor.
E.g. Vin = 5V, Vout1 = 15V, Vout2 = –15V, or
Vin = 5V, Vout1 = 15V, Vout2 = 12V, or
Vin = 20V, Vout = 100V

Buck: Used for stepping down a voltage, e.g. Vin = 10V, Vout = 5V

Buck-Boost:
Used for generating a negative voltage from a positive one without isolation, e.g. Vin = 5V, Vout = –5V

For HELP, press <F1> now.

Fig. 5. Sample help screen from the Switchers Made Simple design package.

Perform thermal check? (Y/N)

Input parameters	Limit values	Component values
Vinmin = 8.00 V	L > 382.00 μH	Cout = 330.00 μF
Vinmax = 30.00 V	Mode = Continuous	Cin = 22.00 μF
Tamax = 40.00 C	Iswpeak = 0.40 A	L = 470.00 μF
Tamin = 0.00 C	Cout > 330.00 μF	D1 = 1.00 A
Vout = 5.00 V	ESRmax < 0.15 Ohm	
Ilmax = 0.30 A	ESRmin > 75.17 mOhm	
Diode = Schottky	Cin > 22.00 μF	
	Vripple = 50.34 mV	

Crossover 7.43 kHz
Phase margin 31.15 Deg

Fig. 6. Regulator design summary screen from the Switchers Made Simple design tool.

Guaranteed system specifications

Another key aspect of creating successful dc/dc converter designs involved the use of system specifications. Hence, the ICs in the Simple Switcher family include guaranteed system specifications when used with the external components recommended by NS. The basic parameter guaranteed for the overall converter is V_{out} – the output voltage. As a system specification, V_{out} is guaranteed over a full range of load, line, and temperature variations.

Ease of design

To convince first-time designers and analogue non-experts that a realisable dc-to-dc converter could be designed efficiently, a simple step-by-step procedure was needed. Also, the design procedure had to be executable either manually, or using a pc with the aid of an expert system provided by National.

For the manual designs we created a one page, three step procedure for the selection of the capacitors, the catch diode, and the inductor. Sources for each of these were provided in the data sheets, together with the corresponding vendor part numbers.

The second support method, aimed specifically at terminal-based designers, was the software called 'Switchers Made Simple.' This is a software tool, designed from the viewpoint of the user, that creates dc/dc converter designs using the Simple Switcher integrated circuits and standard external components. The software runs under dos and is free so sharing is encouraged.

The following section explains the use of the 'Switchers Made Simple' software in designing onboard dc/dc power converters. The current version is 3.3, and supports all available Simple Switcher products and their voltage options – including 3.3V for the step-down and inverting regulators.

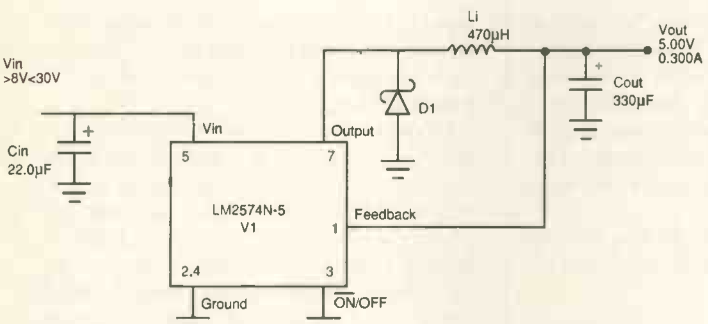


Fig. 7. Switch-mode power supply example incorporating the LM2574 Simple Switcher demonstrates how few external components the device needs.

Fig. 8. Parts list for design example, Fig. 7.

BUCK CONVERTER

Circuit Parameters

Vinmin :	8.00 V
Vinmnx :	30.00 V
Tamax :	40.00 C
Tamin :	0.00 C
Vout :	5.00 V
Ilmax :	0.30
Diode : Schottky	

Misc calculated information

Mode : Continuous
Peak switch current : 0.40 A.
ESRmax : 0.15 Ohms
ESRmln : 75.17 mOhms
Vripple : 50.34 mV
Crossover Freq : 7.43 kHz
Phase margin : 31.15 Deg

Component List

Cout : 330.00 µF
ESR : 0.10 Ohm
Vmax : 20.00 V
678D337M020CG4D : Sprague
Cin : 22.00 µF
Vmax : 43.00 V
L : 470.00 µH
415-0927 : AIE
PE-5311C : Pulse
RL1951 : Renco
D1 : 1.00 A
Vmax : 40.00 V
1N5819 : Motorola
MBR140P : Motorola
11DQ04 :
U1 :
L2574N-5 : National Semiconductor

Beginning the design

The only information a novice power supply designer may have on which to base a design may be the system parameters – input voltage range, output voltage desired, load current range, and ambient operating temperature range. This is the way a linear regulator or module is specified. This is also the way a Simple Switcher design is started.

On the first screen of 'Switchers Made Simple,' **Fig. 5**, following the introductory screens, the user is prompted for regulator type from a choice of step-up, step-down, invert, or multiple-output. If, at this point, the user wants more information, a help file is available via the F1 key.

Once the regulator type is chosen, a screen appears on which the system parameters are entered. Pressing the 'end' key, after entering the parameters, begins the process of doing calculations and choosing components for the circuit.

Component selection

The software asks the user if the standard inductor selection is to be made, and if this is confirmed the software makes calculations for limit values. The results of these calculations are shown in the middle column of the screen. They include values used to select the components used in the final circuit, such as minimum inductor value, peak switch current, and values for the output capacitor and its equivalent series resistance, esr, range.

Recommended component values are then immediately shown in the right-hand column of the screen, and include all external components needed in the design apart from the regulator IC. The user is then given the opportunity to modify the component values.

When the user is satisfied with the component choices, the frequency response of the circuit is estimated, and a window displays the circuit bandwidth, 'crossover', and its phase margin **Fig. 6**.

Thermal calculations

The next step in the design is to determine how much – if any – heat sinking will be needed to keep the internal temperature of the regulator IC within its ratings during its operation over the specified temperature range. The user is asked if thermal calculations are to be done. If so the user is asked to choose the package type from a list of regulator IC packages that fit the application.

If a dual-in-line or surface-mount package is chosen, the pc-board copper will be used for heat sinking, so the user is requested to enter the thickness and type of copper that will be used. The software calculates the minimum copper area required to provide any necessary heat sinking, and allows the user to enter the actual board area that will be used. Finally, the internal junction temperature of the IC and its thermal resistance are displayed in a window.

The design is now finished. The user is given an option to modify the input parameters. If this option is not taken, the user may save the file for further reference. The user can view the schematic on-screen **Fig. 7**, as well as the listing of calculation results and recommended circuit components by part number, **Fig. 8**.

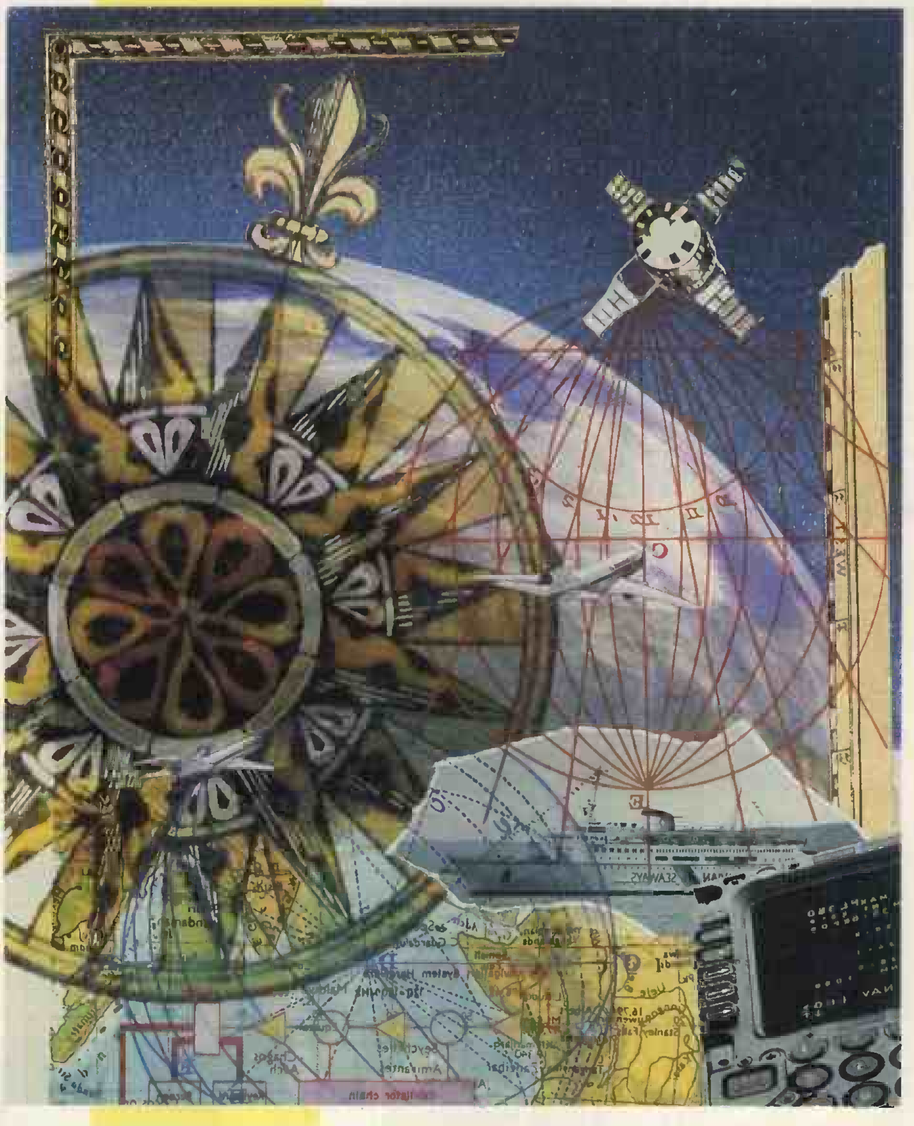
Both the schematic and parts list can be then printed, on either an Epson-compatible dot-matrix or an HP-compatible laser printer. As a last step, the user may return to the beginning of the program to do another design, or end the program.

In most cases, the power supply design can be done in a matter of a few minutes, from specifying the circuit parameters all the way to the final schematic and parts list. This simplicity and speed of design has made the Simple Switchers the choice of many novices and power supply designers alike. In addition, 'Switchers Made Simple' offers experienced power supply designers the flexibility to customise the design, as described in the panel.

Summary

The 'Simple' approach to on-card dc/dc power conversion has taken up its market position between plug-in, pre-manufactured modules, and complex multi-component IC-based converters. Simple Switcher power converters provide a solution that has five or seven total components guarantees overall converter performance, and is well supported by readily available components and design software. ■

Finding GPS



GPS chip sets are still difficult to obtain and use, but there is now a versatile modular receiver system priced under £400. Simon Taylor* describes how this module is applied.

*Simon is an applications engineer with TDC

Much has been written about Global Positioning Systems – GPS – but until now there has been little scope for experimentation with the technology.

Many hand-held GPS devices are available, but few of these have any output or control inputs apart from the standard lcd and keypad. If you wish to create a vehicle location system for example, you will need to get position information from the device over a serial port or similar.

This article describes the construction of a GPS receiver which provides its information over a standard serial RS232 port suitable for input into a pc or other processors, such as a microcontroller.

As its core, the system uses GPS receiver module from Rockwell Telecommunications called the MicroTracker LP. Rockwell is the company contracted by the US Department of Defense to build the current block of GPS satellites, and has made GPS receivers since the positioning system's inception.

The first readily available receiver module – the NavCore V – became available in 1990 and was fairly power-hungry requiring about 1.7W in operation. This module was replaced by the MicroTracker in 1992, which reduced the size and power requirements dramatically, and latterly the MicroTracker LP in 1994.

MicroTracker LP needs only a single 5V supply, and consumes about 900mW, which can be further reduced under software control. The module can supply power to active antennas, but these are not normally needed, as the device has an extremely 'hot' rf front end.

Before continuing with the facilities of this module, it is worthwhile considering some of the relevant parameters in the construction of a GPS receiver.

Time-to-first-fix (ttff)

When a GPS receiver is switched on, it will not be receiving any satellites. At any time, there will be a maximum of twelve satellites in view, out of the full operational constellation of twenty-four. In practice about eight or nine are in the visible sky.

The receiver needs to lock-on to a satellite, and will start to search for other satellites which are in the sky at the same time.

There are many ways in which this can be done. First, a random selection can be made, from any of the twenty-four satellites.

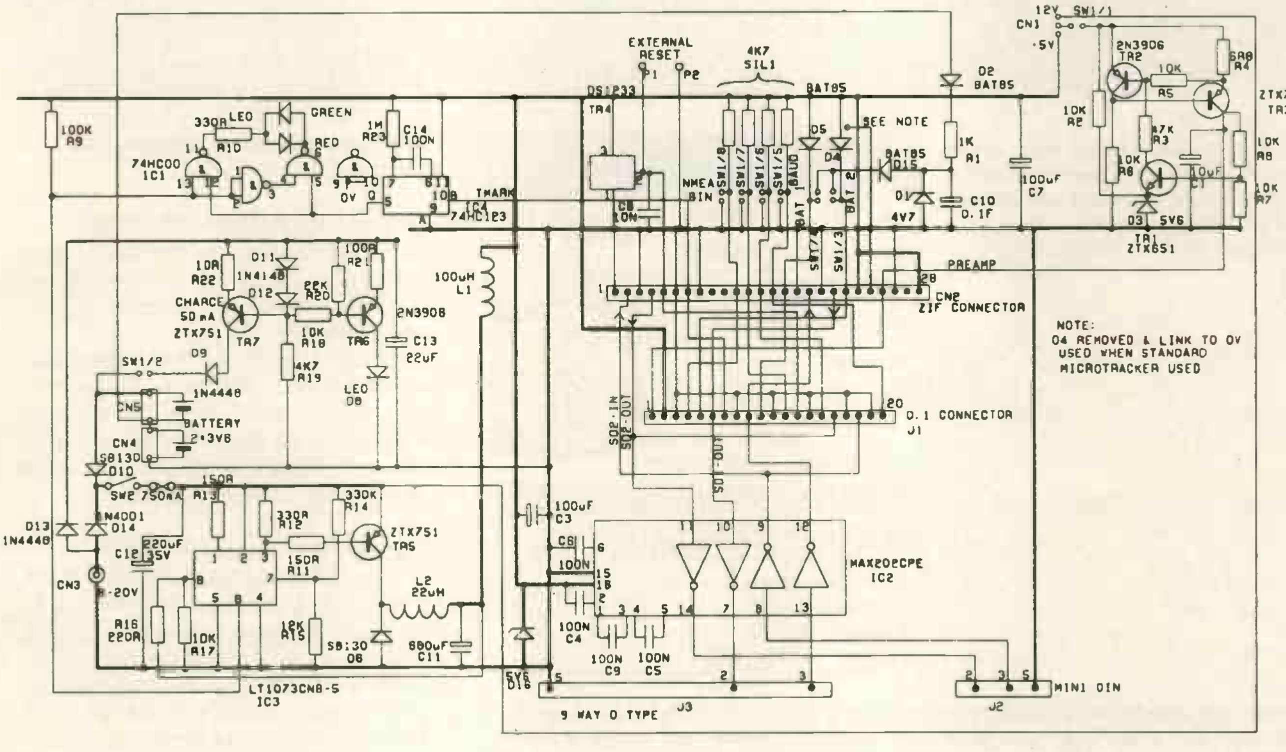


Fig. 1. Circuit diagram of the GPS receiver serial interface board and its power supply.

However, if the receiver 'knows' which satellites *should* be in the sky at that moment, then the search process is much simplified, and hence much quicker.

Periodically, the satellites download an 'almanac' to the receiver, describing their orbital characteristics. This almanac goes out of date, but while still fresh, it is of use. So, if the receiver knows its approximate position – within 100km will not make much difference – and the approximate time, then ttff can be reduced.

Most receivers will not have moved geographically very much while switched off, if at all, and it is usually relatively simple to incorporate a real-time clock into the system.

MicroTracker LP incorporates a real-time clock, and retains its last position in eeprom, so the ttff is quoted as about 30s. When the receiver is completely un-initialised, for example as they are when supplied from the factory, a 'cold start' algorithm can be enabled.

Note for Psion Organiser users

Software allowing the Psion Organiser II to read NMEA data from the Rockwell GPS receiver is available.

In default mode, the Psion display shows latitude, longitude, height in metres, speed in mile/h, heading, horizontal dilution of precision (an accuracy figure), time and differential indicator.

The differential indicator shows 'Diff' when differential corrections are being used by the receiver. MicroTracker LP accepts RTCM-104 corrections directly into an auxiliary serial port and the number of satellites being received.

Key options are :

EXE selects odometer mode, this replaces the height display with an 'odometer' showing the distance covered since the program started. Distances up to half a mile are displayed in feet.

SPACE Display NGR (National Grid Reference). This displays NGR, then waits for a key press before continuing.

DEL Quits program.

Example display with odometer mode, differential corrections applied;

```
N51°16.441'          359.3°
W001°03.916' 37.8
 1.67 Diff 13:07:49
hdop=0.78            Sats 04
```

In this display, 359.3 refers to the heading, 37.8 to speed in mile/h, and 1.67 is the odometer reading. The latter is replaced by a height display in the default mode.

Exclusive EW+WW reader discount

A complete designer's pack is available from TDC at a discount exclusive to EW+WW readers. The kit includes the Evaluation board, MicroTracker LP receiver, Magnetic mount antenna, mains power unit, serial cable, batteries, technical manual and software. Its price – which includes a seat on one of the GPS seminars – is £379.00 exc. VAT and postage. Normally, the combined elements of package would retail at £465.14. All elements of the kit are available individually at special prices to readers quoting this article.

The complete system is tested for GPS reception before despatch. Please add carriage at £12.50 (TNT next day), and VAT to the total.

Please send a cheque/PO or Credit Card details (VISA/Mastercard) to Telecom Design Communications Ltd, Connect House, Stroudley Road, Basingstoke, RG24 0UG.
Tel 01256 332800, fax 332810.

Fig. 2. Screen shot of Binary mode reader program. PC software for this is included in designer's pack. It convert GPS receiver output into user-friendly display form. Note that display includes OSGB coordinates, bottom left.

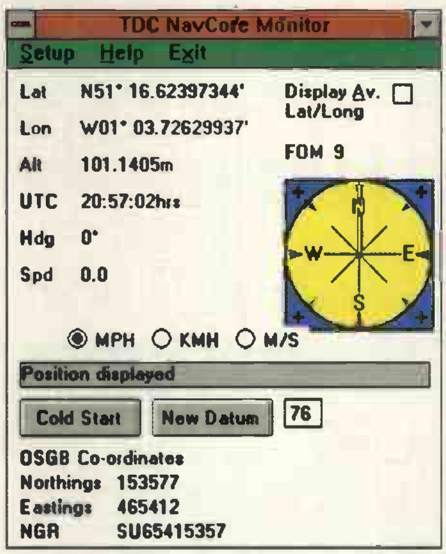
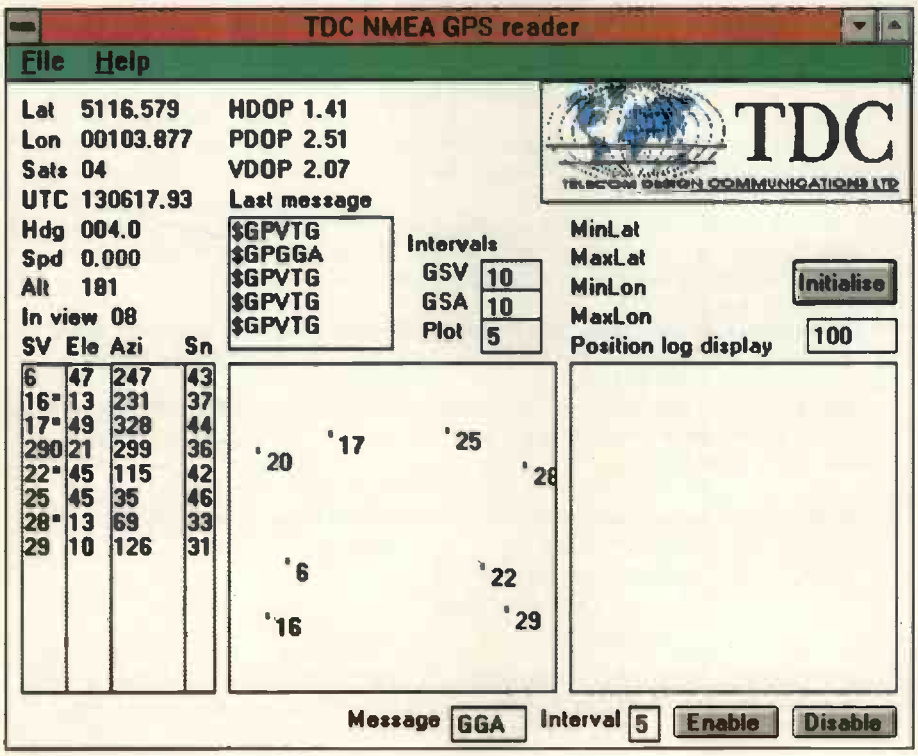


Fig. 3. Screen shot of industry standard NMEA reader program, also included in kit. Displays GPS information on a pc but also shows positions of satellites in the sky and a tracking display showing a history of your position.



This will normally find a position within about seven minutes.

Numbers of satellites required

Only one satellite is needed in order to receive accurate time. All of the satellites transmit UTC (Universal Time Co-ordinated) which is almost identical to GMT.

To compute a position, four satellites are required initially. This gives a position in three planes – X, Y, Z or latitude/longitude/height – but thereafter only three are required to give what is called 2D navigation comprising latitude/longitude only).

Antennae & rf characteristics

GPS signals are transmitted on 1574.42MHz, All of the satellites transmit their data on the same carrier frequency. Sophisticated techniques are used to identify individual satellites from the combined signal, hence the need for a search strategy as described above.

The signal is available to the whole of the world for free use without licensing, so there is no restriction to the use of the information. But the position information is degraded by the US DoD so as to not be useful for military purposes. The accuracy is quoted as better than 100m for 95% of the time, but again has been found to be much better in practice.

The signal is right-hand circularly polarised to avoid distortions and errors due to reflections from buildings etc., but this frequency does of course require a fairly sophisticated antenna. Such antennas are commercially available, and can be obtained from around thirty pounds. It is not advisable to construct an antenna as the signals being dealt with are at very low levels, and adequate performance can be difficult to realise.

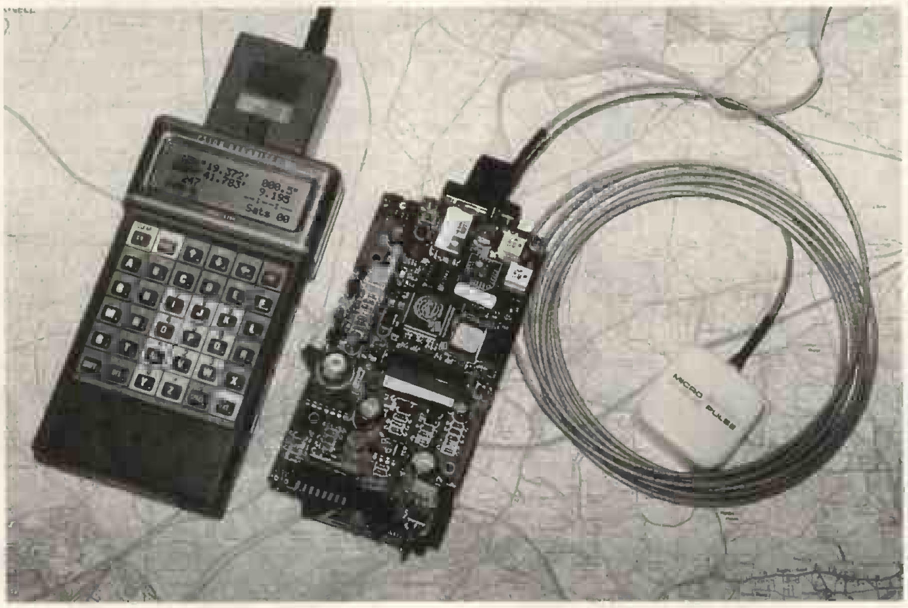
Receiver details

There are two connector options on the Microtracker receiver board. First is the zero-insertion-force connector, which has historically been the preferred type. However, with lower volume manufacture now possible for these systems, the 0.1in connector is becoming more popular due to its cost, popularity and availability.

Connections of interest are on the 0.1in connector are:

GND power supply
5V power supply
BAT1 backup supply for internal ram, which contains up-to-the-minute satellite information, improving ttff.
BAT2 backup supply for the receiver's internal real-time-clock, again, to improve ttff.
TXD data output is at ttl levels and must be level via an RS232 buffer shifted before connection to an RS232 device.

Fig. 3. Photograph of the GPS receiver interfaced to a PSION organiser also shows NMEA reader, and the Micropulse antenna.



RXD data input, again at ttl level.
NMEA/BIN selection for data format output

Data formats to and fro

Information on data formats is of interest to those of you who wish to develop your own software applications for the receiver, but it is possible to obtain position information using software packages readily available from me and on CompuServe.

There are two data formats handled by the receiver. The first is the proprietary 'binary' mode, which consists of a number of defined messages which allow the user to initialise and read data from the receiver.

Binary-mode messages are 'packetised' with a message header, information fields and a checksum. In this way, it is easy to detect if a message has been corrupted, and should be discarded. Data being sent to the receiver is similarly 'packetised'.

The second data format is the universally recognised National Maritime Electronics Association (NMEA) message. This message format is compatible to many marine display systems, but more importantly, outputs data in an ASCII form, which can be easily read using a terminal program.

Writing software to take this data from a serial port, and display in whichever form the user requires, is straightforward.

Hardware detail

GPS is suited to mobile use. Generally, the only exceptions to this rule are timing applications where the 1μs accuracy of the GPS receivers is exploited.

Taking the example of a car, then the most convenient power source is likely to be the car battery, which has a very wide range of possible voltages. These extend from around 11V up to 17V in the event of a fault.

Another example of mobile use is in a portable system, where independent batteries are desirable. As with all portable systems, battery life versus weight is the main concern. Nickel-cadmium batteries are desirable, and it is useful to be able to charge these while operating the device from a car.

MicroTracker requires a regulated power supply of five volts, and active antennas may require twelve volts – again regulated – although some will operate on only five volts.

The circuit shown has two efficient power converters to provide both the five and twelve volt requirements from a wide input supply. Battery charging with charge & low-battery indication are also provided. A dual-colour led flashes once a second while the GPS system is running. It flashes green while the battery level is sufficient, and red when the battery becomes depleted. A red led indicates charge going to the cells.

Using six 700mAh 'AA' size cells, there is enough operating power to supply the MicroTracker LP for about five hours, and to provide backup to the rtc and ram while the device is switched off. If backup is only required for short times, then a 0.1μF capacitor can retain this data for about an hour.

If the operating specification is different, then various other types of battery can be used to suit the application. Power requirements can be reduced by enabling power save modes where a reading is provided only once every five seconds rather than the usual one second update rate.

MicroTracker's serial interface is at ttl levels, so the circuit also includes RS232 level conversion to allow connection to an external computer. A second serial port is included to cater for differential GPS corrections to be input to the receiver to improve the accuracy of the receiver to sub-10m. The primary serial port is in the standard 9-pin format.

The module will bolt to the pcb using stand-off pillars. Either connector type can be used, but if the zif socket is used, then an additional connection cable is required.

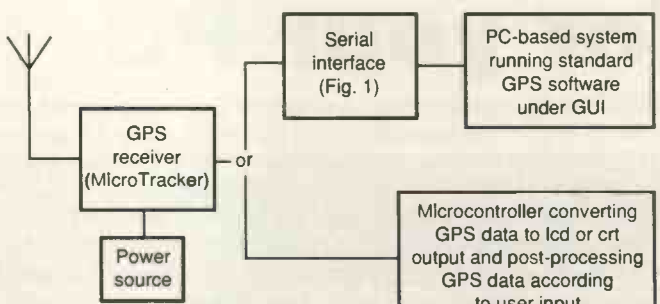
A PCMCIA version of the *MicroTracker* is available, called the *NavCard*, which provides the same functionality, but in a ready-to use PCMCIA format compatible with modern notebook computers. This card does not include NMEA data formats.

Software

If an existing NMEA reader program is being used, then provided that NMEA mode is selected, this should present few problems.

I suggested that in order to confirm correct operation of the device, you should use proven software. The binary mode reader illustrated is such a program. Running under Windows, the program displays position, altitude, heading and speed of the vehicle.

A bonus is that conversion of latitude/longitude to OSGB National Grid References is performed by the program. It is available on CompuServe, or from me at the address mentioned below. I have also developed an NMEA reader for the Psion organiser II, operating through the serial interface. ■



Complete system for reading and presenting GPS data. The GPS antenna, Microtracker and pc software providing the two displays shown in this article are all part of the desgner's kit.

The author can be contacted at TDC's address, mentioned in the special-offer panel, or via E-mail 100043.104@compuserve.com.

130MHz active probe

With a gain of 10 and flat to 130MHz this oscilloscope probe is a useful tool. Nick Wheeler has achieved this performance using surface-mount devices. He's also found an easy way of implementing the design.

Conventional passive probes, usable at even a few megahertz, have to have a compensated divider – usually ×10 – to minimise the effect of the lead capacitance. This division is sometimes inconvenient as it lowers sensitivity to typically 20mV/div.

Much worse, however, is the effect of mismatch, resulting in a pulse of the form of the lower trace in **Fig. 1** being transformed by multiple reflections into that of the upper trace. The lower trace was produced by transmitting from a 50Ω source down a 50Ω cable to a 50Ω termination immediately adjacent to the oscilloscope terminal.

This probe, **Fig. 2**, has a gain of ten, flat to 130MHz and an output impedance of 50Ω. Its input impedance determined primarily by stray capacitance of 5pF. Note that this is half that of a conventional probe. The 100kΩ resistor simply ensures a dc reference for the fet's G_1.

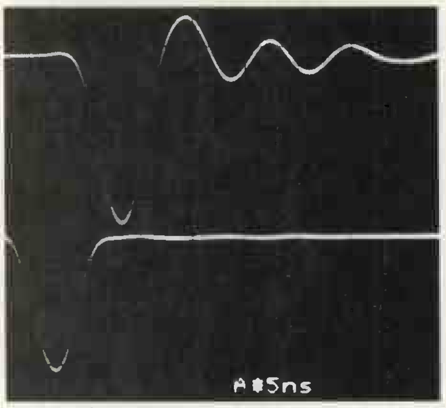


Fig. 1. Multiple reflections caused by a typical probe feeding a 100MHz oscilloscope, upper trace, are almost entirely absent in when the active probe is used, lower trace, in a matched 50Ω system.

SMD prototyping on a shoestring

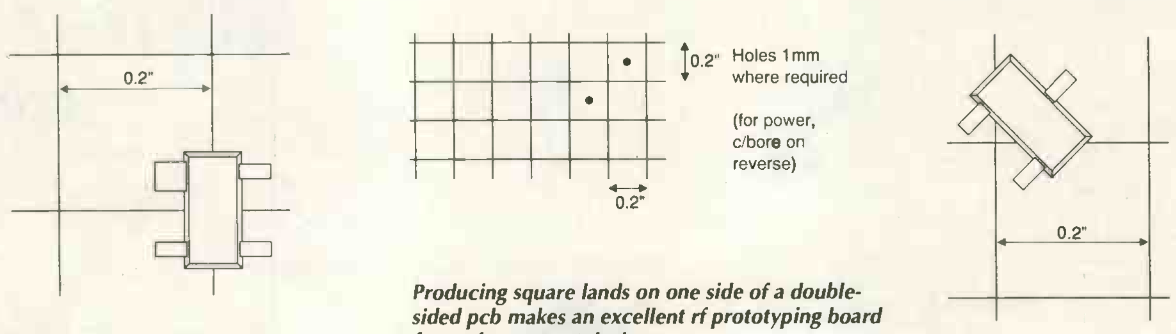
Although this circuit operates to 130MHz, standard double-sided glass-reinforced pcb can be used.

Clean both sides and coat with an etch resistant coating. Engineer's blue works well since it is designed to be scribed with fine lines.

Scribe a 0.2-by-0.2in grid on one side and etch below 50°C. This grid will take most 2, 3 and four-terminal smd parts, with one lead per land. Many other devices can be incorporated by removing a small amount of copper locally.

Where a ground connection is needed, drill a 1mm hole in the centre of the land and solder through.

Assuming 1.6mm glass-reinforced pcb material, each land has about 3pF relative to the ground plane. This can be lowered by removing copper from the ground plane behind the land, but carrying this too far may impair the grounding.



Producing square lands on one side of a double-sided pcb makes an excellent rf prototyping board for surface-mount devices.

The *CF739* GaAs fet, from Siemens, has a transconductance of about 30 and is specified up to 2GHz. Drain load is 100Ω in parallel with the 50Ω input impedance of the *MAR 6* silicon mmic, from Minicircuits. This yields unity gain. The mmic has a gain of 20dB at 100MHz and 19dB at 500MHz. Thus voltage gain is close to 10 down to a few megahertz, at which point the 10nF coupling capacitors become significant.

The fact that this probe has gain is useful in many cases, but it begins to distort signals above 10mV pk-pk and the fet will be at risk of damage at more than three times this level. By trading off some of the gain, various forms of frequency compensated front end protection can be applied. This needs to be done experimentally, as the very small capacitances involved depend on layout detail.

Application

Because the *MAR 6* mmic has to be capacitively coupled, this circuit will not work down to dc. But lower frequencies can be dealt with using ordinary probes. This circuit is intended for examining low-level rf signals and can be used with an input capacitor as small as 1pF. This forms a capacitive divider leading to a gain of about two – still useful.

A suitable type coaxial cable for linking to the oscilloscope is semi-rigid RG 402. ■

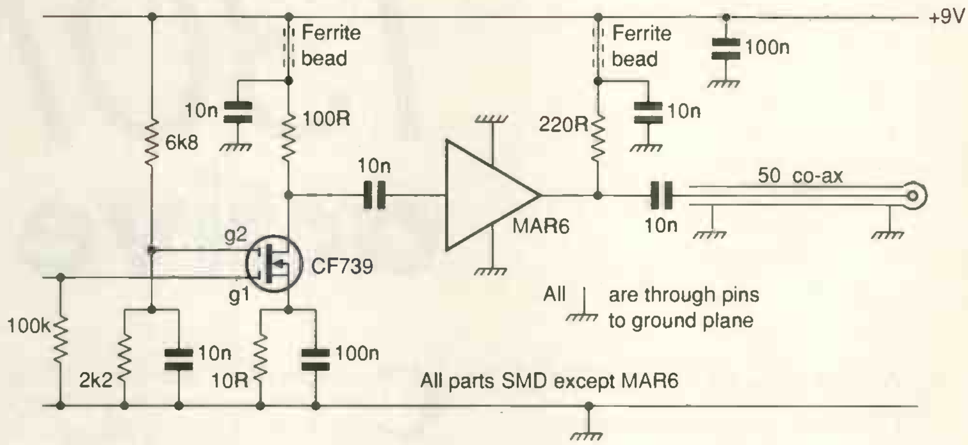


Fig. 2. Surface-mount components are ideal for an rf oscilloscope probe since they allow a physically small design with short connection distances.

Getting more from RS232

Douglas Clarkson builds upwards from a simple RS232 interface for accessing an a-to-d converter through to a multi-drop RS232 communication link.

Increasing use is being made of serial RS232 links to control equipment and read data remotely. Devices such as the Harris *HI7159A* device allow a-to-d conversion to be undertaken with control and data transfer taking place on a single RS232 port. This opens up a range of interesting interface possibilities.

Resolution of the device is $5^1/_2$ binary-coded decimal digits. It is capable of typically eight conversions a second at full resolution and offset compensation, or 60 conversions per second at $4^1/_2$ digits resolution without compensation. The device cannot be considered fast, but it does have the significant advantages of being relatively easy to use and of providing high-precision data capture.

By altering the device's conversion mode, its speed can be increased without compromising accuracy. This allows a series of measurements to be taken on a channel with an error-only value followed by a series of uncompensated values and ending with another error value. Normally the error value is constant and can be subtracted from the uncompensated value.

Such a system is ideal as an environmental monitoring station for example, measuring temperature, wind speed, light level and ultraviolet levels. It could also be used with load-cells to develop medium scale sensitivity weighing balances.

Various options are available for data capture. Serial mode 2 allows separate *HI7159A*s to be addressed independently, **Fig. 1**. In this mode, four separate analogue channels require four *HI7159A*s. Since the device costs around £20, implementing a large number of channels in this way becomes very expensive.

Accessing devices in serial-mode 2 is also slower since part of the serial traffic on the connected system is needed for addressing specific devices. Under mode 2, the total number of *HI-7191*s that can be connected together is 32.

An alternative way of increasing the number of analogue channels is to use standard analogue multiplexing. The single serial link can still be used to control multiplexing. Using four lines for addressing allows the system to accommodate up to 16 separate analogue channels. Having programmable gain in the circuit design is also useful.

Transmitting and receiving

Basically, the *IM6402* converts serial ttl logic data to parallel and vice-versa via two separate 8 bit data ports, **Fig. 2**. While the device can be configured for a range of data bit, parity type and stop bit combinations, in this application it simply has to be configured for compatibility with the *HI7159A*. The format is eight data bits, even parity and one stop bit.

Table 1 is a summary of the control word logic of the *IM6402* uart and indicates the full range of data bits, parity and stop bit permutations available.

Table 1. Control word logic of the IM6402. A wide range of data bits, parity and stop bits are catered for – including the specific 8 data, even parity requirement of the 7159.

Control word CL2	CLS1	PI	EPE	SBS	Data bits	Parity bits	Stop bits
L	L	L	L	L	5	Odd	1
L	L	L	L	H	5	Odd	1.5
L	L	L	H	L	5	Even	1
L	L	L	H	H	5	Even	1.5
L	L	H	x	L	5	Disable	1
L	L	H	x	H	5	Disable	1.5
L	H	L	L	L	6	Odd	1
L	H	L	L	H	6	Odd	2
L	H	L	H	L	6	Even	1
L	H	L	H	H	6	Even	2
L	H	H	x	L	6	Disable	1
L	H	H	x	H	6	Disable	2
H	L	L	L	L	7	Odd	1
H	L	L	L	H	7	Odd	2
H	L	L	H	L	7	Even	1
H	L	L	H	H	7	Even	2
H	L	H	x	L	7	Disable	1
H	L	H	x	H	7	Disable	2
H	H	L	L	L	8	Odd	1
H	H	L	L	H	8	Odd	2
H	H	L	H	L	8	Even	1
H	H	L	H	H	8	Even	2
H	H	H	x	L	8	Disable	1
H	H	H	x	H	8	Disable	2

x=Don't care

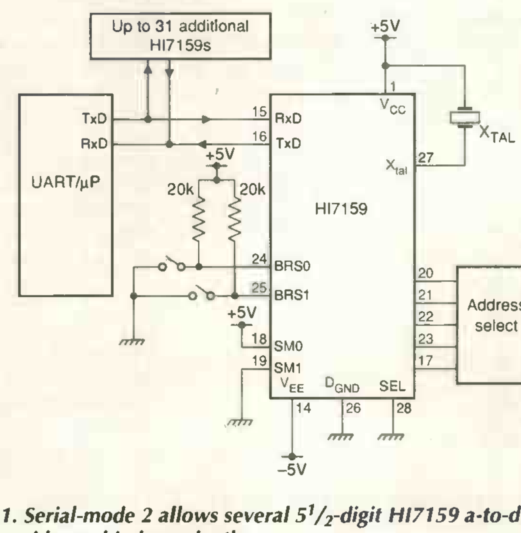


Fig. 1. Serial-mode 2 allows several $5^1/_2$-digit HI7159 a-to-d converters to be addressed independently.

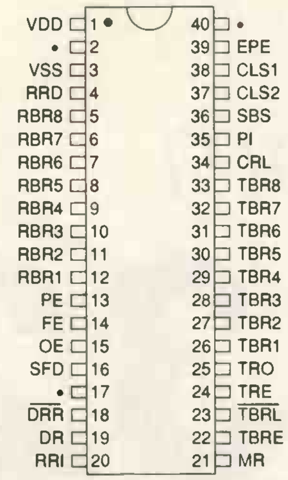


Fig. 2. Conveniently, the 6402 uart has separate eight bit input and output ports.

Table 2: Bit rates and corresponding receive frequency clock of the IM6402

Bit rate	Clock (Hz)
300*	4800
600	9600
1200*	19200
2400	38400
4800	76800
9600*	153600

* indicates frequencies appropriate for *HI-7159*.

Table 3. Truth table of the DG508A.

A_2	A_1	A_0	EN	Switch
x	x	x	0	none
0	0	0	1	1
0	0	1	1	2
0	1	0	1	3
0	1	1	1	4
1	0	0	1	5
1	0	1	1	6
1	1	0	1	7
1	1	1	1	8

Serial data signal levels for the *IM6402* are ttl compatible. A device such as the *NMC232* is needed to translate the logic levels. This device also provides isolation between the ±12V RS232 and ttl levels.

Clocking requirements

While the uart is accepting serial data, the receive clock frequency is 16 times the bit rate. **Table** 2 outlines the appropriate clock frequencies used with commonly occurring bit rates.

Options such as 1200 baud and 9600 baud are readily available from programmable clock oscillators with a base frequency of

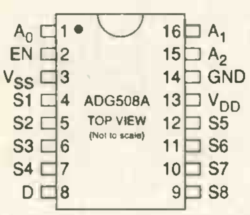


Fig. 3. One-of-eight switching is provided by this eight-way analogue multiplexer.

Listing 1. Simple routine outlining how to control the HI7159A/IM7402 combination via a pc.

```
REM the set of opening DECLARE statements are reserved
REM for the PDQCOMM library routines
DECLARE SUB openCom Action$)
DECLARE SUB ComPrint (Work$)
DECLARE SUB CloseCom ()
DECLARE FUNCTION ComInput$ (Nchar%)
DECLARE FUNCTION ComLoc%
5 CLS: REM clear screen
REM open serial port one at 1200 baud, even parity
10 CALL OpenCom("COM1:1200,E,8,1,RB128,NON")
REM and one stop bit
REM request value of analogue channel to be selected
REM the multiplexer chip use only three active
REM address lines
15 PRINT "INPUT CHANNEL VALUE TO BE READ 0 to 7"
INPUT CVAL
REM define line on IM7402 to be used for latching data
BOF = 64: REM LATCH ENABLE VALUE connected to B6 on HI7159AA
CALL ComPrint(CHR$(128 + CVAL+ BOF))
REM have activated latch to control analogue multiplexer
FOR JJ = 1 TO 100: NEXT JJ REM delay
CALL ComPrint(CHR$(128 + cval))
REM have disabled latch but analogue data being read
REM by multiplexer device
REM proceed to read data, provide options for choice of
REM command byte values all single - not continuous
REM 14 = 51/2 compensated
REM 12 = 51/2 uncompensated
REM 6 = 41/2 compensated
REM 4 = 41/2 uncompensated
REM 2 = error only
REM error only is used when wish to do fast uncompensated
REM conversions
20 PRINT "input command byte": INPUT sb
40 CALL ComPrint(CHR$(sb))
REM now send request byte to check for completion of conversion
50 CALL ComPrint (CHR$(13))
REM wait until data available from conversion
65 GOSUB 500: REM WAIT FOR LOC(1) TO SHOW CHARACTER
     a$ = ComInput$(ComLoc%)
80 IF ASC(a$) < 64 THEN GOTO 50
REM request another status byte to check if conversion complete
REM data now available: request data byte
REM SEND REQUEST FOR FIRST DIGIT PAIR
100 CALL ComPrint(CHR$(1))
     GOSUB 500: REM WAIT FOR LOC(1) TO SHOW data available
     D1$ = ComInput$(ComLoc%)
REM reading second digit pair
200 CALL ComPrint(CHR$(5))
     GOSUB 500
     D2$ = ComInput(ComLoc%)
320 REM READING THIRD DIGIT PAIR
310 CALL ComPrint(CHR$(9))
330 GOSUB 500
340 D3$ = ComInput$(ComLoc%)
350 REM now work out value of conversion
360 b0 = ASC(D1$) AND 15
370 bl = (ASC(D1$) AND 240) / 16
380 b2 = ASC(D2$) AND 15
390 b3 = (ASC(D2$) AND 240) / 16
400 b4 = ASC(D3$) AND 15
410 b5 = (ASC(D3$) AND 48) / 16
420 ovr = (ASC(D3$) AND 64) / 64 : REM overrange
430 pol = (ASC(D3$) AND 128) / 128: REM polarity
440 vlu = b0 + (bl * 10) + (b2 * 100)
     vlu = vlu + (b3 * 1000) + (b4 * 10000) + (b5 * 100000)
     if pol=0 then vlu = -vlu
442 if ovr=1 then print OVERRANGE: goto 15
445 PRINT USING "#######";value = ", vlu
460 GOTO 15
500 REM SUBROUTINE TO WAIT UNTIL DATA IN receive buffer
510 DO
520 LOOP WHILE ComLoc% = 0
530 Nchar%=ComLoc%
540 RETURN
```

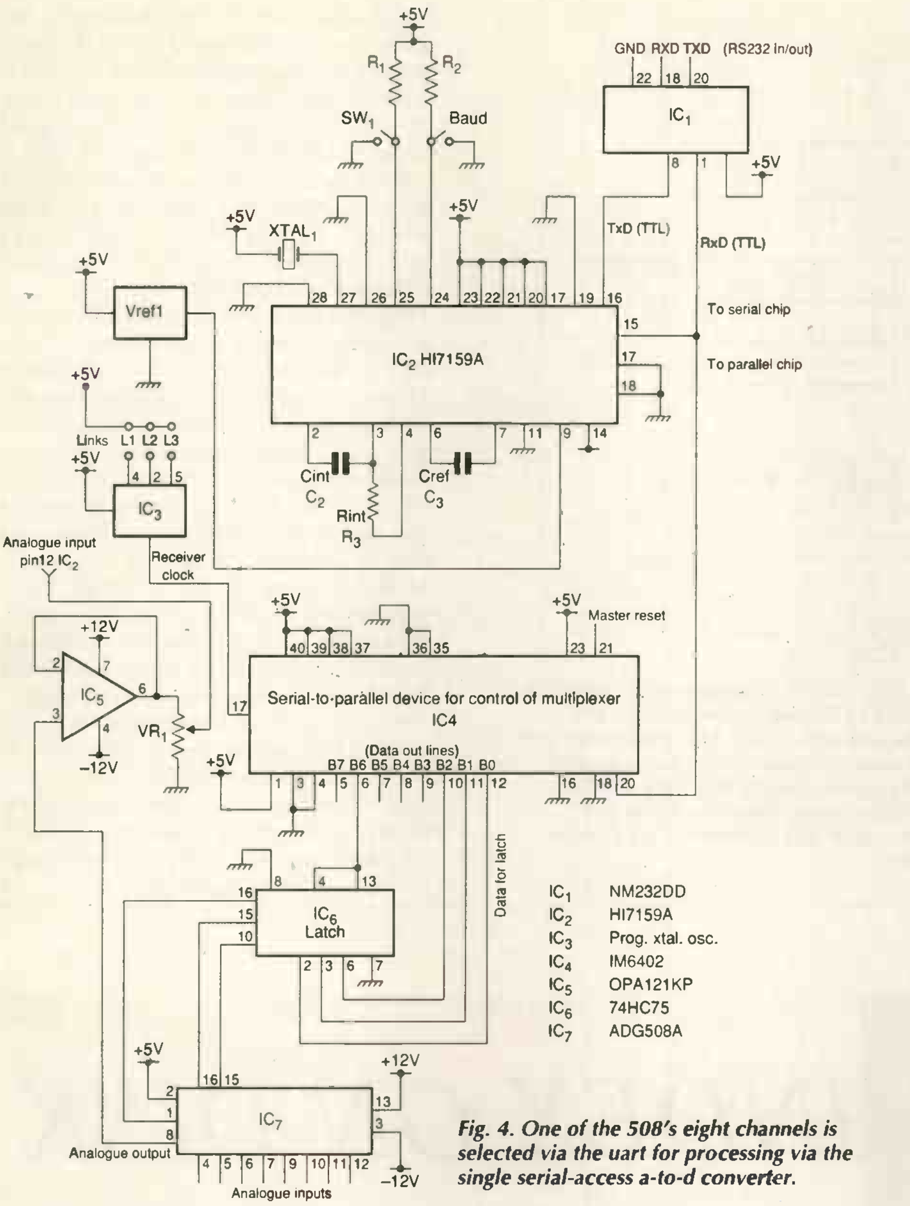


Fig. 4. One of the 508's eight channels is selected via the uart for processing via the single serial-access a-to-d converter.

768kHz. For most systems, 1200 baud is a good compromise.

While it is always cheaper to use specific crystals and divide counters such as the *4020* cmos device, it is usually quicker to buy off-the-shelf programmable oscillators.

Controlling the *IM6402*

Since the *HI7159A* is a slow device, data throughput will not be improved by using low-level language for control. An appropriate high-level language which can set the serial port on a pc to read eight data bits, even parity is required. This specific format cannot be driven by stand alone MicroSoft *QuickBasic*. Routines from the PDQCOMM library have to be linked to make this format an option (available from Grey Matter).

The minor complication of requiring to link routine to another library within the *QuickBasic* environment is more than offset by the ease of programming provided by the language. The interpretive mode of *QuickBasic*, however, is lost. Use of PDQ-COMM, is straightforward. A simple batch file can be constructed to compile and link in a single line instruction.

Where the *HI7159A* shares the same serial input as the *IM6402* there is a potential problem. Sending a byte to the *6402* in order, for example, to control an analogue multiplexer, may cause the *7159A to* latch up.

Where more than one serial port is available, the function of data logging and logic control can be separated. Most pcs which run Windows, however, will require a port for the mouse. Provided that the most significant data bit of command bytes sent to the *IM6402* is set, the *7159* will tend to ignore such input. This effectively leaves seven bits to control an analogue multiplexer and other posible circuitry. Thus any control data sent will be ignored by the *7159*.

There are many devices which can be used for the analogue multiplexing. **Figure 3** indicates the pin out of the *DG508A* and **Table 3** the relevant truth table.

A latch as in **Fig. 4** may be added to free the control line for other functions. Switch S_1 selects the bit rate of the *7159*. Voltage V_{refl} is derived from standard 1.2V reference diode. Master reset on the uart is normally low but can be reset high on power up if required. At the output of op-amp IC_5, potentiometer VR_1 scales the *7159* input voltage to ±2V. Programmable crystal oscillator IC_3 has its bit rate set by links $L_{1\text{-}3}$ as indicated in **Table 4** in order to drive the receive clock at 16 times the bit rate.

There is also the option of implementing variable gain, programmed via data communicated to the *IM6402*. With three digital lines being used for the analogue multiplexer, three lines can be used to drive programmable gain. A programmable gain amplifier suitable for such applications is the *PGA205AP*.

Listing 1 is an example of using the system for data capture. It demonstrates how an analogue channel is selected via a multiplexer, and shows how a command byte is sent to activate the a-to-d converter. The pc then checks for end of conversion and polls the HI7159A for the conversion data, which is held in three separate bytes of data. Subsequently, this data is reconstructed into a signed integer format for processing by the pc.

Multiplexing serial data

It is possible to use the *6402* for multiplexing

Table 4. Pin configurations for 768kHz programmable oscillator for 1200 and 9600baud.

P_1 (L_1)	P_2	P_3 (L_2)	P_4	P_5	P_6 (L_3)	f_{clock}	÷	bit
1	0	0	0	0	1	19200	40	1200
1	0	1	0	0	0	153600	5	9600

Note, 1 represents logic high.

Table 5. Serial-line switching truth table, implemented using the the IM6402 with analogue switches.

Ascii value	B_7	B_6	B_5	B_4	B_3	Mode
128-184	1	0	x	x	x	local HI7159AA mode
192	1	1	0	0	0	serial device 1
200	1	1	0	0	1	serial device 2
208	1	1	0	1	0	serial device 3
216	1	1	0	1	1	serial device 4
224	1	1	1	0	0	serial device 5
232	1	1	1	0	1	serial device 6
240	1	1	1	1	0	serial device 7
248	1	1	1	1	1	serial device 8

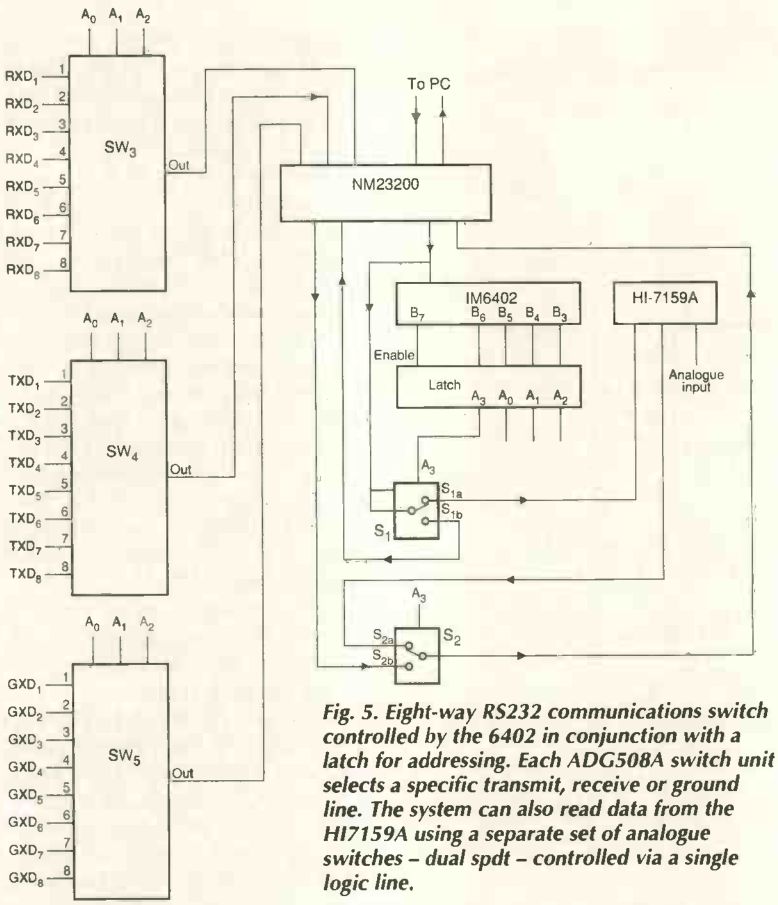


Fig. 5. Eight-way RS232 communications switch controlled by the 6402 in conjunction with a latch for addressing. Each ADG508A switch unit selects a specific transmit, receive or ground line. The system can also read data from the HI7159A using a separate set of analogue switches – dual spdt – controlled via a single logic line.

RS232 serial lines. Where, for example, one serial port is available at a central pc, and a link to a specific peripheral such as a data logger or laboratory intrument is needed, it is possible to select one from several serial devices by use of additional interfacing to the *6402*.

Where serial devices are being connected across only three wires – transmit, receive and ground – a standard analogue switch such as a *DG508A* powered at ±15V will be able to switch RS232 signals between ±12V.

Figure 5 shows how three *DG508A* devices are applied to select specific serial devices. Each of the three is selected via a common address bus, applied via a latch. Additionally, the *7159* can be selected by including switches S_1 and S_2. These may be a dual single-pole, double-throw switch such as the *DG419DJ*.

It is common for serial equipment used for data logging to be relatively intelligent, and able to communicate data in response to 'wake-up' control characters. In this way, a sequence of data capture can consist of selecting a device, sending a serial control character or characters, receiving a data string, closing the link and repeating with another channel.

When high, line B_7 activates the latch. Line B_6 when high enables external serial links 1 to 8 with the addresing of the latch. This is set by $B_{3\text{-}5}$, which control the address lines of the three *DG508A* analogue switches.

Table 5 shows the type of control made possible by this arrangement. ■

LETTERS

GPS vulnerability

Statements made in the letter 'False position on GPS', *EW&WW* March 1995, should be treated with caution. Information presented at the ICAO Communications and Operations Division meeting in March this year, based on work performed by the UK and supported by several states, highlighted the vulnerability of C/A code GPS receivers to interference.

I would also refer your readers test data published in the Institute of Navigation Proceedings, Vol 40 p249, Autumn 1993, and at several ION and Royal Institute of Navigation conferences that confirm such a conclusion.

DRA and NATS have tested a wide range of commercially available C/A code receivers, including Navstar's XRS and equipments certified to the FAA's technical service order TSO-129. The results do not support Mr Leisten's claims of high immunity to interference. In fact all C/A code receivers tested were jammed at a power level between –130dBW and –120dBW at the receiver rf input.

The result have been confirmed during in-flight tests. They indicate a discrepancy of some 60dB between the experimental data and the power levels quoted in your column. It is however very easy to become misled by the mathematics of the spectrum despreading that occurs in a GPS receiver.

Typical receiver noise temperatures of 290K produce a noise power of –203.9dBW/Hz or –137dBW in the C/A code receiver bandwidth. However, as the signal level is some 20dB below the noise level in the predetection bandwidth, it is not until the code has been removed in the correlator-despreader that a positive s/n ratio is achieved.

The theoretical maximum carrier to noise ratio, c/no, with a minimum guaranteed GPS satellite signal power of –160dBW, above a 5° elevation angle, is 43.9dB. In practice the satellites, to every receiver manufacturers delight, are running 4dB hotter than specified with the result that acquisition and tracking appear considerably better than specified.

To ensure the tracking loops remain locked a signal level (c/no) of approximately 30dB/Hz is required in the carrier loop and 22dB/Hz in the code loop. Noise levels that prevent tracking are therefore –160–30, or –190dBW/Hz for carrier and –160–22, or –182dBW/Hz for code.

The correlator C/A code despreading process provides a gain against interference of 63dB/Hz. Noise powers in the receivers predetection bandwidth that significantly degraded the tracking loops measurements are therefore –127dBW/Hz for the carrier and –119dBW/Hz for the code. Generally the carrier loop has a 5 or 6 Hz bandwidth resulting in an interference power at the receiver rf input of –133dBW – A value that agrees well with experimental data.

The problem with the use of GPS for precision approach is that all current methods use some form of carrier phase tracking. To accurately track carrier phase and to be sure that cycle slips do not occur a significantly higher signal to noise ratio in the carrier loop is required.

The US RTCA committee studying the problem are specifying –150dBW as the limit for inband noise. It should be noted that this power is below the thermal noise level. RTCA have set this interference power limit due to the sidebands that exist in the C/A code's correlation function and the resultant degradation of 10 or more decibels in the despreading gain; a 1mW source of inband power could therefore cause interference at 40km.

The problem of the use of pseudolites highlights the interference problem. If power levels and geographic positioning are nor carefully organised they will jam the intended recipient receiver.

The situation is not significantly different for Glonass receivers. Architectures published by some receiver developers would be equally if not more susceptible to interference due to their wide predetection bandwidth.

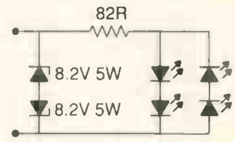
In comparison ILS, and particularly MLS, have significantly higher immunity to interference, although there are concerns over interference into the ILS localiser from fm broadcasts. However at the ICAO (COM/OPS 95) meeting the problem of interference into GPS was identified as a major issue and one that must be solved for the system to achieve the integrity, availability and continuity for use in precision approach and landing operations.

The problem of interference and GPS will be the subject of a workshop held by the Royal Institute of Navigation, 12/13th October in London.

P Nisner, R Farnworth
National Air Traffic Control Services
J I R Owen
DRA Farnborough

On your bike, Steve

Steve Bush, in the July 1995 letters column, writes about the need for better bicycle dynamos. The standard cycle dynamo is not in fact a dynamo at all, but an alternator. To make it operate over a wide speed range, it is not deliberately made lossy (although they are not fabulously efficient), but is a constant current machine. For any given load resistance there is a knee above which increased speed gives little increase in output. The standard set uses a 6V, 100mA bulb at the rear, and a 6V 400mA or 500mA bulb at the front, connected in parallel. The dynamo output is 500mA. This is why failure of the front bulb, or the connection to it, always causes the back bulb to blow, since the entire output of the dynamo then finds its way to the rear bulb.



Efficient red leds would be a better alternative for rear bicycle lamps but using anything other than an incandescent lamp is currently illegal.

Increased output power from the dynamo is entirely possible. If 12V bulbs are used instead, the power delivered is doubled, but this has the disadvantage that the knee voltage is reached at twice the speed. At low speeds the performance is worse, since the bulb filaments are not at full temperature and their resistance is low.

I have had a number or rear bulb

Early transistor response

I read Dr Wylie's letter in the June issue with interest, and totally agree that the history and preservation of early PC transistors is an orphan subject deserving serious attention.

Over the years, I have made an effort to preserve these devices when I find them, so that they now form a small part of my vintage electronic glassware collection.

The most interesting device I have is an early STC PC transistor in the AVO Book (qv) outline 20 form with a paper label 'LS737-crystal triode', which I believe dates from 1950-53 when experimental devices were made under Bell license.

Others include the GET1, Mullard OC50 and infamous unbranded red, yellow and green spot devices in small rectangular aluminium cases. Performance of these devices was, I recall, at best erratic. I suspect that the production spread was such that the colour coding was given in final test. I think the unbranded devices may have been out of specification devices from AEI, who used a similar package although AEI is not listed by AVO as PC type makers.

It's not easy to identify early PC devices, especially if they have failed. I often use the AVO International Transistor Data Manual, which was published early enough to list most of the production PC transistors, together with their manufacturers and clear outlines to help identification. My copy is a third edition, about 1962, (but not dated) for use with the AVO Transistor Tester. Early editions (up to No 6) of the Wireless World Valve Data book also have some PC device listings.

Anthony Hopwood
Upton-on-Severn
Worcestershire

failures in my time, and eventually got so exasperated with this that I put leds in the rear. The enclosed circuit shows what I installed. Unlike many of the pulsing battery powered led rear lamps seen on the roads these days, the light output from this circuit is visibly constant. Four leds are used in an inverse series-parallel arrangement so that two of the leds are illuminated on each half cycle of the dynamo output. Each limb includes one led with a narrow viewing angle, to project plenty of light backwards, and another with a wide viewing angle, to spread light in all directions. In practice this works well. The 8mm ultrabright GaAlAs leds used (RS 577-718 and 577-730) have peak efficiencies at low currents, so pulsing has no advantage in terms of average light output.

The back-to-back 8.2V zeners do not normally conduct, but protect the circuit if the front bulb blows. Two 5W devices are well able to absorb the entire 500mA output from the dynamo, clamping the voltage to about 9V, and the led circuit will tolerate this, though perhaps not for very long periods. Since putting the circuit in I have had a real front-bulb failure, and the rear lamp was undamaged.

In normal use the circuit consumes about 30mA – only a third of the 100mA of a standard bulb, and is every bit as bright. At about £1 each, the leds are not cheap, but were well worth it, in my view. More leds could be used – twelve leds would consume no more power than a standard rear bulb, and produce considerably more light. Led manufacturers are understandably very keen to get device efficiencies up and prices down so they will eventually supersede incandescent bulbs in motor vehicle rear lamps. When this happens, we can expect further improvements.

Unfortunately leds are not yet an option at the front. My own experience is that halogen bulbs are adequate. These are far more efficient than ordinary bulbs, and produce enough light to make road markings and signs visible in dark country lanes. As a means of being seen by other road users, I think they suffice.

In fairness, I should point out that leds are not legal in rear lights, because they are not incandescent. Silly, but true, especially since they are superior and more reliable. Prosecution for the offence of having leds at the back is not, however to be expected. I conclude that this is because in York – one of the country's foremost cycle cities – only about half of the cycles about at night have lights at all, and the police never seem too bothered.

Alan Robinson
York

Thoughts in tandem

I believe that Steve Bush's need for a small efficient dynamo may be satisfied by utilising a small shunt wound electric motor and adapting it to generate current by constructing an electronic regulator to feed the field coil. By experimenting with the device to understand the characteristics of the motor, a fairly efficient dynamo may be built. There should be sufficient residual magnetism in the iron of the device to cause it to self excite.

It may be necessary to polarise the device by running it as a motor from a power supply with the field in parallel with the armature to ensure that it charges with the correct polarity.

All of the car type alternators are constructed as three-phase machines with slip-ring feeds to the rotor for the exciting current. This type of machine would be a good model to copy if one had the machining facilities. Details may be found in most of the Haynes car maintenance manuals, the one for the Land Rover series II I remember as being particularly good on alternators.

It may be possible to find a small device from a piece of war surplus equipment, such as the Bendix Bomb Sight computer that was available from war surplus stores, (showing my age eh?). Devices such as a magsyns or selsyns could probably adapted to work as Steve requires.

There is one other alternative, details of which he may find in a copy of Haynes Motorcycle Maintenance manual for the Honda 400/4, *circa* 1974 to 78, which had a form of variable-reluctance alternator with electronic control. From this he will perhaps be able to design smaller vr machine.

There is another idea, which may prove too wasteful in input energy terms. There are many small permanent-magnet motors available these days. They are often used for electronic seat movements in cars. Such a motor would give a fairly large output, but regulation would have to be a crude zener diode. It may be possible however to arrange a mechanical magnetic shunt to prevent too much energy being absorbed from the bicycle wheel.

My final suggestion is to adopt and adapt a scheme that was prevalent in my 'teens, from a company called Miller, if my memory serves me well. This had a small container strapped to the down tube. It held some dry batteries which took over the lighting when the bike was stationary.

By adapting the scheme to incorporate a set of nickel-cadmium cells, or circular lead-acid cells, and rectifying the existing bike dynamo output Steve could charge his battery at all times while the cycle was moving, have good lighting, and, in case he didn't get enough daylight cycling during the winter, he could unplug the battery pack and trickle charge it. Cell size could be calculated for optimum depending on journey length, charge rate, etc.

Nic Houslip
Dallas Semiconductor Corp Ltd
Birmingham

Power-line reaction

In the May 1995 issue, a letter from Roger Coghill claimed that a review published by the IEE, in June 1994, on the possible biological effects of low-frequency electric fields was

My heart skipped a beat

I read with interest and concern the article entitled Monitoring Heartbeat in the July 1995 issue.

I note your comment concerning the mains adaptor having to comply with medical safety requirements However, I am concerned that there is no apparent isolation from ground at the patient probe inputs.

Under certain single-fault conditions, circulating (low-frequency, typically 50Hz) currents can be present, causing possible fibrillation risk or undesirable muscle stimulation to occur.

There are specific medical equipment safety standards that mandate the maximum permissible leakage currents allowed to flow between the probes. These standards are under the group for Electromedical Equipment and are designated IEC 601-1 for general safety with a further Part 2 standard specifically applying to ECG related equipment.

Michael Brett
Watford
Hertfordshire

Finger on the pulse

It may be worth pointing out that the instrumentation amplifier in Baki Koyunku's heart rate monitor (*EW+WW*, July 95, pp605-7) is probably redundant in his configuration.

Firstly, an 'instrumentation' configuration can provide a very high input impedance. However, the 1MΩ input resistors prevent this feature from being used.

Secondly, the configuration can provide a very high common-mode rejection ratio. However, an elementary analysis will confirm that this is only possible if the first stage has a high gain and the second stage has a very low common-mode rejection ratio. This is because the common-mode gain of the first stage is unity so the circuit relies on a high differential-mode gain, and good rejection of the common-mode component from the first stage.

Koyunku's circuit features a first stage differential gain of only three, which is not good enough. His second stage is bizarre, featuring a different gain and frequency response from the positive and negative inputs. There is absolutely no chance of achieving anything approaching common-mode rejection – especially as his gain control resistor (I think this what it is) affects the gain paths differently.

Readers implementing this circuit will do much better to transfer the gain to the first stage, where it can still be adjusted by varying the single 'central' resistor. Unfortunately, no component references were given in the circuit, so my description necessarily vague.

The second stage should be a unity gain differential buffer with 0.1% resistors, and with capacitors across *both* the feedback resistor and the resistor from the positive op-amp input to ground.

Ideally, the input potential dividers should be altered too. As it is they serve to introduce a common-mode component, or, if you like, they cause common mode interference to generate a differential mode component. Don't forget to allow provision for the bias currents.

These modifications will turn the circuit into an instrumentation-quality amplifier which it quite clearly is not at the moment. A better option may be to use one of the cheap instrumentation amplifier ICs now on the market.

David Gibson
Microsystem Solutions
Leeds

'propagandist material masquerading as science'. He also claimed that the composition of our working committee was heavily weighted with power-utility-related personnel and that we refused to disclose the references on which the report was based.

The IEE has been aware of the public's concern about the possible harmful effects of electromagnetic fields for a number of years. In November 1992 we set up a Working Party, comprising both IEE members and others with relevant expertise in the field to monitor relevant scientific literature. The Working Party has a balanced membership, with only one member from the power-utility-related industry and includes an eminent epidemiologist involved in leukemia research. The literature was retrieved by searches of three major databases and an abstract of each paper was obtained.

When Mr Coghill approached the Institution, for a list of references, he was informed that, because of the contractual arrangements under which they were obtained, we could not supply him with photocopies of the abstracts. Since he had informed us he required the references to carry out his own research, Mr Coghill was provided with details of the databases interrogated, by the IEE, and the search criteria used. We also offered to conduct computer searches, on a commercial basis, for him from the same databases, using the identical search criteria and over the same time period.

These searches would have provided him with identical information that the IEE Working Party considered. The IEE Working Party has continued to monitor relevant scientific literature and there has been no reason to change the conclusions reached in June 1994 that the studies show no firm evidence of biological effects of low-level, low-frequency electromagnetic fields.

J. C. Williams, Chief Executive
The Institution of Electrical Engineers
London

Anti-aliasing filter rework

In my article on antialiasing filters in the June issue, an error escaped me in the preparation of the frequency response graphs: the curves labelled A_n and D_n were named A_x and D_x in the captions, the index 'x' indicating their relation to the mixed-mode response M_x.

I would also like to thank Mr. Self for his comments on my letter ('Lend a golden ear', May '95). I must add that by those examples I was not offering proof, but rather asking questions. Yes, it is difficult to accept that some people have extraordinary abilities of perception. But in the case of my colleague, those resistor values were the only difference I could find, using a simple $3^1/_2$-digit dmm, an HP oscillator and a Tektronix *475* 'scope).

Of course, it is possible that I was fooled. But I also remember reading about a woman being held in a psychiatric clinic for 14 years because she was 'hearing voices', and then, before starting a therapy with new medication, she was sent to a routine audiometry check where it was found out that her low-frequency hearing threshold was about 4Hz. Hopefully, claiming of being able to hear something that others don't will not put anyone under medical surveillance, but the present trend worries me.

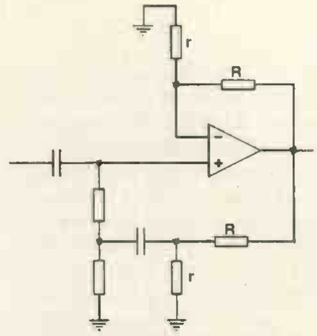
I agree that too many vendors take advantage of the situation. But how come we engineers struggle for every milliohm in the amplifier output impedance? Is it only for stability reasons?

Regarding the communication failure, even we engineers sometimes can not hear ourselves shouting. I believe I have shown (in *EW+WW*, July 1987) one possible way in which 'the unmeasurable avoids to be inaudible'.

For a moment let's put aside the actual phase-switching audibility threshold – which in my opinion is much lower than the crossover spikes audibility threshold. The trouble is that when we use a constant amplitude signal source and take the amplifier input-output signal difference, we intentionally compensate the amplifier phase shift to single out crossover spikes.

When the signal amplitude is lowered, the crossover spikes fade in the noise. The remaining phase error is interpreted as some residual distortion, more easily so if there are other distortion mechanisms in the circuit. However, if the phase is not correctly compensated, a relatively large phase error will appear along with the crossover spikes when the signal level is raised.

I have made this error myself on countless occasions. Interestingly, there was nobody to comment on this. Instead, there are only negative comments on the amplifier model used, despite my clear statement that the model was simply a simulation of a typical popular amplifier.



Possible solution for isolating Self's amplifier's input bootstrapping from its feedback.

It remains to be seen if the described switching phase modulation mechanism is really the cause for the majority of 'scarce definition' marks in subjective evaluations. But engineers keep talking about distortion, so in the mind of inexperienced amplifier buyers, if there is anything audible, it must presumably be 'distortion'.

I have yet to meet a 'subjectivist' who sticks to a Heisenberg acoustic uncertainty principle, which states that even the presence of a single listener modifies the sound field – obviously true, but equally obviously nonsense.

Finally, I would like to offer Mr. Self a possible solution for isolating the input bootstrapping from the feedback in his recent design, in the form of the diagram shown.

Erik Margan
Ljubljana
Slovenia

Mosfets, BJTs and power amps

I have been following Doug Self's articles on power amplifier design and the subsequent flurry of correspondence discussing power amplifier design philosophy. As a commercial designer, I am surprised that so much of the debate appears to revolve around the inherent linearity – or non-linearity – of bipolar junction transistors versus fets. More significant, when considering the design options, is relative price; with a mosfet output device costing pounds rather the pence a bjt commands – even in relatively large quantities.

Further, mosfets have the disadvantage of a finite, and relatively large, on resistance which wastes power and necessitates paralleled devices in order to drive low impedance speakers. These considerations affect 'watts per dollar' calculations considerably and probably account for the relative dearth of commercial mosfet amplifiers more than considerations of inherent device linearity.

But it is not wholly for commercial reasons that the fet is under attack; I do not subscribe to the current swing in intellectual fashion against mosfet amplifiers simply because of their inherent poor crossover distortion characteristics within a Class-B circuit. The mosfet is still worthy of consideration from any engineer designing an audio power amplifier because they possess a number of highly attractive attributes.

I can wholeheartedly confirm Mr Self's empirical observations that the open-loop, transfer-characteristic of the crossover region in a mosfet amplifier is less 'tidy' than the corresponding region in a bjt amplifier when the latter is adjusted for optimum bias. This indicates that fets must be used in an amplifier designed to have adequate, and controlled, open-loop gain to accommodate a relatively high feedback factor. Fashion – and that's all it is – dictates that this is prima facie 'a bad thing'. But the same is true of a bjt amplifier because, while it is true that it is possible to trim a junction transistor output-stage to possess a linear crossover region, real-life operating conditions – and the thermal time-lag in temperature compensation – mean that the bias often departs from the optimal value for the output stage's given operating conditions. Add to this the mosfet characteristics of constant gain-bandwidth product and inherent protection against thermal runaway and secondary breakdown and it is still possible to justify the fet's place in power amplifier design. One exception is perhaps where cost is an overwhelming factor.

Richard Brice
Electric Perception Ltd

Could germanium perform?

I have been following Douglas Self's excellent series on power amplifier design, and the following correspondence with great interest.

One point intrigues me: Mr. Self has made several comparisons between various output topology and devices, bipolar and mosfet – I was wondering how well germanium performs in the same tests. By comparing the I_c/V_{be} curves of silicon and germanium, intuition suggests that the same dips in the output gain curve will be present, but that the slope of the dip will be less steep, suggesting that a smaller bandwidth of gain/feedback stage would be necessary to achieve the same thd performance.

Or I may be entirely wrong.

Ian Benton
Ilkeston
Derbyshire

Final hearing...

In his letter ("Slewing from reality", *Letters*, pp.500-501) Douglas Self denies that music can involve transient hf levels as high as the loudest bass passages, But in this he is himself slewing dangerously outside his field of competence.

He appears to be misusing data on music's spectral composition. This is based on rms, or long-term L_{aeq} measurements and is free from information on slew demands.

Conversely, my comments stem from colleagues who are world class producers of music. I doubt Self has any primary experience of quantifying hf levels of a rock band. The experiment with a supermarket-grade keyboard cuts no ice.

Even the inferred response out to 16kHz enables full level treble seven times higher than the 2.2kHz above which he claims "large" signals do not occur in music.

In fact synthesisers can have far higher sample rates than 44kHz, and analogue synthesisers (which may be digital in their controls) can also have edges extending way above audio. In both instance, increasing use of on-stage mic splitters enables hf edges to be preserved.

I should like to address, in turn, the errors Self alleges in my translation of his circuit.

1. The vas "flush-out" resistor (1kΩ) is not altogether missing. In MicroCap, there is a minimum conductance setting across active device terminals, used to aid convergence. So "re-flush" is always there, just hidden inside the transistor, though admittedly too high in value.

Yet if Self had tried simulation, he'd have seen that stepping the flush-out between 1MΩ and 1kΩ has no effect on the negative slewing, and only a small tidying-up effect on the positive pedestal.

2. Self is adamant that fast signals can never assault an amplifier – here he claims my 100V/µs test signal is too slow! But faster attack has little effect on the slew limit of his topology, and as excess test signal speed also increases the time for a given simulation accuracy, it wasn't necessary.

3. The cvas cs base resistor is different *precisely* because some value stepping was tried, and it had scant effect.

4. Self mentions that all the transistors have been changed, skillfully drawing attention away from the fact that, unlike his own simulations, the transistor model data I have used is shown in its entirety.

Moreover, I ensured that the Self-defined slew-critical bjt parameters are amongst the highest available, So either the poor negative slew performance is true (and his simulation accuracy is questionable); or the circuit schematic is extraordinarily sensitive to transistor parameters.

His comment on portal filtering has not been thought through either. All decent audio sources have source impedances below 1000Ω unless they are the passive preamps or valve preamps used solely by the audiophiles. In such cases, all that will happen with correctly scaled *RC* filtering is that hf will be rolled off prematurely. This will not degrade %thd – apparently the only parameter that matters.

Turning to harmonics, I have made it clear how real sonic issues may be addressed without added parts or presets, but solely through careful design. The resulting values have been very different to those Self believes are perfect. Where is the scientific evidence that supports his assertion that vanishingly low %thd (anything below 0.01%) has any predictable beneficial effect on sonic quality?

Self's low distortion figures may deserve an *A plus* for analogue electronics, but real audio professionals award them *D minus*. The "much increased dc offset" is a non-problem if considered earlier in the design process.

On another topic, I note Self admits that mosfets can have low distortion. In fact he is highlighting the nonsense of trying to compare chalk with cheese. After 17 years of waging war on experienced operators in professional audio on these pages, it is time for Douglas to lay down his weapons.

As for the other correspondents, I thank them for their input. But Mr Davis cannot have carefully read my words. I certainly did not say that capacitor mics were not used 20 years ago. Rather, it takes 'classical' music combined with dynamic microphones and other less hf-responsive technology to explain why Baxandall could go unchallenged for so long.

I take Davis' word on the theoretical limits of certain classic cutting equipment. But the fact remains that ultrasonic signals have been measured at the replay end.

I agree with Mr van den Gevel that vhf %thd sweeps may be used to show slew limiting. On the other hand, steep rises that look like slewing can also be caused by several other mechanisms, and amplifiers can burn out when tested in this way.

The *DIM30/100* test is rather more sensitive than thd, though such measurements will not reveal the impulse time domain information shown in my slew simulations.

Van den Gevel has also ignored my other points, about the reality of rf contamination for example.

Assuming an rf free-environment, Self's asymmetric 40-50V/µs is fine for his putative 20W amplifier. But if you need 1020W (not so outrageous when rescaled in V, and account taken of fidelity-aiding stuff such as 6dB of head-room and listener space greater than shoe-box-sized) 50Vµs may prove most painful and ultimately ear-damaging.

Ben Duncan
Tattershall
Lincoln

Chain letter

Following the Self-Duncan debate – and I must admit I declare on Self's side – how refreshing to read Jeff Macaulay's comments about "the carefully cultivated signal...being mangled..." and "What emerges...is a distorted version of the driving signal, no matter how perfect the input may be."

How about some articles on the whole audio chain to put amplifier performance into its proper perspective? ■

Mark Plews
Herrsching
Germany

Harmony in the *band*

Svetlana Josifovska has been looking at the government's recent spectrum review intended to prepare the way for the introduction of Europe-wide services.

Land mobile radio and other radiocommunications services, and advanced technologies such as T-DAB – terrestrial digital audio broadcasting – are increasing at an incredible rate. One of the effects of these advances is increased pressure for more radio spectrum.

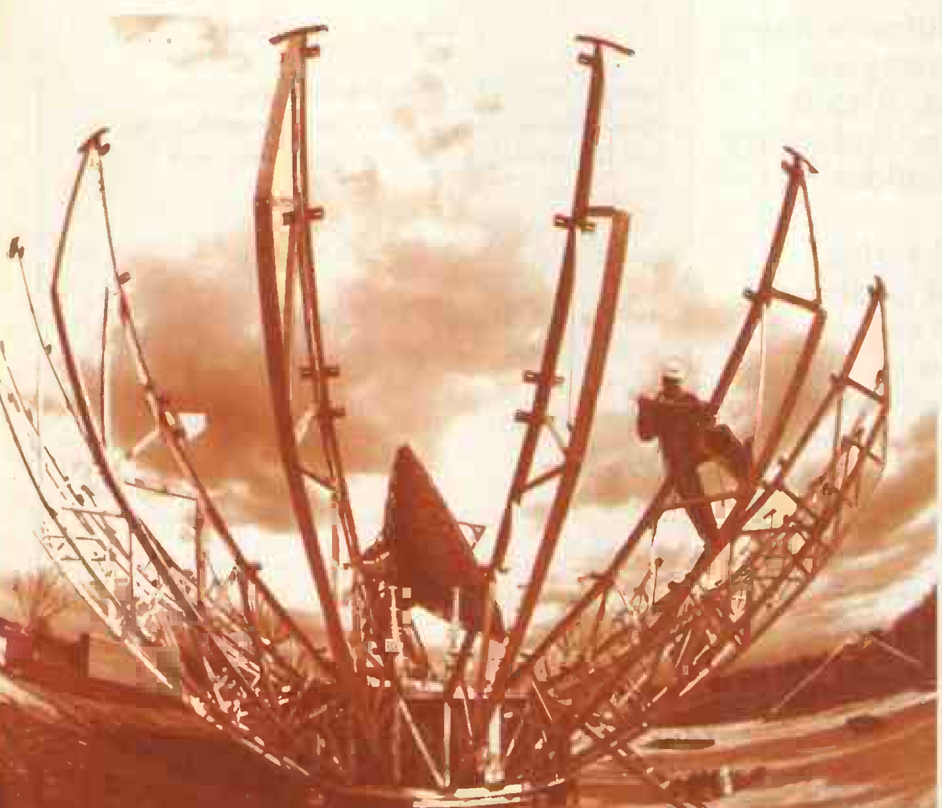
Since the radio spectrum is a limited resource, a logical progression is to reorganise existing services and reallocate frequencies. But, although this may appear simple at first glance, it is actually a delicate game of strategies and tactics. The parties involved – a mixture of government, military and civil authorities – always request more, rather than less, of the radio spectrum.

Stage 3 Radio Spectrum Review committee was set up in July 1992 by the Board of Trade President Michael Heseltine. Last month it came up with a a synopsis that covers arguably one of the most crowded, and hence contentionus, parts of the radio spectrum – the 28-470MHz range.

"The Review Committee made 28 recommendations, 24 of which have been accepted by the Government without reservation. The remaining four are accepted in principle. The recommendations cover both defence and civil use of the Review spectrum and a number of them have already been acted upon," said Heseltine.

Within this frequency range there are allocations that belong to the broadcasting industry and the mobile industry including land, aeronautical, maritime and satellite. These are also space research and operations in the range, that still belong mainly to the MoD. Users of this spectrum range from broadcasters and the government to utilities and emergency-services operators and the civil mobile radio community.

The Stage 3 review of the radio spectrum in the 28-470MHz range completes the government sponsored independent reviews of the radio spectrum over the range 28MHz to 30GHz. The ranges below 28MHz and above 30GHz are subject to internal audits and the results of these are expected to be published later in the year.



More releases between 225 and 400MHz

Amongst the agreed requirements was the role of the MoD in this radio spectrum and its release of more frequencies for civil use. Although the MoD has already released the 225MHz to 230MHz band for T-DAB, it was recommended that it should seek, through NATO, the release of some of the 225MHz to 400MHz frequency band for civil systems in the UK, particularly the 380-399.9MHz band in its entirety. Access to two 5MHz sections in the 380MHz to 400MHz band by 1997 has already been agreed by NATO/ARFA for use by the emergency services and the rest of the spectrum's release is still under review.

"Part of an existing defence frequency band has already been released for a new digital radio broadcasting service for the UK. Further defence spectrum has been released in cooperation with NATO to provide the emergency services with an opportunity to develop a new radio system. This in turn will also assist with the resolution of the long standing continental radio interference problems in the UHF frequency bands in England and Wales," said Heseltine.

These frequencies are of significant importance because it is in line with the CEPT agreement for emergency services to be harmonized across Europe. According to government estimates the emergency services will begin to take up use of this band in 1998 and complete the changeover by the year 2003.

The committee is urging the MoD to adopt a leading role within NATO, through the process of specifying new equipment and through more economical spectrum management, towards a reduction in the total bandwidth required.

The government is also paying more attention to the introduction of T-DAB, by arranging an EU meeting in July to discuss the subject. It encourages broadcasters wishing to use T-DAB to start a service soon. All vhf sound broadcasting should be contained within Band II (88MHz to 108MHz) as soon as possible should the introduction of such service prove feasible.

Improvements to mobile services

On the subject of land mobile services, the government has agreed to a programme of improvements to ensure better management of these services' radio spectrum, which will ultimately benefit users.

The programme would consist of improved methods of frequency assignment, efficient procedures for systematic analysis of licensing data and monitoring results, and efficient spread of frequency spectrum between exclusive Private Mobile Radio (PMR) and shared PMR. The Radiocommunications Agency (RA) has been recommended to oversee the implementation of these improvements and coordinate some areas of planning with the mobile industry. Inclusive of PMR, the other two areas are Common Base Station (CBS) and Public Access Mobile Radio (PAMR) services.

The RA will also take care of the policies and strategic plans for the future use and allocation of the spectrum to meet the needs of the private and public sectors. It will make sure that operators use the given radio spectrum to the best national advantage. This responsibility the RA may share with the MoD, an option which is currently under review.

The RA will also look into deregulating low-power PMR services for indoor use, for example in the area of short-range business radio.

Other aspects of the spectrum that have been included in the review are the arrival of new technologies and the allocation of frequencies to them. One such example is the emerging Digital Short Range Radio (DSRR) and 5KHz narrow-band land mobile radio.

But amongst the important points that have been covered in the committee's proposal and the government's response, is the role of the UK in Europe and the economical benefits it may gain.

"Today's response and the progress which will be made in implementing the recommendations is a further example of the steps we are taking to improve the efficiency of spectrum management and usage in the UK to promote competition and choice for the benefit of all users of the radio spectrum," said Heseltine.

Amongst the examples for a UK lead within the European community is its lead in the talks with NATO, to release some of the frequency ranges for civil use, and its supporting role in harmonizing services with the rest of Europe. These include the emergency services synchronization in the uhf 380-400MHz band, that was recommended by CEPT and which is expected by 1997, and the vhf band for public-sector spectrum requirements for PMR.

Harmonization with Europe of the spectrum use and equipment characteristics is seen as being essential for the UK. The government takes into account that harmonisation does not necessarily always yield efficient use of the radio spectrum, but it will ensure the limiting of allocations where it is essential for Europe-wide capabilities or for border coordination.

These issues have been addressed before in the consultative document, The Future Management of the Radio Spectrum.

"Some of the Review Committee's recommendations addressed spectrum management issues which were also the subject of the Consultative Document on the Future Management of the Radio Spectrum. Therefore it was decided to delay the completion of the Government's response to the Stage 3 Review until after consideration of the submissions to the consultative document," said Heseltine. ■

EXPERT WITNESS

Audio expert John Linsley-Hood rarely enters such debates, but the claims in favour of the bjt over mosfets have challenged his decades of experience.

On several occasions – most recently in the May 1995 issue – Douglas Self has claimed that bipolar junction transistors, bjts, are more linear than power mosfets as audio amplifier output stage devices.

I am reluctant to accept this assertion, since if it were true it would overturn the whole of my experience – in respect of the comparative performance of these devices in audio power amplifiers – over the 20-odd years since the commercial introduction of power mosfets. I feel that the progress of audio design requires that this claim be contested.

There are of course, many other reasons – better hf performance, greater stability, greater robustness and inherent freedom from thermal run-away – which have encouraged a number of manufacturers of audio amplifiers to use mosfets rather than bjts as the output stage devices in their top range designs. But the question which Self raises is that of their intrinsic linearity as an output push-pull pair. This is a quality which I feel can be related – by means of measurements – to the residual harmonic distortion characteristics of audio amplifiers incorporating one or other of these alternative devices in its output stage.

Some years ago, I was asked by a well

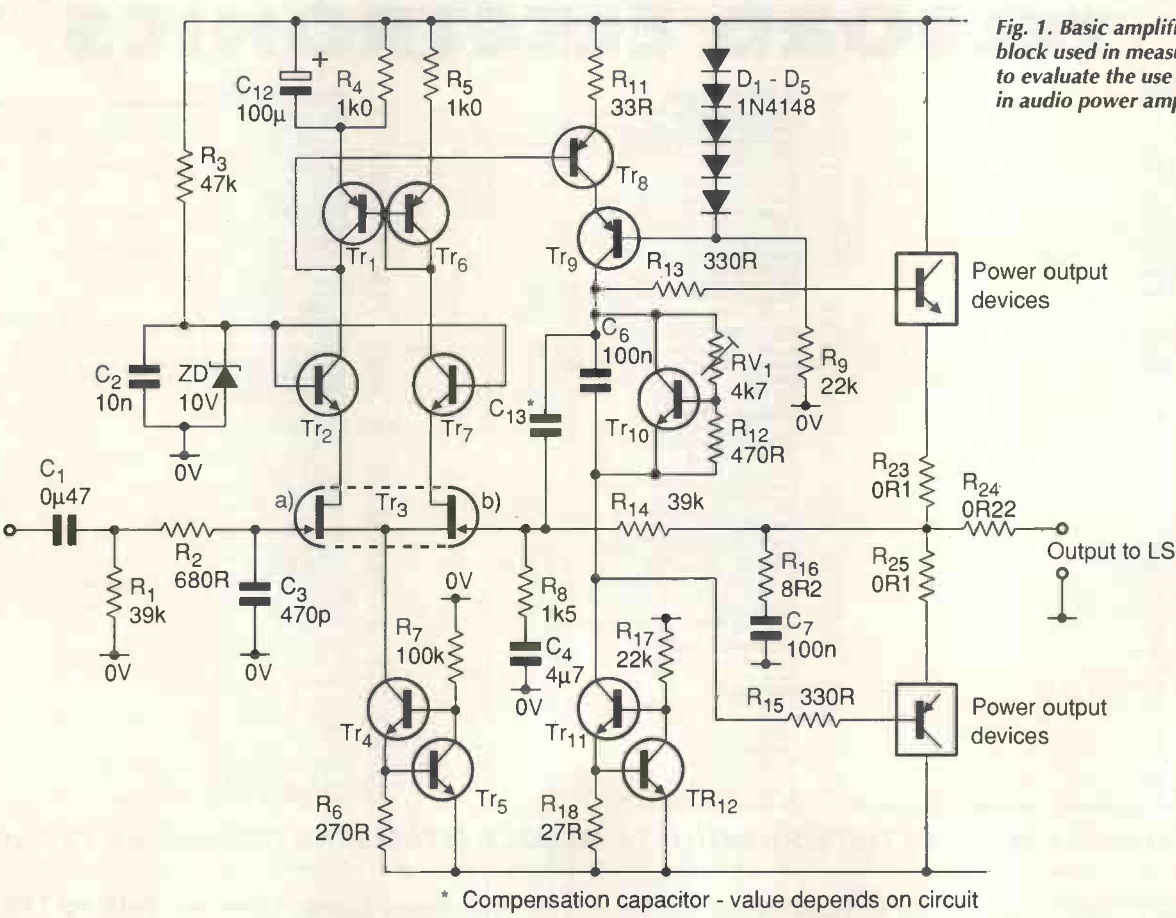


Fig. 1. Basic amplifier gain block used in measurements to evaluate the use of igbts in audio power amplifiers.

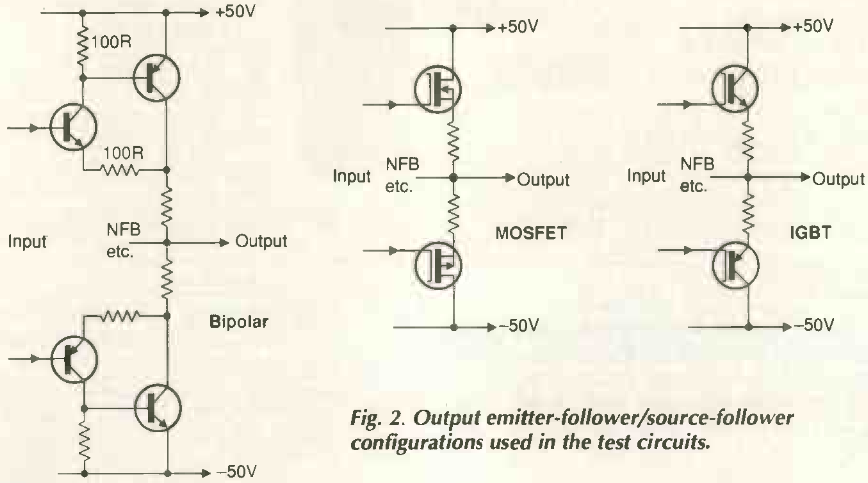


Fig. 2. Output emitter-follower/source-follower configurations used in the test circuits.

known semiconductor manufacturer to explore the use of their recently introduced insulated gate bipolar transistors, igbts, as output-stage devices for audio amplifiers. It appeared that igbts should combine the practical advantages of both bjts and mosfets. To this end, I designed and built a group of audio amplifiers. These used the basic voltage amplifier layout of **Fig. 1** as a gain block. The various output devices were inserted into these gain blocks, in the positions shown as square boxes, using the various emitter or source-follower configurations of **Fig. 2.**

Performance of these amplifiers, operated from a pair of ±50V stabilised dc power supplies, is shown in **Fig. 3**. Measurements were made at a frequency of 1kHz with an 8Ω water-cooled load at power output levels up to the onset of clipping.

In all cases quiescent current of the output-stage devices was adjusted on test to give the lowest harmonic distortion figure for the amplifier. It was noted that this adjustment was less critical with the mosfets than with either the bjt or the igbt devices.

Subsequently, we found that the test oscillator used in these measurements had a residual harmonic distortion at the 1kHz test frequency, of 0.003%. This was mainly third harmonic, so, in reality, all of these measurements had tended somewhat to exaggerate the true residual distortion figure of all of the amplifiers. Nevertheless, the superiority of the mosfet based design is evident in this comparison in which the only variable was the choice of output stage devices.

The reasons for this difference in linearity between the alternative output stage devices are complex. I suspect that two main factors make the negative feedback loop in mosfet amplifiers more effective than in bipolar equivalents. One is the nature of the frequency distribution of the distortion components due to the discontinuity of the transfer characteristics of the output push-pull output pair. The other is the relatively lower internal phase shifts of mosfets.

Whatever the reasons for performance differences, it is clear that they are not adequately revealed by a Spice type computer simulation of the output-pair transfer characteristics. As such, this type of measurement fails to meet Self's own criterion that 'if an apparent fact is repeated many times without number... it deserves to be looked at very carefully indeed'. ■

Further reading

For those of you interested in fuller details of the circuitry used in these experiments, see Toshiba Application Note *X3504*, March 1991.

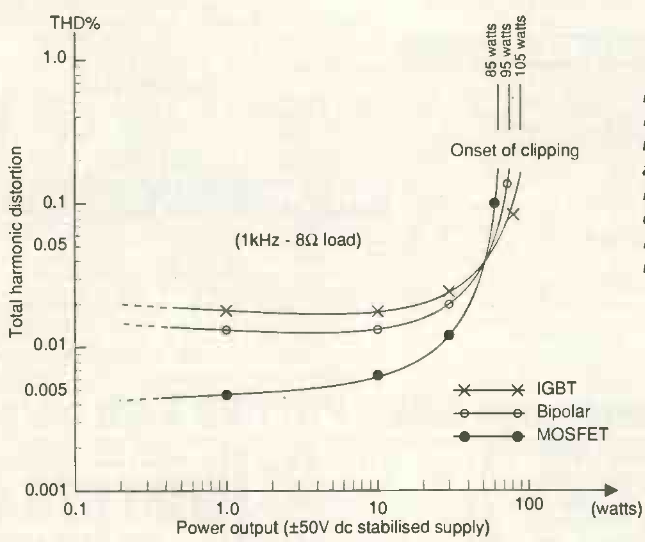


Fig. 3. Relationship between power and total harmonic distortion reveals interesting results.

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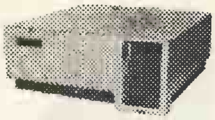

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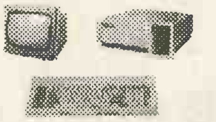

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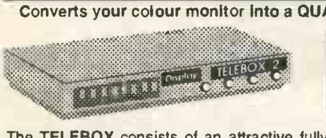

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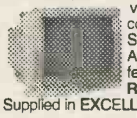

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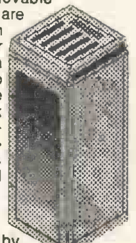

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Temperature to 9bit serial words from one chip.

Temperature bytes

Reading temperature into a digital system means a thermal sensor, signal conditioning and a-to-d conversion. There is now an eight-pin chip that incorporates all this – together with parallel-to-serial conversion and three thermal alarm outputs.

Even given an IC that produces a voltage that directly indicates temperature, you would need to add at least an a-to-d converter in order to interface the temperature signal to a digital processing system. Simplifying the process of reading temperature into digital systems, Dallas Semiconductor has devised an eight-pin chip that not only outputs a 9bit serial stream representing temperature, but also provides three temperature threshold switches. These switches are user programmable via non-volatile ram.

The device measures temperatures from –55°C to 125°C in 0.5°C increments, which is the equivalent of –67°F to 257°F in 0.9°F increments. Conversion time, from a temperature reading to output of a digital word is 1s and data is read from/written via a three-wire serial interface comprising clock, data i/o and reset lines.

Although highly integrated, the *DS1620* is designed for ease of use and requires no external components.

Converting temperature to bytes

Temperature readings from the *DS1620* represent the temperature of the actual device so no external temperature sensor is needed. Having three thermal alarm outputs, the chip can also act as a thermostat.

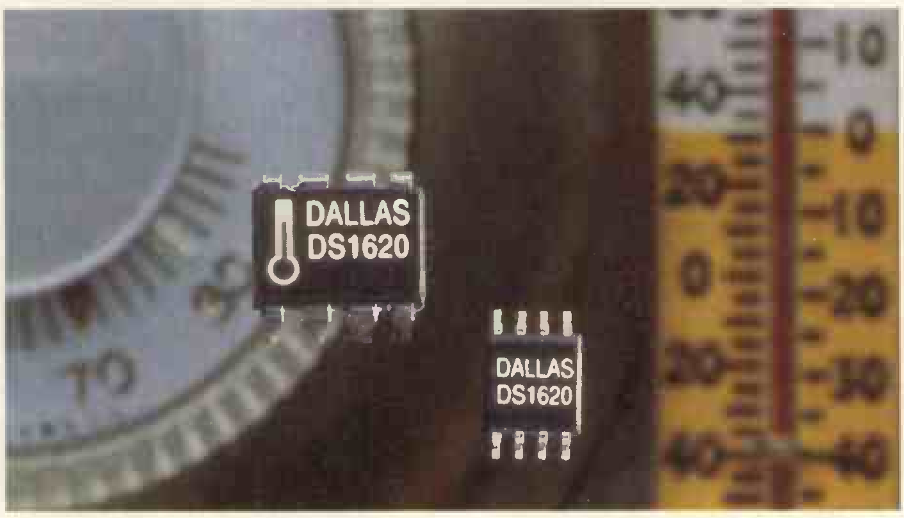


Table 1. Relationship between temperature and the DS1620 2's complement output stream.

Temp	$Output_2$	$Output_{16}$
+125°C	0 11111010	00FA
+25°C	0 00110010	0032
1/2°C	0 00000001	0001
0°C	0 00000000	0000
–1/2°C	1 11111111	01FF
–25°C	1 11001110	01CE
–55°C	1 10010010	0192

Temperature T_{high} is driven driven high if the device's temperature is greater than or equal to a user-defined temperature TH. Similarly, T_{low} is driven high if the device temperature is less than or equal to user-defined temperature TL. Pin T_{com} is driven high when the temperature exceeds TH and stays high until the temperature falls below that of TL.

User-defined temperature settings are stored in nonvolatile memory, so parts can be programmed prior to insertion in a system, as well as used in stand alone applications, i.e. without a cpu. Temperature settings and temperature readings are all communicated to/from the *DS1620* over a simple three-wire interface.

Reading temperature

The *DS1620* measures temperatures using an on board proprietary temperature measurement technique. Temperature readings are provided in a 9bit, two's complement format. **Table 1** describes the exact relationship of output data to measured temperature.

Data is transmitted serially through the 3-wire serial interface, lsb first. The chip can measure temperature over the range of –55°C to +125° in 0.5°C increments. For conversion to fahrenheit, a lookup table or conversion factor is needed.

Since data is transmitted over the 3-wire bus lsb first, temperature data can be written to/read from the device as either a 9bit word, taking /RST low after the ninth bit (msb), or as two transfers of 8bit words, with the most significant seven bits being ignored or set to zero, as illustrated in Table 1. After the msb, the device outputs zeros.

Note that temperature is represented in terms of a $^1/_2$°C, the least-significant bit yielding the following 9-bit format:

msb	lsb
x x x x x x x 1	1 1 0 0 1 1 1 0

Applying the thermostat

Three thermally triggered outputs, T_{high} T_{low} and T_{com} are provided to allow the device to be used as a thermostat, **Fig. 1**. When the temperature meets or exceeds the value stored in the high temperature trip register, output T_{high} goes active high and remains active until the measured temperature becomes less than the stored value in the high-temperature register, TH.

The T_{high} output can be used to indicate that a high-temperature tolerance boundary has been met or exceeded. Alternatively, as part of a closed loop system, it can be used to activate a cooling system and to deactivate it when the system temperature returns to tolerance.

Output T_{low} functions similarly to the T_{high} output. When the *DS1620*'s measured temperature equals or falls below the value stored in the low-temperature register, output T_{low} becomes active, **Fig.** 2. It remains active until the temperature becomes greater than the value stored in the low-temperature register, TL. The T_{low} output can be used to indicate that a low temperature tolerance boundary has been met or exceeded, or as part of a closed-loop system, can be used to activate a heating system and deactivate it when the system temperature returns to tolerance.

Output T_{com} goes high when the measured temperature meets or exceeds TH, and stays high until the temperature equals or falls below TL. In this way, any amount of hysteresis can be obtained.

Operation and control

The *DS1620* must have temperature settings resident in the TH and TL registers for thermostatic operation. A configuration/status reg-

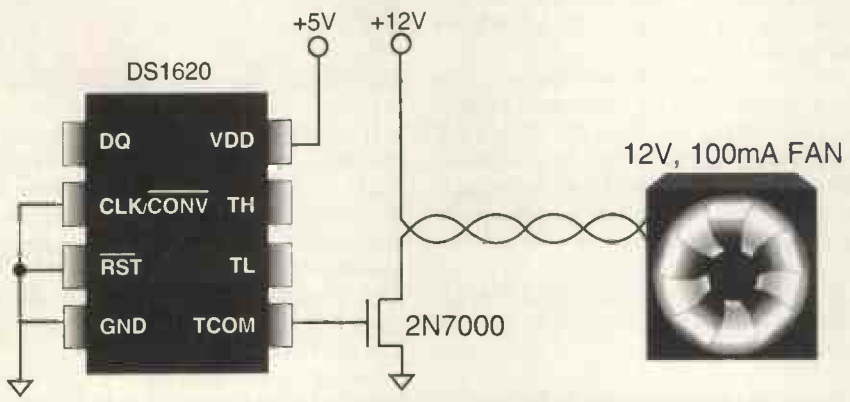


Fig. 1. DS1620 acting as a digitally programmable thermostat. Programming temperature thresholds is a one-off operation since the device incorporates non-volatile memory. Either a 2N7000 or ZVN2106A – both available from Zetex – is suitable for the fan-driving mosfet.

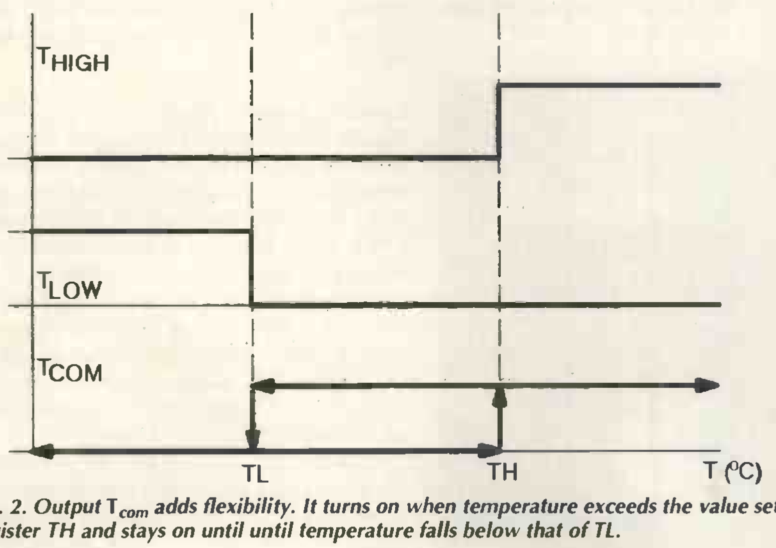


Fig. 2. Output T_{com} *adds flexibility. It turns on when temperature exceeds the value set in register TH and stays on until until temperature falls below that of TL.*

DS1620 command set

Read temperature [AA_{16}] This command reads the contents of the register containing the last temperature conversion result. The next nine clock cycles output the contents of this register.

Write TH [01_{16}] This command writes to the TH temperature-high register. After issuing this command, the next nine clock cycles clock in the 9bit temperature limit, setting the threshold for operation of the T_{high} output.

Write TL [02_{16}] This command writes to the temperature-low register. After issuing this command, the next nine clock cycles clock in the 9-bit temperature limit which will set the threshold for operation of the T_{low} output.

Read TH [$A1_{16}$] This command reads the temperature-high register. After issuing this command, the next nine clock cycles clock,out the 9bit temperature limit which sets the threshold for operation of the T_{high} output.

Read TL [$A2_{16}$] This command reads the value of the TL register. After issuing this command, the next nine clock cycles clock out the 9bit temperature limit which sets the threshold for operation of the TL output.

Start conversion [EE_{16}] This command begins a temperature conversion. No further data is required. In one-shot mode the temperature conversion will be performed, after which the chip remains idle. In continuous mode, this command initiates continuous conversions.

Stop conversion [22_{16}] This command stops temperature conversion. No further data is required. This command may be used to halt a *DS1620* in continuous conversion mode. After issuing this command the current temperature measurement will be completed and then the device will remain idle until a start conversion command is issued to resume continuous operation.

Write config [$0C_{16}$] This command writes to the configuration register. After issuing this command the next eight clock cycles clock in the value of the configuration register.

Read config [AC_{16}] This command reads the value in the configuration register. After issuing this command the next eight clock cycles output the value of the configuration register.

Table 2. Pin functions of the DS1620 thermometer thermostat.

Pin	Symbol	Description
1	DQ	Data input/output for 3-wire communication port.
2	CLK/CONV	Clock input for 3-wire communication port. When the *DS1620* is used in a stand-alone application with no 3-wire port, this pin can be used as a convert pin. Temperature conversion will in on the falling edge of /CONV.
3	RST	
4	GND	Ground pin.
5	TCOM	High/low combination trigger. Goes high when temperature exceeds TH; resets low when temperature falls below TL.
6	TLOW	Low temperature trigger. Goes high when temperature falls below TL.
7	THIGH	High temperature trigger. Goes high when temperature exceeds TH.
8	VDD	Supply voltage, 5V input.

ister is also used to determine the method of operation that the *DS1620* will use in a particular application as well as indicating the status of the temperature conversion operation.

The configuration register is defined as follows:

Configuration/status register bits							
DONE	THF	TLF	x	x	x	CPU	1SHOT

where,

x = Don't care

DONE = Conversion done bit. 1=conversion complete, 0=conversion in progress.

THF = Temperature high flag. This bit is set to 1 when the temperature is greater than or equal to the value of TH. It will remain 1 until reset by writing writing 0 into this location or by removing power from the device. This feature provides a method of determining if the device has ever been subjected to temperatures above TH while power has been applied.

TLF = Temperature low flag. This bit is set to 1 when the temperature is less than or equal to the value of TL. It remains 1 until reset by writing 0 into this location, or by removing power from the device. This feature provides a method of determining if the device has ever been subjected to temperatures below TL while power has been applied.

CPU = CPU use bit. If CPU=0, the CLK/CONV pin acts as a conversion start control, when /RST is low. If CPU is 1, the *DS1620* will be used with a CPU communicating to it over the wire port, and the operation of the CLK/CONV pin is as a normal clock in concert with DQ and /RST.

1SHOT = One-shot mode. If 1SHOT is 1, the device performs one temperature conversion upon reception of the start convert protocol. If 1SHOT is 0, the

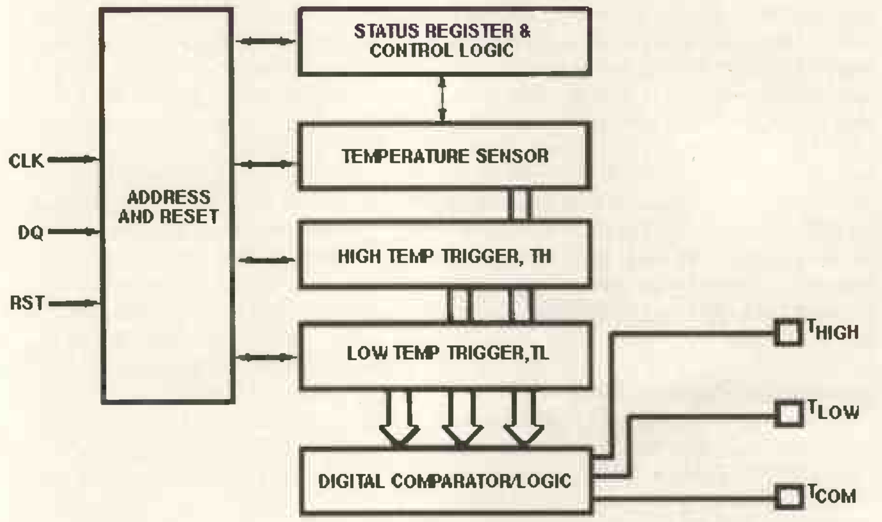
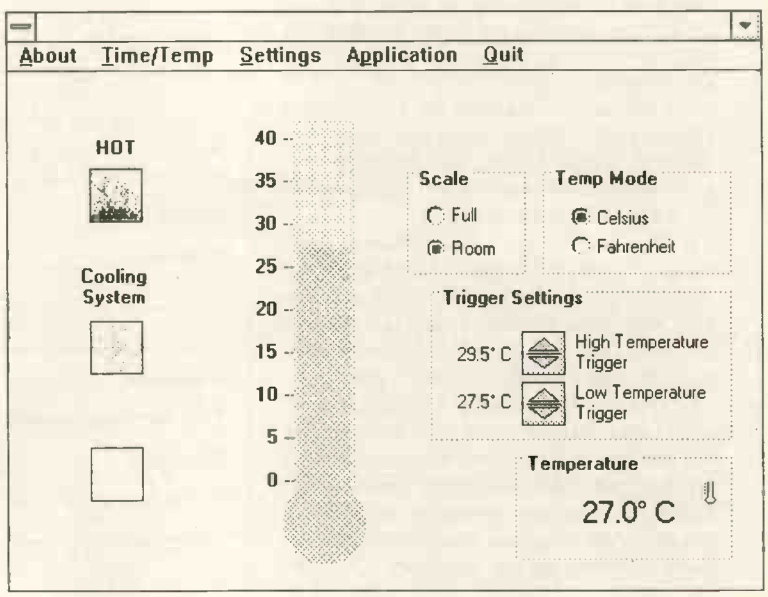


Fig. 3. Programmable via a simple three-wire interface, the DS1620 registers are used to set temperature thresholds and set the operating mode. Programming is non-volatile so the device can act as a stand-alone controller.

PC-based software – part of the designer's kit – allows the DS1620 to be read and programmed via the keyboard. Once temperature thresholds are set, non-volatile memory ensures that they remain set even when the device is removed from its socket.



device continuously performs temperature conversion

For typical thermostat operation, the *DS1620* operates in continuous mode. However, for applications where only one reading is needed at certain times, and to conserve power, one-shot mode may be used.

Note that thermostat outputs T_{high}, T_{low} and T_{com} remain in the state they were in after the last valid temperature conversion cycle when operating in one-shot mode.

Stand-alone measurements

In applications where the *DS1620* is used as a simple thermostat, no cpu is required. Since the temperature limits are non-volatile, the device can be programmed prior to insertion in the system.

To facilitate operation without a cpu, the CLK/CONV pin, pin 3, can be use to initiate conversions. Note that the CPU bit must be set to 0 in the configuration register to use this mode, **Fig. 3**.

To use the CLK/CONV pin to initiate conversions, /RST must be low and CLK/CONV must be high. If CLK/CONV is driven low and then brought high in less that 10ms, one temperature conversion will be performed, after which the device returns to idle mode.

If CLK/CONV is driven low and remains low, continuous conversions will take place until CLK/CONV is brought high again. With the CPU bit set to 0, CLK/CONV overrides the 1-shot bit if it is equal to 1. This means that even if the part is set for one-shot mode, driving CLK/CONV low will initiate conversions.

Communicating over three wires

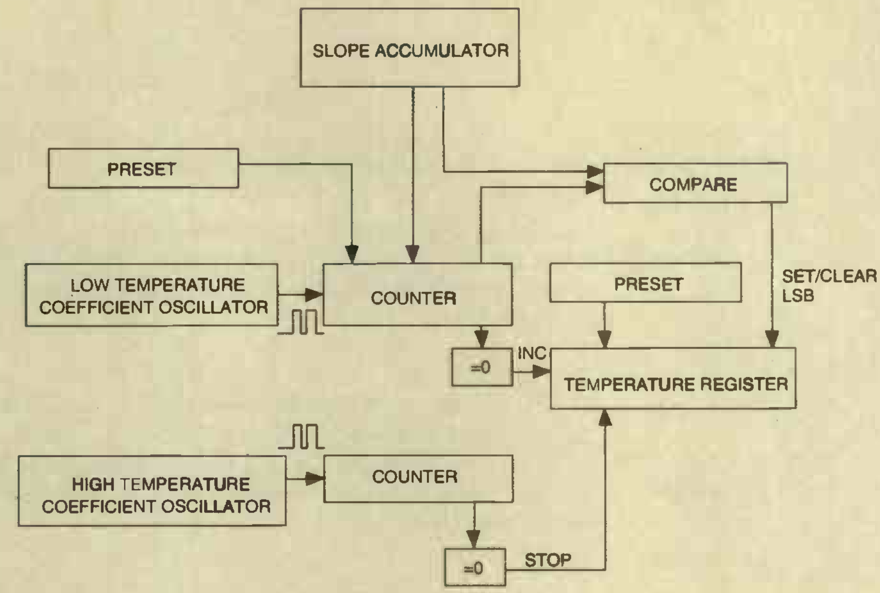
The three-wire bus comprises three signals, namely the reset line /RST, clock CLK, and data signal DQ, **Table 2**. All data transfers are initiated by driving the reset input high. Driving this input low terminates communication.

A clock cycle is a sequence of a falling edge followed by a rising edge. For data inputs, the data must be valid during the rising edge of a clock cycle. Data bits are output on the falling edge of the clock, and remain valid through the rising edge.

When reading data from the DS1620, the DQ pin goes to a high-impedance, state while the clock is high. Taking /RST low terminates any communication and causes the DQ pin to go to a high impedance state.

Data over the three-wire interface is communicated lsb first. The command set for the wire interface, as shown in the panel, is as follows; only these protocols should be written to the device, as writing other protocols may result in permanent damage to the part.

Other chips in the Dallas thermal management range include a two-wire thermometer and thermostat, a two-wire thermometer with memory and a one-wire digital thermometer. Battery temperature managers and battery identification chips providing time/temperature histograms are also available. ■



Direct-to-digital sensing

Direct-to-digital temperature sensors measure temperature through the use of an on-board Dallas proprietary temperature measurement technique.

Each temperature sensor measures temperature by counting the number of clock cycles that an oscillator with a low temperature coefficient goes through during a gate period. This period is determined by a high temperature coefficient oscillator. The counter is preset with a base count that corresponds to –55°C. If the counter reaches zero before the gate period end, the temperature register – also preset to –55°C – is incremented, indicating that the temperature is higher than –55°C.

At the same time, the counter is preset with a value determined by the slope accumulator circuitry. This circuitry is needed to compensate for the parabolic behaviour of the oscillators over temperature. The counter is then clocked again until it reaches zero. If the gate period is still not finished, this process repeats.

The slope accumulator is used to compensate for the nonlinear behaviour of the oscillators over temperature, yielding a high resolution temperature measurement. This is done by changing the number of counts necessary for the counter to go through for each incremental degree in temperature. To obtain the desired resolution, therefore, both the value of the counter and the number of counts per °C – the value of the slope accumulator – at a given temperature must be known.

Give asp a chance

While digital signal processing will do almost anything, Ian Hickman points out that the analogue kind has considerable life left in it and has definite advantages – not least that of lower cost.

Digital signal processing can be as powerful as one wants, or as one's skill in writing algorithms permits, at a price. But very often, analogue circuitry will accomplish any necessary signal processing.

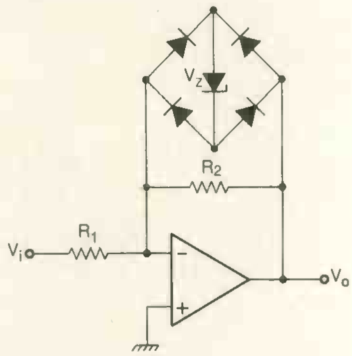
By using analogue signal processing, you avoid the use of adcs and dacs, dsp chips, memory – and the expense of writing algorithms, achieving a much lower power budget. Eliminating clock signals can also be a benefit where small signals are being handled in a physically small enclosure. Analogue signal-processing techniques come in a wider range than many people realise, and this article illustrates some of them.

Bounding, limiting and clipping

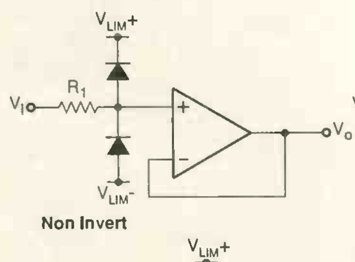
It is often necessary to limit the maximum excursion of a signal, for example when it contains large interference spikes, or when the lower amplitude parts of it need to be amplified for more detailed measurements.

Figure 1(a) shows the traditional way of doing this. If R_s is the slope resistance of the zener diode and the bridge diodes, the gain of the inverting op-amp falls from R_2/R_1 to R_s/R_1 whenever the output tries to exceed $\pm(V_z+2V_s)$, where V_z is the breakdown voltage of the zener diode and V_s is the forward voltage of the bridge diodes. For large signals, the gain falls almost to zero, or minus infinity decibels.

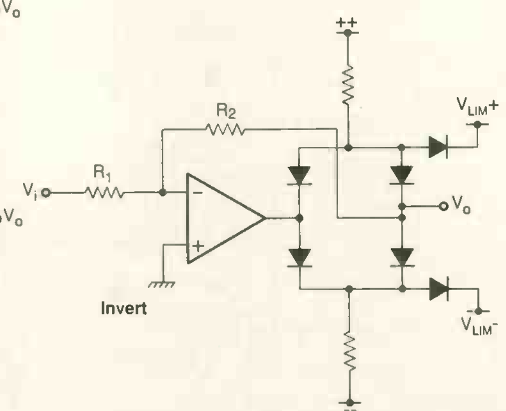
(a)



(b) Non Invert



Invert



(c)

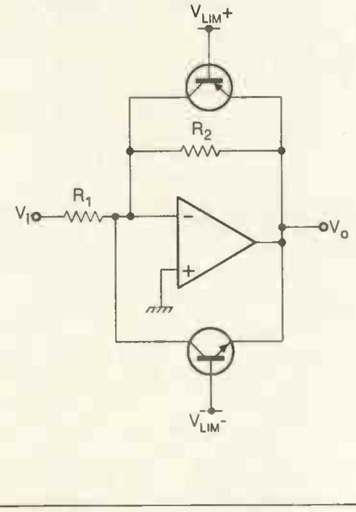


Fig. 1. Circuit providing symmetrical limiting in an inverting amplifier at (a), while at (b) two circuits provide separately adjustable positive and negative limiting levels – though the inverting version is not recommended. Transistors in (c) provide very effective limiting in an inverting circuit.

Disadvantages of the circuit are possible loss of bandwidth. This is due to capacitance associated with the bridge circuit shunting R_2, and the fact that the positive and negative limits are not easily and separately adjustable.

Figure **1(b)** shows a simple non-inverting circuit with separately adjustable positive and negative breakpoints V_{lim+} and V_{lim-}, and a distinctly more complicated inverting version.

The inverting version is not recommended for fast signals, since in the overdriven condition the op-amp is left open-loop. This means that its output will fly off and hit one or other of the supply rails. Recovery of a conventional op-amp from overdrive is a relatively slow process, limiting the bandwidth of the circuit. If you can live with this, it is simpler to go for a larger gain and simply let limiting occur at the rail voltage.

Transistors provide effective limiting in an inverting circuit, providing the output swing keeps within the reverse V_{be} ratings of the devices. As Fig. **1(c)** shows, breakpoints are separately adjustable.

Another example of limiting is shown in **Fig. 2(a)** – a circuit designed to measure the settling time of an op-amp, by using a 'false sum node'. This needs low-capacitance Schottky diodes, with their low forward voltage drop, to enable the 1% settling time to be measured.

In the case of a fast op-amp, the capacitive loading of a conventional ×10 divider probe on the false sum node may limit measurement accuracy, so it is best to use an active probe. Measurement depends critically upon the flatness of the top and bottom of the test squarewave; for measurements of settling time to the 0.1% or 0.01% level, more sophisticated limiting arrangements are called for[1].

If you need a wideband amplifier with symmetrical limiting, the *Linear Technology LT1194*, with its 35MHz –3dB bandwidth, provides a simple and convenient solution, Fig. **2(b)**. This ingenious device[2] provides a limiting level adjustable by means of a control voltage V_c in the range –5V to –1V, with no requirement for any additional components whatever.

An additional advantage of this device is that the gain-defining negative-feedback loop is completed via a second long-tailed pair in parallel with the input pair. This means that the inverting and non-inverting inputs are effectively floating and both present a high input impedance.

Diodes in the circuit of Fig. 1(b) protect the op-amp input circuit, whereas some of the others only limit the output swing. Where a circuit needs protection against really large inputs, use the arrangement of Fig. 2(c); for large positive inputs, D_1 is reverse-biased,

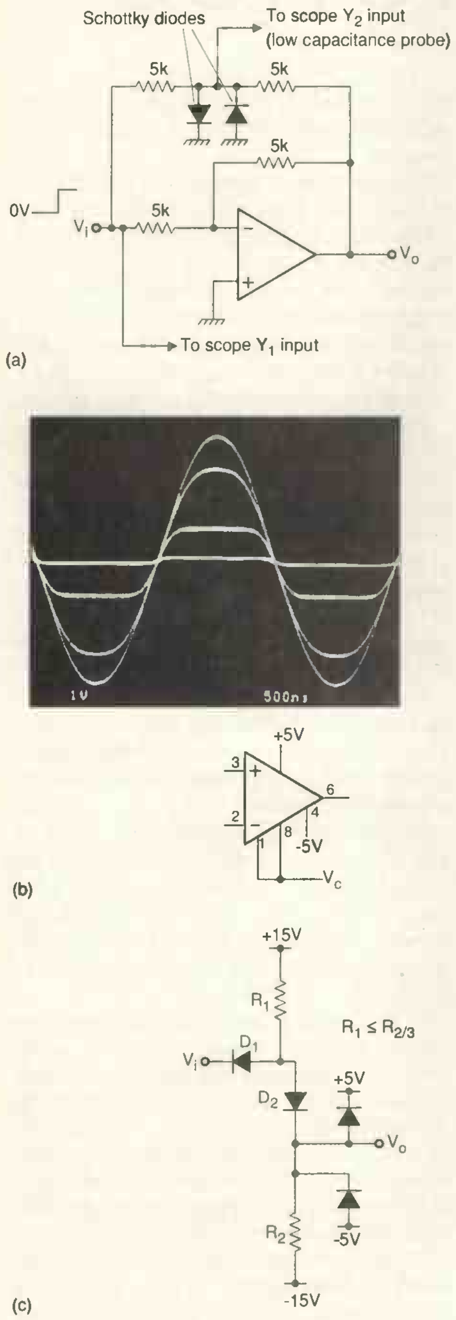
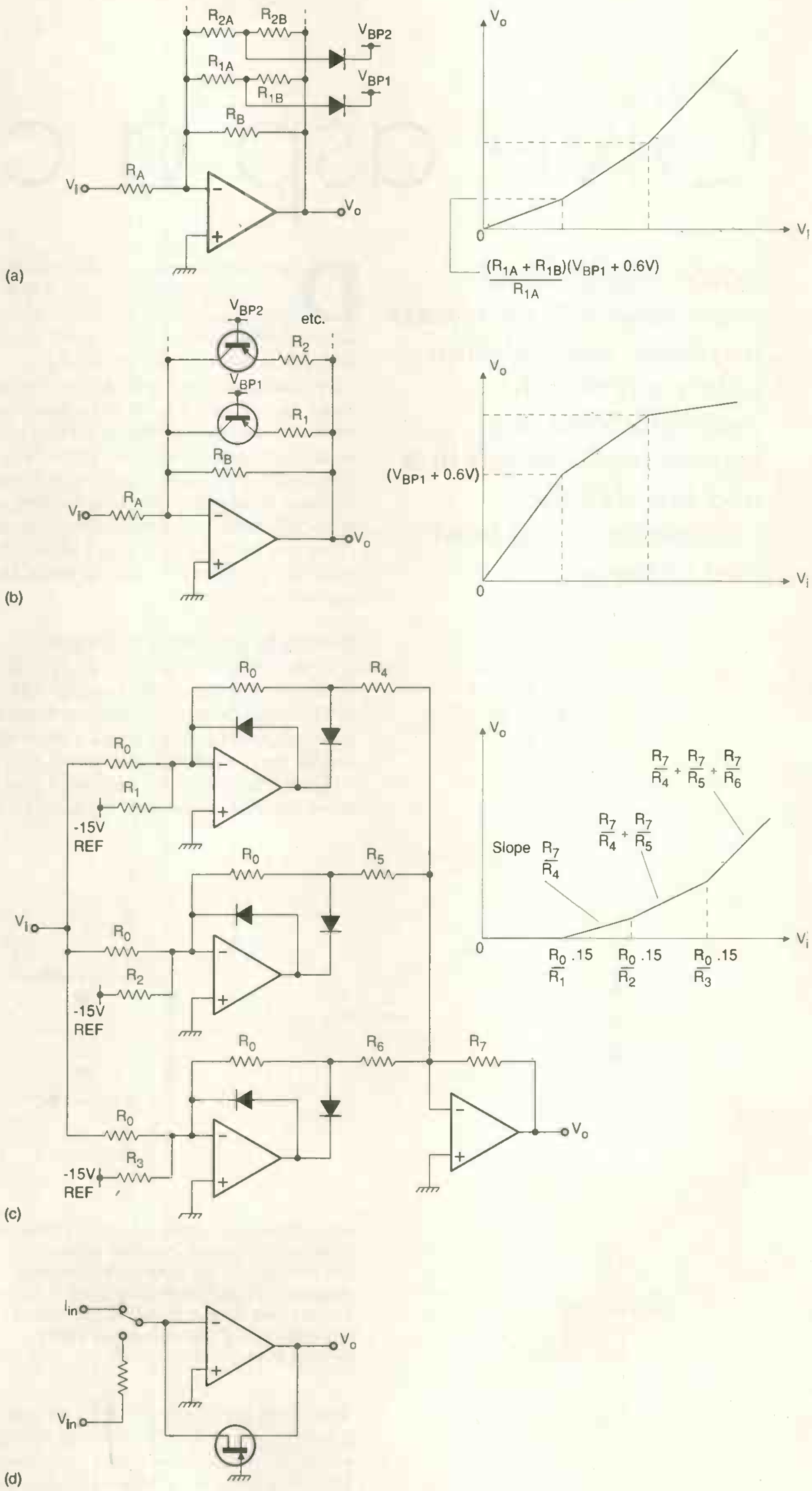
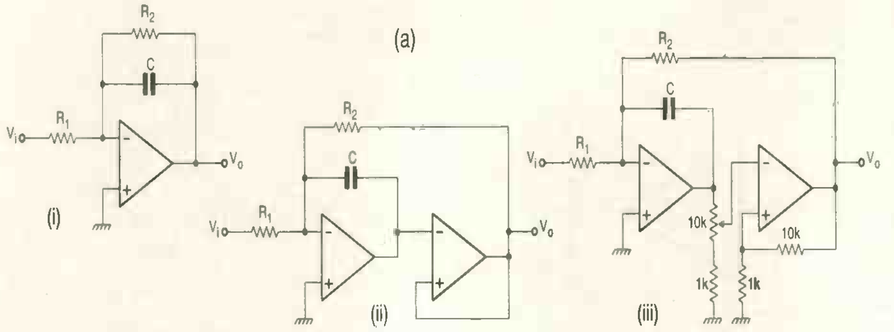
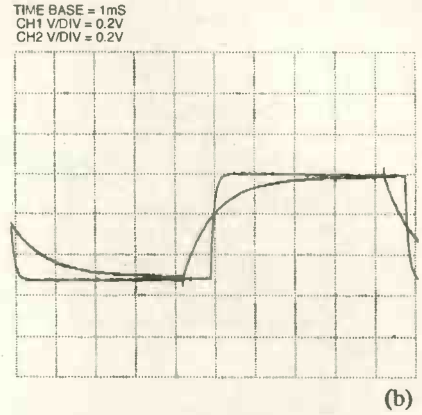


Fig. 2. Diode limiting (a) used in a circuit to measure the settling time of an op-amp. At (b) the LT1194 wideband op-amp circuit with voltage controlled symmetrical clipping level. Reproduced by courtesy of Linear Technology Corporation. Simple bounding circuit in (c) protects following stages from mains inputs.

Fig. 3. Diode breakpoint circuit (a), providing increasing gain with increasing output voltage. Transistor breakpoint circuit in (b), provides decreasing gain with increasing output voltage. Circuit at (c) gives sharp, temperature independent breakpoints and (d) an output voltage proportional to the logarithm of input voltage or current.







while for large negative inputs D_2 is reverse-biased, the output voltage being limited to about ±5V.

Given suitable reverse voltage ratings for $D_{1,2}$ and a large enough dissipation rating for R_1, the arrangement will protect any following circuitry from connection – accidental or otherwise – to 230Vac. In the linear range, for inputs between +5V and –5V, V_o follows V_i.

If R_1 were only equal to $R_2/2$, it could only just succeed in raising the output to +5V. This would leave no spare current to charge the inevitable stray capacitance up rapidly. A lower value avoids a poor frequency response, albeit increasing the dissipation in R_1 for large negative input voltages.

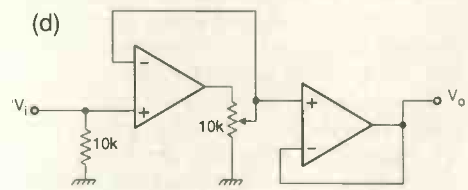
If the circuit is turned upside down and the diodes all reversed, an n-channel mosfet with a 600V drain voltage rating, operating as a constant-current generator, can be substituted for R_1; the dissipation in this 'active R_1' with a large input voltage is only proportional to the input voltage v_{in}, not to v_{in} squared.

Breakpoints and non-linear gain

While Figure 1(a) illustrates an extreme example of a circuit with non-linear gain, there is often a requirement for the gain of a circuit to vary over a range of finite values as the output level varies, rather than suddenly falling to zero. **Figure 3(a)** shows a circuit that provides increasing gain as the input increases in the negative direction. This is because initially, the R_1, R_2 networks are in parallel with R_B, their effect being successively removed as each breakpoint is exceeded.

In Fig. **3(b)**, gain decreases as the output voltage exceeds each successive breakpoint, additional feedback resistors being added in parallel with R_B. With reversed diodes or n-p-n transistors and negative breakpoint voltages, operation is extended to negative-going outputs for both circuits. Both types of breakpoint may be used together to give more com-

Fig. 4(a) 'Leaky integrator' forms a simple slew-rate limiting circuit (i), and a functionally identical circuit at (ii). Adding adjustable gain in the second op-amp stage (iii) provides a variable slew-rate limit, as illustrated in (b). As the gain in the second stage is increased, the required voltage excursion across the capacitance is reduced. This is equivalent to reducing its capacitance, increasing the slew rate shown at (c). This arrangement is effectively a linear amplifier with a high frequency roll-off. Consequently, for a fixed setting of the potentiometer, increasing signal amplitude results in increased slew rate. The circuit can set any desired limit to the maximum slew rate of the largest signal but, as the signal gets smaller, so does the slew rate. (d) This limiter circuit can be used to define the maximum size signal input to the slewrate limiter.



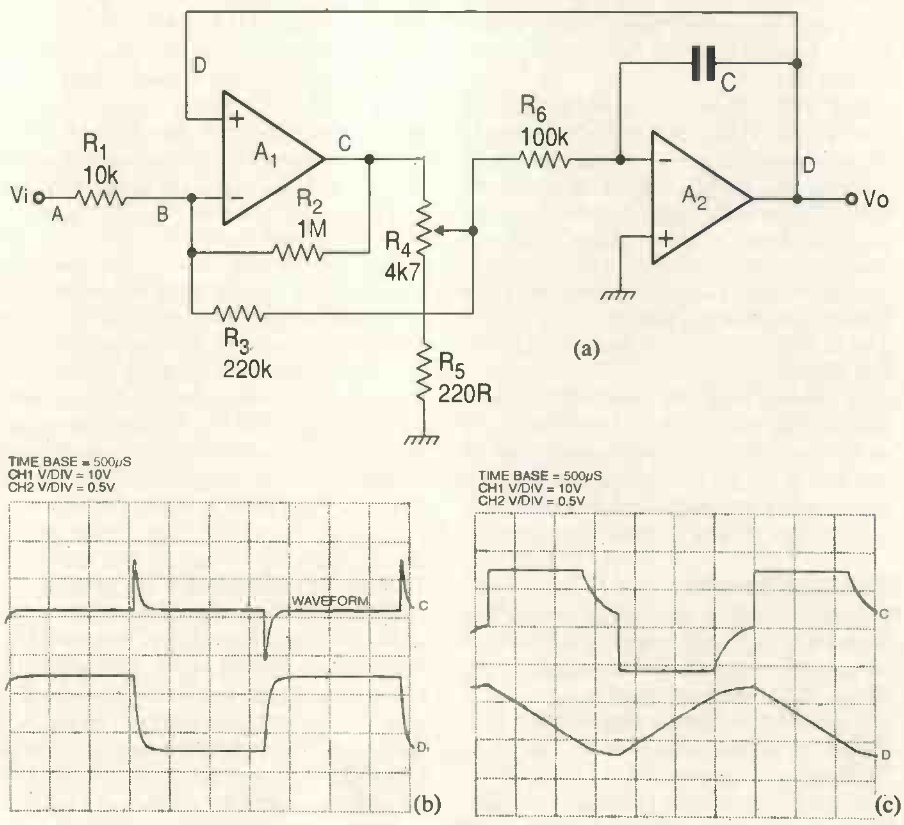
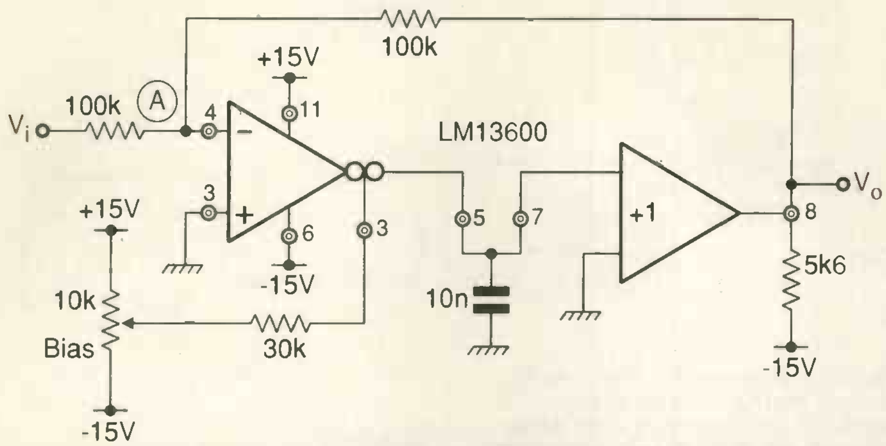


Fig. 5(a). This slew-rate limiter is an improvement on the leaky integrator of Fig. 4, providing a constant maximum slewrate limit, regardless of the signal amplitude. Set for the fastest slewrate (wiper of R_4 at A_1's output) and with the largest signal it can handle linearly, (b) shows that the circuit rapidly settles exponentially. With the same input, but R_4's wiper now at the junction with R_5, (c) shows the circuit settling with a linear ramp, topped off with a slower exponential tail.



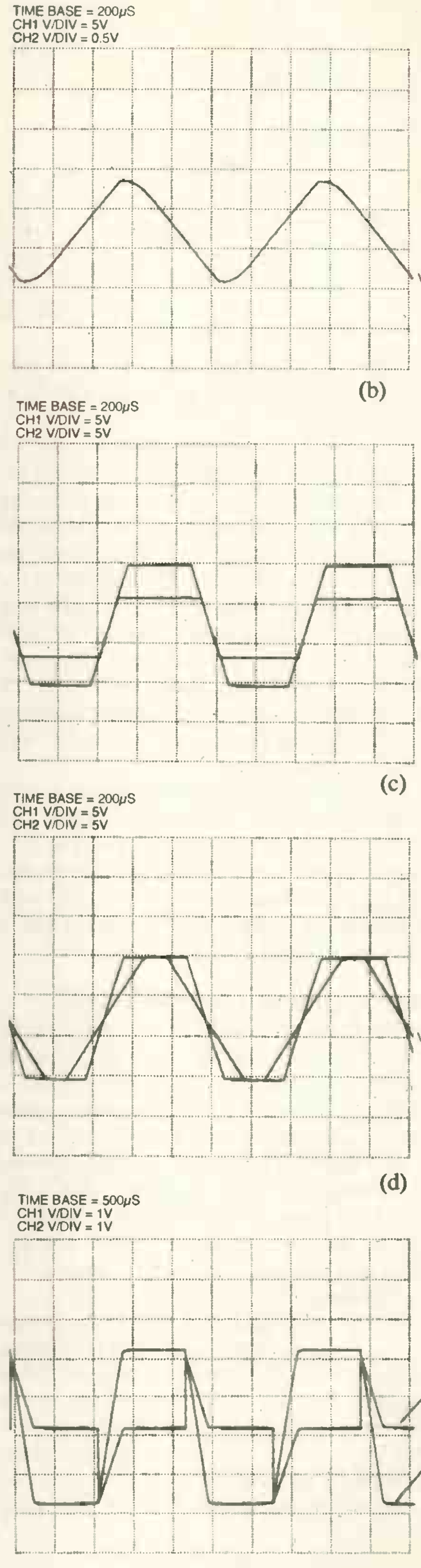


Fig. 6(a) This circuit is a true slew-rate limiter. Output in (b) follows the sinewave input from the peak (where the slope dV/dt is zero) up to the preset slew-rate limit. Thereafter, the set slew rate applies until the loop is again closed, where the output rejoins the ideal waveform, just before the next peak. As the amplitude of a squarewave input in (c) decreases, the slew rate remains constant. Unlike the circuits of Figs 4 and 5, for any given slew-rate limit setting, the slew rate at (d) remains constant until the output rejoins the input squarewave. During the slewrate limited section of the output waveform in (e), the amplifier is open loop. Thus the ota inverting input (point A) during this period is not a virtual earth.

plicated shaping[3].

Circuits Figs 3(a) and (b) provide a gentle transition from one slope to the next, extending over a range of around 100mV or so, as the diodes or base/emitter junctions move from cut-off to conducting. Despite some consequent variation in breakpoint with temperature, this rounding is often beneficial.

Where sharply defined breakpoints free from temperature variations are needed, use the circuit of Fig. **3(c)**. Here, the diode drops are all within the loop and do not affect circuit performance. When this circuit first appeared, the use of one op-amp per breakpoint was considered almost profligate, but high-performance quad op-amps are now commonplace components.

With jfets, either as elements in feedback networks, or to vary the control voltage of a voltage-controlled amplifier, you can achieve smoothly varying gain without discrete breakpoints. Figure **3(d)** is an example of a smoothly varying function of a very specific nature. It exploits the law governing p-n junctions to provide an output voltage proportional to the logarithm of the input voltage or current over a wide range – up to nine decades with suitable devices.

An op-amp with high open-loop gain is needed to keep the base/collector voltage close to zero, or collector leakage current ruins the log. law at the lowest input levels. Parasitic parameters of the transistor cause the circuit to become very slow at very low input levels.

Slew-rate limiting

Circuits for limiting slew rate prevent the rate of change of a signal exceeding some design maximum – whatever the amplitude of the signal. This is often necessary in electronically controlled mechanical systems with large inertia to prevent excessive forces being applied to moving parts.

A 'leaky integrator' can form a simple slew-rate limiting circuit, as shown in **Fig. 4(a)**. Here, circuit (ii) is obviously functionally identical to that in (i). In both cases, the slew rate is limited by feedback via the capacitor.

If now a degree of gain is incorporated in the second op-amp stage, as in (iii), full output swing is obtained with only a reduced swing appearing across the capacitor. This is equivalent to reducing the value of the capacitor, changing the frequency at which the stage's frequency response starts to roll off, without changing the low-frequency gain.

So varying the amount of gain in the second op-amp stage provides a variable slew-rate limit, as illustrated in Fig. **4(b)**. This is simply a linear amplifier with a high-frequency roll-off. Consequently, for a fixed setting of the potentiometer, increasing signal amplitude results in increased slew rate. Thus the circuit can set any desired limit to the maximum slew rate of the largest signal, but as the signal gets smaller, so does the slew rate; see Fig. **4(c)**.

Set the largest signal input by means of a limiter, such as that shown in Fig. **4(d)**; this provides unity gain for small signals but if the potentiometer is set midway, maximum output swing will be limited to half the op-amp's rail-to-rail capability and progressively less as the wiper of the potentiometer approaches ground.

If the integrator of Fig. 4(a) is placed second instead of first and a few other changes made, an improved slew-rate limiter results; see **Fig. 5(a)**. To understand how the circuit works, imagine that the wiper of the potiometer is at the top of its travel. Now, $R_{2,3}$ are in parallel, defining the gain of A_1 as ×18, inverting.

Figure **5(b)** shows the output of A_1, upper trace, and of A_2, lower trace, when a 300Hz squarewave is applied, of just sufficient amplitude to provide the maximum outut swing of which A_1 is capable. Initially, A_1 works as an inverting amplifier, because the voltage across C cannot change instantaneously. So the negative-going edge of the input causes the positive output at A_1, which in turn is applied via R_6 to the integrator, making its output slew negatively.

The circuit settles with A_1 output at zero, otherwise the integrator output would still be changing, and with the output voltage and A_1's non-inverting input at almost the input voltage. Output voltage v_o is given by

$v_{in}R/(R+10k\Omega)$, where R is A_1's effective feedback resistor — in this case, R_2 and R_3 in parallel. As Fig. 5(b) shows, the output settles exponentially to the peak value of the squarewave input.

When the wiper of R_4 is wound down to the R_5 end of its travel, the action is very different and is illustrated in Fig. **5(c)**. Now, only one twentieth of A_1's output is applied to R_3, whose effective value as a feedback resistor is therefore not 220kΩ, but 4.4MΩ. In conjunction with R_2, this gives a demanded A_1 gain of ×81.

With the same input amplitude as before, A_1 is now heavily overdriven and, moreover, the voltage driving the integrator stage is also reduced to one twentieth. So A_1 remains overdriven while the integrator output slews at a constant rate. It does this until voltage at A_1's non-inverting input is so near that at its inverting input that A_1 re-enters the linear range. Thereafter, the circuit settles exponentially as in (b), but on a longer time constant – so long, in fact, that in (c) at this frequency and amplitude the output never quite reaches the peak value before the next edge of the squarewave.

Whatever the frequency, amplitude or waveshape of the input, the slew rate set by the position of R_4 is never exceeded. As long as A_1 is overdriven, the first part of the settling will be at the maximum slew rate – however small the input signal. This is a big advance on the Fig. 4, but the exponential tail to the settling time, visible in Fig. 5(c), remains a disadvantage.

A substantial improvement in this respect, at the cost of a reduced range of slew-rate adjustment, is obtained by connecting a 1MΩ resistor between the output of A_1 and the inverting input of A_2. Although this speeds up the exponential end of the settling tail, it can never be entirely eliminated. An ideal slew-rate limiter would at all times slew at the same rate as the input signal, or at the maximum rate, whichever was the lower.

Figure 6(a) shows the circuit of such a true slew-rate limiter. You can see in **6(b)** how the output follows the sinewave input from the peak – where the slope dv/dt is zero – up to the preset slew-rate limit. Thereafter, the set slew rate applies until the loop is again closed. At this point the output rejoins the ideal waveform, just before the next peak.

If either the frequency or the amplitude is reduced (decreasing the maximum slew rate of the signal) or the control voltage V_c is increased (increasing the ota's maximum transconductance), the sinewave is undistorted. On the other hand, if the amplitude or frequency are sufficiently increased (or V_c reduced) the sinewave is permanently in slew-rate limit – it becomes a triangular wave.

Figure **6(c)** shows how, as the amplitude of a squarewave input decreases, the slew rate remains constant – unlike the effect of the circuit of Fig. 4. Also, the slew rate remains constant until the output rejoins the input squarewave, as in Fig. **6(d)**. During the slew-rate limited section of the output waveform, the amplifier is open-loop and the ota inverting input is not a virtual earth during this period, exhibiting the waveform shown in Fig. **6(e)**. Note the pull-down resistor at the operational transconductance amplifier's Darlington output buffer stage. This is recommended if using the *LM13600*, since the internal biasing of the buffer in this ota is varied in sympathy with the control voltage V_c: the *LM13700*, with its fixed buffer current, might be a better choice in this application.

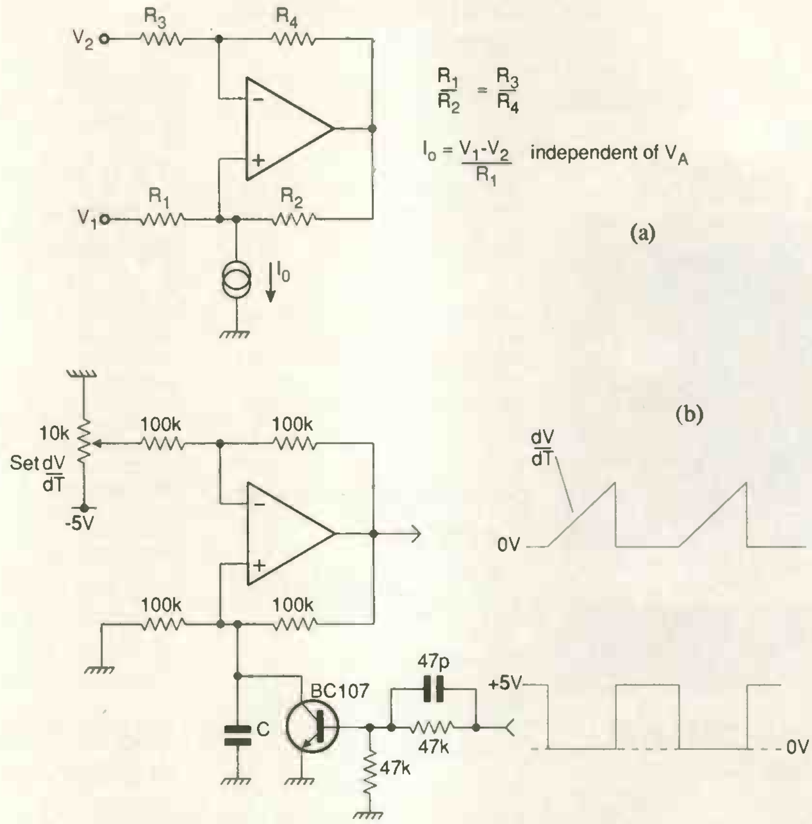


Fig. 7(a). Howland current pump provides an accurate, linear, bipolar, voltage-controlled current source. Non-inverting integrator at (b) is arranged as a linear timebase circuit.

Integrating and differentiating

Basic op-amp circuits for integrating and differentiating are so well known that I won't spend any time on them here. But the Howland current pump[4] in **Fig. 7(a)** is perhaps a circuit that deserves to be better known. It is a voltage-controlled current generator with (ideally) infinite output impedance, and causes a current $(V_2-V_1)/R_1$ to flow in a load.

If the load is a capacitor and V_1 is tied to ground, then the circuit forms an integrator. As such, it possesses two advantages over the more usual op-amp integrator: firstly, it is a non-inverting integrator, and secondly, one end of the integrator capacitor is grounded. Figure **7(b)** shows a linear ramp or timebase generator based on the circuit of Ref. 3.

Some of the circuit techniques mentioned earlier can usefully be combined with integrators and differentiators. For instance, if a band-limited signal is differentiated, clipped and then integrated, the result is slew-rate limited. If, instead of clipping, the differentiated signal is slew-rate limited and then integrated, then its second derivative d^2v/dt^2 is limited.■

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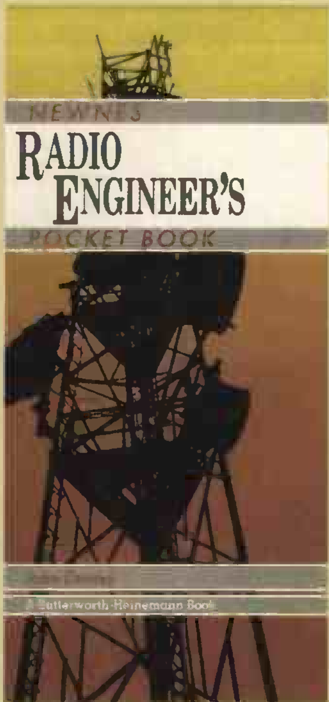

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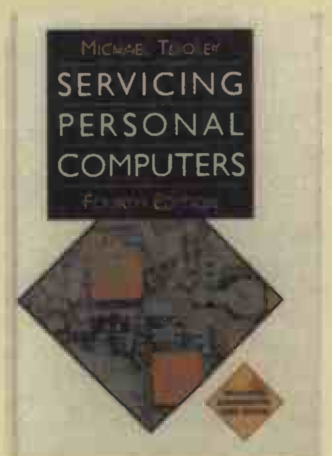

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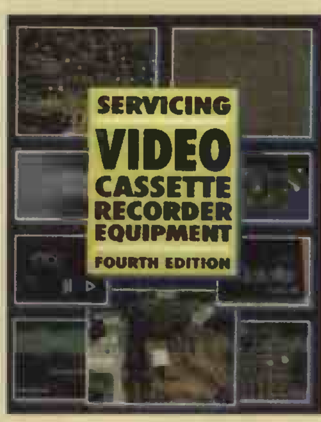

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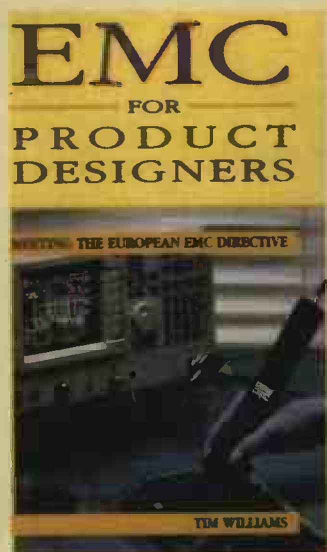

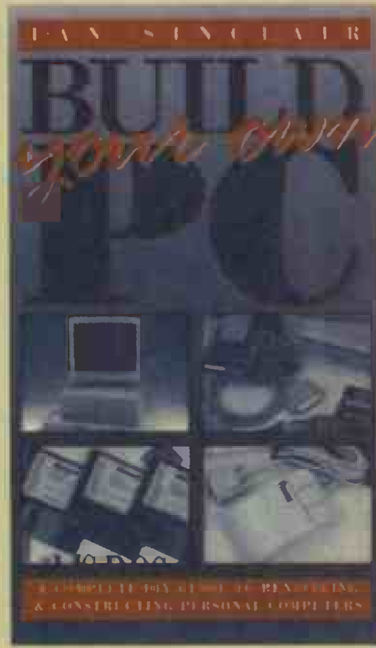

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This antenna design has fewer critical dimensions then most other configurations, and can be easily customised. It comprises a lightweight parabolic shaped plane reflector which is illuminated with a horizontally polarised dipole and reflector feed. Size of the reflector suggested is about the minimum worthwhile for 1.3GHz operation.

Design considerations

Both electrical and mechanical factors have to be considered when deciding on the reflector size and shape. One aspect taken into account in this design was to keep the structure light, and small enough to move around in a family car. The feed boom is removable for transportation. If two wing nuts are used on the mast clamp, it is possible to assemble the antenna in about half a minute and without tools, **Fig. 1**.

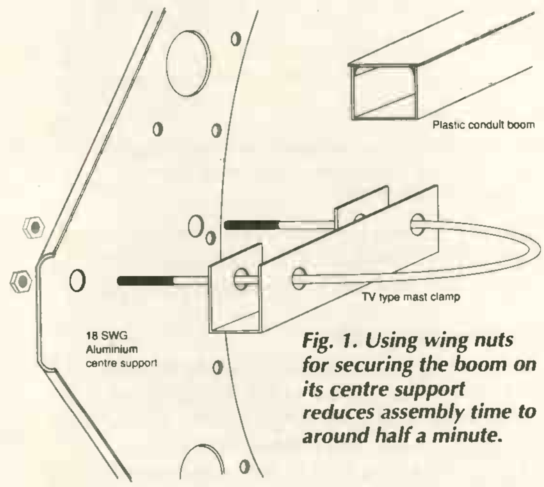


Fig. 1. Using wing nuts for securing the boom on its centre support reduces assembly time to around half a minute.

The distance of 20in from the reflector to the focus was chosen mainly for mechanical reasons. It was arrived at through the use of full size drawings. Performance tests on the antenna show the choice to be acceptable.

The reflector should be formed as accurately as possible, but departures from the mean shape of up to one-tenth of a wavelength are generally tolerable. This is 2.3cm at 1.3GHz – an easily achievable tolerance.

Mesh for the reflector can be seen as a large number of reflecting elements spaced one-thirtieth of a wavelength over most of the reflector surface and one-sixtieth of a wavelength in the area of the overlap at the centre. No figure for actual reflectivity is available. Wind resistance is considerably less for this material than for expanded aluminium mesh.

Test conditions

Gain of the antenna comes from concentrating the radiation in the vertical plane. Claims for antenna gain are notoriously controversial. My antenna was tested in a coastal car park with the sea behind at 1W transmitter power.

The receiving station was about 2.4km inland and used a multi-element yagi antenna combined with a commercial receiver with an S-meter driven from the automatic gain control line. The S-meter was calibrated using external attenuators in the antenna feeder. During the tests, the attenuators were also used to extend the dynamic range of the receiving system.

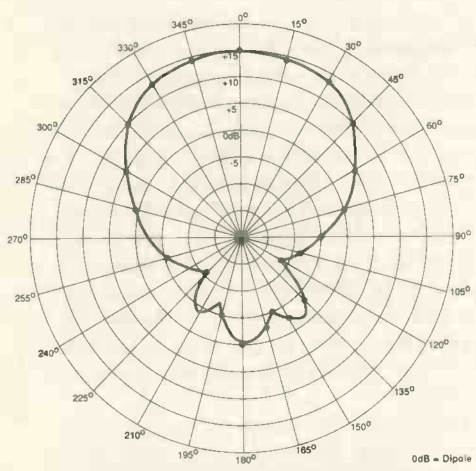
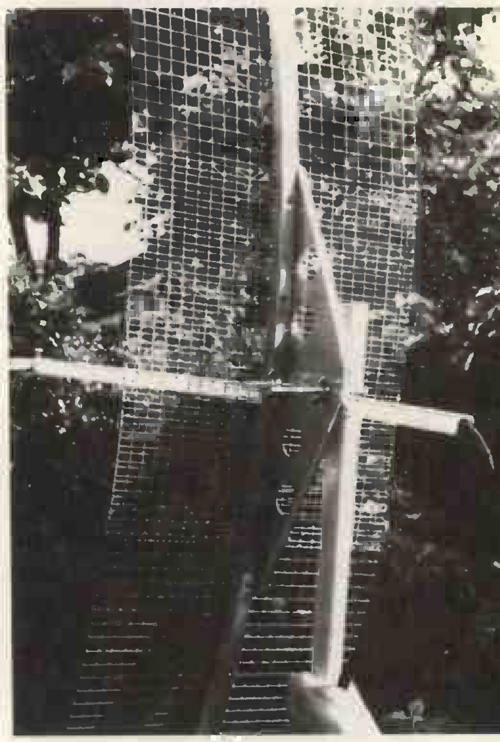


Fig. 2. Horizontal test pattern of the 24cm parabolic antenna. Frequency 1255MHz, swr 1:1.2 and power 1W.

The centre of the test antenna was about 2.4m above ground and the mast was equipped with a scale marked with 15° divisions. A dipole antenna was used to establish the 0dBd reference. The horizontal radiation pattern of the test aerial, **Fig. 2**, was remarkably smooth compared to those of several other aerials tested. An honest attempt to obtain an accurate gain figure yielded a result of 15dBd (*power*×31.6) or, say, 17dBi. The degree of experimental error was not determined.



Featuring wide beam, wide bandwidth and useful gain, the 24cm antenna is easily disassembled for transportation.

Feed design

Wide bandwidth was necessary so that the antenna could be used over as much of the 1.3GHz band as possible. This use includes amateur television.

The feed design satisfies the requirement through the use of few tuned elements and the type of balun chosen. The dipole element is large in diameter relative to the operating wavelength, the size being chosen to bring the centre impedance close to 50Ω.

Dipole to reflector spacing is approximately 0.2 of a wavelength but the actual distance can be varied to allow fine adjustment of the feed impedance. Length of the dipole elements can

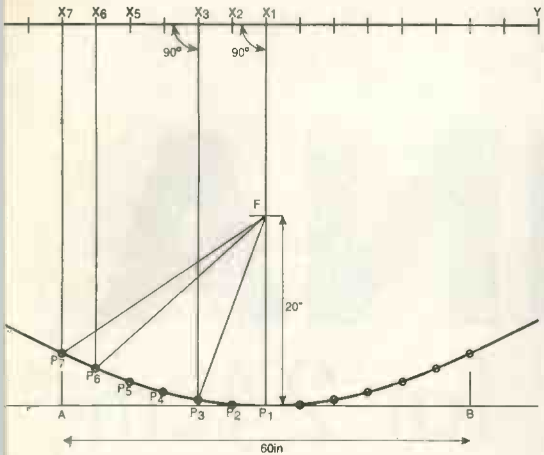


Fig. 3. Antenna's parabola profile can be calculated then plotted, but it can also be drawn without resorting to mathematics by using a pin and string.

be trimmed so that the dipole becomes resonant at the design centre frequency, providing a resistive match at the frequency.

The centre frequency chosen for the antenna is 1.255GHz. Bandwidth of the feed impedance is further increased by making the balun slightly longer than a quarter wavelength electrically. This adds inductive shunt reactance at frequencies below the centre frequency of the design and capacitive shunt reactance at higher frequencies. In this way, it is possible to extend the matching over a wider bandwidth.

Constructional notes

Both imperial and metric dimensions are used in the following description, as the materials used are produced variously in both size units.

A full-size drawing of the parabola profile is needed. Several methods of doing this are shown in handbooks. The beauty of the string and pin method described here is that it requires no mathematics, **Fig. 3.**

First draw line AB, and then at its centre draw line PFX, at right angles to AB. Next either draw line XY parallel to AB, or use a long rule or tape parallel to AB, this must be marked off with regular divisions – X_1, X_2 etc: Now fix one end of a piece of string at point F using a pin. Take the string around another pin at P_1, and then up to point X_1, and mark this length with a knot. Now plot the curve by moving the knot to X_2, and keeping the string at right angles to XY, prick a mark at P_2, and so on, until half the curve is marked out. Repeat for the other half of the curve. Draw a line smoothly rough the pin pricks to show the shape of the reflector surface.

This graphical method illustrates the action of the reflector. If XY is a wavefront, the length of string show lines of constant phase to the focus.

Next the reflector former has to be shaped to fit the drawn profile, **Fig. 4**. The former uses 60in of $^1/_2$in square aluminium tube stock, as commonly used for tv yagi antenna booms.

Use a simple bending machine as shown in **Fig. 5** to avoid crumpling. This consists of a block of hardwood, cut in two by a curved cut with a radius slightly less than the curve at the centre of the required profile. A small scrap of plywood tacked to the underside of one of the two pieces will help to keep things in line.

Shaping of the former is carried out by making a sandwich of the blocks and the tubing in the jaws of a vice. Starting from the centre, squeeze the former gently to shape, moving the metal along an inch or so at a time. When correctly shaped, it can be used to mark out the 18swg aluminium centre support.

Rigidity of the former can be considerably improved by shaping the straight edges round a suitable mandrel of, say, half-inch diameter rod, with a mallet to produce a rolled edge.

A 3-by-2ft sheet of wire mesh called Handy Mesh was obtained from a local hardware store to provide the material for the reflector surface. The mesh consists of half-inch squares formed by 22swg tinned wire. The sheet was cut in half along its length and the pieces overlapped to form a strip 60 inches long. The overlapped section was strengthened by binding some of the coincident wires with 22swg tinned copper wire, then soldering.

A small, off-centre hole will be required to clear the boom, as shown in the drawing. The mesh is fixed to the square tube former by gripping between a strip of three-quarter inch wide 22swg aluminium, with four BA size self tapping screws or one eighth inch pop-rivets.

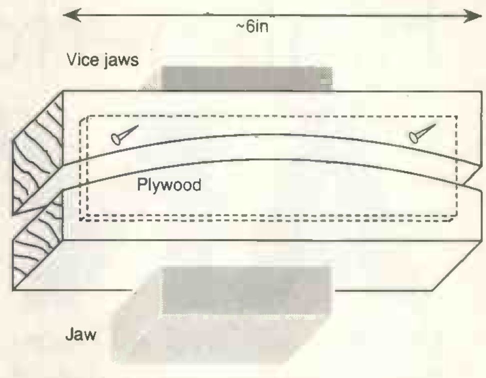


Fig. 5. To avoid having the tube crumple during curve shaping, it is advisable to make yourself a bending fixture.

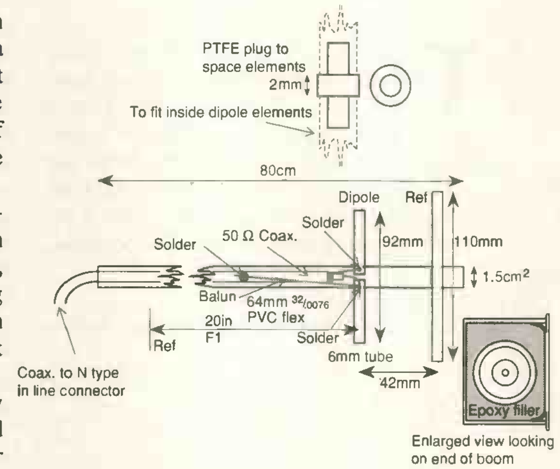


Fig. 6. Plan of antenna boom with feed dipole and reflector. The boom is electrical conduit with a removable lid.

Feed boom details

Dimensions and construction of the feed boom and elements are shown in **Fig. 6**. The boom is an 80cm length of 1.5cm square plastic electrical conduit. This conduit has a press-on capping and is a tight fit in the tv-type square boom mast clamp.

The elements were made from 6mm diameter aluminium tubes, the dipole connections being made with special aluminium solder. Thin-walled brass or copper tube could also be used. The small plug in the centre of the dipole keeps the halves under control during assembly. Each element is held in place with epoxy resin filler. This same material is used to weatherproof the end of the co-axial cable.

The balun uses a length of common 32 strand of 0.0076in PVC-covered flex, joined between the outer and inner of the co-axial feeder cable. To obtain maximum bandwidth, the length of 64mm was arrived at experimentally.

Finishing touches can include replacing the U-bolt clamp fixing nuts with wing nuts, weather-proofing the ends of the square tubes with plastic plugs and varnishing overall. ■

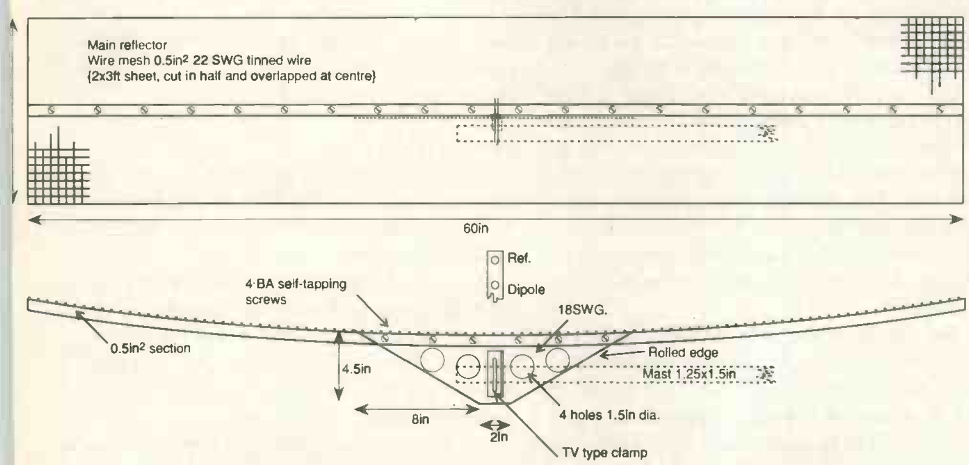


Fig. 4. Reflector detail. Mesh is a commercially-available type comprising half-inch squares formed from 22SWG tinned wire.

NEW PRODUCTS CLASSIFIED

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ACTIVE

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

Low-power, low-cost a-to-d. Analog's *AD876* joins the range of lower-power analogue-to-digital converters, delivering a 20Msample/s conversion rate at a cost of $9.95 in quantity. It possesses a differential track-and-hold amplifier to eliminate the need for an external buffer and digital error correction allows immediate sampling without the need for internal calibration. Power is 160mW at 5V. Analog Devices Ltd. Tel, 01932 266000; fax, 01932 247401.

Discrete active devices

Power mosfets. Two 30V SOIC power mosfets from Siliconix, the *Si4410DY* and *Si4435DY*, possess half the on resistance of any other on the market, says the company. 4410 is an n-channel device giving 13.5mΩ at 10V gate voltage and 20mΩ at 4.5V, while the 4435 has 20mΩ at 10V and 35mΩ at 4.5V. Both use the company's Trench technique to provide transistor densities of 12 million cells per square inch. Siliconix/Temic Marketing. Tel., 01344 485757; fax, 01344 427371.

Low-leakage diodes. Precision diodes by Zetex in the *FLLD200* series of dual devices feature reverse currents of up to 800% better than standard types. Three models are available: the *FLLD263* common-anode type; the 258 common-cathode version and the series-connected 261. Leakage currents are of the order of 3-5nA/100V, with a typical diffusion capacitance of 0.9pF at 1V. Average rectified forward current is 250mA. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5467.

Digital signal processors

Silicon delay lines. *Data Delay Devices* all-silicon delay lines, in SOIC or 14-pin dips, offer five and ten equally spaced taps and give delays from 5ns to 500ns. Operating frequency is 100MHz and ground bounce is said to be low. The devices are compatible with ttl and cmos. BFI IBEXSA Electronics Ltd. Tel., 01622 882467; fax, 01622 882469.

Linear integrated circuits

Dual amplifiers. Current-mode dual amplifiers by Elantec, the *EL2280* and *EL2270* operate from 3V or 12V (±1.5V, ±6V) supplies at low power and with good video performance. Bandwidth at –3dB is 250MHz (70MHz), supply current 3mA (1mA), and slew rate is 1200v/μs (800V/μs) and differential phase and gain of 0.05° (0.15°) and 0.05% (0.15%). Elantec. Tel., 0171-482 4596; fax, 0171- 267 1026.

Video op-amp. Having a 300MHz bandwidth at a 1mA supply current from 5V, the *AD8011* op-amp by Analog Devices is for use in the processing of high-speed video, giving 0.1dB gain flatness up to 25MHz, 0.02% differential gain and 0.06° differential phase error. Worst-case thd is –62dB at 20MHz into 150Ω and the amplifier slews at 3500V/μs, settling to within 0.1% in 25ns. Analog Devices Ltd. Tel, 01932 266000; fax, 01932 247401.

Memory chips

1Mbit synchronous srams. Since the *μPD431232LGF-A8* 1Mbit synchronous static ram has a data access time from the clock of 8ns, it is suitable for use with 66MHz processors and, being organised as 32 by 32, will form a 256K cache memory using two devices. These devices have a burst counter and produce either an interleaved addressing sequence to suit the Intel Pentium *PE54C* and *8046* processors or a linear sequence for MIPS types. NEC Electronics (UK) Ltd. Tel., 01908 691133; fax, 01908 670290.

PCMCIA memory cards. *Centon* sram cards are available with storage capacities from 64Kb to 2Mb, with or without attribute and in flash memory, in 5V/5V or 5V/12V versions, from 256Kb to 32Mb capacity. Also available are ATA flash, Ethernet and fax/modem cards. METL. Tel., 01844 278781; fax, 01844 278746.

Microprocessors and controllers

Filter/codec. *MT9160* is a 5V filter/codec from Mitel that has handset transducer interfaces and programmable sidetone path, with digital gain control, anti-aliasing filters and reference and bias voltage sources. It supports both A-law and μ-law requirements and offers programmable CCITT (G.711)/sign-magnitude coding. There is a differential interface to handset transducers, including a 300Ω receiver driver. Register access is via a serial microport compatible with standard microcontrollers. Mitel Semiconductor. Tel., 01291 430000; fax, 01291 436389.

0.9μ PICs. Microchip has two new PIC risc cpus in a 0.9μnm process. *PIC16C58A* has 2048 by 12byte of program storage and 73 by 8byte of sram for data. With 12 i/o lines individually direction controlled and three timers, the device is fully compatible with other versions. It works on 2.5-5.5V and draws under 2mA. *PIC16C54A* replaces the C54 and is suitable for lithium battery power. Hawke Components Ltd. Tel., 01256 880800; fax, 01256 880325.

'Fastest' a-to-d. Claimed by its makers, Signal Processing Technology, to have the industry's highest sampling rate at 1GHz, the 8-bit *SPT7760* is a full parallel flash design giving a bandwidth of over 900MHz. Because of the wide bandwidth and input capacitance of only 15pF, external track/hold amplifiers are unnecessary in most applications. A proprietary decoding method reduces metastable error to under 1lsb of the Gray code output. Each of the 256 comparators has its own input preamplifier to act as a buffer and to stabilise the capacitance with input voltage range and with frequency. The device is thereby easier to drive than previous designs. All clock inputs, controls and data outputs are ECL-compatible. Signal Processing Technology. Tel. (USA), 001 719 2300; fax, 001 719 2370.

Mixed-signal ICs

Logic power interface. Devices in TI's *Power+Array* family drive three-phase, dc and stepper motors directly from 5V logic-level input. Current ratings are 0.75-1A/channel at up to 60V, and there are low-side and high-side drivers and bridges. Polar Electronics. Tel., 01525 377093; fax, 01525 378367.

PASSIVE

Passive components

Resistor networks. Thick-film resistor networks from BI Technologies are meant mainly for use in inrush limitation, snubber circuits and ups equipment. Four standard types, *BPC-3/5/7/10*, are rated at 3,5,7 and 10W, resistance range being 1Ω to 200kΩ at tolerances between ±0.5% and ±10%. Operating voltages are 300Vac and 500V dc and the networks can handle peak currents of twenty times rated current for 8ms. Temperature coefficient is 100ppm/°C. BI Technologies Ltd. Tel., 01384 442393; fax, 01384 440252.

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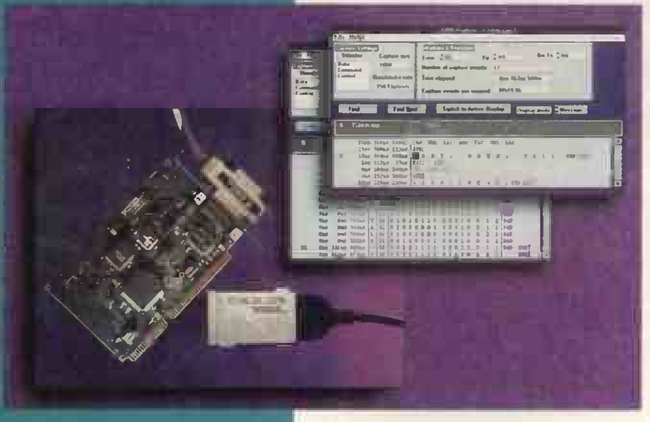


Encoder interfaces. *MBM Electrodynamics* is producing custom interfaces for its range of shaft encoders and scanning motors, thereby presenting the user with a 'ready-to-wear' output at a lower cost than usual. MBM Electrodynamics. Tel., 01273 413981; fax, 01273 425256.

Connectors and cabling

Board/board connectors. On a 2mm pitch, Robinson Nugent *Pak-2* board-to-board connectors have pins designed for through-hole or surface-mounting, vertical and right-angle, in 4-100 ways; optional flanges accommodate various mounting requirements. A variety of stacking possibilities is offered. Robinson Nugent (Europe) Ltd. Tel., 01256 842626; fax, 01256 842673.

Cut flat cable. Sumitomo flat cable is now available cut to length in 1mm and 0.5mm pitch in 100mm and 150mm lengths in short runs at the price normally associated with long production runs, with a next-day delivery. Stripping and reinforcement can be done to order and mating FFC connectors from either JAE or Molex fitted. Flint Distribution. Tel., 01530 510333; fax, 01530 510275.

2mm jumper socket. At the same height above board as a typical IC package, *3M*'s 2mm jumper socket for reconfiguring boards offers an alternative to dil switches and possesses the advantage of easy access for testing. 3M United Kingdom plc. Tel., 01344 858000; fax, 01344 858758.

Displays

Active-matrix module. Citizen has its first active-matrix video module, which uses a new two-mask TFT technique and a four-level drive. It is based on the use of higher-threshold pixels, which are thereby unaffected by stray signal from nearby lines or pixels. Additionally, the small size of the diodes allows a 480 by 230 resolution in the 3in diagonal module, with a contrast ratio of 200:1 and a 40ms response time. Citizen Europe Ltd. Tel., 01753 584111; fax, 01753 582442.

Production equipment

Electric screwdrivers. *Eliza* 20V Electric screwdrivers made by Atlas Copco have a patented mechanical clutch for enhanced accuracy. Screws from 1.2mm-4mm can be tightened to torques of 0.05-3.4Nm and the tools have a very light trigger and reverse switch. Accessories include pistol-grip attachments, vacuum pickup adaptors, a scribe and various insert bits. Hunter AP, Tel., 01934 876028; fax, 01934 876928.

Filters

Control-line filters. BLP Components's *SCF* range of rf filters are designed to protect lines controlling fire and intruder alarms, emergency lighting, elevator controls and the like. They conform to MIL-STD-220A and are suited to Tempest-rated equipment. Both ac and dc versions handle up to 2A and come with two, four or ten-line stand-alone form and 10-line modules for cabinet mounting. BLP Components Ltd. Tel., 01638 665161; fax, 01638 660718.

Hardware

Bobbins. *BFI IBEXSA* has a range of SMD, EFD, EP, RM and EE coil bobbins and bases for high-volume production. As well as custom designs, the bobbins are available in a low-profile range in vertical and horizontal form. Materials are Amoco LVP, Sumitomo LCP and phenolic, all UL approved. BFI IBEXSA Electronics Ltd. Tel., 01622 882467; fax, 01622 882469.

Instrumentation

ESD generator. A low-cost electrostatic discharge generator, the *pQT* from *picoQ* Corporation, tests equipment for immunity to esd and emi. It produces a controlled air discharge through direct contact, needing no ground wire, power source, internal contact-mode relay or polarity switch. E-field and H-field probes are available and the instrument is calibrated to a broadband or human-body model. Lyons Instruments Ltd. Tel., 01992 768888; fax, 01992 788000.

Tv spectrum analyser. Mainly for television work, the *U4341* spectrum analyser in the R&S Advantest range is light and portable and yet highly sensitive, having a built-in demodulator to allow programmes to be shown on the tft screen. Level and frequency measurements may be made, selection being by stored channel frequency, which can be edited for more channels at up to 2.2GHz. Two slots take PCMCIA cards to store data and settings. All national television standards are supported. Rohde & Schwarz UK Ltd. Tel., 01252 811377; fax, 01252 811447.

Electrostatic fieldmeter. *Hand-E-Stat* is an instrument made by the Dutch company Simco to measure the potentials associated with static charges. It is a pocket-sized instrument, offering sample and hold modes, with a light-ranging method of achieving repeatable results. Its case is conductive and there is a ground clip. Autoranging covers ±20kV at a distance of 2.5cm and, as well as the digital display, an analogue output is provided. Accurate measurements can be made in areas using air ionisation. Simco (Nederland) BV. Tel., 0031 5730 88333; fax, 0031 5730 57319.

Energy conservation. Schuter's power management module switches off a pc's monitor when it is idle, bringing it back to life if the mouse moves or someone touches the keyboard. It is a snap-in mounting inlet/outlet plug and socket with the necessary control circuitry and is meant to offer energy saving when large numbers of pcs in an organisation are normally left running all day, being used infrequently. Radiatron Components Ltd. Tel., 01784 439393; fax, 01784 477333.

Lcd panel meter. Needing no external power supply, bypass capacitors or trimmers, the *DMS-30LCD-4/20S* panel meter by Datel is powered by a 4-20mA current loop and measures 2.2in by 0.93in, with a 0.4in high display. An eight-position dip switch selects range, offset and decimal point placing, no jumpers being needed. Datel (UK) Ltd. Tel., 01256 880444; fax, 01256 880706.

Interfaces

GPIB controller/analyser. Two interface boards from National turn existing National interface boards, the AT-GPIB/TNT and the PCMCIA-GPIB card, into GPIB analysers, controlling GPIB instruments as well as sorting-out difficulties such as addressing inconsistency, protocol violation and others. *AT-GPIB/TNT+* and *PCMCIA-GPIB+* also capture HS488 activity. National Instruments UK. Tel., 01635 572400; fax, 01635 523154.

Infrared data link. Conforming to the Infrared Data Association (IrDA) serial infrared comms standard, H-P's

HSDL-1000 provides a link effective up to 1m for communication between mobile computers and devices. It works at 870nm and 115.2kb/s over ±15°, the package consisting of a led and detector, led driver, photodiode transimpedance amplifier, comparator and bias network. Communication is effective in 10klux of sunlight and up to 1klux of incandescent or fluorescent lighting. Hewlett-Packard Ltd. Tel., 01344 366666; fax, 01344 362269.

Sample-rate converter. Analog's *AD1893 SamplePort* is a 16-bit stereo asynchronous sample-rate converter that accepts a digital input sample stream at an arbitrary or varying rate and outputs the stream at the user's chosen rate, thereby also minimising the effect of clock jitter. It is based on the *AD1890*, but also offers an oscillator, single 3-5V operation, a power-down mode and right-justified serial interface format. Thd+n is 96dB minimum, as the dynamic range with a 0-20kHz, –60dB input. Analog Devices Ltd. Tel., 01932 266000; fax, 01932 247401.

Literature

Power mosfets. Several kilograms of paper in the form of International Rectifier's *Hexfet* designer's manual is available free. In 2500 pages, the book contains product data and application notes on the use of Hexfet power mosfets. Solid State Supplies Ltd. Tel., 01892 836836; fax, 01892 837837.

Keyboard. *Model G81-3100* keyboard by Cherry is described in a new brochure. This unit is complete with a magnetic card reader, which will read all the data on all popular cards, and/or a bar decoder that is compatible with common reading devices. Cherry Electrical Products Ltd. Tel., 01582 763100; fax, 01582 768883.

Materials

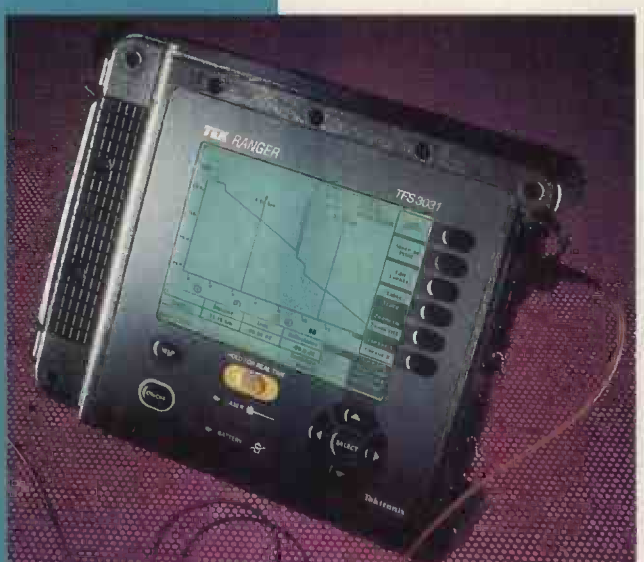
Stickier sticky labels. New from 3M is a pressure-sensitive labelling material that sticks to virtually anything, including the 'difficult' plastics and powder-coated finish. Both *ScotchMark 7871/2* have a polyester face, 7871 in gloss and 7872 in matt platinum, and both are printable by thermal transfer. 3M United Kingdom plc. Tel., 01344 858000; fax, 01344 858758.

Navigation systems

Range of GPS. *Navstar Systems* has a range of global positioning system equipment, being the only British maker of course acquisition GPS receivers with particular experience in professional and defence work. Telecom Design Communications announces that it has concluded an alliance with Navstar. The range includes 2m-resolution, 12-channel differential base stations, six-channel receivers, post-processing software and remote surveying/logging software. Telecom Design Communications Ltd. Tel., 01256 332800; fax, 01256 332810.

Small, low-power GPS receiver. Power-saving facilities in the Rockwell *MicroTracker LP* five-channel receiver engine will, according to TDC, double battery life. More complete integration has reduced component count from earlier Rockwell designs, but compatibility is retained. The board measures about 2in by 3in and delivers differential GPS accuracy to within 5m most of the time, with a time to first fix of 30s from a warm start. *CityTracker* software facilitates use on the move in cities with steadier tracking than before. Although the receiver is optimised for static work, it is able to work at altitudes up to 40,000ft under 4g acceleration. Telecom Design Communications Ltd. Tel., 01256 332800; fax 01256 332810.

Mini-otdr. Tektronix's *TekRanger TFS3031* is miniature optical time-domain reflectometer capable of both single-mode and multi-mode testing of optical-fibre cable in the one instrument. A feature to recommend the instrument for field work is *IntelliTrace*, which adjusts parameters such as pulse width and number of averages as the instrument automatically analyses a fibre link, thereby measuring small and closely spaced events accurately even over long distances. Since the dead zone is smaller than in previous instruments, events down to 5m and up to 100m are testable, results being displayed on a large, backlit lc display. Up to 200 waveforms may be stored and recalled for comparison or analysis and an optional hard disk can be used for archiving and documentation. A power management system is included, battery life being eight hours. Tektronix UK Ltd. Tel., 01628 486000; fax, 01628 474799.



Power supplies

Multi-output smps. A range of very low-cost convection-cooled power units by Coutant Lambda, the *SW series* includes 30W, 40W, 65W and 100W models having one to four outputs from 5V to 24V. The units meet UL, CSA and IEC requirements. Coutant Lambda Ltd. Tel., 01271 865656; fax, 01271 864894.

Small dc-to-dc converters. From Datel, the *XWR* series of 5W converters are in 1in by 1in by 0.45in metal cases and conform to the pinout used by much larger types. They are designed for use in systems with distributed supplies, where the converter can be placed at the load to reduce losses, poor regulation and slow transient response. Outputs are 5V, 12V and 15V unipolar or bipolar and the inputs handle 18-36V or 36-72V. Protection is provided. Datel (UK) Ltd. Tel., 01256 880444; fax, 01256 880706.

Safe 20MHz oscilloscope. *OX801*, a 20MHz analogue instrument by Metrix, has double insulation to IEC 1010-1 Class 2. While offering the usual facilities of an oscilloscope at this level, it also possesses a microprocessor bus system which, among other functions, removes the need for high-frequency signals to be near the front panel. More unusually, the bus has allowed automatic mode indication and remote control and also linking to a computer with software to simulate the oscilloscope screen on that of the computer. Controls are analogue in feel, though digital in operation. Metrix UK Ltd. Tel., 01256 311877; fax, 01256 23659.

Apd supplies. Gardners has a new type of dc-to-dc converter, the *GR84907*, designed to provide variable high-voltage supplies for avalanche photo diodes used in wide-band digital optical communication receivers, as an alternative to the use of fixed supplies and high-value resistors. Output (20V-100Vdc at 0-1.6mA) is isolated and contains only 50mV ripple and noise, voltage being regulated by a control voltage. There is no internal stabilising feedback, since the unit is designed for system use with an external loop, although closed-loop models can be supplied. Gardners Ltd. Tel., 01202 482284; fax, 01202 470805.

UPS batteries. *SmartCell XR* battery packs for Vero's 3kVA and 5kVA *Matrix* ups range reduce cost of ownership, since they replace four of the earlier modules, one *XR* unit giving 45 minutes hold-up time at 3kVA at only 60% of the cost. The units are in 11U high 19in rack modules. Computer monitoring in each SmartCell tests its condition and indicates the need for replacement to the front panel and to the monitor software, also providing a log of periodic checks made by the processor and an accurate indication of run time when the battery is in use. Vero Electronics Ltd. Tel., 01703 266300; fax, 01703 265126.

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Production test equipment

Transformer tester. Voltech's *AT3600* is a comprehensive automatic tester for transformers and other wound components. It is fully programmable to allow an all-in-one test, providing all the functions that normally require up to seven individual instruments, all results being presented either in a single test report or as a simple pass/fail indication. The measuring process is of a closed-loop nature for accuracy, giving a basic 0.2% accuracy on winding resistance from 1mΩ to 100kΩ, turns ratio 10^4-10^{-4} and inductance 100nH-100kH. Windows-based software is supplied to carry out 21 tests, with a further 16 to come later in the year, for which the software is to be supplied free. The user simply enters a schematic diagram of the transformer, creates a 'virtual fixture' by connecting its pins to any of the unit's nodes and selects the tests to be run from menus. Fixtures can be made from kits or built by Voltech. Voltech Instruments Ltd. tel., 01235 861173; fax, 01235 861174.

Pcb test. *CITS200* is a bench-top printed board tester by Polar which verifies the characteristic impedance of tracks, including balanced-line connections using time-domain reflectometry to single-ended or differential measurements. A graphical user interface makes for easy operation and flexibility. A standard pc is linked to a two-channel test system that automates all impedance tests, set-up being carried out by a Windows-based package. In operation, the test file is selected by the operator, who positions the probe and presses a switch. Next the tests are carried out in succession with prompts to the operator to reposition the probe as needed. Results are processed to give a graphical view of impedance, printed and saved to disk. Polar Instruments Ltd. Tel., 01481 53081; fax, 01481 52476.

Switches and relays

3GHz miniature relay. Teledyne's *RF300/303* series of ultraminiature relays provide ±0.1dB rf signal repeatability over the 0-3GHz frequency range. They are meant for low-level rf and dc switching in attenuators, tuning, filters, etc., in which the features of wide bandwidth, good shielding and grounding and high control-to-signal isolation are required. These relays are in TO-5 cans. Teledyne Electronic Technologies. Tel., 0181-571 9596; fax, 0181-571 9637.

Resettable cut-outs. Bi-metallic, resettable cut-outs by Steatite protect against over-temperatures. Ratings are 4A, 6.3A and 10A, the 4A type being BEAB-approved to CO652 and particularly suitable for use in wound components. Contacts are normally closed and are rated at 250Vac at 50Hz. Accuracy of the cut-off is within ±5°C. Steatite Power Ltd. Tel., 0181-778 6611; fax, 0181-778 7722.

Timer switch. Amerace announces the 48mm DIN *Agastat Electronic Timer*, designed for a standard 48mm panel cut-out and to be rail-socket or panel mounted. It has up to eight user-selected timing modes, six timing ranges from 0.1s to 10h at a repeat accuracy of ±0.5%±10ms. All models have 10A dpdt contacts. Amerace Ltd. Tel., 01635 49191; fax, 01635 521641.

Reed switch. Switching 250Vac at 1A, Gentech's *GR19* miniature reed switch has Dimet contacts to give it a life of 100,000 operations. It is available with sensitivities of 20-50AT, an initial contact resistance of 20mΩ, insulation resistance of 10^{12}Ω and a withstand of 700Vdc. Under extreme conditions, the GR19 handles 5A at 20Vdc for a short time. Gentech International Ltd. Tel., 01465 713581; fax, 01465 714974.

Thin, high-voltage relays. Matsushita's *photo-MOS* optically coupled relays control a 400V, 130mA load from a 5mA led input, but are only 2mm high off the board. Two models of this type, the *AQV212S* and *AQV214S* have on resistances of 0.83Ω and 30Ω respectively with leakage current of 100pA at 400V, the led drawing 5mA and dropping 1.14V. The 212 is for load voltages to 60V at 0.35A, the 214 for 400V at 0.1A, both being normally open. In a larger package standing 3.4mm high, the *AQV212A* and *AQV214A* are identical components. Model *AQV614A* has one normally open and one normally closed contact in the 3.4mm package. Flint Distribution. Tel., 01530 510333; fax, 01530 510275.

Rotary switches. New features on the *Grayhill* range of rotary switches are a spring return facility for applications such as 'hold-to-test', and isolated positions in which the shaft must be pulled before rotation to provide a standby or emergency position. To maintain contact in the 'hold' positions, a rotary force must be held, the switch returning to its normal detent position when the force is released. Models having up to six decks are available, with up to six poles per deck. Ratings are 200mA-1A. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Signal re-routeing on a pcb. Lattice Semiconductor's *ispGDS* is a programmable digital switch IC that allows signals to be re-routed on a pc board under software control, eliminating jumpers, dil switches and added bits of wire. Control is by a four-wire, 5V serial interface. Micro Call Ltd. Tel., 01844 261939; fax, 01844 261678.

Transducers and sensors

Optical-fibre sensor. From Honeywell comes the *HPX* series of sensors, which the company claims is in the slimmest (10 by 28 by 60mm) package available. There are three models for use in paper and object sensing, in machinery to detect jams and to count, in electronic gear to count small components and in pharmaceuticals to count pills. Units have a three-turn pot. to adjust sensitivity and an alarm is given if the device needs cleaning. Honeywell Ltd. Tel., 01344 826000; fax, 01344 826240.

Custom load cells. *CorinTech* can now supply thick-film load cells to customers' specification, using techniques developed at the University of Southampton. Piezoresistive thick-film resistors on one-piece metal substrates provide robustness and low cost and the tri-beam design exhibits linear and repeatable output and invulnerability to eccentric loads. CorinTech Ltd. Tel., 01425 655655; fax, 01425 652756.

Displacement controller. *D1* displacement indicators by Control Transducers connect directly to linear sensors to act as a 3.5-digit readout and to pass signal to a pc. Transducers may be rectilinear potentiometers, amplified sensors or current-loop transmitters and the *D1* provides the sensor's power supply. Readout is supplied in the required units, with specified set-points and type of signal output. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

Liquid gas level. Control Transducers's *NGTT* gas level sensor reduces the problems of adapting natural gas to power machines by providing an accurate method of determining the level of liquid in the tank. A constant-resistance pressure gauge is also used to measure temperature, both quantities being needed to calculate level. Output is supplied directly to an analogue or digital voltmeter. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

Vision systems

Digital video encoder. IBM has announced the first chip to compress or encode the data to transmit and store digital video, frame by frame. *MPEG-2 I-Frame Encoder*, the 'I-Frame' referring to the intermittent frame used in high-speed compression and decompression, and an enhanced version of the *MPEG-2 Decoder* for the pc and television consumer market. Blue Micro Electronics. Tel., 01604 603310; fax, 01604 603320.

Video disk recorder. H-P's *HP 4:2:2 VDR* now has double its previous capacity – 24 minutes. The instrument is for workstation use in computer graphics and animation; the new specification allows a half-hour tv programme – the bit between the ads – to be recorded to a hard-disk array for playback, with digital audio. There is a full set of Ethernet commands for file transfer and a SCSI set. The unit is supported by major animation and graphics developers. Hewlett-Packard Ltd. Tel., 01344 366666; fax, 01344 362269.

Windows data acquisition. *Windspeed* software from Windmill provides notebook computers with the facility to log and display data. The package runs under Windows, continuously logging data to disk and charting it in real time so that users can remove and save parts of the record in a choice of formats. Since notebooks do not often possess expansion slots, *Windspeed* uses Microlink data acquisition hardware that connect to a COM port on the computer or to a PCMCIA network adaptor. The system will stream analogue data to a hard disk at 700sample/s via the COM port or at 36,000 sample/s through Ethernet. Hardware is configured from the software by means of set-up files from a library saved by the user. Different channels may be sampled at different intervals within one scan or some channels may be sampled every scan and others every *n*th scan. Windmill Software Ltd. Tel., 0161-833 2782; fax, 0161-833 2190.

COMPUTER

Computer board-level products

PC-104 modules. Pro-Active Control has a range of microprocessor mini-modules based on the *PC-104* for the embedded-systems market. The units run dos software and only need dos and interfaces to work. Top of the range is *PAC-486*, based on a Cyrix *Cx486SLC2* 50MHz processor and including 2Mbyte of dram, with a 4Mbyte option, battery-backed sram up to 512K, clock/timer, serial and parallel ports, keyboard i/o, disk support and power management. Secondly, there is the *PAC-86*, which uses the Chips & Technologies *8680A* pc chip running at 14MHz. This is an 8-bit PC XT computer with full i/o facilities and PCMCIA support, full CGA graphics with crt/lcd drive, 1Mbyte mapped dram, 512Kbyte sram, 512Kbyte eprom, an RS-232 port and a PCMCIA slot.
The third module is the *PAC V-25*, using the NEC V-25 chip to make one of the lowest-cost pc-compatible processor modules available, optimised for embedded applications. Pro-Active Control Ltd. Tel., 01223 300801; fax, 01223 300979.

Optical bus expansion. *PCX-797* is an expansion kit by Fairchild to allow additional ISAbus slots to be linked to a pc using an optical-fibre link. The kit has two cards: a master for the pc and a slave for the remotes passive backplane, with up to 100m of optical fibre between the two. Multiple cards can be used to obtain hundreds of i/o slots for a standard pc. Data throughput is 1.6Mb/s and the 96Mb/s optical transmission avoids bus timing problems. Fairchild Ltd. Tel., 01703 559090; fax, 01703 5559100.

68030 VME card. Syntel's new 68030-based card, the *SYN-VME203*, is a 3U single-board computer for the VMEbus offering good power economy. Facilities on board include a 68882 float co-processor, 1Mbyte or 4Mbyte of 32-bit-wide dram and up to 1Mbyte of eprom. Two RS232 asynchronous serial channels are available with one 16-bit interrupt timer and there is a watchdog timer. The dram subsystem supports fast 16-byte cache burst filling and cacheing is included for all VMEbus operations. Both 8-bit and 16-bit VMEbus transfers are carried out. Auto bus sizing converts 24-bit and 32-bit cpu operations to consecutive 8-bit and 16-bit transfers. Syntel Microsystems. Tel., 01484 535101/2/3; fax, 01484 519363.

Data acquisition

PC measurement modules. Kyle Data Services offers the *MM-232* family of data acquisition modules for computers with RS-232 ports. Units for the measurement of voltage, frequency, digital signals and events are currently available, each with four input channels, two further digital inputs and two digital outputs. Special software is not need and all modules are easily accessible from Basic, Pascal, C and other languages. Modules either plug directly to a 25-pin port or may be remote; they are powered either by the RS-232 line or by external supply. Kyle Data Service Ltd. Tel., 01292 311169; fax, 01292 318005.

Software

Signature analysis. Since 'compatible' ics from different manufacturers can possess marginally differing signature analyses, the technique sometimes produces invalid results. Polar Instruments can now offer *AVR*, which is Advanced Vendor Recognition, to detect the slight differences automatically, avoiding unnecessary fault repair. Polar Instruments Ltd. Tel., 01481 53081; fax, 01481 52476.

Application generator. *Visual Designer* application generator software from Intelligent Instrumentation now supports the Keithley *DAS 16* family of boards, giving access to any analogue or digital i/o functions from the boards including dma, counter and rate-generator functions. The software is Windows-based and allows people to develop application software by drawing block diagrams, with no programming. Virtual sliders, switches and instrument panels appear on screen and control the application. Intelligent Instrumentation. Tel., 01923 896989; fax 01923 896671. ■

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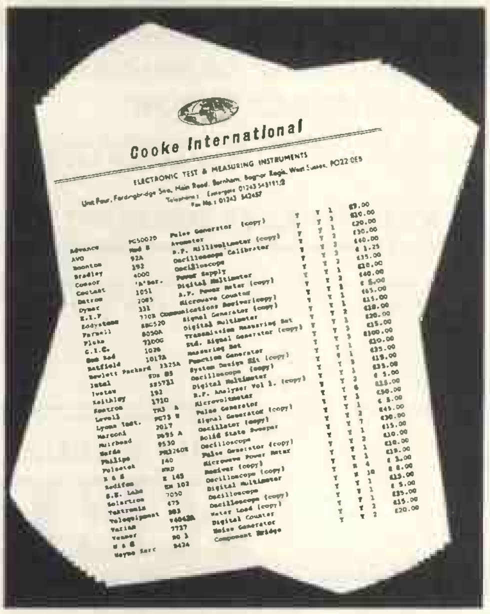

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