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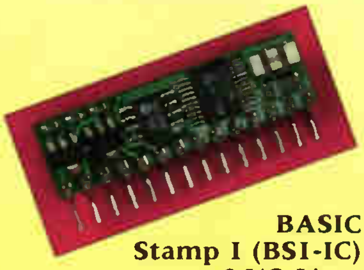

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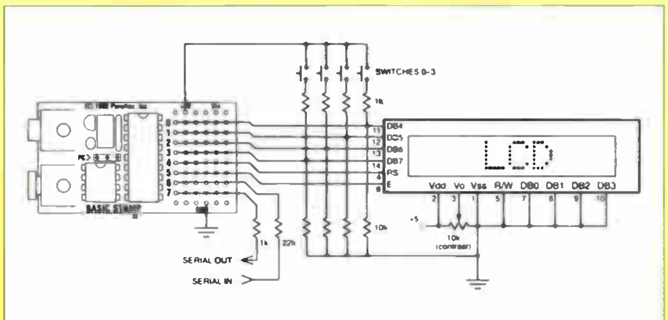

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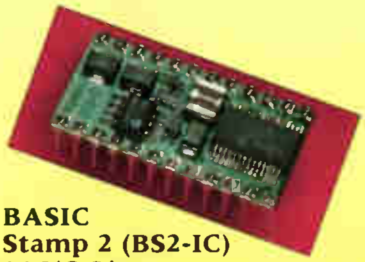

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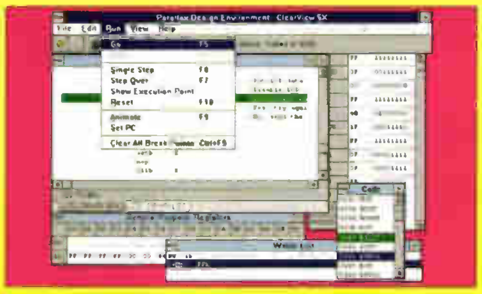

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CIRCLE NO. 101 ON REPLY CARD

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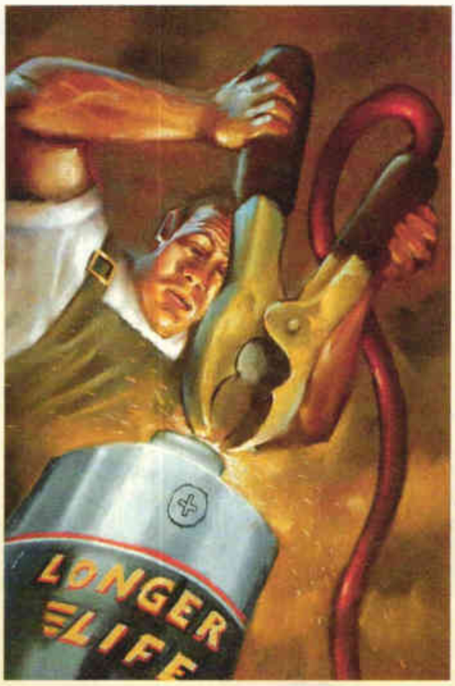

Cover – Hashim Akib

Regulars

Using a spherical mirror, this camera can see in all directions at once – *see page 186.*

New packaging technique means that two-component epoxies can be prepared and applied without mess and mixing proportion errors – *see page 215.*

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REED BUSINESS INFORMATION

Electronics and social change

At one time the UK electronics industry used to perk up with special interest when a new government was expected. Governments went in for intervention: awarding contracts, supporting ailing firms, running national R&D organisations, the Inmos and IT initiatives, the electronics Neddy of the NEDC, and so on. Most of this has been thrown out during the past 18 years, and any Labour government would be unlikely to restore it. But, regardless of the shrinkage of the UK's industry, every new Parliament still has to work within an environment in which worldwide electronics technology brings about changes in the lives and minds of its citizens.

If a Labour government came in, you might expect an administration with a leaning towards socialism to start off by declaring its desired social values – since socialism is based on a moral principle – and then set about finding the practical means for their realisation. This is unlikely to happen. Given the economic constraints, Labour would simply continue to manage the prevailing system of welfare capitalism. But in fact public opinion, in reaction to events, is demonstrating again and again that fundamental human values are the main driving force behind our feelings of how the country should be governed.

There is disgust, for example, at the widening inequality that results from the internal dynamic of market capitalism. From being the original drive behind the labour movement in the 19th century, it is now recognised as a contributor to wider social ills. But growing numbers of people – especially the young – feel excluded from the Westminster political system, which they see as irrelevant to their main concerns. Thus we have the green movement and the women's movement – both outside the traditional arena of Left-Right confrontation – followed by a whole host of causes like human rights, abortion, *in-vitro* fertilisation, animal protection and penal reform. All these are pursued on ethical or philosophical grounds but start from gut feelings.

The intrinsic human values – truth, beauty, virtue, justice, charity, liberty etc. – have little to do with technology. But some of the instrumental values – those that work as means to an end – certainly do play a part. This is because technology is not an autonomous, one-way process, but develops through interaction with human beings. We modify the technology through what we choose to accept or reject, by our responses to whatever products or systems are on offer.

In electronic communications, for example, systems like mobile 'phones, the Internet and personal entertainment equipment tend to reinforce the instrumental values of flexibility and mobility, above the traditional ones of security and stability. This interaction is broadening what the motor car and the telephone started.

The physical distance between communicating individuals brings with it a behavioural and mental distancing. Automation and computing have a similar psychological effect on commercial and manufacturing workers. The work itself tends to become more abstract and intellectualised. It is done largely through the medium of symbols on vdu screens and computer print-outs. Direct sensory experience through physical contact with objects is being lost. Traditional, authoritarian manager-worker relationships are changing. Such automated work

In electronic communications, systems like mobile phones, the Internet and personal entertainment equipment tend to reinforce the instrumental values of flexibility and mobility...

strengthens our respect, already disciplined by the clock, for the values of order, uniformity, precision, calculability, efficiency.

Modern medical technology extends our control over life. Here electronics has become increasingly powerful through diagnostic and research techniques such as computerised tomography, magnetic resonance imaging, positron emission tomography, magnetoencephalography and DNA sequence analysis – in addition to the established radiological, electrophysiological and electro-acoustic methods. Patients routinely talk about having 'scans'. As a result of these and other advances, we are beginning to look at suffering and misfortune almost as technical problems which, *ergo*, can be fixed. Such perceptions tend to wear away our sense of the uniqueness of human life – perhaps justifiably.

Some of these instrumental values could become very strong in our society through the positive feedback of cultural reinforcement. The spread of personal computers and portable phones, for example, is much more the result of imitation and fashion than of sheer practical need. Values influence public opinion, and in a democracy this should theoretically proceed in an orderly, linear fashion into political decision making and legislation. In practice the technology-accelerated values may well sweep past the Westminster lawmaking machine, which will then have to take hasty, ill-considered measures in order to catch up with the resulting human behaviour. ■

Tom Ivall

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Doubts on digital recording quality

Digitally mastered recording leads to perfect replication of sound. This accepted premise is being questioned by a Cambridge-based specialist recording company which is investigating detected differences in cd sound quality.

Prism Sound is conducting a comprehensive research project into the audio differences between different cd pressings which are made from the same material and have identical data, but which are manufactured using a variety of methods.

Ian Dennis, Prism's technical director, said that when he first heard of the problem he was, "very sceptical about the whole thing". But Prism, along with several compact discs pressing plants, produced 14 test cds, including tracks from Pink Floyd and Mariah Carey, all manufactured differently in order to validate the phenomenon.

Preliminary results of listening tests, using sound recording engineers, confirmed that there is indeed a problem. "It's something that offends the ears of people in the industry rather than Joe Public," said Dennis, who wants to extend the tests to include lay people.

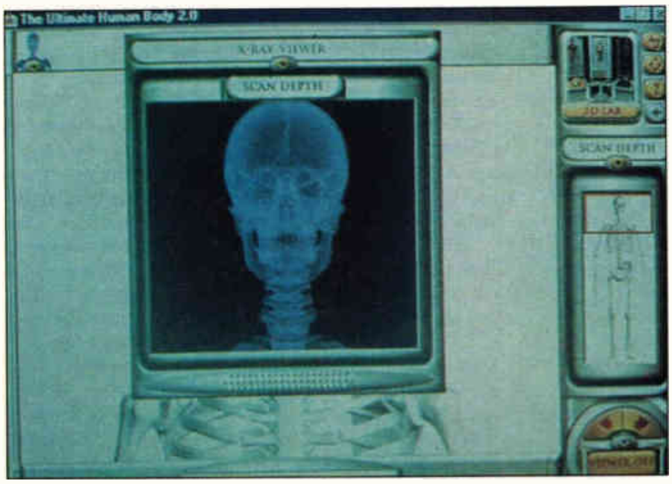
Dennis continued: "It seems that certain cd players are affected by the mechanical properties of the disc."

The differences in audio quality could be due to geometric variations between discs affecting a cd player's servo mechanism which, in turn, affects currents used in the mechanism's motors to cause crosstalk, and hence audible products, or cause jitter in the digital audio output.

Intel releases details of Pentium successors

Intel revealed further details of its x86 roadmap during the launch of its latest MMX Pentium processor, the *P55C*.

The Klamath, the Pentium Pro (P6)-based device, is to be unveiled in the second quarter of the year. Klamath is aimed at high-end desktop and low-end server systems. The Pentium Pro itself, with faster caches, will continue to be used in high-end servers.



Chris Hogg, market development manager for Northern Europe, said that Klamath will first be released at 233MHz, "and other speeds". He would not confirm whether 266MHz devices would be available, preferring to leave some mystery for the launch.

Competitors to Klamath are AMD's K6 and Cyrix's M2. Both devices have multimedia instruction sets (MMX) so as to compete with Klamath. However, both use a Pentium-style bus interface, unlike the Klamath, which has a P6 interface.

To pressure the cloners, Intel plans to exploit its leading process technology. "We're bringing up 0.25μm so we have the capability to go faster," said Hogg. This will allow for the Deschutes chip, a 300MHz Klamath, and also a faster Pentium Pro.

However, Linley Gwennap, editor of *Microprocessor Report*, said: "I'm not sure Intel will increase the speed of Pentium Pro, relying instead on Klamath."

Gwennap believes that, on the desktop, Klamath and Deschutes will hold off AMD and Cyrix.

The move to 0.25μm will also bring faster Pentiums. "Faster P55C parts will probably be only for notebooks," said Gwennap.

A future processor, dubbed Katmai, could enhance the MMX instruction set extensions. MMX2, believes Gwennap, "adds paired floating-point operations, motion estimation, and maybe fixes some of the silly restrictions of MMX1."

Free patent searches on IBM Website

Want to find out if that hot idea of yours is original? IBM is now offering a Web-based patent searching service on its IBM Patent Server site with details of more than 2 million US patents issued over the past 26 years.

Users can search by title, inventor, assignee, abstract, claims, attorney/agent and patent references. In addition, there are a million full images of US patents issued since 1987, with additional images added on a continual basis going back to 1974.

IBM says that the service is part of a capability it developed for its own uses to search US patents and decided it would make a valuable public resource.

The good news is that the service is free to anyone with Internet access. The site is at: **http://www.ibm.com/patents**

GSM navigation without GPS

A Cambridge University researcher has designed a system which allows a GSM handset's position to be calculated without requiring the use of the Global Positioning System (GPS).

Dr Peter Duffett-Smith, inspired by his work in radio-astronomy, has designed a base station which calculates the handset's position within an accuracy of 50m. The technology is based on interferometry, with the positioning of an object determined through multiple readings taken from various points.

The ability to locate handsets will enable numerous additional GSM services to be added. For example, it would enable emergency services to locate broken-down cars or accident sites, allow personal navigation, and even pinpoint a stolen phone or car.

Duffett-Smith believes the technology, dubbed Cursor, could be adopted as early as this year, and he is currently in discussion with major handset and base station manufacturers.

"There is a need for a system like that [location/navigation], and there is definitely a market for it," said one Ericsson spokesman.

The main approach to combining location/navigation technology with GSM is GPS. "GPS is the Rolls-Royce of technology," said Richard Fry from the Technology Partnership. "The GPS infrastructure is already there. The problem with Cursor is that people need to put the infrastructure in and it has to be low cost."

But GPS has four disadvantages when compared with Cursor: it is expensive to fit into handsets, with a current accuracy of 100m it is not accurate enough for urban environments, it is power-hungry, and it takes a while to position itself as it consults geostationary satellites.

With Cursor, the only modification needed in the handset is applied to the software.

Cambridge Positioning Systems, with Duffett-Smith as its technical director, has been set up to exploit the technology's commercial potential. In addition to Cursor, several other similar technologies are currently under development.

Svetlana Josifovska

Employers must say YES to skills training

British engineering employers must continue to invest in skills among the engineering workforce, according to Dr Mary Harris, director general of the Year of Engineering Success (YES).

"It's very important to realise that there is a shortage of engineers now, and that industry has to train skills and continually refresh and update these skills," said Harris.

She made it clear that the main objective of YES, which launched on 22 January, will be to get across to industry, government and opinion formers just how important engineers are to the future economy of the country.

"Engineering supplies about 30% of our GDP, with inward investment supplying about 18%. Companies like Motorola, Fujitsu and Sony come to this country primarily because of our engineering expertise," she said.

Harris pointed out that "we need about 35,000 new engineers every year just to replace those that are leaving, and this year we're only getting about half that number. It's a real problem."

She revealed that a number of YES events that will be taking place in the coming year to "demystify the role of the engineer" to the general pubic.

These will include the BBC's *Tomorrow's World* Live exhibition at the NEC (19-23 March), organised in collaboration with YES, and a national tv and radio campaign.

"We need about 35,000 new engineers each year... and we are only getting about half that number," said Dr Mary Harris, Director General of YES.

US non-profit bodies to get free radio allocations

The US Federal Communications Commission (FCC) has approved the allocation of part of the radio spectrum to enable schools, hospitals and other organisations to use fast, wireless communications technologies without being compelled to paying high prices for radio spectrum use.

The spectrum allocation falls into the 5.150 to 5.350GHz and 5.725 to 5.825GHz ranges, and will allow non-profit and commercial organisations to exploit wireless communications and thereby avoid the potentially high cost of wire installation in buildings.

The wireless devices will have a range as wide as three miles but supporters of the plan, which included Apple Computer, Lucent Technologies, Motorola and Northern Telecom, were disappointed that the FCC did not approve wider wireless links capable of extending as far as 13 miles.

The wider range would have enabled entire communities to share high-speed wireless data links.

The FCC faced opposition to the plan from companies that had paid billions of dollars for radio spectrum licences for Personal Communications Services.

The FCC says that the new wireless technologies will be best suited for rural areas or campus-type environments rather than in built-up areas. This is because the radio channels are adversely affected by buildings at such frequencies.

TI gambles on changed dsp focus

Texas Instruments is to end design and development of its *TMS320C80* and *C82* digital signal processors (dsps), used extensively in videoconferencing and compression.

A spokesperson for the company said: "TI believes it needs to have a more long-term look at visual communication. It is refocussing on ISDN, XDSL and cable modems."

The C80 and C82 will be succeeded, the spokesperson said, by faster, more broadly focused products. Manufacture of the dsps will continue.

The lack of future development is likely to upset the many companies which have designed products based on the devices.

Unless TI brings out a new dsp architecture for video compression quickly, such firms could switch to other vendors' dsps which will have future enhancements, such as Analog Devices' *Sharc* processor.

Fractal antennas "in use within a year"

A US company is predicting that fractal antennas will be used in a commercial product this year.

Nathan Cohen, technical adviser to Fractal Antenna Systems, the company concerned, said: "You will see fractal antennas in transceivers in less than a year." While reluctant to discuss what kind of transceiver, he said it would be "a cell phone or a cordless phone".

A fractal antenna is effectively a track pattern on a pcb which is used as an antenna. The form of the pattern is determined using fractal maths.

Fractal Antenna Systems makes some significant claims for its designs. Its Fractal Micropatch is said to "incorporate the benefits of conventional patch antennas, but at a fraction of the size".

So what is magic about fractal antennae compared with deterministic designs? Cohen says: "There is no magic, we select the fractal equation genetically." He added: "I have yet to see a non-fractal shape that does what our antennas do."

Cohen cited a research project into genetically, non-fractally generated antennae that produced a fractal result. "After all," he said, "fractal patterns are a significant set of all possible patterns."

● A Spanish team has been researching a flat, fractal-tree-like monopole fractal antenna. The work, reported in *Electronics Letters*, describes a broad- or multi-band characteristic, with the distribution of frequencies relating to the distribution of segment lengths over the antenna.

Italians triumph with 0.8V bipolar op-amp

Engineers in Italy have developed a rail-to-rail output, precision input, bipolar op-amp operating with a supply of only 0.8V.

Designing the circuit with bipolar transistors brings with it the problem that the V_{BE} of a bipolar transistor is around 0.7V. This leaves only 100mV for other circuit elements to operate within.

Final measured characteristics, in a sample fabricated by SGS-Thomson Microelectronics, show an input offset of 0.4mV, current consumption of 0.75mA and an output swing of 90% of V_{cc}. The input common mode range, governed by the p-n-p input transistors, is –0.3V to Vcc–0.7V.

Output stage employs a complementary common emitter pair, rather than the usual emitter follower design. This type of output is difficult to drive effectively and doing so requires a lot of transistors. The design uses at least double the usual number.

Giuseppe Ferri, one of the co-designers, is no stranger to low-voltage design. He presented a paper at the 1996 ISSCC describing a 1.3V op-amp in conventional c-mos with rail-to-rail operation at both input and output.

Samsung joins FRAM fray

Samsung is the fifth major semiconductor manufacturer to take a licence for ferroelectric ram (fram) technology from Ramtron of Colorado. The other four are Hitachi, Toshiba, Rohm, and Fujitsu.

Frams are just appearing on the commercial market. Rohm of Japan started making 1000 6in fram wafers a month in September. The company makes 4kbit, 16kbit and 64kbit density devices.

Hitachi has started sampling 256kbit frams.

Toshiba, Fujitsu and NEC are expected to have 1Mbit frams on the market later this year. Both NEC and Hitachi showed 1Mbit frams at the 1996 International Solid State Circuits Conference.

According to Hitachi Semiconductor's executive managing director, Dr Tsugio Makimoto, it will be five years before frams are cost-competitive with drams.

Before then, they will be widely used in portable equipment where their non-volatility and low power makes them ideal memory storage devices.

One key benefit of fram is its endurance. Ramtron spinoff Symetrix has demonstrated frams with 10 trillion read/write cycles. According to Hitachi's Makimoto, that will also be the endurance of Hitachi's commercial frams.

Ferroelectric technology has been pursued for many years without achieving commercial viability. The problem was the quality of the ferroelectric material available. With the development of a new material, layered perovskites, commercial production has become possible.

After wireless IC cards, the next big application for frams, says Makimoto, is in pocket organisers.

Breakthrough in tv-quality video over the Internet

WebTV Networks, the Silicon Valley-based company, says it has achieved a breakthrough in transmitting full-screen, television-quality video over the Internet using standard modems. Called *VideoFlash*, the technology will be incorporated into the company's set-top tv boxes which add Internet browsing features to regular televisions.

The company is unwilling to discuss details of its technology because of its patent applications, but said that it offers a data compression ratio ten times higher than the industry standard MPEG-1.

VideoFlash was demonstrated at the Consumer Electronics Show in Las Vegas recently and is based on a software algorithm rather than dedicated hardware.

"In 1996, WebTV brought Internet capability to the television," said Steve Perlman, co-founder, and CEO of WebTV. "In 1997, with *VideoFlash* technology, WebTV will bring television capability to the Internet." The WebTV boxes are being manufactured and sold in the US by Sony and Philips.

While pre-Christmas sales were low, the company says that sales have been in line with expectations and will grow in 1997. Customers have to purchase a subscription to WebTV's Internet service. All will receive *VideoFlash* automatically downloaded as regular software

The developers of *VideoFlash* include Bruce Leak, COO of WebTV, who helped develop Apple *QuickTime* video technology, and WebTV research fellow Peter Barrett, who invented the widely used *CinePack* video standard. ■

RESEARCH NOTES

Jonathan Campbell

All-round – and up and down – vision

A Columbia University computer scientist has developed a video camera that can see in all directions at once. The Omnicam and special viewing software, developed by Shree Nayar, professor of computer science at Columbia's School of Engineering and Applied Science, and colleagues, gives complete spherical coverage without moving parts.

At the heart of the Omnicam is a video camera that uses a small parabolic mirror to obtain hemispheric views. A miniature off-the-shelf video camera, mounted in a frame, is aimed directly at the apex of the parabolic mirror – a small inverted cup of polished metal enclosed within a transparent hemisphere. So two Omnicams mounted back-to-back can be used to produce views of 360° in a complete sphere.

Special software has also had to be developed to enable multiple images to be displayed on a computer screen in linear perspective at any magnification.

The parabolic optics ensure that the unit has a single effective centre of projection – a single point through which all rays from a scene must pass on their way to the camera's lens. Such a design mimics a camera that takes in only linear perspective, and allows the Omnicam's computer software to generate linear perspective images that are free of distortion.

Vision researchers have previously tried to create omnidirectional vision systems, using fish eye lenses or planar, spherical, conical or pyramidal mirrors. But most of these do not yield the single viewpoint necessary to construct linear perspective images, or, if they do, use moving parts and other complex elements to produce them, says Nayar.

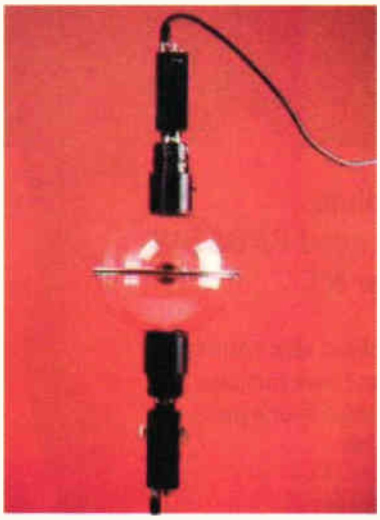
So far, several prototypes of the sensor have been developed, geared towards a variety of applications, including, video surveillance, autonomous navigation, and teleconferencing.

Each prototype is a true video rate (30Hz) omnidirectional sensor that lets a remote user to interact with a dynamic scene. Software allows the user to visually navigate through scene as it changes as well as generate multiple video-rate views of the scene.

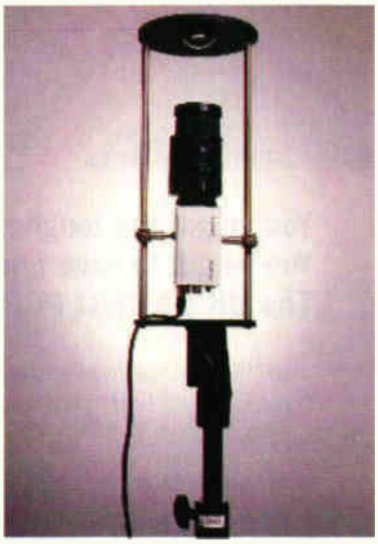
For example, in teleconferences, an Omnicam can show every participant seated around a table, simultaneously, in either hemispheric or linear perspective.

It could also be used to allow a mobile robot to determine its location and direction of travel from local features. Or, placed over a concert stage or at midfield of a sports event, the camera could provide a 360° view to television viewers who could use a set-top box and joystick to bring any frame of that view to their screens.

An online demonstration of Omnicam can be seen at the laboratory's web site at http://www.cs.columbia.edu/CAVE

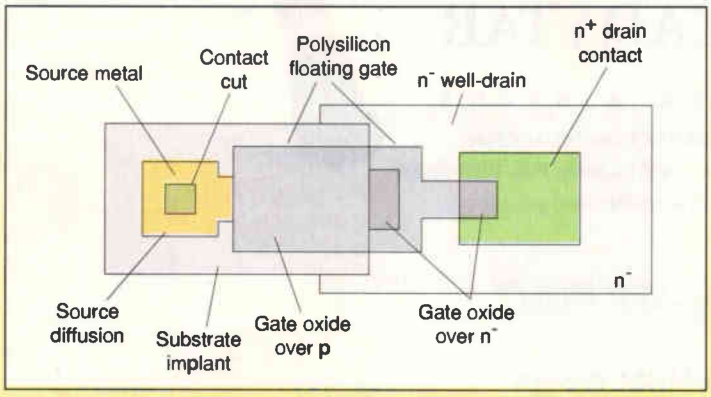


Left: The Omnicam is designed to provide distortion free images. Two units places back to back can give complete spherical vision.



Right: Prototypes have also been developed for specific applications such as navigation.

Silicon device learns how we think

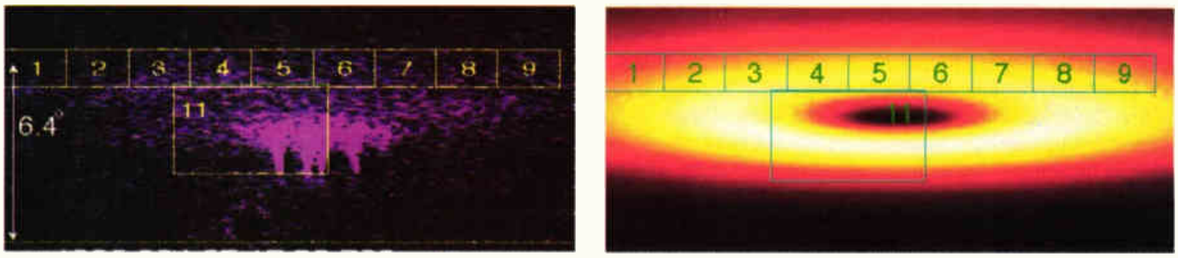


Learning behaviour of brain synapses are now being simulated in silicon by a team at the California Institute of Technology.

Scientists at the Physics of Computation Laboratory, California Institute of Technology, have fabricated a single transistor device that simulates the learning process in the brain. The device decides the result of a stored analogue value and an input in a process analogous to the operation of a brain synapse. By combining devices in an array, the researchers could be on the threshold of development of dense, low power silicon learning systems.

Behaviour of a neural synapse is complex, and the researchers admit that their single transistor can not model its behaviour

Elves that travel faster than light



Integrated model prediction compared with Fly's Eye video image shows the ring-like structure of individual elves. The brief flash begins at the centre and expands outwards. Numbered boxes show show the specific areas where light intensity was measured by the Fly's eye.

Researchers at Stanford's Very Low Frequency Research Group have, for the first time, measured the horizontal structure and dynamics of the new kind of stratospheric lightning that scientists have named 'elves.' The elves are actually high-altitude halos of flickering red light that sometime accompany thunderstorms, and the new measurements, obtained with a specially constructed device called the Fly's Eye, confirm that these flashes take the highly unusual form of luminous rings that spread across the sky at speeds 'faster than light'.

Umran Inan, professor of electrical engineering at Stanford explains that scientists have known for a long time that certain events in the upper reaches of the atmosphere, like solar storms, can affect the lower atmosphere resulting in significant consequences such as power blackouts.

Pilots have reported strange flashes of light above thunderstorms for some time but it wasn't until the late 1980s that scientists these reports were taken seriously.

The pulses of electromagnetic energy generated by the lightning are not anywhere near as strong as those generated by nuclear explosions which can destroy unshielded electrical equipment. But they should carry enough energy to create optical effects.

Researchers proposed that the pulses travel radially outward and upward from the lightning stroke and generate light when they intersect the bottom of the ionosphere – the region above the stratosphere that contains electrically charged atoms.

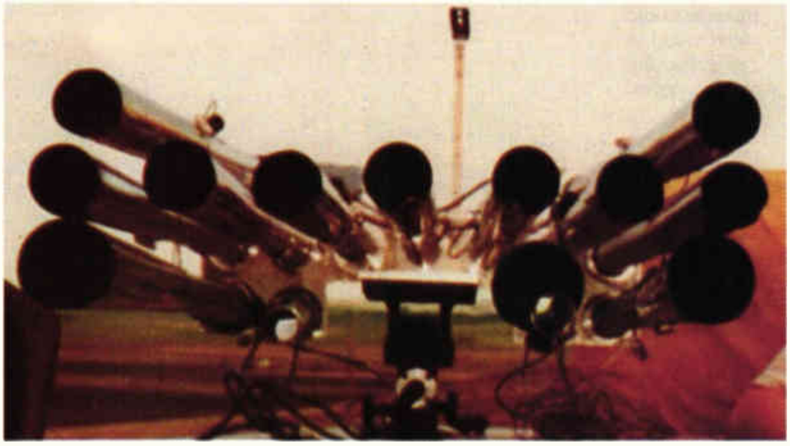
The pulse travels at the speed of light, with the first part of the wave front hitting the ionosphere as a small ring above the lightning stroke. This expands outward as portions of the pulse that travel longer and longer distances strike the ionosphere.

"The ring expands faster than the speed of light for the same reason that waves, when striking the beach at an angle, travel along the shore at a faster speed than the waves move through the water," says Inan.

Because of the ring's superluminal expansion rate, light from its newer parts actually reaches the instrument before the light from the older parts, and this had to be taken into account when interpreting the data.

To gather data on the shape and dynamics of the elves Inan and his colleagues built a special device, christened the Fly's Eye. The instrument has a dozen 45mm barrels, with each barrel pointing to a different part of the sky, and connected to electronics that amplify the incoming light to detectable levels. Because the Fly's Eye has a time resolution of 30μm, it can measure the way elves change over their brief lifetimes.

So far, the Stanford team has been able to record the flickering life cycle of ten elves. All started in a small region centred above the position of a lightning stroke and rapidly expanded outward until reaching sizes as large as 320km across.



Stanford's Fly's Eye is an array of photomultiplier tubes designed to time-resolve the horizontal development of intense sub-millisecond ionospheric flashes that have come to be known as 'elves'.

fully. But it does demonstrate a local learning function (Chris Diorio *et al*, "A single-transistor silicon synapse," *IEEE Transactions on Electron Devices*, Vol 43, No 11, pp. 1972-1980).

The silicon synapse is an n-type mosfet with a poly 1 floating gate, a poly 2 control gate, a moderately doped channel and a lightly doped drain. It uses channel hot-electron injection to add electrons to its floating gate and tunnelling to remove them.

Operation differs from conventional eeprom transistors in that not only does the silicon synapse provide non-volatile analogue memory storage, while computing locally the product of its stored memory value and the applied input, but it also provides simultaneous memory reading and writing, and can even compute locally its own memory updates.

The California team anticipates building synapse-based learning systems in which both the system outputs and the memory updates are computed both locally and in parallel. By contrast, because conventional eeprom transistors are optimised for digital programming and binary-valued data storage, they typically possess few if any of these features, and so have seen only limited use in silicon learning systems.

So far a 2-by-2 synaptic array with a synapse transistor at each node has been tested and is reported to be performing well. The researchers have also developed a synapse learning rule which they believe will enable them to build an autonomous learning system that could form the basis for dense low power systems of the future.

For more information contact Chris Diorio, Physics of Computation Laboratory, California Institute of Technology, Pasadena, CA 91125, USA.

Pixels systems know a thing or two about imaging

Many technologies still need to be perfected before video mail and computer-based teleconferencing become a reality. But Researchers in the School of Electrical and Computer Engineering, Packaging Research Center, at Georgia Institute of Technology look to be well on the road to a breakthrough in at least one key area: development of cheap and powerful smart pixel camera systems.

According to D Scott Wills and colleagues of Georgia Tech ("Processing architectures for smart pixel systems," *IEEE Journal of Selected Topics in Quantum Electronics*, Vol 2, No 1, pp. 24-34), the ideal system, combining high processing performance that scales with vlsi technology advances while achieving high chip efficiency, has not yet been found.

"But a successful solution can have an impact comparable to the introduction of the personal computer, video or fax machine," says the Wills team.

One of the most promising technologies currently being tested in the Georgia Tech labs is the integration of optoelectronics and the pixel system in an architecture given the name Simpil. This incorporates a specialised simd (single instruction stream, multiple data stream) parallel processing architecture with an integrated array of optoelectronic devices.

A 1300nm optoelectronic link allows through-silicon wafer input of digital image data from a detector plane stacked above the processing plane. By reducing the image transfer bottlenecks found in decoupled detector-processor systems, high frame rates are possible without constraining processing power.

Each Simpil node includes a register, an arithmetic logic unit and local memory, and, using the instruction set architecture (isa) to enables addition, subtraction multiplication and multiply accumulation. The node is specifically designed to speed up image processing applications and also includes a-to-d circuitry to convert light intensities to digitally equivalent values.

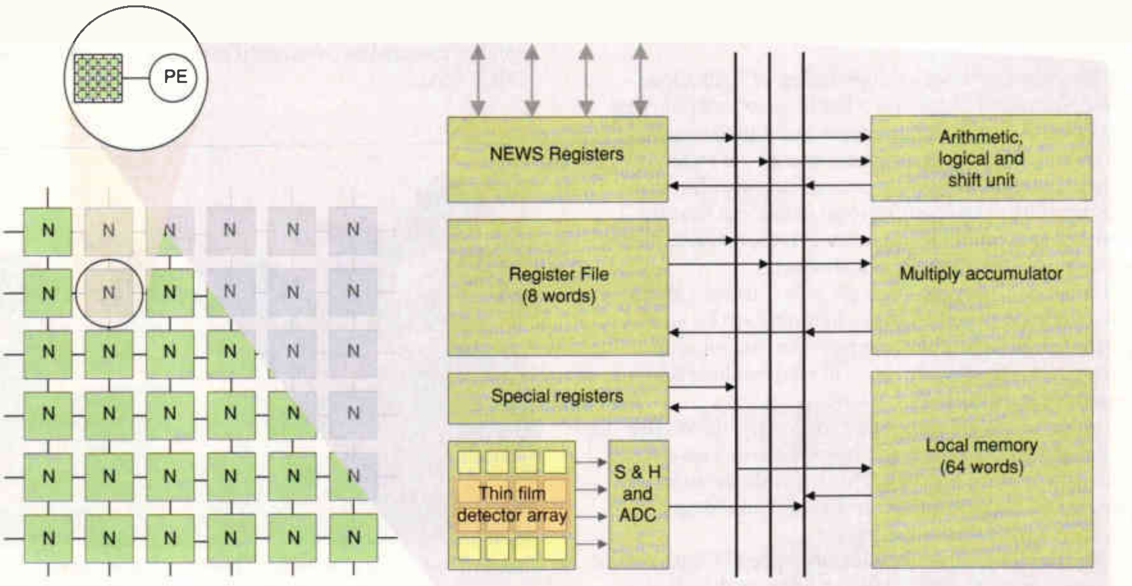
Every node is then interfaced to an array of detectors bonded on top of silicon.

I operation, the parallel processing architecture allows the entire image to be sampled by the system synchronously.

In the Georgia Tech prototype, 64 pixels are connected to each node and the system has so far demonstrated edge detection, convolution and vector-quantisation image compression.

Vector quantisation is a significant application for image processing, and the researchers point out that is clearly demonstrates that high throughput computation can be supported with low memory.

The Georgia Tech chips link processing nodes with subarrays of optoelectronic devices and to neighbouring nodes.



Sensor protects babies from air bag injury

Field sensing technology developed at MIT has been integrated into a smart seating system that can detect the presence of a baby and which way it is sitting. Researchers hope that the system could stop the air-bag being triggered when a baby has been put into the front seat in a rear-facing child seat. So far, in the US, six babies have been killed by the force of a deploying air bag, propelling their seats backwards into the car with explosive force.

Development work on the field sensing technology, which has been taken up by NEC Automotive Electronics, has been carried out at MIT Media Lab's Physics and Media Group may soon prevent these fatalities. Electrodes embedded in a car's seat can distinguish between a rear- or forward-facing baby, and signal the air bag when not to deploy.

Electric field sensing is related to capacitive sensors, such as those found in lift buttons. Though the MIT technology measures the physical quantity of capacitance, it is somewhat different to normal capacitive sensing too.

In capacitive sensing, the amount of current dumped out of a single electrode is measured. For example, as a hand approaches the electrode, the electrode becomes coupled more strongly to ground (through the person).

But in the shunt mode version of the MIT technique,there are distinct transmit and receive electrodes. As a hand approaches, the signal decreases. But since each measurement depends on two electrodes, $n(n-1)/2$ distinct measurements can be made, rather than just n that can be made using conventional capacitive sensing.

While the first application is for baby-seat sensing, the same technology has broader capabilities, and could be used to read any occupant's size and position to determine the most effective air bag action.

Earlier applications of the wide-ranging sensor research have already included collaboration with musicians, creating furniture that can 'see,' and sending data through the human body. ■

CHARGE alkaline cells

If you try to charge an alkaline cell using *any* of the traditional methods – it *will* explode, causing a potentially serious health hazard. But, as Rod Cooper shows here, there is a recipe to recharge them safely – resulting in substantial cost savings if you rely heavily on alkaline cells.

It has been known for a long time that the manganese dioxide/zinc couple is rechargeable. Even before World War II zinc-carbon cells were being given recharging treatment by some of the larger commercial users like cinemas but the techniques used were crude.

In 1953, R W Hallows wrote an article for *Wireless World* with an analysis of these simple methods. He followed this up in 1955 with a review of a device from Holland called the Electrophoor which recharged zinc-carbon cells using a principle called periodic current reversal.

Briefly, this system entailed the use of a continuous train of pulses to charge the cell – a method which had been tried before without much success. But in this case each large forward pulse of current which charged the cell was followed by a smaller pulse discharging it.

The combined result was to charge the cell in a series of steps, two steps forward, one step back. It may seem strange to apply what is in effect an ac charge to a decidedly dc device like a cell, but it works.

Periodic current reversal, or pcr, was already in use by the electroplating industry to prevent the formation of metal dendrites and to give smooth, hard, compact metal deposits. When used to recharge zinc-carbon cells, it had a similar effect on zinc. The main difference between plating and charging was that cell chargers tended to use the mains frequency for pulsing. Hallows pointed out that dendrite formation was one of the main reasons for failure in recharging the zinc-carbon cells. When using dc, pulsed or not, zinc dendrites would quickly form across the separator and short-circuit the cell internally. Periodic current reversal prevented that to a large extent.

Following the Electrophoor, which was a very simple device and easily copied, several other chargers appeared on the market, and to counter their popularity the battery manufacturers started a campaign to convince the public that recharging zinc-carbon cells was unsafe. This was demonstrably untrue. In the mid-seventies there was a well-publicised success by the pro-charging lobby in the Japanese courts to stop this propaganda. This success was not repeated in the United Kingdom or the rest of Europe.

However, these chargers were soon overtaken by events as the popularity of the zinc-carbon cell was already falling – even though a design with an improved format – the zinc chloride type, had appeared. The alkaline-manganese cell, as typified by Duracell, was taking over and these simple chargers could not recharge the new cell. If it was attempted, the cell split and spewed caustic electrolyte everywhere – sometimes explosively.

In the brief period 1981 to 1985, the share of the market taken by alkaline cells increased from 24.5% to 42.6% a process that has continued – though perhaps less dramatically – to this day. The size of the market for alkaline-managanese cell just in the UK was then £230million – so what must the world-wide figures be now?

Rechargeable cells using NiCd have appeared since then but their progress has been dogged by technical shortcomings (see *Wireless World* Jun-Sep '85 'Failure modes in NiCd cells) and they have taken a compara-

Warning

You must not attempt to replicate Rod's charger design unless you understand the concepts discussed in this article in their minutest details – i.e. you are an experienced electronics design engineer with a good knowledge of cell and battery structures. Any attempt to modify any of the components and/or conditions prescribed for the charger circuit could well result in a health hazard due to explosion or burning.

tively small part of the total market compared to the alkaline-manganese primary cell.

Why bother to recharge primary cells?

The amount spent an alkaline cells represents an enormous quantity of raw materials and engineering energy, and normally all of it goes away after just a brief period of use. This situation has always been seen as a good example of the profligacy of technological business by the environmental lobby – with justification.

The economical and ecological implications of this waste are all too clear. At present there is no recycling to speak of – as there is with glass, paper ,and aluminium – despite specific political directives to encourage recovery of the raw materials. Also, these cells are relatively expensive to buy.

In this context, even if a recharging device recycled each cell just once, the saving in raw materials and engineering effort would be enormous. In fact, the P24 design can recharge alkaline cells not just once but several times. The best way to recycle an alkaline cell is obviously not to re-cycle the raw materials but to recharge it.

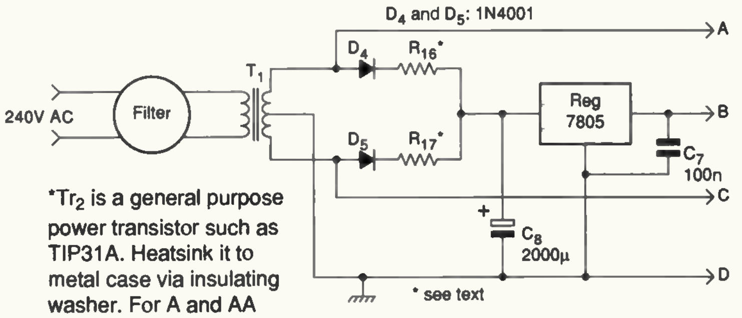
Because of the continuing claims by battery manufacturers that any sort of recharging of alkaline cells is unsafe, this charger has been given two end-of-charge mechanisms and more than usual effort has been put into testing the safety aspect of the design.

The design below was commissioned for commercial use in 1988 before recharging alkaline manganese cells became generally accepted, so before being submitted commercially it was tested with many hundreds of recharge cycles using cells from many different sources. It has proved itself completely safe but only when used correctly.

Recharging alkaline-manganese cells

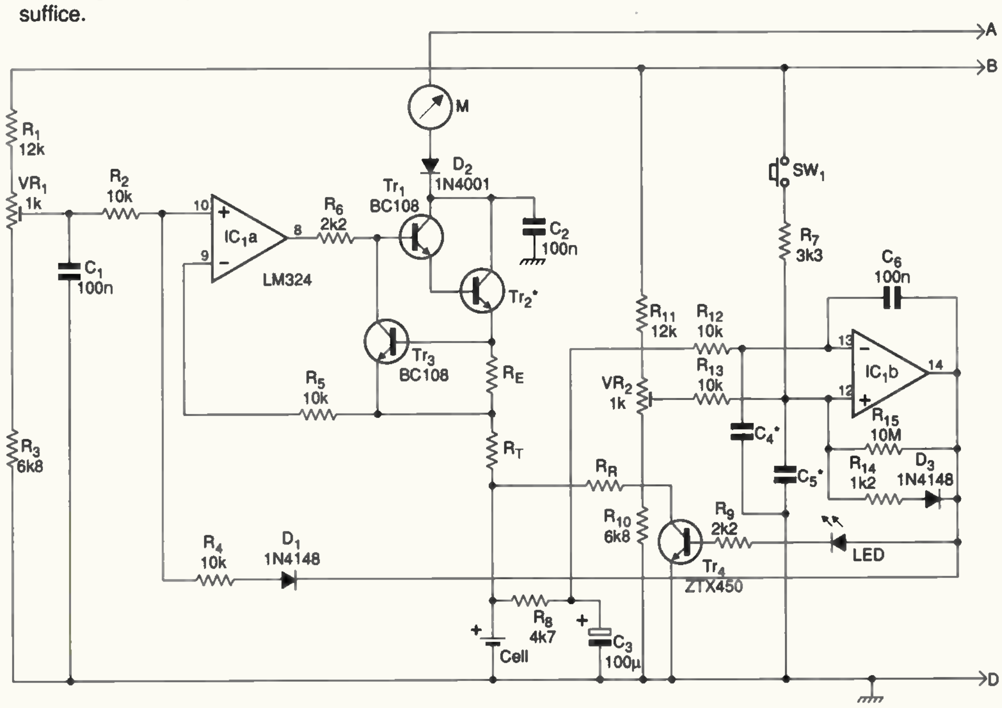
The 'rules' which shaped this design are listed as follows.

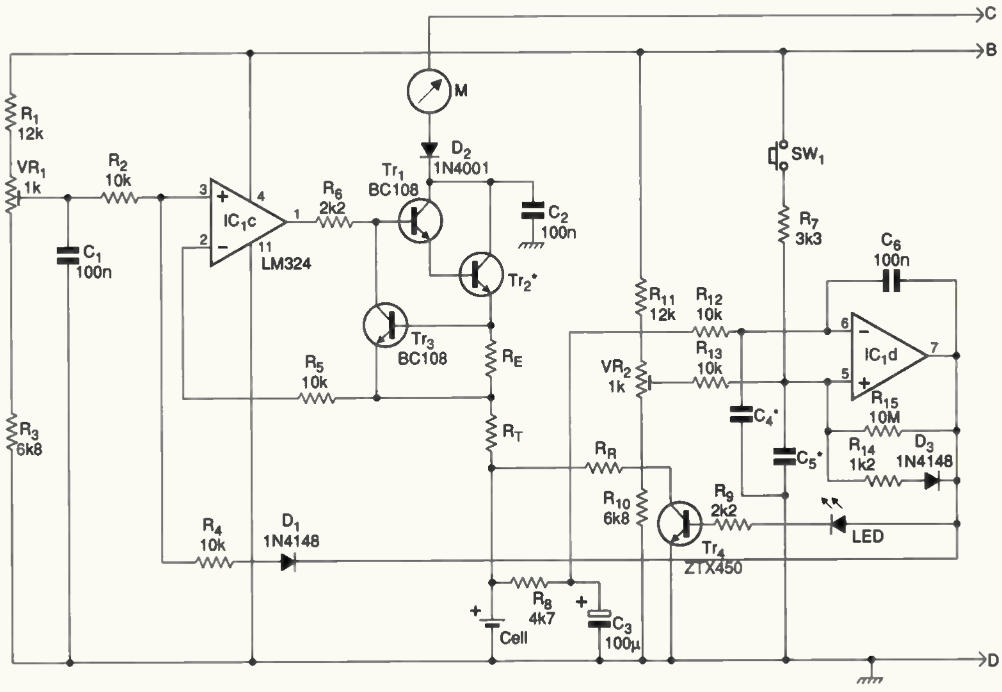
- To be effective, alkaline batteries should be recharged as a single cell, each cell being given its own separate charge regime. Packaged batteries like PP3, PP9 etc. are excluded. Charging these batteries always results in uneven distribution of charge, and leads to reverse charging when discharge takes place. The situation is acute when one cell fails, because this can lead to severe over-charging of the remaining cells in the battery during recharging – which is not safe with alkaline cells – and can result in malfunction of a badly-designed charger. The P24 design charges two cells at a time in



*Tr2 is a general purpose power transistor such as TIP31A. Heatsink it to metal case via insulating washer. For A and AA cells only, a pcb mounting sink may suffice.

Fig. 1. Complete circuit for charging two alkaline cells. Power supply drives both sections and needs a 6-0-6V transformer rated at 1.5A for D cells, 1A for C cells or 0.5A for AA and below. For the values of R16,17, refer to Ray Fautley's article in the February issue.





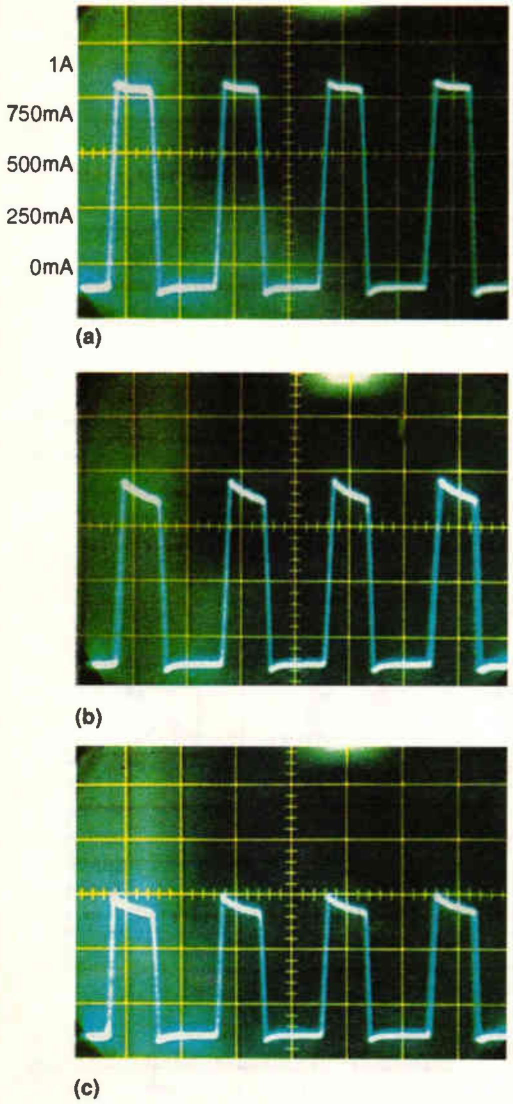


Fig. 2. Oscilloscope views of forward and reverse current in a D cell, a) at start of charge, b) after 1h, c) after 2h, d) after 4h and e) at end of charge, i.e. balance point.

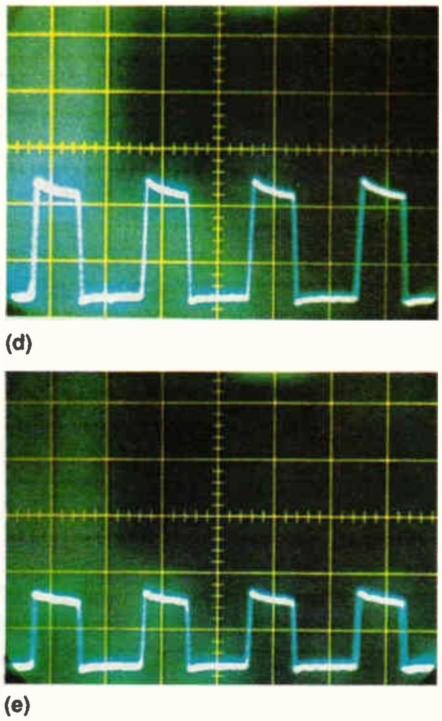


Table 1. Values needed to accommodate the various cell sizes.

	Ah	*Rr*	*Re*	*Rt*
D	15	9.25	0.7	0.3
C	7	20	1.5	0.66
AA	2.25	71	4.9	2.1
AAA	0.8	175	13.5	5.8

separate holders, and treats each cell individually.

● The charging current must be limited to match the charge-acceptance of the cell. One factor affecting the charge-acceptance is the state of charge of the cell. A partly-discharged cell will have a much higher charge-acceptance than a cell approaching the fully charged state – all other things being equal. At full charge the charge-acceptance of an alkaline cell reduces to nothing, for practical purposes. Because of this characteristic, any charging circuit must adapt to give a charge current appropriate to the state-of-charge. A method called taper charging is commonly used with many batteries, which as the name implies, starts at a controlled large current and dwindles as full charge approaches. This technique is used in modified form in the charger.

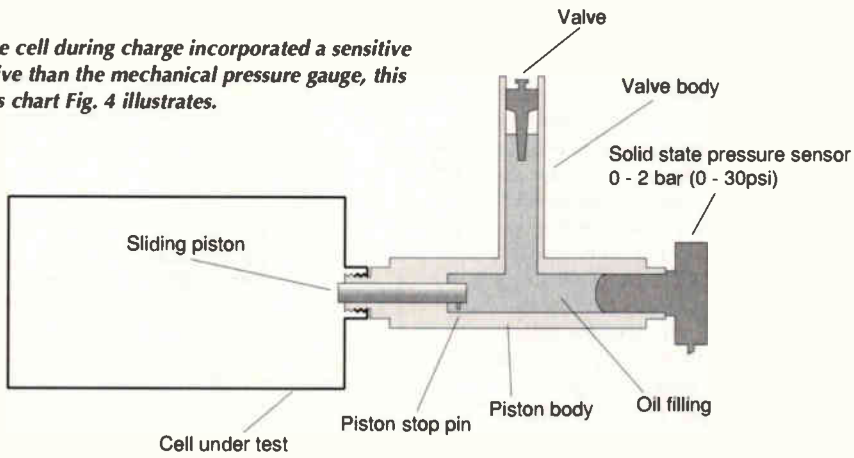
● This design is for use at room temperature. The properties of the alkaline cell vary significantly with temperature. In particular, charge acceptance decreases with falling temperature. Although the commercial version of this design had temperature control, there is no compensation in this version, so a 20°C environment is assumed.

● Current applied to the cell must not exceed a certain maximum charge-acceptance value. For alkaline-manganese cells, this rate is around C/35 to C/40 amps, where C is the capacity of the cell in ampere/hours. There is some variance depending on the origin of the cell. In addition, the charger must be short-circuit proof for safe household use. To deal with both issues, the P24 is made short-circuit proof at the maximum charge acceptance figure. For a D-size cell, the short-circuit current is below 0.5A – safe by any standard.

● Unlike NiCd cells, where overcharge is part of normal operation and is to some extent desirable, no overcharging is permissible with alkaline cells. In the NiCd cell there is a mechanism for recombining the gases produced inside the cell during overcharge, the net result being that the cell merely becomes warm. For the NiCd cell this provides a very convenient overcharge-limiting system. Although a similar mechanism exists for the alkaline-managanese cell, it is not encouraged by the internal structure of the cell, so it cannot be relied on for limiting overcharge. An alkaline-manganese cell could probably be designed with this characteristic but it is unlikely to be produced for obvious reasons. To prevent overcharge, two techniques are used in the P24 charger. First, the charger's taper is made to reduce to nothing well before overcharging can take place. Secondly, a simple voltage cutout operates at a preset voltage slightly below that produced by the taper. One technique complements the other so that if one method fails the other will back it up. This makes the P24 very safe. In many hundreds of cycles, over eight years, it has never burst a cell.

● The charge method must be the periodic-current-reversal type as discussed above to prevent dendrite formation. In this design, pcr

Fig. 3. MkII fixture used for monitoring pressure in the cell during charge incorporated a sensitive electronic pressure transducer. Although more sensitive than the mechanical pressure gauge, this version still showed no significant rise in pressure – as chart Fig. 4 illustrates.



is used at mains frequency for the sake of simplicity, but in a modified form as explained later.

- Lastly – but most importantly – for this charging method to be effective, the alkaline-manganese cell must not be discharged below a certain level. Below this level, irreversible chemical changes take place which render the cell progressively less rechargeable. The actual level is a subject of debate, but in the regime I use, I stop discharge after the top 30% of the total cell capacity has been used. For a typical D size cell of 15Ah total capacity, this gives a usable 5Ah to play with. This is about the same as a D-size NiCd cell provides, but without the snags of the NiCd.

Terminating the recharge cycle

The potential at which gases are evolved from the alkaline-manganese cell is 1.7V at room temperature. Since a voltage slightly less than 1.7V still produces a fully-charged cell, the P24 charger is designed to taper the current to zero at about 1.68V. The cut out operates at 1.62V.

I should mention here in case anyone is tempted to experiment, that this design was not arrived at in a single step. Many circuits were tried in order to provide the 1.68V ceiling – including constant-voltage transformers, electronically stabilised ac supplies, feedback-controlled switch-mode psu and many others. Eventually the circuit of **Fig. 1** was arrived at.

Pulse-balancing

In this technique, instead of charging the cell with a constant ratio of forward to reverse currents, the large forward pulse of the pcr charge is varied and the small reverse charge is kept constant. This gives the taper charge and can be arranged to give a natural balance at the end-of-charge point.

When the cell is in a partly discharged state, i.e. high charge-acceptance, the forward pulse is large but kept within limits by the current limiting circuit. The limit was set at about C/35 amps, but any setting from C/30 to C/40 provides good recharging.

Various criteria were used to fix the maximum limit; the length of charging had to be practical for everyday use; the cell had to show no signs of internal distortion after charging; and the charging components had to be low cost and therefore low power devices.

The maximum ratio of forward charge to reverse charge was set to 4:1 – but in fact any ratio around this figure will work. As charging continues, the size of the forward pulse is gradually lowered to keep within the cell's charge-acceptance limits as explained above, and the ratio of forward charge to reverse charge reduces, until at 1.68V it is 1:1. That is, the energy contained in each forward pulse equals the energy contained in the reverse pulse at this cell voltage. The oscilloscope screens of **Fig. 2** show what happens. Note the period of this quasi-square wave is not quite even.

If left to itself, the cell/circuit combination would settle at a 1:1 pcr ratio at 1.68V and stay there indefinitely. The circuit so far could be regarded as a complete charging circuit, requiring no more components. However, leaving the cell in such a state for any length of time is not desirable because each forward and reverse pulse represents a charge/discharge cycle and there is a limit to the number of these the cell can take.

Also, failure of just one component of the circuit might result in overcharging, so for safety's sake and to preserve the life span of the cell, a further circuit has been added to stop charging. This consists of a simple comparator which effectively stops both forward and reverse pulses just before 1.68V. The voltage chosen is 1.62. This an arbitrary voltage which I found gave a good charge to the cell. Other voltages could be chosen, but voltages lower than this tended to give shorter charging times and not such a good charge. Voltages too close to 1.68V gave erratic turn-off – the reasons for which are mentioned later.

Circuit details

Op-amp IC_{1a}, Tr_1 and Tr_2 provide the 1.68V ceiling voltage for the forward charge pulse. The voltage reference is derived from the 5V power line provided by the *7805* regulator by means of resistor chain R_1, R_3 and VR_1.

Exactly how the ceiling voltage is set is described later, but adjustment is carried out with preset pot VR_1. The forward charging current is provided at mains frequency every half-cycle to this circuit by rectifier diode D_2 from the 6V ac transformer line. On its own, this circuit would generate hf oscillations every half cycle, so this tendency is suppressed by C_1.

Since each of two cells takes a forward pulse every half-cycle, to balance the transformer these two circuits are used back-to-back with centre-tapped transformer.

To limit the maximum current that the above circuit can supply to around C/35, Tr_3 and R_e are added. Transistor Tr_3 simply clamps the base of Tr_2 at a pre-set current level determined by R_e. However, the circuit would try to supply this current at all voltages up to the 1.68V limit, so R_t added to introduce a gradual tapering-off of current as the cell voltage rises. The internal resistance of the cell cannot be relied on for this.

The reverse-charge part of the pcr cycle is

Deciding when to recharge

In practice, imposing a 30% limit does not reduce the usefulness of the method as much as you might think. For example, a typical portable radio running on alkaline manganese C cells can run for 24 hours before needing a recharge, and a flash-lamp with two D cells for about 5 hours.

Moreover, in real life, it seems no more trouble to recharge alkaline-managanese cells than it does when using NiCd. Of course, in contrast to NiCd cells, you can use up the whole of the remaining reserve capacity of an alkaline cell at any time if you wish. This is a big advantage.

A problem arises in deciding when the 30% limit has been reached. With a NiCd cell, it is clear enough when the cell needs a recharge, but the 30% limit is more obscure with alkaline cells. Misunderstanding of the 30% limitation and lack of a method of determining the recharge point has resulted in assessments of this method of recharging, such as the *Which?* magazine survey, giving a negative verdict.

For good results, the current taken by the appliance must be known. Once you know this, you can either judge roughly when the cell needs a recharge from knowing the original Ah capacity, in which case you must put up with the effects of any misjudgement, or you can attach simple integrating timer. This could flash a light-emitting diode or operate a cut-out after C/3 Ah has been reached. A low-current timer design will be shown in a subsequent article.

Safety issues

Safety has to be seen in context. For example, we use that most flammable of fuels, petrol, to power our most popular method transport, but few refuse to travel by car just because of the fire risk. The car makers do not put a warning about fire-risk on cars.

We use a lethal voltage instead of a safe one around the home to power various devices – and we actually hold some of these devices in our hands to operate them!

The risk in both cases is universally accepted despite the fact that people do occasionally come to grief, because the usefulness is great and the risk small when devices are used correctly.

Compared to the two examples above, any risks posed by the techniques put forward must be regarded as miniscule, when approached properly. From the tests conducted, and the continuous use this charger has had over many years, the risks appeared to be acceptable. Any risk would seem to come from random component failure and misuse.

In a well-ordered technology-based society, the warning on the side of alkaline cells Should be amended to, "Do not recharge in an unsuitable charger" It would then be technically correct, which it is not at present.

provided by Tr_4 and R_r. While the cell is being charged Tr_4 is turned constantly hard on via the led and R_g. During the forward pcr pulse, the small reverse-charge current is swamped or cancelled out by the much larger forward current. This is a simpler and cheaper concept than one which uses a second pulsing circuit for the reverse pulse and is easy to turn off at the end of charging.

While charging is in progress, the led is on, and can be used to show when the cell is under charge, but there are better methods as discussed later. Different values of R_e, R_t and R_r are used for cell sizes D, C , and AA. **Table 1** shows values for these sizes.

Op-amp IC_{1b} and its associated components form a conventional comparator to detect the end of charging. It can be pre-set by VR_2 to trip at around 1.62V. Voltage on the cell cannot be monitored directly by the comparator because of the small ripple voltage across the cell when being pulsed. The cell voltage is therefore filtered by R_8 and C_3 before being applied to IC_{1b}.

The comparator is biased towards the 'off' position, i.e. output low, so that once turned off by a voltage over the limit, it cannot be turned back on again except by manually-operated switch. This arrangement is needed because once a fully charged cell has tripped the comparator, the cell voltage soon drops and could turn the comparator back on to give more, unnecessary charging.

Once the comparator is tripped, Tr_4 is turned off via R_9 and the reference voltage presented to IC_{1a} is lowered via R_4 and D_1. This effectively stops charging. In this state only small semiconductor leakage currents around the cell charging circuit are present, which can be disregarded for practical purposes. However it would not be advisable to leave the cell in the charger for long periods, several weeks for example, because of these leakage currents.

The charge sequence can be started again by pressing momentary push-switch S_1 which simply overcomes the bias of the comparator.

The IC chosen for this circuit is the *LM324* quad op-amp – a cheap temperature compensated device which can do both functions of reference and comparator reasonably well. It makes a two-cell charger very economical.

Setting up the circuit

First, comparator IC_{1b} is disabled by being turned on, i.e. output high, by VR_2. A high-value low-leakage electrolytic capacitor is then substituted for the cell. I used 50,000µF computer-grade type which seemed to work very well.

At power on, preset pot VR_1 is now adjusted to give 1.68V on C_2 using a high-impedance dvm. The capacitor acts in approximately the same way as a cell, that is, as a store of electrical energy, so you can say that at 1.68V the energy in from the forward pulse equals energy out from the reverse pulse.

This method ignores the small amount of ripple on the large capacitor about 10mV peak. After this preliminary setting it is best to check the balance point with a real cell to ensure the voltage never rises above 1.68V, before you set the comparator. You could do the setting-up with a cell, but it would then take a much longer time, perhaps several hours, to reach the balance point, whereas you can get to it very quickly using a capacitor.

Having done this, the comparator can be set to trip at around 1.62V by observing the output led while adjusting the voltage on the large capacitor to this value with a variable resistor, While you have the dvm at hand, the comparator voltage swing should be checked to ensure correct operation. This comparator will trip within 10mV of the set voltage.

The charger is then ready to use.

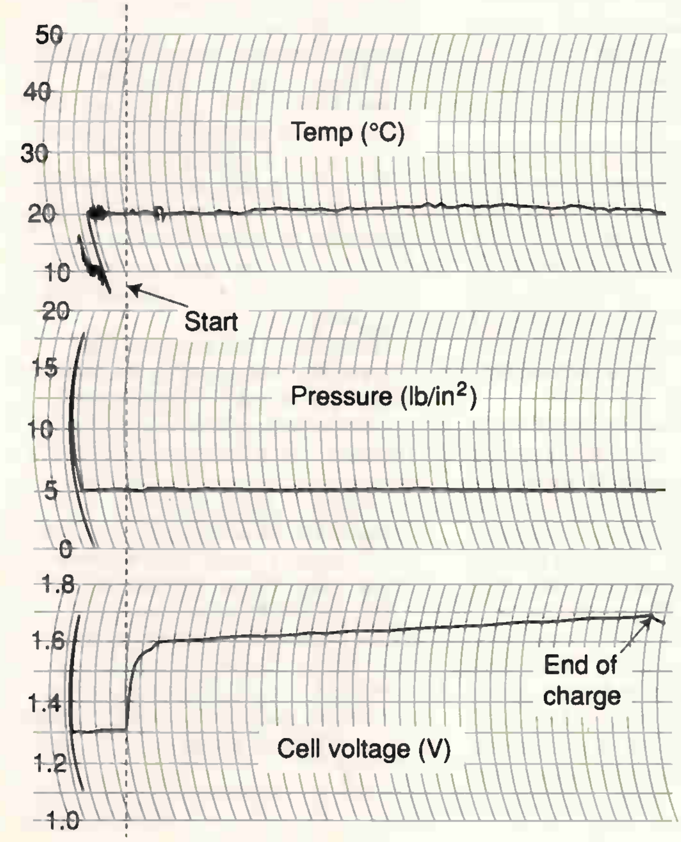


Fig. 4. Traced chart recordings of characteristics of an alkaline call under charge using the circuit of Fig. 1. These show that temperature and pressure rise are negligible.

Charge indication methods

Although the led in series with R_8 can show when charging is in progress, it cannot show what point has been reached. Small inexpensive moving-iron meters are on the market which could be used at point M in the circuit.

A meter is the best indicator with small cells. Alternatively, for D and C cells, a filament lamp-bulb can be successfully used. This is set for full brightness with a discharged cell – i.e. at the short-circuit current – and just glowing at the balance point. A parallel resistor may be needed with some bulbs. For this purpose, I have used a T1 1.5V 200mA, which is difficult to obtain, and a MES lens-end 1.2V type which is more readily available at 200mA and 300mA. Performance can be improved with an ntc thermistor in parallel instead of a resistor.

Using the charger

The charger has been proved from 15°C to 25°C so use outside this range is not recommended. As stated, there is no temperature compensation.

Also, the construction has to be such that the cells and circuit are kept within temperature limits by free ventilation. This means an open construction with plenty of slots for air circulation.

The comparator circuit is prone to trigger prematurely when approaching full charge if current spikes are allowed in from the mains, so a mains filter is essential. For the same reason, the whole circuit should be in a metal box to reduce incoming emi, although the absolute need for free ventilation will nullify most of this benefit. The metal box can be a safety feature if there is any remaining doubt about bursting cells. Also, any thermal effects from Tr_2 can be reduced by using the metal box as the heat-sink.

A spike from a comparator rapidly turning off can trigger a neighbouring comparator to turn off before its time, if that was also in the sensitive region near the end of charging. To prevent an abrupt turn-off, capacitor C_6 has been added.

If you live in an area of high interference, it may be necessary to current-slug the comparator with capacitors C_4 and C_5, and the reference with C_i. The value chosen depends on how severe the interference is. Start with 1µF for *C4* and C_5 if you experience premature turn-off. These capacitors *must* have negligible leakage current.

Drift with age may be a problem if not detected. As a result, it would be reasonable to check the two important voltages with a meter now and then. I suggest a check after the first couple of cycles and then every six or twelve months.

Safety

Because of the likelihood of counter-claims about safety from battery manufacturers trying to protect their market, more than usual effort was put into checking the safety of this design.

There are two main areas that could pose a risk. Firstly, there is the possibility of internal pressure arising from an accumulation of gases caused by electrolysis. This is the the most likely cause of leakage if recharging continues when the cell is fully charged. But it could occur at any point on the recharge cycle if the cell was given a current larger than the cell's charge-acceptance. This could be caused for example by using the charger at low temperatures.

Secondly, a risk could be posed if the temperature during recharge rose to excessive levels. Leakage could come from the expansion of any gases already inside the cell – from reverse-charging for example – or from steam being formed if the cell got hot enough for this. These two potential risks are examined in turn.

Regarding the first point, internal pressure, three methods were used to check internal cell pressure during recharging. I noticed that whenever a cell leaked due to internal pressure from deliberate abuse, it bulged slightly at the ends, and it was possible to measure this small expansion with vernier callipers. In fact, a bulge could be easily detected before any leakage occurred. Both bulge and leak conditions could be readily achieved by deliberately trickle-charging the cell.

This formed the basis for the first method. I measured cells with callipers – with insulated jaws of course – before, during and shortly after being recharged in the P24 but there was never any size increase over many cycles. It was clear that there was no detectable pressure developed by normal recharging.

The second method I used to check pressure was to attach a bourdon type pressure gauge to the cell with an adapter. I did this to be able measure directly the pressure, if any, that was being developed inside the cell. The cell was firstly pressurised via a schraeder valve and left for a few hours to confirm the cell was gas-tight and then given several discharge/recharge cycles.

The gauge never showed any significant pressure increase or decrease. As the volume of the released gases is small, I tried to make this rather clumsy technique more effective by filling the 'dead space' in the gauge with silicon oil, plugged with light silicon grease to stop it escaping and contaminating the cell. But it still never showed any pressure increase.

The gauge was refined in a third, more sensitive test by attaching an electronic pressure sensor to a chart recorder so that the whole cycle could be recorded. The arrangement is shown in **Fig. 3**. This third method again showed no detectable pressure increase.

Regarding temperature rise, it was noted by Hallows that temperature rise with pcr is very small, in contrast to other methods of recharging. My tests confirmed this. A temperature sensor was attached to the metal jacket of the cell during cycling and connected to a chart recorder. As **Fig. 4** shows there was no significant rise in temperature.

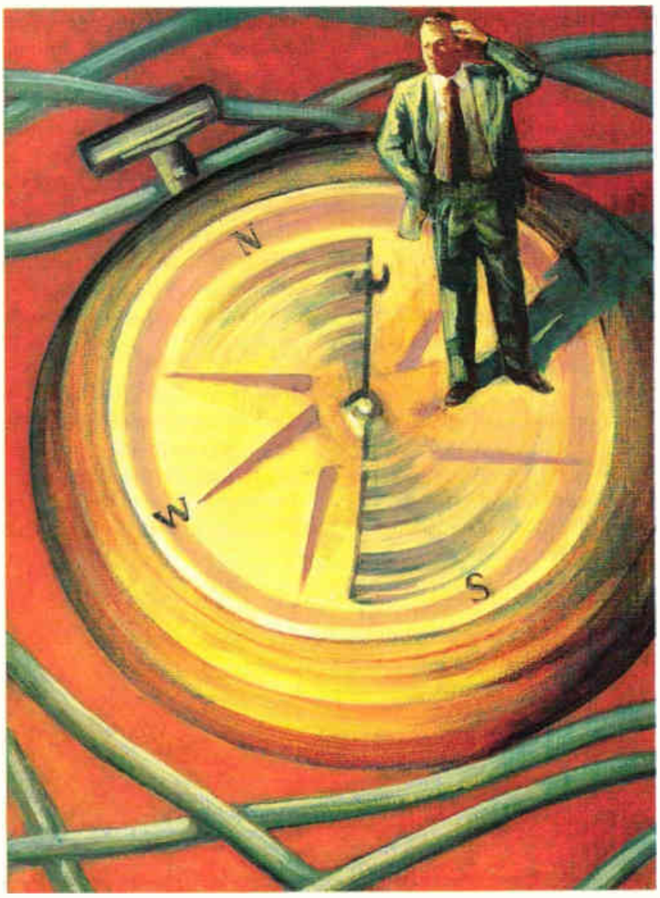
The overall conclusion from these tests was that the P24 simply does not create internal pressure in cells. Cells in many conditions from brand-new to totally dead were used in the tests to cover the range of possibilities in real life.

A post-mortem was conducted on cells that had been cycled to exhaustion to check if there had been any physical changes inside the cell. The cells were sliced up on a miniature milling machine with a 0.5mm slitting saw, which did not disturb the contents too much. In no case was there any tell-tale sign of electrolyte leakage.

No distortion attributable to internal pressure was observed. ■

TRUE orientation

In vector-based orientation schemes, compensating for sensor offset and drift is a major problem. Digital processing not only simplifies compensation, but also allows new techniques such as continuous error removal. David Risk and Richard Noble explain.



Although developed specifically for use with magnetic field sensors, this algorithm is of more general application to any system in which fixed magnitude vector quantities are measured by less-than-perfect sensors. It applies equally well, for example, to the accelerations involved in measuring the gravity vector.

Both types of measurement are of interest in such areas as magnetic compasses, virtual reality devices, sea-bed wreck-finding systems and geophysical surveying.

The algorithm assumes that the sensors used for measurement are substantially linear but may have large, variable and differing zero-offsets coupled with significantly differing and variable sensitivities. This accords reasonably well with reality in that sensor manufacturers usually achieve reasonable linearity, or at least consistent, simply correctable non-linearity, but often suffer from a spread of large zero-offsets which may vary quite alarmingly with temperature, coupled with a spread of sensitivities equally affected by temperature or power supply changes.

In the past, analogue solutions to this type of problem have frequently resulted in a proliferation of trimpots and temperature compensation schemes of varying success.

The currently popular approach of digital processing throughout presents the opportunity to not just simulate these analogue palliatives, but to adopt new, more successful techniques permitting continuous error removal, even if the source of error is not clearly understood.

The approach described here applies specifically to fixed or slowly varying vector measurements, in two or three dimensions, in particular only to vectors which do change their orientation.

The sum of the squares...

The underlying principle used here exploits the fact that the components of a vector, as projected on to the x, y and z axes of an orthogonal coordinate system, are related to that vector by Pythagoras' equation. The square of the total vector magnitude is equal to the sum of the squares of the component values.

Where the vector is of nominally fixed magnitude, the Pythagoras relationship permits the derivation of a unique solution to the sensor imperfections.

Any difference between the sensitivities of the sensors measuring the vector will distort the circle or sphere into an ellipse or ellipsoid, still centred about the origin. The existence of zero-offsets will displace the centre of the ellipse or ellipsoid away from the origin.

Four known points on an ellipse will uniquely define it. Six known points will do the same for an ellipsoid in three dimensions. Hence a succession of real readings from imperfect sensors should permit the calculation of the unknown sensitivities and zero-offsets of each of those sensors.

Details of the algorithm are, for the sake of interest, described in terms of a practical application in which the orientation of the Earth's field is determined using magnetic sensors. The sensor referred here to is the *FGM-3* magnetic sensor developed by Speake & Co, but the principle applies equally well to other types of flux-gate device and to Hall effect or magneto-resistive devices, with appropriate modifications.

The *FGM*-type sensor's output is a large rectangular pulse whose period is proportional to the external magnetic field along its principal axis, within its linear range of operation. Unfortunately, since the output cannot have a negative period, this cannot be a direct proportionality. There has to be a zero field period in the form of a zero-offset large enough to accommodate negative values of magnetic field.

There are then two unknown parameters associated with each sensor, the first being the slope of the relationship between period and magnetic field, and the second, the zero-offset or period which corresponds to zero magnetic field. Both these parameters must be taken into account, when using sensor combinations to determine orientation information using the Earth's field.

Though an attempt is made to reduce the variation in these parameters, no two sensors are alike and some calibration is called for. For small quantities, this calibration is fairly easy to carry out, but in the large scale production of application devices it would be much more desirable to remove the calibration requirement.

Continuous autocalibration

There are some circumstances in which continuous autocalibration is possible. One is the two-dimensional bolt-down type compass magnetometer, in which the sensors are constrained to rotate in a horizontal plane or at least in a fixed plane, which need not necessarily be horizontal. The other is the full three-dimensional sensor combination, used to determine the alignment of the Earth's field with respect to an orthogonal set of sensor axes.

The only other requirement for this type of autocalibration is that the sensor combination should be in continuous or intermittent motion of some sort. If this is the case, it is normally possible to continuously determine and update the values of the two (or three) sensitivities and the two (or three) zero-offsets, using only the readings taken in the normal operation of the orientation-determining device.

The fundamental principles behind the method is the fact that the earth's field can be regarded as substantially fixed in both magnitude and orientation and that the sum of the squares of the orthogonal field components will remain constant regardless of the orientation of the reference axes. In the three dimensional case, for example, if the field components are h_x, h_y and h_z then:

$$h_x^2+h_y^2+h_z^2=|\mathbf{h}|^2$$

If there are any zero-offsets or unmatched sensitivity variations between sensors then this relationship will not hold true and can be made to indicate the required corrections.

For simplicity, a two-dimensional algorithm will be developed first from which the expansion to three dimensions is obvious.

Two dimensional autocalibration

For an Earth field vector, **h**, having orthogonal components h_x and h_y, in the plane of the sensor axes, assume the sensors give output periods of t_x and t_y.

If the sensors have differing sensitivities (slopes of period against field) s_x and s_y and differing zero-offset periods t_{x0} and t_{y0}, then:

$$t_x=s_xh_x/h+t_{x0} \qquad (1)$$
$$t_y=s_yh_y/h+t_{y0} \qquad (2)$$

where the x and y components of the field are assumed to be divided through by h, the modulus of the field. This effectively converts the field components into their direction cosines, which are independent of the field magnitude.

Applying the condition that $h_x^2+h_y^2=h^2$, or more appropriately,

$$(h_x/h)^2+(h_y/h)^2=1$$

gives,

$$(t_x-t_{x0})^2/s_x^2+(t_y-t_{y0})^2/s_y^2=1 \qquad (3)$$

This is the equation, in t_x and t_y, of an ellipse with its centre located at (t_{x0}, t_{y0}) and having principal axes s_x and s_y, **Fig. 1**.

All measured pairs of sensor readings, t_x and t_y, must lie on this ellipse, and hence any four different points are sufficient to define the ellipse completely.

It must therefore be possible to deduce the centre, (t_{x0}, t_{y0}) and the principal axes, s_x and s_y from any four different pairs of sensor readings. Although in theory any four points will do, the precision of calculation with measurements of finite accuracy will be adversely affected if the points are very close together. This should not be a problem with orientation systems which are in constant motion, and the algorithm should be designed to wait until it has collected sufficiently different inputs before proceeding to calculate.

If the four points are denoted by (t_{xi},t_{yi}), i=1, 2, 3, 4 then four equations of type (3) above are available, having the typical form,

$$(t_{xi}-t_{x0})^2/s_x^2+(t_{yi}-t_{y0})^2/s_y^2=1$$

Subtracting these from one another successively will yield three equations of the typical form,

$$((t_{xi}^2-t_{xi+1}^2)-2t_{x0}(t_{xi}-t_{xi+1}))/s_x^2$$
$$+((t_{yi}^2-t_{yi+1}^2)-2t_{y0}(t_{yi}-t^{yi+1}))/s_y^2=0$$

Multiplying each through by s_x^2 and setting $k=s_x^2/s_y^2$ will yield three equations of the form,

$$a_it_{x0}+b_i(kt_{y0})+c_ik=d_i$$
$$i=1, 2, 3$$

where a_i is $2(t_{xi}-t_{xi+1})$, b_i is $2(t_{yi}-t_{yi+1})$, c_i is $-(t_{yi}^2-t_{yi+1}^2)$ and d_i is $(t_{xi}^2-t_{xi+1}^2)$

These three equations are linear in t_{x0}, (kt_{y0}) and k and are therefore soluble for these values, using the determinant method or the Gaussian elimination method of solving linear simultaneous equations.

This will immediately provide the values of t_{x0} and t_{y0}.

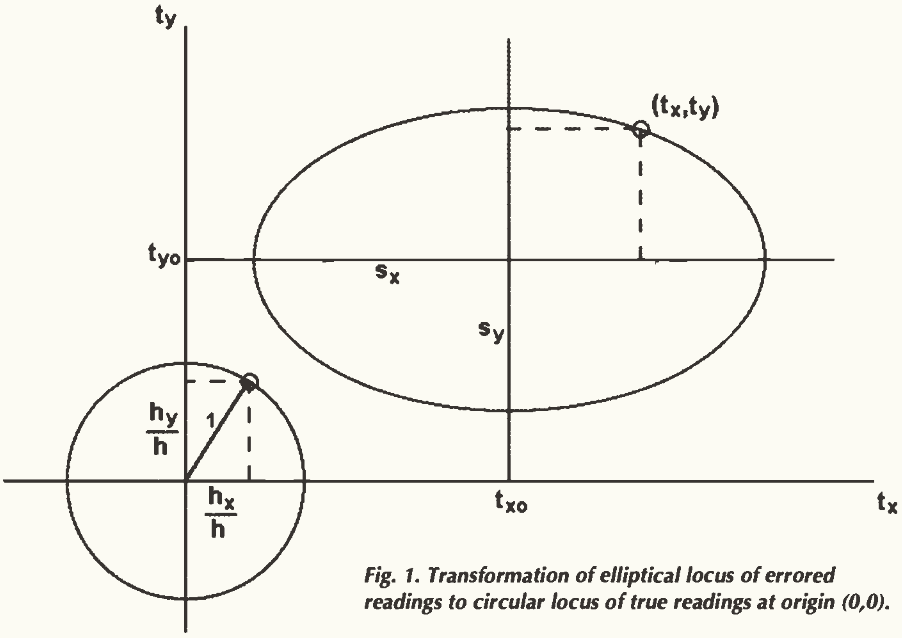


Fig. 1. Transformation of elliptical locus of errored readings to circular locus of true readings at origin (0,0).

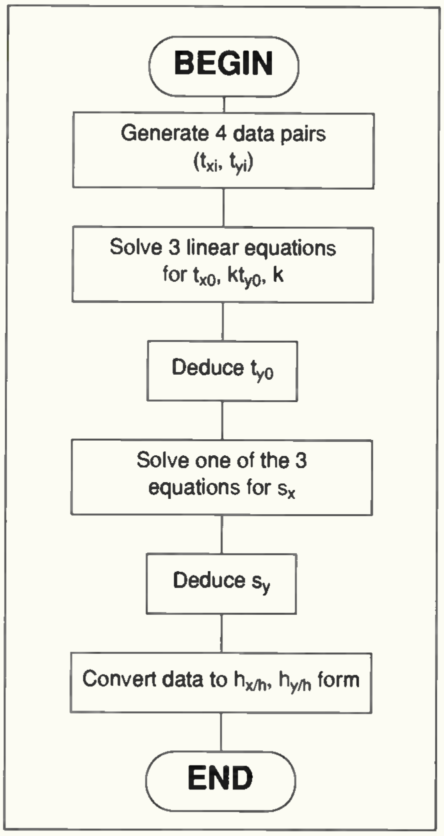


Fig. 2. Sequence of operations in the two-dimensional algorithm

Finally, the last of the type (3) equations, *viz*,

$$(t_{x4}-t_{x0})^2/s_x^2+(t_{y4}-t_{y0})^2/s_y^2=1$$

can be solved for s_x and hence for s_y by setting $s_x^2=ks_y^2$ and inserting the other known values.

This gives the required sensitivities and zero offsets of the individual sensors. They can now be used to correct the incoming readings to give valid direction cosines for the orientation calculations, as follows.

For any pair of readings, (t_x, t_y),

$$h_x/h=(t_x-t_{x0})/s_x$$
$$h_y/h=(t_y-t_{y0})/s_y$$

giving the desired corrected values in terms of known constants and measured values.

Then measuring the orientation angle, θ, of the vector h in the clockwise direction from the y axis, for example, gives,

$$\theta = \tan^{-1}(h_x/h_y)$$

Calibration can be carried out at whatever intervals are considered appropriate to maintain a suitable compromise of stability and speed of data acquisition.

The *FGM* sensors are generally stable enough for orientation purposes without continuous recalibration, if supplied from a stable voltage source, and may only need the initial start-up calibration. The technique can be used, however, to overcome the effects of drift from any potential source.

Moving to three dimensions

The expansion to three dimensions follows the same pattern and ends up defining a three-dimensional ellipsoid using six pairs of differing measurements, yielding five simultaneous equations to solve and giving, finally, three sensitivities and three zero-offsets as corrections.

With this many equations, the determinant method of solution is not very efficient and Gaussian elimination is probably the preferred approach. An excellent description of Gaussian elimination was given by John Hopkins in his article 'DIY Circuit Analysis' in the January 1996 issue of *Electronics World*, p. 31.

Figure 2 is a flow diagram of the sequence of steps required by the algorithm for use in two dimensions.

Practical demonstration

A simple practical demonstration of the mechanics of the algorithm is sometimes useful. This can be carried out by means of a hypothetical experiment. Values for the sensitivities and zero-offsets are assumed and the readings from the sensors are back calculated. The algorithm is then run on the hypothetical readings to show that it can make reasonable estimates of the required corrections.

In practice, the techniques for obtaining readings will vary with individual designers, but a typical method might be to count how many internal processor clock pulses occur during, say, 128 or 256 incoming sensor pulse periods, using either internal hardware count registers or some software equivalent. This generates an arbitrary number which is proportional to the sensor period and therefore to the external field strength. The designer will usually arrange that this number ranges over sizes that stay within some register's capacity for convenience, but which is still large enough to provide the desired precision.

For orientation determinations, the field strength itself does not have to be known, so these arbitrary numbers can go directly into the algorithm as they stand.

Using again the two-dimensional example, suppose that the sensors have different zero-offsets of t_{x0}=2000 and t_{y0}=1850. These are the arbitrary counts obtained at zero field respectively for the two sensors.

Suppose also that the sensitivities differ and can be simulated by slope factors of s_x=880 and s_y=740. Then you can calculate the expected counts for any angle to, say, the y-axis from,

$$t_x=880h_x/h+2000$$
$$t_y=740h_y/h+1850,$$

the values of h_x/h and h_y/h being simply the sine and cosine of the chosen angle. These counts now contain the errors caused by the sensitivity and zero-offset factors chosen.

Using four arbitrarily selected angles, we can make up a table of the field's direction cosines,

θ	0º	30º	60º	90º
h_x/h	0	0.5	0.86603	1
h_y/h	1	0.86603	0.5	0

From this we can calculate a table of errored readings,

t_x	2000	2440	2762	2880
t_y	2590	2491	2220	1850

Using these readings, coefficients of the three simultaneous equations can be calculated,

i	a_i	b_i	c_i	d_i
1	–880	198	–503019	–1953600
2	–644	542	–1276681	–1675044
3	–236	740	–1505900	–665756

In matrix form then the equations are:

$$\begin{bmatrix} -880 & 198 & -503019 \\ -644 & 542 & -1276681 \\ -236 & 740 & -1505900 \end{bmatrix} \begin{bmatrix} t_{x0} \\ kt_{y0} \\ k \end{bmatrix} = \begin{bmatrix} -1953600 \\ -1675044 \\ -665756 \end{bmatrix}$$

Solving by the determinant method gives,

t_{x0}=2000.9 (0.05% error)
kt_{y0}=2604.9
k=1.4086

giving t_{y0}=1849.3 with 0.04% error. From the fourth Pythagoras-type equation using the last set of readings,

s_y^2=548642 and s_y=740.7 (0.1% error)

from which, using $s_x^2=ks_y^2$:

s_x^2=772817 and s_x=879.1 (0.1% error)

The four supposedly unknown parameters have been recovered with a reasonably good accuracy. It may be verified that the errors noted are almost entirely due to using five figure precision for sin60º and cos30º in the direction cosine table. If exact arithmetic is used, as for example in the computer algebra software Derive, then s_x, s_y, t_{x0} and t_{y0} may be recovered with zero error.

Using the corrections above to recalculate the original hypothetical angles from $\theta=\tan^{-1}h_x/h_y$ gives:

h_x/h	–0.001	0.4995	0.8658	1.000
h_y/h	1.000	0.8663	0.5005	0.0009
θ	0.06º	29.97º	59.97º	89.95º
Error	0.06º	0.03º	0.03º	0.05º

In any practical situation, the errors arising from other causes are likely to be much larger than this, but the demonstration does indicate the efficiency of the technique if other errors are small.

This type of hypothetical experiment can be set up as a computer program and usefully exploited to determine the effects of limited digital measurement precision, non-linearity and non-orthogonality on the success of the algorithms when working in less than perfect conditions.

Compass gimbals

An interesting extension of the basic ideas used above applies to certain types of compass, particularly hand-held types, which often require some kind of calibration on a level surface after being switched on. It could also apply to other types of installed compass, used within a limited geographical area in which the earth's field is assumed to vary very little.

For such systems, it should only be necessary to try to ensure a level attitude in one axis, rather than the two normally assumed necessary, since a simple Pythagoras calculation will immediately indicate whether the crossed sensor assembly is level or not, by comparison with the known horizontal component magnitude.

Furthermore, if one axis is known to be level, either by spirit level or gimballing, it is relatively easy to calculate the correct level magnitude of the tilted sensor and even the extent of the tilt. The first case is found by ignoring the output of the tilted sensor and just calculating what it should be by Pythagoras. The second case is found by now using the erroneous tilted output value in conjunction with the calculated true value, in the appropriate trigonometric equation to find the angle between them.

For these restricted cases, then, one gimbal can apparently be discarded, making construction simpler.

Vanishing vehicles?

An even more interesting property of the algorithms, which was neither sought after nor even appreciated originally, is their ability to make surrounding magnetised or magnetisable material virtually disappear from the view of the orientation-determining device.

For a compass, this means that the ship or vehicle in which it is installed vanishes, as far as causing deviations is concerned. While this is not totally true, as will be seen later, it appears to be correct for any reasonably careful installation down to third-order effects. Even for a bad installation it should effect a major improvement.

We are indebted to W. Denne (Extra Master, F. Inst. Nav., Assoc. R.I.N.A.) for the background information and theory on which this analysis of effects is based. For the purpose of the analysis, two types of interfering material are considered.

The first is magnetised material, by which is meant anything which has acquired a degree of remanent magnetisation. This is magnetically 'hard' material with a reasonably high coercivity which has become magnetised by some event in its history and retains this magnetisation, much like a deliberately fabricated permanent magnet. This type of magnetisation can occur during such processes as arc welding or construction and generally does not change much subsequently.

The second type is magnetisable material, by which is meant anything which can acquire a temporary magnetisation as a result of being in a magnetic field. This is magnetically 'soft' material with a low coercivity but reasonably high permeability, allowing it to magnify any local fields which surround it to much higher values. Such material will produce temporary magnets as a result of the Earth's field, for example.

These two types, either singly or in various combinations, account for all the interfering fields experienced by a compass installation and are described by Denne using the following notation.

X, Y and Z represent the true components of the earth's field, X being in the forward direction of the vehicle, Y being in the transverse direction to the right (or starboard) and Z being vertical.

X', Y' and Z' are the disturbed components as seen by the compass, using the same directional significance.

The disturbed values can be expressed in terms of the true values by the following,

$$X'=X+aX+bY+cZ+P$$
$$Y'=Y+dX+eY+fZ+Q$$
$$Z'=Z+gX+hY+kZ+R$$

where the coefficients a to k are attributable to errors arising from the 'soft' material being magnetised by the Earth's field, and P, Q and R are deviations caused by the permanent 'hard' material.

Rearranging the terms gives,

$$X' = X(1 + a) + P + bY + cZ$$
$$Y' = Y(1 + e) + Q + dX + fZ$$
$$Z' = Z(1 + k) + R + gX + hY$$

The first term in each of these equations represents a variation in the amplitude of the field component or, in other words, a sensitivity variation in the measurement.

The second term in each equation is an added offset to the field value, exactly that which we have described as a zero offset previously. If the remaining two terms in each equation were zero, negligible or could be removed algorithmically like the others, then the effect of the disturbing material would be eliminated completely.

It has not proved simple to remove these remaining terms algorithmically, so it is of interest to examine them by a physical interpretation. In any magnetisable body, the direction of magnetisation arising from an external field would generally be expected to have the same direction as the field causing it. For example, in a spherical, totally isotropic piece of soft iron, the direction of magnetisation caused by Earth's field would align itself precisely with the earth's field. In this case, the cross-axis effects would not exist and coefficients such as b, c, d, f, g and h would be zero.

For shapes which are increasingly less symmetrical, then shape-dependent demagnetisation would give rise to some tendency to depart from true alignment of field and induced magnetisation, leading to finite values for the cross-axis coefficients. Nevertheless, for all but peculiarly shaped objects, such as long thin bodies, we would argue that the departure from alignment would represent an order lower effect than the main axis effects.

Magnetic anisotropy could also give rise to similar alignment errors but again, for all but highly anisotropic materials, should give rise to an order lower errors. Provided the initial installation is carried out with some consideration of such errors, it would seem that the cross-axis effects could be of a lower order than the main deviations.

In any case, a considerable improvement in performance might be anticipated from use of the automatic calibration algorithms in most cases.

References

1. Hopkins, J., DIY Circuit Analysis, *Electronics World*, January 1996, pp31-33
2. Denne, W., Magnetic Compass Deviation and Correction, Brown, Son & Ferguson, Ltd, Glasgow G41 2SG (ISBN 0 85174 332 3)

Microcontroller implementation

The algorithms described here are no longer purely academic. Since submitting this article, one of the authors, Richard Noble has successfully implemented a two dimensional version in a microcontroller type chip intended for use as a compass in conjunction with an *FGM-2* two-axis magnetic sensor.

To obtain a reasonable precision, the internal computations are carried out in floating point using a 16-bit mantissa and 8-bit exponent and for economic memory usage the equation solution makes use of Gaussian elimination techniques. Although the *FGM-2* in this type of configuration has zero offsets an order of magnitude larger than the expected signal variations, the immediate result was an orientation precision of one degree and accuracy of two degrees in the final output round the full 360° circle.

The autocalibration is carried out every time the system has collected four sets of input readings which differ from one another by at least three percent.

The 'vanishing vehicles' theory was also put to a rather crude test by attaching a lightly magnetised steel bolt to the sensor. This produced an immediate deviation of about 11°, but after further rotation which triggered a calibration run, the original precision was promptly restored around the full circle.

The work described above was carried out as a FUSE project under the EC IT programme ESPRIT, under the guidance of Bournemouth University as Technology Transfer Node and Staffordshire University as First User Consultants, whose assistance along with that of the EU initiative is gratefully acknowledged.

A production version of the chip will shortly become available from Speake & Co Limited, Elvicta Estate, Crickhowell, Powys NP81DF, Tel: 01873 811281, fax 01873 810958.

Building on Wu and King's landmark paper on continuously loaded antennas, Richard Formato's design procedure for impedance-loaded wideband antennas maximises bandwidth while improving radiation efficiency.

WIDEBAND ANTENNAS

Antennas and resistors are usually like oil and water – they don't mix, at least most of the time. The classic example of an absolutely terrible 'antenna' with an excellent standing-wave ratio is a dummy load. A good dummy load's response is nearly flat well into the uhf range. But because essentially all input power is dissipated as heat from i^2R (Joule heating) losses, for practical purposes its radiation efficiency is zero.

Adding resistance to an antenna invariably reduces efficiency and, as a general rule, adding more resistance makes the antenna worse.

But resistance isn't always bad. As the dummy load shows, resistance can broaden an antenna's response by flattening the variation of input impedance with frequency. Certain types of communication system benefit substantially from wideband antennas, typical examples being spread spectrum, frequency-agile, and ALE, or automatic-link-establishment, systems. In each case, it is desirable to maximise antenna bandwidth while maintaining acceptable power gain and radiation pattern.

One way to accomplish this objective is to add resistance. The question is: how much resistance should be added to strike a reasonable balance between wider frequency response and reduced radiation efficiency?

Adding the correct amount of resistance at the proper location can significantly extend an antenna's frequency range while still providing quite acceptable efficiency and gain. This article describes an improved technique for computing the required loading profile for simple wire antenna elements. A typical monopole antenna is then discussed that provides continuous coverage from about 12MHz to beyond 150MHz with no tuner or matching network.

This is not the first time

The idea of adding resistors to an antenna to improve frequency response has been around for quite some time. In 1953, Willoughby[1] discussed resistively loaded wires in a variety of configurations, including Vees and Rhombics, that provided wideband transmit and receive antennas. The wires were loaded either with discrete resistors or with a gradually tapered resistance profile such that the end nearer the rf source had the lowest resistivity and the end farther from the source had the highest.

Resistance can transform a resonant, standing-wave antenna element into a non-resonant, travelling-wave element, thereby increasing the loaded antenna's bandwidth. The distinction between resonant (standing wave) and non-resonant (travelling wave) antenna elements can be illustrated by considering the centre-fed dipole, or cfd, antenna in **Fig. 1**. In the unloaded antenna, resonance results from the superposition of outward-travelling waves produced by the rf source and reflected waves generated at the impedance discontinuity at the cfd's free ends.

These two oppositely propagating waves combine to produce a standing wave which determines the cfd's resonant frequency. If, however, the outward-travelling wave were not reflected, then no standing wave would exist, and the cfd would not exhibit resonance.

One way to minimise reflections is to add resistance near the ends of the element. The resistors absorb incident energy that has not been radiated away from the antenna, thereby reducing the reflected wave amplitude. This general principle underlies all resistive loading schemes. Of course, there are many ways in which resistance can be added to an antenna, and different approaches can produce dramatically different results.

Altshuler[2] provided the first analysis of the effect of adding a discrete resistance to the cfd. He found that an essentially travelling wave current distribution resulted from inserting a 240Ω resistor in each arm of the dipole a distance λ/4 from the end, where λ is the wavelength. Radiation efficiency was reduced by about 50%, but the input impedance was essentially constant over a 2:1 frequency range.

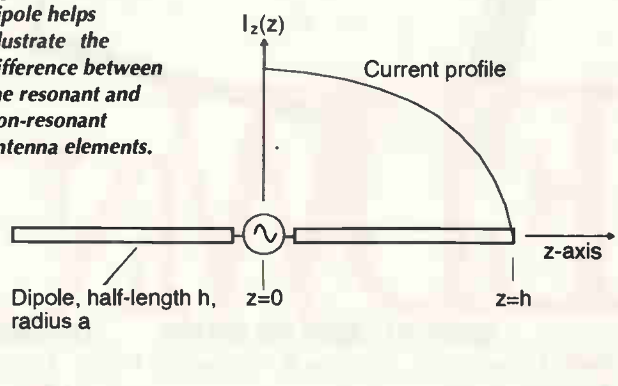
Altshuler's work provided impetus for Wu and King's[3] landmark paper on continuously loaded antennas. Their work forms the basis of recent efforts to improve bandwidth by adding resistance. It is discussed in more detail below.

Some of the results achieved with loaded antennas have been impressive. Kanda[4] built a very small receive-only field probe – a loaded cfd – that exhibited essentially flat frequency response from hf to beyond 1GHz. This sensor was so heavily loaded, however, that its radiation efficiency was far too low for it to be useful as a transmit antenna. Rama Rao and Debroux[5,6] described a 35ft loaded hf monopole with swr≤2 from 5-30MHz and radiation efficiency ranging from about 15%-36%.

This antenna used a fractional loading profile equal to 0.3 times the Wu-King profile and a fixed, lumped-element matching network. Other loading profiles have been proposed that combine resistance and inductance to improve bandwidth and efficiency[7].

This article describes a modification of the original Wu-King profile that increases antenna bandwidth by creating a travelling-wave element

Fig. 1. Centre-fed dipole helps illustrate the difference between the resonant and non-resonant antenna elements.



while at the same time improving radiation efficiency by increasing the antenna's average current. Because the radiated fields are proportional to the antenna's Idl product, a higher average current increases the radiated fields, which in turn improves efficiency. The motivation for this new profile is the realisation that the Wu-King current profile is a special case of a more general travelling-wave current distribution with higher average antenna current.

Wu-King explained

Figure 1 shows a centre-fed dipole antenna consisting of two elements of length h and radius a. Amplitude of the current profile is plotted schematically along one element's length. Maximum current occurs at the rfsource at the feed point, and the magnitude decreases along each arm until it reaches zero at the end. In the Wu-King model, the centre-fed dipole is assumed to have an internal impedance profile along the wire element given by $Z^i(z)=R^i(z)+jX^i(z)$, where Z^i is the (complex) internal impedance per unit length (Ω/m), consisting of lineal resistance R^i and reactance X^i, and where $j=\sqrt{-1}$.

Wu and King develop the differential equation satisfied by the current $I_z(z)$, and then determines by inspection that a travelling-wave current mode exists for one particular impedance profile, Z^i. The Wu-King current distribution is,

$$I_z(z) \approx \underbrace{\left(1-\frac{|z|}{h}\right)}_{\text{linear amplitude decay}} \exp \underbrace{\left(-jk_o|z|\right)}_{\text{travelling wave factor}} \qquad (1)$$

which consists of the product of a linearly decreasing ('straight line') amplitude and a travelling wave propagation factor in the complex exponential term. The wave number is $k_o=2\pi/\lambda$. The propagation factor represents a current wave progressing outward along each dipole arm. There is no reflected wave propagating toward the source to form a standing wave pattern, and consequently no resonance effect.

This current distribution exists *only* when the cfd element has a specific '1/z' internal impedance profile. The required profile is given by:

$$Z^i(z)=\frac{60(\psi/h)}{1-\frac{|z|}{h}} \qquad (2)$$

where $\psi=\psi_R+j\psi_I$ is the complex expansion parameter discussed in Altshuler[2], with real and imaginary parts subscripted R and I, respectively. The ratio of the antenna element's vector potential to current is ψ, and is approximately constant along its length. Because ψ varies with frequency, it is usually evaluated at the fundamental cfd resonance, that is, when $h=\lambda/4$ (see ref. 3). The $1/z$ profile in equation (2) is the basis for the resistive loading used in refs. 4, 5 and 6.

Improved loading profile

An improved loading profile – that is, one that provides better radiation efficiency than the $1/z$ profile – can be obtained by generalising the Wu-King results. The first step is to assume a power law travelling-wave current distribution, of which the Wu-King current distribution is a special case.

The next step is to substitute the assumed current distribution into the current equation developed by Wu and King, which then yields the condition that must be satisfied by the element's internal impedance in order to generate travelling-wave only modes.

This approach is fundamentally different from the one in Wu-King because the loading profile for a particular travelling-wave current mode is now an unknown which is determined by solving the appropriate equations.

The generalised cfd current distribution is assumed to be of the form:

$$I_z(z)= \underbrace{C(h-|z|)^{v}}_{\text{power law amplitude decay}} \exp \underbrace{\left(-jk_o|z|\right)}_{\text{travelling wave factor}} \qquad (3)$$

where C is a complex constant determined by the current at the feed point. Note that the amplitude decay is a power law variation with exponent v. The Wu and King case is recovered when $v=1$, but when $v\neq1$ the more general case is obtained.

The internal impedance profile that produces travelling-wave only currents of the form in equation (3) is determined as follows. The derivatives dI_z/dz and d^2I_z/dz^2 are computed and substituted into the equation satisfied by $I_z(z)$ (Wu and King's equation[11]). This generates the equation that must be satisfied by the auxiliary function $f(z)$ introduced in Wu-King equation (9). Its solution is,

$$f(z)=2v(h-|z|)^{v-2}\left\{1-j\frac{v-1}{2k_o(h-|z|)}\right\} \qquad (4)$$

$f(z)$ determines the impedance profile. Equation (4) generalises Wu and King's equation (12), and recovers their results exactly when $v=1$.

Figure 2 shows several current amplitude distributions parametric in the power law exponent v. It is apparent that values of v less than 1 can lead to significantly higher average antenna currents. Radiating elements with these current distributions are more efficient than those using the $1/z$ loading profile which results by setting $v=1$.

The loading profile resistance and reactance per unit length are computed from $f(z)$ and are given by:

$$R^i(z)=60v(h-|z|)^{v-2}\left\{\psi_R-\frac{(1-v)\psi_I}{2k_o(h-|z|)}\right\} \qquad (5a)$$

$$X^i(z)=60v(h-|z|)^{v-2}\left\{\psi_I+\frac{(1-v)\psi_R}{2k_o(h-|z|)}\right\} \qquad (5b)$$

The corresponding lineal inductance (henry/metre) or capacitance (farad/metre) is given by $L^i=X^i/\omega$ and $C^i=(\omega X^i)^{-1}$, respectively, for $X^i>0$ and $X^i<0$. The circular frequency is $\omega=2\pi f$ where f is the frequency (Hz) at which ψ is computed.

It is apparent from equation (5) that the improved loading profile in

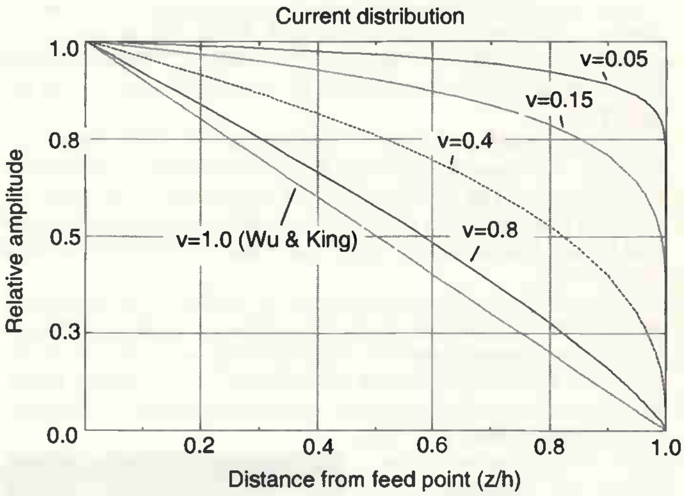


Fig. 2. Values of v less than 1 can lead to significantly higher average antenna currents.

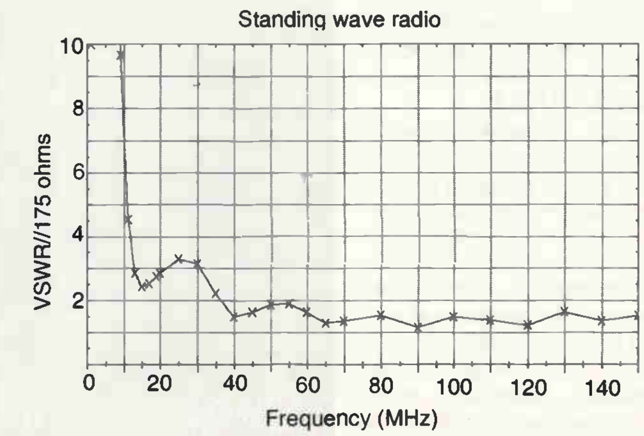


Fig. 3. Compound input swr for a 174Ω feed system, with calculated values marked at points x.

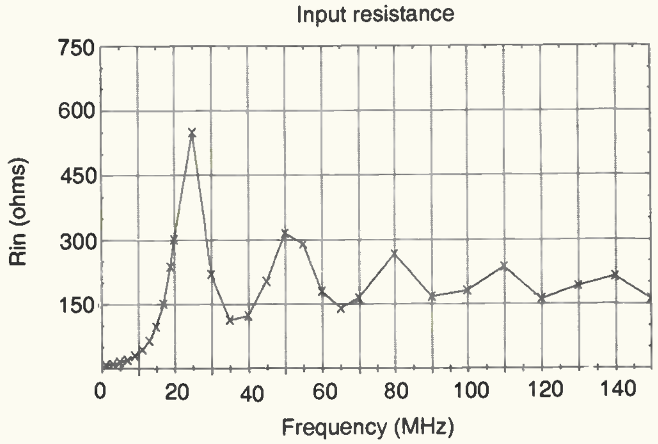


Fig. 4. Monopole feed-point resistance plot.

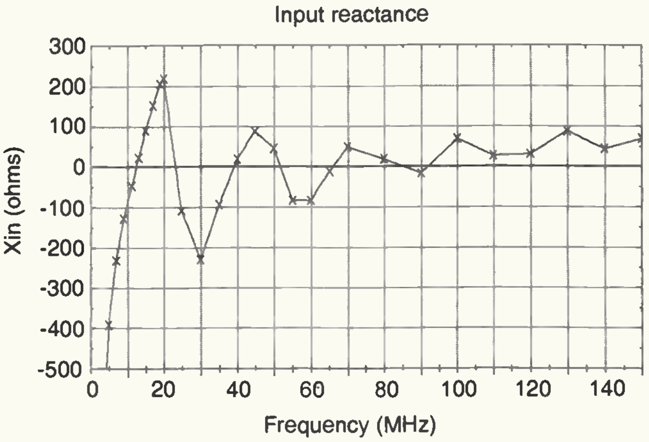


Fig. 5. Reactance plot for the monopole feed-point.

general contains both resistance and reactance. But adding reactance to the antenna, especially capacitive reactance, can complicate construction. As a consequence, many practical designs employ only resistive loading (see refs. 5 and 6, for example), because excellent results are often achieved even without the loading profile's reactive component.

Loaded hf-uhf monopole

To illustrate the degree of broadbanding achievable, a loading profile was computed for a monopole element fed at its base against an infinite, perfectly conducting ground plane. The radiating element height is 5.83m, and its radius 2.54cm. The design frequency for evaluating ψ is 12.86MHz, and the power law exponent v is 0.05. ψ is 8.961–j2.431.

Using equation (5a), a resistance profile was computed for 14 discrete loading points along the antenna, **Table 1**. The profile increases very gradually from 0.419Ω near the base of the monopole to approximately 787Ω near the top. Reactive loading (in this case, inductive) was not included.

The monopole's performance was computer-modelled from 1 to 150MHz. The computed input SWR for a feed system impedance of 175Ω appears in **Fig. 3**; calculated points are marked x. Because swr was computed for a 175Ω characteristic impedance, matching the usual 50Ω coaxial feed requires a 3.5:1 Unun or another suitable broadband transformer.

The monopole antenna's performance is excellent at all frequencies above 36MHz. The swr is below 2 from there to 150MHz (the upper limit for the computer model), and somewhat worse from approximately 12 to 36MHz, reaching a maximum 3.3 at 25MHz. Below 11MHz, swr increases rapidly due to increasing capacitive reactance and decreasing radiation resistance. This behaviour is characteristic of electrically short antennas, and is evident in the monopole's feed point resistance and reactance plots in **Figs 4** and **5**, respectively. The data in these curves were used to compute the swr plot in Fig. 3.

The monopole antenna's impedance bandwidth is remarkably good – especially considering that there is no matching network and only discrete resistive loading is employed. In addition, no attempt was made to further improve the loading profile by, for example, modifying computed resistance values or adding reactance. Adjustments such as these can frequently yield even better performance, but they are not considered further.

Table 1. Resistance profile for 14 discrete loading points along the antenna.

Height (m)	Resistance (Ω)
0.208	0.419
0.625	0.489
1.041	0.581
1.458	0.699
1.874	0.859
2.290	1.080
2.707	1.401
3.123	1.889
3.539	2.689
3.956	4.129
4.373	7.131
4.789	15.102
5.205	49.473
5.622	786.910

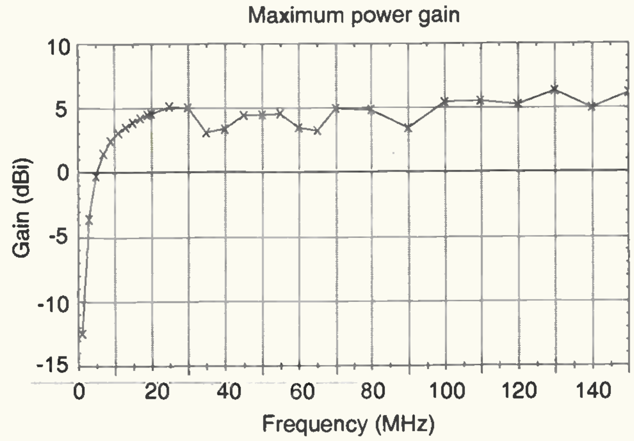


Fig. 6. Maximum gain of the wide-band monopole antenna.

The following observation illustrates how dramatic the effect of loading an antenna can be. For an *unloaded* monopole, the bandwidth for swr≤2.5 (50Ω feed) is typically 15-25% of its λ/4 frequency, depending on the length-to-diameter ratio. A monopole λ/4 high at 12.86MHz, such as the one considered here, would show a bandwidth of less than 3.2MHz. Increasing the bandwidth to greater than 115MHz, as the improved loading profile does, is indeed a very substantial improvement.

Of course, as the dummy load example teaches, impedance bandwidth alone does not a good antenna make. Two other key measures of the loaded monopole's performance appear in **Figs 6** and **7**, maximum gain and radiation efficiency, respectively.

Power gain, computed as the product of directive gain and efficiency, is plotted in dBi (decibels relative to an isotropic radiator). For comparison, the maximum power gain of a half-wave cfd in free space is 2.15dBi. The loaded monopole's gain at 10MHz is nearly 3dBi, and from 10 to 150MHz it is mostly in the 4-6dBi range. The monopole with the improved resistance profile thus exhibits power gain figures that are typical of similar antennas with no loading at all.

The point was made at the start of this article that the fundamental issue in choosing a loading profile is the trade-off between bandwidth and radiation efficiency. The merit of a particular profile is determined primarily by these performance measures. An examination of Fig. 3 showed that the monopole's swr curve is more or less flat from 36 to 150MHz, with somewhat higher but still acceptable swr from 12 to 36MHz.

The second measure of merit, radiation efficiency, is plotted in Fig. 7. The efficiency is generally above 60% over the entire range of 10 to 150MHz, with only minor dips below 60%, and some regions where it is near or above 70%. Even the minimum efficiency value of 45% or so near 35MHz is quite acceptable.

The improved resistive loading profile has produced an antenna with exceptionally good swr bandwidth, relatively high power gain, and very acceptable radiation efficiency.

In summary

Adding resistance to an antenna can dramatically improve bandwidth, but doing so reduces radiation efficiency. The trade-off between greater bandwidth and efficiency is not arbitrary. Some loading profiles are much better than others for creating wideband antenna elements.

Previous theoretical calculations of suitable profiles provide a sound basis for loaded element design yielding very good results. But these studies considered only a special case of a travelling-wave current distribution. The improved loading profile described in this article results from extending the previous work to a power law travelling-wave current mode.

Typical computer modelling results show that the improved profile provides better performance than previously used profiles. The technique for calculating the improved element loading promises to yield still better antennas in terms of bandwidth and efficiency, and may be put to good use to accomplish this goal. ■

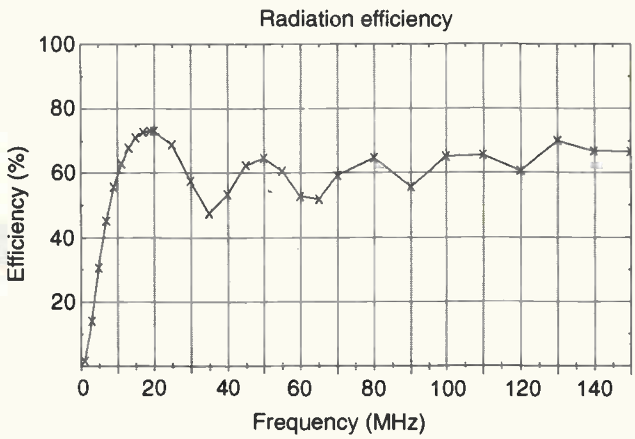


Fig. 7. Radiation efficiency of the loaded wide-band antenna example.

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2.5GHz power source

The fruits of research into applying microwaves to geological samples, this pulsed 2.5GHz power source is very inexpensive relative to laboratory sources since it uses mass-produced parts, as John Share and John Hakes demonstrate.

Dr John A.Share and Mr. John Hakes are with the Department of Earth Science at University of Liverpool

Generating significant power at microwave frequencies is inherently expensive. Producing 500W at 2.5GHz was, for us, prohibitively costly. Fortunately, mass production of a domestic appliance has resulted in 600W, 2.5GHz generators with integral power supplies being available for very little financial outlay. Second-hand units are readily available at even lower cost.

At first glance, a domestic microwave oven may not seem a likely candidate for conversion. One manufacturer suggested that the demise of the magnetron would be the inevitable result.

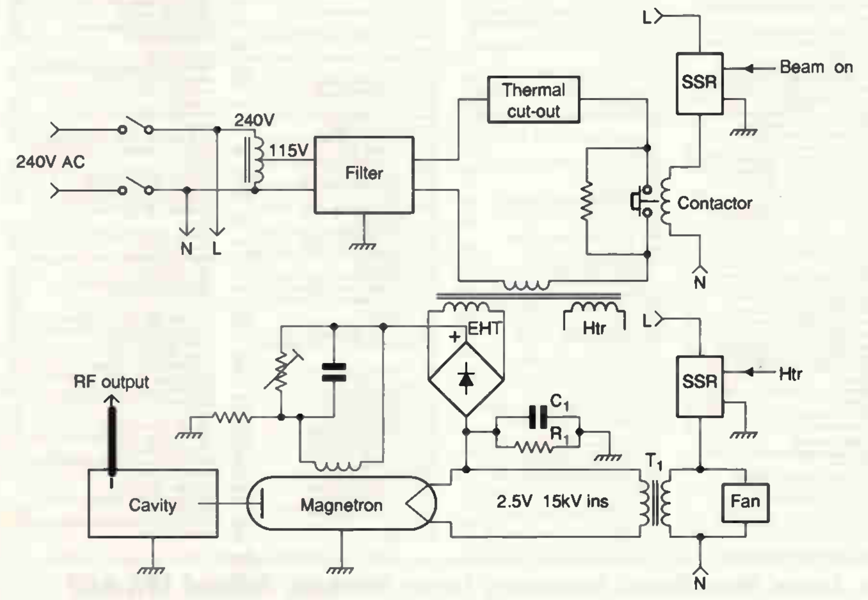
Essentially, the design of a domestic microwave oven consists of a magnetron, a cavity assembly, for the anode tuned circuit, an eht power supply interlocked with a heater supply and a blower. It also includes various control and timer circuits.

Operating frequency is governed by three factors – dimensions of the cavity, the magnetic field applied to the magnetron, and the actual value of the eht. Varying any one of these causes the frequency to alter. Significant changes will result in the magnetron ceasing to function.

Starting at component level without considerable expertise would be foolhardy. However, the conversion of a domestic microwave oven into a microwave power source is relatively straightforward.

Mechanical modification requires extracting the entire magnetron/cavity assembly from its domestic usage enclosure. This reveals that the oven space and the anode cavity are two separate items, and that there is an aperature between the two to allow coupling.

The first stage is to remove the oven with a hacksaw and fit a plate over the aperture. A sheet of thin metal and numerous self-tapping



This 2.5GHz power source is based on a reconfigured microwave oven. Components T_1, C_1, R_1 and the solid-state relays are part of the reconfiguration.

Warning

Very serious hazards exist both within this apparatus, due to the presence of lethal power supplies, and from biological damage caused by the microwave energy generated. Duplication of this research work must not be undertaken by anyone who does not fully understand the dangers of microwaves and very high voltages. Extreme caution must be exercised during development and operation. Access to a certificated microwave radiation detector is absolutely essential.

screws – no more than 2cm apart – will ensure a tight rf seal. The signal is extracted into RG8U by a quarter-wave E-probe, installed precisely on the centre line of the cavity at a distance of a quarter wavelength, i.e. 3.33cm, from the end wall.

Initial trials using N connectors and a probe, soft-soldered to the socket, resulted in numerous flashovers and carbon tracking. By drilling out the socket and arranging the plug so that the centre insulator and conductor of the RG8U coax passed directly into the cavity, far greater reliability was achieved. Trimming the projection to obtain maximum power output was then a very simple cut-and-try procedure.

A variable vane within the cavity was obviously intended by the manufacturer to compensate for manufacturing tolerances. This was reused to compensate for the presence of the new coupling.

In its original form, the heater winding was included on the eht transformer assembly and inevitably there was an indeterminate delay between operating the unit and the appearance of rf output. Our specific requirement was that the unit would produce bursts of rf under computer control for programmable periods of one to ten seconds duration.

An additional heater transformer, T_1, was required so that the rf output could be controlled by application of the eht. As with many high-frequency devices, the anode is external and is usually connected to the metalwork to assist cooling. This requires elevating the cathode by several kilovolts, and the heater supply must of necessity also be at several kilovolts, relative to ground. Suitable transformers are available as surplus; however, custom-built transformers are not out of the question.

Interlocked with a blower and thermal trip circuitry from the original unit, the heater supply remains active continuously while the eht can be controlled by solid-state switching of the transformer primary.

Interfacing to a pc-compatible used a commercial parallel i/o card and some custom-designed logic using 74-series devices. This operated the solid-state relays and provided interlocks and safety cutout features to provide automated software control.

Spectral purity of the output is not as horrendous as we first feared; in fact, it is surprisingly clean, but frequency modulation due to eht ripple was very evident. Adding smoothing, namely C_1 and R_1 resulted in a significant improvement – an 8μF paper capacitor and 1MΩ glass-encased high-tension resistor were available, but these values do not seem to be critical.

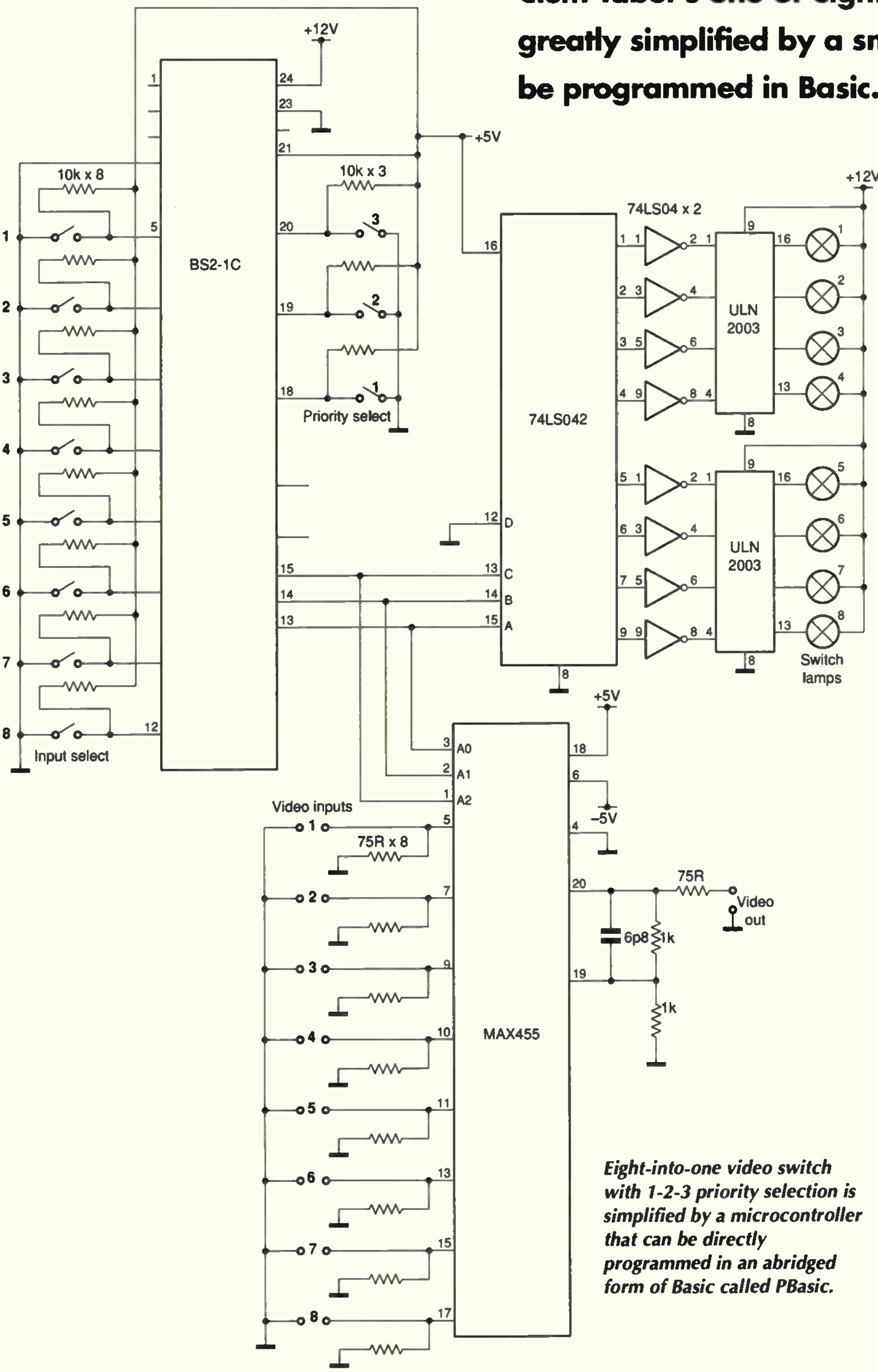
The purpose of this system was to excite spin waves in rock samples. For this, it is necessary to achieve maximum current coupling into the rock sample. A coaxial cavity was used to achieve this. Such passive devices offer exceptional selectivity – even when crudely made to modest workshop machining tolerances.

The system was operated without failure for more than two years, and has only been made redundant by advances in experimental techniques and knowledge of the magnetic structures of the rock samples themselves, necessitating higher frequencies of excitation. ■

Initial funding for this project was provided by the University of Liverpool RDF Fund, British Petroleum and the Natural Environmental Science Research Council.

VIDEO channel selector

Clem Tabor's one-of-eight video channel selector is greatly simplified by a small microcontroller that can be programmed in Basic.



Eight-into-one video switch with 1-2-3 priority selection is simplified by a microcontroller that can be directly programmed in an abridged form of Basic called PBasic.

Recently, I was asked to produce an eight-into-one video switcher with the added complication of three priority switches for inputs 1, 2 and 3. These select the associated input for as long as the switch is depressed. On release the output reverts to the previously selected input.

To construct such a system using hard-wired logic was daunting, so I decided to use a Basic Stamp *BS2-1C* microcontroller module to provide some intelligence.

The circuit diagram shows eleven of the Stamp's sixteen i/o lines used as inputs, taken low by the momentary contact input select and priority select switches. The three lines configured as outputs control the *MAX455* video switch and the ics driving the input selection indicator lamps. The video switch circuit was taken from a Maxim applications manual and the lamp driver components were used because they were to hand.

The *Pbasic* program is shown below. The program is split into several sub-routines. 'Loop' scans all the switches until one is closed. Then a jump is made to the routine shown at the end of the corresponding 'button' instruction. If switch 5 is pressed, the program jumps to 'SIX'. Binary 5 is placed on the three-bit output port and the number 6 is stored in memory location 0. A jump is then made back to 'loop', to wait for another switch closure.

If one of the switches 13, 14, or 15 is closed, the program jumps to the related 'PRIO' routine. In the case of switch 13 (priority 1), the jump is to 'PRIO 1'. Binary 0 is placed on the output port and the program returns to the start of 'PRIO 1' until the the switch is opened. At this point the number previously stored in memory location 0 is placed in variable 'last'. The 'if Last' lines ensure that the next jump is back to the routine which switched the last input selected before the priority switch was operated.

Readers familiar with *Pbasic*, will probably see ways of improving this program, but I think it illustrates that a 'jobbing' technician like me, with modest programming experience, can quickly produce quite complicated systems using a *Basic Stamp*. ■

Pbasic listing reads the eight push switches and selects the appropriate video channel, subject to the information on the priority-select switches.

```
'8 way switch driver. Binary outputs with three priority switches.
'Clem Tabor. 12-12-95.

last var nib    'declare variable "last"

goto ONE        'select output 1 on power up

loop:
button 0,0,200,100,b0,1,ONE        'detect button 0 & go to next stage
button 1,0,200,100,b1,1,TWO        'detect button 1 & go to next stage
button 2,0,200,100,b2,1,THREE      'detect button 2 & go to next stage
button 3,0,200,100,b3,1,FOUR       'detect button 3 & go to next stage
button 4,0,200,100,b4,1,FIVE       'detect button 4 & go to next stage
button 5,0,200,100,b5,1,SIX        'detect button 5 & go to next stage
button 6,0,200,100,b6,1,SEVEN      'detect button 6 & go to next stage
button 7,0,200,100,b7,1,EIGHT      'detect button 7 & go to next stage
button 13,0,200,100,b13,1,PRIO1    'detect priority button 0 etc
button 14,0,200,100,b14,1,PRIO2    'detect priority button 1 etc
button 15,0,200,100,b15,1,PRIO3    'detect priority button 2 etc

goto loop       'back to start if no button change

ONE:
low 8           '0 out, output 1
low 9
low 10
write 0,1       'store latest button press
goto loop       'back to start

TWO:
high 8          '1 out, output 2
low 9
low 10
write 0,2
GOTO loop

THREE:
low 8           '2 out
high 9
low 10
write 0,3
goto loop

FOUR:
high 8          '3 out
high 9
low 10
write 0,4
goto loop

FIVE:
low 8           '4 out
low 9
high 10
write 0,5
goto loop

SIX:
high 8          '5 out
low 9
high 10
write 0,6
goto loop

SEVEN:
low 8           '6 out
high 9
high 10
write 0,7
goto loop

EIGHT:
high 8          '7 out
high 9
high 10
write 0,8
goto loop

PRIO1:
low 8
low 9
low 10
button 13,0,200,100,b13,1,PRIO1   'detect priority
                                  'button 0 state & stay
if b13>0 then PRIO1               'there until it is released
read 0,last                       'read last button press
if last=1 then ONE                'if butt. 0 last,
                                  'the back to ONE
if last=2 then TWO
if last=3 then THREE
if last=4 then FOUR
if last=5 then FIVE
if last=6 then SIX
if last=7 then SEVEN
if last=8 then EIGHT
goto loop

PRIO2:
high 8
low 9
low 10
button 14,0,200,100,b14,1,PRIO2   'detect priority
                                  'button 2 state & stay
if b14>0 then PRIO2               'there until it is released
read 0,last                       'read last button press
if last=1 then ONE
if last=2 then TWO                'if button 1 last,
                                  'then back to TWO
if last=3 then THREE
if last=4 then FOUR
if last=5 then FIVE
if last=6 then SIX
if last=7 then SEVEN
if last=8 then EIGHT
goto loop

PRIO3:
low 8
high 9
low 10
button 15,0,200,100,b15,1,PRIO3   'detect priority
                                  'button 3 state & stay
if b15>0 then PRIO3               'there until it is released
read 0,last                       'read last button press
if last=1 then ONE
if last=2 then TWO
if last=3 then THREE              'if button 2 last,
                                  'then back to Three
if last=4 then FOUR
if last=5 then FIVE
if last=6 then SIX
if last=7 then SEVEN
if last=8 then EIGHT
goto loop
```

Light reading for COM 1

Edward Buckley Stamp-based subsystem for PCs reads the value on a dedicated light-to-frequency converter and transmits it to the pc via an RS232 link.

In the January 1996 edition of *Electronics World* an article by Claus Kuhnel showed how easy it is to use the *TSL230* with the *Basic Stamp 1* microcontroller. The *Stamp* measured the width of one output pulse and converted this into a power flux measurement.

In application notes, Texas Instruments discusses two main measurement systems when using the *TSL230*. One is fast light measurement, the other longer term measurements offering a higher resolution. Slower measurements methods also incorporate the ability to reject spurious effects such as tube flicker.

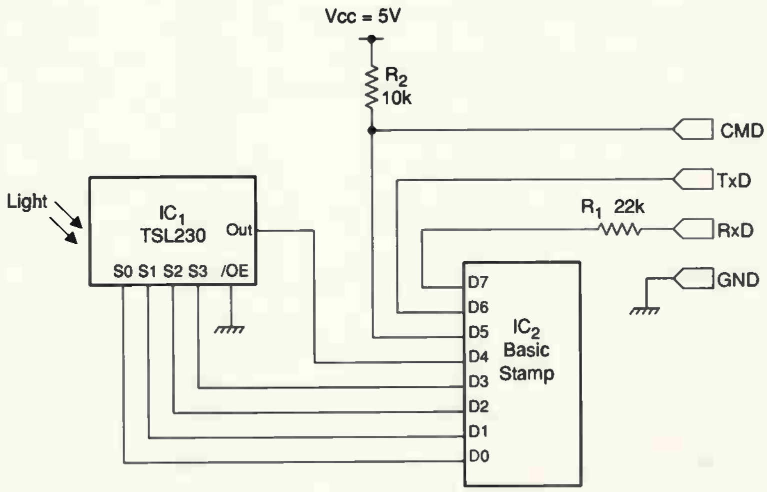
Claus's article, based on the *Stamp I,* used the fast method – measuring the width of the IC's output pulse. The newly introduced *Stamp 2* microcontroller has the benefit of several commands additional to those available on *Stamp I.*

One of the new commands is COUNT. *Stamp 2* 'counts' the pulses coming from the

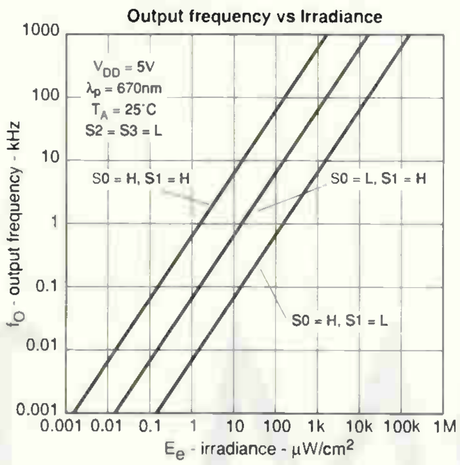
Programming the Stamp

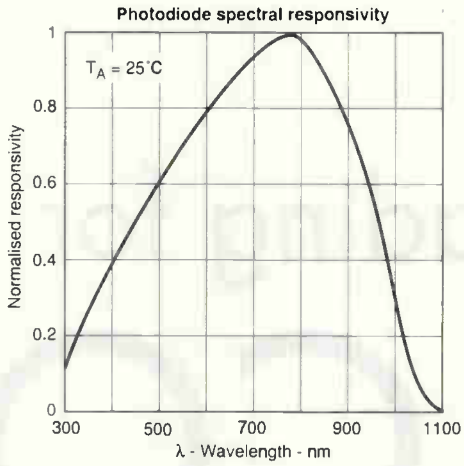
Basic Stamps comprise a PIC microcontroller, eeprom, oscillator and regulator. The PIC is incorporates a Basic interpreter. Programs similar to the ones shown in these articles are stored in eeprom, making them non-volatile and allowing them to be changed within seconds as often as required.

Programs are constructed on a pc using special low-cost text editor software. Once completed, they are downloaded into the Stamp via a COM port in the case of Basic Stamp 2, or and LPT port for the Basic Stamp 1.



Interfacing to the pc via a COM port this circuit, together with the software shown, forms a subsystem for sensitive light measurement.





Technical support
Basic Stamp is distributed by Milford Instruments at Milford House, 120 High Street, South Milford, Leeds LS25 5AQ, tel. 01977 683665, fax 01977 681465.

Stamp2 program for reading the TSL230 light-to-frequency chip.

```
'Stamp2 first determines sensitivity range using Pulsin command
'Then switches to COUNT command for high- resolution measurement.
'The Stamp automatically switches sensitivity and frequency to give
'a power flux measurement range from 10nW to 100mW/cm².

'Variables
sensitivity var byte  'TSL230 sensitivity range
freq_range var byte   'Frequency scaling ratio
flux var word         'Light power flux result,16 bits
puls_in var word      'result from Pulsin command, 16 bits
cou_nt var word       'result from Count command, 16 bits

'Constants
  so        con 0    'so connected to pin 0
  s1        con 1    's1 connected to pin 1
  s2        con 2    's2 connected to pin 2
  s3        con 3    's3 connected to pin 3
  CE        con 4    'Chip-enable pin (active when low)
  data_in   con 5    'data input line to pin5

'Initialisation of pins
'Pins read right to left

  dirl=%00011111  '5 outputs and 3 inputs

'First section determines the sensitivity
'and frequency range for optimum operation.

Start:
  'Start at x1 sensitivity
  outi=%00001001              'F/10 and x1 sensitivity, chip enabled
  sensitivity =1
  pulsin data_in,1,puls_in    'wait for high pulse-2µs increments
  if puls_in<300 AND puls_in<>0 then measure   'if true goto measure

'Increase sensitivity to x10
sensitivity =10
outi=%00001010

pulsin data_in,1 ,puls_in     'wait for high pulse again if
puls_in<3000 AND puls_in<>0 then measure      'if true measure

'Increase sensitivity to x100
sensitivity =100
outl=%00001011
pulsin data_in,1,puls_in      'wait for high pulse again

if puls_in<>0 then measure    'if true, measure

'Increase output frequency to F/2
freq_range=5
outl=%00000111

'Now go to the COUNT command.
'Count gives a higher resolution than the pulsin command and rejects
'flicker effects but takes longer.
'500ms count period taken to avoid overflow in count variable

Measure:

  count data_in,500,cou_nt    'Measure number of pulses in 500ms

'Now sort out the units before writing to the screen!
'Due to the high dynamic range of the TSL230 need to re-arrange the
'power flux calculation depending on the sensitivity/freq_range used
'otherwise may result in overflow in "flux" variable

  if sensitivity>1 then skip1
  flux=cou_nt/10*2/10*13/100         'convert count to flux
  'note maths sequence to prevent overflow to variable
  debug "Power Level=",dec flux,"milliwatts/cm2" 'Display answer
  goto start

skip1:          if sensitivity>10 then skip2
  flux=cou_nt*13/10*2/10
  debug "Power Level=",dec flux,"microwatts/cm2"
  goto start

skip2:          flux=cou_nt*2*13/freq_range
  debug "Power Level=",dec flux,"nanowatts/cm2"
  goto start
```

TSL230 over a period of time and uses this number to calculate the power flux. Being based on a large number of pulses, the result is now smoothed of spurious signals.

Both COUNT and PULSIN are used in this *Stamp 2* version of Claus's article. The command PULSIN helps quickly find the best sensitivity and output frequency range, thereafter COUNT returns a smoothed result.

Variables and constants are defined in the usual way and the output pins set to produce the default sensitivity, namely ×1, and a frequency division of *f*/10.

The *Stamp 2* takes the first measurement, at Pulsin in the program, following which it decides whether to increase the sensitivity or to progress on to take a full measurement. Conditional statement puls_in<300 checks whether a better result could be obtained by increasing the sensitivity and the puls_in<>0 checks for no result at all. If there is no input, a very low light level indeed is being measured. No input only occurs when the *Stamp 2* has timed out at 131ms without any reading, resulting in the writing of 0 to the variable.

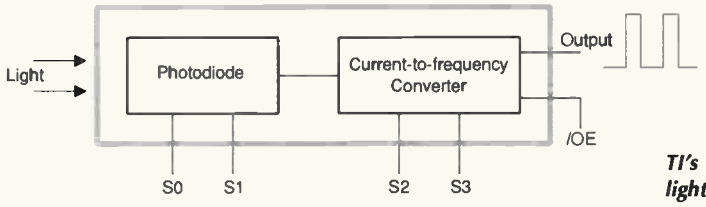
The above is repeated to check whether a jump to ×100 sensitivity would be beneficial. If very low light levels are being experienced, a further attempt to optimise the measurement conditions is considered by the *Stamp* by increasing the *TSL230*'s output frequency from *f*/10 to *f*/2. This 'hunting' for a suitable operating point is indicated on the accompanying frequency/flux diagram for the *TSL230*.

Once 'optimum' conditions are in place, the *Stamp* proceeds to take the actual power flux measurement using the Count command. It measures the number of cycles from the *TSL230* in a 500ms period.

The *Stamp 2* can measure up to a maximum of 65 536 cycles at a maximum frequency of 125kHz. The start-up conditions/hunting procedure and count period should ensure the count is readable by *Stamp 2*.

Finally the actual power flux is calculated and written to the computer screen using the Debug command. A *Stamp* standard chunk of code may also be introduced here to send the result to either an liquid-crystal display, remote computer or store the result in electrically erasable prom.

The high dynamic range of the *TSL230* and the 16-bit integer maths capability of the *Stamp* mean that some care has to be taken with the arithmetic to avoid either overflows or loss of accuracy due to rounding errors. ■



S1	S0	Sensitivity
L	L	Power down
L	H	1x
H	L	10x
H	H	100x

S1	S0	FO scaling (divide-by)
L	L	1
L	H	2
H	L	10
H	H	100

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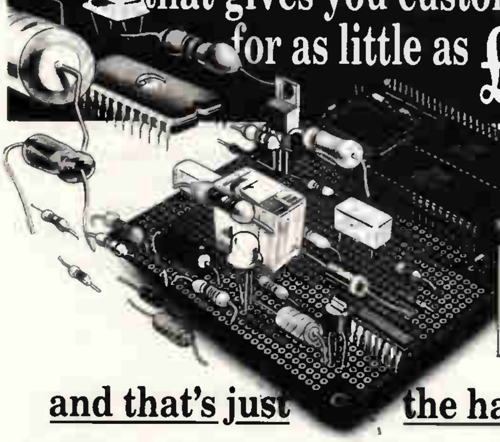

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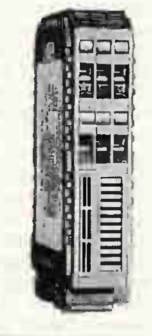

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CIRCLE NO. 122 ON REPLY CARD

With this novel packaging, the components of any dual-component adhesive can be mixed and dispensed with ease and safety.

SOLUTION FOR STICKY PROBLEMS

Anyone who has mixed high-performance epoxies will know that any glue finding its way on to a surface where it is not wanted can be a real nuisance to remove. With this new concept, the epoxy is mixed in a sealed bag and applied from the same bag via a snip on the corner. The result is thorough, precise mixing with much less mess.

It is a fact of life that some of the best adhesives are the trickiest to handle. They often have two components which must be mixed thoroughly in the correct proportions for optimum performance – not too easy with viscous liquids. Then there is the problem of delivery. The correct amount of adhesive must reach the repair surface without picking up contamination or sticking to everything else on the way – including your fingers.

These difficulties are overcome by a new approach where choice of adhesive is linked with a custom packaging and delivery system, designed and developed by Resintech.

A typical *TwinPack* sachet is shown in the photograph, top left. It contains 5ml of an epoxy resin adhesive. Contents of the sachet could also be polyurethane or silicones, with applications from adhesives to sealants, potting compounds and encapsulants. Pack sizes can vary from 1 to 1000ml with resin to hardener ratios easily varied.

Components are accurately preweighed and kept free from moisture or other contamination right up to the time of use. This leads to quicker, easier, cleaner safer installations compared with conventional packaging methods. A more reliable bond is generated time after time.

High-strength adhesive

This epoxy sample is a high strength adhesive with a work life of 8-10 minutes. It is suitable for most substrates including metal, wood, china, ceramics, glass and hard plastics. Make sure surfaces to be bonded are clean, dry and free from grease, dust or dirt.

Remove the red clip and bar, which until now have acted as a barrier to separate the resin and hardener. Now, the innovative part of the packing is revealed as the sachet becomes one pouch, allowing the two components to be freely mixed. Contents still remain sealed during this process against moisture, air or particle contamination.

To help mix, use the red bar to push components together. Kneed the pouch with your fingers until a uniform mix is obtained. This takes about a minute.

The plastic pouch is made from a strong, tough laminate which will withstand vigorous handling. To apply the adhesive, simply cut off a corner of the pack and squeeze out the contents. The corner cut can be varied to provide the ideal bead size for the repair area. Allow adhesive to set before handling which can be 20 minutes to 2 hours depending on the application.

Any unused residue still in the pack will cure to a solid and in this form does not constitute a waste hazard,

In engineering, prototyping, maintenance and servicing, the convenience of small quantities, ease of handling and storage are most important. The *TwinPack* can be carried around in your pocket, bag or toolkit, at home or at work, without fear of leakage.

No associated equipment is needed. *TwinPack* is self-contained and ready for use.

And other adhesives?

Given the correct choice of adhesive, then the packaging and delivery system have a major impact on factors such as convenience or installed cost. The mess and hassle associated with mixing from separate tubes or tubs can be reduced to a minimum by using a clip separated sachet – *TwinPack*, or a double barrelled cartridge called *DuoSyringe*. These two systems are largely interchangeable with the *DuoSyringe* preferred for non-sag sealants and *TwinPack* for easy spread adhesives.

For industrial applications benefits come through productivity gains, cost saving, improved quality and safety. These arise because quantities of adhesive and mix ratios are precisely controlled. Waste is minimised, and anyone handling the adhesive is protected from exposure to hazardous material.

TwinPack can be produced in almost any size and quantity to match production requirements. There is a choice of materials for most adhesive, potting encapsulation or sealant applications.

Alternative delivery

As an alternative to the *TwinPack*, Resintech also offers the *DuoSyringe* . Here, the resin and hardener are contained in a side-by-side syringe. The material is discharged through a mixing nozzle directly on to the substrate, using a dual action dispenser gun.

DuoSyringe has specific application in handling adhesive sealants, where high degree of thixotropy is designed in to the system to enable a non-slumping bead to be formed. A typical *DuoSyringe* volume is 50ml. Resin to

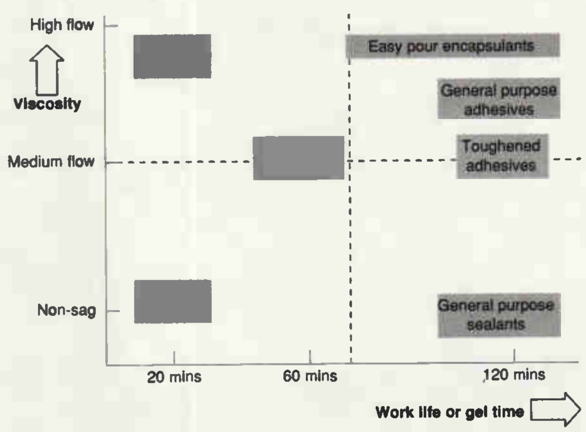


Fig. 1. Relative flow performance of available epoxy resin formulations.

*Due to varying regulations in different countries regarding gifts with magazines, this offer is limited to UK readers only. However, Resintech has generously offered to supply a free sample to any overseas reader, on receipt of a written request, while stocks last.

hardener ratios can be varied up to 10:1 using different barrel diameters.

Selecting the right adhesive

Determining the right adhesive for the job is very much depends on the substrates to be handled and performance specification. More often than not the choice is subject to some trial and error before the right balance between bond strength, production requirements and cost can be established.

Some guidelines are offered here to help narrow the choice and these are backed up by a materials design and sample service from Resintech for those all important preliminary tests and trials.

Process variables

Figure 1 shows relative flow performance of available resin formulations at the incipient point of mixing. Once reaction starts, at a rate which can be varied according to desired work life, viscosity increases in all systems until first a gel and then a solid state is reached.

A vertical bond would need a low flow adhesive, a product in the lower half of the chart. A potting compound for intricate components may need a high flow material from the upper half of the chart. Printed-circuit board mounting, or other sequential tasks, would need a fast cure product from the left hand side of the table.

Work life is defined as the approximate length of time a 25g standard mass of adhesive requires to begin to harden – known as the gel-point – at room temperature. The rate of reaction is very dependent on temperature. Oven cure is widely practised. As a rule of thumb for each 10°C rise in temperature work life or cure time will halve.

As a note of caution, an accelerated reaction may influence final properties.

Performance matrix

It is common to measure the mechanical performance of a bond in terms of shear strength and peel strength. Requirements will vary according to application. The higher up the chart, **Fig. 2**, the greater the shear strength, and the stronger the bond. The further to the right indicates more flexibility. Toughened systems have a good balance of both properties.

Environmental factors such as fluids, heat and sunlight may dictate that a different chemical system be chosen.

Resintech can offer custom formulation to achieve the desired balance of properties from any system. In addition to the options mentioned previously these might include high clarity for optical systems, colour matching,

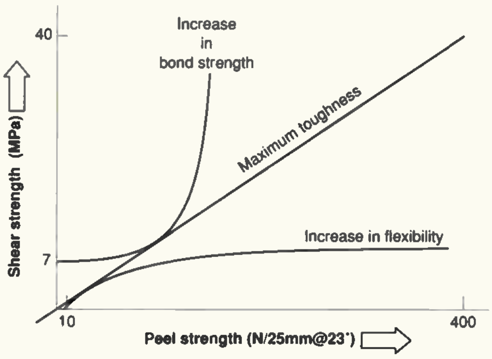


Fig. 2. Performance balance between various epoxies highlights the importance of choosing the right product for the job.

flame retardancy, zero halogen components, high purity electrical grades and more.

When combined with specific packaging and delivery systems such as the *TwinPack* or *DuoSyringe* a high performance adhesive can be used with minimum mess and hassle. ■

Resintech, manufacturers of a wide range of adhesives, is at Horcott, Fairford, Glos GL7 4BX, on 01285 712755 fax 713036.

Filters based on frequency-dependent negative resistors offer the performance of *LC* filters but without the bulk and expense. And component intolerance is lower than for other lf filters types. Ian Hickman explains.

FILTERS using negative resistance

When it comes to filters, it's definitely a case of horses for courses. At rf the choices are limited; for tunable filters covering a substantial percentage bandwidth, it has to be an *LC* filter. If the tunable elements are inductors, you have a permeability tuner; alternatively, tuning may use variable capacitance, or varactors. Fixed frequency filters may use *LC*s, quartz crystals, ceramic resonators or surface-acoustic-wave (SAW) devices, while at microwaves, the 'plumbers' have all sorts of ingenious arrangements.

At audio frequencies, *LC* filters are a possibility. However, the large values of inductance necessary are an embarrassment, having a poor Q and temperature coefficient, apart from their bulk and expense.

One approach is to use '*LC*' circuits where the inductors are active circuits simulating inductance. There is a number of these, and **Fig. 1** is an example. For high-pass filters, synthetic inductors with one end grounded, Fig. **1a**), suffice, but for low-pass applications, rather more complicated circuits simulating floating inductors are required, Fig. **1b**).

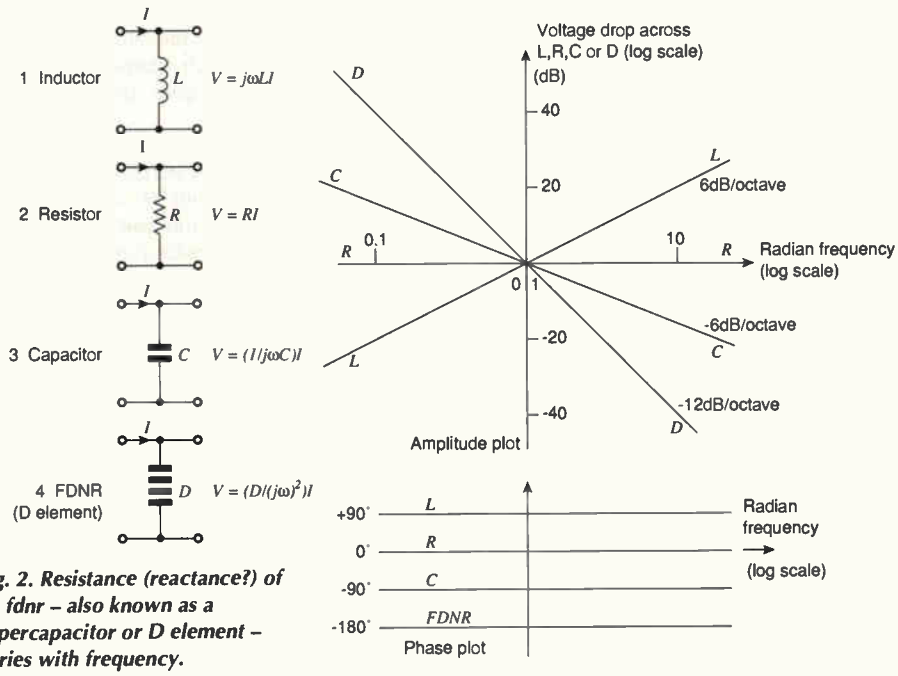


Fig. 2. Resistance (reactance?) of an fdnr – also known as a supercapacitor or D element – varies with frequency.

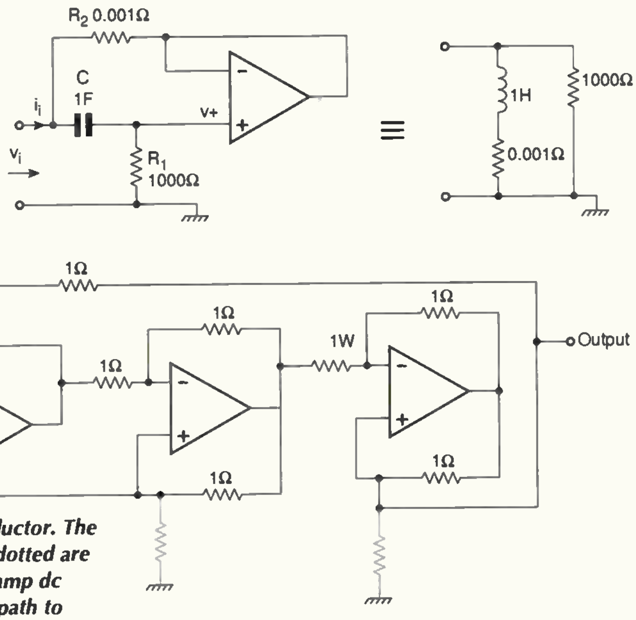


Fig. 1. Synthetic inductors.
a) A 1H 'inductor' with one end grounded. Q is 10 at 0.00159Hz, and proportional to frequency above this. Below, it tends to a 0.001Ω resistor, just like the corresponding real inductor.

b) Floating synthetic 1H inductor. The high-value resistors shown dotted are necessary to define the op-amp dc conditions if there is no dc path to ground via input and output.

More recently, switched-capacitor filters have become available, offering a variety of filter types, such as Butterworth, Bessel and Elliptic. These vary in complexity up to eight or more poles.

For narrow band-pass applications, a strong contender must be the *N*-path filter. This scheme uses switched capacitors but is not to be confused with switched capacitor filters; it works in an entirely different way. However, both switched capacitor and *N*-path filters are time-discrete circuits, with their cut-off frequency determined by a clock frequency. Hence both types need to be preceded by an anti-alias filter – and usually followed by a low-pass filter to suppress clock frequency hash. That is the downside: the upside is that tuning is easy – simply change the clock frequency.

The cut-off or centre frequency of a switched capacitor filter scales with clock frequency, but the bandwidth of an N-path filter does not.

Where a time-continuous filter is mandatory, various topologies are available, including Salen and Key and Rausch. An interesting and useful alternative to these and to LC filters, with either real or simulated inductors, is the fdnr filter, which makes use of frequency-dependent negative resistances.

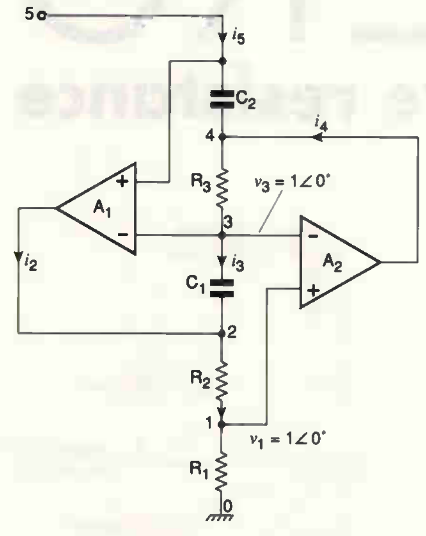


Fig. 3a). Circuit diagram of an fdnr. If V_1 is the voltage at node 1, etc., then V_1=V_3=V_5. Also, i_1=i_2+i_3 and i_3=i_4+i_5.

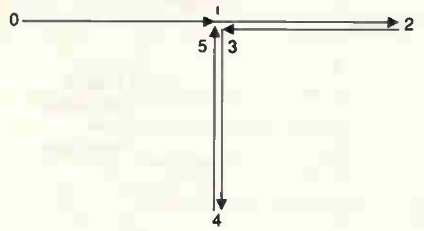


Fig. 3b). Voltage vector diagram for (a) when R_1=R_2=R_3=R, C_1=C_2=C and f=$^1/_{2\pi}$CR.

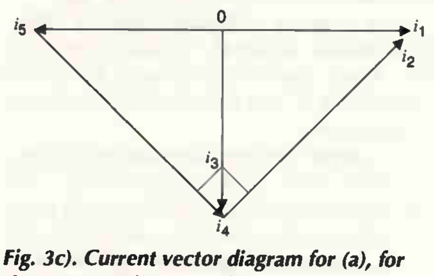


Fig. 3c). Current vector diagram for (a), for the same conditions as (b).

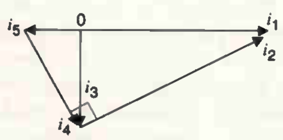


Fig. 3d). As (c) but for f=$^1/_{4\pi}$CR. Note that i_2 and i_4 are always in quadrature.

What is an fdnr?

A negative resistance is one where, when you take one terminal positive to the other, instead of sinking current, it sources it – pushes current back out at you.

As the current flows in the opposite direction to usual, Ohm's law is satisfied if you write I=E/–R, indicating a negative current in response to a positive potential difference, or pd. This would describe a fixed, or frequency independent, negative resistance. But fdnrs have a further peculiarity – their resistance, reactance or impedance, call it what you will, varies with frequency. Just how is illustrated in **Fig. 2**.

With inductors, the voltage leads the current by 90°; with capacitors, it lags by 90°. Combining these with resistive terminations, where the voltage leads/lags the current by 0°, you can make filters. Such filters may be high-pass, band-pass, low-pass, or whatever you want.

It was pointed out in a famous paper[1] that, by substituting for L, R (termination) and C in a filter, components with 90° more phase shift and 6dB/octave faster roll than the L, R and C, exactly the same transfer function could be achieved.

Referring to Fig. 2, L, R and C are replaced on a one-for-one basis by R, C and fdnr respectively. An fdnr can be realised with resistors, capacitors and op-amps, **Fig. 3**.

So how does an fdnr work?

Analysing Fig. 3 provides the answer. Looking in at node 5, you see a negative resistance; but what is its value?

First of all, note that the circuit is dc stable. At 0Hz, where you can forget the capacitors, A_2 has 100% negative feedback via R_3, and its non-inverting input is referenced to ground.

Likewise, A_1 has its non-inverting input referenced to ground, assuming there is a ground return path via node 5. It also has 100% negative feedback; A_2 is included within this loop.

The clearest and easiest way to work out the ac conditions is with a vector diagram; just assume a voltage at node 1 and work back to the beginning. Thus in Fig. 3, assume that $V_{1,0}$, i.e. the voltage at node 1 with respect to node 0 or ground, is 1V ac, at a frequency of 1 radian per second (1/(2π) or 0.159Hz), and that R_1=R_2=R_3=1Ω, C_1=C_2=1F. Thus the voltage at node 1 is represented in Fig. **3b)** by the line from 0 to 1, of unit length, the corresponding current of 1A being shown as i_1 in Fig. **3c)**.

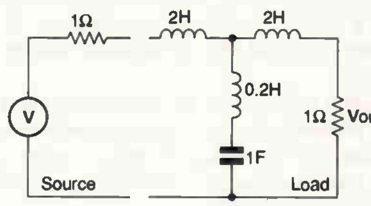
Straight away, you can mark in, in b), the voltage $V_{2,1}$, because R_1=R_2, and node 1 is connected only to an (ideal) op-amp which draws no input current. So $V_{2,1}$ equals $V_{1,0}$ as shown.

But assuming A_2 is not saturated, with its output voltage stuck hard at one or other supply rail, its two input terminals must be at virtually the same voltage. So now $V_{3,2}$ can be marked in, taking one back to the same point as node 1. Given $V_{3,2}$, the voltage across C_1, whose reactance at 0.159Hz is 1Ω, the current through it can be marked in as i_3 in Fig. 3c).

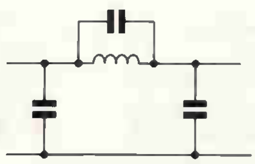
Of course, the current through a capacitor leads the voltage across it, and i_3 is accordingly shown leading the voltage $V_{3,2}$ by 90°. Since i_1=i_2+i_3, i_2 can now be marked in as shown.

As i_3 flows through R_3, $V_{4,3}$ can now be marked in, and as the voltages at nodes 5 and 3 must be equal, $V_{5,4}$ can also be marked in. The current i_5 through C_2 (reactance of 1Ω) will be 1A, leading $V_{5,4}$ as shown. Finally, as i_3=i_4+i_5, i_4 can be marked in, and the voltage and current vector diagrams (for a frequency of $^1/_{2\pi}CR$) are complete.

The diagrams show that $V_{5,0}$ is 1V, the same as $V_{1,0}$, but i_5 flows in the opposite direction to i_1 – the wrong way for a positive resistance. Fig. 3d) shows what happens at f=$^1/_{4\pi}CR$, half the previous frequency. Because the reactance of C_1 is now 2Ω, i_3 is only 0.5A, and therefore $V_{4,3}$ is only 0.5V. Now, there is only $^1/_2$V ($V_{5,4}$) across C_2, but its reactance has also doubled. Therefore i_5 is now only 0.25A; not only



Figs. 4a), above, and b), below. A low component count elliptic low-pass filter with a minimum attenuation of 36dB from twice the cut-off frequency upward, the price being as much as 1dB pass-band ripple. The minimum capacitor design of 4a) is more convenient than 4b) for conversion to an fdnr filter.



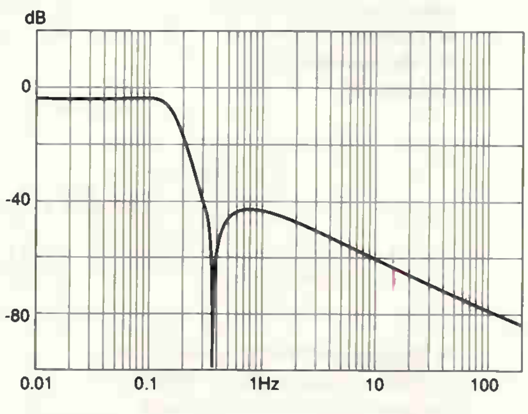


Fig. 4c). The frequency response of the filter.

is the current negative (a 180º phase shift), it is inversely proportional to the square of the frequency, as shown for the fdnr in Fig. 2.

Pinning down the numbers

Looking in at node 5, then, appears like a -1Ω resistor at 0.159Hz, but you need to know how this ties up with the component values. The values of the vectors can be marked in, on Figs. 3b) and 3c), starting with $V_{1,0}$=1V. Then $V_{2,1}=R_2/R_1$, and $V_{3,2}=-R_2/R_1$. It follows that $i_3=(-R_2/R_1)/(1/j\omega C_1)=-j\omega C_1.R_2/R_1$. Voltage $V_{4,3}=R_3 i_3=-j\omega C_1.R_2.R_3/R_1$, and $V_{5,4}=-V_{4,3}$.

So $i_5=-V_{4,3}/(1/j\omega C_2)=j\omega C_1 j\omega C_2.R_2.R_3/R_1$. Looking in at node 5 the resistance is $V_{5,0}/i_5=V_{1,0}/i_5$, where $V_{1,0}$=1V. So finally the fdnr input looks like,

$$fdnr=R_1/(j\omega C_1 j\omega C_2.R_2.R_3)$$
$$=-R_1/(\omega^2.C_1.C_2.R_2.R_3) \qquad (1)$$

With 1Ω resistors and 1F capacitors, this comes to just −1Ω at omega=1rad/s, or 0.159Hz. To get a different value of negative resistance at that frequency, clearly any of the *R*s or *C*s could be changed to do the job, but it is best to keep all the *R*s equal – at least roughly – and the same goes for the *C*s.

As a cross check on equation (1), note that it is dimensionally correct. The units of a time-constant *CR* are seconds, while the units of frequency are 1/seconds, be it cycles or radians per second. Thus the units in the denominator cancel out and, with a dimensionless denominator, the expression has the units of the numerator R_1, which is ohms.

A practical example

Designing an fdnr filter starts with choosing an *LC* prototype. Consider a simple example – a low-pass filter with the minimum number of components, which must reach an attenuation of 36dB at little more than twice the cut-off frequency. This is a fairly tall order, but a three-pole elliptic filter will do the job, if you allow as much as 1dB pass-band ripple.

A little experimentation with a CAD program came up with the design in **Fig. 4a**). This has nice, round component values, although the cut-off frequency is just a fraction below the design aim of 1 radian per second, but never mind, it will do for starters.

If you were designing an *LC* filter as such, you would certainly choose the π section of Fig. **4b**), rather than the tee section, as the π section is the minimum inductor version. But for an fdnr filter, the minimum capacitor version is preferable, as the *C*s become fdnrs (fairly complicated), whereas the *L*s become *R*s and are therefore cheap and easy.

But before passing on to consider the fdnr, note that the computed frequency response of the normalised 1Ω impedance *LC* filter is as shown in Fig. **4c**).

The low frequency attenuation shows as 6dB rather than 0dB. This is because the 1Ω impedance of the matched source (a 2V emf ideal generator behind 1Ω) is considered here as part of the filter – not as part of the source. To the 2V generator emf, which is what the CAD program models as the input, the source and load impedance appear as a 6dB potential divider.

The fdnr version of the filter is shown in **Fig. 5**. Not only do the *L*s become *R*s and the *C*s fdnrs, but the source and terminating resistors become capacitors. In an *LC* filter, the source and terminating resistors would usually be actually part of the source and load respectively. But an fdnr filter at audio frequencies will be driven from the 'zero' output impedance of an op-amp and feed into the nearly infinite impedance of another. As a result, you must provide the terminations separately if you want the response to be the same as the prototype *LC* filter.

In Fig. 5, the inductors have been replaced with resistors on an ohm-per-henry basis, and the *R*s and *C*s converted to *C*s and fdnrs similarly. As it happens, the required fdnr value is −1Ω, so values of 1Ω and 1F in the circuit of Fig. 3 will do the job.

If one had used the tabulated values for a 1dB ripple, 35dB A_s three-pole filter, e.g. from Ref. 2, **Fig. 6**, the required value of C_2 in the tee-section version, would have been 0.865F.

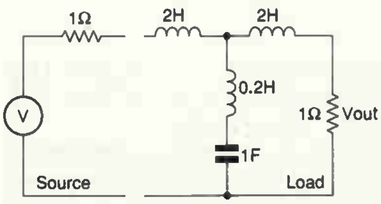
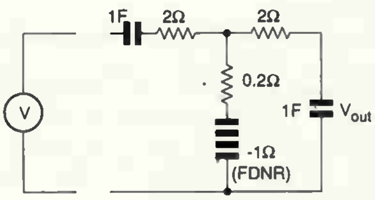


Fig. 5. An fdnr version of the low-pass filter.



Accordingly, from equation (1), R_1 in Fig. 3 would become 0.865Ω, or you could change R_2 and/or R_3 to achieve the same effect. Alternatively, you could scale C_1 and/or C_2, but it is best to leave them at 1F – the reason for this will become clear later.

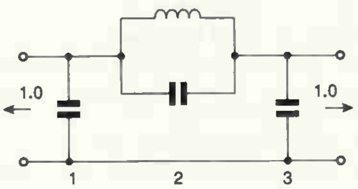
Having arrived at a 'normalised' fdnr filter design, i.e. one with a 0.159Hz cut-off frequency, the next step is to denormalise it to the desired cut-off frequency – let's say 10kHz in this case.

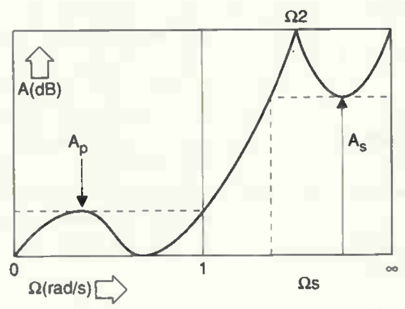
There is no need to change the *R*s at this stage, but to make the fdnr look like −1Ω, or −0.865, or whatever, at 10kHz, the capacitor values must be divided by 2π times 10,000. And since the termination capacitors must also look like 1Ω at this frequency, they must be scaled by the same ratio.

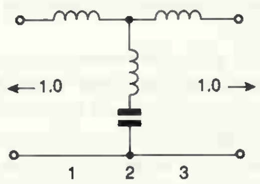
You now have a filter with the desired response and cut-off frequency, but the com-

Fig. 6. Tabulated normalised component values for three-pole 1dB pass-band ripple elliptic filters with various values of A_s at Ω_s. Ω here means the same as ω elsewhere in the article.

Ω_s	A_s (dB)	C_1	C_2	L_2	Ω_2	C_3
1.295	20	1.570	0.805	0.613	1.424	1.570
1.484	25	1.688	0.497	0.729	1.660	1.688
1.732	30	1.783	0.322	0.812	1.954	1.783
2.048	35	1.852	0.214	0.865	2.324	1.852
2.418	40	1.910	0.145	0.905	2.762	1.910
2.856	45	1.965	0.101	0.929	3.279	1.965
Ω_s	A_s (dB)	L_1	L_2	C_2	Ω_2	L_3







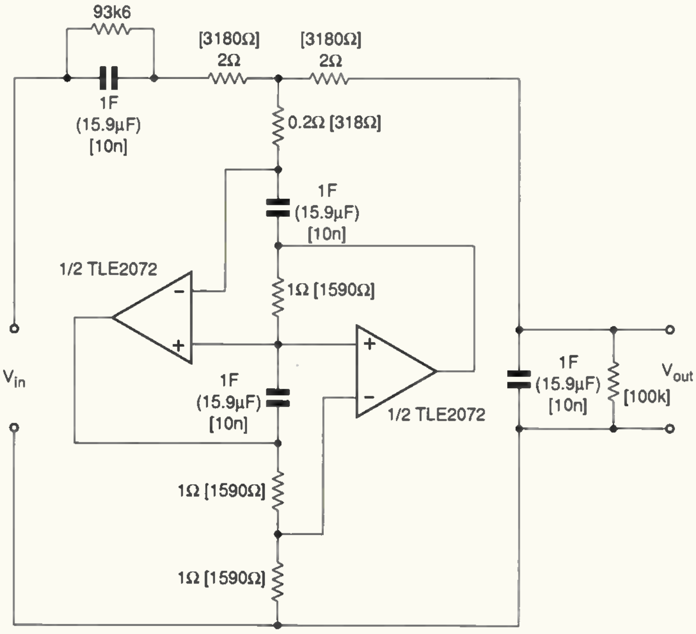
ponent values shown in round brackets in **Fig. 7**, are a little impractical. This is easily fixed, by a further stage of scaling.

Since resistors are more easily obtainable in E96 values and 1% selection tolerance, it pays to scale the 15.9μF capacitors to a round value, such as 10nF. So all impedances must be increased by this same ratio: N=1590 – the resistors multiplied by N and the capacitors divided by N.

Conveniently, the *C*s in the fdnr are the same value as the terminating capacitors, if, as recommended, any change in the required normalised fdnr negative resistance was effected by changing the *R* values only. The resultant practical component values are shown in square brackets in Fig. 7.

One peculiarity of an fdnr filter is due to its use of capacitive terminations. The impedance of these varies with frequency and, notably, becomes infinite at dc (0Hz). Thus any practical fdnr filter would have infinite insertion loss at this frequency.

This is remedied by connecting resistors in parallel with the terminating capacitors, to determine the 0Hz response. They are shown in Fig. **7a)** and have been chosen, taking into account the two 3,180Ω resistors, to provide 6dB attenuation at 0Hz. This is done to match the filter's pass-band 6dB loss. With the addition of these, Fig. 7a) is now a practical, fully working low-pass filter, the computed frequency response of which is shown in Fig. **7b)**.



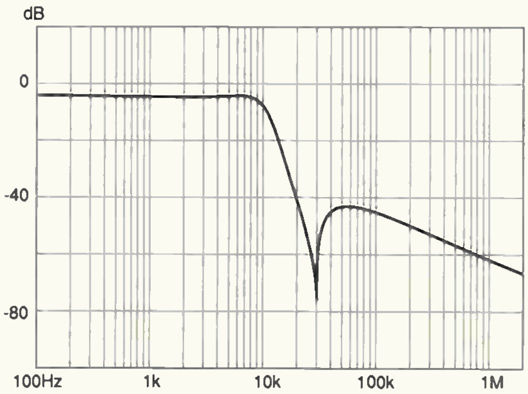


Fig. 7a). Complete fdnr filter with 10kHz cut-off frequency. The values in round brackets are a little impractical, but are easily scaled to more sensible values in square brackets.
Fig. 7b). Computed frequency response of the above filter. The cut-off frequency (at –1dB) is a shade below the intended value, as was that in Fig. 4c).

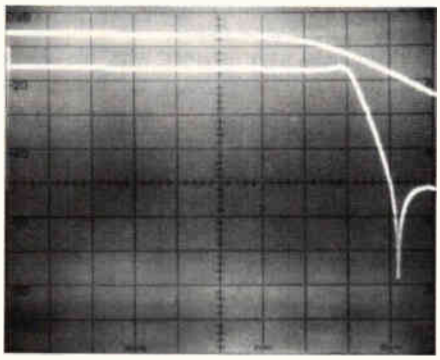


Fig. 8. Actual frequency response of the circuit of Fig. 7. The horizontal scale is logarithmic frequency, the left-hand vertical being 20Hz, the third, sixth and ninth vertical graticule lines representing 200Hz, 2kHz and 20kHz respectively. Horizontal graticule lines are at 10dB intervals.
Upper trace, generator reference level top of screen, representing the source emf. This trace was recorded with the shunt leg of the filter open circuited with the 318Ω resistor removed. Lower trace, response of complete filter withthe 318Ω resistor replaced. Reference level has been moved dowan one graticule division for clarity.

Log sweeps and IF bandwidths

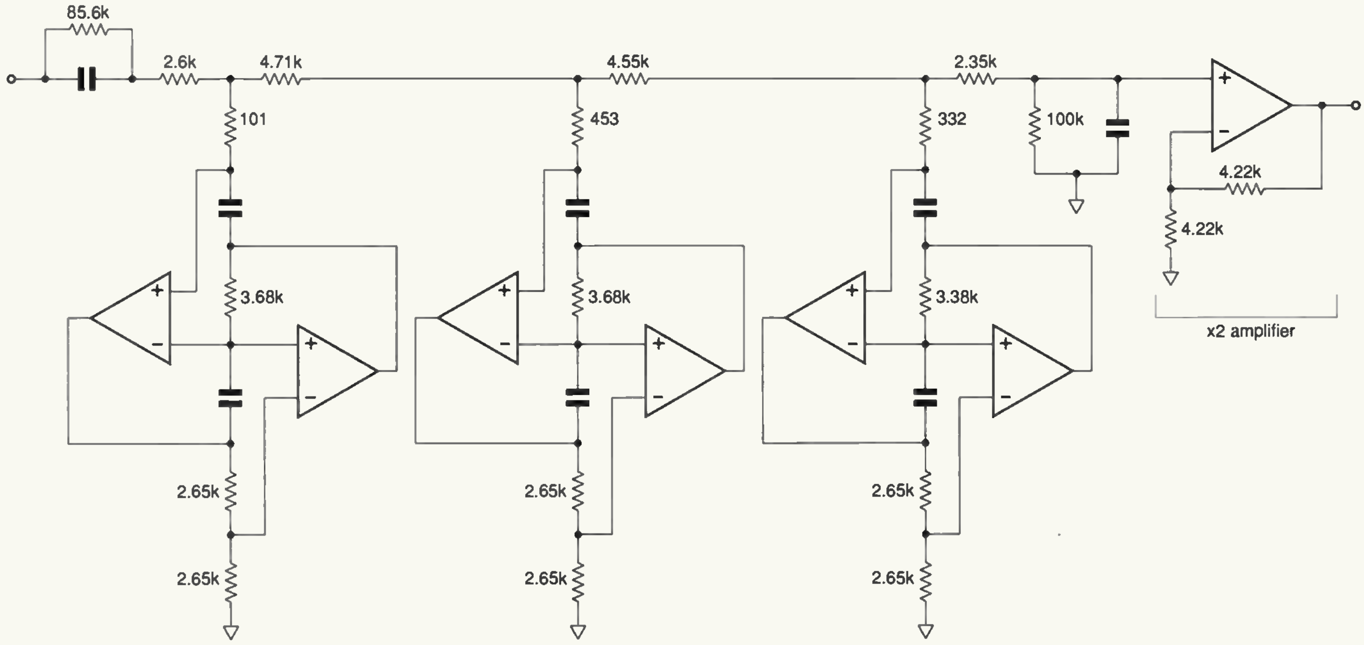
The response shown in **Fig. 8** was taken using the log frequency base mode of the *HP3580A* 0-50kHz spectrum analyser. In this mode, the spot writes the trace across the screen at a steady rate, taking about 6 seconds to sweep from 20Hz to 44.3kHz. Thus the sweep rate in hertz per second increases greatly as the spot progresses across the screen.

This means that if a resolution bandwidth narrow enough to resolve frequency components encountered near the start of the sweep (e.g. 1Hz or 3Hz bandwidth) is used, then near the end of the sweep the analyser will be passing through any signals far too fast to record their levels even approximately.

On the other hand, if a bandwidth such as 300Hz – wide enough to accurately record signal amplitudes in the 20kHz region – is used, the zero frequency carrier breakthrough, response will extend half way across the screen. So, when using log sweep mode to record the amplitudes of stationary signals, compromises must be made.

But this is not the case in Fig. 8, for here the only signal of interest is the output of the tracking generator, to which the analyser is, by definition, always tuned. So the analyser is at no time sweeping through a signal and in principle it might seem that the 1Hz bandwidth could be used. There is a restraint on the bandwidth, however, set by the rate at which the signal amplitude changes. This can get quite fast in the vicinity of a notch, and accordingly the trace in Fig. 8 was recorded with a 30Hz resolution bandwidth. At 10Hz bandwidth, the notch appeared shunted slightly to the right and its full depth was not recorded.

On the other hand, at a 100Hz bandwidth, the notch response was identical to that shown, but the left hand end of the trace, representing 20Hz, was elevated slightly, due to the zero frequency carrier breakthrough response. If, due to a fortuitous conjunction of component tolerances, the actual notch depth had been much deeper than it actually was, the 100Hz bandwidth would have been necessary to capture it. In that case, it would be better to switch back to linear frequency base mode, and make the notch measurement at a span of 100Hz, or even 10Hz, per horizontal division.



This is all fine in theory, but does it work in practice?

Proof of the pudding

Ever of a pragmatic – not to say sceptical – turn of mind, I determined to try it out for real. So I made the circuit of Fig. 7a) up almost exactly as shown, and tested it using an *HP3580A* audio frequency spectrum analyser.

The circuit was driven from the *3580*'s internal tracking generator. There were minor differences. Whereas the plot of Fig. 7b) was modelled with *LM318* op-amps, these were not to hand, so a *TLE2072CP* low-noise, high-speed j-fet input dual op-amp was used. This is a handy Texas Instruments device with a 35V/μsec slew rate and accepting supplies in the range ±2.25V to ±19V.

The required resistor values were made up using combinations of preferred values, e.g. 82kΩ+12kΩ for 93.6kΩ, 270Ω+47Ω for 318Ω, etc, all nominal values thus obtained being within better than 1% of the exact values. 100kΩ+12kΩ was used for the terminating resistor, to allow for the 1MΩ input resistance of the spectrum analyser in parallel with it. The resistors were a mixture of 1% and 2% metal film types, except the 47Ω, which was 5%. The four 10nF capacitors were all 2.5% tolerance polystyrene types.

Although the circuit worked, its response was not exactly as hoped, due to being driven from the *3580A*'s 600Ω source impedance. So a *TLE2027* single op-amp – not to be confused with the *TLE2072* dual device used for the main circuit – was used as a unity gain buffer to drive the filter from a near-zero source impedance. Its output level was set at the top of the screen, **Fig. 8**.

First, the filter action was disabled by removing the 318Ω resistor, leaving a straight-through signal path. The upper trace shows the 6dB loss due to the terminations mentioned earlier. It also shows a first-order roll-off due to the effect of the terminating capacitor at the load end, with the two 3180Ω resistors.

Response of the complete filter, with the 318Ω resistor replaced, is shown in the lower trace. The reference level has been moved down one graticule division for clarity. The –1dB point is at two divisions in from the left, which, given the horizontal scaling of three divisions per decade, corresponds to 9.3kHz – pretty close agreement with the predicted performance of Fig. 7b).

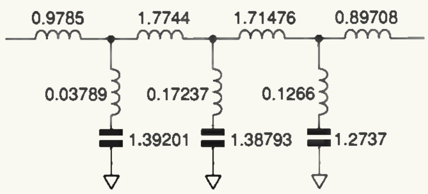
In logarithmic frequency mode, the analyser's bandwidth extends only up to 44.3kHz. But this is far enough to see that the notch frequency and the level of the return above it, 36dB below the l.f. response, also agree with the computed results.

Others like it, too

Various applications have been found for fdnr filters, especially in measuring instruments. The advantage here is that the response is predictable and close to the theoretical.

Some other active filter sections, when combined to synthesise filters of a higher order, show a higher sensitivity to component tolerances. This is a disadvantage where the filters are used in the two input channels of an instrument, which requires close matching of the channel phase and amplitude responses. For this reason, fdnr filters were used in the input sections of the *HP5420A*, **Fig. 9**.[3] ■

Fig. 9a). Normalised seven-pole elliptic* LC *prototype filter, and b), below, derived fdnr input antialiasing filters used in the HP5420A.



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2. Hickman, I., 'Newnes Practical RF Handbook', 1993, ISBN 0 7506 0871 4, p245.
3. 'Front-end Design for Digital Signal Analysis. Patkay', Chu and Wiggers, *Hewlett Packard Journal*, Vol. 29, No. 22, October 1977, p9.

Hands-on Internet

Focusing on inductor design this month, Cyril Bateman's has found that his searches are faster thanks to the new European version of *Yahoo*.

Since its inception, the Internet has seen a doubling in its usage each year, but in 1996 this growth tripled. This increased growth was in part due to the growth of access in Europe. Modem speeds also doubled during the year, but because of the growth in usage, real download speeds at the end of the year for the UK were little faster than those found a year ago when modems were generally slower.

Since the Internet is of US origin and US based, many accesses from Europe to European sites still require to be routed to North America and back across the Atlantic, which inevitably causes cable bottlenecks. Changes to routeings to permit direct European access to European sites are planned for this year. This should remove the bottleneck, resulting in improved download speeds for *all* users.

The extremely popular original *Yahoo* search engine can become overloaded, making it slow to respond. A new subsidiary, optimised for the needs of UK and Ireland, is www.yahoo.co.uk.[1] At the moment, it is much less used, and responds rapidly. Should your search request find no matches within *Yahoo*, then the search engine transparently defaults to search within the *AltaVista* site for you – which is very comprehensive – **Fig. 1**.

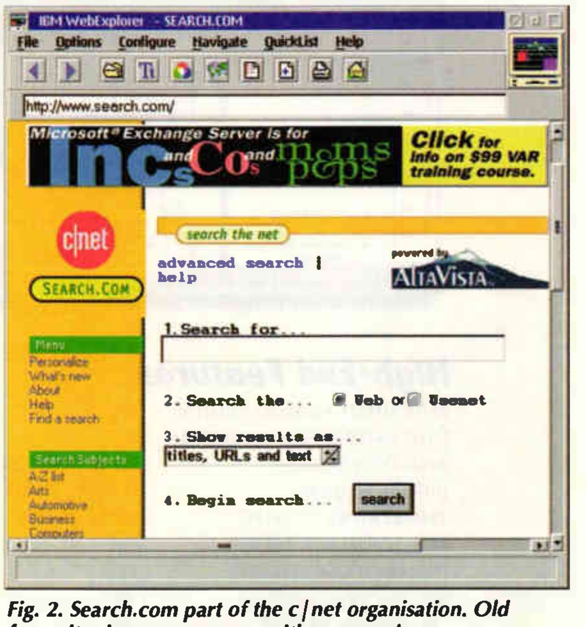

Fig. 2. Search.com part of the c|net organisation. Old favourite shareware source with a new role.

Fig. 1. This popular US search engine has been reworked for local European needs.

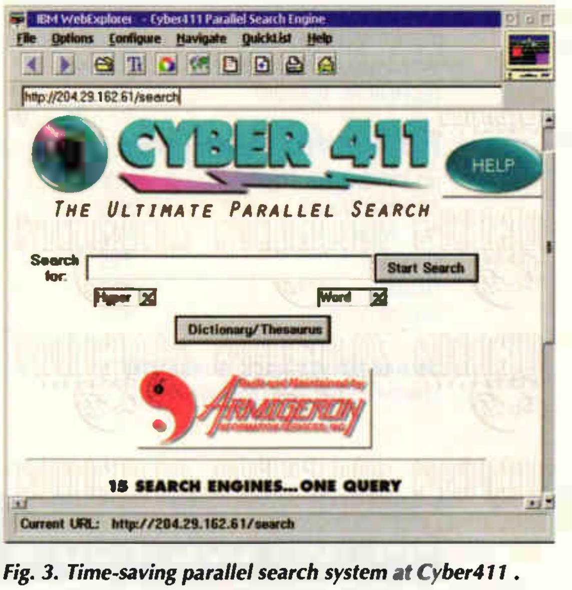

Fig. 3. Time-saving parallel search system at Cyber411.

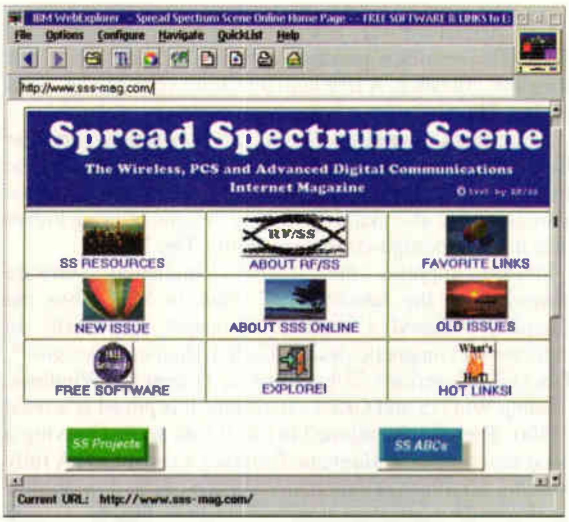

Fig. 4. On-line magazine and forum for designers. Good source of free rf and design software.

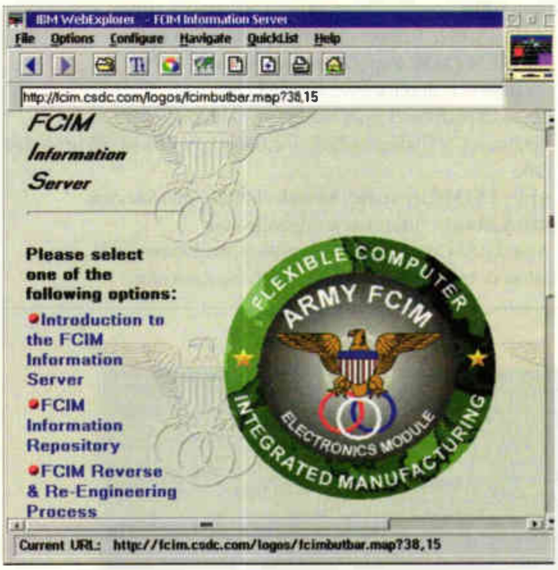

Fig. 5. The FCIM site data bank of MIL standards and part numbers – caters for all electronic parts – not just inductors.

A new search tool

Long established as a shareware source, Cnet.com, is a new dedicated Internet search engine. Its subsidiary, Search.com[2], is powered by the AltaVista search engine, but it has custom refinements. Two forms of search are possible. The top level search, **Fig. 2**, is based on the *AltaVista* system. Further down the page is a second search enquiry box which allows preselection of your choice from the eleven most popular search engines.

Typical of the latest style search engines is CYBER 411[3]. It performs your search request by searching fifteen different search engines simultaneously. The default search option collates all results and removes duplicated items, before displaying its findings to you. If you prefer to maximise search speed and accept duplicate finds, then select the 'hyper' search mode, **Fig. 3**.

Simulation software

Having followed the various articles for amplifier design and loudspeaker systems over the past year in *Electronics World*, you may be interested in further exploring measurement or design of a loudspeaker. A lengthy twenty-page FAQ, Software for Speaker Design, lists shareware, commercial

Fig. 6. The dTb home page is a useful source of slightly different design software.

software packages and hardware. It can be downloaded from http://harmony-central.com/Other/speakerfaq.

Spread Spectrum Scene[4] – also known as RF/Spread Spectrum – is an on-line Internet magazine providing a meeting place for designers to exchange ideas and problems. Its consultants' corner holds reviews of simulation software packages and industry news.

Featured this month is a freeware program from Hewlett Packard called *AppCAD*. In addition to rf design and analysis routines, this package also includes an excellent tutorial on thermal analysis of semiconductor packages. I regularly use and recommend this package[5]. It can be downloaded using FTPSearch Norway.

Spread Spectrum Scene has a carefully selected library of free software. Some, such as the rf and Smith chart packages, are usually very difficult to find, but all are easily downloaded from this page. Especially interesting is Mathsoft's *Mathbrowser* which allows use of *Mathcad*'s live worksheets, in a browser, over Internet, **Fig. 4**.

Models for passives

Simulation models for most semiconductors and integrated circuits are readily available, but simulation models for capacitors and inductors are much less common. In previous articles I have detailed sources of capacitor modelling data but not those covering inductors.

Both in practice and in modelling, inductors and magnetic devices are generally less well understood than capacitors. Realistic generic spice models for inductors seem not to exist anywhere on the Internet.

Alternative source for inductor information

While it is not usually practicable to make your own capacitors, inductor cores are readily available, as is wire. But people wanting to fabricate their own inductors or transformers can find data sheets on magnetic materials difficult to interpret.

If you have to design inductors regularly and have standardised on a small selection of core materials, it is relatively easy to write a dedicated calculation software routine. It can incorporate the magnetic design data parameters to automate the design process. It can even be made to calculate effective inductance with dc load current.

Much of the basic design information on inductors is old. Many of the books on the subject are no longer in print. Many years ago I found in a second-hand store a 1946 copy of Radio Designer's Handbook, once distributed by *Wireless World*, which I treasure because it includes many useful inductor design tables.

Fig. 7. Miniature inductors and transformers. Speciality surface mount devices for all needs.

Fig. 8. Design of magnetic circuits, inductors and transformers made easy. The demonstration package is free.

Recent searches on inductors and magnetics have identified a few sources of helpful information. One which is easy to use and understand is part of the US military stores catalogue. It is the Army FCIM Electronics Module[6]. This lists a large number of US military-specification approved inductors. More usefully, it also includes design tables, which can be down loaded for local use. You will need to register yourself before you can access this site though, **Fig. 5**.

Commercial software is also available. A company called dTb Software[7] offers packages covering inductor design and wire size guidance. A free demonstration suite can be down loaded, **Fig. 6**.

Associated Components Technology[8] – a large volume Far Eastern inductor maker – provides detailed descriptions of the construction of potted air-cored surface-mount chip inductors. This company also claims to supply designs offering higher than normal through-current capability, **Fig. 7**.

Intusoft supplies the *ICAP/4* simulation software mentioned in the January 1997 issue. In May 1996, the company released a package dedicated specifically to inductor and magnetic design. Called *Magnetic Designer*[9], this tool is a serious 32-bit software system for Windows, needing Win32S and OLE2 extensions. It is priced at around $1300. The page mentioned in ref. 9 lists an FAQ giving a good description of *Magnetic Designer*'s capability. A fully working 4.5 Mbyte demonstration version can be downloaded, or obtained by postal service, on four discs, **Fig. 8**. ■

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Scrambler for data

Based on a pseudo-noise sequence generator, this scrambler scheme for digital bit streams combines good data security with low error rates. Designed by Wasim Ahmad and Mohiuddin Bhat

Scramblers are a simple and effective means of making digital data communications secure. A simplified, generalised diagram of a scrambler and the corresponding descrambler are shown in **Fig. 1**.

For scrambling, an n-stage serial-input-parallel-output shift register, or sipo, is used. Its output taps a_1, a_2...a_n, are added to give m_k''. The resulting bit sequence is added to the message bit sequence m_k to give the bit sequence m_k', ready for for transmission.

Output taps from the shift register are gated by a 0 or 1. Any tap gated with a 1 results in a connection being made to the corresponding shift register stage. When the gating is zero, no connection is taken from that stage.

The receiver has to know the pattern required for unscrambling the transmitted bits. Unscrambling becomes difficult if the number of shift register stages, n, is large. One problem associated with this technique – especially, for large values of n – is that of multiple errors may result in the unscrambled output. A single erroneous bit in the received message m_k' can cause a sequence of n erroneous bits in the unscrambled message. This problem of error propagation becomes even more serious if the transmission channel is noisy.

Various techniques have been reported for reliable and secure data transmission[1-3]. Usually, the scrambled bit stream is transmitted using some form of channel coding, but this naturally complicates the system.

Our method for scrambling the data is novel. It need not use large numbers of shift-register stages, yet it makes unscrambling very difficult for hackers.

A new scrambling scheme

Figure 2 shows the proposed scheme. It is similar to that shown in Fig. 1, but the tap gating values are continuously changed with the help of a pseudo noise sequence. Inputs Q_A, Q_B, etc, to the and gates are the outputs of the pseudo noise sequence generator[3] shown in **Fig. 3**.

In order to unscramble the received message, the receiver has to know the number of shift register stages and the pseudo noise sequence. It also has to know the starting pattern of Q outputs from generator used in the scrambler. This makes unscrambling very difficult for unauthorised receivers while minimising the problem of error propagation by allowing a smaller value for n.

We have tested the scheme with n=4 and a pseudo noise sequence of 15 bits long. Performance was good. The system works equally well when the and gates of Fig. 2 are replaced by nand gates.

For a practical implementation, all the bistable devices of the pseudo noise sequence generator are initially cleared. This must be done at both the transmitting and receiving ends.

First, a synchronising word is passed. Since the Q outputs of Fig. 2 are zero, m_k'' will be zero and the synchronising word will pass directly. At the end of synchronising word, the bistable devices of each pseudo-noise generator are loaded with the starting values of the pseudo-noise sequence. This generates the data sequence, with clock pulses, and the process goes on.

At the transmitting end, if the whole pseudo-noise sequence is registered – say a 31-bit sequence handled by a 31-bit circulating shift register – many binary signals could be scrambled separately using different groups of Q outputs. These signals could then be transmitted using time division multiplexing. Each receiver will obtain its corresponding Q codes from pseudo-noise sequence generator. In this way, only the proper receiver will receive its corresponding message.

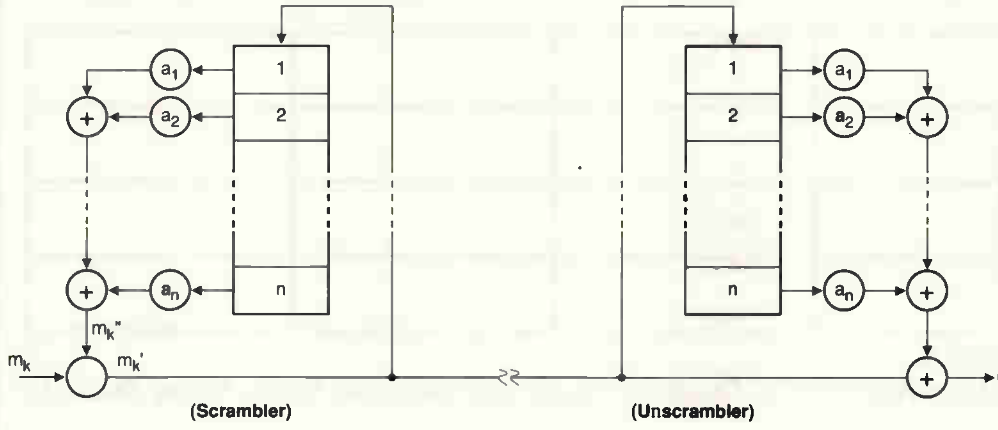

Fig. 1. In the conventional scrambler and descrambler scheme, the data stream to be transmitted is added to a coding string produced by a shift register. The same register arrangement is used to remove the coding at the descrambler end.

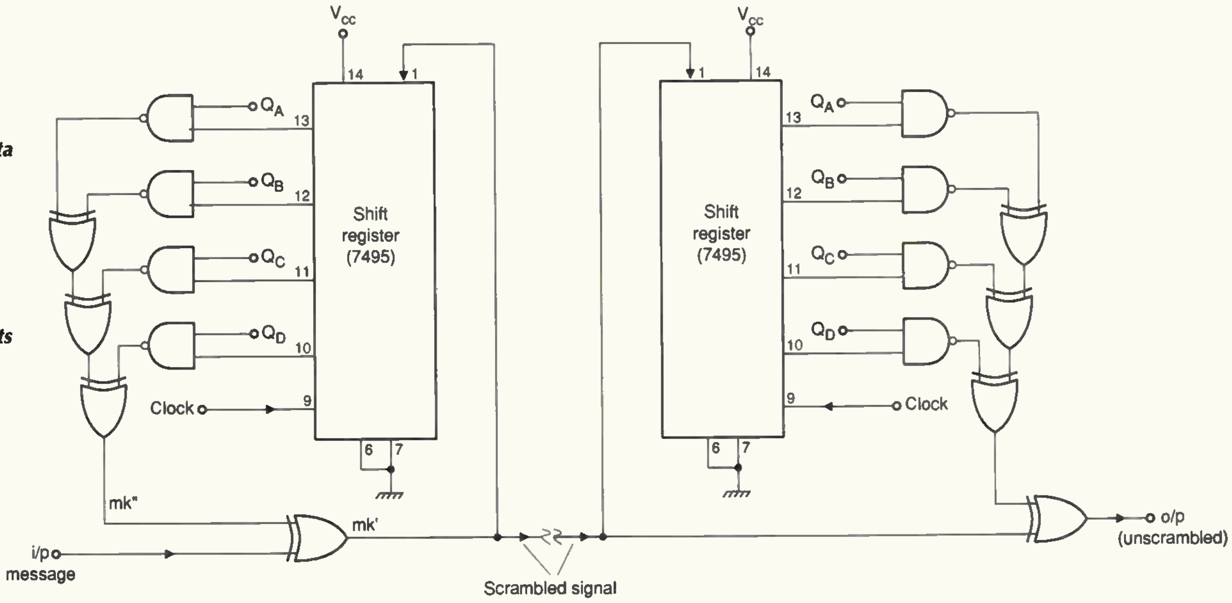

Fig. 2. Enhanced scrambler and descrambler. Data is made more secure without increasing the shift register length since the code fed to the And gate Q inputs is derived from a pseudo noise generator.

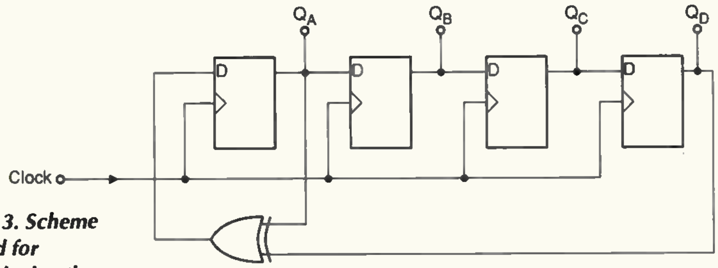

Fig. 3. Scheme used for producing the pseudo random sequence feeding the and gates of Fig. 2.

In summary

We have found this to be a reliable and secure method of data transmission. Scrambling is efficient and the scheme reduces the problem of bit-error propagation relative to the traditional method described earlier.

The concept has been tested with one 15-bit pseudo-noise sequence used for the message and another one for scrambling/unscrambling at the transmitting and receiving ends. The scrambled message was passed through a cable producing negligible inter-symbol interference and noise. Unscrambling was found reliable.

In a proper implementation of the scheme, a synchronising word should precede the message. It is important to d

Detecting the ending point of the synchronising word is important. Therefore, a clock recovery circuit is needed at the receiving end if the channel can suffer from inter-signal interference or zero-crossing jitter. ■

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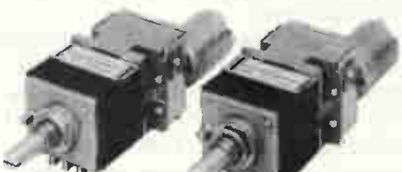

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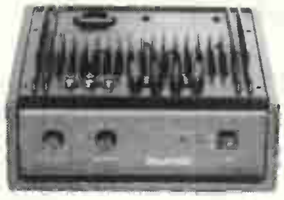

This attractive module consists of a low pass filter and power amplifier ready for you to mount in a suitable sub-woofer cabinet. The combined unit can then be combined with any new or existing hi fi or home cinema speaker system to add in the real bass punch missing from most setups.
The ASM 100 module comes as a ready-to-mount unit on a solid diecast aluminium frame/heatsink. Input signal can be at line or speaker level for easy system integration. There are three separate stereo inputs at line level and the unit will use any signal presented or mix all inputs to add bass to any signal. The speaker level inputs are used by simply wiring the unit in parallel with the existing speakers to provide them with strong bass support. Crossover frequency can be selected to 50, 100 or 200Hz and the bass level can be adjusted by a front panel control. The 'Green' power supply switches the unit to standby if no signal is present. Drawings are included free for the compact 418 x 380 x 303mm cabinet.
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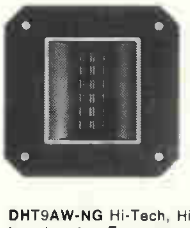

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APPLICATIONS SUMMARY

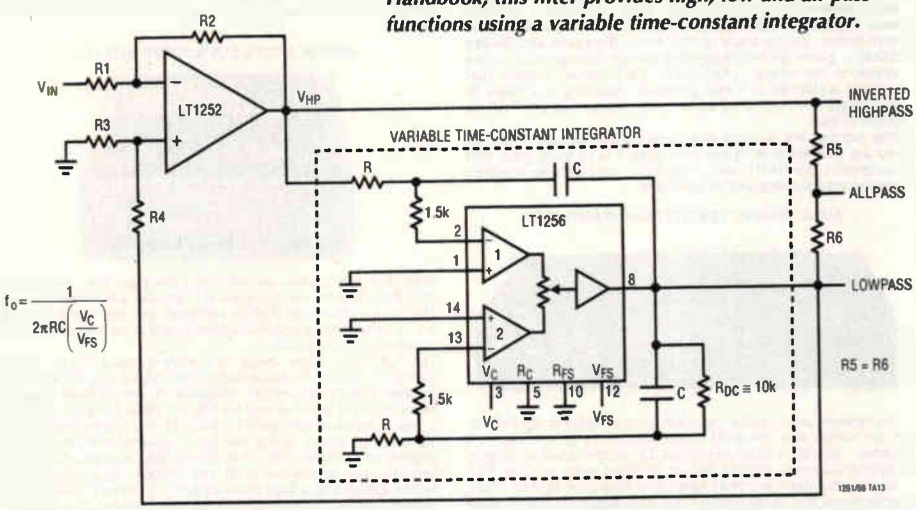
Voltage-controlled low-pass high-pass and all-pass filter

Taken from Linear's High-Speed Amplifier Solutions Handbook, this filter provides high, low and all-pass functions using a variable time-constant integrator.

An integrator within the feedback loop of an *LT1252* current feedback amplifier creates a filter with three output functions – low-pass, high-pass and all-pass.

By using a variable time constant integrator as the feedback element, the cut-off frequencies of the low-pass and high-pass outputs are adjustable, set by control voltage (V_c) on **pin 3** of the *LT1256*. On the all-pass filter output, adjusting the control voltage alters the phase of the output. Resistors $R_{5,6}$ set output impedance of the all-pass filter. For the *LT1252* R_2 should satisfy the cf amplifier's minimum resistance (750Ω) and all four resistors must satisfy $R_2/R_1=R_3/R_4$. Pass-band gain for all three outputs is $-R_2/R_1$.

***Linear Technology**, The Coliseum, Riverside Way, Camberley, Surrey GU15 3YL, tel. 01276 677676, fax 01276 64851.*



Send colour 1000 feet over low cost twisted pair

Composite video signals can be sent appreciable distances on low-cost twisted pair – in two directions. The cost advantage of this technique is significant. Standard 75Ω RG-9/U coaxial cable costs between 25¢ and 50¢ per foot, but PVC twisted pair is only pennies per foot. This means hundreds of dollars are saved in installations as short as 1000 feet, easily paying for additional electronics. The system also provides for 'drops' or receiver taps along the pair.

This bidirectional 'video bus' consists of the *LT1190* op-amp and the *LT1193* video difference amplifier shown in **Fig. 1**. The two top-left *LT1190*s generate differential signals to drive the line, which is back-terminated in its characteristic impedance. The twisted pair receiver is an *LT1193* video difference amplifier, bottom right, and it converts signals from differential to single-ended.

Because of the *LT1193*'s unique topology, it is possible to provide cable compensation at the amplifier's feedback node as shown. In this case, 1000 feet of twisted pair is compensated with 1000pF and 50Ω to boost the –3dB bandwidth of the system from 750kHz to 4MHz. Attenuation in the cable can be compensated by lowering the gain-setting resistor R_G. At top right, another pair of *LT1190*s provides cable termination via low output impedance and generates differential signals.

A good indication of the system's ability to pass colour video is **Fig. 2**. This multiburst pattern was passed through 1000 feet of low-cost PVC twisted pair; it contains a 3.58MHz chroma subcarrier and a 4.5MHz sound subcarrier. Although the photo shows these frequencies attenuated about 3dB, a clean picture is present at the end of the twisted pair.

This and the above circuit are taken from Linear Technology's High-Speed Amplifier Solutions Handbook.

***Linear Technology**, The Coliseum, Riverside Way, Camberley, Surrey GU15 3YL, tel. 01276 677676, fax 01276 64851.*

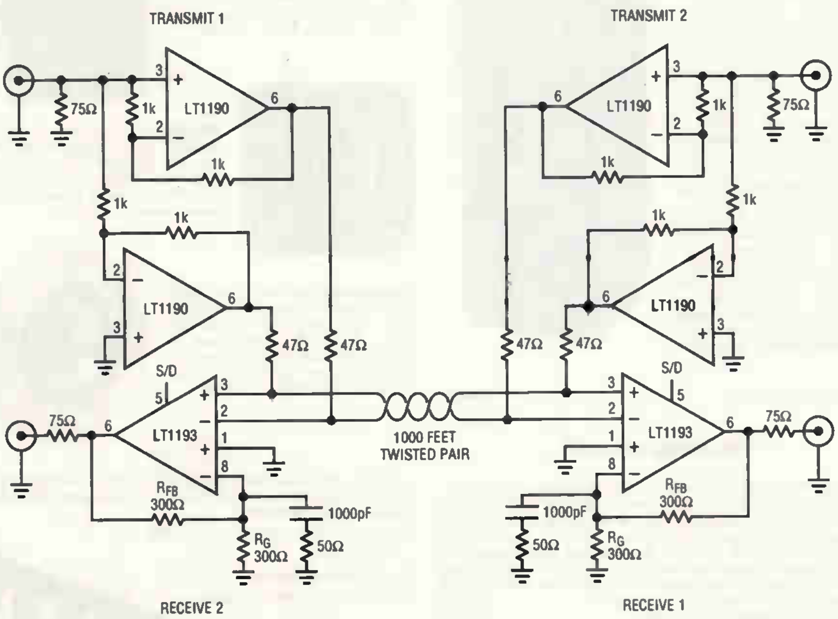


Fig. 1. Video transceiver uses low-cost PVC twisted pair.

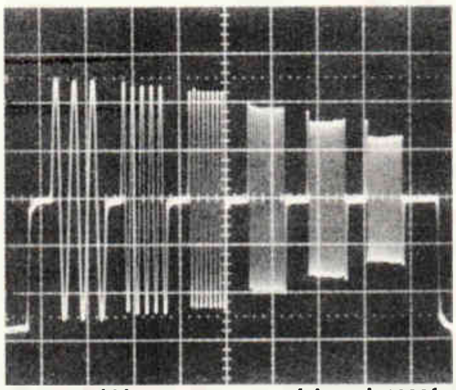


Fig. 2. Multi-burst pattern passed through 1000ft of twisted pair. Although –3dB at around 4MHz is indicated, colour video pictures are clear.

Sensitive carbon-monoxide sensor

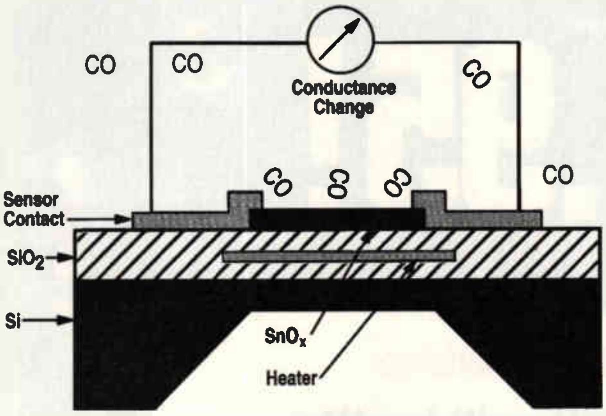
A new CO gas detector, namely the Motorola *MGS 1100*, is soon to become available. Tentatively priced at $10 in small quantities, the detector will eventually be supported by application-specific hardware. But the published preliminary data, summarised below, should be enough to construct a complete CO detector.

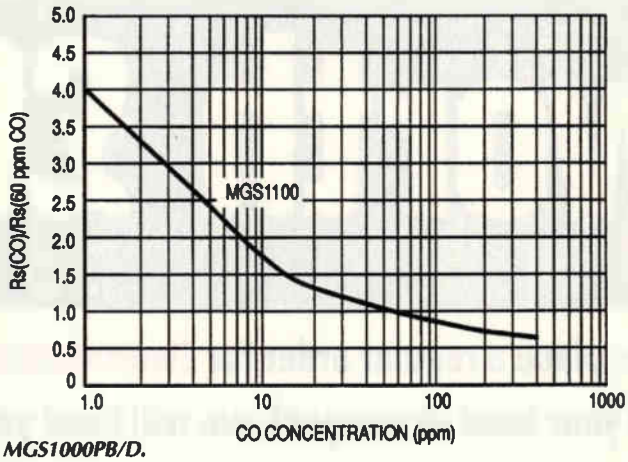
Incorporated into the device is a 5V heater taking a maximum of 50mA. It requires a 48 hours pre-conditioning period. This is intended to be a mains operated device. The associated circuitry needs to supply 5V at 50 mA, which is easily done. The recommended mode of operation is to switch the heater from 5V for 5 seconds to 1V for 10 seconds.

In still air the device has a nominal resistance of 10kΩ, which reduces by a factor of 1.8 at a CO concentration of 100ppm. The measurements should be taken at consistent points in the on/off cycle, e.g after 4.5 seconds high and 9.5 seconds low. Two sample-and-hold circuits need to be gated at the appropriate times and compared.

Although unlikely to prove fatal unless exposure is prolonged, 100ppm is a toxic level.

***Motorola**, tel 01354 688040, fax 01354 688248 (please note that this number is not for general enquiries).*





Sensitive at low gas concentrations, the MGS1100 is described in document MGS1000PB/D.

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049

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PM-128 digital panel meter

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100MΩ input impedance
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Maximum display 1999 counts, auto polarity
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£100 WINNER

Frequency-programmable analogue band-pass filter

Centre frequency of this filter is linearly proportional to the digital inputs of an a-to-d R-$2R$ ladder converter.

Two main elements comprise the circuit: a gyrator and a band-pass filter. **Figure 1** shows the gyrator, the input impedance of which is given by,

$$z_{in}(s) = sR_1R_2$$

which appears as an equivalent inductor with a value of $L_{eq}=LR_1R_2$.

Figure 2 is a basic band-pass filter, whose transfer function is,

$$H(s) = s/CR'/(s^2 + s/CR' + 1/LC)$$

compared with that of a second-order band-pass, which is,

$$H(s) = \omega_0 A_0 s/Q(s^2 + s\omega_0/Q + \omega_0^2)$$

Pass-band gain is 1 and centre frequency

$$f_0 = 1/(2\pi LC) = \omega_0/2\pi.$$

Bandwidth is $f_0/Q = 1/(2\omega CR)$.

An analogue-to-digital converter is represented in **Fig. 3**, in which b_{n-1}, b_{n-2}... b_0 are the digital inputs to an n-bit R-$2R$

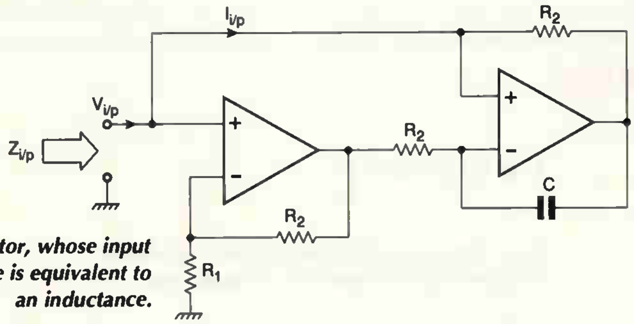


Fig.1. A gyrator, whose input impedance is equivalent to an inductance.

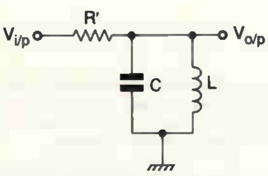


Fig. 2. Band-pass filter.

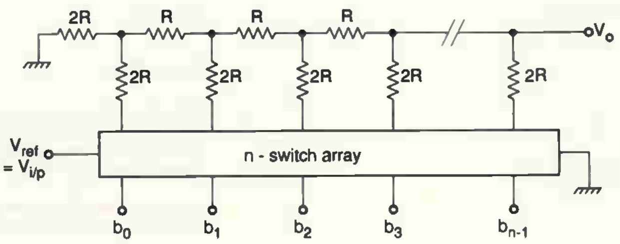


Fig. 3. N-bit ladder network used in an analogue-to-digital converter, the input voltage being represented by the resistor switches.

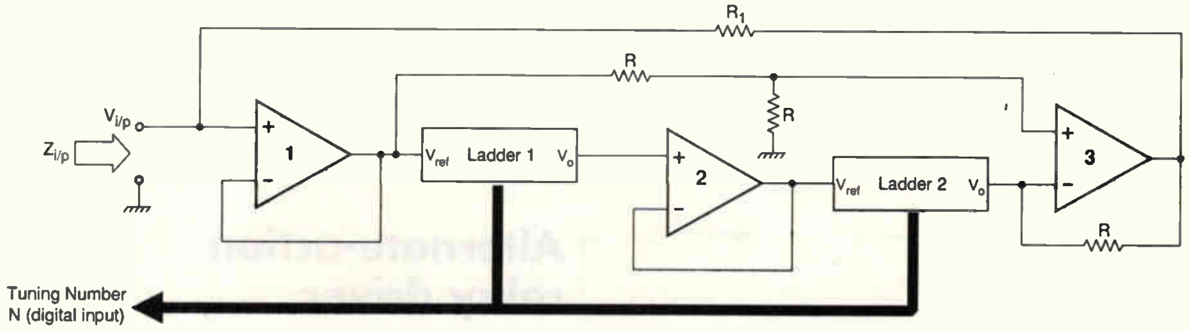


Fig.4. Input impedance of this arrangement is resistive and varying inversely with N².

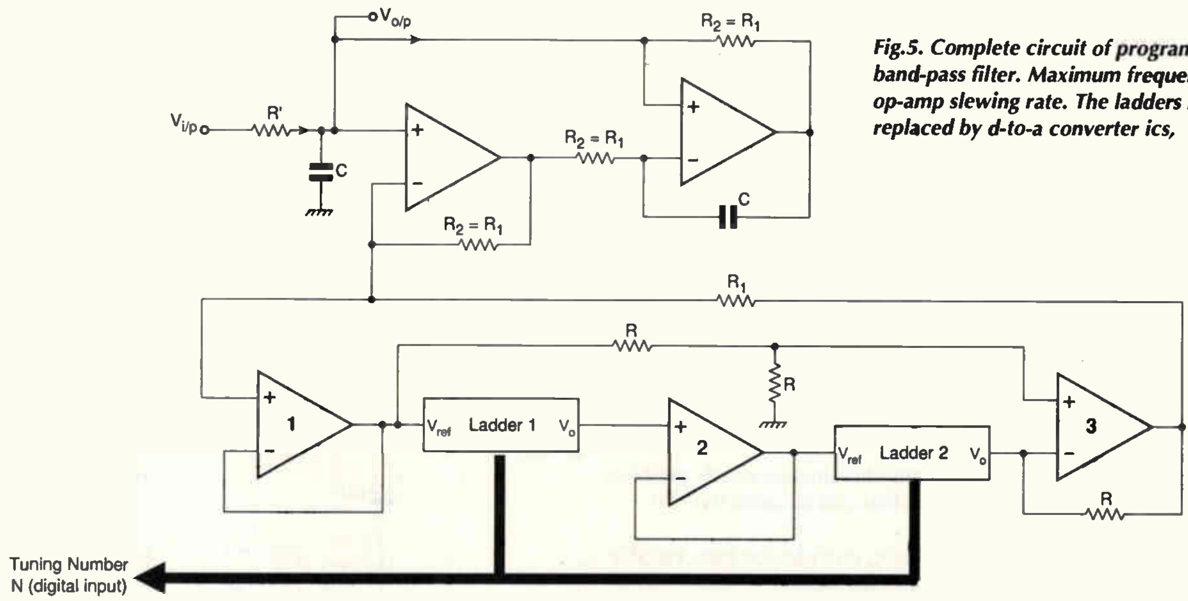


Fig.5. Complete circuit of programmable band-pass filter. Maximum frequency is set by op-amp slewing rate. The ladders may be replaced by d-to-a converter ics,

ladder network, whose output is $v_o = N/2^n v_{in}$. Output impedance is R.

Figure 4 is a development of this, in which the op-amps are simply buffers and both ladders have the same digital input. Output voltage of ladder 1 is $(N/2^n)v_{in}$, which is also the input to ladder 2, whose output is therefore $(N/2^n)2v_{in}$. The output impedance of the ladders being R, op-amp 3 output voltage is Av_{in}.

Now connecting R_I from input to output produces an input impedance,

$$z_{in} = R_1/(1-A) = R_1/(1-(1-N/2^n)^2)$$
$$= R_1(2^n/N)^2$$

which is a resistance varying inversely as the square of N.

If now the resistor R_1 in the gyrator circuit of Fig.1 is replaced with the circuit of Fig.3 and this new gyrator is used as the inductor in the circuit of Fig.2, the whole being shown in **Fig.4**, the centre frequency becomes,

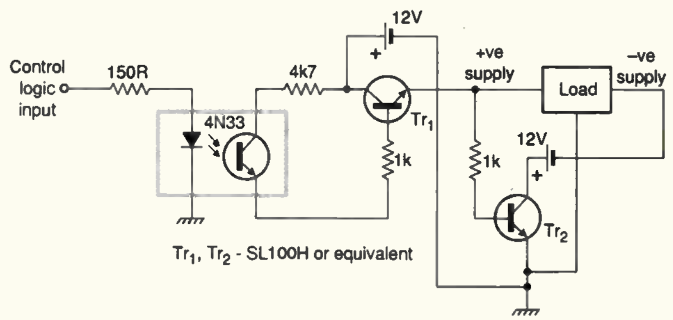
$$f_0 = 1/(2\pi(R_2R_1(2^n/N)^2 - C)^{1/2}).$$

If $R_2 = R_1$,

$$f_0 = N/(2\pi R_1 C * 2^n),$$

the centre frequency being proportional to the tuning number N, independently of bandwidth.

S. Santhosh Kumar
Kerala
India



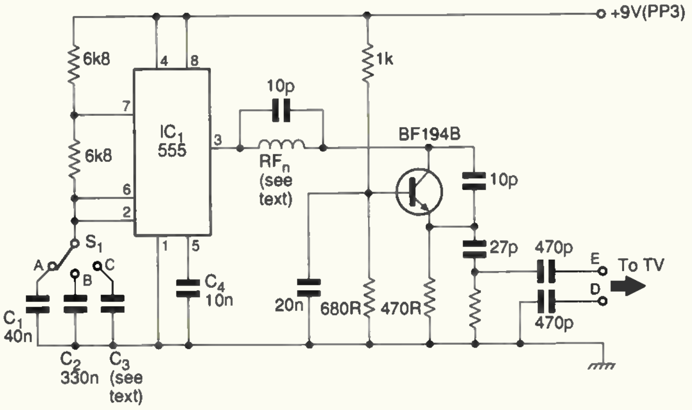
Both rails of a dual power supply are switched by logic input.

Switch two supply rails at once, digitally

Logic-level signals control the on/off switching of dual power supplies.

Input switching logic signals go to an opto-isolator which, when conducting, supplies base current to Tr_1 and completes the path from the +12V supply through the load to ground. At the same time, Tr_2 receives base current via Tr_1 and provides the path for the negative 12V supply. In the absence of drive to the opto-isolator, both transistors are off and neither power supply is connected.

V Lakshminarayanan
Centre for Development of Telematics
Bangalore, India



TV bar generator. Assemble it on strip board with the coil about 10mm above the board. Use a PP3 battery rather than a battery eliminator to avoid ripple.

One-chip television bar generator

An uncomplicated circuit, this simultaneously produces a pattern of horizontal bars and an audio tone for servicing purposes.

A *555*, used as an oscillator, provides the bars, variable in number by the switched capacitors C_{1-2} from 15 to 4; C_3 is about 10pF, but will need to be selected to produce vertical bars.

Output from the *BF194B* rf oscillator, whose coil is not cored, falls in the vhf range. Its coil is eight turns of 24g enamelled copper wound on a small former, which is removed later.

Connect points E and D to the television aerial socket. Switch C_1 or C_2 into circuit and set the tv to channel 4, adjusting the coil spacing to obtain a clear pattern and audio. You might need to tune the tv to get a good pattern.

Raj K Gorkhali
Kathmandu
Nepal

Obviously not for UK use, Ed.

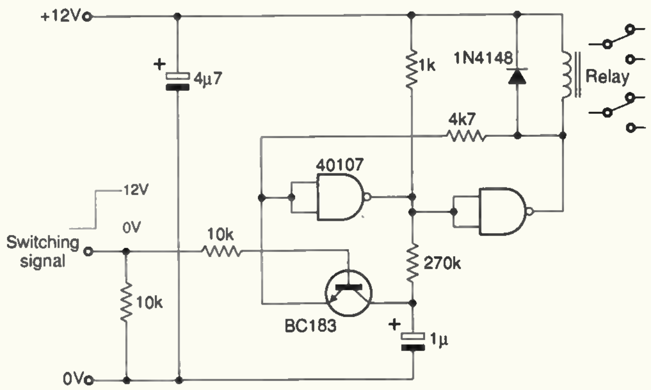
Alternate-action relay driver

A relay was required to change state when a positive voltage of variable duration, and possibly noisy, was received.

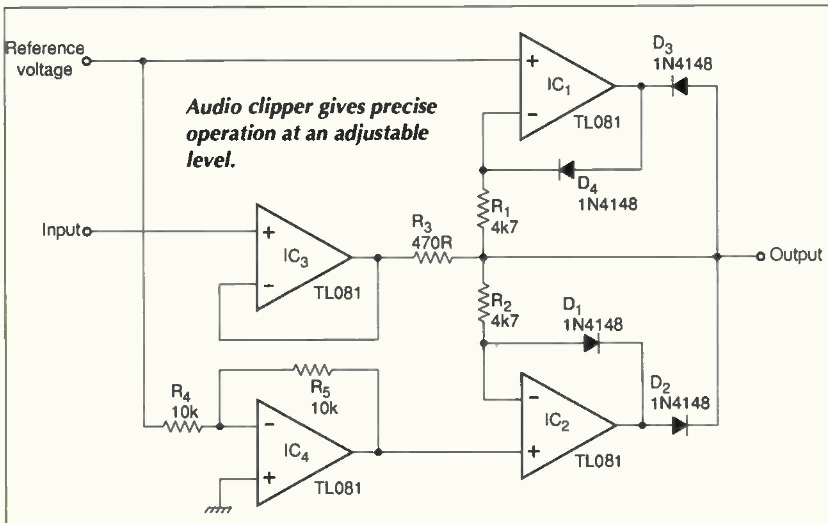
Connecting a push-button switch in the position of the transistor in the circuit shown achieves the effect needed, since it is able to conduct in both directions, and the *40170* will drive the relay directly.

If a transistor is only required to switch, it can often be used with emitter and collector reversed. Here, an n-p-n type, a *BC183*, acts alternately in the conventional direction and in reverse, depending on the state of the circuit, and responds to a positive input to the base in both conditions.

L S Whitlock
Taunton
Somerset



Transistor is used to replace a push-button switch to produce relay action on receipt of a positive voltage input to the circuit.



Audio clipper gives precise operation at an adjustable level.

Symmetrical audio clipper

Virtually self-explanatory, the diagram shows an audio clipper which operates symmetrically at a level set by the reference input. Since the op-amps are never in saturation, the circuit operation is fast.

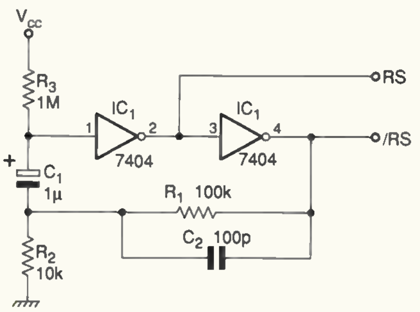
The only other point to make is that, when not clipping, the circuit attenuates by a factor $R_3/R_1\,R_2$.

J A Burnill
Camberley
Surrey

Clean power-on reset

These two inverters give a clean power-on reset, the reset time being determined by R_3C_1 and the amount of hysteresis by $R_{1,2}$. Capacitor C_2 is simply a speed-up component.

J A Burnill
Camberley, Surrey



Simple circuit to provide power-on reset with hysteresis.

RS 232-to-parallel conversion

Under control of a pc by way of Com1 or Com2 ports, the circuit takes in RS 232 serial data and converts it to parallel form.

Data comes in to a *74LS164* serial-to-parallel shift register, together with a pulse to set the latch. Data rate B in baud is around $11.5/C_1$kHz, where C_1 is in nanofarads. Divisor value for the pc is $(1.8432\times10^{6})/16B$, the pc being programmed for that rate by using this value in 16-bit form by means of Turbo C statements for Com 1:

```
outport(0x2FB,131);      /* 2FB for baud setting */
outport(0x2F8,0xX);      /* lsb byte of divisor */
outport(0x2F9,0xXX);     /* msb byte of divisor */
outport(0x2FB,3);        /* 2FB for transmission */
```

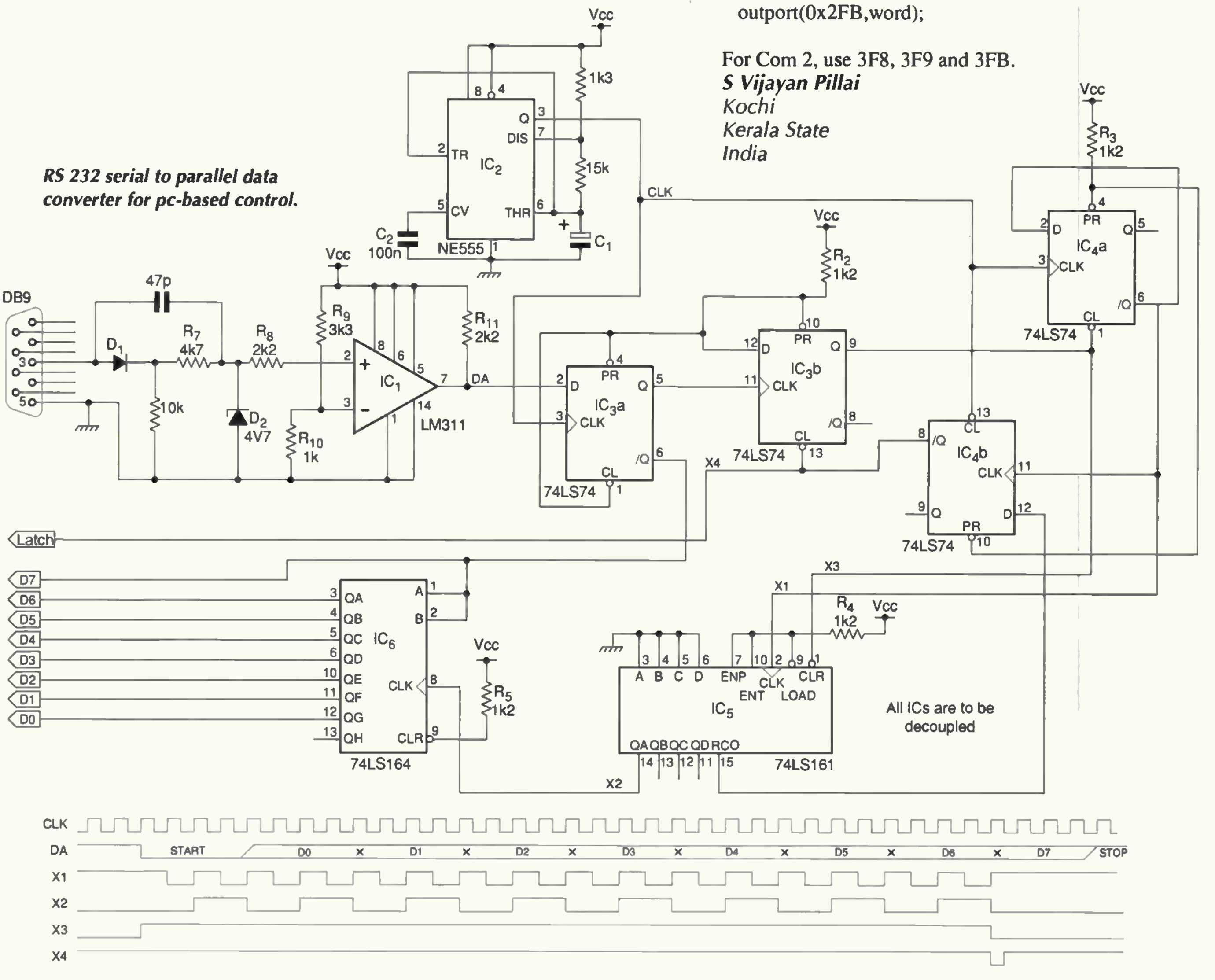
Data in eight-bit form may be sent to pin 3 of the 9-pin D connector by the statement:

```
outport(0x2FB,word);
```

For Com 2, use 3F8, 3F9 and 3FB.

S Vijayan Pillai
Kochi
Kerala State
India

RS 232 serial to parallel data converter for pc-based control.



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Philips PM5390 RF syn – 0.1 – 1GHz – AM + FM – £1000.
S.A. Spectral Dynamics SD345 spectrascope 111 – LF ANZ – £1500.
Tektronix R7912 Transient waveform digitizer – programmable – £400.
Tektronix TR503 + TM503 tracking generator 0.1 – 1.8GHz – £1k – or TR502.
Tektronix 576 Curve tracer + adaptors – £900.
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Tektronix 1502/1503 TDR cable test set – £1000.
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Nicolet 3091 LF oscilloscope – £400.
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Time Electronics 2004 D.C. voltage standard – £1000.
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CIRCLE NO. 133 ON REPLY CARD

Programmable logic primer

Geoff Bostock explains how logic elements are formed in various types of programmable logic chips, in this second extract from his book 'FPGAs and Programmable Logic'.

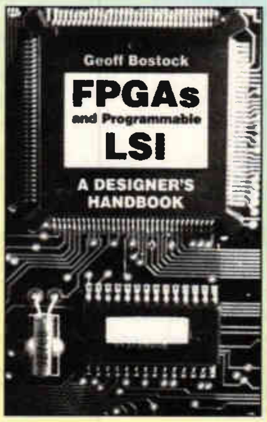
The two main technologies used for standard logic are bipolar, for the transistor-transistor logic, or ttl, families, and complementary metal-oxide semiconductor, or c-mos, for the 4000 and HC families. Examination of the basic circuit diagrams of a standard gate built in the two technologies, **Fig. 1** for ttl and **Fig. 2** for c-mos, illustrates the main differences in performance and application.

Invariably, ttl has a direct current path from V_{cc} to 0V; when any input is low, current will flow through R_1 and via the output transistors of the driving stages. If all the inputs are high, T_2 is switched on with a standing current determined by the values of R_2 and R_3. Also, during switching, both output transistors conduct momentarily, causing a substantial current spike to be drawn from the power supply.

On the other hand, c-mos always has either the lower n-channel ladder turned off when any input is low, or all the upper p-channel transistors off when all the inputs are high. The only current which flows is a charging/discharging current when any of the nodes changes level; this is due to the capacitance associated with any node.

In terms of power consumption, then, there is a substantial difference between the two approaches to standard logic. Consuming anything from less than 1mA to more than 6mA, ttl is a relatively power-hungry technology. Consumption depends on the resistor values, which are chosen according to

This article is derived from Geoff Bostock's new book 'FPGAs and programmable LSI – a designer's handbook'. The work covers designing FPGAs, large PAL structures, RAM and antifuse-based FPGAs and FPGA selection. Comprising 215 pages, this book is available by sending a postal order or cheque with a request for the book to Electronics World, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. The fully-inclusive price is £27.50 UK, £30 Europe or £33 rest of world. Alternatively, fax your full credit card details and address on 0181 652 8956 or e-mail jackie.lowe@rbi.co.uk.



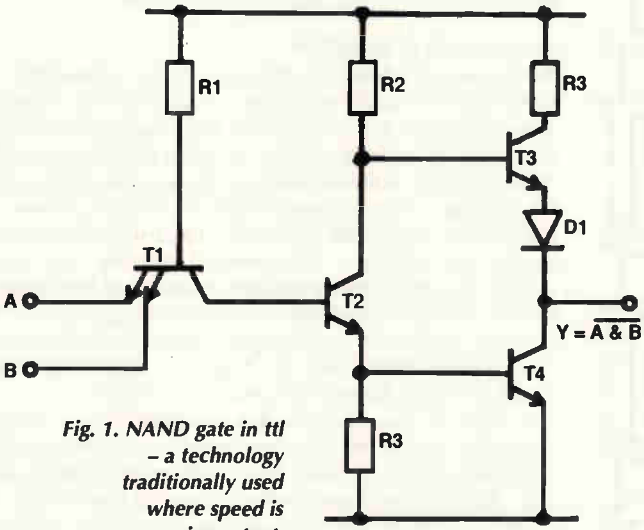


Fig. 1. NAND gate in ttl – a technology traditionally used where speed is important.

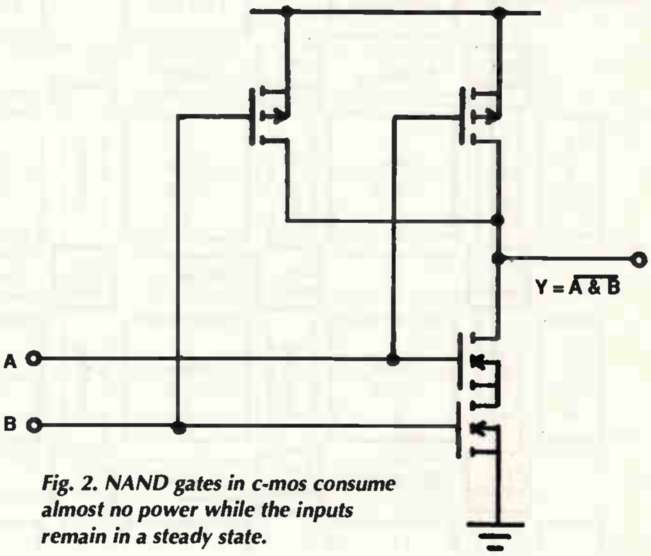


Fig. 2. NAND gates in c-mos consume almost no power while the inputs remain in a steady state.

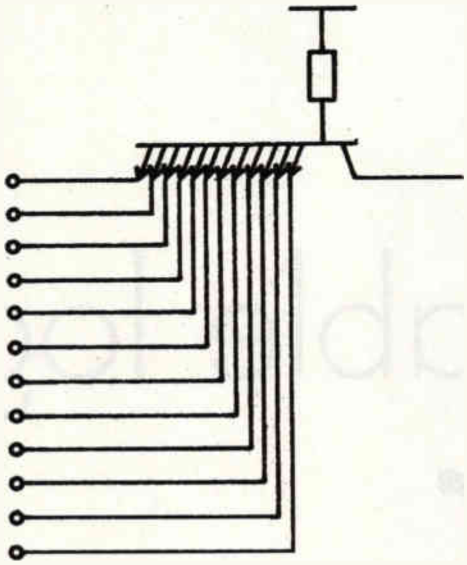


Fig. 3. With ttl, there is almost no limit to the number of inputs that can be attached to a single gate.

the speed of operation required for the particular ttl family. Current consumption also increases with operating frequency because of the current spike at each switching of the outputs.

In a cmos gate, when no inputs are being switched, the only current flowing will be leakage current through the mos transistors, which are in their off state. As a result, c-mos, consumes virtually zero power at zero frequency. As with ttl, the supply current increases with operating frequency as both internal and external capacitances are charged and discharged more often.

Importance of operating speed

The other major parameter which is usually important to circuit designers is operating speed. A crude measure of this is the propagation delay between the inputs and outputs of a logic gate. The most important part of this delay is the time taken to charge and discharge the node capacitances inside and outside the circuit. Another component is the time needed to remove stored charge from transistors in the circuit.

The time for a change in voltage is $(C \times V)/I$ where C is the node capacitance, V is the voltage change and I is the charge/discharge current. A low delay time is thus achieved by having a low capacitance and voltage change, but a high current.

Internal capacitance depends largely on feature size, which is likely to be a common factor between ttl and cmos. External capacitance depends only on the packaging of the integrated circuit and the pcb layout into which it is inserted. Both of these should be independent of the technology. The voltage change for ttl is 2.0V to 0.8V, allowing for noise margins for most families. At 5V, cmos noise margins dictate a voltage change of 3.5V to 1.5V. This is slightly more than for ttl, but the actual voltages at which switching occurs are not defined exactly. As a result, there will probably be little significant difference between the two technologies.

Current proves to be the real point of divergence. The ttl families are driven by transistors which are saturated, or held on the verge of saturation by Schottky diodes, while cmos current sources are more nearly resistive. For a given scale of technology, then, ttl families are traditionally the faster. Differences in the geometry of bipolar and cmos transistors have made c-mos transistors easier to scale down in size, so the current situation is that cmos devices can be as fast as ttl.

Another aspect of device speed is the output slew rate. If the rise or fall time of an output is comparable with the physical delay along the pcb tracks, reflections can become a problem and outputs may need to be terminated. The faster ttl families could fall into this category in some circumstances, while this was not usually a problem with the older c-mos devices. Now, however, with smaller geometries and faster edges, c-mos can need the same remedial treatment in some circumstances.

One other important difference between the two technologies lies in the input structures. A ttl input is a compact diode structure and there is virtually no limit to the number of inputs which can be attached to a single gate, as can be seen in **Fig. 3**.

In a c-mos gate, transistors must be stacked up, one transistor for each input. If there are too many transistors in the stack, the voltage drop across individual transistors becomes too small for correct operation and the gate fails to work. The consequence of this is that, for example, the 12-input gate in **Fig. 4** must be made by using two stages of gating, where one stage is possible in ttl form.

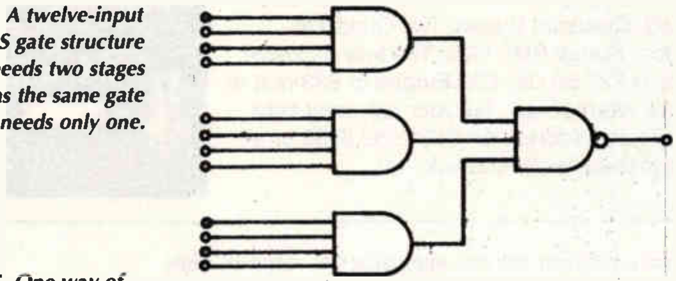


Fig. 4. A twelve-input CMOS gate structure needs two stages whereas the same gate in ttl needs only one.

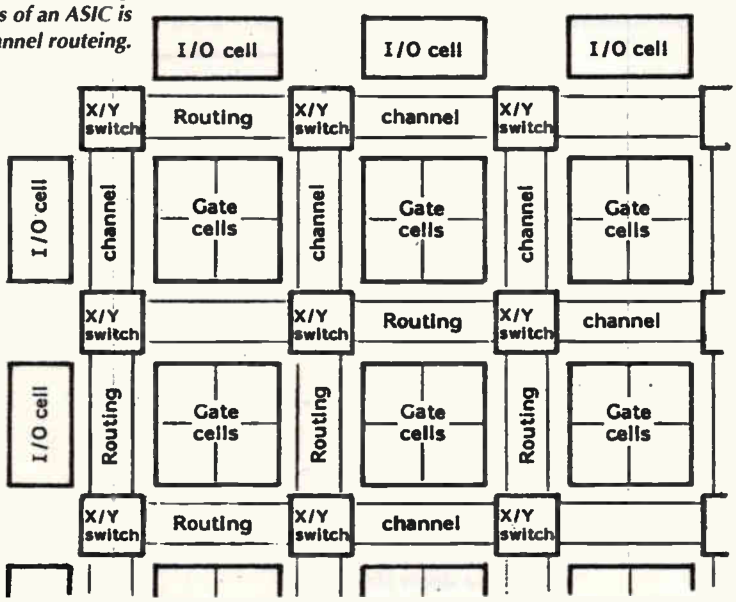


Fig. 5. One way of interconnecting components of an ASIC is channel routeing.

Large-scale integration

As processes improved to the point where a thousand or more transistors could be laid on a single chip, lsi, or large-scale integration, became feasible. The situation is different from medium-scale integration, in which functions can still be looked on as building blocks with universal application. For example, a four-bit counter might be used in a computer, a cd player or a digital multimeter. large-scale integration circuits are usually a self-contained function, the most prolific example being the microprocessor, apart from which most lsi functions are specific to a particular application. For example, a universal asynchronous receiver/transmitter, or uart, will normally only be found in communications equipment and a frequency synthesiser in tuners.

At first it was thought that microprocessors would drastically reduce the volume of small and medium-scale integration chips being used, but two effects made the reverse true. Although microprocessors operate at frequencies in excess of 100MHz, a simple logic function, such as ANDing two bytes, may need several operations to acquire the data, perform the function and then provide an output. The total cycle may occupy more than 100ns, compared with less than 10ns in readily available logic chips. The processor is also prevented from performing other tasks during this period, so it makes sense to continue with hard-wired logic or small-scale integration for simple logic functions.

The other effect is the need to interface the microprocessor to the outside world. Nearly every application, for example, needs an address decoder so that data and process instructions can be routed to and from the processor. This, and the need to customise many of the general-purpose peripheral circuits, adds to the number of discrete logic circuits surrounding the microprocessor.

In principle, these added chips can be combined into a single chip. Economies of scale dictate that this is not a practical approach unless the circuit is going to be used in upwards of a hundred thousand. These custom circuits are used in some applications where the quantities allow the cost of designing the chips, making masks, designing test sequences, and so on, to be amortised over a sufficiently large number.

Most applications are not large enough to benefit from this approach, but we can examine other ways in which the circuit designer can condense his hard logic into a small number of LSI chips.

Application-specific integrated circuits

The standard process for manufacturing integrated circuits involves growing a layer of silicon dioxide on a silicon wafer, etching windows in the surface layer and then introducing a controlled amount of impurity into the windows by vapour or electron-beam deposition. Successive layers of different impurities laid down through differently sized and shaped windows build up the active components in the surface layer of silicon.

Connections between the components are made by evaporating a conductor, usually aluminium or polycrystalline silicon, over the silicon dioxide. Previously etched windows allow contact to be made to the desired components. The conductor is then etched into tracks to define the circuit connections. Two or more conductor layers can be used by sputtering layers of silicon dioxide between the conductor layers.

In a standard or custom lsi circuit the layout is made by placing components on a 'floor plan' according to the schematic circuit diagram which defines the function of the chip. It will be laid out to minimise the area of the finished circuit, bearing in mind the design rules for the process. Usually, it is desirable to make the chip with a particular aspect ratio, often square, in order to aid assembly. Also, components which are close on the schematic will need to be close on the chip to make the conductor tracks as short as possible.

Clearly the arrangement of components is suitable for only the particular circuit under consideration; if any changes need to be made, or a new circuit laid out, the component positions will need to be changed as well. However, in a gate array – the simplest type of masked application-specific integrated circuit – the components are laid out in a predetermined pattern and the conducting layers tailored to the circuit schematic diagram.

Usually, the circuit components are set on a rectangular grid. Connections between components may be along muting channels in the gaps between components or, if no gaps are left, by routing across unused components. **Figs 5** and **6** show the channel routing and 'sea-of-gates' approaches, respectively.

The other choice to be made in a gate array is what the basic component should be. The simplest component is a two-input NAND or NOR gate. In principle, any logic circuit can be built from basic gates; you have already seen, in Figs 3 and 5 of last month's article, how an exclusive-OR gate and a D-type bistable device may be constructed. On the other hand, using two-input gates to build a 16-input composite gate, which may be needed to decode a microprocessor address, would take five levels of logic plus at least 15 gates. This would result in a long propagation delay and use a significant proportion of the gate array resources.

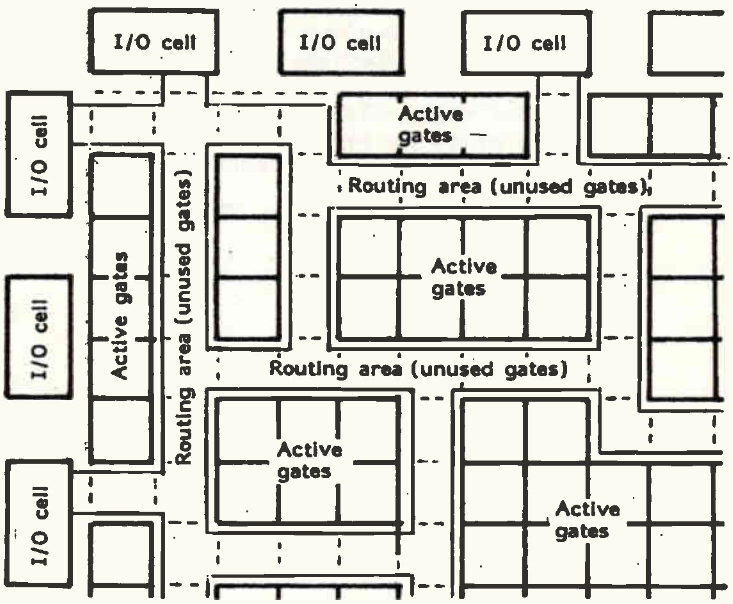


Fig. 6. In the 'sea-of-gates' asic, interconnections are made across unused components.

An alternative approach is to use a more complex cell. One example is a cell containing four p-channel and four n-channel transistors. The cell can be configured in several different ways – a four-input gate, a three-input gate plus inverter and two two-input gates are examples. The transistors can also be made into transmission gates which form the basis of bistable ICs.

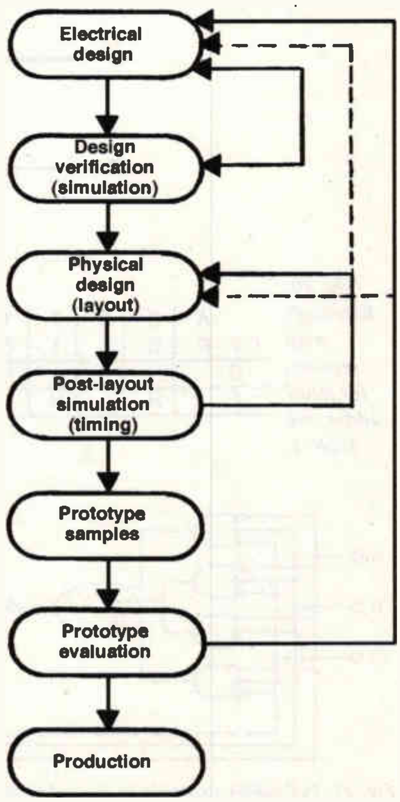
Gate arrays exist with a variety of cell designs and interconnection methods, but the ways in which designs are entered are, on the whole, quite similar.

Designing an asic

Designing a logic circuit is virtually independent of the physical form in which the circuit will be implemented. There are differences in the way in which tri-states and state machines, for example, are designed in different end products, but multiplexers, counters and other standard logic functions may be used to build up any logic system.

In effect, the designer is presented with a library of building blocks which are connected together to produce the desired result. If ttl or cmos standard logic is being used, the library will be listed in a data book which presents the relevant features of each device. These include the pin-out showing which functions appear on which pin, dc parameters showing how each device interfaces with any other, and ac parameters indicating how fast signals will pass through the system. The ac performance will be modified by

Fig. 7. ASIC design procedure.



the way the devices are physically connected; the lengths of pcb tracks and the number of inputs driven by each output affect the load capacitance and, hence, the delay through each chip.

An asic designer will also work from a library. In this case the library may also take the form of a data book but the information will be presented in a different way. For a start, there is no need for pin-out information. The circuit is most probably being drawn on a computer-aided design system. Standard symbols for gates, bistable devices and more complex structures will be used.

Fig. 8. Diode array is the simplest form of AND gate.

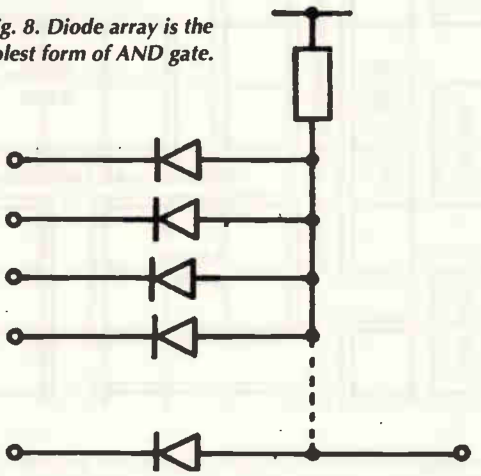


Fig. 9. MOS transistor alternative to the AND gate of Fig. 8 consumes far less power.

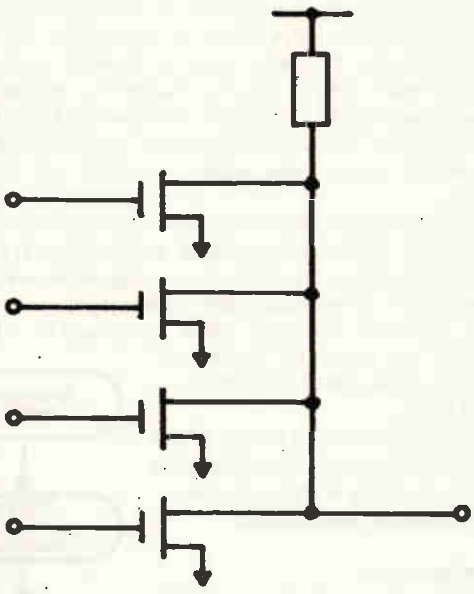


Fig. 10. Karnaugh map sections for a full adder, see Table 1.

CI \ A B	0 0	0 1	1 1	1 0
0		H		H
1	H		H	

S

CI \ A B	0 0	0 1	1 1	1 0
0			H	
1		H	H	H

CO

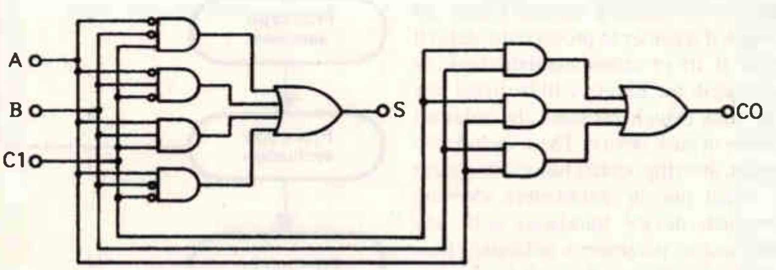


Fig. 11. Full adder example in discrete logic.

The asic equivalent of a standard logic function is a macro. This is one or more cells with the wiring to produce the required function. When a signal path joins two cells in a macro, the physical location of the electrical nodes is built into the macro description so the designer does not need to know where the connections have to be made on the chip.

Parameters for dc and ac are quite significant in asic design. Standard logic functions have built-in buffers. These ensure that all devices in a given family will drive each other with compatible voltage levels and current drains. They also guarantee a maximum propagation delay when driving a specified load. In asics there are no internal buffers, so fan-out becomes an important factor when making a design.

Simulation is an important part of the asic design process. This is not only to ensure that the logic function of the finished device meets the original goal, but to check that the design rules have not been violated. Part of the skill lies in ensuring that all parts of the design are testable, that every part of the circuit can be exercised by applying signals to the various inputs, and that the results can be seen by changes at the outputs. Bed-of-nails testing is not feasible for an asic and simulation will show any deficiencies in device testability.

Macros for ASICs are usually more 'fine grained' than standard logic circuits. To use an inverter in ttl or c-mos means either specifying a chip with six inverters or wiring a NAND or a NOR gate as an inverter, if one is conveniently spare. This may mean wasting five inverters or, if they can be used elsewhere in the circuit, running long tracks to and from another part of the pcb. Similarly, bistable devices usually come in pairs, counters in four-stage blocks, and so on.

These restrictions do not apply to asics; an inverter will, at worst, take a single cell, and the inverter may be able to be included in a cell which includes another function. Likewise, a counter with five, six or seven stages can be included without wasting any cells; moreover, features such as fast lookahead carry can be designed in to give a better performance than might be possible with discrete logic chips.

The downside is apparent when the whole design is considered. Gate arrays are made with a fixed number of cells so a design will have to be built in the device with the next higher cell count than the actual number needed. If the smallest array has 1000 cells and the next highest 1500 cells, and a certain design needs 1010 cells, then more than 30% of the chip will be wasted.

Having selected a suitable array, the next stage is to map the logic diagram onto the physical cells. This is now normally automated, the process being referred to as place-and-route. The place function involves assigning cells to each of the macros in the design, routing being the connection of the cells according to the connections between the macros already specified.

With a channel architecture, it is usually possible to use 95% or more of the cells, depending on the number of connections allowed in each routeing channel. The sea-of-gates type of array is often limited to about 60% utilisation as connections have to be made over unused cells. These usually give shorter track lengths, on the whole, as the connections can be made by more direct paths.

The final stage in the design process is post-layout simulation. The first simulation gives an idea of the delays and timing performance of the circuit, by including fan-out and a nominal delay for connections. Until the routeing has been completed, though, an accurate measure of the delays cannot be obtained. The post-layout simulation includes an estimate of the extra delays due to the actual track lengths and should give an accurate picture of the performance of the final device. If the performance does not measure up to the requirements, there is usually an opportunity to make manual changes to the layout to reduce the delay in critical paths.

A typical design flow is shown in **Fig.** 7.

Applying asics

Once a design is finalised, it will have to go to the manufacturing stage. The asic is a semi-customised integrated circuit; it starts out being processing as a standard component, because all the diffused components are independent of the final design. The final stages are, however, customised – each different design is application-specific and requires its own pattern to be imposed on the upper levels of the chip. A stock of partly finished wafers can be held in readiness for customisation.

Customisation is achieved by making masks to create the connections already specified by the place-and-route step in the design process. Four or more masks may be required – two to define the interconnection tracks to be etched into the two-track metallisation layers, and two to allow contact holes to the silicon and between the two metal layers to be made.

Mask-making and processing are both time-consuming and costly. Frequently, these activities take place in a different country from that in which the design is created, and the finished chip may be packaged in a third country. All is well if the final device performs as expected but, whether due to a design mistake or an imperfect specification, if changes have to be made, much delay and expense can result.

The non-recurring engineering, or nre, costs involved in masked asics make them more suitable for projects with expectation of long, stable, high-volume runs. As with all aspects of design and engineering, it is sometimes expedient to make trade-offs. Savings in packaging plus the need for high performance in a confined space might make it economic to use a gate array for an expected run of only a few hundred. But it is more usual to be looking at a volume in excess of 10 000 for gate array designs.

It seems from the analysis so far that circuit designers are stuck with only two choices for building logic circuits, standard families and masked asics. Fortunately, technology has evolved a third option – the programmable switch.

Programmable logic devices

There are four types of programmable switch which have been used in any volume on programmable logic devices.

The simplest form of AND gate is a diode array, **Fig. 8**. If all the inputs are near V+ then none of the diodes conduct and the output will also be pulled up to near V+. Any input taken to 0V will pull the output to a diode drop above 0V.

In a standard logic circuit, all the inputs are available at device pins, but in a programmable logic device, or pld, a programmable switch is placed in series with the input. This allows the user to select which signals will affect the gate.

A metal fuse was the first type of switch and is traditionally associated with bipolar plds. An alloy, such as nichrome or tungsten-titanium, is evaporated onto the surface of the chip and etched into small strips about 5μm wide and 20μm long. A current pulse of about 50mA is sufficient to vapourise the metal, which fuses into the overlying silicon dioxide, leaving an open circuit at the fuse site.

An alternative fuse in bipolar technology is the AIM, or avalanche-induced migration, device. This is a small transistor with a floating base, so that the emitter-collector path is normally high impedance. If the emitter-base junction is deliberately overstressed, the aluminium from the emitter contact migrates into the junction, causing a short circuit. The emitter-collector path is now a diode and can be used in its own right as a gating element; in this case, then, the fusing process is used to establish the required inputs to the gate.

In mos technology the transistors are, themselves, very efficient switches which can be turned on and off by applying high or low signals to their gates. By adding a second gate, floating between the control gate and the conducting channel, the transistor threshold can be varied by charging or discharging the second gate. In the low-threshold condition, the transistor acts normally, but in the high-threshold state the channel is held off permanently.

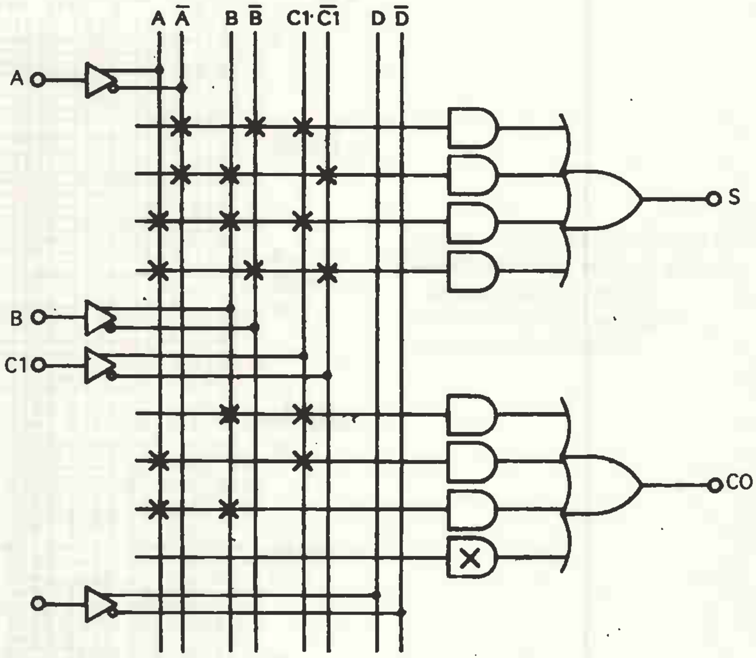


Fig. 12. Full adder of Fig. 11 implemented in a PAL4H2.

The floating gate can be charged electrically but needs ultra-violet light to discharge it. By adding silicon nitride to the sandwich, the surface states can be defined for positive and negative charges and the floating gate can be predictably discharged electrically. The transistors themselves are used for logical gating, as in **Fig. 9**.

A later development is the antifuse. This is simply a thin layer of silicon oxide/nitride sandwiched between two conducting layers, which may be either silicon or metal. A short voltage pulse of 15-20V ruptures the insulating layer and the heat alloys the two layers together. A resistor of less than 1kΩ results, sufficiently low to appear as an on switch to signals in a cmos environment.

Programmable array logic

The most commonly used plds are programmable array logic, or pals. Although not the first plds, they are the easiest to use and took the largest share of the programmable logic market.

They are based on the idea that any combinational logic function can be represented by a 'sum-of-products' equation. Sometimes, AND functions are referred to as product terms, by analogy between logic equations and arithmetic equations, and OR functions as sum terms; sum of products means just the OR combination of a number of AND terms. The justification for this concept is the Karnaugh map.

A Karnaugh map, or K-map, is constructed by taking all the inputs to a given function and drawing a grid containing all the possible combinations of high and low for those inputs. Conventionally, each axis of the grid is numbered with Gray code, with half the inputs expanded along the x-axis and the other half in the y-direction. This can be illustrated by examining the truth table and K-map for an adder; the truth table is shown in **Table 1**.

The K-maps for S and CO are shown in **Fig. 10**; each needs four AND terms on first inspection, but consider the two cells circled together in the CO

Table 2 Truth table for PAL adder

A	B	C1	S	CO
L	L	L	0	0
L	L	H	1	0
L	H	L	1	0
L	H	H	0	1
H	L	L	1	0
H	L	H	0	1
H	H	L	0	1
H	H	H	1	1

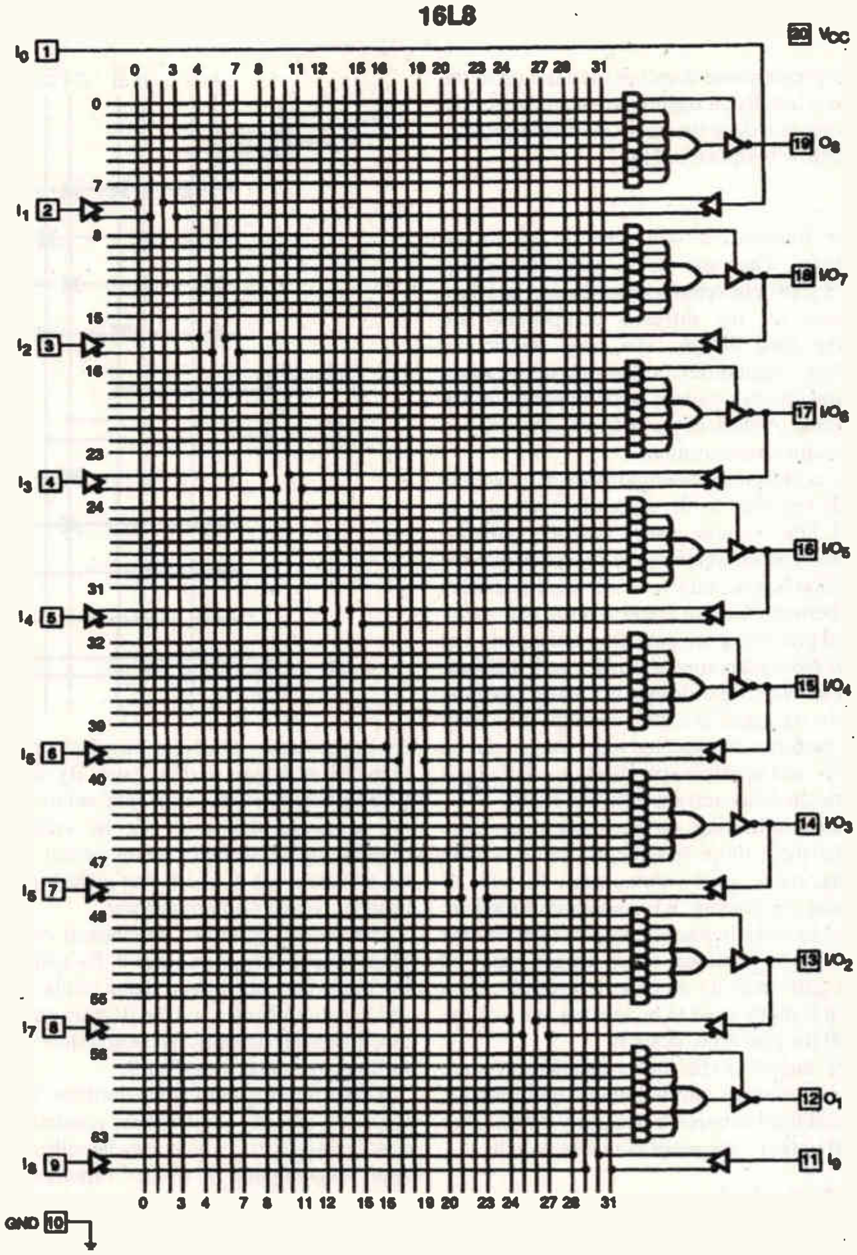


Fig. 13. PAL16L8 equivalent circuit – the simplest combinational pal currently in manufacture – reproduced by permission of Advanced Micro Devices.

map. These represent the AND terms:

!A&B&CI
A&B&CI.

This simplifies to just B&CI, because input A can take either sense. It can therefore, be eliminated from the equation. In terms of logic analysis, this is because !A#A=1.

The full equations for the adder are, therefore:

S=!A&!B&CI#!A&B&!CI#A&B&CI#A&!B&!CI
CO = B & CI # A & CI # A & B

In discrete logic these functions can be built as in **Fig. 11**.

The structure of a simple combinational pal is very similar to this circuit. **Figure 12** shows how this same function could be incorporated into an imaginary *PAL4H2*. Numbering of the pal is quite logical; in this example, the '4' refers to the number of inputs to the AND array, Letter 'H' means that the outputs are high when one of the AND terms is true (active-high) and '2' is the number of outputs from the PAL.

The eight vertical lines in the figure carry the buffered/inverted signals from the four inputs – that is, A, !A, B, !B, CI, !CI, D, !D, although D and !D are not used in this example. Each crossing point between a vertical signal line and the input line – which is really an eight-bit bus in this case – into each AND gate has a programmable switch which determines whether or not the signal is connected to the AND gate. The diagonal crosses indicate those fuses which are left intact for this application.

The simplest combinational pal currently in production is the *PAL16L8*, **Fig. 13**. As with the first example, the numbering means that this pal has 16 array inputs although, as six of these are fed back from outputs, the device has just ten dedicated inputs; in addition, there are eight active-low outputs.

Because of the feedback, any configuration from ten inputs/eight outputs to 16 inputs/two outputs can fit in to this pal. Alternatively, some of the feedback pins can be used to make functions such as latches, or to augment the seven product terms per output for more complex function.

By adding bistable devices to the outputs, it is possible to make pals capable of containing sequential functions. The *PAL16R8* has all eight outputs registered and fed back to the AND-array, hence there are just eight direct inputs plus a common clock for the bistables and a tri-state enable to allow the outputs to be connected to a bus.

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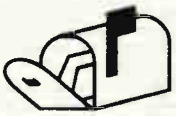

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MAGNETIC MARBLES They have been around for a number of years but still give rise to curiosity and amazement. A pack of 12 is just £3.99 ref GI/R20

NICKEL PLATING KIT Proffesional electroplating kit that will transform rusting parts into showpieces in 3 hours! Will plate onto steel, iron, bronze, gunmetal, copper, welded, silver soldered or brazed joints. Kit includes enough to plate 1,000 sq inches. You will also need a 12v supply, a container and 2 12v light bulbs. £39.99 ref NIK39.

Minature adjustable timers, 4 pole c/o output 3A 240v,
HY1230S, 12vDC adjustable from 0-30 secs. £4.99
HY1260M, 12vDC adjustable from 0-60 mins. £4.99
HY2405S, 240v adjustable from 0-5 secs. £4.99
HY24060m, 240v adjustable from 0-60 mins. £6.99

BUGGING TAPE RECORDER Small voice activated recorder, uses micro cassette complete with headphones. £28.99 ref MAR29P1.

POWER SUPPLY fully cased with mains and o/p leads 17v DC 900mA output. Bargain price £5.99 ref MAG6P9

9v DC POWER SUPPLY Standard plug in type 150ma 9v DC with lead and DC power plug. price for two is £2.99 ref AUG3P4.

COMPOSITE VIDEO KIT. Converts composite video into separate H sync, V sync, and video. 12v DC. £8.00 REF: MAG8P2.

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6"X12" AMORPHOUS SOLAR PANEL 12v 155x310mm 130mA. Bargain price just £5.99 ea REF MAG6P12.

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ROCK LIGHTS Unusual things these, two pieces of rock that glow when rubbed together! belived to cause rain! £3 a pair Ref EF29.

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ELECTRONIC ACCUPUNCTURE KIT Builds into an electronic version instead of needles! good to experiment with. £7 ref 7P30

SHOCKING COIL KIT Build this little battery operated device into all sorts of things, also gets worms out of the ground! £7 ref 7P36.

FLYING PARROTS Easily assembled kit that builds a parrot that actually flaps its wings and flies! 50 m range £6 ref EF2.

HIGH POWER CATAPULTS Hinged arm brace for stability, tempered steel yoke, super strength latex power bands. Departure speed of ammunition is in excess of 200 miles per hour! Range of over 200 metres! £7.99 ref R/9.

BALLON MANUFACTURING KIT British made, small blob blows into a large, long lasting balloon, hours of fun! £3.99 ref GI/E99R

9-0-9V 4A TRANSFORMERS, chassis mount. £7 ref LOT19A.

2.5 KILOWATT INVERTERS, Packed with batteries etc but as they weigh about 100kg CALLERS ONLY! £120.

MEGA LED DISPLAYS Build your self a clock or something with these mega 7 seg displays 55mm high, 38mm wide. 5 on a pcb for just £4.99 ref LOT16 or a bumper pack of 50 displays for just £29 ref LOT17.

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ACTIVE

A-to-D and D-to-A converters

Low-power 8-bit a-to-ds. A range of analogue-to-digital converters with serial i/o interfaces – the *WM0831/2* – are available with differential, dual, four or eight inputs, use little power and run from a single 3V or 5V supply. The *0831* has differential input, which can be single-ended, and its conversion result emerges under the control of clock and chip-select inputs. It uses a three-wire interface and a separate voltage reference input. The *0832* is arranged by the data-in input as two single-ended inputs or a differential input; voltage reference is ratiometric and is connected internally to the supply pin. Pronto Electronic Systems Ltd. Tel., 0181 554 5700; fax, 0181 554 6222.

Linear integrated circuits

1200V ic. International Rectifier claims a world's first for the *IR2233*, a 1200V, three-phase gate driver for inverters operating from 460V ac, its main advantages over earlier circuitry being that no opto-couplers or other peripheral circuits are needed; the design confers high immunity from d*v*/d*t* noise from the igbt switching transients; matched delays and timing control of the three channels; and integrated protection circuitry; and a driver stage designed for igbts. The chip contains three each independent high-side and low-side gate drives; current sense amplifier; timing and logic circuitry and interfacing for analogue and digital inputs. Gate drive output is up to 500mA. International Rectifier. Tel., 01883 732020; fax, 01883 733410.

Dtmf transceiver with micro interface. Mitel's *MT88L89/85* are 3V integrated dtmf transceivers having an adaptive microcontroller interface, the *85* also having a full-chip, power-down pin. The receiver section is based on the *MT8870* dtmf receiver, the transmitter using a switched-capacitor d-to-a converter for low distortion and accurate switching. The use of internal counters makes for precisely timed tone bursts and a call-progress tone detector, which is sensitive to –30dB, allows a microprocessor to analyse the tones and have adjustable guard time. Mitel Semiconductor. Tel., 01291 430000; fax, 01291 436389.

Cellphone amplifier. *RF2131* by RF Micro-Devices is designed mainly for use in cellular telephone handsets. It is a gallium arsenide heterojunction bipolar device 1.3W output device (on 6dBm input), working in the 800-950MHz band on a single supply of between 4V and 6V, from which it draws 400mA at 1.2W output or 10μA in standby. Power control from minimum to maximum is a 0.2-4V voltage, the process taking 100ns. The device has been tested with both capacitive and inductive duplex filters for compatibility; it is protected against overvoltage and has temperature compensation. Anglia Microwaves Ltd. Tel., 01277 630000; fax, 01277 631111.

Microprocessors and controllers

Controllers with analogue features. Microchip has new 8-bit, one-time programmable controller families that possess analogue circuitry for car electronics and appliance application. *PIC16C642/662* are the first in the 64 and 66 families and contain a 4V brown-out protection, two voltage comparators and a voltage reference module. There is 4K of on-chip program eprom and 176byte of ram. Full development support is provided by Microchip's development tools. Arizona Microchip Technology Ltd. Tel., 01628 851077; fax, 01628 850259

Mixed-signal ICs

Single-chip baseband converter. Harris has introduced the first single-chip if/quadrature modem integrating the functions of up and down conversion and quadrature modulation/demodulation. The *HFA3724* converts in both directions between baseband and .if. over a 10-400MHz frequency range and is intended for wireless lans and PCMCIA wireless network cards, time-division duplex binary and quad.-modulated comms systems, tdm multiple-access packet protocol radio, etc. Power consumption is 390mW at 5.5V and 4.4mW in standby. Harris Semiconductor UK. Tel., 01276 686886; fax, 01276 682323.

20-bit audio codec. *AK4520* from AKM is a 20-bit stereo delta-sigma codec with data converters on-chip. The a-to-d converter has 64 times oversampling and a dynamic range of 96dB at 5V and 92dB at 3V operation, while the d-to-a converter uses 128 times oversampling. Component count is reduced by the inclusion of the second-order switched-capacitor filter and continuous time filters on the chip, which also increases tolerance to clock jitter. Single-ended input and output also help to reduce the number of components and sampling rate can be from 30kHz to 50kHz, with master clock frequency at 256 or 384 times the clock. Power supply needed is 2.7/3.6V or 4.5/5.5V and power dissipation at 3V is 100mW. DIP International Ltd. Tel., 01223 462244; fax, 01223 467316.

Miniature drill. A small, 12V, 100W drill for modelling or repair work, the *MB1012* from Minicraft is available in a drill kit, which contains the *MB730* transformer with speed control and 40 accessories for cutting, grinding, routing, polishing and drilling, all in a storage case. The drill chuck takes bits from 0.3-6mm and the drill can be held by its pistol grip, as a pen or in the palm. Minicraft. Tel., 01388 420535; fax, 01388 817182.

Power semiconductors

High-current half bridges. *Super Magn-a-pak* modules by IR contain a pair of semiconductors to make a half-bridge rectifier consisting of two thyristors, a thyristor/diode combination, two diodes or two fast-recovery diodes, the modules being rated at 400-600A and meant for large motor drives and similar applications. These modules use proven IR semiconductors, formerly in B-puks. International Rectifier. Tel., 01883 732020; fax, 01883 733410.

PASSIVE

Passive components

Power chip resistors. Philips' *PRC201* 1W power chip resistors are now available in a 1% tolerance and in values down to 0.02Ω for current measurement uses. Values in the 0.02Ω-1MΩ range are made. Philips Components. Tel., 00 31 40 2722790; fax, 00 31 40 2724547.

450V electrolytics. *Nichicon PJ* series electrolytics afford high reliability and high ripple current handling. Values offered cover the 0.47μF-15000μF range in the E12 progression to within ±20% at 120Hz and 20°C. Impedance of, for example, a 1000μF 16V capacitor is 0.072Ω at 100kHz and its tolerance to ripple current 1.4A rms between 10kHz and 200kHz. The 450V versions at, say, 4.7μF handle a 55mA ripple at 120Hz. Case sizes are 10-16mm diameter and 16-31.5mm in length. Easby Electronics Ltd. Tel., 01748 850555; fax, 01748 850556.

Connectors and cabling

Mains connectors. Rendar has a new series of rewireable connectors, the *478* model, which use pillar terminations, floating, captive contacts, reversible and captive clamp and three-point cover fixing. They are easy to wire and conform to the relevant safety standards. Rating

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is 10A or 15A in the USA and the connectors take three-core cable up to 14awg. Models for hot or cold conditions to Class I or II, in straight or right-angled forms and in black or white finish are available. Rendar Ltd. Tel., 01243 866741; fax, 01243 841486.

S-m connectors. *PAK-50, P50L* surface-mounted, low-profile connectors have a mated height of 7-8mm and are designed for smt board stacking application, being available in vertical, horizontal and parallel board stacking forms. Insertion and withdrawal forces are low, due to the tapered contacts, being 3.36oz per pin and 0.49oz per pin respectively. and the contact design makes for low wear. Contact resistance is 40mΩ and isolation resistance 1GΩ. Robinson Nugent (Europe) Ltd. Tel., 01256 842626; fax, 01256 842673.

Hand-held dvms. Tektronix announces the *DMM910 Series*, an extension of the company's range of hand-held, true rms digital multimeters. The three instruments in the range are *DMM912/914/916*, the last two allowing the reading of two measurements at once and measuring temperature in both fahrenheit and celsius. In addition, the *916* has a 1ms peak hold function to capture fast transients. Of the other instruments, *DMM912* is the most basic and the 914 a mid-range type. Tektronix UK Ltd. Tel., 01628 403300; fax, 01628 403301.

Waterproof connectors. Should your interests necessitate the immersion of connectors under six feet of water, you could do no better than to consider the *Conxall Mini-Mizer* range of control and sensor connectors. They meet the requirements of UL, CSA, NEMA, ANSI and NFPA and use epoxy-sealed, all-moulded construction, with anodised, machined aluminium coupling ring and receptacle shell, which shrugs off most solvents, oils and other industrial hazards. The connectors come in two-pole to six-pole types, all having a lengthened earth pin for first-make, last-break connection. They handle 16g or 18g cable in lengths from 2m to 5m or to order. Astralux Dynamics Ltd. Tel., 01403 240055; fax, 01403 25565

Displays

Colour lcd for vehicles. Philips Flat Panel Display Co. has introduced an lcd monitor for use in vehicles as navigation displays and fleet management systems, which need not be mounted on the dashboard and incur expense for dashboard redesign. The display is based on the 5.1in *LDE052T* active-matrix colour lcd module, designed for use in tough conditions and with a backlight giving a brightness of 300cd/m². Monitors may have an RGB, NTSC or PAL interface or a 320 by 240 interface for digital indications. In addition, it can be supplied with a speaker, audio amplification, infrared remote-control receiver and backlight control. It comes in a plastic case with a rounded shape and can be mounted almost anywhere, since it has a very wide viewing angle. Philips Components. Tel., 00 31 40 2722790; fax, 00 31 40 2724547.

Filters

Inlet filter. *FN9310* inlet filter modules can be fitted in a few seconds and take up only 24.05mm behind the front panel, using a 'lock and shield' fixture, whereby the inlet automatically locks in place when it is pushed through the mounting hole, no screws being needed. Metal fingers make contact between the panel and filter earth. If attack by the muscle-bound is envisaged, an extra locking nut can be used to give more security. The inlets come in current ratings of 1-15A and, as an example, a 3A version gives a minimum of 40dB differential and common-mode interference attenuation in the 3-30MHz band. Schaffner EMC Ltd. Tel., 0118 9770070; fax, 0118 9792969.

Elliptic low-pass. Linear has introduced its second eighth-order, elliptic low-pass filter in an SO-8 pack, the *LTC1069-6*, which draws 1mA at 3V and gives 72-80dB dynamic range. It needs no other components, apart from bypasses. To allow frequency variation, the clock is external, the cut-off being 1/50 the clock frequency. Pass-band ripple is ±0.1dB, stop-band attenuation 42dB at $1.3f_{cut\text{-}off}$, 66dB at twice that frequency and 72dB at 2.1 times the cut-off frequency, which can be up to 14kHz at 3V or 20kHz at 5V. Linear Technology (UK) Ltd. Tel., 01276 677676; fax, 01276 64851.

Hardware

Emc-proof cabinets. Vero's *VERAK EMC* range of 27-42U cabinets provide over 80dB of attenuation at 100MHz. Available in plan sizes from 600/600mm to 800/800mm, the *Verak EMC* is rated at 750kg static load, being built in 2mm steel framework. Knife edges that meet the panel-mounted gasket to give the seal are part of the verticals, being self-located in slots in the end frames to give accuracy and repeatability, earth continuity being maintained throughout the cabinet. Doors can be hung either way. Steel or glass doors are available, both being emc protected and a new type of mechanism maintaining the seals. Vero Electronics Ltd. Tel., 01489 780078; fax, 01489 780978.

VXIbus chassis. Racal's *1269 VME/VXIbus* chassis allows the use of both C-size VXIbus and B-size VXI/VME boards in one unit. It complies with both VXIbus and VXI*plug&play* specifications, having six VXI C slots and three VME B slots, no adaptors being needed. The slot 0 interface allows the use of VXI or VME controllers simply by changing the orientation of a small card. Power available is 720W, with cooling, and an option is a systems monitoring facility with a front indicator. Racal Instruments Ltd. Tel., 01628 604455; fax, 01628 662017.

Test and measurement

Mobile 'phone tester. *CTS 55* from Rohde & Schwarz is a modular test set for the quality assurance and servicing of GSM and DCS1800/1900 mobiles. Main operations are call setup and cleardown for incoming and outgoing calls, free selection of channels, power and channel change, echo test, power measurement, Rx sensitivity using bit error rate measurement, phase/frequency error and power ramp v. time; none of all this needs any specialised GSM knowledge. Indications, selected from a menu, are on a tft colour display, results being presented graphically. Rohde & Schwarz UK Ltd. Tel., 01252 811377; fax, 01252 811447.

Counter. Thurlby Thandar's *SR620* is a universal timer counter to measure all the usual parameters and also pulse width, phase and rise and fall times. In addition, it is capable of statistical work, including mean, maximum, minimum standard deviation and Allan variance, all being shown on a 16-digit display. Standard timebase is 10MHz ageing at 3×10^{-6}/year, an optional oven-controlled oscillator providing 5×10^{-10}/day. Time interval is measured to 25ps resolution, intervals up to ±1000s to within 50ps relative accuracy. There is start/stop hold-off, both of these inputs functioning as start or stop. Frequency range is 1.3GHz with gates from 1μs to 500s. The instrument can send histograms and strip charts to an oscilloscope with an X input and to HP-GL plotters and printers. GPIB and RS-232 interfaces allow remote control. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Multimeters. Clarke Power Products has a new range of eleven multimeters, covering between them the requirements of kitchen-table amateurs to professionals, at prices from £6.95 to £64.95. Clarke Power Products. Tel., 0181 986 8231; fax, 0181 986 6512.

Cable tester. *Microscanner* by Microtest Inc. is the first in a family of cable test instruments, this one designed to verify and fault-find in twisted-pair cabling. By combining wiremap with length measurement, it

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confirms continuity, wiring connections, indicates fault location and identifies wires. This is a hand-held unit, which will measure the length of a cable using tdr, identifying shorts, open circuits and connections to hubs, as well as tracing hidden cables in walls and floors. Microtest Europe. Tel., 01293 894000; fax, 01293 894008.

Transient capture analyser. Endevco offers the *Model 28989* transient analyser, which has four channels and works to a host pc. Its four voltage inputs store up to 50,000 samples/channel, each having a 12-bit analogue-to-digital converter for simultaneous, selectable sampling rates to 20kHz with one trigger channel active. Triggering is internal, external or via Windows software, repeated runs being performed by a mouse-controlled setup button. Both mains and dc-powered versions are available. Endevco UK Ltd. Tel., 01763 261311; fax, 01763 261120.

Microwave counters. *Escort 3000* counters is a range of four instruments covering in total 0.001Hz-3GHz on an eight-digit display. The four models are EFC-3305 giving 0.001Hz-100MHz and 80MHz-3GHz; the EFC-3303 reading 0.001Hz-100MHz and 50MHz-1.3GHz; 5Hz-100MHz and 50MHz-2.4GHz on the EFC-3203A; and 5Hz-175MHz with the EUC-3200. Various functions are available on some of the models, such as autoranging, ac/dc coupling, pulse-width average and rev/min measurement. Feedback Test and Measurement. Tel., 01892 653322; fax, 01892 663719.

Interfaces

***Combiscope* adaptor cable.** Fluke offers the *PAC33* print adaptor cable, which enables any Fluke instrument, such as the *PM 3394B Combiscopes*, with a standard RS-232 interface to be connected to any printer having a Centronics interface. It uses standard RS-232 and parallel connectors, a 9V battery and will take hard use. Fluke UK Ltd. Tel., 01923 240511; fax, 01923 225067.

Literature

Emc testing. Seaward Electronic can provide a free booklet on the subject of emc testing in-house. It contains a guide to the EMC Directive, standards and methods of ensuring compliance; there are lists of basic, generic and product standards and descriptions of the Declaration and Technical Construction File. Technical descriptions are provided of rf, harmonic and flicker emission tests, esd, eft voltage dips and surge immunity tests. Seaward Electronic Ltd. Tel., 0191 586 3511; fax, 0191 586 0227.

Materials

Emc gasket. Holland Shielding Systems BV can provide gaskets that give good shielding, extremely low closing force, 80% compression and negligible compression set. Dimensions range from 2 by 2mm to 40 by 60mm, all being available on a 200m roll. The gaskets can be supplied with an environmental seal and/or a self-adhesive layer, and there are flame-retardant versions. Holland Shielding Systems bv. Tel., 0031 78 613 13 66; fax, 0031 78 614 95 85.

Printers and controllers

Thermal printer. Seiko's *LTP4242* lightweight thermal printer measures only 86.6 by 52 by 22mm, has a 75mm/s printing speed and is meant for journal and receipt printing. It is designed for ease of use, having simple paper insertion, rapid head cleaning and replacement and a new static electricity protection method by grounding the frame, all metallic components being in ground contact, so preventing damage. Resolution is 432 dots per line, printing width 54mm on paper 58mm wide and the supply needed is 5V and 24V. Craft Data Ltd. Tel., 01494 778235; fax, 01494 773645.

Double-Y pen recorder. Kipp and Zonen of Delft produces what it says is the only such instrument available with CE approval. *BD90/91* flat-bed recorders both have single and dual-pen *XY-t* facility and are for general-purpose use. They have modules for voltage, timebase sweep and can be remotely controlled. Kipp & Zonen. Tel., 01727 843939; fax, 01727 842185.

Production equipment

Ovens. A range of electric ovens by Hedinair is batch-produced, in contrast to the common arrangement in which companies build them to order; in this way, cost is minimised. The range contains ovens operating at up to 200°C, 300°C and 425°C, all of these being available in sizes up to 915mm cube internally. There is sufficient power to give a rapid heat-up time, air recirculation providing even heating, and a good exhaust rate means that solvent-based coatings on components can be tolerated. For these, a separate 50m³/h exhaust fan can be used, with an optional switch to turn off the heating if the air flow stops. Hedinair Ltd. Tel., 0181 590 2090; fax, 0181 590 0262.

Power supplies

80A dc-to-dc converter. Vicor has a dc-to-dc converter that accepts an input of 250-425V and produces an output programmable from 1V to 5.5V at 80A and 84% efficiency. Power density is 80W/in³, so that the converter helps to reduce power-supply size and zero-current switching has reduced input conducted noise and output ripple. Dimensions are 4.6 by 2.2 by 0.5in and the shape of the converter allows it to be recessed in the board to give a height of 0.43in. Vicor UK. Tel., 01276 678222; fax, 01276 681269.

600W switcher. From Coutant Lambda, the *Alpha 600* switched-mode power supply, the latest member of the range, which provides between one and seven outputs to customer configuration and meets all relevant standards. Input voltage is universal at 47-63Hz and start-up time is under 0.5s. Thermal, overcurrent and overvoltage protection are standard, as is power-factor correction. Output is adjustable; load regulation for a 100% change in load is <0.5% with remote sensing and <2% without. Options include curve B emc filtering, remote sensing, power good, ac failure parallelling and power supply inhibit. Coutant Lambda Ltd. Tel., 01271 865656; fax, 01271 864894.

Open-frame, 55W. Computer Products offers the *NAN55* open-frame, 55W (with a fan) power supply in the same size (5 by 3in) as many 40W units on the market. Input is universal and there are single, dual and two triple output models providing 5V, 5.2/12.1V, and 5.2/12.1/–12V. The 5V or 5.2V output of each is regulated to ±2% and is protected. All the relevant standards, directives, laws, regulations and edicts have been respected. Computer Products, Power Conversion Ltd. Tel., 00353 24 93130; fax, 00353 24 03257.

Emc-testing antenna. Luminaires have their own European standard, EN55015, which requires that a large loop antenna be used for emc measurement. Laplace has a 2m diameter loop for this purpose covering 9kHz-30MHz, which is supplied with antenna factor data to enable its use with any emc receiver or spectrum analyser capable of factor compensation. Mounted on a collapsible wooden frame, it is a three-axis antenna with a switch to select each loop and screened housings for the current transducers. Software supplied with the kit automatically compensates for antenna characteristics, displaying EN55015 limits corrected for the 2m loops. Laplace Instruments Ltd. Tel., 01692 500777; fax, 01692 406177.

Protection devices

Resettable fuses. Raychem has additions to its range of polymeric resettable fuses, the *miniSMDC110/075/050/020* which extend the range of protection from 200mA to 1.1A, while enabling visual inspection of solder joints on a board. The low-resistance 110 has a 1.1A hold current and is rated at 6V, the new package helping with joint inspection. The *075* and *050* have the same performance as *miniSMD075/050* with hold currents of 0.75A and 0.5A, trip times being under 0.5s. The *020* holds at 200mA and trips 30% faster than the earlier SMD type. Raychem Ltd. Tel., 01793 572692; fax, 01793 572209.

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Switches and relays

Reed switches. Gentech *Series GR10* miniature Form A pressurised reed switches have a maximum switching voltage or 500Vdc, maximum current 1A, dc contact rating of 15W and initial contact resistance of 0.1Ω, contacts being of ruthenium. Diameter of the switches is 2.54mm and they are 20mm long, with an uncut length of 53mm and minimum length of 25mm cut. They can be supplied potted for pcb mounting and sensing use. Gentech International Ltd. Tel., 01465 713581; fax, 01465 714974.

Transducers and sensors

Optical encoder. Intended for use in applications needing what is effectively a highly accurate potentiometer, Control Transducers' optical encoder meets the requirement. The *Model DP100BT Digital Potentiometer* can be mounted on a board or a panel and produces 100 ttl two-phase square waves per revolution, which corresponds to 400 2-bit codes/rev., other arrangements being available. The device is rated for over five million revolutions. A free designers' guide is on offer. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

'Smallest' 50W open-frame. Vero's *BVS45/BVM45* open-frame, switched-mode power supplies measure 108 by 64 by 29mm and come in single, dual or triple output versions. Inputs are universal 47-440Hz ac and outputs various combinations of 5V, 12V and 15V. Overcurrent, overvoltage and short-circuit protection is provided and the units meet the relevant standards requirements for radiated and conducted noise. Hold-up time is over 70ms at 230V ac. Vero Electronics Ltd. Tel., 01489 780078; fax, 01489 780978.

3000A current sensor. Davtrend has a ±3000A current sensor for the measurement and control of static or dynamic ac or dc in plant and machinery control systems. It operates from dc to 25kHz to an accuracy of ±0.25% of range, with a gain drift of ±0.01% and zero offset drift of ±0.01% of range per degree. Response time is 10μs, coping with *di/dt* of 300A/μs and output is 0 to ±10V for 0 to ±3000A at 100Ω; these characteristics can be varied by the company to suit other uses. The sensor accommodates a 50mm busbar. Davtrend Ltd. Tel., 01705 372004; fax, 01705 326307.

Position sensors/systems. *MTN/1E* position sensors and electronics by Monitran are made in stainless steel and sealed against attack by all manner of dirt and fluids, connectors being of the heavy duty type, sealed and, as an option, armoured. Various fittings such as spring-loaded ends, rod end bearings and threaded extensions are available. Ranges cover ±0.5mm to ±550mm, outputs being ac, dc or 4-20mA. Instrumentation can be supplied for most engineering parameters. Monitran Ltd. Tel., 01494 816569; fax, 01494 812256.

Coded car keys. Philips' newly announced car immobiliser keys are designed to foil those lurking in car parks and trying to intercept the radiated codes as people unlock their cars. Both key and car process a random number generated by the car; both have to match before the car lets you in and, since the code is not transmitted, it cannot be intercepted, the transmission consisting only of random numbers. Keys cannot be copied. You can educate the car to recognise several keys so that it will go through its performance of adjusting mirrors and seats to suit whoever has come in. If you address it rudely, it will probably explode. Philips Semiconductors (Eindhoven). Tel., 00 31 40 2722091; fax, 00 31 40 2724825.

COMPUTER

Computer-aided design

Image acquisition. National Instruments introduces the *IMAQ PCI-1408* image acquisition board for Windows 95/NT, which is accompanied by *IMAQ Vision* acquisition, process and analysis software and the *NI-IMAQ* driver software. The board works with several video standards and the software with National's virtual instrumentation packages. Used together, the package assists in on-line inspection, gauging, process control, component sorting and handling and bar code reading. National Instruments UK. Tel., 01635 523545; fax, 01635 523154.

Computer board-level products

Backplane. Vero has a new range of 10-layer, 3U backplanes for industrial use. *CompactPCI* is electrically compatible with the pc desktop specification, so that existing applications can be run in a rack-mounted computer. A Hard Metric 2mm connector gives 235 pins per slot, 25 being user-defined. The backplane supports 32 and 64bit PCI transfers at up to 132Mbyte/s and provides independent power rails for 5V and 3.3V, with decouplers. Throughout the connector, 28 pins are ground and defined early power pins enable hot swapping. Vero Electronics Ltd. Tel., 01489 780078; fax, 01489 780978.

Single-board VME. The 3.3V *B10* single-Eurocard cpu from Men GmbH of Nuremburg offers scalable performance up to 100Mips by the use of a 68040 or 68060 with a 68360 risc processor in companion mode to handle Ethernet and all other serial input/output; the Ethernet interface supports AUI, Cheapernet and twisted pair. The four serial ports can be RS232, 422 or 485 and, as an option, optically isolated. Memory takes the form of 32Mbyte of dram, 8Mbyte of flash, 2Mbyte of sram and a 1Mbyte eprom, the dram being in a PS-2 simm and therefore proof against shock and vibration. External mass memory devices use a SCSI-II controller supporting direct memory access and data transfer to 10Mbyte/s. Tellima Technology Ltd. Tel., 01484 866806; fax, 01484 866816.

Dsp starter kit. NEC has a starter kit, the *EB-77016STARTER*, for the company's μPD7701X family of dsp processors. The kit has the μPD77016 ram dsp processor, workbench and debugger software running under Windows. It is connected to the pc's printer port, program development proceeding by way of the JTAG debug interface of the processors. Programs are then down-loaded into the built-in flash memory and booted into the μPD77016 to allow operation independently of the host pc. Also in the kit is a stereo a-to-d-to-a converter, power supply, cables and documents, with some programming examples. NEC Electronics (UK) Ltd. Tel., 01908 691133; fax, 01908 670290.

OS-9-to-dos/Windows. Version 3.2 of BVM's *PCLink*, the networking package for OS-9 to dos/Windows 3.x, supports dynamic data exchange to allow real-time updating from an OS-9 machine to a standard Windows application running on a pc, so that packages such as Scada on a pc can be used to store and display system parameters with no complicated programming being needed. The OS-9 machine can access all pc peripherals and the pc has similar access, supporting most OS-9 calls and remote OS-9 commands, including remote login. *PCLink* allows networking of the two types of system over a number of media, and coexists with Microware's ISP and other Ethernet protocols. It can also be used on RS-232 or RS-485 ports at up to 2Mb/s. BVM Ltd. Tel., 01489 783589; fax, 01489 780144.

Data acquisition

1MHz acquisition boards. *Win-30* data acquisition boards by United Electronic Industries Inc. are multi-function analogue and digital i/o boards supporting Windows 3.1, 95 and NT software. Throughput on all boards is 1MHz using data packing and bursting dma or 32-bit rep string operations, sampling up to 16 channels at 62.5kHz or 1 channel at 1MHz. Software controls input ranges, dma and interrupt levels. Other boards in the series have a variety of facilities, including various input types, input amplification, buffering and programmed gain. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 01273 570215. ■

MEASURING low resistance

Simple low-resistance measurement usually relies on high direct current to produce a voltage drop, stressing the component under test. Reducing the current and amplifying the voltage drop can result in offset problems. Frantisek Michele explains the advantages of using ac to produce the drop.

Measuring resistors with low values can be tricky. The obvious method is to measure the voltage drop across the unknown resistor using a known current and then calculate its value.

Because the drop depends on the current through the resistor, the current needs to be large enough to produce a measurable voltage. For example, the voltage drop is only 10mV if the measured resistor is 0.1Ω and current through the resistor is 100mA.

Large currents supply large voltage drops. However, in many cases, the measured components will not tolerate such large currents.. Also, the heat generated by within the component due to the large current can cause measurement errors.

This problem can be solved by amplifying the voltage drop so that less current is needed. If the amplifier has a 60dB gain, the output will be 0.1V if the current is 1mA and the resistor is still 0.1Ω.

The ac alternative

Operational amplifiers however have a dc input offset voltage. This offset causes an error when the input level is very low. An ac amplifier technique circumvents this problem.

Referring to Fig. 1, IC_A, C_1, and $R_{1\text{-}4}$ form a square waveform generator operating at around 300Hz. Diode D_1 limits the square wave to 6V peak-peak. Because the values of the measured resistor R_X and additional resistor R_A are much less than R_6, current through R_X will be,

$$I_X=6/R_6=2\text{ mA}$$

Then, IC_B's input is,

$$V_{IN}=I_X\times(R_X+R_A)=0.002\times(R_X+R_A)$$

Amplifier IC_B supplies the ac gain $R_8/R_7=10$. Diode D_2 with IC_C converts the ac signal to dc with gain of $1+R_{11}/R_{10}=10$. DC amplifier

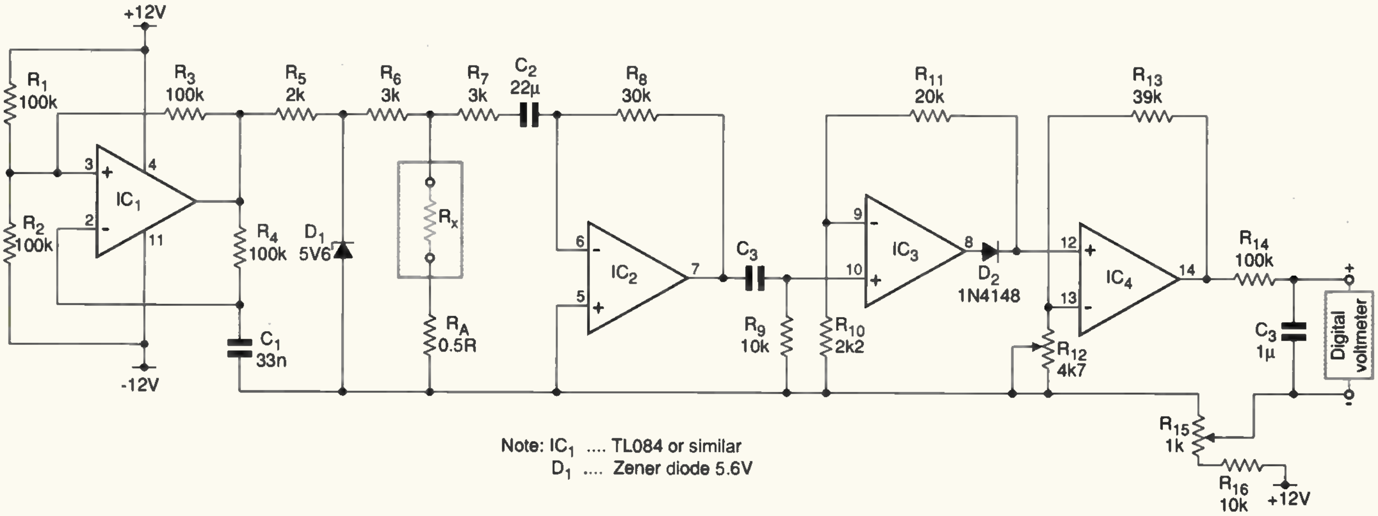

Fig. 1. Oscillator IC_A produces an ac voltage over the unknown resistor, which is then amplified, rectified and displayed. Since the stimulus is ac, high current and drift associated with dc schemes for measuring low resistance are removed.

IC_D has a gain of $1+R_{13}/R_{12}$. As a result, the output is,

$$V_O=0.5\times VIN\times 10\times 10\times(1+R_{13}/R_{12}$$
$$=0.1\times(R_X+R_A)\times(1+R_{13}/R_{12})$$

where 0.5 is the conversion efficiency for a 50% duty-cycle waveform. After the dc output is smoothed by R_{14} and C_4, a digital voltmeter can measure R_X.

Resistor R_A supplies a base signal for the amplifiers. When R_X=0, R_A sends a 1mV peak-peak signal to IC_B. If R_A=0 and R_X is very small, IC_B's noise may swamp the weak input. To compensate for the output offset due to R_A, R_{15} calibrates the digital voltmeter to zero when R_X=0. Adjusting R_{12} makes the scale 1Ω/V. Thus, a 2V digital voltmeter can measure resistances from 0.001 to 1.999Ω.

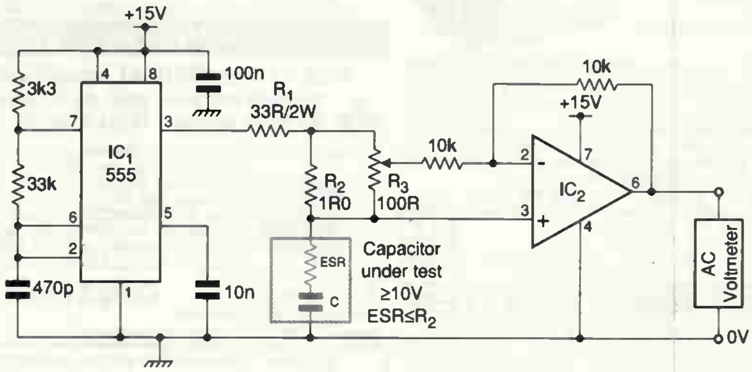

Fig. 2. Equivalent series resistance in a capacitor is particularly important in circuits such as switch-mode power supplies. Operating at 50kHz, this circuit drives about ±180mA into the capacitor and looks at the resulting ac voltage over the 1Ω resistor via an op-amp.

Measure resistance of a capacitor

The equivalent series resistance, or esr, of a capacitor can be measured using **Fig. 2.**

Oscillator IC_1 forms a 50kHz square-wave generator. It drives a current waveform of about ±180mA into the capacitor under test through R_1 and R_2.

When R_3 is adjusted to the proper value, the voltage drop across the equivalent series resistor is precisely nulled by the inverting amplifier IC_2. Thus, V_O is the pure capacitor voltage which is the minimum voltage that can be produced at V_O.

To make an ac voltage measurement, adjust R_3 until V_O is minimised. Then note the position of the potentiometer and multiply it by the value of R_2. That product equals the capacitor's esr.

The capacitor is biased about to 7.5V. Lower-voltage capacitors cannot be measured with this circuit. Changing the value of R_2 allows other ranges of.esr to be measured. However, for small R_2 values, the current level should be increased to keep a reasonable voltage across R_2. This will require some sort of buffer.

The circuit is intended for capacitors greater than 100μF. Ripple voltage becomes large for smaller values and accuracy decreases. ■

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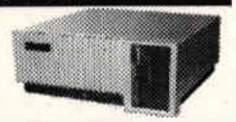

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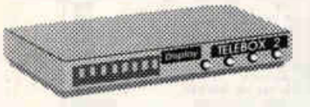

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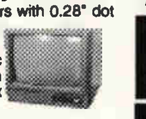

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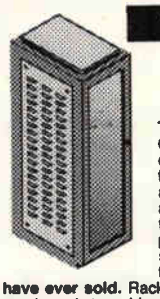

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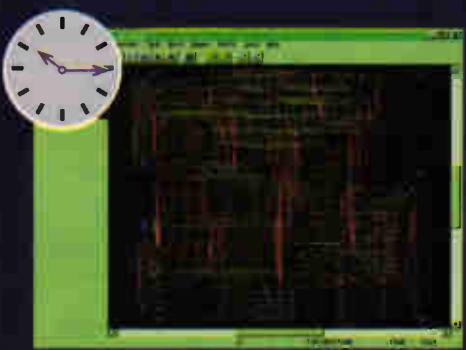

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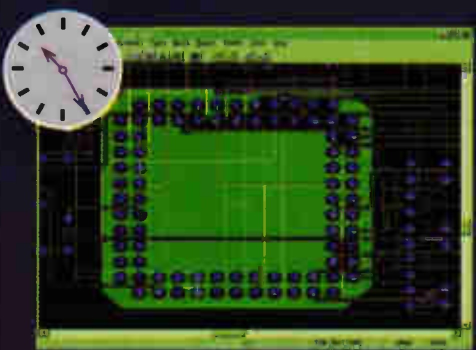

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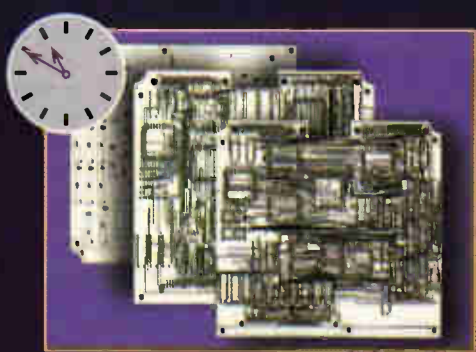

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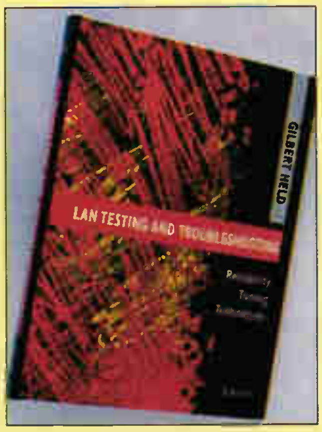

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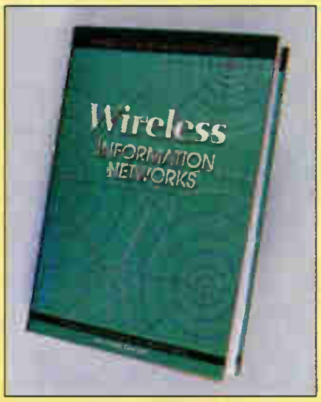

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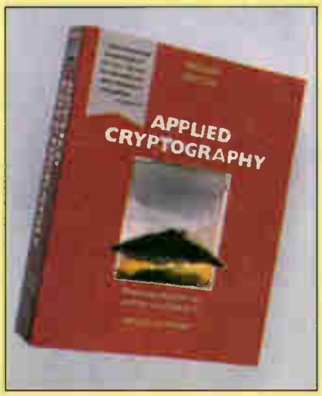

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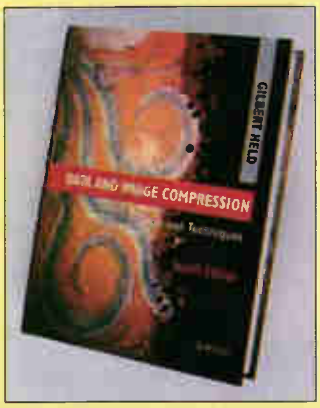

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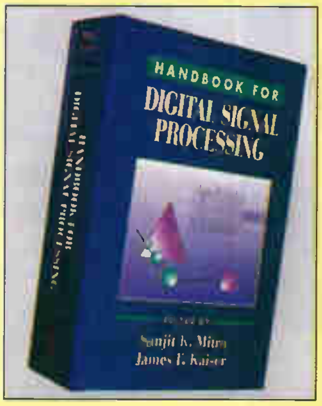

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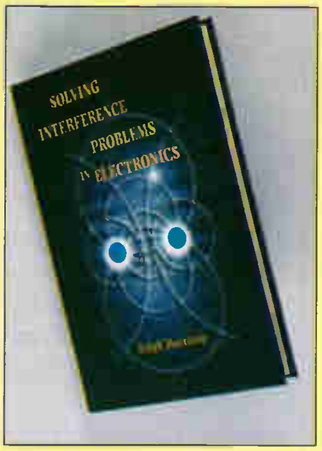

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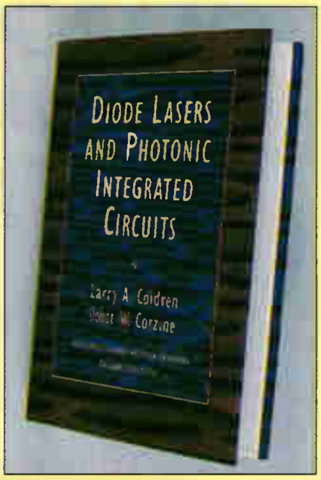

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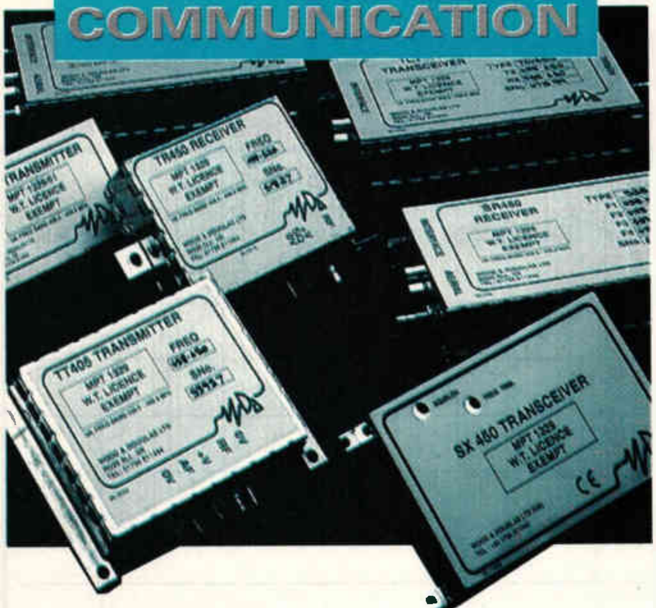

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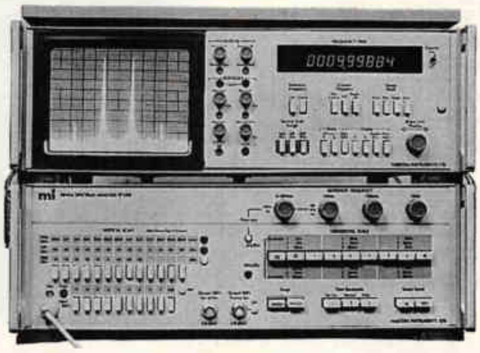

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LETTERS

Letters to "Electronics World"
Quadrant House, The Quadrant,
Sutton, Surrey, SM2 5AS

Splitting phase

I have been following with interest the recent correspondence regarding the concertina phase-splitter circuit. Like Mr McFadden, I too was a designer of valve audio amplifiers in the 1950s.

In the July/August 1996 issue of *Electronics World*, Morgan Jones attempted, unsuccessfully, to derive the Langford Smith formula for the voltage gain. In fact, the derivation is quite straightforward.

The diagram shows the usual linear equivalent circuit for a triode amplifier.

From this circuit:

$$V_{gk}=V_i-V_k \quad (1)$$

$$i(R_a+r_a+R_k)=\mu V_{gk}=\mu(V_i-V_k) \quad (2)$$

Also:

$$i=V_k/R_k \text{ and } i=-V_a/R_a \quad (3)$$

Substituting separately for these values of i gives, respectively, the gain to the cathode, G_k, and the gain to the anode, G_a.

$$G_k=V_k/V_i=\mu R_k/[R_a+r_a+R_k(1+\mu)] \quad (4)$$

$$G_a=V_a/V_i=-\mu R_a/[R_a+r_a+R_k(1+\mu)] \quad (5)$$

From these two equations, you can see that the respective output impedances are,

$$R_o(\text{cathode})=(R_a+r_a)/(1+\mu) \quad (4a)$$

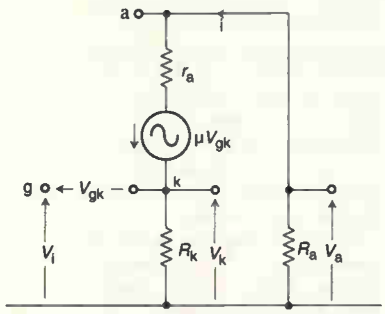
$$R_o(\text{anode}) = r_a+R_k(1+\mu) \quad (5a)$$

For the concertina phase-splitter, the anode and cathode loads are identical, $R_k=R_a=R_L$

Substituting for R_L in equations (4) and (5) gives the Langford Smith gain equation,

Gain, $G=\mu R_L/[r_a+R_L(2+\mu)]$

Clearly, it matters not whether the load R_L is a pure resistance or a complex impedance: the gain from both the anode and cathode will be identical. Further, providing the two loads are identical, gain is always less than unity, and typically 0.9.



Linear equivalent circuit for a triode amplifier

Consider what happens, however, if the anode and cathode loads are not identical. This situation could easily arise if a Class AB_1 push-pull amplifier is overdriven so that grid current flows during part of the positive half-cycle. I have taken an example of an *ECC82* triode operating with anode and cathode loads of 50kΩ and a mean anode current of 2.5mA. Under these conditions μ is 20 and r_a about 17kΩ. It is assumed that, for part of the positive output swing, when grid current flows in the power valves, the load resistors are shunted on alternate half-cycles by an additional 50kΩ. The changes of gain at the anode and cathode are shown in the table below.

You can see that variations in the anode and cathode load resistance have very little effect on the gain to the cathode but a significant effect on the gain to the anode. Furthermore, when the shunt load is across the cathode during its positive half-cycle, enhanced gain at the anode can saturate the valve.

	G_a	Gk
No shunt load	−0.895	0.895
50Ω shunt, anode	−0.458	0.916
50Ω shunt, cathode	−1.689	0.845

These problems can be minimised by operating the valve with lower values of load resistance, eg 10kΩ, and with a proportionately higher anode current. A better solution, however, is to use a different phase-splitter, such as a long-tailed pair.

Keith Thrower
Reading,
Berkshire

Improving power factor

I read with considerable interest Irving Gottlieb's article 'Improving Power Factor' in *EW* July/August 1996. It reminded me of my first job 30 years ago, designing tv deflection and high tension circuits.

Deflection current is roughly sawtooth-shaped, and resetting the current after each line period causes a large di/dt. The resulting induced voltage pulse in a secondary winding on the deflection transformer was rectified to generate the 25kV required for those days' colour tvs. Since the voltage pulse was narrow (I think some 10μs) with a prf of 15,625Hz, the high-voltage had a rather high source resistance, causing it to drop with increasing picture brightness, which of course 'blew-up' the picture.

We decreased source impedance by tuning the transformer to the 3rd harmonic of the tv line frequency of 15,625Hz. For even better performance we timed to the 5th harmonic, making the pulse shape even more rectangular. (See Gottlieb's Fig. 3).

Interestingly, the inductance part of the resonant tank was formed by the coupling induction between the primary and the secondary. Only a capacitor was then necessary to complete the circuit.

This method may not be practical for mains frequencies. However, tuning to higher harmonies might improve the power factor and also allow use of (physically and electrically) smaller tank parts. I

Was LVD message from outer space?

Regarding John Woodgate's letter, I don't think many readers of *Electronics World* who are on the receiving end of the new EEC directives be impressed by his pedantic distinction between the 'old' LVD of 1976 with which they will be familiar and the 'new' LVD of this year, which presents an entirely new dimension. The 'new' LVD uses the CE Mark as a type of public announcement of a legal commitment and woe betide anyone who falls foul of it. One wonders what the motive was in trying to make such a distinction.

I find his bland assertion that, "The chance of any large company trying to use a Directive against a very small competitor is minimal," quite astonishing. Is Mr Woodgate living on the same planet? One only has to pick up a newspaper to read every day of some new incident involving sleaze, criminal dishonesty and malpractice in business at all levels.The new LVD presents such a tempting opportunity for amoral big business to defeat competitors that it will be irresistable. Unfortunately, with this country's obsession with secrecy no-one will ever know how much of this connivance will be taking place, because you can be sure a clock of confidentiality will be drawn over it.

Thirdly, Mr Woodgate states that it is, "important for UK companies not to shut down for fear of the directives." By using the word fear he shows an almost total lack of understanding of an important part of this sorry business. If he had read the letter from Greece (Ann Baker, Letters in *EW* Jan '97) he would have realised that it is not some irrational dread of the directives themselves, nor even of the draconian punishments threatened, although they are bad enough, but genuine anger and disgust at the way the directives have been implemented.

This is what all the fuss is about. The disproportionate cost to small companies of implementing the directives is of most concern. These two directives make life particularly difficult for small firms but hardly affect large companies, and I speak from my own experiences. This extra overhead will put up the cost of all electrical products made in the EEC by small firms, and will deter all grass-roots innovators. As the letter from Greece points out, this will backfire on the EEC as the bureaucratic and financial burden becomes too much to make new projects worthwhile for small companies. No doubt a few will persevere, but this is not good enough. As for third world and Far East exporters to the EEC, they will have a field day – and if any unworldly reader really wants to know how this will be done I can enlighten them.

I am sure small firms and designers on the brink of introducing new projects that need to pass the two hurdles will be asking the same question that I have been asking myself; is it any longer worth the trouble?

I can confidently repeat my assertion that the outlook for these firms in Europe is bleak, and has been made deliberately so by the EEC.

If Mr Woodgate spoke on behalf of a small manufacturing firm, his words would be more believable. Instead, his letter could have come straight from an EEC Ministry of Propaganda.

Rod Cooper
Sutton Coldfield
West Midlands

wonder if Mr Gottlieb would care to comment on this.

Jan Didden
Heerlen
The Netherlands

Twisted cables?

With reference to Cyril Bateman's article in the February issue, as far as I can understand it, he claims to have proved that three hitherto unknown factors dominate loudspeaker cable performance:

- Addition of approximately 50kHz ringing to a special test signal when a load resembling no known loudspeaker is used.
- Generation of harmonic distortion by copper cables. This would be revolutionary if true, but it isn't.
- Generation and/or pickup of inharmonic distortion by copper cables. This would be amazing if true; it isn't.

While the first of these 'discoveries' appears to be merely irrelevant to audio, the second two seem to be important enough that some details should be given. However, Mr Batsman uses a high-distortion oscillator, which is cleaned up by an unspecified amount by an unspecified filter. He gives no details of the harmonics generated, or the nature of the thd residual. Nor, as far as I can tell, does the test set-up used correspond to any of the circuit diagrams given, which hardly aids those who might be concerned to replicate his experiments. The Blameless amplifier design he uses is certainly stable with any normal source impedance and should give less than 0.001% thd at 1 kHz, if properly constructed.

I can see no reason why anyone should feel obliged to repeat Mr Bateman's experiments before commenting on them. As an attempt to stifle debate, it is unsubtle. And it is hardly reasonable to insist on replication if insufficient details are given to make this possible. Nonetheless, I have found time for a few minutes of my own experiments, and as a result I think I can help Mr Bateman with "the resolution of the matter" of these mysterious distortions.

Figure 1 shows a standard method for measuring a speaker impedance curve. The speaker is driven by an Audio Precision low-distortion test-set through a high resistance compared with its impedance, so the voltage measured across it – with a suitably chosen drive voltage of 6V rms – can be read off directly in ohms-magnitude. The unit I tested was a conventional two-unit reflex design of bookshelf size, **Fig. 2.** Bass resonance is double-humped due to the tuned port; the peak around 2kHz is more likely to be due to the crossover rather than tweeter resonance.

If a loudspeaker is so driven, non-linear distortion can indeed be detected at the loudspeaker end of the 600Ω driving impedance. With the values shown thd is about 0.3% around 500Hz to 1kHz, falling off as frequency increases above 1kHz. Clearly, the only possible source of non-linearity is the loudspeaker. Prime suspects are the BiL non-linearity, and cone suspension force/displacement non-linearity of the bass unit; which is more important is not clear at this point.

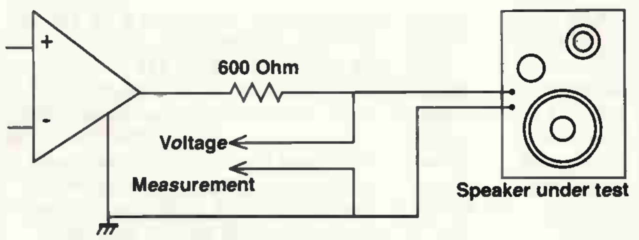


Fig. 1. Standard method for measuring a speaker impedance curve.

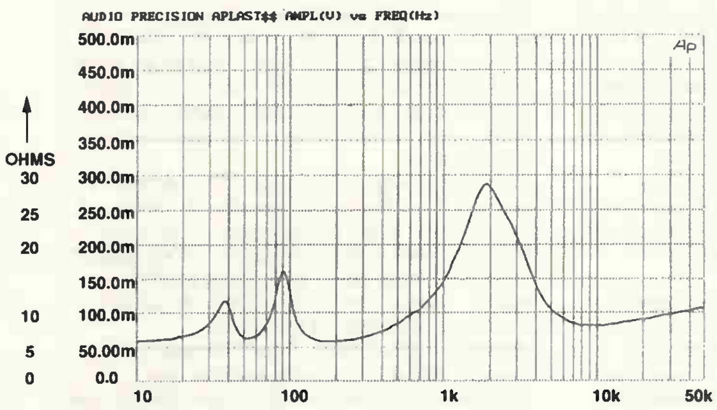


Fig. 2. Impedance curve of conventional two-unit reflex design of bookshelf size. Bass resonance is double-humped due to the tuned port; the peak around 2kHz is more likely to be due to the crossover rather than a tweeter resonance.

In real speaker use, the driving impedance will be at least two orders of magnitude lower and the thd measured correspondingly less; in any case it is not exactly obvious whether this distortion would improve or impair the linearity of the overall voltage-in/sound-pressure-out relationship of a loudspeaker.

As for 'inharmonic distortions', my experiments suggest that their origin is the loudspeaker acting as a microphone and adding a room-noise signal to the harmonic distortion described above. A computer in the same room generates enough noise to swamp thd readings at low frequencies; traffic rumble is also a problem.

I believe that this resolves the matter of the mysterious distortions.

Having produced some numerical rankings – which I decline to accept – Mr Bateman then seems to leap to the conclusion that the vital parameters for speaker cable are minimal resistance, inductance, and characteristic impedance. Since the latter is meaningless for less than one-sixth of a wavelength – ie about 2 miles at 20 kHz – it can be ignored. Minimal resistance and inductance remain as criteria, and these are perfectly reasonable, and wholly conventional requirements, which appear to have been reached by a very convoluted route.

What is painfully absent is any estimate of just how low these parameters should be before they cease to cause any audible modification. His preference for coaxial cable appears to be solely based on its lower inductance, and there seems to be no evidence that the substantial extra expense over lawnmower cable is justified.

My conclusion must be that I cannot discern any new discoveries at all in Mr Bateman's complicated investigations."

Doug Self
London

Cyril replies...

The three conclusions Douglas has drawn bear no similarity with any conclusion I hold as a result of what eventually became several hundred hours experimentation and reporting. Douglas has closed his mind to this work, preferring to act as self appointed judge, jury and executioner on this topic.

I will try to deal with each of his points in turn.

As to my simulations using a capacitive load 'resembling no known loudspeaker', I can only commend that he measures both impedance and phase of speakers, by whatever method he prefers: and calculate the equivalent reactive loading, since ignorance of this can result in much confusion. Figure 4 in the February issue clearly shows my test speaker equating to 9μF in parallel with 10.97Ω at 3k3Hz, while Fig. 5 clearly shows an inductive reactance at 1kHz of 3.3Ω or 5.3mH for the same speaker.

As to my distortion measurements, I clearly identified the distortion level of my 'cleaned up' generator and test amplifier when loaded with 8.2Ω in the text, as 0.045%. I also made it clear that comparative, conventional thd was measured, page 122 column 1 para 2, 3. Furthermore on the two distortion results, Figs 6, 7, I deliberately included measurements at both amplifier output and speaker terminals on the plots in order to make absolutely clear any changes in my test source behaviour.

The reason I suggested readers should repeat these experiments was to encourage informed debate – certainly not to 'stifle debate'. For this reason, each test's conditions was carefully detailed, including PSpice netlists, components and circuits used if other than completely conventional and obvious to all. I seem to have failed in both where Douglas is concerned.

Since phase measurements were crucial to the investigation: Douglas's simple text book impedance test was obviously quite unsuitable, although I did describe it in a few words on page 121.

As to his distortion measurements I note he carefully avoided measurement of his speaker around 3kHz, when he might well have observed rather more interesting results. Also, he carefully avoids stating whether or not a speaker cable was used; if one was, perhaps he will enlighten us further.

As to his final nonsense regarding traffic rumble or the emanations from a computer in the same room, I have the fortune to live in a village in the Norfolk Broads which is by-passed to through traffic and has low density housing. My home is 35m from the nearest, and now quiet, road. My house is double glazed with low 'E' glass and well soundproofed, since before the bypass was built: traffic noise used to intrude.

My computers are located distant from my test workshop and neither were powered. Nor were any fluorescent lights, vacuum cleaners, washing machines, microwave or central heating systems, while measuring distortion. I tried deliberately to replicate normal household listening conditions, while completing a measurement of all cables in one session.

As to other computers locally, the nearest known to be in use was at the local Dentist's surgery some 300m distant. The loudest extraneous noise for these distortion tests were birds also a squirrel eating hazelnuts in a nearby tree.

As an emc filter engineer of many years, I asked for a loudspeaker model, not help with the non-harmonically related noise. The most obvious source was wideband radiation from the domestic mains. This was picked up by the 0.54mH solenoid inductor in the crossover, which acted as an aerial, and by the speaker cables used – even though care was taken to avoid proximity to the ring mains layout.

As to Douglas' re-iterations of his disbelief in impedance having any bearing, he accepts cable resistance and inductance are important and while he doesn't mention capacitance as such, fails to realise that with low resistance, low inductance and some line to line capacitance, this exactly describes a low impedance cable. I, for one, find it much simpler to use one number – impedance – which is all embracing, rather than grapple with three or four interdependent parameters.

Having rejected impedance, Douglas offers no alternative explanation for the behaviour from 1kHz to 10kHz of the lower impedance cables in both Figs 1 and 2 of the January article, page 55. Certainly inductance and resistance alone cannot explain these curves and 1, 2, 3, 10 and 20kHz were among the actual spot measurement frequencies for all tests.

As to his belief impedance or cable lengths less than 1/6 wavelength is meaningless, I could suggest some good text books for his education. For my part, I use the Reference Data for Radio Engineers (pub Sams) – commonly called the ITT (STC) Handbook, both being the same. STC for whom I once worked at a senior level, was a totally owned UK subsidiary or ITT. In this handbook the transmission line education I used for articles 1 and 2 is defined as a 'fundamental equation'.

His final critique – the excessive cost of co-axial cable – is equally false. Maplin sells RG58 at 35p/m, CT/FT100 at 57p, URM67 at £1.35, and 79 strand at 54p. The nearest lawnmower cable with low resistance 2.5mm^2 section, costs 52p/m in 100m reels from Farnell. Jenving's cable is £3.78/m.

As to the test amplifier, I built two samples using boards purchased from *EW*, with components exactly according to the parts list. All solder joints used Multicore low-melting-point silver alloy 62.36.2 solder. This is a true eutectic with coincident liquidus and solidus temperatures of 179°C. Solder joints were made using an Antex TCS 50W temperature controlled iron set to 290°C, fitted with a 3mm tip. To suggest the amplifier was built other than correctly is at best insulting.

When first built, each sample was tested powered from my Advance 30-0 30V current-limited bench supply with each rail decoupled using 10,000μF 63V capacitors adjacent to the pcb. With bias set, the amp was left to stabilise, driven by the 60mV, 1kHz square wave from my oscilloscope, supplied via 0.5m of RG58 cable. Almost immediately, the bench supply shut down. I replaced the fuses with 1A F types and repowered but with no input drive. Following some evaluation using a 10kΩ preset, inserted between RG58 inner and amp input, while observing the output on my oscilloscope, I determined that the first sample oscillated at rf, blowing the 1A rail ruses with less than 2.1kΩ. Sample 2 did the same with less than 2.4kΩ, hence my use of a series 4.7kΩ.

Since I anticipated Douglas may well react to this, at the completion of all cable tests I carefully repeated the above, using both bipolar (Panasonic) and various makes of polarised 10μF electrolytic capacitors for C_1. All combinations gave the same results.

I did not elaborate on this earlier since it was irrelevant to my results. However, I find it hard to believe that no one else found this problem. If Douglas wishes, I will gladly ship an amplifier for clarification to an independent judge. Perhaps Ben Duncan would like to volunteer for this?

Invention – or implementation?

With reference to the article, "Stepping Out" in your October edition, about a decade ago, much like Ian Hegglun, I followed the same steps of reasoning and arrived at the full-wave multiplier. It is not 'improved' over a conventional series multiplier because it requires double the terminal voltage from the transformer over its single-ended counterpart. It also necessitates the use of diodes of double the voltage rating (and half the current).

That said, it does lend itself to higher power systems. I have used it since 1989 at powers up to 35kW and voltages of 225kV for ion-beam generation, electron-beam welding, capacitor charging and x-ray systems.

As a plea to engineers in general; do not confuse implementation with invention or worse still, assume that because you may not have seen your idea before that it is a novel one. In the same way that Ian Hickman's parallel multiplier is 'old hat', so is this latest offering. ■

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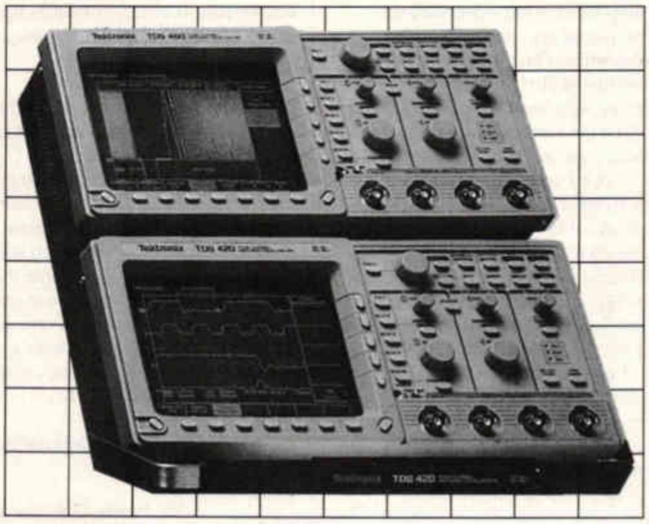

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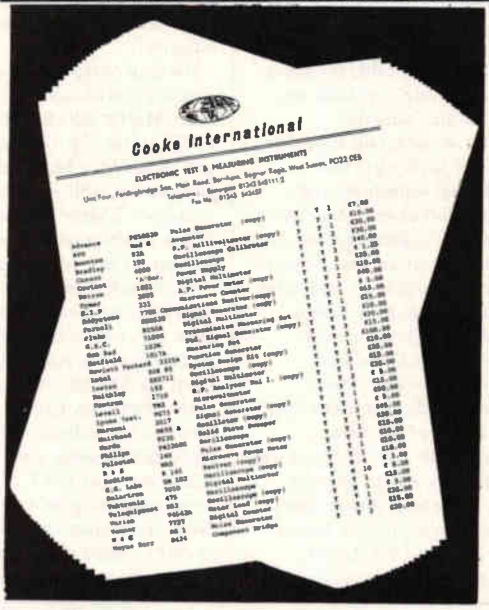

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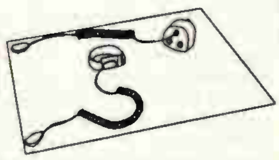

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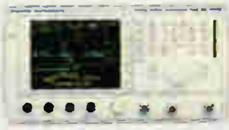

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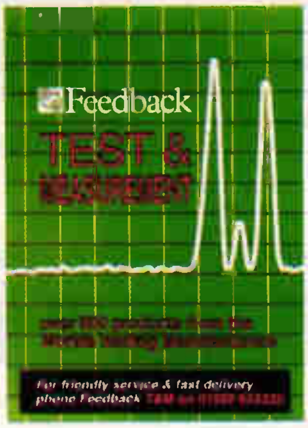

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