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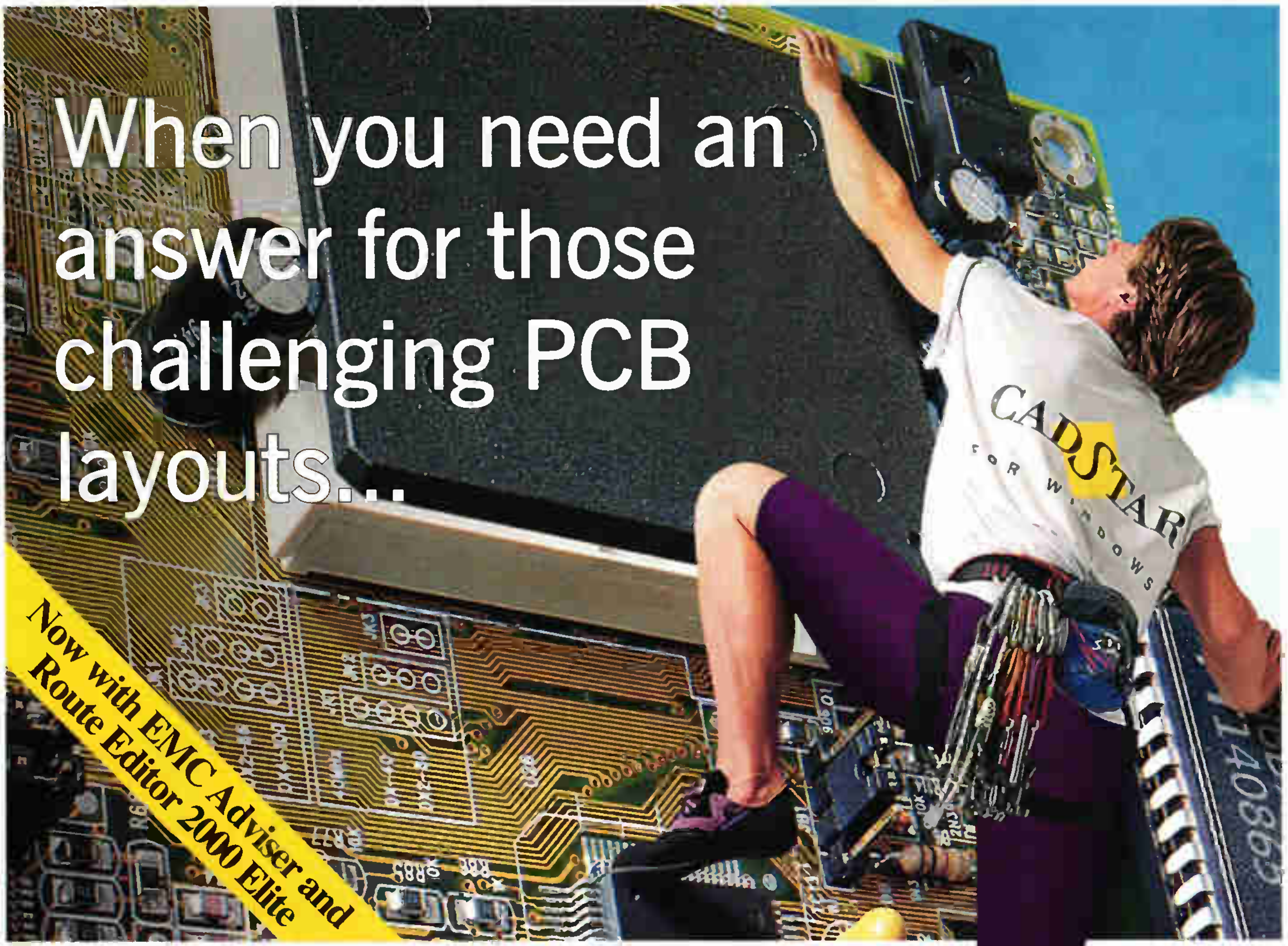

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Cover – Jamel Akib

Regulars

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15% reader discount on temperature-controlled soldering stations is extended – see page 905.

Using the CD on this month's cover to simulate a novel analogue solution could win you one of six TRAC development systems worth £600 each.
Find out more about this design competition, and how you can use your free demonstration CD on page 935.*

Smart, temperature-sensitive fire detectors that can sense the signature of an uncontrolled fire are being researched at Purdue. See Research Notes on page 891.

DECEMBER ISSUE ON SALE 6 NOV

Design competition

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The CD on this month's cover* contains a fully-working demonstration of Fast Analog Solutions' design and simulation tools for TRAC – a unique programmable analogue chip. See page 935.

*only available on UK issues.

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Technology – shackles or wings?

Electronics and nomads don't seem to have a lot in common, but the talk is of a nomadic age of electronics as MIPS/watt and MIPS/$ ratios in microprocessors continue to soar.

In the early eighties, typical microprocessor clock rates were 4-10MHz. Today typical clock rates are 100-200MHz and superscalar techniques allow more than one instruction to be processed per clock cycle.

That means we now have microprocessors capable of 200/400MIPS – a performance improving by a factor of about 30 every decade. Based on this level of performance increase, we should have GIPS processors by 2000.

At 1990 current price/performance levels of $1 per MIPS, GIPS micros would be expensive but the $/MIPS ratio has been improving by a factor of 100 every decade and by 2000 we'll have a ten cent MIPS.

The MIPS/watt ratio has been improving by a factor of ten every decade – largely as a result in decline of operating voltages – from 5V, to 3.3V. 2.5V, 1.8V and even to 0.9V. Every time you cut the operating voltage by a third, you halve the power consumption leading to dramatic reductions in the size and weight of portable products.

So what kind of products does it make possible? The way things are going it looks as if a combined workstation, camcorder, digital-camera, wireless phone, tv, Internet browser, scanner, fax, speech recognition system and speech digitiser, organiser and games machine can be squeezed into a notebook computer sized portable form factor early in the 21st century.

All those functions will require about 250MIPS of processing power which the ten cent MIPS will deliver for $25. Control and memory chips will take the total bill of costs for chips to $100, suggesting that the final cost of the product will be about $500. That's because the chip cost of advanced electronic products is already accounting for around 15 per cent of final product cost and the percentage is increasing.

So what's all this got to do with nomads? Well, possessed of such a tool you could be as connected to all your information and entertainment sources in the middle of the Sahara as you are at home or in the office.

You wouldn't actually need a fixed office, which means you wouldn't need a fixed abode. Does anyone want to live without a fixed abode? Well, no one really knows because mainstream human lifestyle has revolved around having fixed abodes for the past 10,000 years – ever since humans settled down to farm.

But, for millions of years before that man was nomadic – following herds of animals, or herding their flocks of animals from pasture to pasture. The fact that many people still practise nomadic lifestyle shows that nomadism is an urge that still beats strong in humans.

Bedouin still roam the Sahara and the Middle East, in Africa, the Kalahari Bushmen, Masai and Fulani are modern nomads, several million Gypsies still wander over Europe, until recently aborigines in Australia and Red Indians in America practised nomadic lifestyles.

And what of nomadic sophisticates? The executive who lives out of a suitcase, the retired who go on six month vacations, the students who back-pack around the globe. All these groups suggest that there is an intrinsic desire to travel that suggests the primeval nomadic urge survives in the human psyche.

> ...possessed of such a tool you could be as connected to all your information and entertainment sources in the middle of the Sahara as you are at home or in the office.

"Travel is the world's largest industry in terms of gross output, approaching $3.4 trillion. It employs 204 million people worldwide, or one in every nine workers – 10.6 per cent of the global workforce," wrote John Naisbitt in his 1994 book Global Paradox.

Last week a company announced it was building a ship in which the cabins – or rather apartments for they had many rooms – would be sold for several million pounds each for people who wanted to cruise permanently.

The airline business grows at around five per cent a year, at any one time there are 300,000 people in flight over the USA – a permanent flying city, China plans to build 25 new airports by 2002. Tourism grows relentlessly.

The question is, "are people natural nomads or are they natural settlers who like a spot of travelling?" If it is the former, then they could just be waiting for the opportunity provided by ever-cheaper travel and by ever-improving communications and computing technologies to cut loose from the bonds of home and office and adopt nomadism as a mainstream lifestyle. ■

The writer, David Manners, is co-author, with Dr Tsugio Makimoto executive managing director of Hitachi Semiconductors, of Digital Nomad to be published in the autumn by John Wiley and Sons Ltd.

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UP DATE

New UK chip fab on the horizon

Deputy Prime Minister John Prescott has given the go-ahead for Birmingham's first-ever wafer fab. Philips Semiconductors is now deciding whether it still wants the site.

When Philips showed interest early last year in building a billion dollar wafer fab on a green field site in Birmingham's green belt area, the West Midlands Development Agency applied for planning permission.

Local objections, led by the area's MP Sir Norman Fowler, produced a Public Enquiry. The final decision went all the way to the Secretary of State for the Environment and Deputy Prime Minister, John Prescott, before getting approval late last month.

Now it is up to Philips whether to go ahead with its plans or not. Making the decision at Philips will be Stuart McIntosh, chief operating officer. "When I got back from vacation last week I saw that the Public Enquiry into the site was published and there was renewed speculation about who would go onto the site. We never confirmed that we would use that site. It was one of several sites we looked at," explains McIntosh.

It is thought Philips is also considering sites in northern England like Durham and Washington where it already has factories. "We have to decide whether to go to an existing site or a greenfield site," said McIntosh.

However, Philips is also considering a joint venture. "When you think of the scale of fabs today, for those people who are not in memories, it makes a lot of sense to work with partners to share the risk of filling a fab," said McIntosh.

Likely partners are Taiwan Semiconductor Manufacturing Company (TSMC) in which Philips has a 30 per cent stake, IBM with whom Philips has a joint venture fab in Sindelfingen, Germany called Sub-Micron Technology, and SGS-Thomson Microelectronics with whom Philips shares a fab.

Future cars – higher rails, lower engines

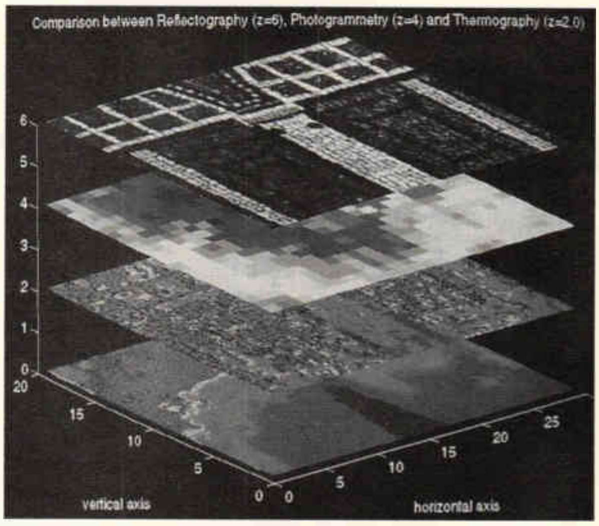
Revolutionary change in the electronic architecture of cars is on the way in coming years with the introduction of dual-voltage systems, multimode networks, and the abolition of the steering column, said Dr Karl-Thomas Neumann, director of Motorola's Advanced Systems Laboratories, at the company's Horizons technology conference in Florida recently.

Neumann said that future cars may use 42V systems as well as the traditional 12V systems. Forty two volt systems would power such functions as heated windscreens and catalytic converters, while the 12V system would power the usual functions such as lights.

These systems are getting considerably more complex with emerging top-end cars such as the Mercedes S-class and the next generation E-class. With each car system controlled by increasingly powerful microprocessors, they will come to resemble a network of PCs, said Neumann, who added: "They will have more than 30 networked systems."

This trend is already filtering down market to less expensive cars, said Neumann. "The VWPassat has a networked system and it's coming to all other cars across the spectrum," he added.

The abolition of the steering wheel column is another important innovation, he said: "You'll have a steering wheel but no steering column. Without the steering column you can get the the engine lower down in the car." This puts the weight lower in the car chassis giving better handling.



Tile tools... Cambridge Control has announced an updated set of image processing and analysis tools for Matlab 5 from The Mathworks. The image processing toolbox gives access to a wide range of algorithms and analysis tools. It is used by companies in aerospace, astronomy, remote sensing and research.

Cisco technology cuts congestion on Internet

Technology to help relieve Internet congestion has been unveiled by Cisco Systems.

Cisco showcased two technologies: a cacheing system which stores popular US-based Web pages on European servers, and a family of high-speed switching routers designed to speed up backbone networks provided by Internet services providers (ISPs).

The company's Cache Engine will provide European ISPs with the ability to store information held on parts of the Web located on US computers and servers. "The key problem is that US users and Europe-based users always want to access the same Web pages," said Olivier Cognet, Cisco's infrastructure marketing manager. Cisco's solution is to store significant chunks of the US-based Web pages most frequently accessed by European users.

"Around 80 per cent of a page- the main text and pictures- will be stored on the European network," said Cognet.

Cisco's Series 12,000 switching routers will provide ISPs with between five and 60Gbit/s switching capacity. "This is between five and twenty times more than classical routers are capable of," said Cognet.

DVD launch will miss Xmas stocking

DVD video players and software are not going to be launched in Europe in time for Christmas. This is despite the commitment shown to the product by the major consumer electronics makers and Hollywood studios.

At the consumer electronics show Internationale Funkausstellung in Berlin recently, 14 hardware companies and movie moguls confirmed DVD video launch for spring 1998.

"It has been a long hard run. We have fully agreed on a single standard and [will be] rolling all this rich content in the first quarter of 1998," said Doug Dunn, Philips' Video and Sound chairman and CEO. "This product will transform the consumer market which has been pretty flat at $75bn per year for years"

Despite the importance of DVD as a consumer product, only a limited number of titles will be released by the time of its planned launch next year. Some industry watchers believe the DVD Video market will not take off before 1999. However, other problems have been linked with the delayed launch, including the copying of DVD content onto analogue formats such as VHS tapes.

"There was a delay introduced by the MacroVision copyright protection mechanism used in DVDs which affected the picture quality. Now they are pursuing a better standard," said Ralf Jakobitz, technical support for Pioneer Germany.

At IFA the companies confirmed acceptance of MPEG-2 audio for Europe. But, Panasonic and Thomson have already been shipping players with Dolby Digital (AC-3) which are incompatible with MPEG-2. "You can play both discs on our equipment but the multi-channel capability will be affected" said Matsushita.

Red light for green outlaws

Some 1000 companies in the UK have missed the registration deadline for the latest packaging recycling regulations and are now breaking the law.

So claims the Environment Agency which estimates that 3500 companies throughout the UK are liable under the regulations. To comply, they must have registered during August either with the Environment Agency or with recognised schemes, such as from Valpak and waste disposal firms.

Of the liable companies, 500 have registered directly with the Environment Agency, and around 2000 with recycling schemes, leaving a shortfall of 1000. In the electronics sector, these are believed to include medium sized enterprises and importers."We do expect to get a number of late registrations," said an Environment Agency spokesperson.

In brief

Digital set-top sales could reach $3.2bn

The burgeoning global demand for digital satellite set-top boxes will cause sales to reach the $3.2bn-mark this year alone, claims In-Stat Multimedia Services, a US market research group.

In-Stat forecasts that the market for direct broadcast satellite (DBS) will rise to above $4bn in 1999 with the advent of digital satellite technology in geographical regions Europe, Japan, Asia Pacific, Latin America and the Middle East.

"As are result of these launches, the worldwide market for DBS set-top boxes will be the start of the digital age of convergence technologies," said Gerry Kaufhold, senior analyst at In-Stat.

Trafficmaster en route to first profits

Three deals worth up to £30m will drive Trafficmaster, the road traffic information specialist, into profits for the first time since its stock market flotation in 1994.

These are an extended deal with the car maker Vauxhall for its Oracle units, a licensing agreement with Mannesmann Autocom in Germany and a deal with Cellnet, the mobile phone operator, for the supply of traffic information over mobile phones.

Last week, Trafficmaster reported sales that more than doubled to £2.88m and cuts in pre-tax losses from £1.42m to £594,000 for the first six months ended June 30, 1997.

Videoconferencing hits 85pc

Eighty-five per cent of Europe's leading 500 multinationals have implemented videoconferencing technology, according to a report commissioned by Sony.

The report, released this week, was based on a survey of 53 of the 500 European companies. It found that the average company from the sample had 29 videoconferencing units – group systems or desktops – installed.

The companies using videoconferencing cited travel cost reduction and improved speed of decision-making as the major benefits of the technology.

Of those not using it, 58 per cent expect to being doing so within the next two years.

Wharfdale and Quad have new owners

Verity Group is to sell its loudspeaker and hi-fi design and manufacturing groups Wharfedale and Quad to Pointfield.

Verity entered into a conditional agreement with Pointfield and its parent company IAG under which it will sell 60 per cent of Wharfedale and Quad for £4.8m. However, Verity will have a 40 per cent shareholding in Pointfield, an interest which is expected to end within five years of completion.

The transaction is a direct result of the success of NXT flat panel speaker technology developed under the umbrella of Verity.

Polaroid shoots license fraudsters

Polaroid has unveiled what it says is the world's first driver's license facial recognition system. The system is designed to reduce fraudulent attempts to reissue a driver's license by automatically checking an applicant's identity.

US state West Virginia says it will be the first to use the Polaroid system in conjunction with an automated fingerprint recognition one.

The system works by taking a digital picture of an applicant and matching it with future renewals. However, Polaroid did not say how the system will deal with changes in facial features due to ageing, or use of eye glasses or facial hair.

Racal contracts

Racal has picked up MoD contracts worth around £100m. Three contracts involve a satellite-based battlefield radio terminal, electronic warfare system for Type 23 frigates and a transportable command and control centre for the RAF.

Sega's 3D deal powers up law suit

Smarting after losing a potentially lucrative graphic chipset contract for Sega's forthcoming 64-bit video games console, 3Dfx Interactive has filed a lawsuit against Sega and NEC.

According to analysts' reports, UK graphics firm VideoLogic and NEC - which manufactures and sells VideoLogic's PowerVR 3D graphics chip – have won the Sega contract although neither Sega nor NEC/VideoLogic have publicly confirmed the arrangement. A spokesperson for 3Dfx, however, said: "Sega told us that NEC had the contract."

In a \$155m damages lawsuit, 3Dfx is accusing NEC of inducing Sega to break the contract. The lawsuit alleges that Sega, an investor in 3Dfx, entered into the contract knowing that it would break it, in order to gain access to its proprietary technology.

"Sega terminated our contract without justification. We lived up to all of our commitments, yet they terminated the agreement. It's a clear breach of contract," said Greg Ballard, president and CEO of 3Dfx.

"We have many relationships with companies and until something is officially announced, we have nothing to say," was the NEC's spokesperson's response. He added that NEC had not yet been served with the lawsuit and had no response until it could study it, but he said that NEC was surprised to be named.

Sega is an investor in 3Dfx, owning slightly under five per cent of the company. Sega's spokesman said it would defend itself fiercely against the suit.

Orientation by mobile phone

The first off-the-shelf mobile phones that transmit user location details will be available within two years, according to BT.

BT Labs is working on MOSA (Mobile Social Alarm), a device that uses Global Positioning System (GPS) satellite technology. The company expects the phone to be bought by people who are concerned about becoming lost, or have health problems, like diabetics.

Steve Furner, engineering psychologist at BT Labs' Human Factors Unit, explained that the idea for MOSA came about last year when the research facility held a discussion between BT technologists and representatives from the public. "They were looking at ways in which telecommunications technology could be used to reduce vulnerability in the community," he said.

According to Furner, MOSA is being developed because, unlike with fixed telephones, mobile users are more likely to "have to find out some way of describing their location, which can be difficult".

One issue to be resolved is how receivers of calls make use of position signals being sent to them. A possible solution being considered is to set up call centres which hold the MOSA owner's details such as age, health problems, and contact details of next of kin.

UK BDB licences in the balance

The European Competition Commission is reported to be addressing the issue of whether to withdraw British Digital Broadcasting's (BDB) licences to supply digital terrestrial television services in the UK. The EC Commission is believed to be concerned about the extent of BSkyB's involvement in its supply of programming.

"We know these issues. They have been considered before we received the licences," said one BDB spokesman, responding to the reports.

In brief

K6 yields down

Advanced Micro Devices says its has had problems getting good production yields on its K6 microprocessor, resulting in far fewer chips than it originally forecast. The problems will result in lower revenues for its current financial quarter.

IBM shuns NetPC

IBM has given up plans to make NetPCs – the halfway house between PCs and network computers, NCs – even though it produced a NetPC prototype three months ago. NetPCs were an attempt by Intel and Microsoft to counter the NC concept of 'bare-bones' personal computers.

PowerPC embedding plan

Motorola's PowerPC microprocessor division is expected to soon announce a new strategy for the PowerPC, with more attention being given to embedded microprocessor markets. The shift is a result of the PowerPC systems' market continuing to shrink following Apple's problems in expanding its markets.

Encryption stifles exports

California politicians have asked the US government to lift export barriers on powerful encryption technologies saying that California's high tech companies are losing billions of dollars in business because of the stringent rules.

Mentor DSP heads off

Control of Mentor Graphics' DSP Station signal processing design software is being transferred to its Dutch subsidiary, Frontier Design. Development of DSP Station will continue at Frontier. A spokesperson said Mentor Graphics will concentrate on other aspects of the EDA business. ■

RESEARCH NOTES

Jonathan Campbell

Space gamma detector has place on earth

More sensitive brain and heart examinations, and even improved environmental monitoring, and non-destructive evaluation, could be the dividends of a super-sensitive gamma-ray detector array developed by space researchers.

The array, invented by a team at Nasa's Goddard Space Flight Center, for gamma ray astronomy, is composed of multiple semiconductor detectors made from cadmium zinc telluride (CZT).

The reason for the exceptional accuracy of these detectors is the extremely fine strip electrodes attached to the front and back of each CZT chip. Each strip in the electrode is separated by only 100µm from its neighbour. When a gamma ray strikes the chip, it produces an electric charge in the strip electrode. Since so many strips are packed into such a small area, the scientists can determine very accurately where the gamma ray comes from."

The novelty of the new array is that it has many more detectors than ever packed together before. The area of the strip detector array is nearly 3660mm^2 with over 500000 pixels – 100 times more pixels than any other array currently available.

One of the non-space potential applications of the array is in medical imaging, where Goddard and the University of Arizona are collaborating to use a CZT pixel array, developed at the University of Arizona, in a brain imaging instrument.

Diagnostic applications will allow better characterisation of epilepsy and stroke. The array could also bring a breakthrough in an entirely new class of high resolution nuclear medicine scanners, for such clinical tasks as heart imaging to assess the amount of tissue damage subsequent to a heart attack, and for improved detection of breast cancer.

Unlike other gamma ray detectors, the arrays do not require cooling with cryogenic fluids, so they are portable. This allows them to be used wherever accurate detection of gamma rays is necessary. For example, they can be used to monitor radioactive waste storage sites or to verify nuclear treaties.

Gamma rays are also used to probe the interiors of structures and equipment, non-destructively looking for faults. An accurate detector is essential for constructing a high resolution image and for reliable fault identification.

Nasa says it even has a soft drinks company that plans to use gamma rays to determine if its bottles are filled to the correct level.

Extremely fine strip electrodes, indistinguishable here, attach to the front and back of each CZT chip. When a gamma ray strikes a CZT chip, it produces an electric charge in a strip electrode. (Picture Nasa Goddard Space Flight Center)

Quantum computing not ready to freeze out

First experimental demonstration of a transistorless approach to computing, called quantum-dot cellular automata (QCA), has been reported by University of Notre Dame researchers.

In their experiment, a single electron was used for the first time to control the position of another electron.

Conventional microelectronic technology has relied on shrinking transistors to produce increasingly-smaller, faster and more powerful computers. But, the laws of physics prevent conventional devices from working below a certain size, so that method is nearing its physical limits.

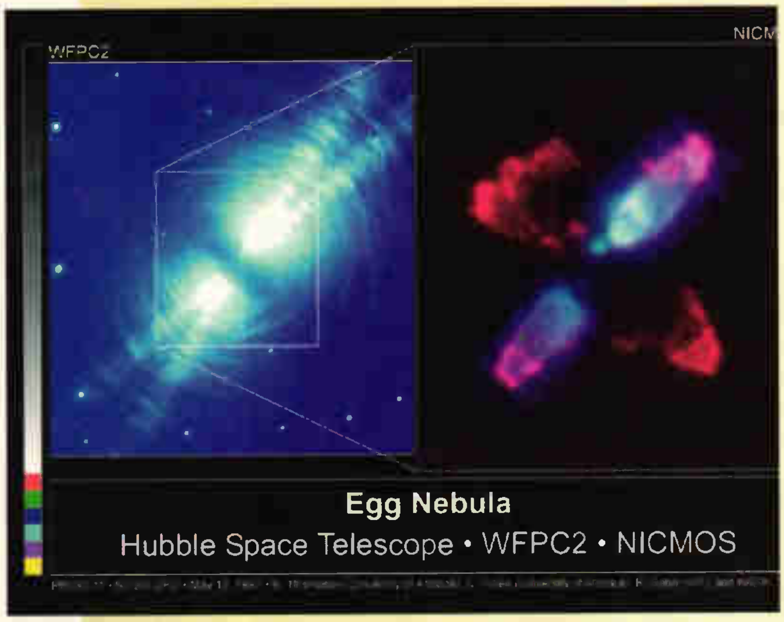
QCA leap-frogs that barrier by using the quantum dot, a tiny structure in which an electron can be confined. These quantum dots can be created and arranged into cells through microelectronic techniques, and in turn the cells can be lined up end to end to form binary wires or arrayed to form switches and various computer logic devices.

If successful, a 100mm-square QCA chip could contain as many as 40trillion devices, as opposed to the 6million devices in the most advanced conventional chip. And since it does not rely on flowing electrons to transmit a signal, no electric current is produced and heat problems are avoided.

In this first demonstration of a basic cell, the device studied was made of four metal dots, connected with tunnel junctions and capacitors. One immediate problem is that the device needs to be operated at a temperature less than 50mK, very near absolute zero.

Researchers believe it will be another 20 years before computers and electronic equipment will incorporate QCA.

Contact: Gregory Snider,a assistant professor of electrical engineering, University of Notre Dame, Tel: 001 219 631-4148 or more information can be found at http://www.nd.edu/~lent/QCAhome.html.



Dying sun reveals star birth: ***A long-held model for dying Sun-like stars is that they eject matter in a slowly expanding spherical wind. But observations by Joel Kastner, a research scientist at MIT's Center for Space Research, made colleagues at the Jet Propulsion Laboratory and Vanderbilt University, on the 3000 light-years distant Egg Nebula, are forcing a shift in this theory. Their work shows that stars also eject matter at high speeds (preferentially along their poles), and may even have multiple jet-like outflows from their surfaces.The picture shows the Egg Nebula as it appears in visible light with the Hubble space telescope's wide field and planetary camera 2 (left) and (right) as it appears in infrared light with Hubble's near infrared camera and multi-object spectrometer. Blue corresponds to starlight reflected by dust particles, and red corresponds to heat radiation emitted by hot molecular hydrogen. The distance between the tip of each jet is approximately 200 times the diameter of our solar system (out to Pluto's orbit). Studying the death of Sun-like stars is important for understanding how two of the elements crucial for human life – carbon and nitrogen, formed from hydrogen and helium inside stars – are expelled into the interstellar medium, where they eventually become the building blocks of new stars and planets..***

Contact: Joel Kastner, Massachusetts Institute of Technology, Cambridge, Massachusetts.

Fire signature will trigger alarm

Smart temperature-sensitive fire detectors that can sense the signature of an uncontrolled fire could improve safety in industrial and domestic applications, following building of a prototype system at Purdue University. The system does not have to be looking directly at the flame to see it, but can pick up reflections off the walls, giving it a very fast response time compared to a smoke detector.

The idea of using near infrared to detect fires is not new. But the advance made by combustion research specialist Jay Gore at Purdue, is to apply a unique discrimination algorithm to the process to eliminate false alarms.

Over the past two years, Gore and his team have examined the near-infrared radiation given off by several standard types of fires, covering a wide range of combustible materials.

"We have analysed flames based on the way the intensity of the light fluctuates as they burn," Gore explains.

"For example, we may see a peak in intensity every tenth of a second, and then the pattern is repeated. That type of fluctuation frequency is characteristic of an uncontrolled fire."

Once a flame has been detected, software analyses the fluctuations in its near-infrared intensity – its infrared signature – and determines whether to sound an alarm.

The new device uses fibre optics to detect radiation from a fire, so optical fibres could easily be run from a central detection unit to each room in a multi-story building, providing blanket coverage.

In the home, the team has managed to teach the detector not to respond to common household flames such as candles, gas stoves and cigarette lighters. It also does not respond to fluctuations from hot plates, solar radiation or fluorescent light, which are different than those from uncontrolled flame.

"We still have a problem with the alarm going off from a fireplace, which is an uncontrolled flame, but we are developing ways that may make the device blind to certain spots in the room," says Gore.

Another bug yet to be worked out is the difficulty in picking up smouldering fires – an area where smoke detectors are also slow to respond because of the time it takes for the smoke to reach the detector. At present, the intensities obtained for smouldering fires are too low for the detector to discriminate from background noise. But the team is hopeful that more sensitive infrared technologies are now available that might boost its capability.

"I believe that in the next few years, the home PC will control the home security and safety systems and that a fibre optic communication network will carry such signals in addition to voice and data. This detector could be easily integrated into such a communications network," says Gore.

Contact: Jay Gore, Purdue University, West Lafayette, Indiana.

Jay Gore, Purdue professor of mechanical engineering (foreground, below) and his doctoral student Xian-Chua Zhou use laser technique called particle imaging velocimetry to investigate how air flows into a fire (right). Understanding the characteristics of uncontrolled fires is fundamental to the new fire alarm. (Pictures Purdue University, Vincent Walter.)

Environment-friendly fridge is sound advance

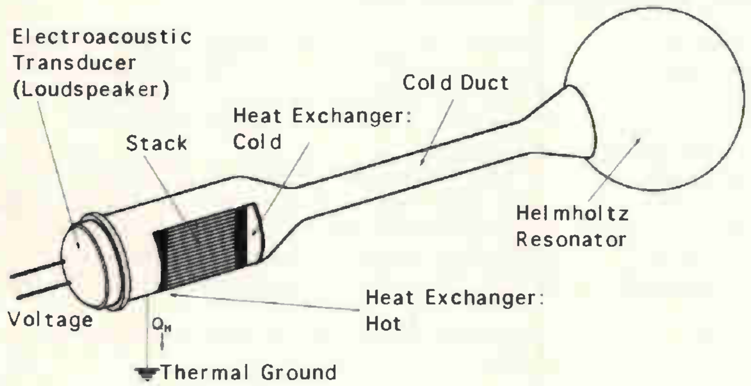
A vibrating acoustic driver (right) causes gas atoms inside the device to oscillate, setting up pressure fluctuations, or sound waves, which in turn cause temperature fluctuations. The result, one end of the device becomes cooler than the other.

Researchers have built a prototype of a fridge that uses sound rather than traditional coolants to reduce temperature. The concept, called thermoacoustic refrigeration, could be used as the basis for refrigerators and cooling systems with low manufacturing and maintenance costs, new kinds of air-conditioning systems, and portable systems for liquefying natural gas.

The theory for using sound waves to cool was actually developed back in the 1960s and a few researchers in Europe and the United States began working on the idea in the 1980s. But since that time, scientists have gained a better understanding of these systems and useful devices are now being built.

The prototype has been built by Luc Mongeau, assistant professor of mechanical engineering at Purdue University, and other groups at Johns Hopkins and Penn State universities, the University of Mississippi, the University of Texas, Los Alamos National Laboratory, Ford Motor Co., and the US Naval Postgraduate School are all working in related areas.

At the heart of the latest design is a hollow metal tube, 150mm in diameter and 600mm long, with a large round cavity on one end, called a Helmholtz resonator. Attached at the opposite end of the tube is an acoustic driver – a vibrating diaphragm similar to a loudspeaker, but sturdier and more powerful. As the diaphragm vibrates, gas atoms pressurised to 20 atmospheres inside the enclosure oscillate back and forth, setting up pressure fluctuations.



Brian Minner, a doctoral student, and Luc Mongeau, perform an impact test on the acoustic driver, a component of their prototype thermoacoustic cooling system, below. (Picture Purdue University, David Umberger).

Fluctuating pressures inside the cavity are accompanied by a fluctuation in temperature, as a gas compressed quickly becomes warmer, while quickly decompressed it becomes cooler.

So the gas particles within the device are becoming alternately hot and cold, dynamically, at a typical frequency of 200 hertz.

Gas atoms transfer their heat to and from a piece of porous material called a stack, which is located near the acoustic driver. The end result is that heat is pumped toward the driver, cooling the side of the stack farthest away from the driver.

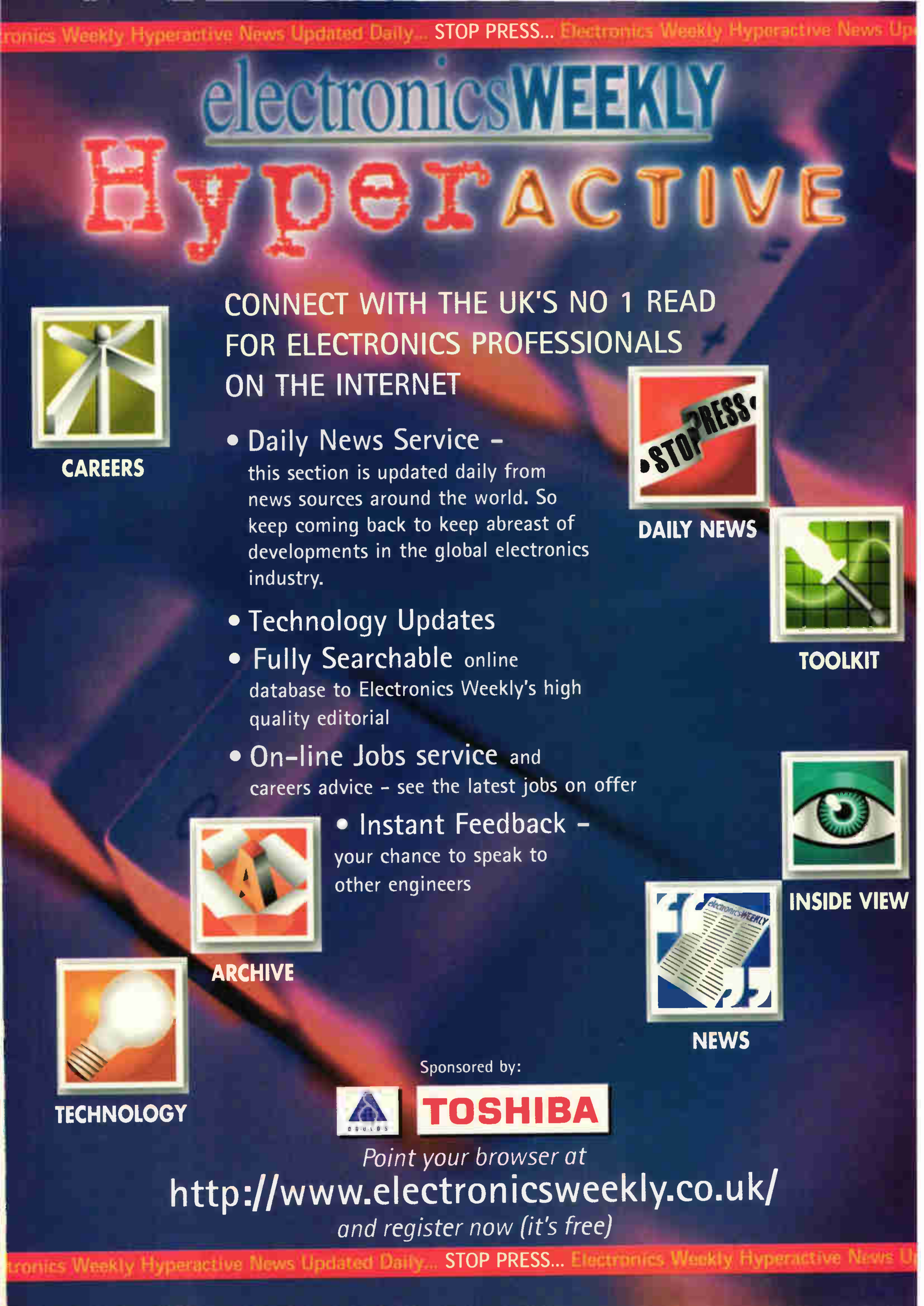
The pressure fluctuations propagate as very loud sound waves – around 180 decibels – but if designed properly a thermoacoustic refrigerator shouldn't be any louder than a conventional appliance, and may be even quieter.

As in a conventional refrigerator, an appliance cooled by a thermoacoustic device will still require coolant to circulate through pipes. But this could be water or a combination of water and glycol. The main advantage to this type of cooling system is that there are no phase-change refrigerants involved, such as chlorofluorocarbons, which are harmful to the environment,.

With thermoacoustic refrigeration, all the elements are environmentally benign, including the coolants and the gas inside the device, which is a mixture of inert gases like helium, argon or xenon that present no hazard to the atmosphere.

Another benefit is that the absence of mechanical compressors and lubricants used in conventional refrigerators means less maintenance. Also, when you don't need a lot of cooling, you just need to turn the volume down.

The researchers admit that a home appliance based on thermoacoustic refrigeration is still years away, but they are working on several aspects of the technology. For example, the technique could be used for cooling computer chips and other electronic equipment, and one researcher has a patent on a thermoacoustic device to cool seismic instruments in the Earth's crust. ■



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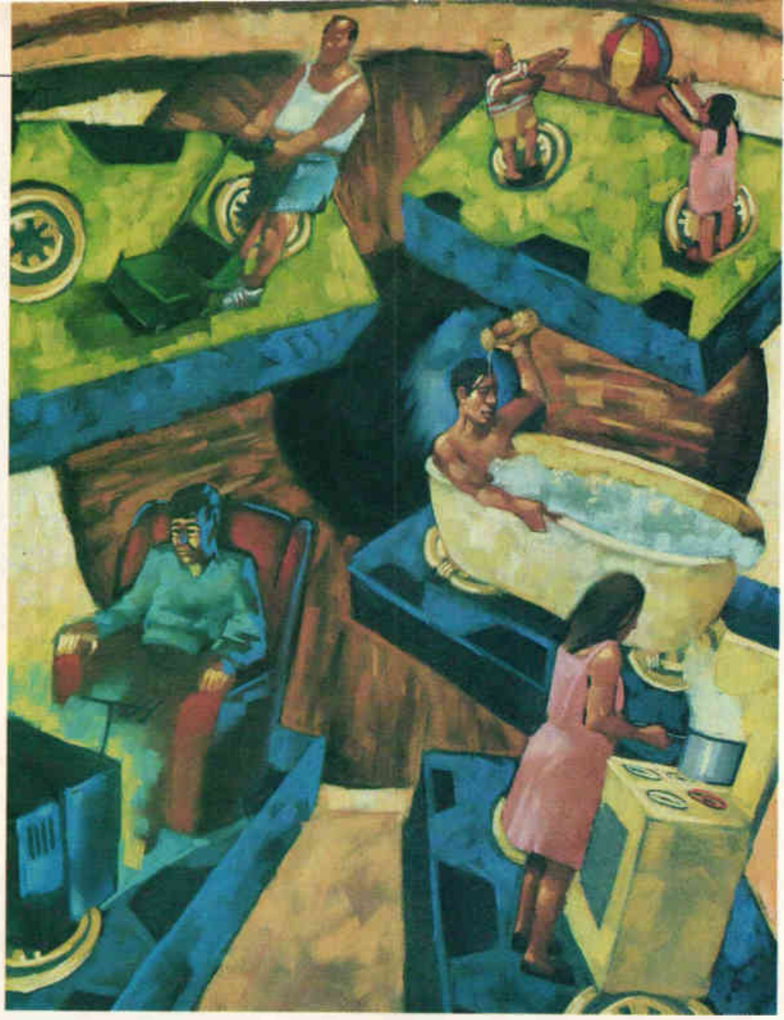


Ian Hickman's video surveillance security system can handle up to six camera/microphone combinations. Designed with cost in mind, it doesn't need a special video recorder – nor does it require any modifications to a standard domestic vcr.

Video surveillance system

Unfortunately, security is a pressing consideration both for business and domestic premises nowadays. I have recently been experimenting with a video surveillance security system, built around economical ccd cctv cameras. These are controlled by a small but powerful microcontroller, easily programmed in a version of the popular and widely used Basic programming language.

The system is up and running, and only awaits the installation of the cameras and passive infra red, or PIR, motion detectors in their final positions. The development process was both challenging and absorbing, with – as is usually the case - the hardware development proving less problematic than the software.

Design goals

The first task was to decide exactly what the system should consist of, and just what it should do. The wish list went something like this,

- The system should deter burglars.
- It should be reliable and affordable.
- It should record any activity in the region of any one or more of its PIR detectors, via the associated cctv camera(s), cycling around each in turn if more than one is active.
- To conserve tape it should record only when a PIR detector senses motion.
- It should possess a degree of 'survivability' to an attack aimed at disabling it. So it should sense if the lead to any camera is cut, and record sequentially from each of the remaining cameras in turn.
- It should be possible to start with an installation of just two cameras, but be readily expandable up to six cameras without hardware modifications.

The ability to deter burglars demanded that if a prowler be detected when the system is armed and active, a good loud alarm bell should sound – with the object of sending the prowler on its way. Much better to avoid a break-in in the first place, rather than rely on the recorded evidence possibly leading to the apprehension of the burglar.

But a short delay was deemed useful, on two counts. Firstly, it would reduce the likelihood of annoying false alarms due to a passing dog, fox, bird or what have you. Secondly, it would provide a record of who – or what – triggered the system.

Suppliers

Black and white ccd cctv cameras used in this design come complete with lens and microphone and are type *A-921-S*, from Anchor Supplies Ltd. Tel. 0115 986 4902 or 4041, The Cattle Market Depot, Nottingham NG2 3GY.

The *Basic Stamp2* computer with 16 i/o lines, up to 600 program lines, comms to 19200 Baud, is available from Milford Instruments Tel. 01977 683665, Milford House, 120 High Street, South Milford, Leeds LS25 5AQ.

The *Stamp2* microcontroller

The *Stamp2* consists of a *PIC16C57* microprocessor U_1, a 2Kbyte eeprom, U_2 and sundry peripherals, **Fig. A**. Eeprom U_1 contains 32 bytes of ram, six of which are reserved for i/o control, setting which pins are inputs, which are outputs, and what their states are. The remaining 26 bytes of ram are available to the user.

The eeprom's permanent memory is programmed with a Pbasic interpreter, and so is not available to the user. The user's program is stored in U_2, starting at byte 2047 and working back towards the beginning. Since the available ram space is limited, further data, variables, etc., may be stored in U_2, starting at byte 0 and working upwards. This is all looked after by the operating system, as is the allocation of variables and data to appropriate storage locations. In the event of the space required by the user's program plus any data being more than available in U_2, the *Stamp2* programming software, running on the host pc, will issue a warning message.

The user's program is written in Pbasic on the host pc, using the software supplied which is on a single 3.5in floppy. The software is very economical on space, as it runs under MS-DOS, not Windows, and it does not need to be loaded on to your hard disk. On receipt of the software, make a working copy, using dos command 'diskcopy,' Xtree or your favourite file manipulator, and bank the original as your back-up.

To write your application, from the dos C:> prompt type,

A: <ENTER>
CD STAMP2 <ENTER>
STAMP2 <ENTER>

This takes you into the editor ready to start writing code. Typing (ALT)L brings up a list of programs on the disk – any you have written plus "EXAMPLE.BS2" supplied. From C:> you can just type A:STAMP2\STAMP2 which is marginally quicker, but a subsequent (ALT)L (LOAD a file) does not bring up a list of the programs on the disk – it simply brings up a box asking you to name a file.

While writing your program, don't forget to regularly save the latest version. Having written your program, (ALT)R stores it in U_2 program memory – assuming that it compiles satisfactorily, otherwise an error message results. But I prefer first to run (ALT)M, which brings up the first of three map pages showing memory usage – namely the ram allocation.

The second page shows the whole of U_2's eeprom usage and the third shows detailed eeprom usage on a byte by byte basis. The up and down cursor keys scroll through this, while successive space bars cycle round the three pages.

Pressing (ALT)R stores your compiled program in eeprom as a series of 'tokens', as in other flavours of Basic. At compile time, any declared constants are substituted for their labels and at run time, U_1 wades through the tokens and interprets them into the appropriate actions.

Once programmed, the *Stamp* can be removed from the carrier board which is connected by a cable to the pc's serial port, and inserted in the target system. I found removing it from the vice-like grip of the 24-pin 0.6in DIP socket on the carrier board a little difficult. I found an Augat 24-pin zero-insertion-force socket which was no wider that the original and fitted this to the carrier board instead.

Readers who are familiar with one of the fuller versions of Basic will have to get used to some of the limitations of Pbasic. For instance, in a FOR... THEN construct, the only thing that can follow THEN is an address. FOR... THEN is in fact here just a conditional branch instruction. And another example is, to test the condition of an input pin, IF IN3=1 THEN... tests pin P_3 – assuming it is set as an input – and is ok. But to follow IN with a variable rather than a constant, a more complicated syntax is required – IF INSBIT0(M)= etc tests pin 0+M. Note that M cannot be computed at run time; IF INSBIT0(N+6)= etc is NOGO; add a preceding line of code M=N+6.

Once these limitations are taken on board, the *Stamp2* microcontroller, with its development system is delightfully easy to use, and surprisingly powerful into the bargain. It just requires a little ingenuity to circumvent some of the limitations of such a small, neat system.

An example is the implementation of an elapsed time indicator running at a steady rate regardless of variations in program looping, incorporated in Figs 3 and 4.

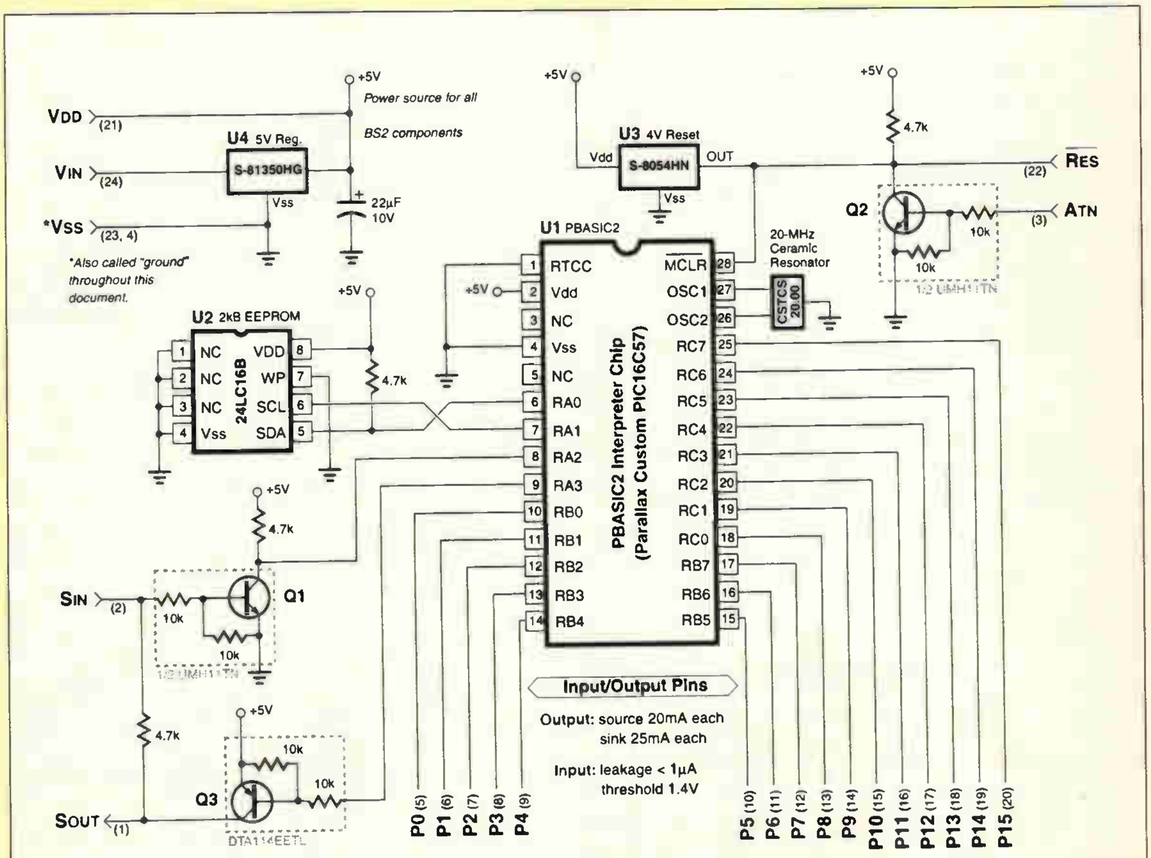


Fig. A. Few peripheral circuits are needed to produce a microcontroller with non-voltatile memory and programmable in Basic.

Hardware requirements

With the requirements more or less mapped out, I started on the hardware, as this would define just what the system could and could not do. With the software, you can do just anything you want – subject to the required quantity of code fitting into the available program storage space.

The hardware consisted of the two cctv cameras, two PIR detectors, and a board full of circuitry which was constructed for speed and convenience on a commercial DIP breadboarding pcb.

For the prototyping, I added my domestic video recorder to the system. Of course it needed to be under the control of the microcontroller which formed the heart of the system. To avoid having to modify the vcr's remote control unit, I bought a separate unit, with a view to modifying it as necessary. The mature system will use a dedicated vcr in conjunction with the new remote control.

The remote controller I chose was a '4-in-1 light-up remote' from Tandy – a universal type which will work with any of about 400 models of vcr from 170 or so different manufacturers. It is simply programmed with the appropriate three digit code for the particular vcr in use.

This remote controller also copes with an even wider range of televisions, and takes satellite receivers and cable converters in its stride for good measure.

Circuit requirements

Figure 1 shows the suite of stabilised power supplies, among other things. The cctv camera modules draw 100mA each at 12V, while the rest of the circuitry draws about 60mA from the +5V and –5V supplies.

On the face of it, the 30VA transformer shown should power the whole complement of a six camera system. But a standard 6V+6V transformer would not then provide sufficient raw supply voltages for the stabilisers. On the other hand, a standard 9V+9V transformer would be overkill, but bespoke transformers apart, there are few other options. If you just happen to have a 7.5V+7.5V transformer, that would be ideal.

The circuit is arranged to use a bridge rectifier for the main load, the raw supply for the +12V stabiliser. The full-wave rectified raw supply for the +5V stabiliser is via D_7. The circuit doesn't really manage to provide full-wave rectification with just a single diode; D_7 works on each half cycle in conjunction with one or other of the diodes in the bridge rectifier BR_1.

The raw supply for the –5V rectifier is provided by a half-wave voltage doubler circuit working from one of the two 6V windings. For this reason, much larger electrolytics are needed at C_5 and C_{10} – a reasonable price to pay for the convenience of getting all the supplies from a single stock transformer.

During testing, note that with this kind of interlinked raw supplies, the load current on the +12V supply must exceed that on either of the others.

Each of the cameras is provided with its own +12V supply via a diode and resistor, e.g. D_{9A}, R_{17A}. An associated comparator, run from the +12V supply, monitors the drop across the diode. The bridge of four resistors is necessary, since the comparator's input common mode range does not include the positive rail.

If the camera is drawing its expected 100mA, the comparator produces a logic 1 at its open collector output, its load resistor R_{24A} being returned to +5V.

Op-amp IC_{12} monitors up to four cameras. A second such op-amp can be fitted if expansion of the system to the full complement of six channels is envisaged at some stage in the future.

Sections of IC_{12} monitoring channels not yet fitted with a camera can be forced to return a logic 1 'camera ok' signal, by closing S_{3A}, S_{3B}, etc., as appropriate.

While a cut lead would obviously put a

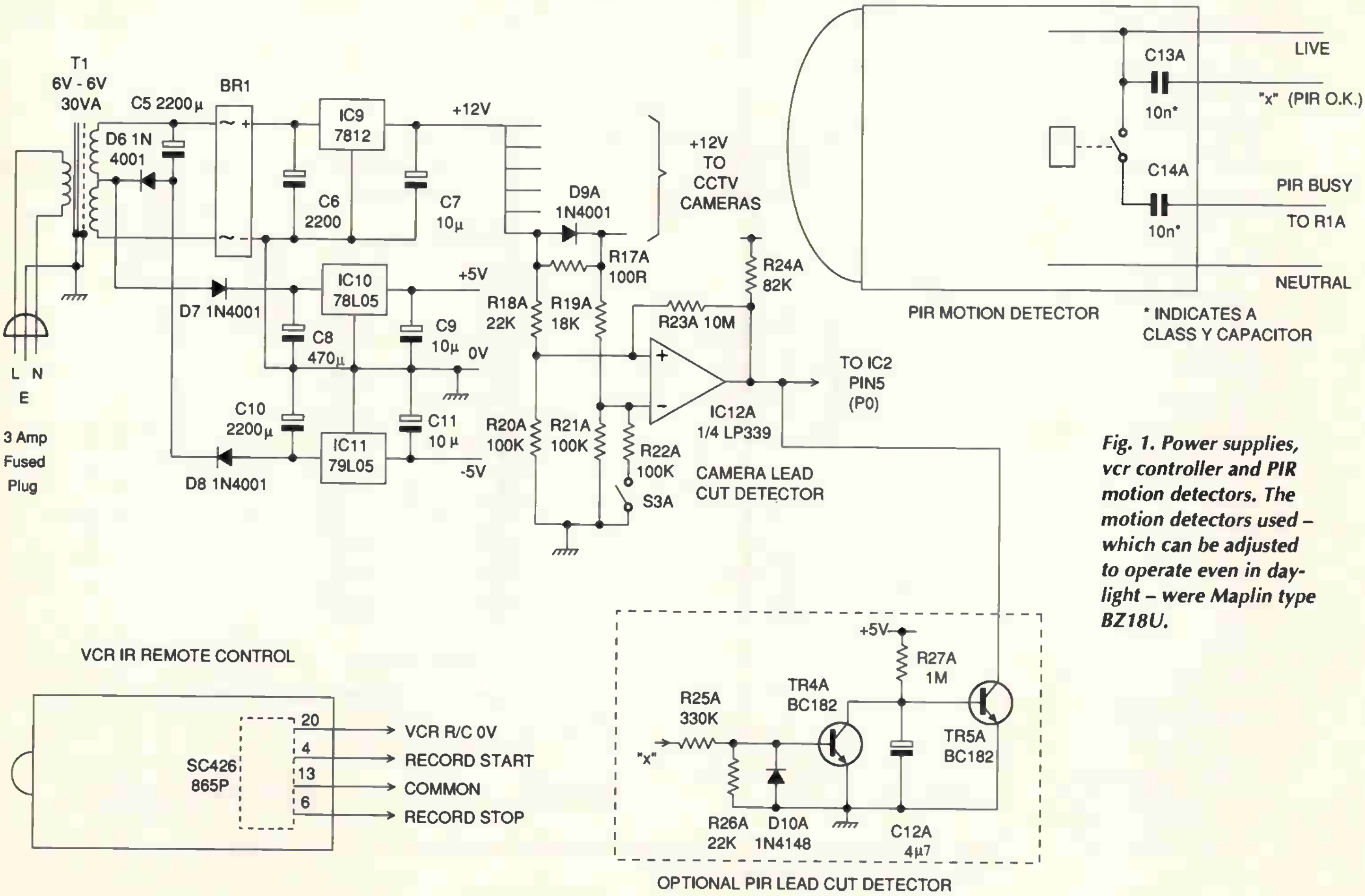
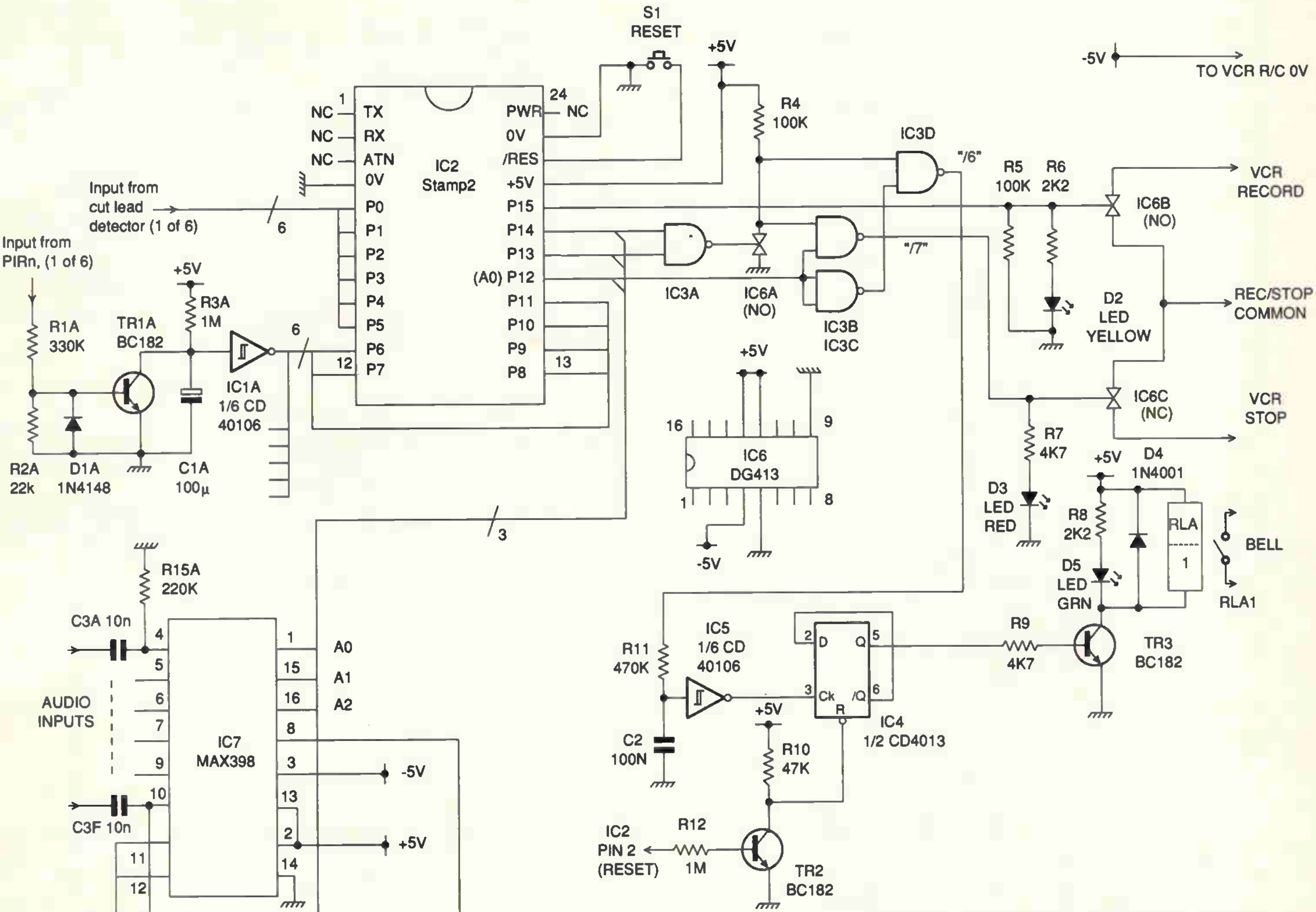


Fig. 1. Power supplies, vcr controller and PIR motion detectors. The motion detectors used – which can be adjusted to operate even in daylight – were Maplin type BZ18U.



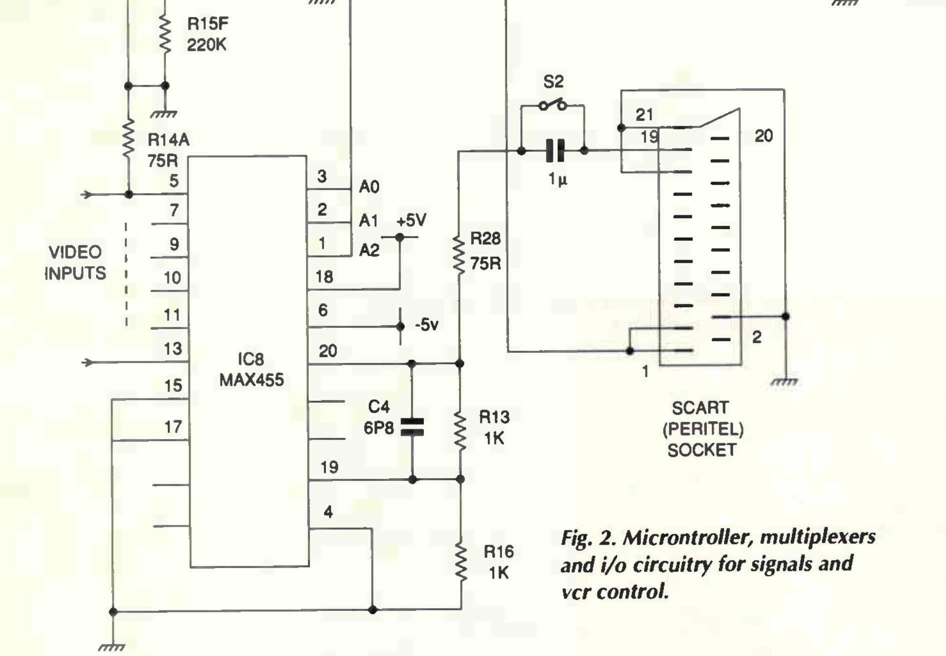


Fig. 2. Microntroller, multiplexers and i/o circuitry for signals and vcr control.

camera out of commission, a cut lead to the PIR detector also represents an attack on the system. So Fig. 1 shows an optional PIR monitor circuit. In the event of the PIR detector's lead being cut, Tr_{4A} turns off while Tr_{5A} turns on. It this signals a 'camera' fault via the wired-or connection at the output of IC_{12}.

Note in this connection that C_{13A} and C_{14A} must be class Y capacitors. These are special capacitors designed for interference suppression circuits, where they are connected between mains live and earth.

Clearly, in the event of such a capacitor failing short circuit *and* the earth lead of an equipment becoming disconnected – admittedly a double fault – the case of the equipment could become live. So class Y capacitors must be used. These are special self-healing high reliability types. A residual-current circuit breaker is always a useful addition.

If the earth lead becomes disconnected, but the capacitor has not failed, then current can flow through the body of anyone touching the equipment, of magnitude limited in the worst case by the reactance of the capacitor. For this reason, class Y capacitors fitted between live and earth are usually limited to 10nF. The reactance then limits the worst case shock one could receive to about 750μA. This is enough to feel but it should be too little to cause harm. Nevertheless, it would very unwise to experiment in this area.

If you don't like this scheme, you could always use an opto-isolator instead. In fact, judging by the healthy sounding click from the PIR when it triggers, there is a conventional relay inside. So with a little dismantling, it is possible to free the contact tied to the live side and bring it out separately.

Figure 1 also shows the modified video-recorder remote control handset, the heart of which is a Motorola 40-pin IC, designated *SC426865P*. A vcr remote control, being in effect permanently on, uses straightforward level sensing of the contacts, rather than multiplex scanning as in a computer keyboard. This simplifies the interfacing task enormously.

A c-mos switch across the contacts usually

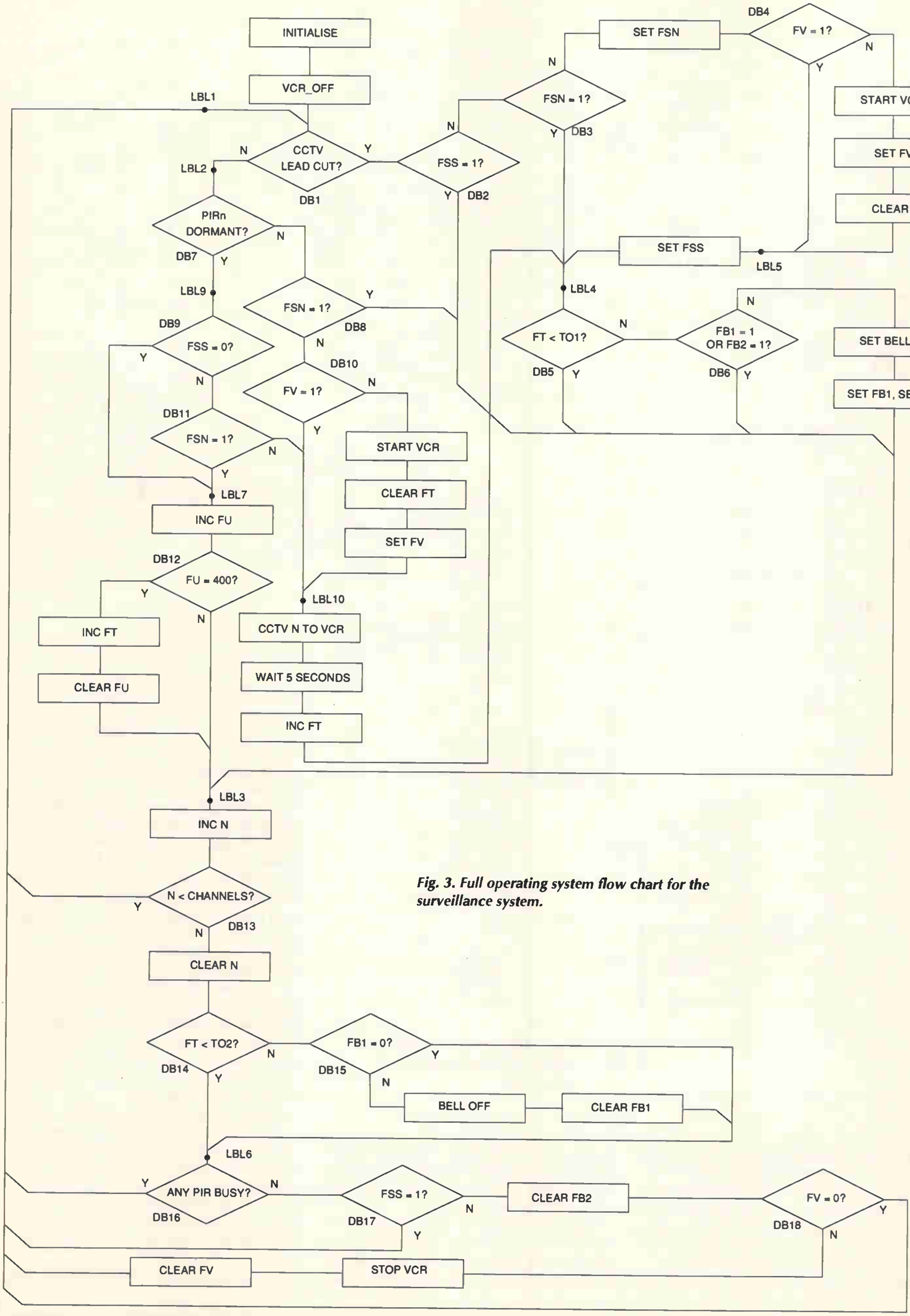


Fig. 3. Full operating system flow chart for the surveillance system.

bridged by a button does the job. The only minor complication is that the c-mos switches used run from ±5V rails of the equipment, whereas the remote control uses a four-cell 6V supply. So the common negative rail of the remote control was simply strapped to the –5V rail of the equipment.

Figure 2 shows the greater part of the hardware, including the microcontroller. The controller is represented here as IC_2, but it is in fact a small pcb with pins that fit a standard 24-pin 0.6in DIL socket. On this pcb are a number of components, see the panel.

Inputs P_{0-5} (pins 5-10) monitor the six camera cut-lead detectors. Inputs P_{6-11} monitor the outputs of up to six PIR detectors. If only, say, three of the the PIR detectors are fitted, the corresponding inputs at R_{2D}, R_{2E} and R_{2F} can simply be left open circuit: they will then be read as not busy.

Pins 17 to 19 (P_{12-14}) are used as address lines, to control which of the video outputs is switched by the eight-way video multiplexer IC_8 through to the video recorder. Likewise, IC_7 selects the corresponding sound channel, both outputs going to a Scart, or Peritel, socket.

A Scart plug to plug lead carries the signals to the vcr, most of which nowadays are provided with such an input. A lead with the full Scart implementation is not necessary. One with just composite video and audio suffices.

Whereas the *MAX398* at IC7 is a straightforward eight-way multiplexer, the *MAX455* at IC_8 includes a wideband output amplifier. Resistors R_{13} and R_{16} set the gain of this to ×2, so that it can deliver unity gain to a 75Ω load from a back terminated source.

Switch S_2 requires some comment. It was included in the early stages of testing, when the composite video was piped in to a 12in monochrome portable television. This had been used as a monitor in the days of home computers – remember the *UK101*? – in the early eighties.

The signal was simply fed into the output of the set's video detector, which worked fine if the signal was ac coupled. If it wasn't, the bias level is upset and the picture a perfect blank. When feeding the vcr it was not needed, and was switched out of circuit, as on a particularly contrasty picture, it tended to cause line tear and even occasional loss of frame lock.

Lines P_{12-14} are also used, in conjunction with P_{15}, to control the operation of the system. Line P_{15} is dedicated to switching the vcr on. The remote control used includes a feature to minimise the hazard of accidentally recording over a tape which carries a wanted programme, but where the record-protect tab has not yet been broken out. Thus starting the recorder requires two presses of the record button, i.e. two closures of normally-open switch IC_{8B}. These are indicated by flashes of diode D_2. Together with the other leds, this provides a very useful function monitor during testing.

With a maximum of six channels, numbered 0 through 5, lines A_{0-2} provide two spare codes, namely 6 and 7. These are decoded by IC_3 and one section of IC_6, used as an inverter, as the inverse codes, six bar and seven bar. Thus the output of IC_{3B} is normally high, and normally lit led D_3 acts as the equipment's on indicator. Briefly asserting a 7 on lines A_{0-2} douses the led, and ceases to hold IC_{6C} open, thus signalling the vcr to stop.

Briefly asserting a 6 causes a negative pulse at IC_{3D}'s output. This is inverted by one section of IC_5, toggling IC_4 and – via Tr_3 and the relay – turning the bell off or on as required. With a function controlled by a toggle, there is the possibility of it getting out of step due to glitches or race conditions, if these are not taken into account at the design stage. Here, the bistable device cannot toggle till long after the outputs of the various decode gates have settled, due to R'_{11} and C_2.

Additionally, bistable IC4 is reset at switch on, and whenever the reset button is pressed. D_5 acts as the bell on indicator, and usefully substitutes for the bell, ensuring blissful silence, during setting up and testing.

Designing the software

The software design started with a flowchart, which defined what the system was to do, in response to what stimuli and circumstances, and in what order. The flow chart was then implemented by coding it in the subset of Basic used by the *Stamp2* controller.

Once coded, the program is stored in the eeprom on the *Stamp2* device. It is transferred from the pc's serial port via an umbilical to the controller. Once programmed, the controller is inserted in the target system, in this case socket IC_2 of Fig. 2.

Naturally, the first stab at the flow chart proved to be faulty, the system not behaving as intended. Modifications followed, only to reveal different sets of bugs. The development was further complicated by minor modifications to the requirements listed earlier. After undergoing numerous changes, the flow chart finally looked like **Fig. 3**.

Each version of the flowchart spawned a corresponding version of the source code. Deriving this from the flowchart is relatively straightforward, once you have mastered the limitations and syntax of the Pbasic programming language used by the *Stamp2* microcontroller. But it is not entirely straightforward, since the source code comes out in a long linear form, **Fig. 4**, and not in the two dimensional form of the flow chart.

It would be instructive, if time consuming, to recast the flowchart into a single straight line, corresponding to the source code. It was certainly easy to get each section of code, correct in itself, in an unfortunate order, with the result that program execution ran round in an endless loop.

To facilitate cross referencing between the flow chart and the source code listing, the labels which are the subject of GOTO commands in Fig. 4 are marked in on the flow chart in Fig. 3, while the decision boxes on the latter appear as comments in the source code.

How the system operates

When success was finally achieved, the system operated as intended, which is as follows. The program cycles round all six channels in turn, looking at each to see if the PIR detector is 'busy', or if the camera lead has been cut. The course of action is decided at each turn by the result of tests, indicated on the flow chart as 'decision boxes', numbered DB_1, DB_2, etc.

If the PIR associated with camera channel N is found to be busy, the vcr, if not already on, is turned on and that channel is switched through to the vcr input. The channel is recorded for five seconds, after which the other channels are tested in turn.

If only the one channel is busy, it remains switched to the vcr, which records the channel uninterrupted. If two or more channels are busy, they are each recorded in five-second segments in turn.

As soon as a channel becomes busy and recording starts, an elapsed time counter – initialised to zero at vcr-on time – starts to count up. As the *Stamp2* does not provide a timer-counter or rtc, this has to be a function of your own program.

When time-out TO1 is reached the outside alarm bell is turned on, the time being a compromise between catching the intruder on tape, and not giving him – or her – time to break in at leisure. Time-out TO1 is settable in the software to any value you choose, some 20 seconds is currently shown in Fig. 4.

When the last active channel ceases to be busy, the vcr is turned off, provided no camera lead has been cut. The bell is turned off at the expiry of time-out TO2, regardless of whether any channel is busy or not, as this is an anti-nuisance requirement of most local authorities. However, when the last active channel ceases to be busy, the bell is re-enabled – provided no camera lead has been cut – ready in case a channel subsequently becomes active again.

What happens if the cable is cut?

All of the foregoing assumes that no camera lead has been cut. A cut camera lead is taken as evidence of a deliberate attack on the system, and results in an entirely different mode of operation.

A camera lead cut flag FS(N), indicating that camera N has been disabled, is set. A general camera lead cut flag FSS is also set. This results in the vcr being started, and five-second segments from each camera in turn – except of course the camera with the cut lead – being recorded indefinitely.

This continues until the system is reset or the tape runs out, after up to eight hours with a C240 tape and the video recorder in long-play mode. The bell starts ringing after time-out TO1 and stops after TO2. It is not available to ring again until the system is reset.

The various timeouts depend upon elapsed time counter FT being incremented after every five seconds of record time. This results in the elapsed time being little more than five seconds times the count in FT, since program execution proceeds at several thousand operations per second.

However, if a PIR were active only just long enough to set off the bell, and FSS were not set, then the route via DB8 and LBL10, Fig. 3, would not execute, FT would not be incremented and the bell would stay on for ever. This could easily be remedied by incrementing FT every time the (no recording) route

Fig. 4. Pbasic program listing for the complete security system. This routine is downloaded from a pc into the Stamp's non-volatile memory.

```
                  'VCR_SECY.BS2 video security system
TO1 CON 5         '01 TIMEOUT1, BELLON AFTER TO1X5 SECONDS
                  '02 OR OTHER HOLDOFF IF PREFERRED
TO2 CON 15        '03 TIMEOUT2, BELLOFF AFTER TO2X5 SECONDS
                  '04 SAY 15 FOR INITIAL TEST ONLY, ELSE 200
CHANNELS CON 3    '05 NUMBER OF CCTVs FITTED
M VAR NIB         '06 CURRENT PIR NUMBER
N VAR NIB         '07 CURRENT CHANNEL NUMBER
P VAR NIB         '08 PIR INDEX
FSS VAR BIT       '09 GENERAL CAMERA LEAD CUT BIT
FS VAR BIT(6)     '10 CHANNEL CAMERA LEAD CUT BIT
FV VAR BIT        '11 VCR FLAG, 1 = ON
FB1 VAR BIT       '12 BELL CONTROL BIT
FB2 VAR BIT       '13 ANOTHER BELL CONTROL BIT
FT VAR WORD       '14 ELAPSED TIME COUNTER - 5-SECOND UNITS
FU VAR WORD       '15 AUX. ELAPSED TIME COUNTER, FAST LOOP
OUTD = 0          '16 CLEAR P12 - P15 OUTPUT LATCHES
DIRD = 15         '17 SET P12 - P15 AS O/P (P0-P11 I/P)
FSS = 0           '18 CLEAR GENERAL CAMERA LEAD CUT FLAG
FV = 0            '19 CLEAR VCR FLAG
FB1 = 0           '20 CLEAR BELL CONTROL FLAG
FB2 = 0           '21 CLEAR BELL CONTROL FLAG
FT = 0            '22 ZERO ELAPSED TIME COUNTER
M = 0             '23 ZERO CURRENT PIR NUMBER
FOR N = 0 TO 5    '24 LOOP TO -
FS(N) = 0         '25 ZERO -
NEXT              '26 ALL SIX CAMERA LEAD CUT FLAGS
N = 0             '27 CLEAR CURRENT CHANNEL NUMBER
P = 0             '28 CLEAR PIR INDEX
GOSUB VCR_OFF     '29 TURN OFF VCR - JUST IN CASE
PAUSE 10000       '30 DON'T CLASH WITH LINE 37
LBL1:             '31 LABEL 1 - SEE FLOWCHART
IF INS.BIT0(N) = 1 THEN LBL2      '32 DB1: CCTV N LEAD CUT?
IF FSS = 1 THEN LBL3              '33 DB2: ALREADY FLAGGED?
IF FS(N) = 1 THEN LBL4      '34 DB3: IS CHANNEL FS FLAG SET?
FS(N) = 1                   '35 IF NOT THEN SET
IF FV = 1 THEN LBL5         '36 DB4: VCR ON?
GOSUB VCR_REC     '37 IF NOT THEN TURN ON
GOTO LBL5         '38 ALREADY ON
LBL5:             '39
FSS = 1           '40 SET GENERAL CAMERA LEAD CUT FLAG
LBL4:             '41
IF FT < TO1 THEN LBL3 '42 DB5: STILL TO REACH TIMEOUT 1?
IF FB1 OR FB2 = 1 THEN LBL3     '43 DB6: BELL ALREADY ON?
OUTD = 6          '44 IF NOT THEN TURN IT ON
PAUSE 500         '45
OUTD = N          '46 RESTORE CURRENT CHANNEL NUMBER
FB1 = 1           '47 SET BELL ON FLAG
FB2 = 1           '48 DITTO
GOTO LBL3         '49 FINISHED LOOKING AT THAT CHANNEL
LBL2:             '50
M = N + 6         '51 CAMERA INPUT CORRESPONDING TO N
IF INS.BIT0(M) = 0 THEN LBL9    '52 DB7: CH. N PIR DORMANT?
IF FS(N) = 1 THEN LBL3 '53 DB8: IF NOT, IS CAMERA LEAD CUT?
IF FV = 1 THEN LBL10   '54 DB10: IF NOT, IS VCR ALREADY ON?
GOSUB VCR_REC          '55 IF NOT THEN START VCR
GOTO LBL10             '56
LBL9:                  '57
IF FSS = 0 THEN LBL7        '58 DB9: IF GEN. CAM.
                             LEAD CUT FLAG CLEAR
IF FS(N) = 1 THEN LBL7      '59 DB11: IF CHANNEL CAMERA
                             LEAD CUT FLAG NOT SET -
GOTO LBL10        '60 (SWITCH CURRENT CAMERA TO VCR)
LBL7:             '61
FU = FU + 1       '62 INCREMENT FAST LOOP COUNTER
IF FU < 400 THEN LBL3 '63 DB12: 400 OF FU EQUALS ONE FT
FT = FT + 1       '64 SO INCREMENT FT -
FU = 0            '65 AND ZERO FU
GOTO LBL3         '66
LBL10:            '67
OUTD = N          '68 SWITCH CURRNT CAMERA TO VCR
PAUSE 5000        '69 RECORD FOR 5 SECS BEFORE PROCEEDING
FT = FT + 1       '70 NOTCH UP ANOTHER 5 SECONDS ELAPSED
GOTO LBL4         '71
LBL3:             '72 ALL FINISHED WITH CURRENT CHANNEL
N = N + 1         '73 START NEXT -
IF N < CHANNELS THEN LBL1  '74 DB13: DO NEXT CHANNEL -
N = 0             '75 OR CHANNEL 0 IF THROUGH
IF FT < TO2 THEN LBL6  '76 DB14: BELL ON TIMEOUT EXCEEDED?
IF FB1 = 0 THEN LBL6   '77 DB15: IF SO IS BELL OFF ALREADY?
OUTD = 6          '78 IF NOT, TURN IT OFF
PAUSE 500         '79
OUTD = N          '80 RESTORE CURRENT CHANNEL NUMBER -
FB1 = 0           '81 AND CLEAR BELL FLAG 1
GOTO LBL6         '82
LBL6:             '83
FOR P = 6 TO 11   '84 IS ANY CAMERA CHANNEL -
IF INS.BIT0(P) = 1 THEN LBL1    '85 DB16: STILL -
NEXT              '86 BUSY?
IF FSS = 1 THEN LBL1    '87 DB17: GENERAL
                         CAMERA LEAD CUT FLAG SET?
FB2 = 0           '88 IF NOT CLEAR BELL ON FLAG 2
IF FV = 0 THEN LBL1    '89 DB8: IS VCR STILL ON?
GOSUB VCR_OFF     '90 IF SO, THEN TURN IT OFF
GOTO LBL1         '91
VCR_OFF:          '92 ROUTINE TO TURN OFF VCR
OUTD = 7          '93 ("PRESSES BUTTON" -
PAUSE 500         '94 ON VCR CONTROLLER -
OUTD = 0          '95 RELEASES BUTTON)
FV = 0            '96 CLEAR VCR ON FLAG
RETURN            '97
VCR_REC:          '98 ROUTINE TO TURN ON VCR
OUT15 = 1         '99 THIS "BUTTON" -
PAUSE 500         '100
OUT15 = 0         '101 REQUIRES -
PAUSE 500         '102
OUT15 = 1         '103 TWO -
PAUSE 500         '104
OUT15 = 0         '105 PRESSES
FV = 1            '106 SET VCR ON FLAG
FT = 0            '107 ZERO ELAPSED TIME COUNTER
RETURN            '108
```

from LBL9 to LBL3 executed. But then, with a six-camera installation, there would be a delay of up to 30s in looking at a given channel again, after its PIR became busy.

To avoid both problems, an auxiliary counter FU is used. This is incremented every time a channel is found not busy (DB_7 via LBLs 9, 7 and 3). When it reaches 400, FT is incremented and FU reset to zero. This makes FT increment at a more or less steady rate, however many or few channels are busy.

Setting up and testing

If the microcontroller has been successfully programmed as per the listing in Fig. 4, and the hardware correctly built and connected to cameras, PIRs and vcr per Figs 1 and 2, the system should work first time at switch on – you should be so lucky!

In practice, it is better to test the system stage by stage as each is completed, starting with the suite of regulated power supplies. Note the earlier comment that the loading on the 12V supply must exceed that on the others.

The camera cut-lead detectors can be tested without the cameras in circuit, by operating S_{3A}, S_{3B}, etc. Likewise, the multiplexers and the video-recorder control circuits can be checked by temporarily patching pull-down resistors to 0V and switches to +5V to address lines $A_{0\text{-}3}$, before ever the microcontroller IC_2 is fitted. The leds indicate the appropriate functions as they occur.

The experienced electronic engineer will already know the wisdom of this stage-by-stage approach to equipment commissioning. If my advice is followed, the less experienced will not have to learn it the hard way.

With all the peripheral circuitry working, the cameras, PIRs, vcr and the remote control for the latter can be connected, and IC_2 installed. At switch on, red led D_3 should light and then briefly wink off, indicating the transmission of a vcr off command. This is included as a standard part of the reset/initialisation routine, because the vcr may already be on when the reset button is subsequently pressed.

After a ten second delay, to avoid confusing the vcr by sending conflicting commands in quick order, program execution continues. Note that as C_{1A}, C_{1B}... are initially uncharged, all PIRs signal that they are busy: consequently yellow led D_2 now flashes twice, indicating a vcr on command.

At this stage, the vcr should start and as many cameras as there are PIRs fitted, should be connected at five second intervals to the vcr. If no PIR is connected to $R_{1(N)}$, that channel will be read as not busy. Five seconds was chosen as a suitable sample length for each camera, allowing a good look at anyone in the field of view, while not unduly extending the delay before the next look, in a multi-camera system.

Early video recorders could take up to a second or more to synchronise with a video input signal, and in this system of course, the composite video from the cameras is not synchronised. Modern vcrs lock up very quickly when the video source is switched, and are apparently, unlike their older predecessors, not harmed by repeated resynchronising.

After a delay of about 0.7CR, each channel will cease to be busy. The CR in question is $R_3 C_1$, which with the values shown comes to about ten seconds. This large value was chosen for convenience in testing and C_1 may be reduced in value by a factor of ten or more if desired. There will in any case be a minimum PIR busy period set by the relevant control on the PIR. The other control on the PIR mentioned has the useful facility to be set so that detection occurs even in broad daylight.

After time out TO1, the green led D_5 should light, and when the last channel ceases to be busy, D_3 should briefly douse to signal the issue of a vcr stop command. After the delay set by TO2, diode D_5 should extinguish, indicating bell off.

If the reset button is now pressed, D_3 should indicate the sending of vcr off, even though the machine is off already, but there should be no subsequent vcr on indication from D_2, as all the $C_{1(N)}$ capacitors are already charged up. Now, waving a hand in front of one of the PIRs should start the vcr and switch the corresponding video signal to it, followed by the usual sequence of time outs.

For development purposes, the system was set up with nominally three channels, although only two cctv cameras were fitted. This was so that switch S_{3C} could be opened, signalling that the lead to camera three had been cut. This causes the software to cycle around indefinitely, connecting the remaining two cameras to the vcr alternately.

Installation

The PIR units mentioned above are listed as suitable for outdoor installation, although the leaflet that comes with them recommends that a sheltered location, such as under eaves is preferred.

The camera units are bare, and so will need weather protection. An enclosure from the RS range of clear lid boxes sealed to IP65 is the obvious choice. A hole in the underside, covered with mylar film, may be needed to permit sound to reach the camera's built in microphone, although the microphone is remarkably sensitive. And of course, a hole will be necessary to pass the supply lines to – and the signal lines from the camera.

It is recommended that the mains supplies to the PIRs should be from sockets mounted on the central control equipment itself. This will ensure that the case is earthed whenever the supply to a PIR is on.

The PIR unit is completely enclosed in plastic, there being no available earth terminal in it. So the third lead of its supply cord was used to bring back the PIR busy line to R_{1A}.

If the optional PIR lead cut detector circuit is implemented, then a four core mains lead will be needed. With the two metre mains lead to the PIR used during development, there was some residual current via the PIR busy lead even when the PIR was not tripped. This is presumably due to capacitance across the relay contacts and between cores in the mains lead, and could be a problem with a long run to a PIR unit.

As mentioned earlier, a dedicated vcr is preferred, as this can be permanently connected to the controller, both being mounted out of sight. After all, if a burglar does gain access, any video recorder on view will be the first thing to go.

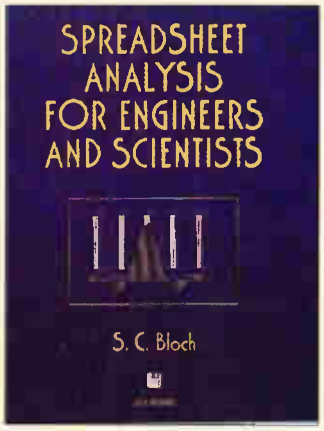
Customisation

Most of the parameters used by the control program, such as the number of channels and the length of time outs, are defined as constants at the start of the listing. This makes them easy to modify to suit individual requirements or preferences, without the tedium of combing through the listing to change them wherever they occur. More drastic changes to the software, such as modifications affecting the flow diagram, are possible, but you are then on your own.

I cannot guarantee that the listing shown is 100% bug free – most programs aren't anyway. For instance, if a particularly nimble thief were to manage to locate and cut all N camera leads in less than TO1 seconds, the alarm bell would never be turned on. In this case, only the odd glimpse of what was going on would be captured, from some of the last cameras to be cut. However, oddballs such as this apart, no malfunctions have been observed in practice.

Likewise, the hardware can be adjusted to fit your requirements. For instance, if you envisage that the maximum eventual number of installed channels will not exceed five, then IC_5 in Fig. 2 can be one section of IC_1 – as is the case in my system.

On the other hand, the system is easily expanded to eight or more cameras and PIRs. This can be achieved by using multiplexers, addressed by $A_{0\text{-}2}$, to select which PIR busy – and PIR lead cut detector – signal is routed to one of the *Stamp2* microcontroller input pins. Thus instead of a six channel system requiring twelve input pins, as in Fig. 2, an eight channel system would require only two. Additional address leads would cope with further multiplexers, permitting 16 or 32 channel systems to be built. ■



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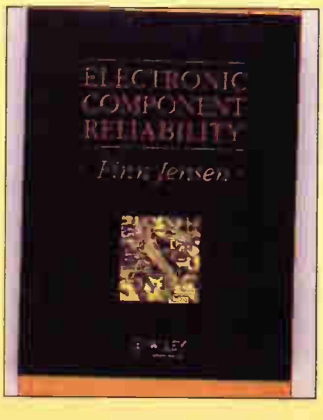

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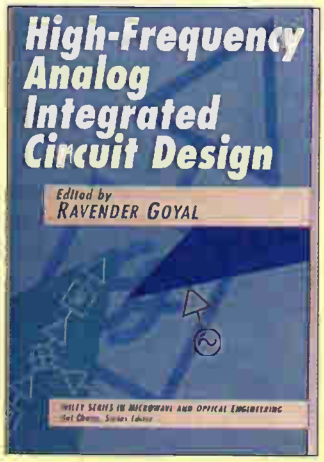

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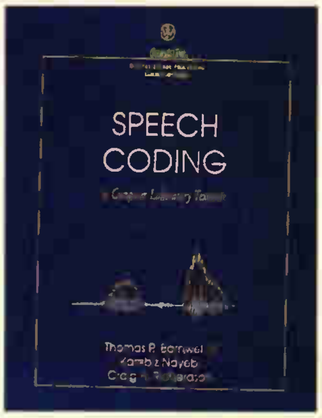

perform a wide variety of speech coding and speech processing experiments. The text presents and explains a set of basic speech coders analytically and in terms of the specific parameters controlling each coder. The manual leads the student through the experimental process of understanding how speech coders work and sound via over 70 exercises and projects. The class-tested menu-driven, Dos-based software can be operated by students with little or no training.

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Fazed by phase?

Having looked at numerous existing phase measurement circuits, Cyril Bateman found that none of them could guarantee 0.1° resolution at frequencies to 100kHz – so he designed his own.

At the simplest level, the phase angle of two waveforms is the time difference at their zero crossings as a proportion of their periodic time. Multiplied by 360, this becomes their phase angle expressed in degrees. This principal is used by many dedicated test instruments which measure phase.

But other techniques exist. Perhaps the most accurate method involves digitising both input signals, then performing a Fourier transform to derive their phase information. Essential to this method is 16-bit, or better, resolution, and time-coincident, dual-channel analogue-digital conversion.

Below 20kHz, suitable low-cost integrated circuits as used for digital audio are available, but not so for higher frequencies. While this method works extremely well, it requires significant amounts of hardware and software.

I recently needed a low cost, portable, easy to use phase meter, capable of working to 100kHz with at least 0.1° resolution. To start with, I researched my library and the Internet for previously published circuits that could be quickly built. References for a representative selection of the various techniques that I found is presented later.

While other circuits undoubtedly exist, none of the ones found suited my needs. Some required a known phase source for calibration. Others gave insufficient phase resolution, were limited to low frequencies or simply did not offer the desired accuracy.

One particularly common failure was a measurement ability to 180°, but without identification as to whether leading or lagging phases were being measured.

Ways of measuring phase

A common method for measuring phase involves the use of op amps or comparators to produce either a square wave having rising and falling transitions at the input waveform zero crossing, or a ttl compatible waveform. A second identical channel is provided for the reference waveform.

The time difference between these two square or ttl waveforms is then extracted using an exclusive-or gate. The width of this pulse represents the time difference to be measured.

Having generated the pulse, two techniques were used to measure its width compared to the periodic time of the waveform. The most common was to integrate or average this pulse over many time intervals, giving a direct measure of the phase difference. Less common was measurement of the actual time duration of this 'difference' pulse, in comparison with the periodic time, **Fig. 1**.

Integration method. A difficulty with any integration method is its dependence on the accurate generation of a 'squared-up' version of the two input signals – especially equality of rise and fall times – and ensuring a consistent pulse amplitude. At audio frequencies this is simple but at 100kHz and above, becomes more difficult.

While ls-ttl circuits can provide reasonably fast rise and fall times, the mandatory 5V supply combined with their permitted variation of output voltage swing, requires the circuit be accurately calibrated with signals of known phase. Ideally any circuit should be self calibrating or be acceptably accurate when calibrated using dc only.

One slightly unusual approach to this integration problem was found in a *Wireless World* article by C. Hodgson[1]. This circuit used a number of nand gates to generate a push-pull output together with full wave recti-

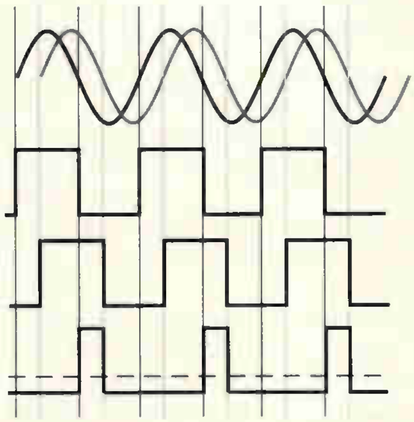


Fig. 1. Two common and practical phase-measurement methods. Measurement of the time difference between signals or integration of this phase difference pulse, give identical answers.

Fourier transform background

The Fourier-transform integral was devised by John Baptiste, Baron de Fourier in his Analytical Theory of Heat[5], published 1822. In practice a modified technique – the fast-Fourier transform – is more commonly used. This straightforward method permits conversion to and from the time and frequency domains. It can be performed using a dedicated spreadsheet routine or simple computer programs,[6] as well as by dedicated digital signal processing integrated circuits.

In the context of this article, a time-domain repetitive waveform can be transformed into its frequency domain components, both amplitude and their relative phases. Assuming two single-frequency sinewaves of differing amplitudes and phase have been accurately measured in the time domain, for example by digitising, then their relative phases can be accurately calculated.

This technique is used in certain high-frequency, superheterodyne vector network analysers to ascertain both reference and measurement signal phases. Furthermore, it is a most useful technique[7] when evaluating the power impressed by complex waveforms, on both capacitors and inductors.

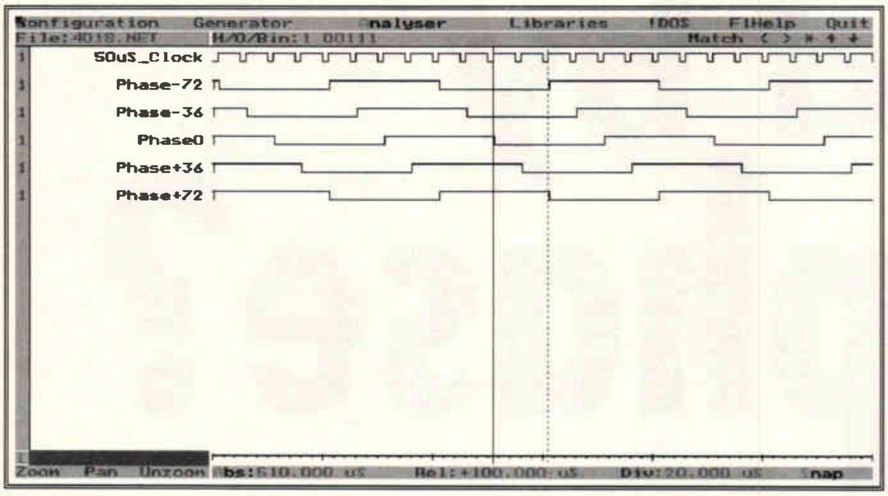


Fig. 2. Simulation showing the five different lower frequency phases output with the 4018 IC set to divide by ten. Also clearly shows that due to edge transition switching, the clock signal mark/space ratio is unimportant.

fication and addition of the two signals.

The integration method can provide unlimited resolution of the measured phase, but remember that increased discrimination – while desirable – does not of itself improve the measurement accuracy.

Time measurement method. My searches found two approaches were commonly used to measure time difference. Both used counting techniques together with either a fixed high frequency clock or a phase locked loop generated clock frequency 360 times higher than the frequency being measured.

Superficially, this time measurement method looks the more attractive, especially for low or audio frequencies. However to provide even one degree of resolution at 100kHz, requires clocking the circuit at 36MHz. This is too fast for c-mos or ls-ttl and approaching the limits of high speed c-mos. To provide a 0.1° resolution needs even faster clocking and ecl technology.

Is a different approach needed?

With these problems in mind, the application notes for the Analog Devices *AD630* looked most interesting. This precision op amp has two independent differential input stages together with a precision comparator used to select which input stage is active. The combination of this rapid response comparator with high slew rate fast settling amplifiers, minimises distortion when switching channels.

Having benefited from the hindsight gained by examining the techniques used in the circuits found by my searches, I decided to investigate my own circuit solution, based on some newer integrated circuits.

Alternative ideas

The *AD630* can be used as a balanced modulator or demodulator. In the balanced modulator configuration, its output is directly proportional to signal amplitude and phase differences. Using squared up constant amplitude reference and measure signals, applied to the modulation and carrier inputs, the output after integration, is a precise measure of the phase difference between these input signals.

This circuit is particularly attractive for examination of quadrature signals, since signals having either 90 or 270° phase difference, produce a zero output.

Following PSpice simulations, a circuit based on this IC was bread-boarded, and initially worked quite well. Unfortunately as testing progressed, to provide the needed performance at higher speeds, the circuit became much larger and more complex than expected or desired. So I sought an alternative.

A new approach

Burr-Brown makes a higher speed, but otherwise similar, switched input amplifier, the *OPA678*. Again, I simulated a circuit using this chip. While overcoming some of the speed problems found with the *AD630*, I concluded that it was not the solution that I was looking for.

A problem common to both the *AD630* and *OPA678* circuits stemmed from my use of the *AD790* precision high speed comparator. This outputs a ttl level signal that can be disabled or reset according to the voltage applied to pin 5.

Ideally, the *AD630* and *OPA678* require signals balanced to earth, and hence accurate level shifting. This could be as simple as ac coupling using a *CR* network, but it must avoid exceeding the capacitance load limits for the comparator. My searches failed to find a more suitable zero crossing comparator or one having an output balanced to earth; all seemed to provide ttl compatible outputs.

One difficulty when developing a phase meter is how to generate stable signals of accurately known phase, for calibrating the circuit. Obviously, generating either zero or 180°, is no problem, but intermediate phases which are reliably known and repeatable, are rather more difficult. The lack of suitable test signals having known phase, halted this project for a time.

Producing test signals

while leafing through my c-mos handbook, I found a method for generating suitable test signals by chance when I noticed the output waveforms for the *HEF4018B* integrated circuit.

The *4018* is a presettable divide-by-*N* counter that can be used to divide a clock signal by ten. At this ratio, each output goes low, and remains low for five clock cycles, then remains high for a further five clock cycles. The chip has five outputs, which change state in turn with each rising clock transition.

This circuit only responds to positive going clock transitions. As can clearly be seen in the Pulsar simulation, the equality of the mark/space ratio of this clock signal is unimportant, the output waveforms always have a unity ratio, **Fig. 2**.

In this way, selectable and equal mark/space ratio signals at each 36° of phase at the lower frequency can be generated. These could be used directly to trigger ttl or c-mos circuits – or equipped with suitable low pass filters – could provide acceptable quality sine wave

More phase meters

Examples of circuits found in my library search.

Circuits using multiple gates:
Hodgson, C, 'Stereo Power and Phase Meter' *Wireless World*, May 1978.

Circuits using xor or nand gates:
Wyatt, M, 'Phase meter uses just two chips' *Electronic Design*, June 1989.

Jones, D, Audio Frequency and phase comparator, *New Electronics*, October 1981.

Nixon, S, Economical Phasemeter for Education use, *New Electronics*, September 1981.

Boreham, N, 'Phase Meter,' *Wireless World*, August 1979.

Circuits using counting or Phase Locked Loop Systems.
Sabah, N, 'PLL performs accurate phase measurements' *Electronics*, June 19 1980.

Jarman, H, 'Digital Phase Display,' *Electronic Product Design*, Feb. 1988.

Ward and Walsh, '360 degree phase detector,' *Electronic Engineering*, April 1987.

Lucassen, R, 'Digital Phase Meter,' *Elektor Electronics*, June 1991.

Circuit found on Internet:
Simple phase meter operates to 10 MHz, AN9637, Aug. 1996 Harris Semiconductors.

signals of known phase relationship.

The phase difference of these signals is accurately maintained, subject only to use of a stable clock working within the c-mos clock rate limits and the intrinsic rise and fall time variations of the IC itself, **Fig. 3**.

Accepting the *AD790* comparator's ttl output, it seemed sensible, having once generated a ttl signal to seek a digital solution, satisfying my requirements for accuracy, resolution and ease of calibration. Ideally the final output would give ±1.8V representing ±180° of phase, providing 0.1° resolution with a 3.5 digit or 0.01° with 3.75 or 4.5 digit meters.

Renewed attempt

Having now the means to verify my circuit ideas, I decided to examine the circuit of a frequency/phase detector often used in phase locked loop applications[2] using a dedicated logic circuit simulator called Pulsar.[3] This did not provide a solution, but further examination of phase-locked loop circuits suggested the circuit finally adopted which provides one special benefit.

Like the Hodgson circuit, it has a push-pull output. Used as a phase detector, it gives a clear distinction between leading and lagging phases. Needing no frequency halving of the signals, it similarly eases subsequent integration. The simulation results looked very good, exhibiting no unstable states.

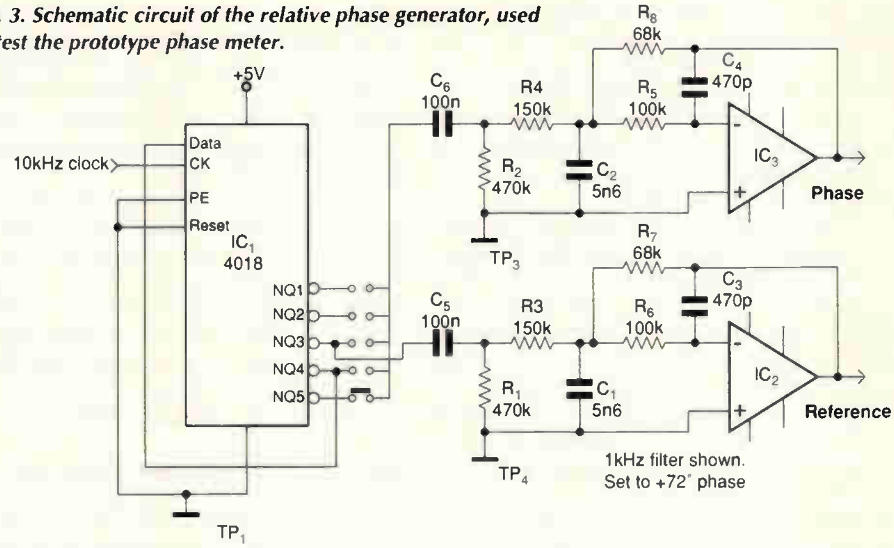
This phase detector triggers only on falling-edge transitions of the reference and measurement input signals, so does not require signals having unity mark/space ratio. Many other circuits use frequency division to ensure signal equality, aggravating integration at low audio frequencies.

With two discrete input and output ports, depending on whether the input or reference signal is leading, one output stays high while the other gives a negative pulse of width corresponding to the time difference between the two signals. These output signals can be accurately integrated using a unity gain instrument amplifier, which automatically provides level shifting to ensure an output balanced about earth. Common mode and differential *CR* filters, which pre-integrate and slow the very fast rise and fall times output by the phase detector, facilitate choice of the in-amp used.

By way of demonstration, this Pulsar simulation uses duplicated circuits to simultaneously show the outputs for both lagging and leading 90° phases. With measurement lagging, out_R has a negative going pulse exactly one quarter the width of the repetition period, while the out_M remains high. With a leading phase signal, the corresponding outputs out_{R2} and out_{M2} exhibit exactly similar but interchanged waveforms, Fig. 4.

PSpice simulations of a Burr Brown *INA118* instrumentation amplifier confirmed the accuracy of this approach. The phase detector's push pull output signals were pre-filtered in *CR* circuits using 1% resistors and 1% capacitors. Capacitors should be either polypropylene or polystyrene, to ensure minimal dielectric absorption.

Fig. 3. Schematic circuit of the relative phase generator, used to test the prototype phase meter.



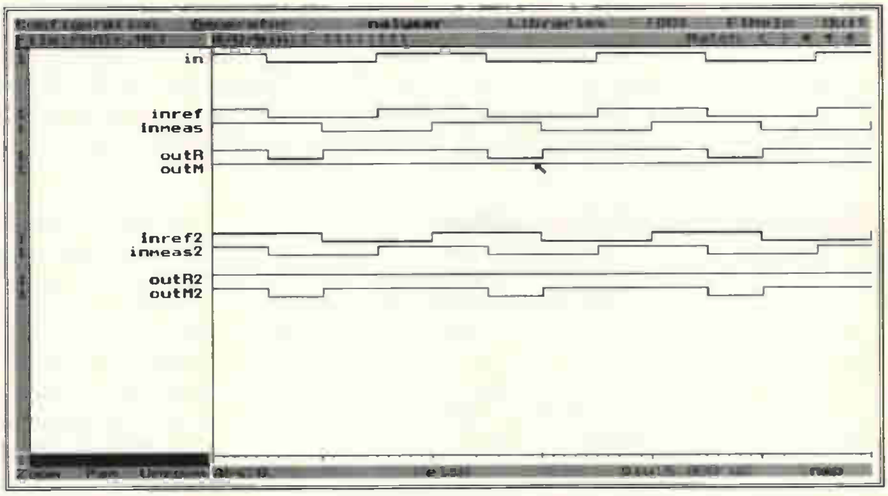


Fig. 4. Simulation showing phase detector outputs, out_R and out_M for 90° of lagging phase, while with leading phase the output pulses interchange, see out_{R2} and out_{M2}.

Practical implementation

Originally, to simplify calibration, I planned to use 74HC integrated circuits powered from a nominal 3.6V supply rail. These HC op-amps can use 2 to 6V supplies yet swing within 100mV of the supply voltage and earth. When powered at 5V, they exhibit equal rise and fall times of around 4ns.

When bread-boarded, reducing the supply to 3.6V gave two problems. The rise and fall times were slowed and with the numbers of gates used, switching simultaneously, resulted in considerable supply rail and ground-plane noise levels, for 100kHz and faster, signals.

This noise is caused by both output transistors of a gate conducting simultaneously, albeit for less than 2ns while switching. It causes an intermittent near short circuit on the supply rail. Obviously this supply rail noise also appears as substantial ringing on the required output pulses, causing a glitch to appear on those outputs which remain at the high level.

The solution was to leave the phase detector running on 5V then to clean up and adjust the levels of the final output waveforms. To do this I used a Schmitt triggered buffer running on its own 3.6V nominal, adjustable, independently stabilised supply line. This line was increased as necessary to provide a 3.6V output swing, giving a most useful 10mV-per-degree correlation.

I considered using the new 3.6V LVC integrated circuits which are 5V input tolerant. But I found that I needed to increase the supply above 3.6V, sufficient to considerably exceed the 3.6V maximum voltage rating of LVC.

The final solution

Both 74HC and 74AC integrated logic circuits have an input overdrive limit of 0.5V or 20mA. These limits are exceeded when driving a chip supplied at say 3.8V, from an HC circuit having a 5V supply. To prevent reverse conduction of the chip input diodes exceeding this 20mA limit, 100Ω series protection resistors were used on both linking inputs.

To clean up the phase detector's push pull signals, a 74AC14 Schmitt input hex inverter, which has very fast rise and fall times and can both source or sink 24mA from each output, was chosen.

The *CR* filters used to pre-filter the in-amp, present a large capacitive load albeit decoupled by the 10kΩ series resistors, to the 74AC14 outputs. To minimise any effect, two inverter outputs were used in parallel, for each push pull output signal. The complete circuit used for the prototype can be seen in the schematic drawing, **Fig. 5**.

With the phase detector part of this schematic already drawn for use in the Pulsar simulations, the final circuit was translated into a double sided printed circuit board using *Easy-PC Professional XM*[3] to minimise redrawing. To minimise supply line noise, this design used one board face to ensure the least possible ground plane impedance. A wide and short 5V supply, low impedance transmission line, was used to power the logic chips. All tracks, and components were mounted on the second side together with eight wire links needed to avoid disturbing the ground plane.

Following initial tests, to further reduce supply line noise, 0.1μF 1206 size ceramic surface mounted chips were directly soldered to each integrated circuit power pin and to the ground plane – as close as possible to the chip.

Each integrated circuit, whether analogue and digital, was thus doubly decoupled using 0.1μF radial-lead multilayer ceramic capacitors on the component side of the board, together with 0.1μF surface mount 1206 size multilayer ceramics soldered directly to each power pin and the ground plane. This reduced noise to acceptable levels.

To optimise circuit speed, all integrated circuits used in the phase detector were soldered directly to the double-sided printed circuit board, interconnected as far as possible, using 100Ω impedance signal tracks. To avoid the inductance found with conventional sockets, but permit chip interchange, all other integrated circuits were mounted using Harwin ultra low profile individual sockets.

How did this prototype shape up?

Using ±15V supplies for the op amps, the circuit triggers reliably with signals down to 10mV, and can accept signals up to 10V. Since I use this meter with both 1MΩ oscilloscope probes and 50Ω measurement impedances using coaxial cable with external terminators, no attenuators could be built into the prototype. Used with switchable divide ten oscilloscope probes the usable range becomes 10mV to 100V, which should suffice.

The prototype was housed in a die cast box of 150 by 80 by 50mm together with the PM-128 digital panel meter recently offered via *EW&WW* and the ±9 volt batteries used for power. This provides portability and ground isolation but restricts the input voltage to 6V without prior attenuation.

While not revealed by my simulations, this phase detector can occasionally exhibit an anomalous start up mode with both push-pull voltages inverted. Instead of one output having a negative pulse with the other output high, this second output goes low and the output pulse becomes positive going hence the built in 2V full scale meter shows over-range.

While not intended or desired, this mode facilitates calibration. In this condition, and with no signals, adjust the nominal 3.6V supply rail using an external meter until the output circuit reads precisely 3.6V. This compensates both for the push-pull output voltage swing of the particular 74AC14 used, and any variation of the instrumentation-amp filter circuit 1% resistor values.

This anomalous mode is easily corrected. Simply switch off then re-power the circuit.

Applying the phase meter

An obvious use of the meter is for looking at phase change measurements in audio amplifiers and loudspeaker systems. In addition, I have used it, together with my reflection bridge circuit,[4] to measure capacitor and inductor behaviour at various frequencies and ac test voltages. This is to me a natural application, and was the underlying reason for needing both 1MΩ and 50Ω input impedance levels.

Perhaps less obvious, but also related to component measurement, it has been used to measure the phase component in conventional four-terminal volt/amp impedance measurements. Using 50Ω coaxial cable with quality through-terminator and R_{sense} resistors, connected to the device under test with minimal lead lengths, this technique is valid to high frequencies. It is limited only by the volt and phase meter frequency limitations.

The simple correlation shown on the diagram below is possible since the current common to both the device under test, i.e. DUT, and the R_{sense} resistor must have the same phase. These equations also assume that V_2 probe's shunt resistance is much larger than R_{sense} and its capacitive reactance is also very large compared to the device under test's reactance and R_{sense}. For most component values, both conditions are easily met by using a divide ten oscilloscope probe.

Hence,

$$esr_{DUT}=\cos(phase\ angle)\times|Z|-R_{sense}$$

$$reactance_{DUT}=\sin(phase\ angle)\times|Z|.$$

For negative reactance,

$$capacitance=\mathrm{abs}(1/(2\pi\times frequency\times reactance))$$

and for positive reactance,

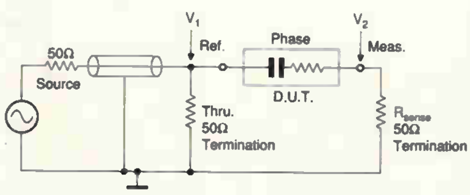
$$inductance=reactance/(2\pi\times frequency)$$

As with all capacitance and inductance measurements, accuracy of the minor loss term effective series resistance depends almost totally on the phase measurement accuracy and resolution. While the R_{sense} resistor used should be of known value, non-inductive and of minimal capacitance, its absolute accuracy has a lesser effect than any errors in phase measurement. Thus ensuring a good accuracy for the reactive term being measured.

To minimise measurement errors, the source impedance and R_{sense} should be of similar, low impedance. Much lower than the resistance and capacitive reactance loading of the test probes used. In practice using source and R_{sense} resistances of 50Ω, with divide by ten low capacitance oscilloscope probes, gives acceptable accuracy when used in the above four terminal circuit.

There are two easy ways to obtain R_{sense}. The preferred choice is an SMA 50Ω termination. These are generally made using thin film chip resistors and rated for 2W power. The completed termination is specified for voltage-standing-wave ratio to very-high frequencies, yet for its quality, is inexpensive. BNC/SMA adapters can be readily purchased if required.

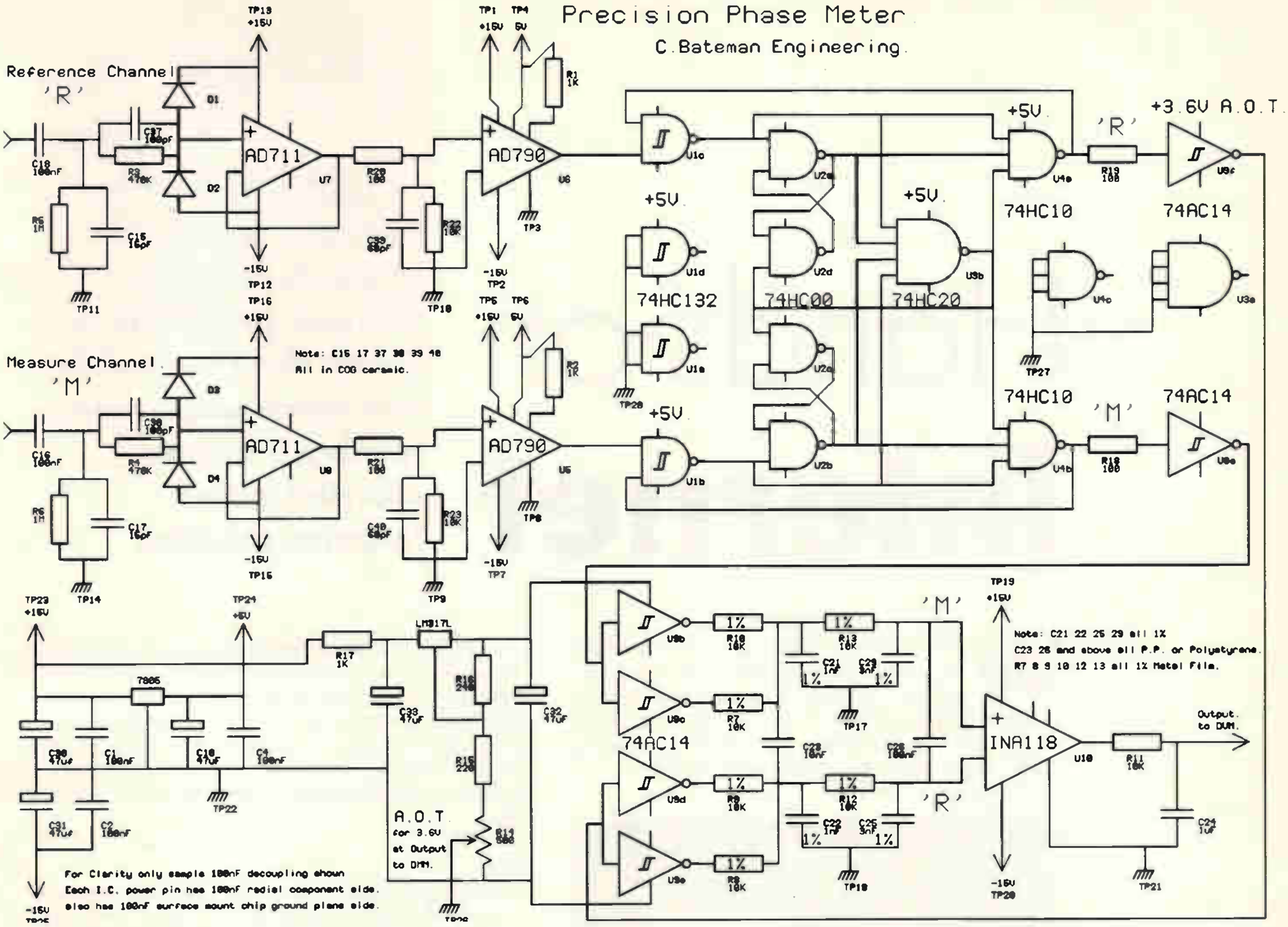
A poor second choice is 1% surface-mount chip resistors as these typically have 1nH of inductance. Note that many low-cost BNC terminations, unspecified for vswr at frequency, are built using 51Ω conventional resistor rods, hence are suitable only at low frequencies.



$I_{DUT}=V_2/R_{sense}$

$|Z|(DUT+R_{sense})=V_1/I=V_1\times R_{sense}/V_2$

By adding R_{sense} and through terminator, the meter can examine phase component in conventional four terminal volt/amp impedance measurements.



With no input signals or with the reset switch set, the output meter should display 0V.

Calibrated as described, two frequency and accuracy limitations result from the logic circuit rise and fall times and supply-rail noise related transients.

The 74AC14 integrated circuit has rise/fall times of typically 2ns. With some allowance for jitter, this represents perhaps 0.1° of phase at 100kHz. Supply line noise, particularly the approximately 2ns wide switching glitch appearing on the high outputs, causes a similar error. Both contribute to a small linear overstatement of the measured phase angle as frequency rises.

Given equal signal and reference voltage levels, the zero crossing accuracy using the *AD790* without offset trim is excellent.

However, even with an input hysteresis of only 0.5mV, should the two input signals differ greatly in amplitude, then some error is inevitable due to the difference in signal slew rate at zero crossing. The solution is obvious – try to equalise input signal levels.

Any offset in the input-buffer amplifier has a similar effect, so any protection diodes used must be matched for reverse leakage current. For simplicity, the prototype circuit was built without either protection diodes or trimming of *AD711*, *INA118* or *AD790* offsets.

I used the *4018* circuit, with its outputs selected to provide alternately ±36 and ±72° of phase difference test signals, as a test source for the dc-calibrated prototype phase meter. Measurements of this combination, from 100Hz to 100kHz input frequency, show negligible phase measurement variations to 10kHz, increasing to less than 0.5° at 100kHz – inclusive of the *4018* chip's error contribution, **Table 1**.

Due to its c-mos maximum clock rate limits, the *4018* circuit cannot be used to quantify errors above 100kHz. Measurement at 1MHz of a common signal source connected to both reference and measure inputs using equal 0.6m lengths of RG58 cable, reads zero degrees.

Fig. 5. Schematic of the final phase meter prototype as built and tested. Good results could be further improved by adding offset adjustments to the AD711, AD790 and INA118 stages, combined with balancing of the 1% components.

Increasing the cable length of one input by 1.25m resulted in a phase reading of 2.20°, very close to the theoretical 2.25°. This accuracy more than suffices for my present needs. ■

Table 1. Test results of prototype phase meter. Tests used the 4018 phase-generator circuit. Results measured with the 4.5-digit external dmm used for dc calibration.

Test frequency	Test signal –36°	Test signal +36°	Test signal –72°	Test signal +72°
100Hz	–35.97	35.99	–71.98	72.00
1kHz	–35.98	36.00	–71.99	72.00
10kHz	–36.02	36.05	–72.04	72.05
100kHz	–36.32	36.45	–72.22	72.38

References

1. Hodgson, C, 'Stereo Power and Phase Meter,' *Wireless World*, May 1978
2.Maxim Engineering Journal Vol. 19, Maxim Integrated Products UK, 01189 303388.
3. Pulsar digital logic simulator, Number One Systems, St. Ives, 01480 461778.
4. Bateman, C, 'Looking into impedance,' *Electronics World*, August 1997
5. Joules Watt, 'A series of remarks on Fourier,' October 1987.
6. Omer, W, 'Faster Fourier Transforms,' *Electronics World*, June 1986.
7. Bateman, C, 'Power dissipation in capacitors,' *EW+WW*, April 1995.

Readers interested in pcb layout should send an sae marked 'Phase Meter' to the editorial offices. Add £1.50 for photocopies of layout or nothing for details of pcb film price. Depending on demand, Cyril may make a pcb available.

Hands-on Internet

Cyril Bateman looks at CAD software on the net biased towards solving thermal problems in circuit design. And he has some good news for those of you worried about Internet security.

When reporting new Internet features, sometimes the ink has barely dried before the words are outdated. Last month, I mentioned AltaVista's 'Live Topics enhancement, which entered Beta testing at the Palo Alto site, in February. I had deferred discussing 'Live Topics' until it was established at all of the AltaVista sites.

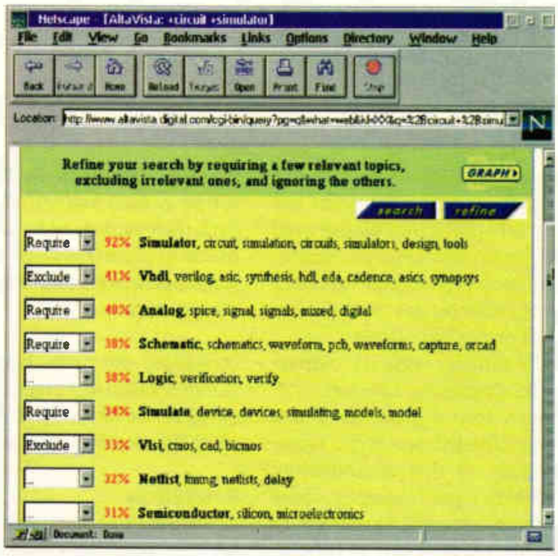


Fig. 1. Replacement AltaVista 'Live Topics' search wizard. My initial search using +circuit +simulator resulted in 9000 hits. The refined search produced 100 high quality results.

On 26 July, AltaVista introduced a revised 'Live Topics' user interface, now accessed by clicking on the 'Refine' button at the far right of the display. This latest interface also presents a similar list of key words, for selection or deselection. You must now open a selection box, in order to select or deselect its keywords. An optional graphics interface is also available by simply clicking on 'Graph', **Fig. 1**.

Bugs

Currently, the Internet is alive with two main news topics – namely the Presley anniversary and 'bugs'. Many reports describe two major bugs which potentially affect most Internet users. The 'Out of Band' bug, which exploits weaknesses in the Microsoft TCP/IP stack, is variously known as 'SSPING', 'JOLT' or 'NUKE'.

This bug first appeared in the Internet's relay chat 'IRC' channels, as a means to deliberately stop others talking, by shutting down their computers. It seems that if a Windows-based machine, on-line via a modem, receives a string of invalid instructions addressed to its port 139, then Windows immediately shuts down, leaving a blank screen. It is claimed this bug affects all Windows operating systems, Windows NT, Windows 95 even Win 3.11.

According to a 'ClNet' news report[1] dated 12 May, these problems have resulted in glitches or shutdowns on Microsoft's own website NT machines, among many others. In true Internet fashion, this bug is variously and

Where to surf

1. Hole in Windows 95, NT fixed
   http://www.news.com/News/Item
2. Microsoft Knowledge Base.
   http://www.microsoft.com/kb/articles/q168/7/47.htm
3. WinNuke Testing Ground.
   http://206.148.240.160/~dirk/winnuke.html
4. The Singapore Privacy Bug.
   http://home.netscape.com/info/security-doc.html
5. Macinsearch.com
   http://www.macinsearch.com
6. Hands on Internet, *EW&WW*, March 1996
7. Micro-Cap V version 2.0,
   http://www.spectrum-soft.com/news/winter97
8. Spice runs thermal analysis.
   Wyatt, M, Design Ideas, *EDN*, 18 August 1994.
9. Flomerics Inc.
   http://www.flowtherm.com
10. Thermal Engineering Associates.
    http://www.thermengr.com
11. Optics, Automation and Instrumentation.
    http://www.oainet.com
12. Thermal Solutions Inc., http://www.sauna.com
13. Simulation – Electro-thermal.
    http://whirligig.ecs.soton.ac.uk/~mz/da/p212.htm
14. Tera Analysis Co.
    http://www.tera-analysis.com
15. Lakeview Software – Therm.zip
    http://ftpsearch.ntnu.no/ftpsearch

widely reported, including several sites which offer 'patches' and some even the facility to deliberately test your machines vulnerability, if you so request.

A quick search of AltaVista against 'WinNuke' identified a very large number of Web pages offering comment and advice, including items in Microsoft's knowledge[2] base. Windows 95 users should download the knowledge base article shown from Microsoft together with the file Vtcpupd.exe. When run, this file updates two files Vtcp.386 and Vnbt.386, **Fig. 2**.

Several of the 'Nuke' test sites identified by AltaVista, now only have messages explaining they had closed due to external pressures. 'WinNuke' however was working when I used it deliberately to try to 'Nuke' my own machine.[3] Since I run only native OS/2 Warp 4 Internet software and not Windows, it was not affected. However if you are using Windows and feeling really brave, why not give 'WinNuke' etc., a try?

And for Netscape users...

I said two bugs. The 'Singapore Privacy Bug' affects the Netscape Communicator browser software. Apparently it allows a Hacker to use Netscape's 'Live Connect' feature to monitor your Web activity, URL's visited, and data and passwords you enter into HTML forms. It even allows monitoring of data placed in a Netscape cookie file.

Some reports have suggested this monitoring could encompass details of any credit card details you entered on Internet forms and the contents of your e-mail messages. Netscape has issued a document[4] dated 25 July which accepts that Communicator has this problem but says Navigator 2.x and 3.x are not affected, **Fig. 3**.

Macintosh users now have a search page dedicated to their special needs. Macinsearch.com[5] hosts the Mac Web Network of Macintosh dedicated sites together with links to conventional Internet search engines.

One reader reported difficulty accessing certain addresses I quoted. All have been taken direct from the page and have been tested and work. However on occasions I too have subsequently found that a site's full address may not work depending on particular server routeing.

Deleting the last segment out of a long address, it then does. Once at the page, re-adding the deleted part address, it then also works perfectly. Occasionally this technique can be needed to follow up a link, from say AltaVista, should clicking on a link fail, with error 404.

I opened by noting how quickly some aspects of Internet change. Use of Web pages with a browser to download or upload files, now dominates Internet usage. I was recently forcibly reminded that the core operating systems for Internet remain unchanged, even though the latest browsers may hide this from the user.

Using my Netscape 2.02E browser on a page optimised for the latest browsers, my download attempt failed and I was forced to use my dedicated FTP client[6] instead. This had gathered dust all year and not been used since upgrading my operating system to OS/2 Merlin.

Having recalled how to drive this client, FTP worked faultlessly, performing the failed transfer much more quickly than is normal when using the Web browser file transfer. Originally I used this FTP client for all file transfers, but stopped when many sites only supported browser transfer, and no longer supported the use of an FTP client.

Simulation and design software

The *Micro-Cap V* simulator,[7] version 2.0, available from June, now includes the BSIM3 Version 3 mosfet models. Presentation of simulation results has been enhanced, now having 'Performance' and '3D' plotting features.

Performance plotting is a method for displaying specific data from an analysis that contains multiple runs, on to an X-Y plot. The example shown is of rise time change with resistance change, for this simple *LCR* circuit. The 3D plotting allows simultaneous display of three analysis variables. This example also shows the affects of stepping

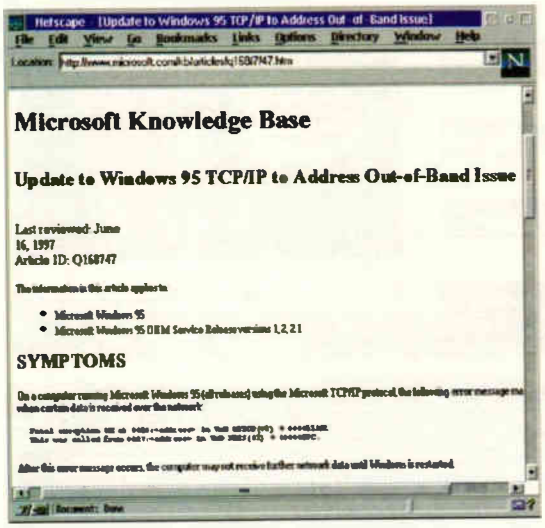


Fig. 2. Microsoft's official response to the Windows 95 'WinNuke' problem. To become 'bombproof', download this article and the file 'Vtcpupd.exe'.

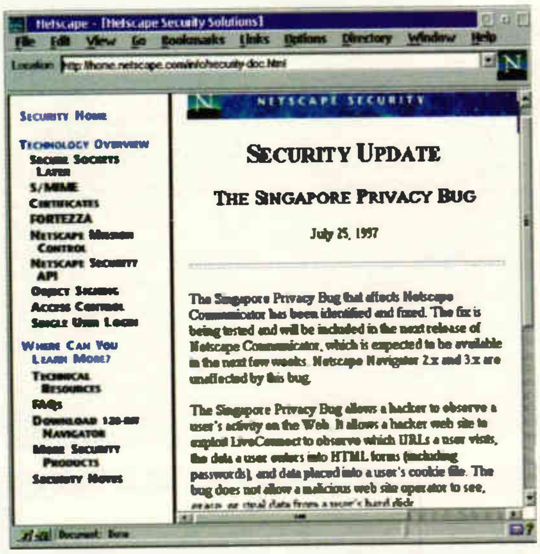


Fig. 3. Netscape's 'Singapore' security problem. Since no 'intrusions' have been reported, is claimed to be a minor problem.

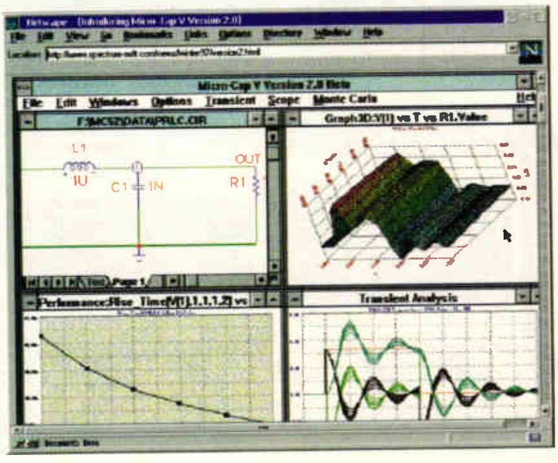


Fig. 4. Micro-Cap V version two, now with BSIM3 version 3. Update to this popular 'alternative' Spice simulator.

Fig. 5. Flomerics Inc. suppliers of the Flowtherm thermal simulator. Hosts a large reference library and the 'Cooling' magazine.

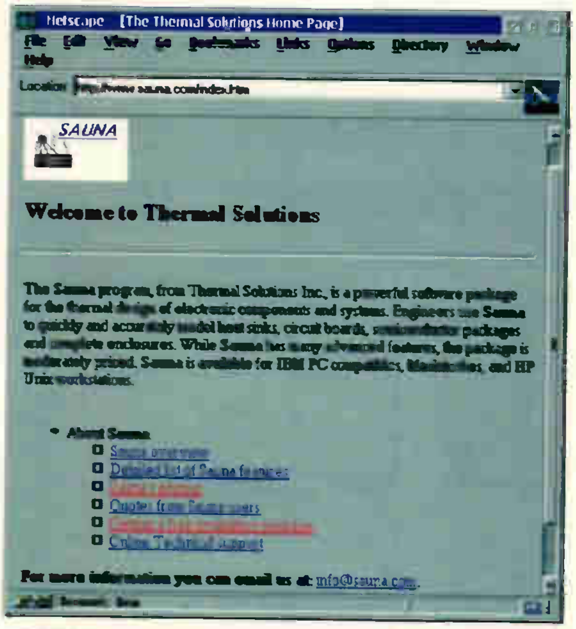


Fig. 6. 'Sauna' by name, this version aims to cool your products. This 3D simulator provides easy linking of components to printed board layouts. Uses the traditional thermal resistance grid of resistors method.

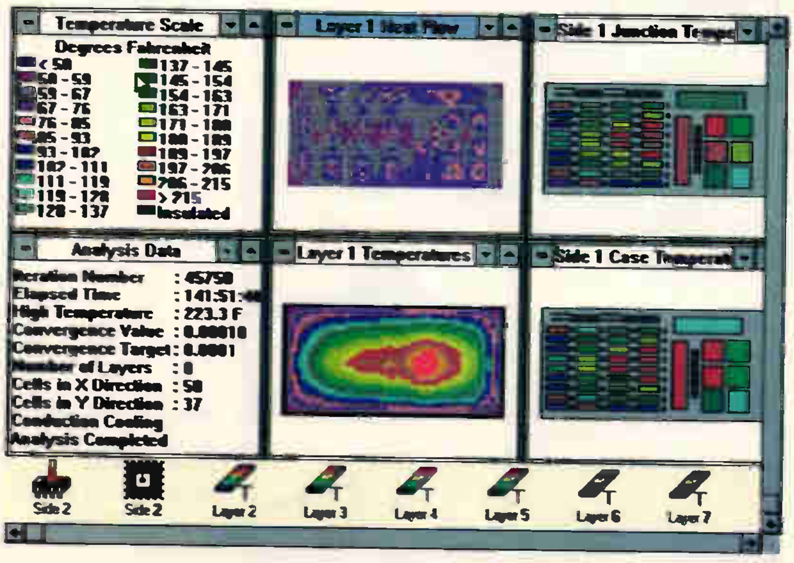


Fig. 7. Demonstration pcb example from Therm.zip. Explore this multilayer board example for yourself.

the resistor value, and plots node 1 Voltage v time v resistance, **Fig. 4**.

All electronic circuits generate heat, which must be dissipated to ensure satisfactory service. While some estimates using thermal resistances can be made, a final solution might require 'cut and try' methods.

Since a close analogy can be drawn between heat dissipation and electrical circuit design, some limited analysis[8] is also possible using a Spice simulator. Larger problems however are generally tackled using finite-element, finite difference or computational-fluid-dynamics methods, for which specialist thermal simulation packages have been developed.

Flomerics[9] has been working with the 'Delphi' and 'Seed' European projects to develop methods to handle both board and device-level modelling. Their *Flowtherm* simulator merges the traditional thermal resistance network with fluid-dynamics methods, offering benefit as the fluid-dynamics analysis grid becomes less detailed. The company's page hosts a very large library of thermal reference papers, some available by download, together with their *Cooling Magazine*, both well worth studying, **Fig. 5**.

Certain lower cost – even free – evaluation programs and support data are available for designers wishing to explore these techniques, before committing to one or other method. Thermal Engineering Associates[10] hosts a Web page with conference details and links to many sites. One especially useful introduction is the Data Handbook offered free on request,[11] linked from their 'freebies' heading.

An advocate of the thermal resistance network method, Thermal Solutions[12] offer its 'Sauna' 3D program which runs on pcs, Macs and HP Unix workstations. Sauna claims special benefits for its 'Gray radiation' analysis method used to analyse modules within boxes. Sauna v3.0 now has the capability to simulate transient and duty cycle thermal simulations. A working evaluation version is available on request, **Fig. 6**.

Most colleges have thermal simulation projects, many now being targeted specially towards electronics needs. As part of its Design Automation Group, the Electrothermal Circuit Simulation project[13] is working to develop accurate thermal macromodels for dissipative power devices. These are based on the devices actual physical geometry, with particular emphasis on power Hexfets and thermal feedback simulation.

2D thermal modelling

One possibly simpler introduction to the subject is by way of a 2D thermal modelling package. I mentioned the excellent *Quickfield* finite element simulator in the May issue. Optimised for electronic simulations, *Quickfield* is particularly useful when evaluating an individual component's thermal characteristics. A working evaluation version[14] can be downloaded.

Targeted more towards board-level analysis, 'Therm.Zip' is an older shareware package, quite suitable for exploring the basic concepts of thermal analysis. Therm is written by Lakeview Software, and being only 390k when zipped, can quickly be downloaded from Winsite[15] or most Winsite mirrors, **Fig. 7**. ■

Radio reflections

St Petersburg is the home of one of the world's oldest technical museums. And within its walls is a wealth of early British wireless apparatus – as Khatskel Ioffe reveals.

The Alexander S. Popov Central Museum of Communications in St. Petersburg was founded in 1872 as a telegraph museum. It was a closed establishment affiliated to the Department of Telegraphs, Ministry of the Interior.

In 1884, a postal department was set up and the establishment started to be referred to as the Post-and-Telegraph museum. Wireless equipment has been collected by the museum since 1910.

Eventually, the museum became a research and educational centre for the history of all kinds of communication – postal communication including postage stamps, telegraphy, telephony, radio and television.

The museum has a collection of British equipment dating from the earliest stage of the development of wireless.

We have in our collection Marconi coherers in their original packing cases. The coherer played a decisive role in the advent of wireless telegraphy by acting as a radio wave detector in the receiver. Many of you will recall that the coherer exhibits a dramatic increase in conduction under the effect of incident electromagnetic waves.

The phenomenon was described by the French physicist Edouard Branly (1846-1940) in 1890. It was extensively studied by many scientists, in particular Sir Oliver Lodge (1851-1940). It was he who coined the term coherer to describe an insulating tube with two electrodes and containing metal filings between them. The coherers of Guglielmo Marconi (1874-1937), starting from his first patent of 1896, had an evacuated glass tube.

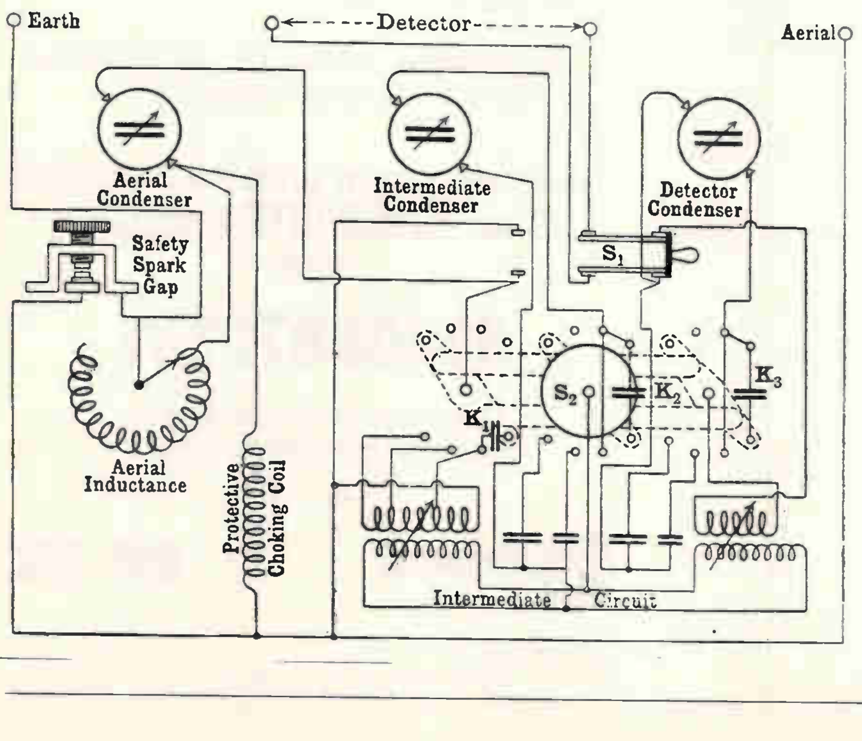


Fig. 1. Early circuit diagram of multiple tuner shows that the system comprises three sections – the antenna circuit, the intermediate circuit and the detector.

We have eight coherers of Marconi's Wireless Company. Each is housed in a metal case to protect it against electromagnetic radiation. We believe they date from about 1900.

These coherers comprise an evacuated glass tube, 60mm long and 4mm in diameter. Sealed into the tube are two platinum wires, each terminated in an obliquely cut cylindrical silver electrode ground tightly to the inner wall of the tube.

A small quantity of filings comprising 96% nickel and 4% silver was poured into the wedge-like inter-electrode gap, with a minimum width of 0.5mm. For protection, the coherer is secured to an ivory stick, with which it was mounted on the receiver.

Marconi's magnetic detector

In 1902 the Italian government placed the warship 'Carlo-Alberto' at Marconi's disposal. With King Emmanuel III on board, she came to Kronstadt on an island in the Baltic Sea, 29km westwards of St. Petersburg. Marconi established long-range communication with, among others, a Poldhu station on the southern coast of England.

In his receiver installed on the warship, he employed a magnetic detector whose principle had been suggested by the physicist Ernest Rutherford (1871-1937) in 1896, **Photo. 2**. This detector depends on the ability of electrical oscillations to reduce magnetic hysteresis.

In Marconi's magnetic detector, two grooved wooden wheels are brought into uniform rotation by a clock mechanism, or spring motor. An endless flexible band, comprising several dozens of thin iron wires, is stretched between the wheels. The band moves in the inside of a small glass tube with a speed of around 8cm/s under the poles of two horseshoe permanent magnets whose like poles are brought close together.

Slipped over the tube, one over the other, are two coils of insulated copper wire. The primary coil is placed into the aerial-earth circuit and the secondary is connected to a telephone receiver.

Travelling under the magnet poles, the iron wire band is magnetised first in one direction and then in the opposite direction. A reversal of the sense of magnetisation occurs under the inside like poles, not exactly at the instant the band passes under them but with a delay due to hysteresis. The rf field created by the primary coil during the advent of a signal momentarily reduces the hysteresis, causing a shock reversal of magnetisation in the band.

from Russia

The rapid displacement of the magnetic lines of force produces an instantaneous current in the secondary, bringing about a sound in the telephone receiver. This re-occurs with a frequency equal to the interruption frequency of the interrupter of a spark transmitter as long as the telegraph key is in the down position. The telephone of the receiving station produces a sound during all this time.

The magnetic detector responds to the peak value of an electromagnetic wave. As a result, this detector was suitable for stations with heavily damped oscillations. It was not until the mid 1990s that it was replaced by the contact detector.

We have two examples of Marconi's magnetic detectors which we believe were manufactured between 1908 and 1910.

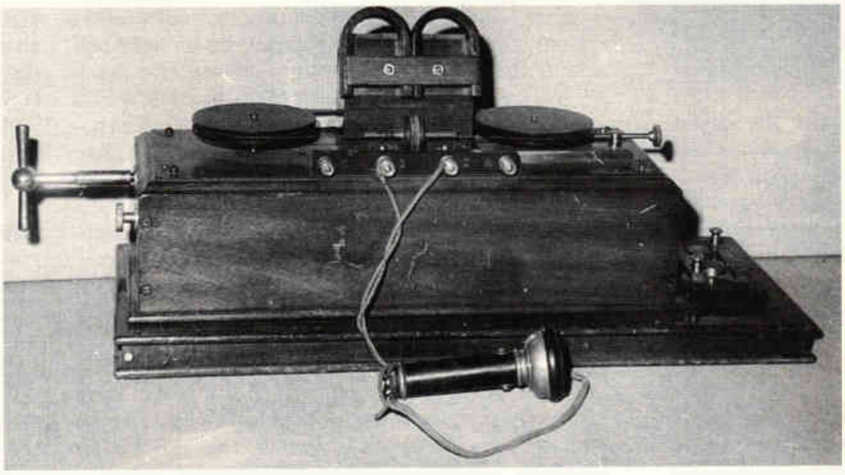


Photo 2. Marconi's magnetic detector, circa 1909, relied on the fact that electrical oscillations reduce magnetic hysteresis. The band around the wheels is iron wire, which passes between magnetic poles as the wheels rotate.

Tuning gear

Photo 3 shows a multiple tuner[1] – a unit for tuning a receiver, operated in conjunction with a magnetic detector. It was used in Marconi's coastal and ship wireless stations in the early 1900s.

Figure 1 shows the circuit diagram of the tuner as represented at that time[2]. The tuner is essentially made up of three separate circuits termed the aerial circuit, the intermediate circuit and the detector circuit. The variable coupling between the circuits is of the transformer type.

Switch S_1 brings the unit to rough or fine tuning. In the former case, the aerial circuit only is tuned. In the latter case, all the three circuits are tuned. Switch S_2 has four fixed positions for the four subranges in the intermediate and detector circuits. Smooth tuning is performed by variable capacitors.

Varying the detector coupling allows the intensity of the detector output signal to be changed and facilitates fine tuning to a possibly weaker signal.

This instrument marked a great step forward in the development of practical wireless through a substantial increase in selectivity.

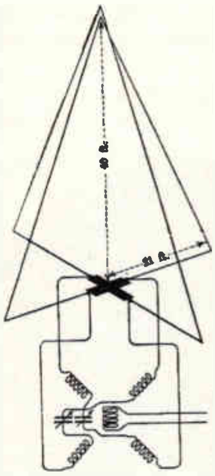


Fig. 2. Schematic of the Marconi–Bellini–Tosi direction finder. Two variable capacitors were used to allow coarse and fine tuning.

The earliest valve

Fleming's diode, **Photo 4**, is the earliest kind of vacuum valve[3]. It is a cylindrical glass envelope, about 50mm in diameter, filled with rarefied gas and provided with a Swan type base.

Located in the envelope are a carbon filament in the form of an arc – the cathode – and a nickel cylinder enclosing it – the anode or plate.

The anode is secured on a glass stem. Two wires run from the base through the stem to the cathode. The plate lead passes through the lateral surface of the envelope. When placed in a receiver circuit, the valve conducts in one direction only, from the heated filament to the cylinder (the electronic current), and thus serves as a detector.

Photo 3. Multiple tuner from circa 1909 was used to tune receivers relying on magnetic detectors.

Wireless orientation

Shown in **Photo 5** is the direction finder of the Marconi–Bellini–Tosi system. The Marconi Company invited all interested persons to send in relevant invention and patent and offered for them high rewards.

Among such patents purchased by the Marconi Company was the patent for the 'wireless compass' of the Italians Bellini and Tosi[4]. This was one of the most remarkable inventions in the first decade of practical wireless. In 1908-1909 it hit the headlines of all leading electrotechnical journals.

We believe that our example is a specimen of the first implementation of this invention. It dates from about 1910. Its circuit diagram[5] is shown in **Fig. 2**.

The apparatus was designed for ships to determine their location from the known coordinates of fixed wireless stations. The principle of directed transmission and reception of Bellini–Tosi is reduced to the use of a radiogoniometer. This device consists of two fixed field coils located at right angles to one another. The coils are connected respectively to two symmetrical frame antennas, perpendicular to one another, and to a moving 'search' coil pivoted inside the fixed coils.

The search coil connects to a receiver, and the reception strength will depend on the position of this coil. When its plane coincides with the direction of coming waves, the reception is strongest. Coupled with the pivot of the moving coil is an azimuth pointer.

Located on the top panel of the instrument are terminals for connecting the antennas and the receiver. There is also the goniometer knob which has a pointer over a circular scale of azimuths in degrees. The knob was for tuning the antenna to a station whose bearings are to be taken. Antenna switches on the top allowed the unit isolated while the local transmitter was being used or during a thunderstorm.

On the side wall of the instrument are the knobs of two capacitors for separate adjustment of the antennas, as well as the 'Billi condenser' for fine tuning. In the latter capacitor, two brass tubes slide over two other brass tubes along a hard-rubber bushing, the capacitor dielectric.

The goniometer is located behind a hinged front wall of the case. This is a hollow cylinder having two bases of an insulating material on different but close levels at both top and bottom.

Wound round two formers at right angles to one another are two coils of copper wire forming rectangular turns. Variable capacitors are placed in the middle of each coil. Set true with the axis of the cylinder is the search coil on its own former, the coil shaft being terminated in the knob for taking bearings.

The operator tuned into a station, then turned the goniometer knob to the left until minimum audibility was attained. Next, the the knob was turned in the opposite direction and a similar position found. The position between these points corresponded to the direction to the station. For higher accuracy, the operator strived for lowering the intensity of received signals using the detector coupling control on the receiver.

Photo 4. The simplest type of valve – Fleming's diode – from circa 1907. It has a carbon filament forming the cathode and a nickel anode in the form of a cylinder enclosing it.

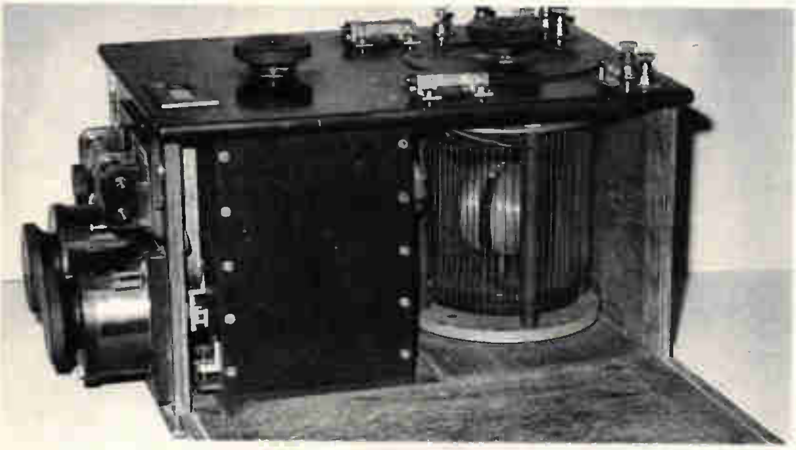


Photo 5. Direction finder of the Marconi–Bellini–Tosi system from circa 1910 helped ships find their position by indicating the directions of known transmitters.

Direction-finding receiver?

Photo 6 shows the intermediate and detector circuits of a receiver. I believe this unit was designed for use in conjunction with a Marconi–Bellini–Tosi direction finder and was built at the same time.

Using two terminals at its top, the intermediate circuit was connected to the corresponding terminals (R, R) on the direction finder to receive a signal from the search coil of the goniometer.

The intermediate circuit contains an induction coil and a variable capacitor – the 'intermediate tuning condenser' – which could be connected in parallel for long wave or in series for short wave using a switch.

Coupled with the coil of the intermediate circuit via a variable transformer is a detector circuit coil whose control is brought out to the front wall of the unit to allow variation of the detector coupling. The detector circuit is tuned using a Billi condenser.

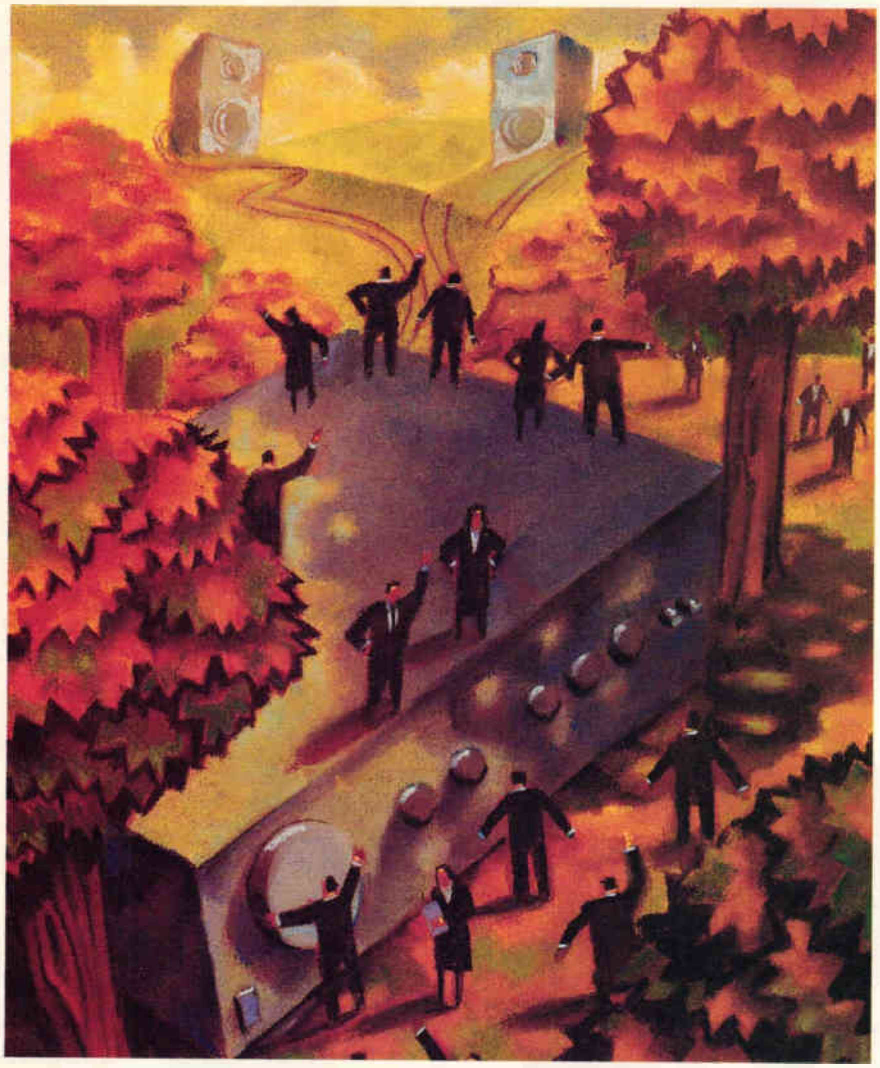
Fleming's diode or a contact detector could be selected as the detector using a switch. The filament of the diode was powered by an external battery through a rheostat. The regime of the contact detector was set by a potentiometer. ■

Photo 6. Intermediate and detector circuit unit for direction finder – probably – from circa 1910.

In a subsequent article, Khatskel outlines Marconi transmitters and receivers from around 1915 and describes the museum's version of Fleming's cymometer.

References

1. 'The Physical Society's Annual Exhibition of Apparatus', *The Electrician*, 1 Jan 1909, p. 468.
2. 'The Marconi System of Wireless Telegraphy', *The Electrician*, 3 May 1912, p. 134.
3. J.A. Fleming, Improvements in instruments for detecting and measuring alternating electric currents. Brit. Pat. No. 24850. Date of application 16 Nov 1904. Complete specification left 15 Aug 1905. Accepted 21 Sept 1905.
4. E. Bellini and A. Tosi, Luftleiteranordnung zur Ausstrahlung elektrischer Wellen in einer einzigen Richtung und zur Aufnahme elektrischer Wellen aus einer einzigen Richtung zwecks Bestimmung der örtlichen Lage der unbekannten Geberstation. D.R.P. Nr. 221242, 22.VIII.1908 (German Patent). The corresponding British Patent was not in my reach.
5. 'Directive Wireless Telegraphy', *The Electrician*, 27 Dec 1912, p. 587.



Doug Self looks at how loudspeaker impedance – which varies dramatically with frequency – interacts with amplifier output and speaker cable. This final article follows two previous related discussions by Douglas, one on the importance of the amplifier's output stage and the other on loudspeaker cable effects.

Speaker impedance matters

The most important feature of electromechanical loudspeakers, from the amplifier/cable point of view, is their non-uniform impedance with frequency. This has several consequences.

The resistance of the cable from the power amplifier is non-zero, giving rise to response variations in the voltage at the loudspeaker end as speaker impedance varies with frequency. There may be important effects on the amplifier's distortion performance. Impedance dips may activate overload protection with cacophonous results.

While an understanding of all the factors that combine to make the impedance curves of multi-way loudspeaker complex and variable is not strictly necessary for amplifier design, it does give useful insight. For many years, DIN and ISO standards specified that the impedance over the audio band should not vary by more than ±20% of the average value, so that an 8Ω speaker should not fall below 6.4Ω, but this appears never to have been taken seriously.

When the electrical model of a single-unit load replaces the standard 8Ω resistive load, something remarkable happens; high-frequency distortion virtually disappears, as shown in **Fig. 1**. This was obtained by wiring a 0.33mH inductor in series with the 8Ω resistive load.

A Blameless amplifier driving 8Ωs or more exhibits only crossover distortion, increasing with frequency as the negative feedback factor falls, and the magnitude of this depends on the current drawn from the output stage; with an inductive load this current falls as frequency rises.

Single speaker units

While single-driver designs are now rare in hi-fi applications, many public-address, disco and sound reinforcement applications still use single, so-called full-range, drive units. The impedance variations of single speaker units can be modelled with fair accuracy.

Figure 2 shows an electrical model of a single speaker unit. Resistance and inductance of the voice coil are represented by R_c and L_c respectively. Components L_r and C_r model the

Misconceptions

In many audio fields the conventional wisdom is wrong. The amplifier-cable-speaker system is no exception, and the misconceptions unearthed in the writing of this set of articles are summarised below:

- Amplifier output impedance has no effect on loudspeaker damping. 'Damping factor' is essentially a meaningless parameter.
- Amplifier output impedance is often not determined by the amplifier itself, but by the output inductor winding resistance.
- Output ringing in square-wave testing with capacitive loads is not from the amplifier and gives no information on its stability.
- The amount of ringing observed is affected more by the test square-wave rise-time than any circuit parameter.
- The shunt capacitance of any normal speaker cable is trivially small, and can have no significant effect on frequency response or anything else.
- Reactive loads are 'more difficult to drive' only in the sense that they put more electrical stress on the output stage than a resistance. They do not cause instability if the usual precautions are correctly implemented.

electromechanical resonance of the cone mass with the suspension compliance and air-spring of the enclosure; R_r controls damping.

Components R_r, L_r and C_r have no physical existence, but give the same impedance characteristics as the real resonance, determined by mechanical inertia and compliance. At 0.1mH, voice-coil inductance here is fairly low, which might apply to a modest 8in unit; 18in bass units go to 1mH or more.

The input impedance magnitude this network presents to an amplifier is shown in **Plot A1**. The peak at 70Hz is due to the cone resonance; without a sealed enclosure, the cone restoring force is lower and the free-air resonance is at a lower frequency.

The rising impedance above 1kHz is due to the voice-coil inductance L_c. This part of the model is not very accurate, as in real life eddy-current losses etc make the voice-coil inductance more of a 'semi-inductor' with approximately 45° phase shift[1,2,3]. This is more important for crossover design rather than for assessing amplifier loading at high frequencies.

The phase angle varies from +35 to –35°, passing through zero at the resonance point. Speaker loads do not show pure inductance or capacitance, which would give a 90° phase angle.

This model can be directly related to the electro-mechanical and mechanical-acoustic domains by means of conceptual 'transformers' that handle the scaling of the units of measurement.

Multi-way speaker systems

The impedance curve of a typical multi-way hi-fi loudspeaker is much more complex, with multiple humps and

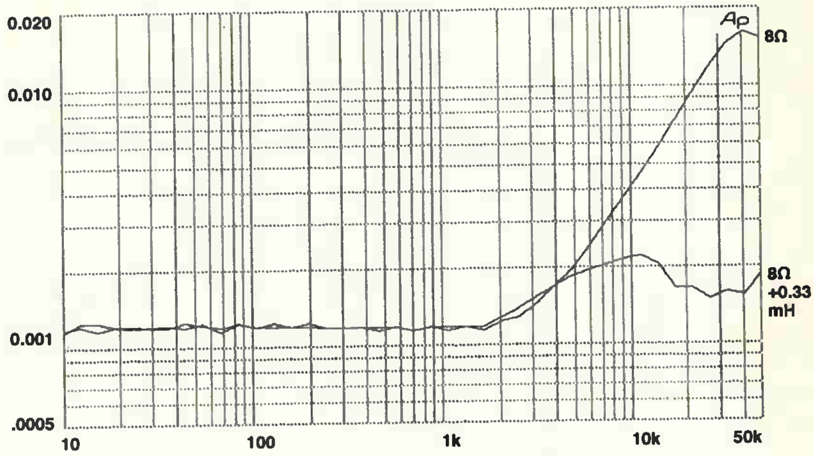


Fig. 1. Crossover distortion is dramatically cut if the load has an inductive component that reduces current at high frequencies.

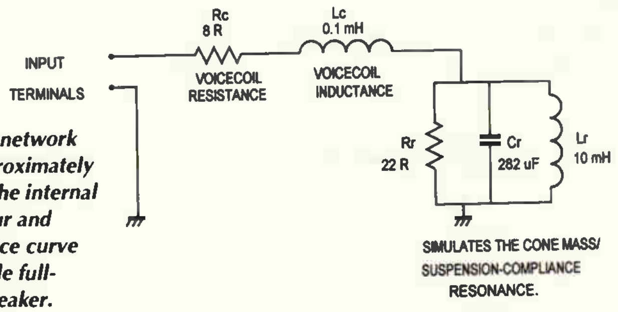


Fig. 2. A network that approximately models the internal behaviour and impedance curve of a single full-range speaker.

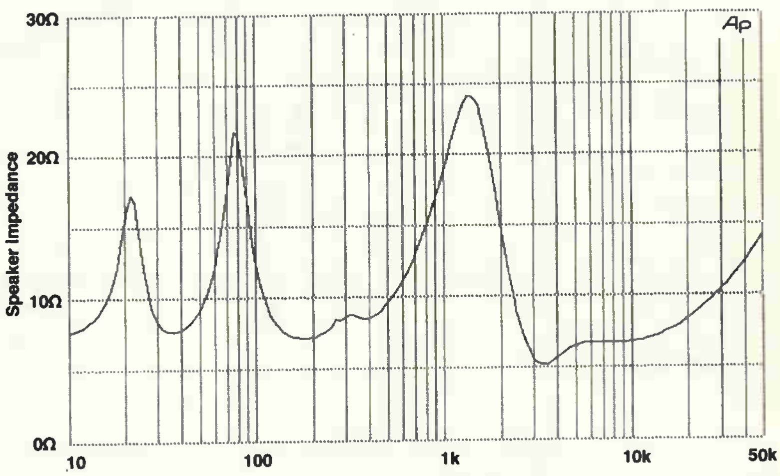


Fig. 3. Measured impedance curve for Kef Cara SP3076 two-way speaker with auxiliary bass radiator.

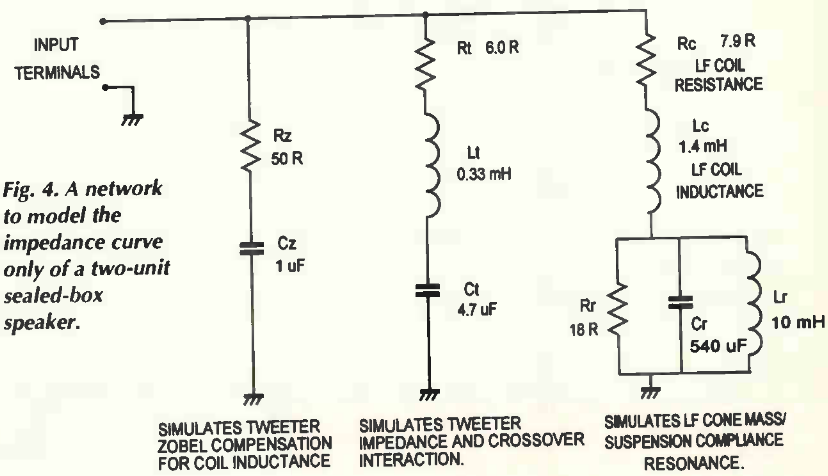
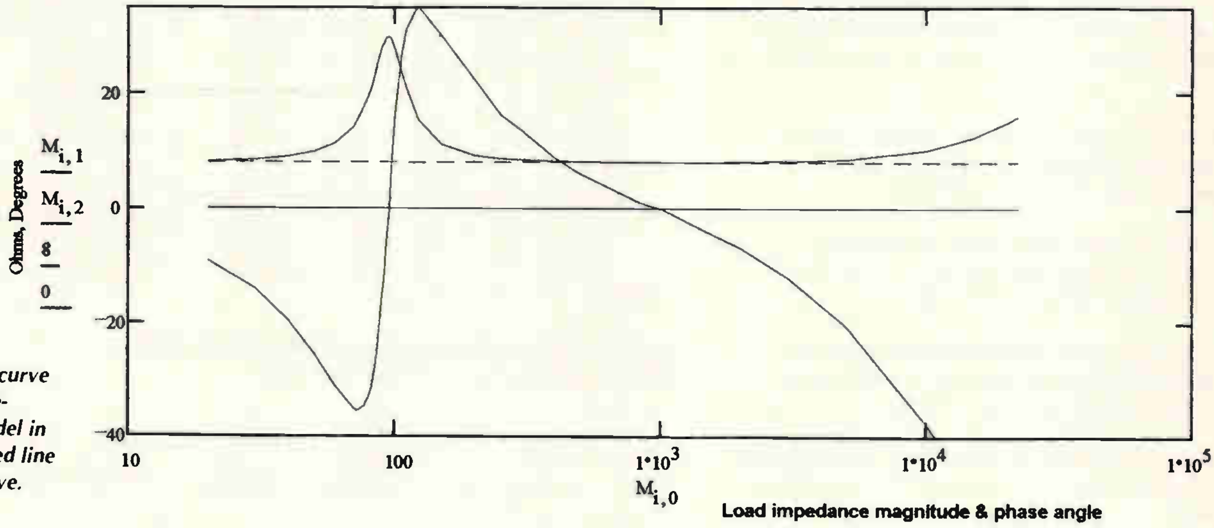
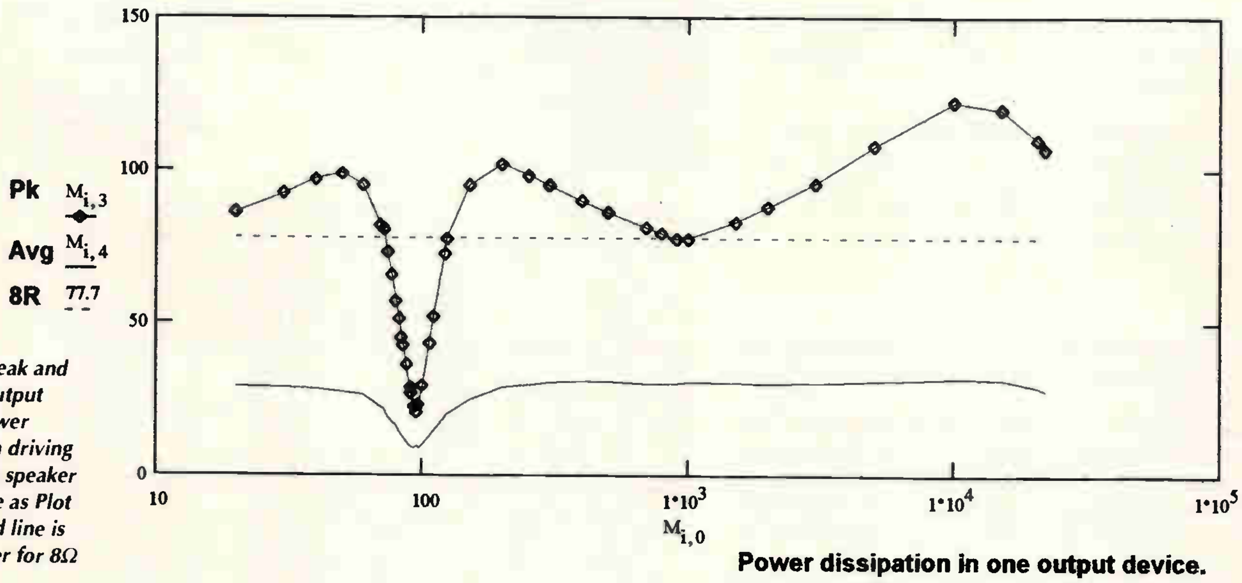


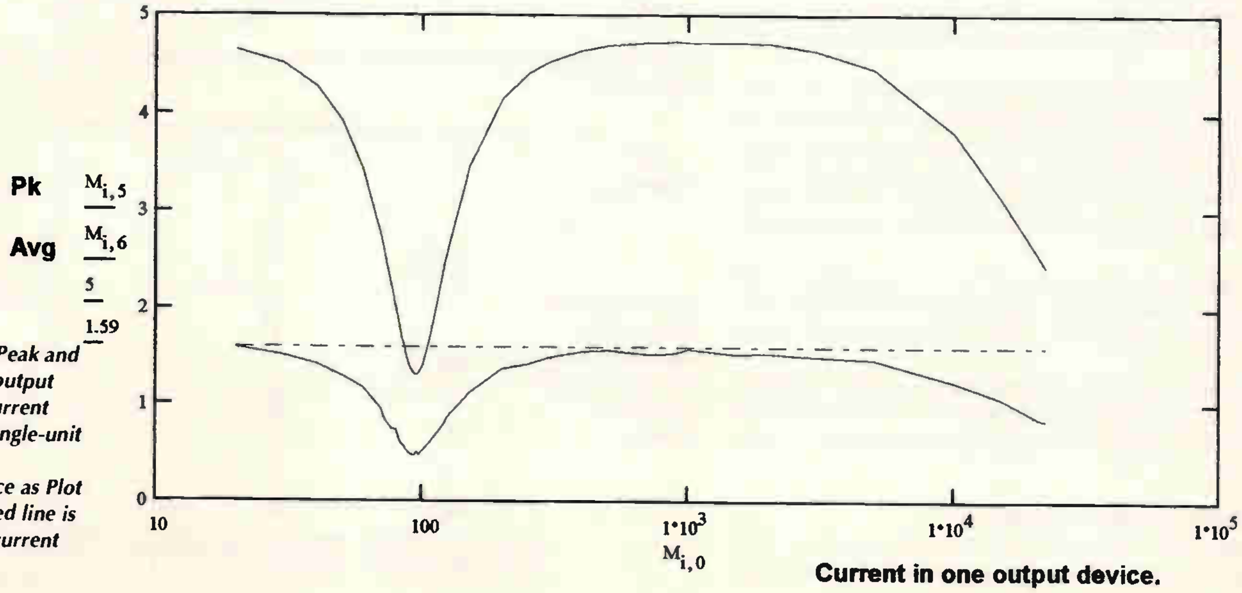
Fig. 4. A network to model the impedance curve only of a two-unit sealed-box speaker.



Plot A1. Impedance curve of the single-speaker model in Fig. 2. Dotted line is 8Ω resistive.



Plot A2. Peak and average output device power dissipation driving single-unit speaker impedance as Plot A1. Dotted line is peak power for 8Ω resistive.



Plot A3. Peak and average output device current driving single-unit speaker impedance as Plot A1. Dotted line is average current into 8Ω.

dips representing various features of the speaker.

At the low-frequency end these are fairly predictable; the bass unit resonance will give a significant hump in low-frequency impedance, with associated phase changes. Reflex, i.e. ported, enclosure designs have a characteristic double-hump in the low-frequency, with the middle dip corresponding to the port tuning. The impedance in this dip does not normally fall below the nominal impedance of the speaker. A typical plot of impedance against frequency is shown in **Fig. 3** for the Kef *Cara* – a two-unit system with an auxiliary bass radiator.

The high-frequency region of speaker impedances is much more variable, depending in a complicated fashion on the drive units and their interaction with the crossover components. There are significant non-linearities varying across the audio band.

The 'semi-inductor' behaviour of the voice-coil inductance also becomes important. This means that simulating the electro-acoustic equivalent of a multi-way loudspeaker accurately enough to predict its impedance curve is a major enterprise. It is probably best tackled with specialised loudspeaker simulation software rather than a general electrical simulator such as Spice. An excellent example is Abakab[4] produced by J W Panzer.

Nominal speaker impedance specifications must be treated with a good deal of scepticism; they do not represent the minimum impedance and probably not the 20-20kHz average either. One example[5] has three impedance minima, of 4.3, 3.9, and a frightening 2.7Ω, yet still claims to be an 8Ω speaker. I am aware of one British "8Ω" design that falls to 1.6Ω at low frequencies.

For amplifier design purposes, there is no reason why an impedance model should express the physical reality of a speaker in terms of electro-acoustic equivalents; it is sufficient to model the impedance curve itself.

Figure 4 shows an example of this approach, while **Plot B1** gives the impedance curve. This assumes a simple crossover network that includes Zobel compensation R_Z. C_Z for rising tweeter coil impedance, and is based on a network proposed by Ken Kantnor in *Stereophile*.[6]

Amplifier reviews in *Stereophile* are currently undertaken with this load as well as pure resistances. The major effect that comes to light is frequency response variation due to excessive amplifier output impedance. Deviations of 1dB have occurred even with solid-state designs; these have presumably had resistors introduced into the output circuit in pursuit of unguessable Subjectivist ideals.

Crossover design

Design of the crossover probably has the most important influence on speaker impedance. A two-way speaker system is assumed at this point, as the extra complications of a three-way crossover are serious.

The design process passes through the following three stages. First, choose the alignment of the crossover system. In other words, select the order of filters, filter Q, and offset of filter frequencies that will give the best summation of the acoustic output of each speaker unit.

In this respect, it is common to talk of Butterworth or Bessel alignments, and so on. This results in the simplistic circuit of **Fig. 5**, where the speaker impedances are assumed to be purely resistances.

The second design step is to correct for non-resistive speaker impedances. Driver impedances are far from pure resistances, so filter responses that assume they are will be wrong due to mistermination.

As **Fig. 6** shows, the main resonance of the hf unit can manifest itself as an unwanted peak in low-frequency output just below the crossover frequency.[7] This can be corrected by cancelling the impedance rise at resonance by placing a

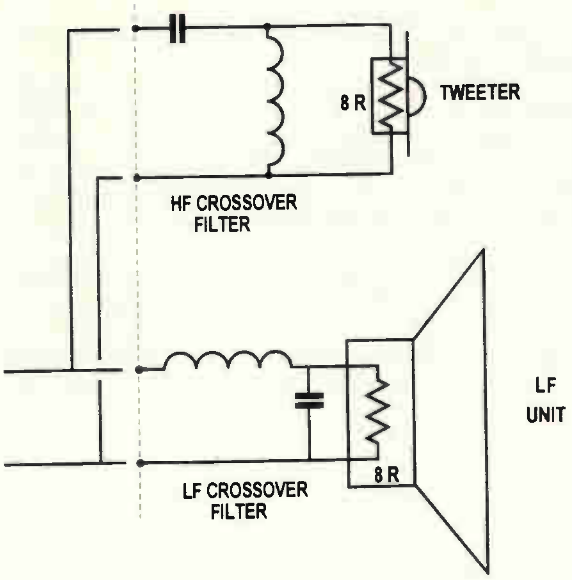


Fig. 5. Simple two-way crossover unit. This would only work properly if the impedance of each speaker unit was purely resistive; it is not.

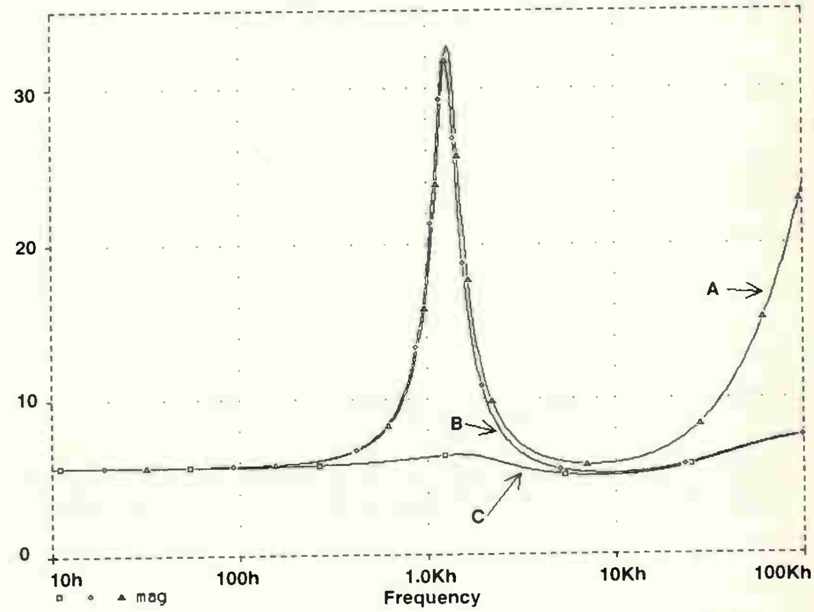


Fig. 6. A is a typical tweeter impedance curve; resonance is at 1.1kHz. The high-frequency rise is at a higher frequency than Fig. 2 as voice-coil inductance is much lower at 37µH. B is after correction for voice-coil inductance, while C adds resonance correction.

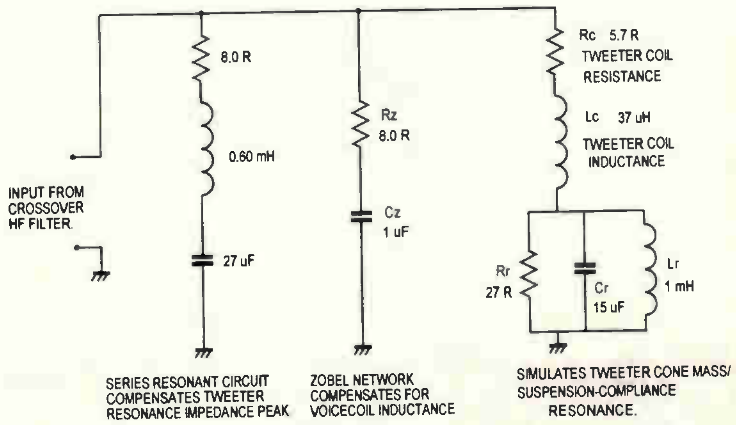


Fig. 7. Tweeter impedance compensation circuitry. The Zobel network R_Z, C_Z compensates for voice-coil inductance; series-resonant circuit RLC corrects impedance at resonance.

series-resonant *RLC* network in parallel with the high-frequency unit, **Fig. 7**.

For amplifier designers, an unhappy side-effect is that the impedance seen by the amplifier is reduced. This correction is not required for the low-frequency drive unit, as the bass resonance and its associated low-frequency roll-off are a fundamental part of the speaker design, and there is no electrical crossover filter at the bottom of the low-frequency range to be misterminated.

It is sometimes also necessary to correct for rising driver impedance due to voice-coil inductance, by shunting it with a Zobel network R_z, C_z, as in Fig. 7; Fig. 6 shows this is very effective. Unfortunately this also reduces the impedance seen by the amplifier, at high frequencies – where it this is most undesirable. The important point is that both corrections reduce the average impedance seen by the amplifier.

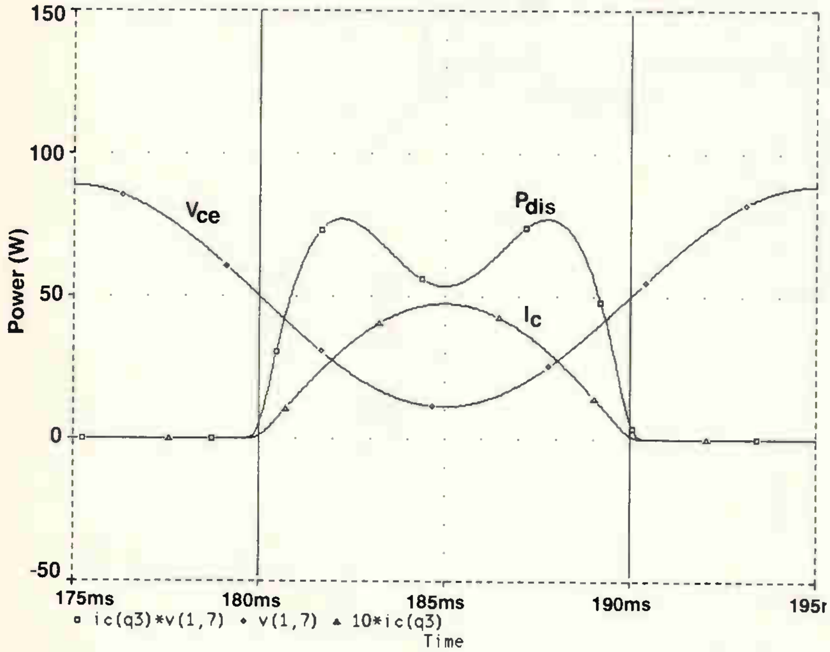


Fig. 8. Instantaneous V_{ce}, I_c, and P_{diss} in an output transistor driving 8Ω to 40V peak at 50Hz, from ±50V rails. Device dissipation peaks twice at 77W in each half-cycle.

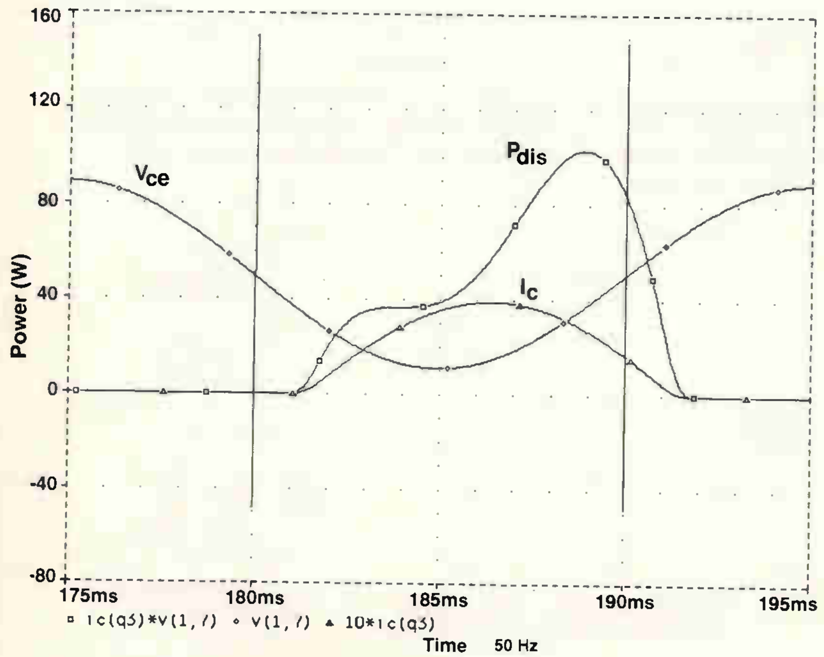


Fig. 9. As Fig. 13, but driving 50Hz into the single-speaker load. At this frequency the load is partly inductive so current lags voltage and the instantaneous power curve is asymmetrical, peaking higher at 110W towards the end of the half-cycle.

An excellent example both of the response irregularities caused by using crossover filters designed for purely resistive loads when the real load is a speaker impedance, and its correction by the addition of a Zobel network is given in ref. 3. Another example is ref. 7.

Equalising the drive units – if necessary – is the third step in the crossover design process. If the response of a drive unit is not as flat as required, equalisation networks are used. A gentle peak in the response is usually nulled by a parallel-resonant circuit in series with the drive unit. This increases the average impedance. If however the response of a drive unit needs to be increased over a given frequency region, there is likely to be an impedance dip in this area.[8]

While this gives insight into why impedance curves can be complex, an amplifier must still be capable of driving the impedance at the deepest dip in the curve. It is often assumed that this will be no lower than 4Ω for a nominal 8Ω multiway speaker, and this is in general true for what might be called mainstream designs. It is less likely to be the case for 'high-end' equipment, which may have precipitous drops to 2Ω or even less.

Loading may be much lighter over the rest of the audio band, but this does not allow any meaningful economies in the design of the amplifier output stage, heatsinking, or power supply. This is regrettable as these are by far the most expensive parts of a power amplifier.

Loads for amplifier testing

Amplifiers are almost universally designed and tested running into purely resistive loads. But they spend their working lives driving loudspeakers containing important reactive components and electromechanical resonances.

At first sight this is a nonsensical situation; however, testing into resistances is neither naïve nor an attempt to avoid the issue of real loads; there is little alternative.

Loudspeakers vary greatly in their design and construction, and this is reflected in variations in the impedance they present to the amplifier on test. It would be necessary to specify a 'standard speaker' for the results from different amplifiers to be comparable. Secondly, loudspeakers have a notable tendency to turn electricity into sound, and the sinewave testing of a 200W amplifier would be a demanding experience for all those in earshot – soundproof chambers are neither easy nor cheap to construct.

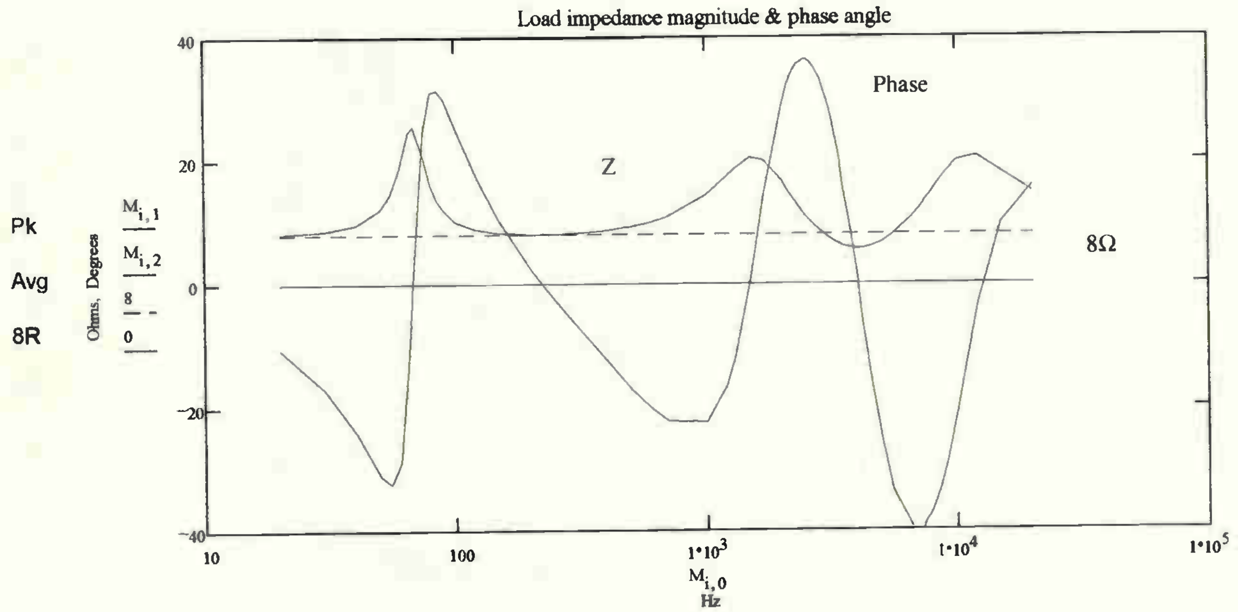
Thirdly, such a standard test speaker would have to be capable of enormous power-handling if it were to be able to sustain long-term testing at high power; loudspeakers are always rated with the peak/average ratio of speech and music firmly in mind, and the lower signal levels at high frequencies are also exploited when choosing tweeter power ratings.

A final objection is that loudspeakers are not noted for perfect linearity, especially at the low-frequency end. If the amplifier does not have a very low output impedance this speaker non-linearity may confuse the measurement of distortion. Amplifier testing would demand a completely different sort of loudspeaker from those actually used for listening to music; the market for it would be very, very small, so it would be expensive.

Now to those difficult loads

There is an assumption that any reactive load is more difficult for an amplifier than a purely resistive one; it is devoutly to be wished that people would say what they mean by 'difficult''. It could mean that stability margins are reduced or that the stresses on the output devices are increased.

Both cases may be true, but I have a deep suspicion that this is based on anthropomorphic thinking. It is easy to assume that if a signal is more complex for a human brain to contemplate, it is harder for an amplifier to handle. This is



Plot B1. Impedance curve of model of the two-unit speaker in Fig. 4. Dotted line is 8Ω resistive.

not true; it is not necessary to understand the laws of physics to obey them. Everything has to anyway.

When amplifiers go unstable it is always at ultrasonic frequencies – assuming you are not contemplating some sort of historical curiosity with ac coupling in the forward path – and never in the middle of the audio band. This is so, despite the fact that many speakers have major convulsions in their impedance curves around this region.

Reactive loading can – and does – imperil stability at high frequencies unless the usual precautions are taken, but does not cause oscillation or ringing mid-band.

Output stage stresses are certainly increased by reactive loads. If you are driving a single speaker unit, with the impedance curve of Plot A1, impedance magnitude never falls below the 8Ω nominal, and is substantially greater in some regions; this suggests overall amplifier power dissipation would be less than for an 8Ω resistive load.

Unfortunately, this is far from the truth. When pure resistance is driven, the voltage across an output transistor falls as the current through it increases, and they never reach a maximum at the same time.

Figure 8 shows a Class-B amplifier with an 8Ω resistive load. Instantaneous power is the product of instantaneous current and voltage drop, and has a characteristic two-horned shape, peaking twice at 77W during the conducting half-cycle.

When a typical single-speaker load is driven at 50Hz, the impedance is resistive-inductive, at 8.12+3.9jΩ. This means that the current phase-lags, altering the instantaneous product of voltage and power to that shown in **Fig. 9**.

Over the Class-B half-cycle, the average dissipation is slightly reduced, but the peak instantaneous power increases by 30% due to the voltage/current phase-shift. This could have serious results on amplifier reliability if not allowed for.

Note that this impedance is only equivalent to 8.5Ω in series with 10.8mH at 50Hz. Driving this replacement load at any other frequency, or with a non-sinusoid, will give wholly wrong results. Not every writer on this topic appears to appreciate this.

In the same way, if the single-speaker load is driven at 200Hz, on the other side of resonance, the impedance is resistive-capacitive at 8.4–3.4jΩ, and the current leads. This gives the same result as Fig. 9, except that the peak power now occurs in the first part of the half-cycle. The equivalent load at 200 Hz only is 10.8Ω shunted with 35μF.

When designing output stages, there are four quantities to compare with output device ratings. These are peak current, average current, peak power and average power.

I simulated a conventional emitter-follower output stage driving the single-speaker load with a 40V peak sinewave, with ±50V rails. Separate transient simulations over many cycles were done for 42 spot frequencies 20Hz to 20kHz, and the peak and average quantities recorded and plotted.

Many cycles must be simulated as the bass resonance in the impedance model takes time to reach steady-state when a sinewave is suddenly applied; again, not everyone writing on this topic appears to appreciate this.

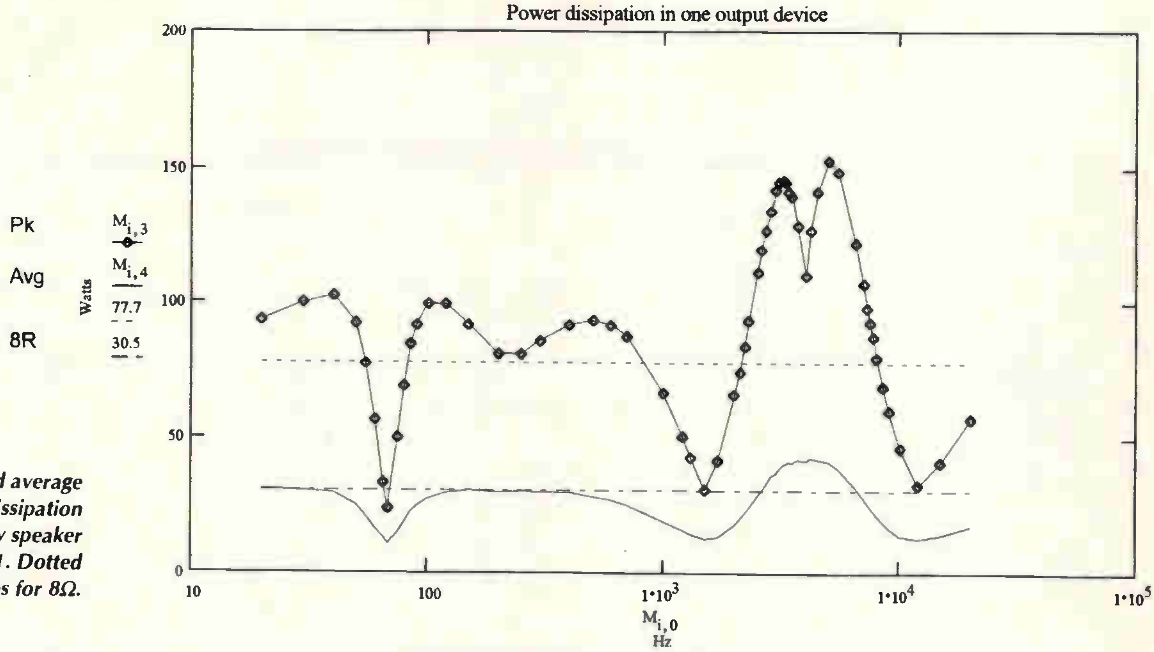
The steady sinewaves used here are a practical approach to simulation and testing rather than a good approximation to music. Arbitrary non-cyclic transients can be investigated by the same method, but the number of possibilities is infinite. It is also necessary to be careful with the initial conditions.

Plots A1, A2 and **A3** are the distilled result of a daunting amount of simulation. Plot A2 shows that the gentle foothills of the impedance peak at bass resonance actually increase the peak instantaneous power stress on the output devices – despite the reduced current drawn. Peak dissipation is only reduced below that for an 8Ω resistor, shown dotted, around the actual resonance peak, where it plummets by 75%.

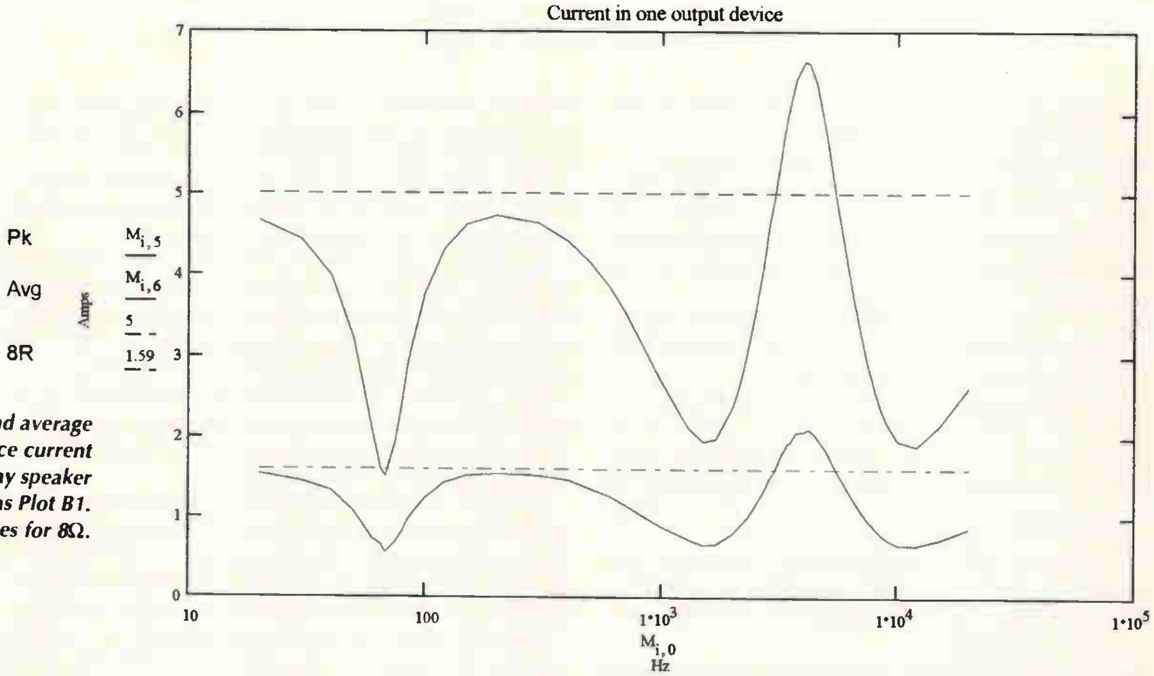
Likewise, the rise in impedance at the high-frequency end, where voice-coil inductance is significant, causes an even more serious rise in dissipation peaks to 50% more than the resistive case. The conclusion must be that for peak power, the phase angle is much more important than the impedance magnitude.

The effects on average power dissipation, and on the peak and average device current in Plot A3, are more benign. For this kind of load network, all three quantities are reduced when impedance increases, despite the phase shifts.

As far as I am aware this diagram is produced here for the first time. It shows at a glance that the most dangerous areas for the amplifier are the sides of the resonance hump where phase-shift is greatest. The critical quantities for semicon-



Plot B2. Peak and average output device power dissipation driving two-way speaker impedance as Plot B1. Dotted lines for 8Ω.



Plot B3. Peak and average output device current driving two-way speaker impedance as Plot B1. Dotted lines for 8Ω.

ductor safety are the peak instantaneous values; for heatsink design average power is what counts, while for the power supply average current is the significant quantity.

Plot B1 shows the impedance plot for the simulated two-way speaker load at 59 spot frequencies. The curve is more complex and shows dips below the nominal impedance rather than peaks above, typical of multi-speaker designs. It is no surprise that these dips give the greatest output device stress as they combine increased current demand with phase-shifts that increase peak instantaneous dissipation.

Once again in **Plot B2** the bass resonance causes increased peak dissipation due to phase shift; the other quantities are reduced. In the high-frequency region there is an impedance dip at 6kHz which nearly doubles peak power on its lower slopes. The bottom of the dip itself sharply reduces peak power where the phase angle passes through zero, giving the notch effect.

Average power, and peak and average current, **Plot B3**, are also increased by the impedance dip, but to a much more modest extent. Peak power looks like the quantity to watch.

Power device ratings often allow the power and second-breakdown limits – and sometimes the bond-wire current limit – to be exceeded for brief periods. If you attempt to exploit these areas in an audio application, you are living dangerously, as the longest excursion specified is usually 5ms, and a half-cycle at 20Hz lasts 25ms.

Output stages are often designed by drawing load-lines on diagrams of the safe operating area of the device. An example is given in **Fig. 10**, where ABCD defines the limits of the safe operating area.

Line XY is for an 8Ω resistive load, and line ZY is for 4Ω. Reactive loads turn these lines into ellipses that are more likely to intersect the power-limit BC. The width and orientation of such ellipses are set by the amount of load reactance, and this is another independent variable.

To simplify the diagram I have just taken the 4Ω-plus-reactance case for all possible reactances, and shown the envelope made up of all the closest approaches to the safe operating area limit.[9] This is another straight line, drawn from the maximum current point Z to a point W at twice the rail voltage. It represents 4Ω plus any amount of inductive or capacitive reactance. Note that this is not the same as 'all loads with a modulus of 4Ω' which is much more demanding.

In the case of a wholly reactive load, such as a pure capacitance of impedance –4jΩ, the load-line is a circle of radius 10A – or 40V – centred on Y; this is clearly a very severe test, as all the amplifier power is dissipated in itself and none in the load. It is also quite unrealistic. Opinions differ as to how reactive a load it is reasonable to drive to full output; both 45° and 60° have been suggested.[10,11 resp.]

For safe operation the output protection locus must fit between the load-line envelope and the safe operating area limits.[12]

And for non-sinusoid signals?

The emphasis so far has been on sinewave drive. You have to start somewhere. It is plausible that time-varying signal amplitudes and frequencies, or alternatively arbitrary waveforms, could produce more demanding situations for the amplifier, particularly in terms of peak current.

Cordell[13] has demonstrated that waveforms specially designed to exploit the energy storage in the reactive elements can increase the currents drawn, by timing a voltage reversal so that it aids the back-emf due to cone velocity. I can confirm this is so, but the correlation between such waveforms and reality remains uncertain.

I conclude that a truly 'difficult' load is one with lots of small humps and dips giving significant phase-shifts and increased peak dissipation across most of the audio band. Low impedances at the high-frequency end – above 5kHz – are particularly undesirable as they increase amplifier crossover distortion.

Summarising the system

Looking at the amplifier-cable-speaker system as a whole, I have already shown in my previous articles[14,15] that the amplifier and cable impedances have the following effects with an 8Ω resistive load.

• A constant amplitude loss due to the cable resistance forming a potential divider with the 8Ω load. The resistive component from the amplifier output is usually negligible.

• A high-frequency roll-off due to the cable inductance forming an *LR* low-pass filter with the 8Ω load. The amplifier's output inductor – included to give stability with capacitive loads – adds directly to this to make up the total series inductance.

The main factors in speaker cable selection are therefore series resistance and inductance. If these parameters are below 100mΩ and 3μH, any effects will be imperceptible. This can be met by 13A mains cable – especially if all three conductors are used.

When an amplifier is connected to a typical speaker impedance rather than a pure resistance the further effects are:

• The frequency response of the voltage at the loudspeaker terminals shows small humps and dips as the uneven speaker impedance loads the series combination of amplifier output impedance and cable resistance.

• The variable loading affects the amplifier distortion performance. High-frequency crossover distortion reduces as load resistance increases above 8Ω; even 68Ω loading increases high-frequency distortion above the unloaded condition.

For loading heavier than 8Ω, crossover may continue to increase, but this is usually masked by the onset of Large Signal Nonlinearity[16].

• Severe dips in impedance may activate the overload protection circuitry unexpectedly. Signal amplitudes are higher at low-frequency so impedance dips here are potentially more likely to draw enough current to trigger protection. ■

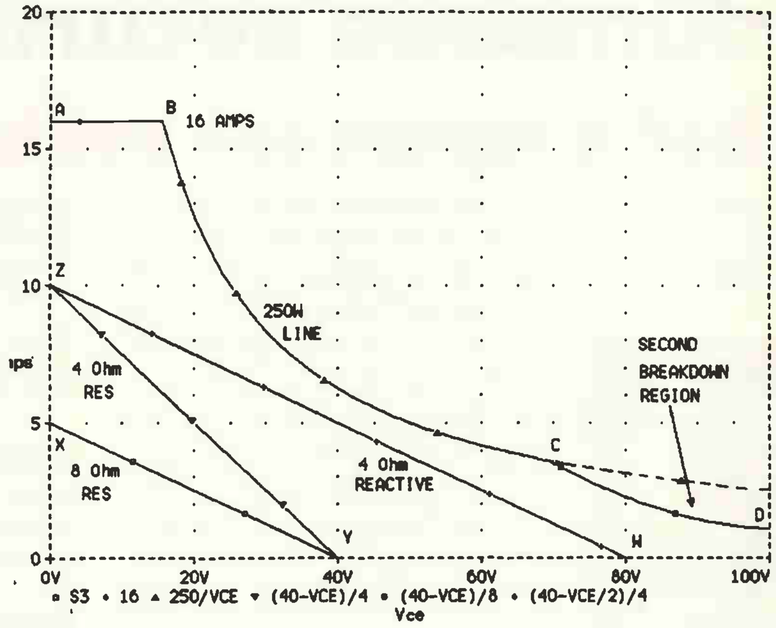


Fig. 10. Safe Operating area for the MJ15024 power transistor. AB is the 16A I_c limit, BC the 250W P_{diss} limit, and CD the edge of the second breakdown region. The V_{CE} limit is way off the picture to the right, at 250V. ZW is the envelope of loads of 4Ω plus any amount of reactance, for ±40V rails.

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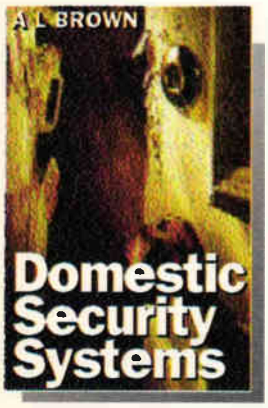

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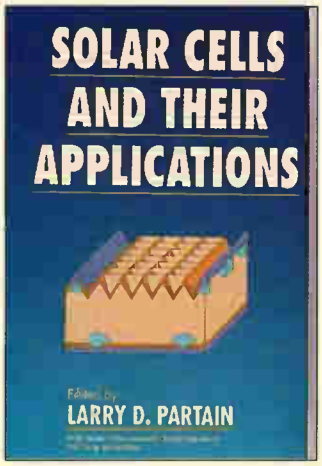

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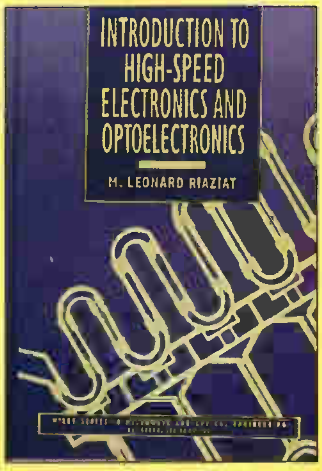

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On TRAC

Field-programmable logic chips are versatile building blocks that can be configured to perform almost any digital processing task – and many types can be rapidly reconfigured to suit changing needs. But is there such a solution available to analogue circuit designers? David Winch explains that there is.

All businesses want to confirm that their new product ideas work as soon as possible, and to make sure that finished products get to market without delay.

In digital electronics, these desires have been partly met by the adoption of a structured, top-down design discipline in both software and hardware. In addition, the programmability of digital hardware has enabled the evolution of universal circuit boards and ICs which further cut down the design cycle time.

By contrast, analogue design has essentially always been a bottom-up discipline – structureless and with very little computerised design support, backed by analysis tools. This has led to a general perception that the quickest solution will always be digital.

This situation is compounded by the fact that the tools available to date have made digital engineering far easier to teach. And, because it is a new use of 'arithmetic', is regarded as 'more exciting'. As a result, universities produce more designers of digital circuitry than they do analogue designers – which helps to reinforce the perception.

Providing analogue designers with the same benefits available to field-programmable gate-array users, a device called *TRAC* – an acronym for totally reconfigurable analogue circuit – is intended to address these problems.

The real world is analogue

The introduction of *TRAC* means that businesses no longer need to make substantial investments in digital designs in order to solve what are essentially analogue problems.

By defining the core problems and embracing the top-down design concept, *TRAC* enables new ideas and opportunities for creative solutions. Claimed to be one of the first exponents of a structured, top-down approach to analogue design, *TRAC*, like its digital fpga counterpart, offers rapid prototyping, fast time to market and easy to use programming tools.

TRAC has other advantages over digital solutions. It removes the overhead of anti-aliasing filters, sample and hold circuitry, and a-to-d and d-to-a conversion. It also works in

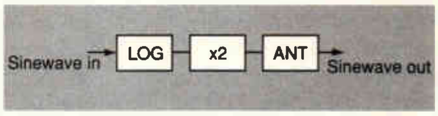


Fig. 1. These functions produce a frequency doubler. Digital frequency doublers are also easy to produce – unless you want a sinewave output.

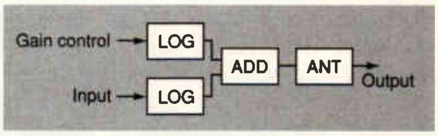


Fig. 2. Log, add and antilog functions combine to make an amplifier with infinitely variable programmable gain.

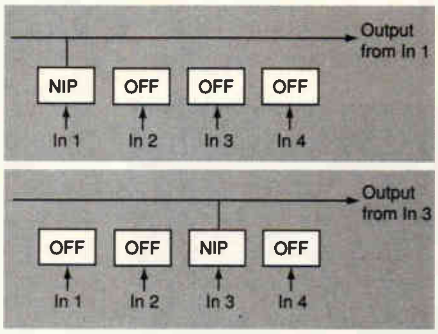


Fig. 3. Analogue multiplexing with TRAC makes use of the device's non-inverting pass function and its almost instantaneous reprogrammability.

David is Programmable Products Manager at Fast Analog Solutions – a Zetex Group company focussing on programmable analogue devices.

continuous real time.

Though not an absolute equivalent of an field-programmable gate array, *TRAC* offers all the benefits associated with such a device. It is supported by quick to learn and easy to use Windows-based drag-and-drop design software, ensuring rapid product development.

Design simulation is completed in a few seconds and provides results in both graphical and tabular outputs. Downloading to silicon via the parallel port takes a few microseconds,

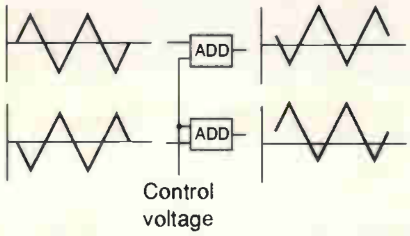


Fig. 4. In the smps design, inverted and non-inverted outputs from the integrating oscillator feed two adders. The adders' second inputs connect to a control voltage.

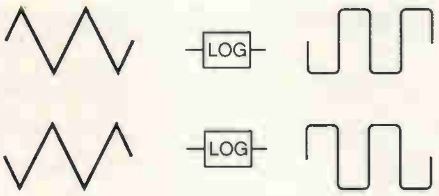


Fig. 5. Logarithmic elements are used to amplify the sudden change in polarity in the triangle waveform.

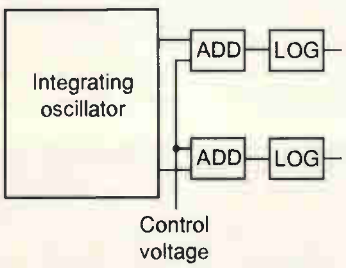


Fig. 6. How main elements of the smps controller come together. Combined, the adder and logarithm functions do the job of the traditional smps error voltage comparator.

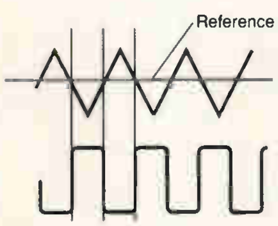


Fig. 7. Outputs of the integrating oscillator, as used in the smps example.

to create *TRAC* silicon, producing a device that closely matches the simulation.

Uses of the device are limited only by the designer's imagination. Whatever the application – be it power management, motor driving, control and instrumentation or phase measurement – *TRAC* can implement the function.

Think laterally

The key to making the most of *TRAC*'s programmable analogue design method lies in the ability to think laterally. You need to 'think outside the box', as they say. If you have enough flexibility of vision, almost anything is possible using analogue design.

What about doubling a frequency for example? Your initial reaction may be to suggest the use of digital signal processing. Remember though that $\sin 2x$ can be written in terms of $\sin^2 x$. Now, log, multiply and antilog functions available in *TRAC* make squaring analogue signals easy, **Fig. 1**.

Can *TRAC* act as a programmable gain amplifier ? Yes it can, and since the solution is an analogue one, gain can be made infinitely variable between two values. And how about a voltage to set the gain? This is simply an analogue multiplier. Multiplication is easy with *TRAC*'s log, add and antilog functions, **Fig. 2**.

And analogue multiplexing? Don't fall into the trap of implementing existing designs. The core problem is connecting the right input to the output at the right time. With *TRAC*'s non-inverting pass function and its infinite, virtually instantaneous re-programmability, this task becomes almost trivial, **Fig. 3**.

TRAC can implement the function, whether it be a switched mode power supply, proportional-integral derivative controller, multiplexer or any other application.

Implementing an smps with *TRAC*

Linear power supplies contain a series regulator and a mains transformer, making them heavy and inefficient. With linear supplies efficiency can be as low as 30% whereas efficiencies for switched-mode designs can be 80% and higher.

Switched-mode designs offer a reduction in size of the power supply because when used at high switching frequencies a smaller transformer is required. Reduction in size and improvement in efficiency makes switching to smps very attractive.

The benefit of using the *TRAC* device as the core of an smps is that the design can be readily tailored to the voltage and current required. With its low offsets, the technology ensures that good matching is achieved but the best feature of the device is the ease of re-configurability of the design. If changes are required for a particular solution, then the device can be altered almost instantly.

The timing parameters – i.e. the switching period and the width of the output pulse – can readily be changed because the smps design is re-shaped by simply changing external components.

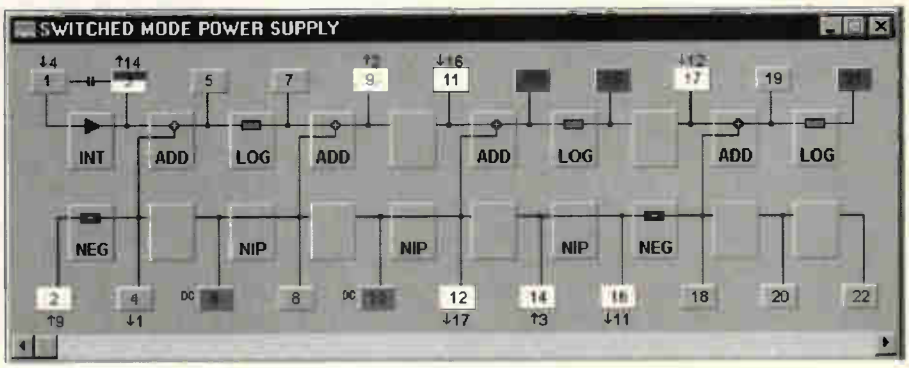
Behind the smps design

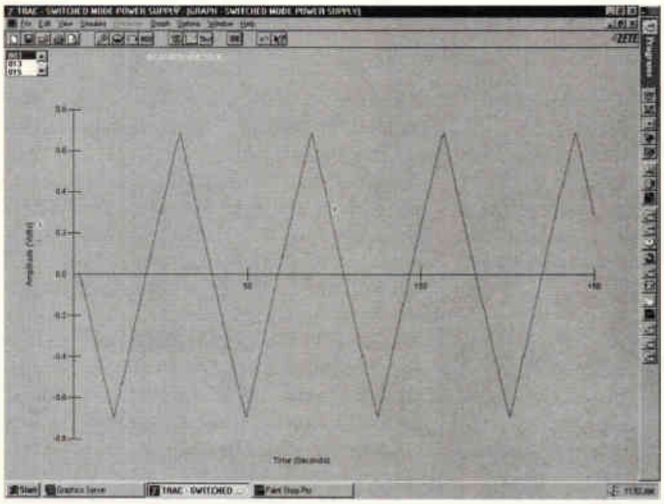
In this design an integrating oscillator forms the basis for the switched mode power supply. The output from the integrating oscillator is fed into an adder, this signal is also inverted and is fed into another adder. The other input to the adders are connected to an external control voltage, **Fig. 4**.

When the output from the integrating oscillator equals the control voltage on the rising edge, the output from the adder switches polarity. The sudden change in polarity is amplified via a logarithmic function which outputs $+V_{be}$. The change of polarity also occurs on the falling edge of the integrating oscillator output when the signal output switches to $-V_{be}$. An output pulse also occurs on the inverted output of the integrating oscillator, **Fig. 5**.

Comparator operation is provided by the adder and logarithmic amplifier functions, **Fig. 6**. The integrating oscillator is a useful tool

Fig. 8. Configuring the device as a switch-mode psu controller under Windows involves little more than dragging, dropping and clicking. Once simulated and verified, the design is downloaded directly to TRAC and the implementation is complete.





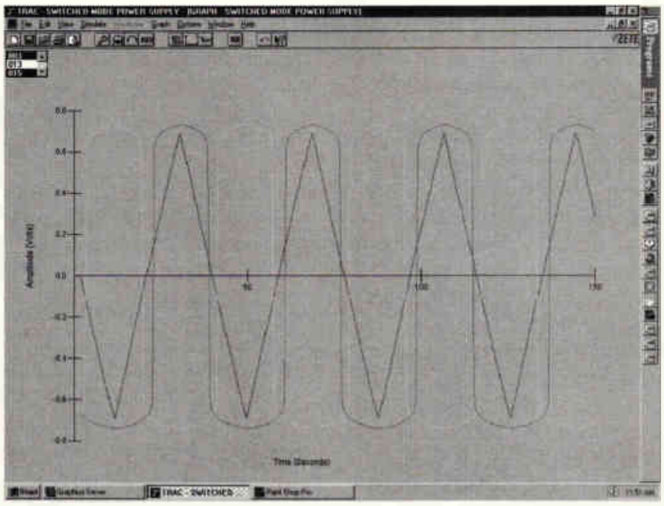


Fig. 9 & 10. Simulation results for the switched-mode power supply. Fig. 9 shows output from the integrating oscillator while 10 represents output signals interleaved.

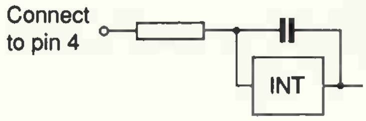


Fig. 11. In the simulation, pin 4 connects to the integrator's virtual ground (diagram directly above). In the real hardware, a resistor is inserted into the connection.

and will be of benefit in many other applications.

The smps design includes a integrator followed by an adder which in turn is followed by a logarithmic amplifier. There is a positive feedback loop from the output of the log function to the input of the adder. There is also negative feedback from the output of the log function to the input of the integrator.

Output from the integrator switches between V_{be+} and V_{be-} and is represented by,

$$E_o = \frac{1}{CR}\int V_{be}dt$$

Since V_{be} is ideally constant the integral becomes,

$$E_o = \frac{-V_{be}t}{CR}$$

At time T the inputs to the adder become equal and opposite, and will have moved through $2V_{be}$,

$$2V_{be} = \frac{-V_{be}T}{CR}$$

where T=$2CR$

Also at time T the output from the log function switches polarity and the integrator switches over. A complete cycle comprises two T periods and the oscillation frequency is then given by,

$$f = \frac{1}{2T} = \frac{1}{4CR}$$

The integrating oscillator can output in either triangular or square waveforms, **Fig. 7**.

Programming

On cell 6 in **Fig. 8**, the dc input is 0.0001V. This ensures that the integrating oscillator starts up when simulated. On the real hardware, this extra dc input is not necessary as noise will cause the oscillator to start.

The switched-mode power supply circuit shows the versatility of designing with the *TRAC* concept. The method of a string-like architecture and the use of links is clearly an advantage when manipulating signals. If the design is made more complex, then a further

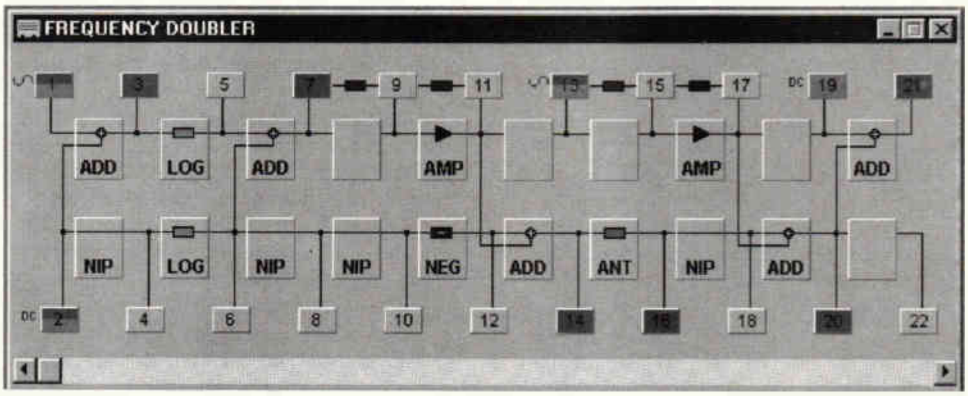


Fig. 12. Cell page showing the layout of the mathematical operations to create a frequency doubler.

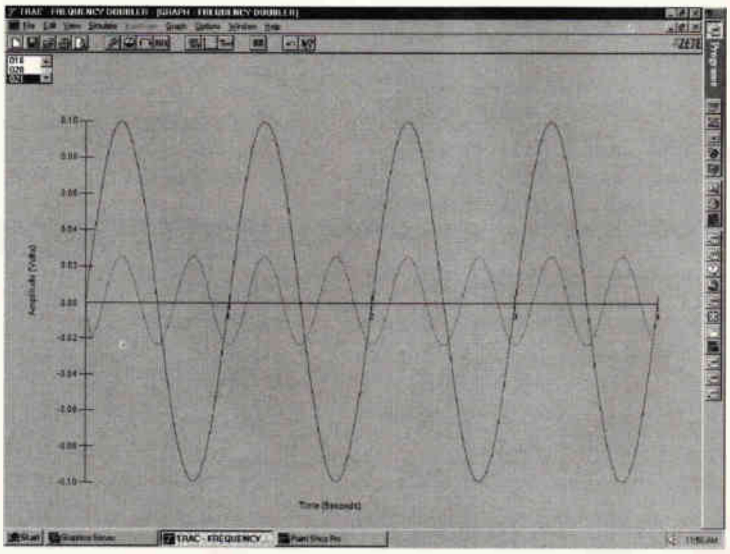


Fig. 13. Results of the frequency doubler simulation, showing that the unwanted ac and dc content has been removed.

TRAC device can easily be cascaded to extend the design.

The simulation results for the switched mode power supply design are shown in **Figs 9** and **10.** Figure 9 shows the triangular waveform generated by the integrating oscillator at pin 3.

In Fig. 10, both interleaved output signals are shown mapped over the triangular wave. You will notice that the output signals change polarity when the triangular wave amplitude is plus or minus the control voltage. For this simulation, the level is set at 0.1V.

SMPS hardware details

The integrator function is performed using an auxiliary function and external components. The link from pin 4 does not connect directly to pin 1, the virtual ground of the integrator, as shown in the simulation. This link is instead replaced by a resistor.

Because the integrator is positioned in cell 1, the external components have to be positioned accordingly. **Figure 11** shows the positioning of the external components. Suggested nominal values for the resistor are between 1kΩ and 10kΩ. The nominal value for the capacitor is from 1nF.

The values for the external components can be set in the software. When defining the auxiliary function as an integrator, a value is required to allow the software to simulate the design. This 'value' is equal to the *1/RC* time constant which sets the operating characteristics of the integrator.

Frequency doubling

A number of applications rely on frequency doubling, including signal conditioning and encoding and decoding. A non-linear circuit is often used to generate a signal that is a multiple of the input signal frequency. This technique is frequently useful if high frequencies are involved and good stability is required.

The *TRAC* is especially suited to frequency doubling because of its flexibility and accuracy. Requirements of conditioning signals may alter and the *TRAC* device can be programmed

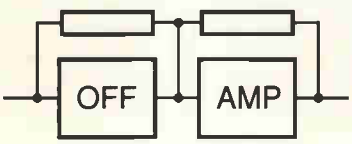


Fig. 14. Positions of the external components needed for the frequency doubler design.

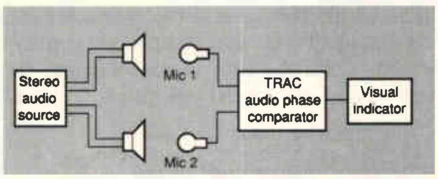


Fig. 15. Elements of a loudspeaker phase comparator incorporating TRAC.

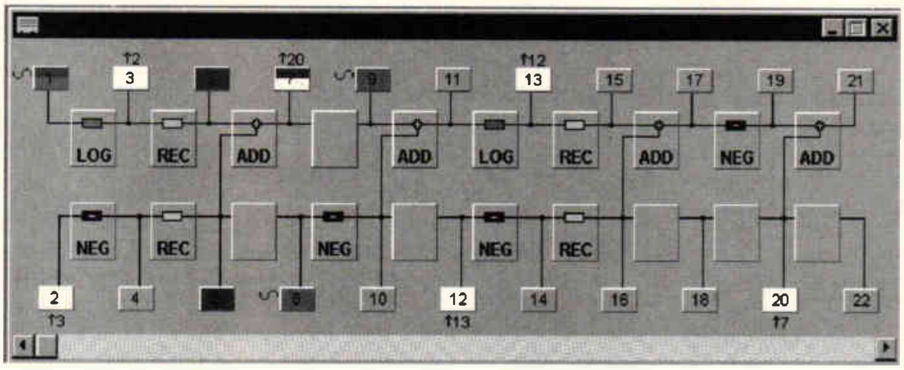


Fig. 16. Functional blocks required to implement the speaker phase comparator.

to change its characteristics accordingly.

When used as a frequency doubler, the device produces an output signal with low harmonic distortion. The chip also has the benefit that it can be re-configured in a matter of minutes.

The doubler described here is based on a standard trigonometric function. This function is implemented by decomposing the operation into a series of mathematical steps. These are emulated in the *TRAC* software to give the required simulated response. This design can then be realised using the accompanying *TRAC* hardware.

Frequency doubling in theory

The frequency doubling operation is realised by a mathematical approach. This involves taking the log of a signal and then doubling the resulting signal. The final stage is to take the antilog of the doubled signal.

The first stage is to take the log of an input signal $E_o.\sin(\omega t)$. The signal is then doubled to give $2\log E_o.\sin(\omega t)$. This is equal to,

$$\log E_o^2 \sin^2(\omega t)$$

Finally when the signal goes through an antilog function,

$$E_o^2 \sin^2(\omega t)$$

Using the trigonometric identity,

$$\sin^2(\omega t) = \frac{1-\cos(2\omega t)}{2}$$

gives,

$$E_o^2 \frac{(1-\cos(2\omega t))}{2}$$

This shows that the frequency is doubled in the term $\cos(2\omega t)$.

This is a simple procedure, but performing this function generates unwanted dc and ac signals that have to be removed.

Designing the doubler

Below are some of the mathematical steps linked with the *TRAC* cell diagram shown in **Fig. 12.** Each pin number refers directly to the pins shown in the diagram.

Alternating input at pin 1 is $E_o.\sin(\omega t)$ while dc input at pin 2 is E_{dc}. These signals are added. This is done to move the input signal off the zero dc level. This eliminates problems that would occur due to cross over distortion.

The function on pin 3, $-(E_o\sin(\omega t)+E_{dc})$, logs the result and adds a further log dc signal. This is necessary in the latter stages of the design to allow the subtraction of an unwanted dc signal component.

Pin 7 carries the function,

$$-(\log(E_o\sin(\omega t)+E_{dc})-\log E_{dc})$$

At this point the signal is doubled and the trigonometric identity is used. Pin 11 is now represented by,

$$\log\left(\frac{E_o^2}{2(1-\cos(2\omega t))}+2E_{dc}E_o\sin(\omega t)+E_{dc}^2\right)-\log E_{dc}^2$$

This signal is used to remove unwanted ac content. Pin 13's function is $E_o\sin(\omega t)$. The result of the antilog function gives, at pin 16,

$$\frac{E_o^2}{2E_{dc}}(1-\cos(2\omega t))+2E_o\sin(\omega t)+E_{dc}$$

Removing the unwanted ac signal gives:

$$-\left(\frac{E_o^2}{2E_{dc}}(1-\cos(2\omega t)+E_{dc})\right)$$

at pin 20. After removing the unwanted dc signals the final result appears on pin 21 as,

$$\frac{E_o^2}{2E_{dc}}\cos(2\omega t)$$

Simulating the doubler

After the design has been configured in the *TRAC* cell page, the only job left is to press the simulation button in the software. The results are presented graphically in **Fig. 13.** This shows that the unwanted ac and dc signals have been removed to leave an output signal twice the frequency of the original signal.

Designing the doubler's hardware

The prime function of the *TRAC* development hardware is to provide a tool for the electrical evaluation of the chip's design. This can be done by connecting the external inputs, output monitors, any necessary external components to the *TRAC* device and then programming it to the given design.

The design of the frequency doubler requires two auxiliary functions to be configured as ×2 amplifiers. This can be done by inserting 1 and 2kΩ resistors in the appropriate sockets on the development board.

External component positions for the frequency doubler are shown in **Fig. 14**.

Audio phase indicator

Developed for audio applications, this phase indicator determines whether two loudspeakers are operating in or out of phase.

Incorrect connection at the terminals of an amplifier or loudspeaker can result in the signals being out of phase. This reduces the sound level – particularly at low frequencies – and causes other more subtle problems.

To use the indicator, two microphones are placed at equal distances from two separate loudspeakers. The signals from the microphones are processed using a single *TRAC* device, **Fig. 15**.

A suitable indicator at the *TRAC* output would be a bi-colour led or a moving coil meter. The *TRAC* carries out the analogue processing. An external driver would is needed if the bi-colour led option is chosen, as the output swing of the *TRAC* is slightly below that necessary to drive the diodes.

The phase indicator circuit can be used with signals derived from a generator – producing, for example, a sine wave – or from audio sources. It gives an indication of an in or out of phase relationship, but does not give numerical indication of any phase difference that may exist. This could easily be achieved by extending the circuit.

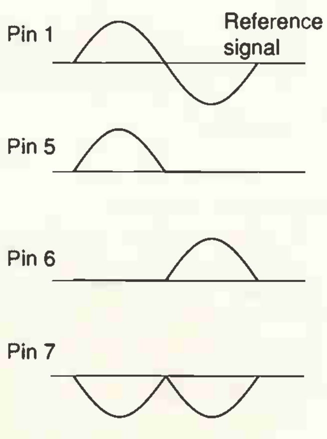


Fig. 17. Waveforms found on the phase comparator's pins 1, 5, 6 and 7 are a function of the reference signal only.

Detecting phase difference

This design is based on the assumption that the inputs from the microphones will either be in phase or 180° out of phase. The requirement is to generate an output that is a function of the relative phase of the two input signals.

This design benefits from the fact that, at any particular instant, both inputs have the same frequency, although they may have opposite magnitude. The output is the reference signal, full wave rectified; positive for 0° phase difference and inverted for 180° phase difference.

In practice, the design must be tolerant of amplitude differences and phase variations due to inaccurate positioning of the microphones. The output does switch polarity for a small duration when there are minor phase differences, in the region ± 20°, but a smoothing capacitor on the output would eliminate this effect. Amplitudes differences have no major effect on the circuit performance.

Producing the software

The *TRAC* circuit is shown in **Fig. 16**. Note, the output of any *TRAC* function, excluding the OFF and NIP, is inverted with respect to the input.

The reference signal is applied to pins 1 and 9, while the sample signal is applied to pin 8. Output is taken from pin 21.

For the purpose of the simulation, the reference signal is a sine wave. The out of phase sample signal can be created by defining a phase offset within the *TRAC* simulation menu.

Assume here that both signals have equal amplitudes. Firstly, consider the reference signal applied at pin 1. This signal is passed through a LOG and the output is processed through two different paths. One path contains a REC and the other contains a NEG and REC

Note that a REC function is a form of antilog function that gives a zero output when the input is positive and a positive output with a negative input. The results – pins 5 and 6 respectively – are added together giving a sum which is the reference signal full wave rectified and inverted.

Figure 17 illustrates the waveforms at pins 1, 5, 6 and 7. These waveforms are a function of the reference signal only and are independent of the phase relationship of the reference and sample signals.

The sample signal applied to pin 8 is inverted by a NEG function. This inverted sample is added to the reference and the sum is seen at pin 11.

When the reference and sample are in phase, the sum is equal to zero and when out of phase, the sum is equal to the sample amplified by two. **Figure 18** shows the signals at pins 9, 10 and 11.

Next, the sum is processed in a similar fash-

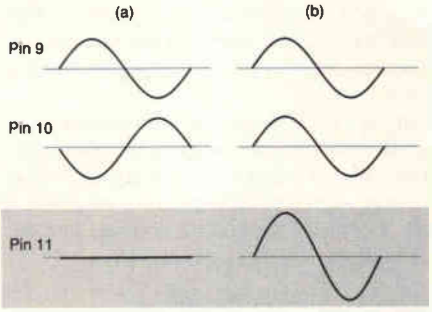


Fig. 18 When the reference and sample inputs are in phase, the sum is zero while when they are out of phase, the sum is twice the amplitude of the sample.

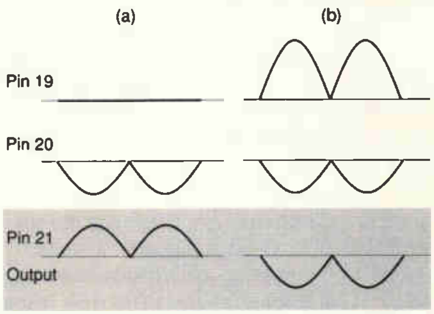


Fig. 19. Intermediate and output waveforms, for in phase and 180° out of phase conditions.

ion to the reference signal at pin 1, where the waveform at pin 17 is the signal at pin 11 full wave rectified and inverted. As the signal at pin 11 is zero when the circuit inputs are in phase, this processing path has no effect and pin 17 will also be equal to zero.

An external link connects pin 7 to pin 20, thus the reference signal full wave rectified and inverted is seen at both. This signal is then added to the signal at pin 19, which in turn is pin 17 inverted.

Pin 19 is equal to zero for in phase relationship and for 180° phase difference, pin 19 will be effectively the reference signal doubled in magnitude and full wave rectified. The circuit output is pins 19 and 20 added together, Fig. 19.

In summary, the output is a full wave rectified signal, being positive or negative depending on the phase relationship,

For preliminary evaluation purposes, the input and reference signal that would normally come from the two microphones can be derived from two signal generators. ■

For more information, visit Fast Analog Solutions' web site at http://www.fas.co.uk

Design competition

Will your TRAC design win you one of six development kits worth £600?

Competition rules

No purchase is necessary to enter this competition • Entries will be judged jointly by Zetex and *Electronics World* • The judges' decisions will be final and no correspondence will be entered into • Fast Analog Solutions reserves the right to publish competition entries in its applications literature • There is no cash alternative to the prizes offered • Multiple entries are welcome: each will be judged on its individual merits • Employees and agents of Zetex, Zetex related companies and Reed Business Information are not eligible for entry • The closing date for receipt of entries is 30 November 1997 • Posting your entry represents acceptance of these rules • The winners' names will be published in this magazine, and notified in writing.

Win one of six TRAC development kits worth £600 each

TRAC – from Fast Analog Solutions, FAS, the programmable product arm of Zetex – is a unique analogue building block. It provides a structured, top-down approach to solving design problems using analogue techniques – rather than resorting to digital signal processing. The uses of TRAC are limited only by the designer's imagination. The device has the potential for enabling entirely new analogue solutions.

In return for your TRAC design idea, FAS is offering one TRAC development kit to each of the six best entries. In addition, the best of the designs will be published in Electronics World.

Entries will be evaluated on their creativity and ingenuity as well as their ability to improve over existing solutions in terms of component count, circuit complexity, board space requirements, development time, assembly time and end-product market edge.

Since the development suite on the CD ROM includes simulation, there's no need for you to make a prototype circuit. Simply work out your design using the demonstration CD and send us the resulting output file together with any notes you think necessary to tell us about the design and why you think it is a winner.

Details of how to implement the TRAC device are given in the article on page 930 and in the form of examples on the demonstration CD, available to UK readers only.

Entries should reach Electronics World's editorial offices no later than 30 November 1997and the envelope that they arrive in must be clearly marked FAS Competition. Our address is Electronics World Editorial, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.

Using the TRAC demonstration CD

On the CD presented free with this month's edition of *Electronics World* is not only the *TRAC* demonstration design software, but also numerous design examples and a data sheet for the *TRAC* chip.

The software is Windows based and fully functioning, except that the capability to download a design to the *TRAC* chip is disabled. Note that to enter the competition, only a simulation of your design is necessary. You are not obliged to verify the hardware.

With Windows 95, the install prompting routine loads automatically on closure of the CD drive tray. Simply follow the instructions of abort the procedure.

If you have Windows 3.1x, use File Manager to select the *TRAC* CD ROM, which will usually be drive D:. Next, find the file START.EXE in the main directory and either double click on it or select RUN from the start menu and type START.EXE into the box.

Uninstall facilities are included.

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049

CIRCUIT IDEAS

Over £600 for a circuit idea?

New awards scheme for circuit ideas

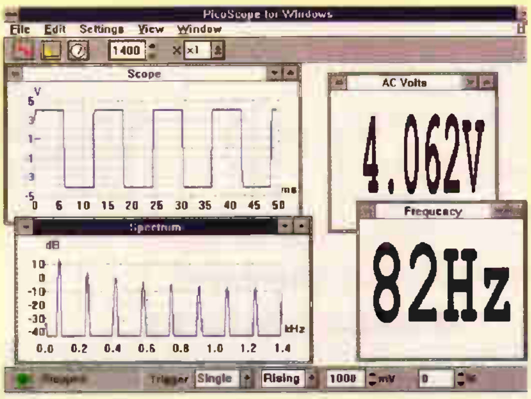
- Every circuit idea published in Electronics World receives £35.
- The pick of the month circuit idea receives a Pico Technology ADC42 – worth over £90 – in addition to £35.
- Once every six months, Pico Technology and Electronics World will select the best circuit idea published during the period and award the winner a Pico Technology ADC200-50 – worth £586.

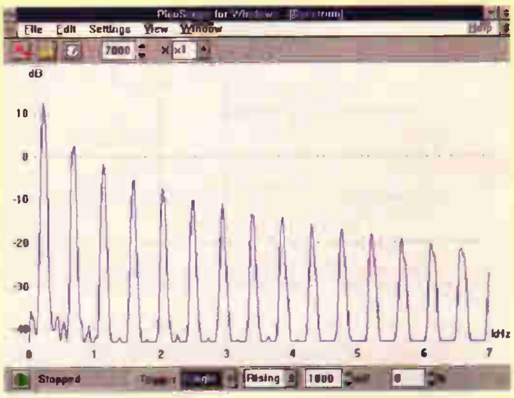
How to submit your ideas

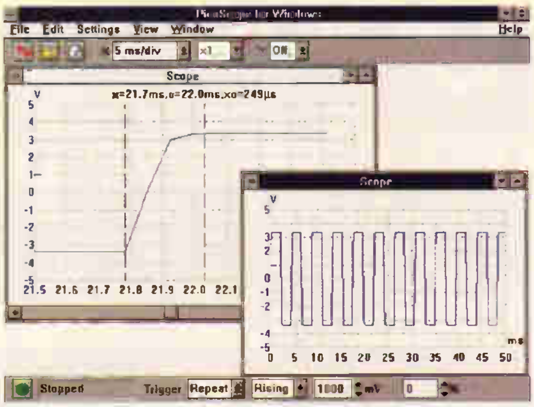
The best ideas are the ones that save readers time or money, or that solve a problem in a better or more elegant way than existing circuits. We will also consider the odd solution looking for a problem – if it has a degree of ingenuity.

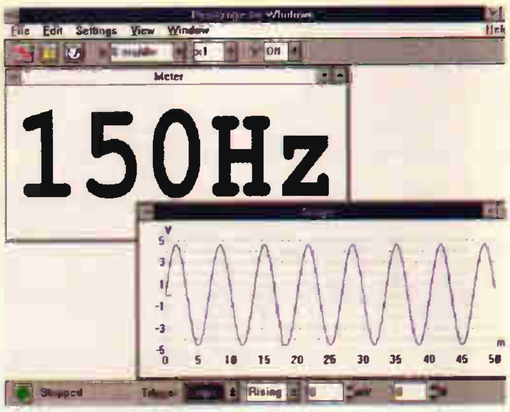
Your submission will be judged on its originality. This means that the idea should certainly not have been published before. Useful modifications to existing circuits will be considered though – provided that they are original.

Don't forget to say why you think your idea is worthy. We can accept anything from clear hand writing and hand-draw circuits on the back of an envelope. Type written text is better. But it helps us if the idea is on disk in a popular pc or Mac format. Include an ascii file and hard-copy drawing as a safety net and please label the disk with as much information as you can.









Turn your PC into a high-performance virtual instrument in return for a circuit idea.

The ADC200-50 is a dual-channel 50MHz digital storage oscilloscope, a 25MHz spectrum analyser and a multimeter. Interfacing to a pc via its parallel port, ADC200-50 also offers non-volatile storage and hard-copy facilities. Windows and DOS virtual instrument software is included.

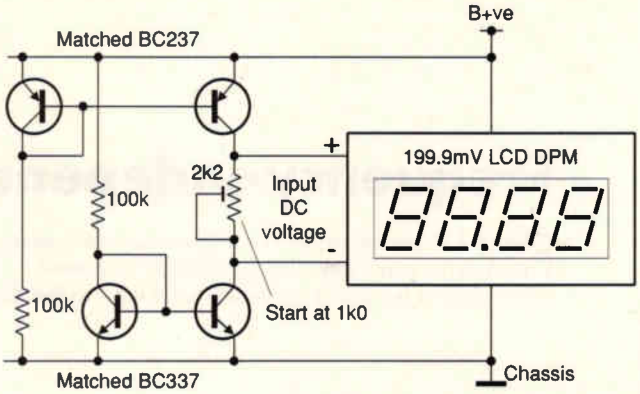
ADC42 is a low-cost, high-resolution a-to-d converter sampling to 12 bits at 20ksample/s. This single-channel converter benefits from all the instrumentation features of the ADC200-50.

Dpm monitors its own supply voltage

Without input dividers or range multipliers, this inexpensive circuit measures up to 19.99V dc. It is currently in use as a permanent car battery monitor.

The two matched pairs of transistors, used as current mirrors, should be on the same board and away from heat and air currents. If the circuit is calibrated at 12.6V, accuracy between 10V and 15V is better than 1%.

Graham Maynard
Newtownabbey
Northern Ireland
(A22)



Input circuit for a commercial lcd digital panel meter allows measurement to 19.99V dc without the need for input dividers.

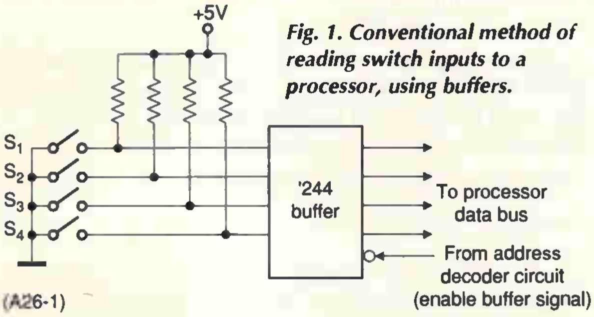


Fig. 1. Conventional method of reading switch inputs to a processor, using buffers.

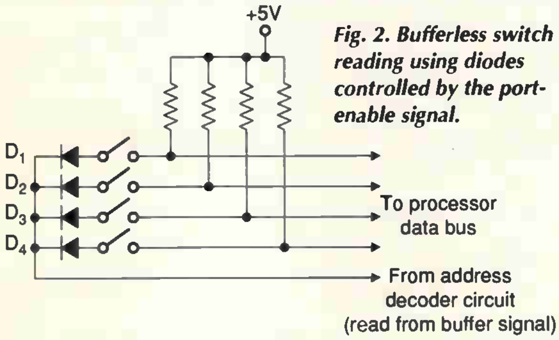


Fig. 2. Bufferless switch reading using diodes controlled by the port-enable signal.

Microcontroller reads unbuffered dip switches

Needing to read a set of dip switches to an *8051* and to save space and cost, I decided to omit the usual octal buffers, shown in the 'conventional' circuit of **Fig. 1**. Instead, the circuit of **Fig. 2** is now in use, although this a general circuit diagram, since an *8051* does not use the pullup resistors, the data bus needing its own. The circuit will read as many switches of any kind as may be needed and may be used with any processor with an external data bus. The omission of buffers does increase capacitive load on the bus.

Since the lower end of the switches sees the active-low port-enable signal, the diodes isolate the data bits when the enable signal is inactive, pulling the switches low when when enable is active, the data bus and hence the state of the switches now being read by the processor. It is advisable to use HCT/HC port-enable signal output with a guaranteed 0.1V logic low voltage. Diodes are general-purpose, small-signal types of the *1N4148* type.

Dhananjay V Gadre
Inter-University Centre for Astronomy and Astrophysics
Pune

Voltage-controlled oscillator

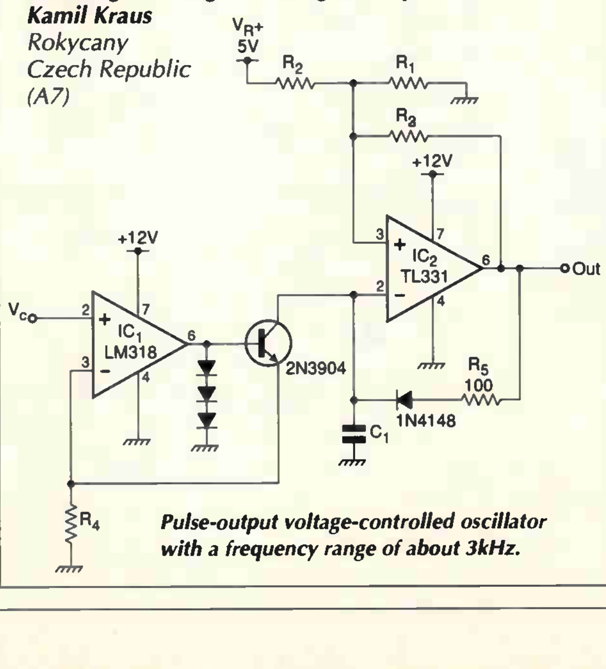
Producing pulses over a 3kHz range, this vco is simple and the control/frequency relationship linear over this limited range.

The *TL331* is used as a Schmitt trigger. When its inverting input is below the trigger voltage at the non-inverting input, output is high and C_1 charges through the diode and R_5 until the voltage across it reaches the trigger voltage, at which point the output goes low and feedback via R_3 and the voltage divider sets a new, lower trigger voltage. The capacitor now discharges through the current sink formed by the transistor and R_4 under the control of the input op-amp. As the capacitor voltage reaches the lower trigger voltage, the Schmitt output again goes high and the cycle repeats.

The result is a series of pulses at the output whose frequency is controlled by the speed of discharge of the capacitor, in turn controlled by the input op-amp, the relationship being $f=V_c/kR_4C_1$, where k is the difference between the two trigger voltages.

If upper and lower trigger voltages are chosen to be 4V and 2V, the ratios R_2:R_1 and R_3:R_1 are 0.515 and 0.33 respectively. Making R_1 10kΩ produces R_2=5.1kΩ and R_3=3.3kΩ. Choosing C_1=0.1μF, R_4=10kΩ and V_c zero to half the supply voltage, the frequency range is around 3kHz, higher voltages affecting linearity.

Kamil Kraus
Rokycany
Czech Republic
(A7)



Pulse-output voltage-controlled oscillator with a frequency range of about 3kHz.

Frequency-independent 90° phase shifter

Figure 1 shows the circuit of a voltage-controlled phase-lead shifter using an analogue multiplier with input E. The multiplier behaves as a variable resistance, so that the resistance from point (a) depends on E. Gain is unity and phase

$$\phi = 2\arctan\left[\frac{(1-E/10)}{\omega CR}\right],$$

to a maximum of

$$\phi = 2\arctan\left(\frac{2}{\omega \mathrm{CR}}\right).$$

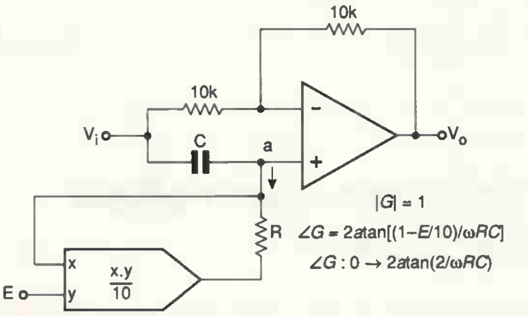


Fig. 1. Phase-lead shifter in which shift is controlled by input voltage **E**.

Substituting a multiplying digital-to-analogue converter for the multiplier allows digital control of phase, as seen in **Fig. 2** and interchanging the R and C turns the circuit into that of **Fig. 3**, a phase-lag shifter with

$$\phi = -2\arctan\left[\omega CR(1-E/10)\right],$$

to a maximum of

$$-2\arctan(2\omega CR).$$

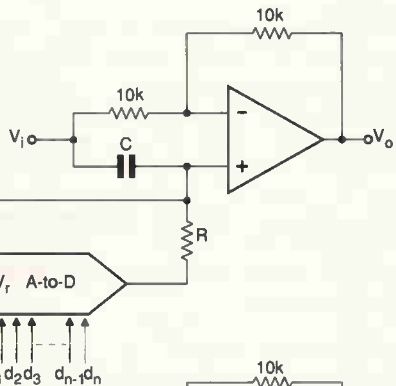


Fig. 2. Multiplying d-to-a converter provides digital shift control.

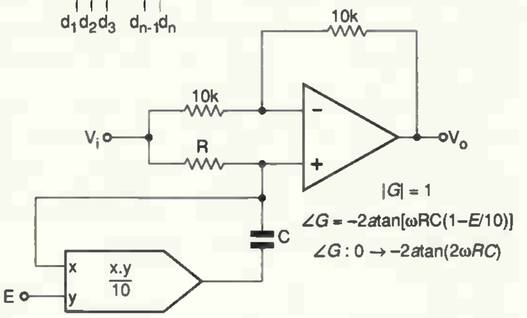


Fig. 3. Simply reversing the positions of the **R** *and* **C** *makes a phase-lag shifter.*

Figure 4 shows the circuit of the final circuit – a 90° phase-lag shifter that is independent of frequency over about two decades, assuming error and offset in the integrator op-amp can be kept to about 3mV. It uses the lag circuit and a second multiplier to provide a direct voltage proportional to phase shift. The integrator in the feedback loop attempts to zero this voltage and the result is a 90° shift, assuming

$$f > f_1 = 1/4\pi CR,$$

with $< 1°$ error if

$$f < f_2 = 1/36CR\Delta E,$$

where ΔE is integrator error. Clearly, the lead circuit of Fig. 1 can also be used in the 90° shifter.

Mai Saleh
Damascus
Syria
(A3)

ADC-42 WINNER

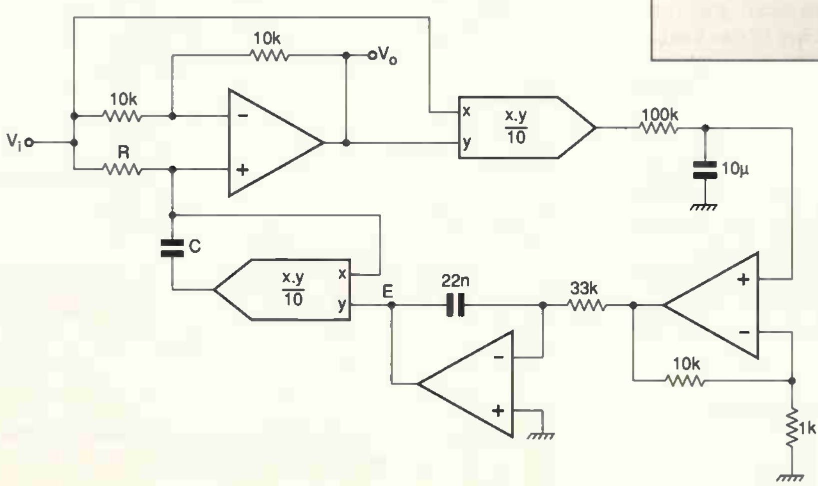
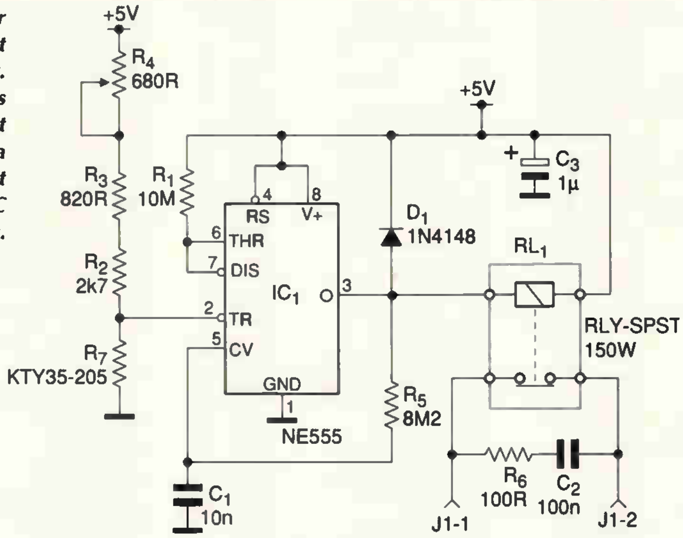


Fig. 4. Circuit of the frequency-independent 90° lead or lag phase shifter, which operates over two frequency decades with good-quality op-amps.

Simple thermostat with hysteresis

555 timer wears yet another hat. In this circuit it behaves as a thermostat with 0.25°C hysteresis.



With a relay and a few other components, a *555* timer functions as a precise thermostat with hysteresis.

Starting with the relay on, its contact open and no current to the heating element, the action of the *KT86-206* silicon temperature sensor is to reduce the voltage at the trigger input of the 555 to below $V_{cc}/3$, at which point the 555 begins to oscillate at a frequency determined by the value of R_1 and stray capacitance: about 2kHz.

As a result of the oscillation, pin 3 of the 555 goes positive with a duty cycle of about 98%, which switches the relay off, its contact closing and passing current to the heating element, whereupon the input to the trigger pin increases and stops the oscillation. Switching hysteresis is determined by the value of R_5, in this case 0.25°C.

With the input chain shown, adjustment between 10°C and 25°C is possible, the resistors concerned being metal-film types for stability.

Sam de Jonge
Geel
Belgium

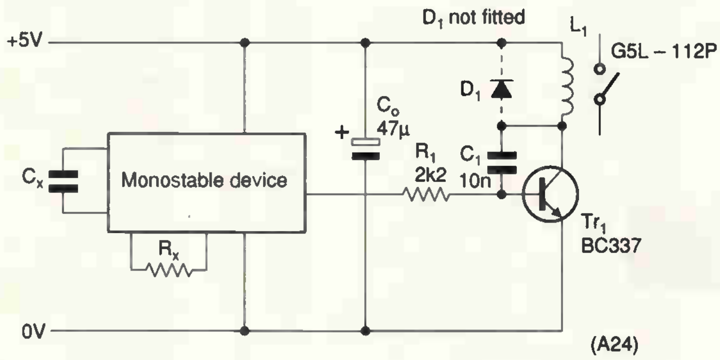
Relay noise suppressor

As a result of jittery noise on the output of a cmos monostable, which was driving a relay by way of an n-p-n transistor, high-frequency oscillation made its appearance on the relay coil. Decoupling the rather poor supply brought some improvement, but connecting a capacitor between collector and emitter of the transistor stopped the noise.

At this point, I realised that it would be more effective to use the transistor as a Miller integrator by connecting a much smaller capacitor from collector to base, as shown here. This has the additional advantage that the usual catching diode may be omitted, since the value of the capacitor can be such that the spike from the relay coil stays well within the transistor's collector voltage rating.

Relay operation is slowed down a little, but this can be compensated elsewhere in the circuit, for example by reducing the time-constant of the monostable.

C J D Catto
Cambridge



This solution to noisy resulting in spiky relays also renders the catching diode redundant.

Telephone privacy

When several extensions use one telephone line, it may become inconvenient to allow another extension to overhear your conversation. This very simple circuit arrangement prevents that possibility, cutting off all except the one in use, with no discernible effect on performance.

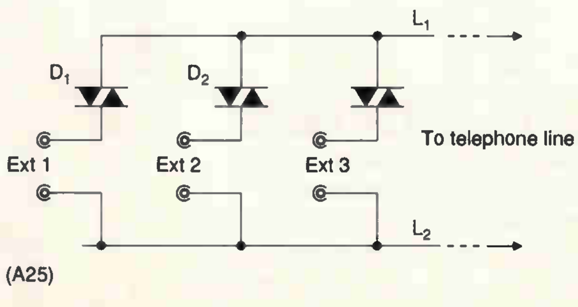
As the diagram shows, each handset is connected via a diac. Line voltage is 48V dc when all handsets are on-hook and around 6V dc when one is lifted. Diacs start to conduct when the applied voltage is above the diac's breakover voltage of about 35V dc and continue to conduct when voltage drops to a low level, ceasing conduction when current is too low to sustain it. (Diacs are used to conduct the ringing tone, since they conduct both ways).

If, then, one extension is lifted, breakover occurs at that diac and the extension is operative since the diac is still conducting at the 6V now across it and with the current through the handset of around 26mA. The others cannot break over at this voltage.

Since different handsets exhibit various impedances, it may be that large currents could flow, in which case small resistors may be needed in series with the diacs.

If two handsets are lifted simultaneously, only the one having the diac with the lowest breakover voltage will be activated.

Rahal Jayawardene
Nugegoda
Sri Lanka



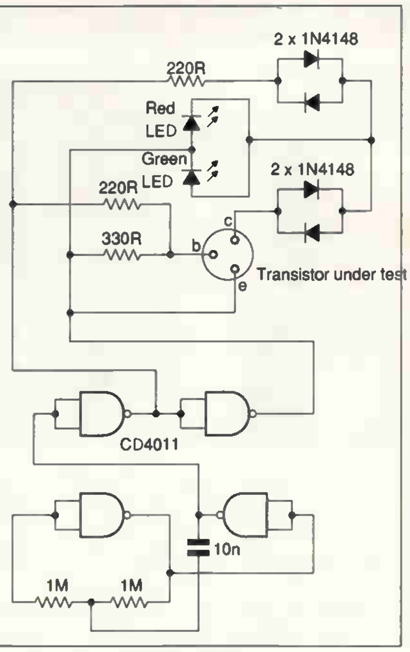
Only the telephone extension in use will operate in this arrangement, the others being cut off.

Transistor polarity checker

Blinking red or green leds indicate p-n-p or n-p-n transistors.

An oscillator built on the *4011* quad 2-input nand drives the gate of the transistor whose type is unknown. Depending on its polarity, the red or green led in its collector circuit will conduct; red for n-p-n. If there is no transistor, they both blink alternately.

Raj K Gorkhali
Kathmandu
Nepal



If you are unsure of which kind of transistor you have, these leds blink to indicate p-n-p or n-p-n types.

Amplifier polarity reversing switch

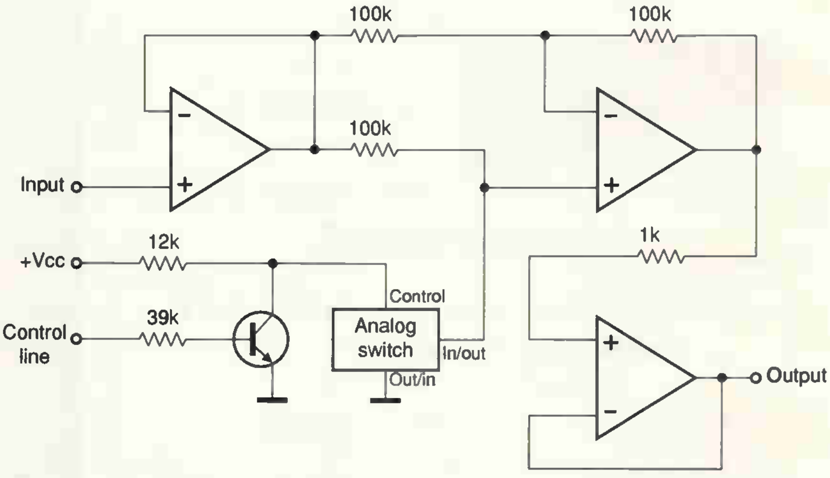
Figure 1 shows the circuit of a level-controlled reversing switch to produce at the amplifier output an in-phase or anti-phase version of the input. It was originally used in a linear, ground-referenced vco and is now used in three-state pulse generators and lock-in amplifiers.

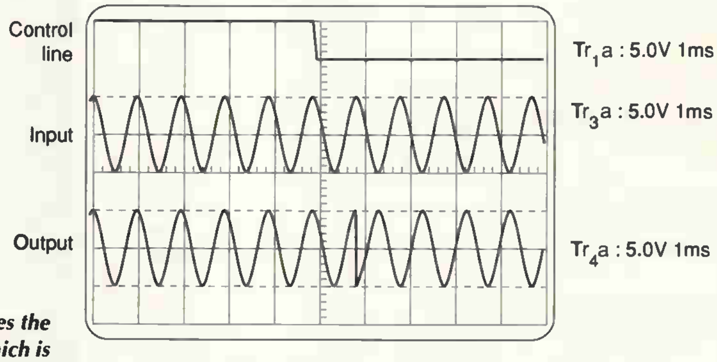
It depends on the state of the analogue switch; if the switch conducts, amplifier 2/4 (one element of a quad fet-input op-amp) becomes a unity-gain, non-inverting amplifier, reverting to an inverting type when the switch is off.

Figure 2 shows the operation; reversal takes place cleanly as the control input goes low. In **Fig. 3(a)**, the circuit is in use as a three-state pulse generator and **Fig. 3(b)** shows the control circuit truth table.

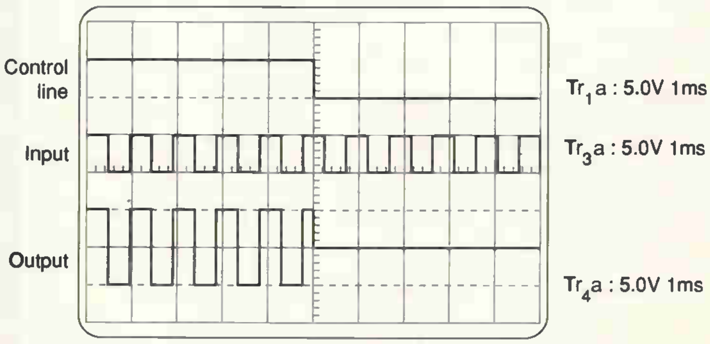
The analogue switch should be a low-resistance type for high accuracy.

Ashwani Karnal and
Hari Om Mahawar
Centre for Advanced Technology
Indore
India
(A10)





Logic-level control input determines the polarity of output from this amplifier, which is also used as a three-state pulse generator. Fig. 1 is top, Fig. 2 middle, Fig. 3 bottom and Fig. 4 below.



Control line	Input	Output
X	0	0
1	1	1
0	1	–1*

* –1 corresponds to –5V

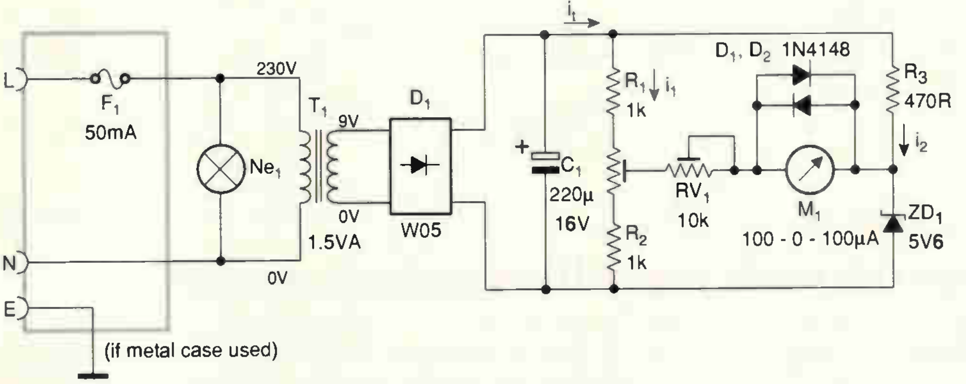
Mains deviation monitor

Since, post 1995, the European mains voltage is 230V+10%, –5%, it appears that drops of up to about 11V could happen without notice. Anyone to whom this might be a problem may like to try this monitor, which reads percentage deviation directly to an accuracy within 0.5% of reading.

The transformer, rectifier and smoothing capacitor provide an unregulated direct voltage varying directly as the mains input. On the left of the bridge, the resistor chain balances the bridge at 230V, the centre point varying with the input. On the right, the zener provides a stable reference, sourcing or sinking current when the input varies above or below 230V; this current is small and linearity is preserved.

Diodes across the meter are for protection on extreme input excursions, RV_1 balances the bridge and RV_2 provides calibration.

N K Goodman
Westfield
East Sussex



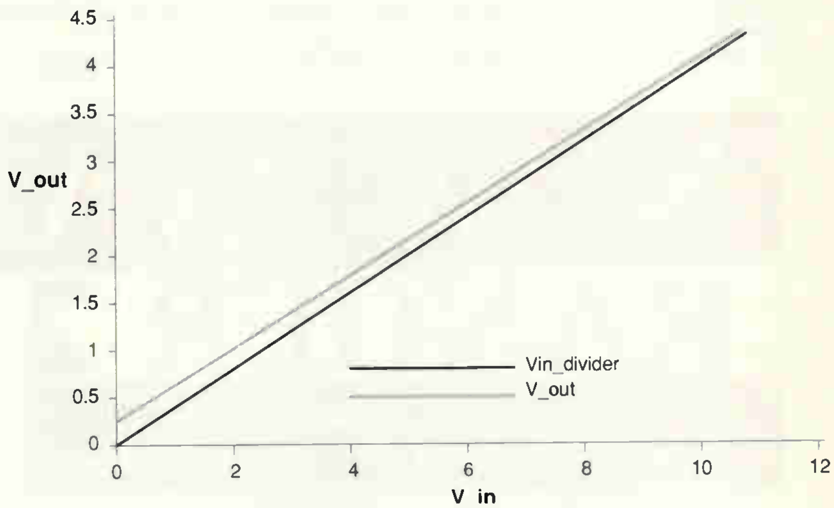
Isolated dc motor control

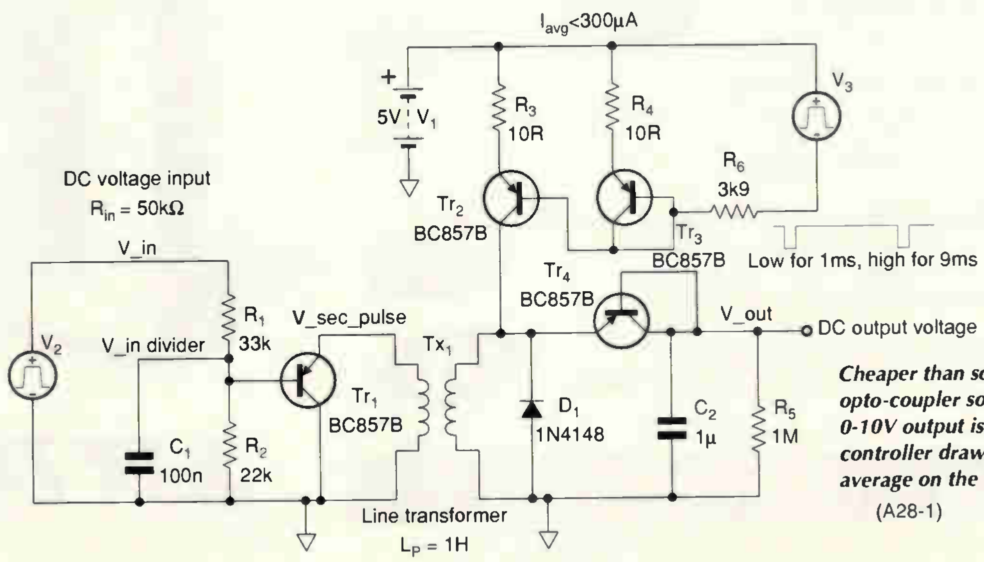
Having the requirement to control the speed of a fan motor, operating at mains potential, by an isolated, low-power, 0-10V dc input, we looked at a number of methods including isolating amplifiers, dc-to-dc converters and optocouplers. Since precision was not a requirement, this simple circuit was developed instead.

A microcontroller, operating at line voltage, was available and was used to supply a 100Hz train of negative-going pulses with a duty cycle of 10%. On receiving a pulse, the current mirror $Tr_{2,3}$passes a current of about 1mA through the primary transformer winding to charge C_2 via Tr_4. On the low-voltage side of the circuit, Tr_1 is a voltage follower which clamps the secondary voltage to 4V maximum, controllable by the variable input voltage, V_{in}. Since the isolating transformer has a turns ratio of 1:1, the variable-height pulses are reflected back to the primary, rectified and smoothed to provide the dc output voltage.

The circuit draws only 300µA average on the primary side. It is reasonably stable with temperature and, although there is some offset due to transformer resistance and current differences through base-emitter junctions of $Tr_{1,4}$, it can be removed and scaled in the controller's program.

Stefan Hansen
Sønderborg
Denmark
(A28)





Cheaper than some opto-coupler solutions, this 0-10V output isolated motor controller draws only 300µA average on the live side.
(A28-1)

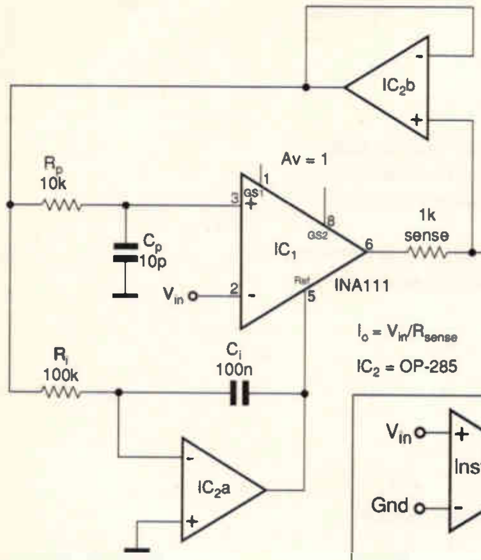


Fig. 2. Use of integrator feedback sets the dc operating point to centre the output.

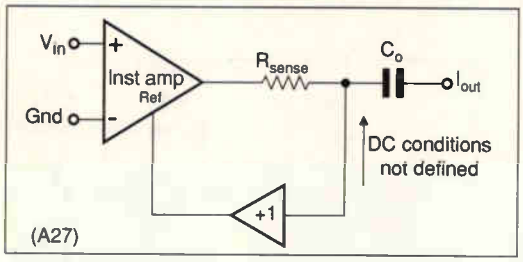
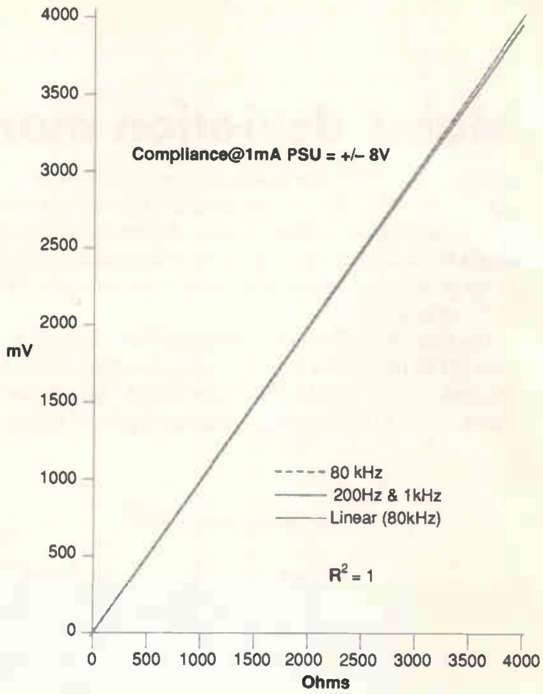


Fig. 1. Current source with capacitive load coupling has no defined dc operating point and is therefore unpredictably unstable.



Stable current source with capacitive output coupling

The ac current source shown in **Fig. 1**, having capacitive coupling to the load, is unstable, since its dc operating point is undefined. Output exhibits a tendency to wander off towards a rail.

In **Fig. 2**, the output is centred on zero by means of the integrator feedback to set the operating point, the non-inverting input and ref terminals having been transposed so that dc conditions are set by way of the output stage of the *INA111*. Phase margin is thereby improved and maximum voltage compliance (the range in which output voltage may lie in order to maintain constant current) boosted to 10V rms at ±15V.

Gain of the integrator shown in Fig.2 is suitable for operation from a few hundred hertz to over 100kHz, its time-constant being at least ten times the period of the lowest frequency for good voltage compliance. Components R_p and C_p compensate the control loop. Power supply needed is ±8V.

Alex Birkett
London SE22

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ACTIVE

Discrete active devices

Low-power, high-gain rf transistor. Motorola's *MRF917T1* low-power rf, n-p-n transistor is meant for use in low-noise, wide dynamic range front ends in pagers and cellular telephones. Gain-bandwidth product is 6GHz typical at 6V/2A and noise figure 1.7dB at 500MHz (2.3dB at 1GHz). This is a surface-mounted device in SC-70/SOT-323, supplied on tape and in reels. Motorola. 001 602 244 6108; fax, 001 602 244 4597

Schottky arrays. Zetex announces two Schottky diode arrays, in both s-m and through-hole versions, designed to reduce noise caused by reflections, cross talk and switching on high-speed parallel data lines. *SDA 12* (six diodes) and *SDA 32* (twelve) clamp between two supplies, such as V_{cc} and ground, their high peak forward current and low voltage and capacitance proving effective in the protection of data lines. Packaging is 8-pin and 20-pin SO or dil for through-hole mounting. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5467.

Linear integrated circuits

400MHz op-amps. *MAX4308/9* are voltage-feedback operational amplifiers from Maxim which combine high speed with very low distortion, meant for low-level, 12-bit to 16-bit application in medical imaging and instrumentation. They are decompensated versions of the *4108/9* and need closed-loop gain of 5 (*4308*) and 10 (*4309*). The 4308 has a 220MHz –3dB bandwidth, a 1200V/µs slewing rate and dynamic range free from spurii of 93dBc at 5MHz and R_L of 100Ω; the 4309 has a bandwidth of 200MHz. Further characteristics include 90mA output current, 0.004% differential gain and 0.008° differential phase. Settling time is 8ns to 0.1% and 12ns to 0.01%. Maxim Integrated Products UK Ltd. Tel., 01734 303388; fax, 01734 305511.

Instrumentation amplifier. Burr-Brown's *INA22* precision differential instrumentation amplifier takes only 60µA quiescent current from a single or dual 2.2-36V supply and its input common-mode range extends to 0.1V below ground. A single external resistor sets voltage gain from 5 to 10000, offset drift from the 250µV maximum input voltage offset being 3µV/°C. Output swing is within 100mV of the rails. Burr-Brown International. Tel., 01923 233837; fax, 01923 233979.

Microprocessors and controllers

Risc microcontrollers. Holtek's *48x00* range of microcontrollers feature risc architecture and have a maximum of 224byte of general-purpose data ram, 16 by 8Kb of program ram and special-purpose data memory. There are 63 single-word instructions, 56 bidirectional i/o lines and internal and external interrupt sources. Additionally the chips possess a programmable watchdog timer with a built-in oscillator and, on some of the range, voice synthesis and tone generation. A low-cost Windows-based development package is available. Joseph Electronics Ltd. Tel., 0121 643 6999; fax, 0121 643 2011.

Mixed-signal ICs

Video capture and decoding. Brooktree have two new devices for the capture of analogue and decoding of NTSC/PAL/SECAM video, both in single chips. *Bt829* is a low-cost device for pc capture applications such as image processing, videoconferencing and interactive video, supporting both square pixel and CCIR 601 resolution, with a pixel port for a number of interfaces. Other features include chroma/luma filtering, scaling and closed-caption decoding, with programmable hue, brightness and contrast and window cropping. *Bt848* is intended for pc television, Intercast, H.234 desktop conferencing, motion video capture, still capture and VBI data. As a bus master, no local memory buffers are needed to store pixel data. It is independent of the PCI bus. Telecom Design Communications Ltd. Tel., 01256 332800; fax, 01256 332810.

Motors and drivers

Rectifier kits. Kits for use in the design of drives for 230V ac or 460V ac motor drives in the 370W to 11kW region can be had from International Rectifier. *PoweRtrain* kits contain all power conversion and control circuitry, with or without braking, in modules significantly smaller than many other motor controllers. Features include an integrated power stage, control stage, a bill of materials, complete circuit diagrams, Gerber plots for board layouts and a system specification. International Rectifier. Tel., 01883 732020; fax, 01883 733410.

Oscillators

Vcos. *Vari-L* surface-mounted, voltage-controlled oscillators use a supply of 3.3V or 5V (some at 1.2V, 3.5mW) and are available in 15 centre frequencies between 773MHz and 992MHz and one at 1960MHz. All are single-ended, producing phase noise of under –108dBc/Hz at 1GHz, suppression of second harmonic better than –12dBc and of third harmonic better than –24dBc. Acal Electronics Ltd. Tel., 01344 727272; fax, 01344 424262.

PASSIVE

Cameras

Very small camera. *DM340* by Dyna Image is claimed to be the smallest ccd inspection camera with built-in lighting currently available. The lens is mounted on a flexible metal tube up to 300cm long or on a 300cm cable, if lighting is not provided; a mirror-image side-view attachment can be used. Output is EIA or CCIR with 512 by 492 or 512 by 582 pixels respectively. Auto exposure is 1/60 to 1/10000 or 1/50 to 1/10000 in the two formats. Case size of the main unit is 55 by 30 by 151mm. Selectronic Ltd. Tel., 01993 778000; fax, 01993 772512.

Passive components

Square pots. Rotary potentiometers from *Taiwan Alpha* have a square body measuring 9mm per side and a snap-in board mounting. The pots come with one, two, three and four tracks in values from 5kΩ to 200kΩ, all having metal bushing and shafts and, as options, push-lock mechanisms, rotary switches and/or momentary push switches. Ratings are 0.05W and 50Vac. TW Electronics Ltd. Tel., 01635 278585; fax, 01635 278122.

Miniature transformers. Dual-output, board-mounted mains transformers by Stontronics offer the choice of two inputs and five outputs. Each has two independent secondary windings and independent primaries to handle 115V or 230V input voltages. All conform to IEC742 and BS415 and have double-section bobbins for maximum insulation between windings. The units are designed to confer flexibility but the company points out that special requirements can also be met. Stontronics Ltd. Tel., 01734 311199; fax, 01734 311145.

Small trimpots. More surface-mounted trimming potentiometers are added to the BI range in the form of three 4mm square, sealed, single-turn types, the *23G, 23GL and 23J* models. The first two are gull-wing types with a mounted height of 2.54mm and 2.31mm, while the 23J is a j-hook version of 2.1mm height. All three are available in values of 10Ω-2MΩ and are rated at 200V dc at 0.25W. Temperature range is –55°C to

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125°C. BI Technologies Ltd. Tel., 0116 2781133; fax, 0116 2781199.

Transformer core kits. Philips offers a range of planar E cores in kits to simplify design and development of low-profile transformers for switched-mode power supplies, in which the 'windings' are printed on the board. Thermal resistance is reduced by about 50% over conventional wirewound units. Hawnt Electronics Ltd. Tel., 0121-784 3355; fax, 0121-783-1657.

Latching solenoids. Permanent magnets are used in Densitron's new range of solenoids for a latching action without the use of holding power. For some applications, medical or chemical processes, for example, the lack of power supply and therefore heat is essential. The solenoids run at high speed and stop immediately when power is removed with very little temperature increase.

Low-*Z* electrolytics. Samwha's *RZ* range of electrolytics show very low impedance at high frequencies, a 200µF, 50V type having an impedance of 0.075Ω at 100kHz and 20°C. Case size is only 10mm diameter by 25mm, but allowable ripple is 1040mA at 105°C and 100kHz. Under those conditions, its life is 5000 hours, but at 7°C it would be 40,000 hours. Range of values is 1-15000µF at ratings between 6.3Vdc and 63V dc. All values are to within ±20% and components are protected against cleaning fluids. Samwha Electric UK. Tel., 01748 850555; fax, 01748 850556.

Densitron Europe Ltd., 01959 700100; fax, 01959 700300.

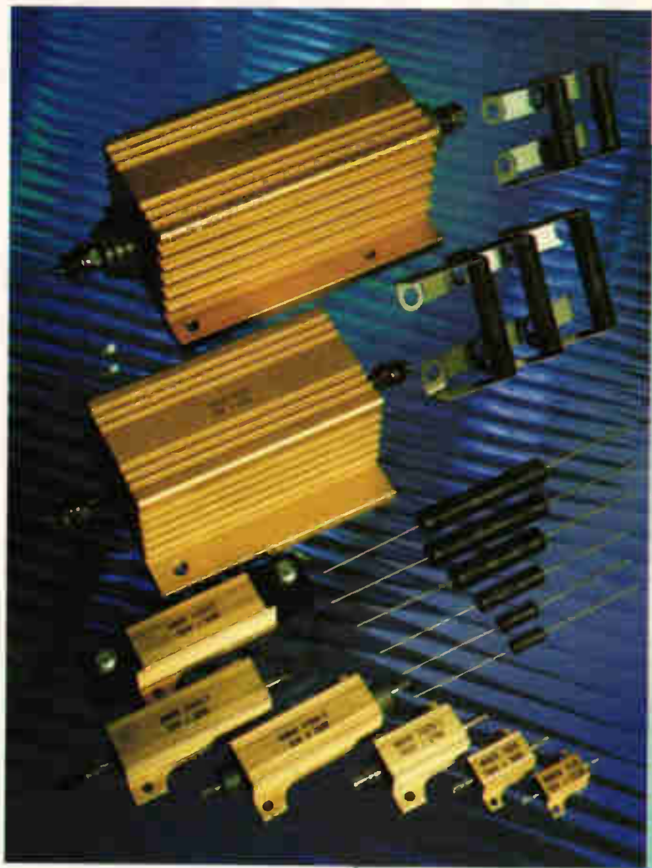
Ntc thermistors. *Shibaura NTC* thermistors are now available in the UK from BFI Ibexsa. The thermistors are for use at high temperatures up to 500°C and use a fine ceramic chip and glass coating, which makes for stability and reliability. BFI IBEXSA Electronics Ltd. Tel., 01622 882467; fax, 01622 882469.

Surface-mounted inductors. Magnetix *CM32/45* general-purpose, s-m inductors are in standard 1210 and 1812 packages, are wirewound and epoxy encapsulated. The range of values is 0.1-1000µH in 10% E6 values, E12 to order. They withstand wave and flow soldering and come taped and reeled. Anglia. Tel., 01945 474747; fax, 01945 474849.

Axial inductors. Total Frequency Control makes a range of miniature inductors on ferrite cores with axial leads, the *EC36* range, which covers the 0.1-1000H range of values. Termination technique used ensures mechanical strength and an epoxy coat protects the devices against humidity. Leads are normally straight, but can be preformed, and the inductors are available loose, taped and reeled or 'ammo' packed. Total Frequency Control Ltd. Tel., 01903 745513; fax, 01903 742208; e-mail, eddie@tfc.co.uk.

Chip inductors. *Type LEM* wirewound, surface-mounted chip inductors by Taiyo Yuden handle currents up to 720mA and are made in hf, high-current and standard versions from heat-resisting resin. Self-resonant frequency of the hf type is up to 4.3GHz with *Q* at 1GHz to 60 and in values from 10nH to 100nH at ±10% or ±20% tolerance. The miniature high-current inductors have values in the 1-33µH range to the same tolerance and take up to 475mA; srf is a minimum of 165MHz at 1µH and *Q* 10-20. In the standard version, values are 0.12-220µH with the addition of 5% tolerance, current 720mA and srf over 600MHz at 0.12µH; *Q* is 10-50. Taiyo Yuden UK. Tel., 01494 464642; fax, 01494 474743.

Carbon-film resistors. Including the *RKL2* at 0.8mm by 2.2mm, the *RKL* range of miniature carbon-film resistors have no metal caps and have the leads soldered in with high-temperature solder. Power ratings are from 35mW to 70mW at 70°C and working voltages 60-100V. The smallest *RKL2* range of components comes in 5% values from 24Ω to 200kΩ in E24 and 10Ω to 10MΩ in E12 10% values; *RKL5* resistors have higher values up to 22MΩ. Long-term stability is better than ±2%. Rhopoint Components Ltd. Tel., 01883 717988; fax, 01883 712938.



Power resistors. Working voltages to 2.3kV in harsh surroundings are featured in the *ATE* family of precision, wirewound resistors, designed for reliable operation over the range –55°C to 350°C. There are silicone-coated types in resistances from 0.01Ω to 82kΩ, rated at 3-15W, and a range in extruded, gold-anodised aluminium housings in values to 120kΩ in ratings from 5W to 250W, all available in 1% tolerance and all meeting or exceeding the relevant MIL-R specs. Kestronics Ltd. Tel., 01727 812222; fax, 01727 811920.

Connectors and cabling

Optical-fibre contacts. *BPO* optical contacts and accessories for a single multi-mode optical fibre are intended for use with high bit rate or broadband telecomms, military and broadcast systems where signals must be taken through equipment back panels. The BPO contacts, an extension of the company's EC range, snap into DIN41612 housings in 3U Eurocard and 6U pcb backplanes and can be combined with coaxial and signal contacts in the same housing. Single-mode insertion loss is 0.5dB and return loss over 60dB. Transradio Ltd. Tel., 0181 997 8880; fax, 0181 997 0116.

Digital video cable. Alcatel *Serial Digital Video Cable* is smaller than usual in diameter, has a 0.6mm solid plain copper core and cellular polyethylene insulation, shielded with aluminium/Petp tape and a 0.15mm tinned copper wire braid. Outer is a low-smoke, zero-halogen jacket. Capacitance is 56pF/m, resistivity 62.5Ω/km. Wadsworth Electronics Ltd. Tel., 0181-268 6500; fax, 0181-268 6565.

Flexible circuit connectors. A rotary locking mechanism eliminates poor mating and reduces assembly time on a range of high-density, low-profile, surface-mounted connectors by JAE for the connection of flexible circuits to boards. *JAE IL-FHR* 0.5mm pitch connectors contain 6-50 contacts and protrude only 1.9mm above the board. The locking mechanism provides rapid, zero-force insertion, apart from making for increased security in the presence of shock and vibration. There are straight and right-angled variants and a type to fit into a cut-out in the side of the board to make the assembly even thinner. Ratings are 50V ac and 500mA, contact resistance being 20mΩ. JAE Europe Ltd. Tel., 01276 21717; fax, 01276 66165.

Crystals

Quartz and ceramic resonators. A full range of frequency-standard components from *Railtron* comprises everything from simple chip resonators to rubidium and caesium frequency standards. Ceramics cover the 2-6MHz frequency range to within ±3% in a variety of mounting styles, while in crystals there are ultra-thin

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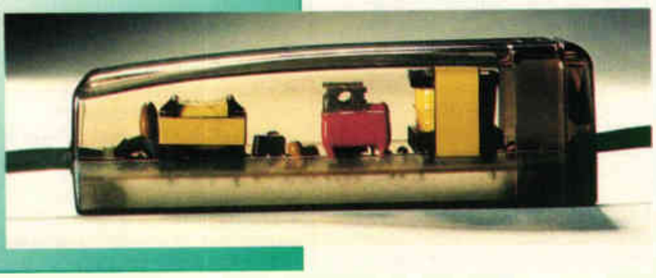
devices, clocks, oscillators and crystal units to 622MHz vxcos, 500MHz txcos, oxcos and filters. High-accuracy crystals are made in the range 800kHz to 360MHz to between ±3ppm and ±50ppm. Sematron UK Ltd. Tel., 01256 812222; fax, 01256 812666.

Ceramic resonators. Murata's CSA, CSAC and CST ranges of *Ceralock* ceramic resonators are only about a third the price of crystals which, unsurprisingly, makes them popular in clock generators. They can also be used in the updating of older designs and some are available with built-in capacitors. Frequency range is 1.26MHz to 50MHz in leaded and s-m versions to a tolerance of ±0.25% over –40°C to 125°C. Surtech Interconnection Ltd. Tel., 01256 351221; fax, 01256 471180.

Displays

Edge-lit panels. *Comled* edge-lit displays by Comtronic are intended for use in aircraft, road vehicles, trains and the like, offering improved performance over filament or electroluminescent types. Their advantages include low power consumption, long life, variable light output and blue, green, yellow, orange or red illumination. Any light source from 2V to 120V ac or dc may be used for readability in sunlight. Acal Electronics Ltd. Tel., 01344 727272; fax, 01344 424262.

'Digital' halogen lamp driver. Universal Transformers' *Digital Driver* halogen lamp controller is said to eliminate many of the drawbacks of conventional drivers. It regulates lamp voltage to 11.8V maximum over a wide input span. The importance of this? A 10% overvoltage is said to cause a 40% reduction in lamp life. And power loss is reduced from 20% in a conventional high-frequency transformer to 10%, reducing heat stress and saving energy. The driver's life is 15 to 20 years whereas traditional controllers have a life expectancy of 5-7 years. Although it has a regulated output, the unit can still be used with a conventional dimmer. Capable of driving lamps to 60W, *Digital Driver* is mains driven, isolated, and overload protected. It measures 124 by 42 by 39mm. Universal Transformers, 2 Hurworth Road, Newton Aycliffe, Co Durham DL5 6UD. Tel. 01325 317429, fax 01325 311081.



Test and measurement

Waveform generator. Waveforms from the TTi *TGA1230* arbitrary waveform generator may be specified by front-panel editing controls, from a digital storage oscilloscope or by down-loading waveforms from a pc; Windows-based software is provided for the creation and editing of waveforms on a pc. Frequencies in the range 1mHz to 10MHz may be displayed and defined with 12-bit vertical resolution and 4-65536 horizontal points. A ram will store up to 50 waveforms. The *TGA1230* also operates as a conventional function generator, providing sine, cosine, haversine and havercosine signals in the 1mHz-10MHz range and square waves to 15MHz. It will also generate triangle, ramp and sin*x*/*x* waveforms. Several types of modulation can be applied. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Hand-held spectrum analyser. Meant for installation and servicing work, the *Model 3200 RF Field Analyser* marketed by Datalines works on signal in the 100kHz-2GHz range. It is powered by rechargeable batteries, has a liquid-crystal display and an RS-232 interface to allow displays to be saved on a pc. Datalines Communications Ltd., Tel. and fax, 01908 370011.

High-voltage probes. Test and Measurement Marketing can supply a range of instrument probes to cope with voltages up to 40kV, with a bandwidth of 100MHz in the *PHV 4000-3*, or 6kV at 400MHz for the *PHVS* series. All the probes are adjustably frequency compensated and attenuation is 60dB, other ratios being supplied to order. Input impedance is 50MΩ and 2pF for the *4000-3* or 50MΩ and 6pF for the *PHVS*. Test and Measurement Marketing. Tel., 01626 333455; fax, 01626 53053.

CE wattmeter. Feedback's *EW1604* electronic wattmeter, already in common use worldwide, has been upgraded to allow it to exceed the European standards and obtain the CE mark. Its power range is 250mW-10kW fsd at 0-20kHz. Input voltage is 5-1000V and current 50mA-10A, for any waveform. Feedback Test and Measurement. Tel., 01892 653322; fax, 01892 663719.

FFT plug-in. Tektronix has introduced the *TDS2MM*, a plug-in module for the *TDS200* range of real-time digital oscilloscopes. The module adds fast Fourier transform analysis and the functions of automatic rise-time, fall-time, positive and negative pulse-width measurement to the oscilloscopes and provides RS232, Centronics and GPIB ports for printing and remote control. Tektronix UK Ltd. Tel., 01628 403300; fax, 01628 403301.

Literature

Reed relays and sensors. New brochures from *Crydom* are concerned with ranges of reed relays and sensors for flow and level measurement. Reeds are available in rf types for 0-30MHz and in high-voltage versions for 0-15kV, in addition to those for general-purpose use. Flow switches provide pressures to 10bar, switching up to 1.5A. Crydom. Tel., 01202-897 964; fax, 01202-891 918.

New from Fluke. Fluke has the latest catalogue of test gear and data acquisition equipment, and also can supply several other catalogues covering digital multimeters, Scopemeters, the new Industrial Scopemeter 123 and others in their range. Fluke UK Ltd. Tel., 01923 240511; fax, 01923 225067.

Cs and Rs. Capacitors and resistors marketed by Hawnt are the subject of a product guide, in which are described components from seven manufacturers: Arcotronics, AVX, Beyschlag, Nippon Chemi-Con, Welwyn and Wima. There are selection charts for leaded capacitors and resistors and separate section for surface-mounted devices. Hawnt Electronics Ltd. Tel., 0121-784 3355; fax, 0121-783 1657.

Livingston Rental. Livingston presents a brochure on *Tektronix* oscilloscopes and logic analysers, the oscilloscopes including exotic machinery with a bandwidth of 8GHz, sampling at 50Gsample/s and with Instavu acquisition. The logic analysers handle clock speeds to 200MHz. Livingston says it can deliver any of them, usually within 24 hours, for short or long-term hire. Livingston Rental Ltd. Tel., 0181-943 5151; fax, 0181-977 6431.

Rf connectors and assemblies. 3M manufacture connectors and cable/connector assemblies to work at rf in telecomms, GPS and consumer equipment. Surface-mounted female connectors, supplied in tape and reel packaging, have a low c.g. to ease vacuum placing on the board and contacts for wave soldering or ir temperatures. The assemblies are for 50Ω use up to 6GHz or 75Ω to 1GHz and use shielded, low-profile, straight or right-angled plugs, with cables such as RG178, which is a shielded pliable cable, or a 1mm diameter type, or several other choices. There is also a shielded board-to-board adaptor. Arrow-Jermyn. Tel., 01234 270027; fax, 01234 214674/791501.

Rf test sockets. Aries Electronics offers a brochure on its range of sockets for manual, automatic and burn-in of rf devices, both surface-mounted and through-hole types. The sockets are precision machined and use the company's patented Microstrip contacts, which give minimum signal loss and long life. Electrical characteristics, mounting and associated adaptors are shown and described.
Aries Electronics (Europe). Tel., 01908 260007; fax, 01908 260008.

Switches. EAO-Highland has a 34-page catalogue of all its switches, indicators and pcb-mounted components, including emergency stop switches, led indicators, keylocks, selector switches, membranes and buzzers. There is also information on linear power supplies and alarms made by EAO. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Please quote *"Electronics World"* when seeking further information

Materials

Sticky labels. CILS has a large range of labels that will not turn a hair when attacked by IR flow, wave soldering, pc board washes, cleaning agents, solvents, oils, water and humidity and that cope with temperatures from −80°C to 400°C. Labels come in a variety of materials and with adhesives, both permanent and removable, for various surfaces; there are also those which indicate that someone has tried to remove them. Any size or shape of label can be made, with bar codes, serial numbers and the like and they will take additions by laser, dot matrix, thermal transfer or inkjet printer, if required. Computer Imprintable Label Systems Ltd. Tel., 01273 681000; fax, 01273 681144.

Power supplies

High voltage. The *597* range of high-voltage supplies by Brandenburg delivers up to 10W from two precise outputs of ±5kV or ±10kV with respect to ground. Outputs are controllable from zero to the maximum by either a control voltage or by front-panel controls and can be made to track each other for a balanced supply. An lcd shows voltage and current. Brandenburg. Tel., 01384 393737; fax, 01384 440777.

Long-life dc-dc converter. The *MacroDens PKF2000* 6W converters are said to have a mean time before failure of 4.9 million hours, which should see them still fairly sprightly in February, 2557, give or take. Efficiency is 84%, inputs 18-36V dc and outputs 3.3V dc or 5.5V dc. Units take up 47.8 by 23.5mm of board space and are only 8mm thick. Casing is epoxy and the units meet all the relevant standards. Campbell Collins Ltd. Tel., 01438 369466; fax, 01438 316465.

Programmable 300W. *Xantrex HPD* switchers, a family of 300W interfaced power supplies, are fully protected against the usual disasters, use remote sensing and have a 10-turn pot. for accurate control of voltage and, as an option, current outputs. Noise and ripple over 20Hz-20MHz are less than 1mV rms and 2mA rms. RS232 and GPIB interfaces may be used and a single-address multichannel interface allows single-channel IEEE-488-to-multichannel optical-fibre control of up to 31 units. The interface, when installed, activates a front-panel led. Three models give various combinations of 15V/20A, 30V/10A and 60V/5A. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

50W dc-dc converters. Astec's *AA50A* 50W converters are meant for use in inhospitable positions in industrial electronics. Input is 20-72V dc, outputs on single models being 5V/10A, 12V/4.2A and 15V/3.4A. There are triple-output versions to give 5V/6A, 12V/0.75A and −12V/0.75A and a model with 15V and −15V instead of the 12V outputs. Internal output filtering reduces reflected ripple. The case measures 4 by 2.6 by 0.5in. Chloride Powerline. Tel., 0118 9868567; fax, 0118 9755172.

Pc power supply. Tri-MAP's *SP2-4400* 400W supply fits the PS/2 arrangement and has four outputs: 5V, 50A, 12V, 20A and, for analogue expansion boards, −5V and −12V, both 1A. The unit has IEC mains input and monitor output connectors and a remote power switch to mount on the pc case. Input is a universal 90-264V ac and two autosensing ranges are switch-selected. All relevant directives are complied with. Tri-Map International Ltd. Tel., 01705 424800; fax, 01705 424801.

High-energy cells. Varta can offer more models in its range of nickel-metal-hydride cells, which now contains eight types in capacities from 580mAh to 4000mAh. The new models include the *VH650*, which is an AAA cell and the *VH4000*, a 4/3 type giving the highest capacity in the range and meant for use in laptop computers. Varta Ltd. Tel., 01460 73366; fax, 01460 72320.

Voltage regulators. Five new regulators in the Zetex *ZR78L* family, which are enhanced versions of the standard 78L type, provide outputs of 4.85, 5.2, 7.0, 8.5 and 12.0V, all delivering 200mA in SOT23 or TO92 chips instead of the more commonly used larger packages. Quiescent current is 350µA and initial voltage output tolerance is within 2.5%. Thermal overload shuts the regulators down, recovery automatically taking place on cooling. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5467.

Radio communications products

Hf ssb transceiver. The *AEL 2020* high-frequency, single-sideband transceiver has ten programmable channels and is suitable for mounting on a vehicle or as a fixed station for operation by unskilled users. Frequencies, operating mode, emergency channel and antenna socket selection are programmable at the factory or by a dealer using pc software. Two individually tuned half-wave dipole base station antennas or two whips for two-channel networks connect to two separate sockets. Audio processing increases 'average talk power' by 6dB over earlier models. Maximum transceiver output power is settable between 25 and 125W pep. South Midlands Communications Ltd. Tel., 01703 255111; fax, 01703 263507.

Protection devices

Motor protection. *Motor Load Control* from Load Controls Inc. prevents damage to industrial electric motors. It is surface mounted and shows true power, including power factor, on a led, high and low power trips with adjustable delays and trip timers being standard; the high-power trips handle overload, jamming, bearing trouble and blockages, while the low-power one detects low load or no-flow conditions. A 4-20mA analogue output and individual trip relays allow pump or motor control and process performance monitoring. The unit is compatible with three-phase motors up to 240kW on 230-460V ac. Vidas International Marketing. Tel., 01252 718466; fax, 01252 718467.

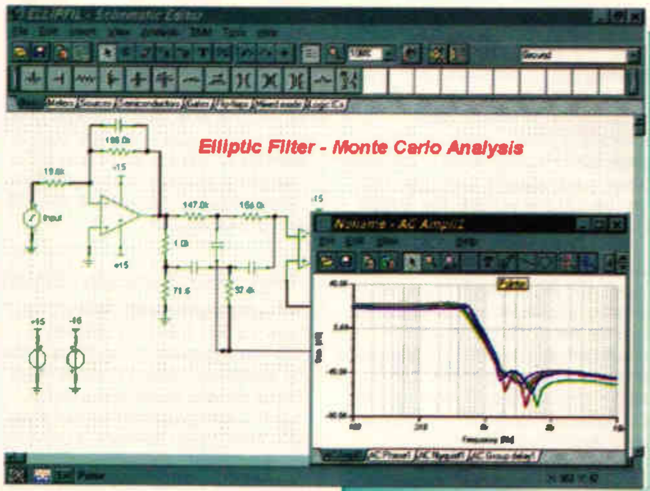
Voltage suppressors. Semtech offers a new line of transient voltage suppressors, the *SMS05C/12C/15C/24C family*, which protect five unidirectional and four bidirectional lines, having common anodes. Protection for data, signal and supply lines meets IEC 1000-4-2 Level 4 (ESD) and IEC 1000-4-4 (EFT). Peak pulse power is 300W and clamping voltage 9,8,19,24 or 43V; reverse standoff voltage is given by the designation. Package is the SOT-23. Semtech Ltd. Tel., 01592 773520; fax, 01592 774781.

Switches and relays

500A inrush relays. Three relays by Eberle take inrush current of 100-500A. *40947* has twin contacts: first to touch is of tungsten and takes the inrush, while the silver cadmium oxide second one does the switching; inrush current for this one is 500A for 10µs. *40957* has one contact of silver tin oxide to handle 120A for 20ms; this and the *40947* having 720Ω coils for 24V dc. Lastly, the *42903* has a coil of 1400Ω at 24V dc and lower power consumption of 0.41W, its single contact taking 100A for 20ms. Radiatron Components Ltd. Tel., 01784 439393; fax, 01784 477333.

Little relay. *ASPF Series* single-pole, solid-state relays by Crydom take up a mere 0.63in^2 of board space and yet perform reliably with loads of 3A, having an integral heatsink, the Powerfin. Packages are single in-line and zero and random-switching versions are available. Transient overvoltage is 600V pk. Power switching is by back-to-back scrs to provide high d*v*/d*t* rating to eliminate the need for a snubber and higher current and thermal ratings. Control voltage is 4-10V dc and output operating voltage 12-280V rms. Crydom. Tel., 01202-897 969; fax, 01202-891 918.

Telecomms relays. *PVT322* by International Rectifier is the first series of two-pole commercial electronic relays to need no load derating when both channels conduct. Maximum load voltage is ±250V dc or ac pk, and max. load current 170mA/channel at a resistance of 10Ω maximum. Actuation needs only 2mA of led current and i/o isolation is 4kVrms. The relays are in the standard 8-pin



Better Tina. *TINAplus* is an improved version of Tina, the circuit design, simulation and analysis software package, coping with analogue, digital and mixed circuitry. New to this version are faster algorithms for mixed-mode and larger circuits, new component models and an 'immediate' mode for testing incomplete circuits. There are over 4000 components including cmos and ttl in the standard, user-expandable library and tools and hot keys for simpler operation. Analysis includes ac, dc and transient analysis, Fourier, digital step-by-step and timing analysis, noise, pole/zero analysis and many more. A range of virtual instruments is also included to drive and test simulated circuits. Tandem Technology Ltd. Tel., 01243 576121; fax, 01243 576119.

dip plastic package, with either through-hole or gull-wing pins. International Rectifier. Tel., 01883 732020; fax, 01883 733410.

Transducers and sensors

Weighing indicator/controller. To be used as the link between load cell and indicator, these devices from Control Transducers will also resend output data to a pc. *Model L1* has an isolated power supply, the readout being specified in the user's units, together with single or dual set-points and resend outputs in voltage or 4-20mA current loops. The company says that you can install the unit according to tailored instructions, confident that it will work as planned. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

Cable extension meter. Celesco's *Model PT8600* transducer is an industrial linear-to-rotational device calibrated to Dynapar encoders. In this type, the rotation is obtained by extending a stainless steel cable from the casing, up to 60in in length; as the cable extends, the encoder rotates to give the position of the cable end, which is fixed to the object in question. A spring motor maintains tension in the cable. Variohm Components Ltd. Tel., 01327 351004; fax, 01327 353564.

MMX notebook. *IN Lite* is claimed to be the first Pentium MMX ruggedised notebook to become available in the UK. It has either a 10.4in tft display that can be seen in sunlight or a 12.1in tft-XVGA type for a wide viewing angle. A lockable metal case contains a full AT format slot and a built-in mains power supply/charger. Keyboard is detachable and dos users will appreciate a new graphics chip that allows the whole screen to be used. Kontron points out that this equipment is certified for flight-testing use by Boeing. Kontron Elektronik Ltd. Tel., 01923 421521; fax, 01923 254118.

COMPUTER

Data acquisition

Data cards. Fourteen new data acquisition cards by LeCroy for ISA and PCI buses provide sampling at u to 80Msample/s at 12-bit resolution and 8-bit sampling at 500Msample/s, with memory lengths to 16Mbyte per channel. PCI versions of the *SigAcqCard* are meant for users wanting data with a high pulse repetition frequency, transferring the data from on-board memory to storage in microseconds at over 100Mb/s. All the cards give a high s:n ratio and the design is such that the frequently met 'hostile environment' does not affect the ratio. Software drivers are available in a number of forms, including the various incarnations of Windows, LabVIEW and LabWindows. LeCroy Ltd. Tel., 01189 344882; fax, 01189 348900.

Development and evaluation

Flash development. Smart supports users of its asynchronous flash memory modules with the introduction of programming tools to develop code for simm and dimm modules. *Code Developers' Kits* give read/write capability for the modules so that users can develop test code on a pc. Kits contain a flash simm or dimm adaptor, programmers' guide, a 4MB flash memory, a PCMCIA reader/writer connecting to a pc ISA bus and all software. Kits support flash modules from all major suppliers. Smart Modular Technologies. Tel., 01908 234030; fax, 01908 234191.

Mass storage systems

192Mbyte flash. A single-height VMEbus flash memory from PEP, the *VNEM-F1*, has a capacity of 192Mbyte per module, making it the largest VMEbus board in 3U form. There are 64Mbyte on the base board and another 128Mbyte on a piggyback module, write protection being set in banks of 4Mbyte and writing being possible in blocks of 128Kbyte. PEP Modular Computers. Tel., 01273 441188; fax, 01273 441199.

Software

Fuzzy logic design. *Fuzzy Logic Design for G* is announced by National Instruments. This is an interactive design software package for fuzzy control applications that can build virtual instruments for BridgeVIEW (G being its graphical development language) and LabVIEW software and is also compatible with National's data and image acquisition hardware in industrial automation applications. Systems are designed by a point-and-click process, accessing what is virtually an expert system. National Instruments UK. Tel., 01635 572400; fax, 01635 524395. ■

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Access

VISA

Seeing through noise

Patrick May's analyses reveal that we may not know all there is to know about 1/f noise in semiconductors, and they provide an insight into low frequency noise in general.

Analysing the effect of diffusion of carriers instigated by the recombination of holes and electrons at a semiconductor surface using mathematics reveals a component of carrier concentration that varies with time.

When electrons and holes recombine at the surface of a semiconductor, equal electron and hole diffusion currents result. The magnitude of the diffusion current does not depend on concentration gradients within the electron/hole systems in the bulk. It is governed by an independent system – the 'fast surface states' in the surface.

An interesting discovery is made when the solution to the diffusion equation defining the problem is examined closely. In a layer near the surface – or interface – the concentration of holes and electrons is reduced.

The amount by which this concentration is reduced varies with time. What is more, the square of the modulus of the Fourier transform of this time varying function has a spectrum defined by,

$$\frac{1}{\sqrt{\frac{1}{\tau^2}+\omega^2}}$$

where τ is the lifetime of carriers near the surface and ω the angular frequency. From now on, I will refer to this as 'surface diffusion noise'. It describes the low-frequency noise in silicon junction field-effect transistors, or j-fets, and when $\tau \rightarrow \infty$, it becomes the well-known $1/f$ noise: it solves the mystery regarding the absence (Buckingham, 1983) of $1/f$ noise in silicon junction fets and provides an explanation for $1/f$ noise in other devices.

Surface diffusion noise is, therefore, a universal source of noise in semiconductors. A brief mathematical analysis of the special case when the lifetime of the carriers is infinite has already been presented (May 1988).

Mathematical analysis

The mathematical analysis appertaining to the diffusion process is based on the continuity equation,

$$\nabla \bullet J = -\frac{\partial \rho}{\partial t} \qquad (1)$$

where J is the current density and ρ is charge density.

In order to apply the analysis to the metal-oxide silicon transistor, the gate is visualised as a strip of n-type crystalline semiconductor, **Fig. 1**.

The thickness of the strip is of the order of 3μm. The top face of the strip represents the oxide/semiconductor interface into which the electrons or impulses are ejected from the bulk. The diffusion current which results, is in one dimension and perpendicular to the surface or interface if the surface states are evenly distributed.

If the current density, J, is assumed to be a diffusion n type current, eqn. (1) for currents in the x-direction perpendicular to the main faces becomes,

$$\frac{dn}{dt} = \mathrm{D}_n \frac{d^2 n}{dx^2} \qquad (2)$$

where $n=n(t,x)$ is the electron density and D_n is the diffusion constant. The solution to eqn. (2) for a unit impulse is given by,

$$n = \frac{1}{(\pi D_n t)^{\frac{1}{2}}} \exp\left[-\left(\frac{x^2}{4D_n t}\right) - \left(\frac{t}{\tau_n}\right)\right] \qquad (3)$$

where τ_n is the life time of the electron carriers. The square of the modulus of the Fourier transform of n is,

$$|N(\omega,x)|^2 = \frac{1}{D_n\sqrt{\left(\frac{1}{\tau_n^2}+\omega^2\right)}} \exp\left[-x\sqrt{\left(\frac{2}{D_n}\right)}\sqrt{\left(\frac{1}{\tau_n}\right)+\sqrt{\frac{1}{\tau_n^2}+\omega^2}}\right] \qquad (4)$$

For the condition,

$$x\sqrt{\left(\frac{2}{D_n}\right)}\left(\sqrt{\left(\frac{1}{\tau_n}\right)+\sqrt{\frac{1}{\tau_n^2}+\omega^2}}\right) < 1 \qquad (5)$$

in expression (4) the approximation below applies,

$$|N(\omega,x)|^2 = \frac{1}{D_n\sqrt{\left(\frac{1}{\tau_n^2}+\omega^2\right)}} \qquad (6)$$

when the electrons are strong majority carriers ($\tau_n \rightarrow \infty$) expression (5) becomes,

$$x\sqrt{\left(\frac{2\omega}{D_n}\right)} < 1 \qquad (7)$$

and expression (6) becomes,

$$|N(\omega,x)|^2 = \frac{1}{D_n\omega} \qquad (8)$$

Corresponding expressions for hole carriers are obtained in a similar manner.

At the surface – or silicon/silicon oxide interface – the electron diffusion current is made up of impulses due to the electrons. Each ejected electron, or impulse, reduces the electron concentration near the silicon/silicon oxide interface at the surface by the amount expressed in (3) which is a function of time.

The square of the modulus of the Fourier transform of this function gives the spectral characteristic expressed in equation (6).

On n^2, the time varying component is many orders of magnitude lower than n^2 in the sample. It depends on the spectral expression given in equation (6), the rate of surface recombination per unit area, v, and the thickness of the sample, L. It is given by

$$S_n(\omega) = \frac{J}{D_n L}\frac{1}{\sqrt{\frac{1}{\tau_n^2}+\omega^2}} \qquad (9)$$

If $S_v(\omega)$ is the power spectral density when the mean square of the voltage across the sample is $\overline{V}^2$ then,

$$\frac{S_v(\omega)}{\overline{V}^2} = \frac{S_n(\omega)}{n^2}\frac{1}{n^2 D_n L}\frac{1}{\sqrt{\frac{1}{\tau_n^2}+\omega^2}} \qquad (10)$$

The expression in equation (10) becomes that for 1/*f* noise when $\tau_n \rightarrow \infty$.

Relevance of the mathematical analysis

Note that the expression in equation (3) is capable of being fully integrated (van der Ziel, 1979). As a result, Carson's theorem (van der Ziel, 1979) can be applied so that equation (3) leads to the conclusion in equation (10).

Also, the noise expressed in equation (10) is not inversely proportioned to N – the total electron population – but is inversely proportioned to the thickness. This corresponds with an important observation made by Weissman, 1988.

Prevailing theories on the origin of 1/*f* noise are based on the well established premise that the noise is produced by fluctuations in conductivity (Voss et al, 1976). The carrier number theory first proposed by McWhorter, 1957, has been modified in various ways. In essence, it relies on the precept that surface states act as carrier reservoirs. These reservoirs allow an interchange of carriers between the bulk – where they contribute to conductance – and the surface where they are held immobile in the surface states.

The interchange is assumed to occur in 'pulses'. A suitable weighting of the time constants gives rise to the 1/*f* noise spectrum. This implies that all the carriers released by the surface states take up their positions in the bulk so as to contribute to a change in conductance which faithfully follows the dictates of the original pulse. No attempt has been made to explain how this is accomplished. It also implies that these carriers are immune to the process of diffusion produced by the gradients of concentration of the 'injected' carriers.

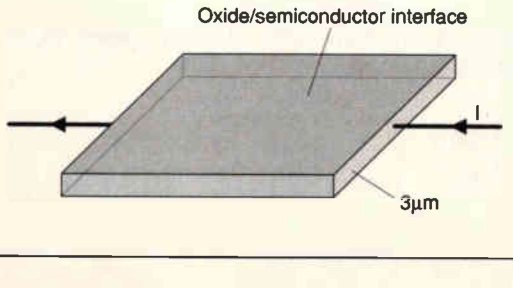

Fig. 1. The n type semiconductor strip represents part of a metal-oxide-silicon transistor. Holes and electrons diffuse through the strip to the oxide/semiconductor interface where they recombine.

Bearing in mind the necessary long time constants of the pulses, these implications detract from the credibility of McWhorter type models. The surface diffusion theory I propose here takes the opposite stance; here the short pulses produced by the 'fast surface states' give rise to a diffusion process which is itself responsible for the production of the 1/*f* noise spectrum when the diffused carriers are strong majority carriers.

Evidence that there is correlation between 1/*f* noise and fast surface states has been repeatedly reported (Sochava et al, 1960; Macrae, 1962; Klaassen, 1971; Broux et al, 1975) and the lack of correlation between the slow surface states and 1/*f* noise has been recorded (Sochava et al, 1960).

Practical implications

The validity of the proposed theory is made evident when it is applied to semiconductor devices by using the criteria in expressions (5), (6), (7) and (8) to explain the presence or otherwise of 1/*f* noise in common semiconductor devices.

Enhancement mode mos transistors produce nearly pure 1/*f* noise when operating in their linear mode. This is because the carriers in the channel are strong majority carriers giving rise to the spectrum expressed in equation (8).

Equation (7) indicates that there is an upper frequency 'cut off'. This effect has been observed before (Freight *et al*, 1971). If the diffusion constant, D_n, is taken as $4.4\times10^{-3}m^2s^{-1}$ in expression (7), the upper frequency limit for a channel thickness of 2.5µm is greater than 50MHz. In practice, the channel thickness is less than 2.5µm.

When the mos transistor is biased into saturation, the carriers in the 'throat' are no longer strong majority carriers and their lifetime is finite. This gives rise to the band limited white noise described in expression (6). When this component is added to the 1/*f* noise spectrum the typical noise characteristic with a 'bump' in it results.

When forward biased, the p-n junction diode produces 1/*f* noise as well as other superimposed components (Kleinpenning, 1978). In this case the 1/*f* noise is produced in the emitter where majority carriers prevail and the criteria in expressions (7) and (8) apply.

As the current enters the depletion region, the criteria expressed in (5) and (6) apply. Hence the spectral shape due to both components is similar to that of the mos transistor in saturation.

Silicon field effect transistors do not produce significant 1/*f* noise (Buckingham, 1983). Yet according to the empirical formula given by Hooge (Buckingham, 1983; Hooge, 1969) the j-fet is an ideal medium for 1/*f* noise.

The surface diffusion noise theory proposed here precludes the presence of 1/*f* noise in the device. The channel is bounded by two depletion regions so its carriers are of finite lifetime. This means that the criteria in expressions (7) and (8) do not apply. But the general diffusion noise as described in expression (6) does apply.

Evidence for the theory

Buried-channel metal-oxide silicon transistors are ideal for illustrating the validity of the surface diffusion theory proposed here because their channel properties can be altered by adjusting the gate source potential, V_{GS}.

The way in which the noise spectrum in an n-channel buried-channel mos transistor varies with V_{GS} has been studied experimentally (Carruthers et al, 1987). For positive values of V_{GS}, the observed behaviour is similar to that in enhancement (surface channel) mos transistors; it has a strong 1/*f* noise characteristic.

When V_{GS} is positive, the channel is very close to the silicon/silicon oxide interface and has a strong electron concentration with $\tau \rightarrow \infty$. This results in the $1/\omega$ spectrum described in expression (8).

When V_{GS} is made negative, the observed spectral behaviour is similar to that of the junction fet. A depletion layer separates the channel from the silicon/silicon oxide interface and the channel itself is bounded by depletion layers. The electron concentration in the channel is decreased and the carrier lifetime takes a finite value, τ.

The spectrum takes the form described in expression (6) as is the case for the junction fet. Note that low negative values of V_{GS} produce a thin depletion layer at the semiconductor/oxide interface, which gives rise to Lorentzian spectra attributed to Generation-recombination (Buckingham, 1983).

In summary

Surface diffusion noise is a universal term. It accounts for both $1/f$ noise and the low frequency type of noise present in junction fets. Low-frequency noise is also prevalent in other devices. It usually appears when part of the channel in the device has carriers with finite lifetime and has been passed off as a bump in the $1/f$ noise characteristic.

I hope that this discussion of the mechanism of low-frequency noise will lead to improvements in the way that devices are designed, and that it will contribute to a better understanding of 'fast surface states'.

My thanks go to the University of Exeter and its School of Engineering for providing library facilities. ■

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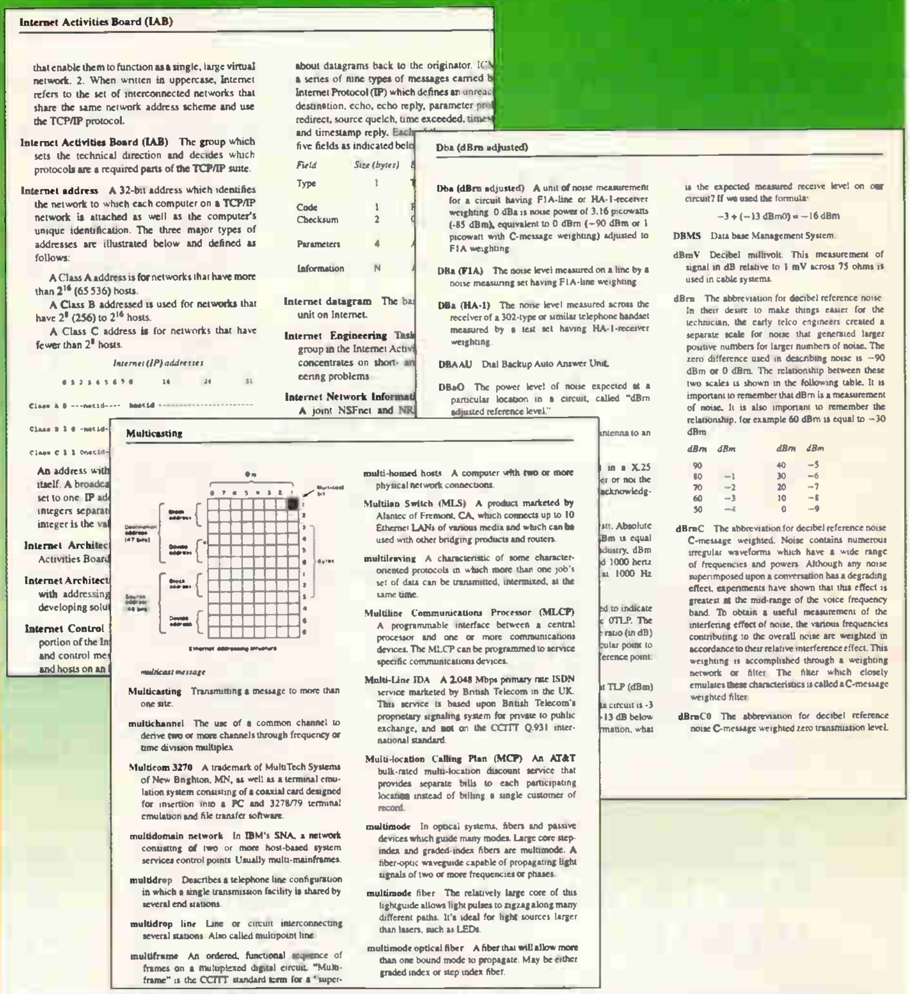

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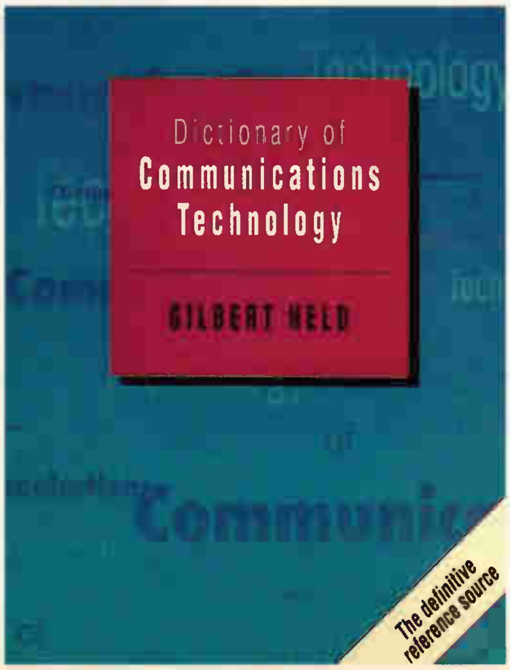

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LETTERS

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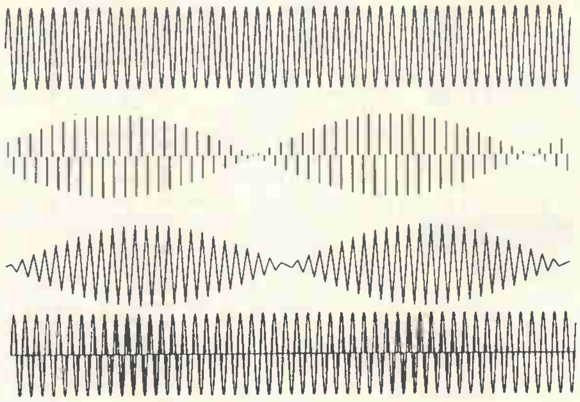
Beating frequencies

Chris Bullman's letter on sampling a 19.999kHz signal at 44kHz in the September issue prompted an overwhelming response. There follows a sample of the first few that came in. Space permitting we will present more in the next issue.



What you see is not what you thought. The signal can be reconstructed from the samples – but the samples are not the signal.

Chris Bulman raises the question of sampling a signal at close to the Nyquist frequency, and how this introduces what looks like amplitude modulation, see diagram on right.

The answer to Mr Bulman's question lies in his own letter, where he says "unless the filter had a time delay of over a second". The filter required to reconstruct a signal as close to Nyquist as in the example will indeed have a ringing response that lasts over a second.

The Nyquist theorem only says that if you sample fast enough the samples contain information for the original signal to be reconstructed. It does not say that the samples will like the original signal.

The proper reconstruction involves passing the sequence of samples, as impulses, through a low pass filter with cutoff at the Nyquist frequency. Each impulse hitting the filter causes ringing. The sum of these successive rings reconstructs the signal. If the signal frequency is very close to the Nyquist, then the filter has to have a very sharp cutoff indeed – so it will have a very long impulse response, hence the 'memory' needed to carry over into regions where the signal 'looks' small.

Chris Bore
Woking
Surrey

Chris Bulman's letter in your September issue, in which he postulates 'beating' as a result of sampling a 20kHz signal at 44.1kHz set me thinking. This, plus the recent equipping of my workshop pc with a sound card stirred me to action.

I wrote a simple C routine to generate a minute's worth of digitised sine wave at 20kHz as a .RAW file.

Having done so, I examined it with Creative Technology's Wave Studio utility. Sure enough, the waveform looks totally chaotic, with samples apparently scattered all over the screen. But, when I pressed the 'play' button and displayed the *Soundblaster* output on an oscilloscope, I found a clean 20kHz sine wave. No trace of 4.1kHz either aurally or on the 'scope.

While my mathematical skills are nowhere near enough to rise to the challenge of understanding what is going on, I can only assume that the effect of kicking the output filter with the timed and scaled pulses of energy from the d-to-a converter causes the reconstruction of the 'input'.

Chris Miller
Sevenoaks

In response to Chris Bulman of Bedford, I believe I have a non-mathematical way in which to explain the problem he has with sampling theory. Consider the signal emerging from the digital-to-analogue converter: the high pitched whine going 'wuhwuhwuhwuh.'

As Chris may be aware, two sinewaves of different frequency when mixed together will cause a similar effect, i.e. they will 'beat.' The signal he describes can be thought of as a high-frequency sinewave (in this case, at half the sampling frequency) multiplied by a lower frequency sinewave (at a frequency equal to the difference between half the sampling frequency and the audio signal that was sampled initially).

Alternatively, it can be considered as the sum of two sinewaves: one at the audio frequency and one at the difference between the audio and sampling frequencies. This is a purely mathematical result.

The upshot of this is that if the signal is considered as two beating sinewaves, then it can be seen that as long as the audio signal is below the Nyquist frequency, then the high frequency component of the beat will have a frequency higher than the Nyquist limit. In the first example he gives it will be 20.001kHz, in the second it will be 24.1kHz.

This is why it is important to use a high-order filter with a cut-off at the Nyquist limit if these aliasing tones are not to be heard. And, of course, such high-order filters introduce phase shifts if not necessarily actual time delays.

D Chrastina
Scunthorpe
North Lincolnshire

Chris Bulman raises the question of sampling a signal at close to the Nyquist frequency, and how this introduces what looks like amplitude modulation.

The answer to Mr Bulman's question lies in his own letter, where he says "unless the filter had a time delay of over a second". The filter required to reconstruct a signal as close to Nyquist as in the example will indeed have a ringing response that lasts over a second.

The Nyquist theorem only says that if you sample fast enough the samples contain information for the original signal to be reconstructed. It does not say that the samples will like the original signal.

The proper reconstruction involves passing the sequence of samples, as

A drive on the high side

A thought always worth bearing in mind when designing gate drive circuits for power switching mosfets and igbts is that once these types of devices have been turned on, there is no need to sustain the gate drive. The device gate capacitance will maintain the applied gate voltage until it is time to turn the device off – using appropriate gate drive circuitry.

I refer to Ian Hegglun's article 'Driving on the high side,' from the August issue, in which he describes an elegant way of using low cost gates to drive the top transistor in bridge configurations. I am impressed with Ian's idea and offer one of my own on this.

A single pulse of short duration applied to the gate via a diode can be used to turn on a mosfet or igbt, and another pulse applied to a shorting device across the gate can provide turn-off. The gate can be left unattended in between.

While the pulses can be made quite short – less than 1μs in some applications – they must be long enough to see the device through the on and off switching transitions to prevent any adverse reaction due to the Miller effect. Once turned on in this way the device will safely handle any current in accordance with the specification sheet despite the fact that the gate is left floating.

The gate drive circuit is required to provide energy only during the switching transitions. Isolation between gate drive circuit and switching device can be provided by means of a small hand-wound ferrite-bead pulse-transformer – a transformer weighing less than five grammes sufficient to drive a 100A igbt, for example.

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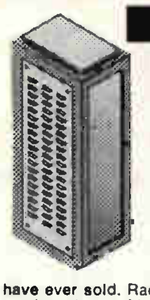

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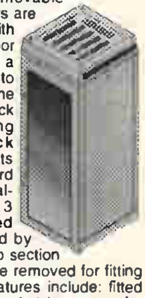

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impulses, through a low pass filter with cutoff at the Nyquist frequency. Each impulse hitting the filter causes ringing, and it is the sum of these successive rings that reconstructs the signal.

If the signal frequency is very close to the Nyquist, then the filter has to have a very sharp cutoff indeed – so it will have a very long impulse response, hence the 'memory' needed to carry over into regions where the signal 'looks' small.

Chris Bore
Woking
Surrey

Silver and the Bulman whistle

In the October issue Bob Pearson suggests that the reason why rf circuitry is silver-plated, although the reduction in resistance is modest at 7%, is that oxidised silver is conductive.

This is not so. Firstly, silver oxide does not form on silver because it is relatively unstable and cannot be made by heating silver in air.

Secondly, I am pretty sure it is totally non-conductive. The brown/black tarnish film that makes your teaspoons unsightly is actually silver sulphide, which is very definitely non-conductive. This derives from atmospheric hydrogen sulphide pollution – probably from diesel engines – and can have catastrophic results on small-signal switch contacts. This problem appears to be getting worse, and the only cure I know is to abandon silver, although it is otherwise ideal, and stump up for gold-flashed contacts.

As far as I can determine, copper in rf circuits is silver-plated not to reduce losses but to prevent corrosion. Skin effect means that any plating must have resistivity as low or lower than copper. Silver is the only metal that meets this; it is not immune to corrosion but it corrodes much more slowly than copper.

Turning to Chris Bulman, I can verify that bipolar transistors do sing when passing audio. The apparent high-frequency emphasis may be due to the small size of the radiating element – presumably the transistor die – but in any case this seems to rule out thermal effects. The absence of thermal distortion in discrete amplifiers also argues against a thermal origin.

Sound level appears to increase with device current rather than voltage or power, but the mechanism is apparently unknown; I have never seen a reference to the effect in any textbook or manufacturer's data. Does anyone have an answer?

Doug Self
London

Light service

May I suggest a simpler solution to H T Wynne's ingenious Porch Light Saver (*EW* Oct '97). That is to just use a low-power fluorescent light.

The simplest method is to plug in one of the modern compact electronic units. A 7W one should be sufficient but for a large porch an 11W might be required. My own porch is lit with a conventional, non-electronic, 4W straight fluorescent tube. I originally fitted a 6W one but found it brighter than necessary.

Although the average life of a conventional fluorescent light is only 5000 hours compared to 10000 for the electronic and its light output about 20% less, the conventional one is likely to work out a little cheaper in the long run due to the rather lower replacement cost.

P F Gascoyne
Wantage
Oxon

Calculating resistances

Some of my comments on the article 'Resistors in C' from the April issue were answered by the author in the August letters page, but I think the points he raises require a further response.

John emphasised that his printed program listing had merits in its size and simplicity, but set against those must be the lengthy processing time involved. I still cannot see the program getting much use if the calculation time can run into minutes – as stated in the original article.

For routine E24 calculations it would be better to have all 1728 permutations listed on a few sheets of paper, this is the reason why I encouraged readers to use a more efficient approach, and not rely on pure processing power to make a poor algorithm look good.

As to the algorithm contained in the 'more complex' version of the program – available via the special offer associated with the article – being different from that in the listing, all I can say is the article only indicated changes made to the data entry and presentation of results.

The order of results in the sample output is exactly the same as that produced by the simple version so I incorrectly assumed the search sequence was the same.

When querying the usefulness of "higher order" parallel combinations of resistors with values such as 0.01R and 10M, I was surprised they were included in the search range as I am not aware of any part of the industry using such combinations.

John might have guessed that I have come across the parallel resistor problem before. The eight-bit microprocessor I mentioned in the June edition was in fact the humble BBC Microcomputer of the early 1980s, on this I had a program to design resistive matching pads and within it a routine to convert the results to practical resistor pairs.

Just in case I am accused of being behind the times on this, I have tried transferring the original routine into a Windows program. For those interested in the results it is available from the "File Download" section on the web site:

http://ourworld.compuserve.com/homepages/david_markie

John concluded his letter with: "...it is easy for a person to belittle another's ideas, but it is something else to have the idea and do something about it. What was it they said about Marconi."

Well I am sure quite a lot was said about him, but what specifically did John have in mind?

David Markie

EMC rules

I applaud the courage and honesty of Ivor Catt. His display of courage begs me to come clean, even though I would prefer to keep the head down and pretend that I was thriving. I can personally vouch for the fatal impact the new regulations have had on my small company and on me, in that the game became not worth the candle.

Thank goodness I was convinced in time to contribute to a half decent pension fund and that I am old enough to draw it down. Helping friends and neighbours isn't all that bad. For example, I have become a world leader at repairing wall warts and I may prepare a paper for the IEEE/IEE on the technology involved.

Perhaps some kind reader, up to speed on the regulations, will let me know if it is within the law to use *Super Glue* to stick my wall warts back together after repair. There are no approval markings, that I can see, on the new bottle of glue I opened today. ■

Ted Crowley
Co. Wicklow Ireland

Fields, lines and tremors

Dominic de Mario is to be commended for investigating natural phenomena usually ignored as too basic to be worth investigating by mainstream science.

I have recorded the natural electric field since 1980, and can confirm that the wavetrain at 1.87Hz is probably a mains artefact caused by some sort of machine imposing pulses on the normal 50Hz mains waveform. The most likely source is a sensor controlled machine like a compressor or pump.

Mr de Mario's electric field sensor is elegant but unnecessarily sophisticated, and likely to detect rfi because it is chip based.

Natural atmospheric electric field activity is energetic and can show swings of hundreds of volts on a modest wire antenna. High input impedances and low noise devices are *not* needed with a reasonable antenna because the natural voltage gradient approaches 150 volts/meter above ground.

Since I wrote an *EW* feature on natural electric field and ionospheric interaction in October 1989, I have continuously run a 6Q7GT valve antenna electrometer in cathode follower mode as a geoelectric field detector rather than low voltage silicon devices .because the valve is lightning and rfi proof To keep the natural electric field within the 600V valve power supply, the antenna is loaded down by switched resistors For a horizontal wire antenna 20 meters long and 5 meters high, the antenna load of 300MΩ gives a voltage span of 100-200V on most days. The induced mains voltage of about 15 volts pk/pk riding on the atmospheric potential is turned to good account by using it as a 'chopper' to modulate the atmospheric potential, to give greater sensitivity – if required – than a signal taken off the valve cathode.

This high sensitivity signal is taken off the double diode anodes in the valve, and the 50Hz mains component filtered out using simple series *R* and *C* because the frequencies of interest are from dc to about 1Hz.

Using the sensor, it is possible to record lightning activity and cloud charges, as well as solar flares. The voltage on the antenna relates to the average height of the lower ionosphere, modulated by relative humidity- it also carries transients caused by meteors burning up overhead – but that's another story!

Mr De Mario's device will certainly work, but be unduly sensitive to local interference from mobile phones and vehicle radios as well as people moving near the antenna, whereas the loaded long wire antenna on a cathode follower valve will be rfi and lightning proof and can be accurately calibrated to directly read the atmospheric field voltage gradient.

Anthony Hopwood
Upton upon Severn
Worcestershire

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