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CIRCLE NO. 102 ON REPLY CARD

Contents

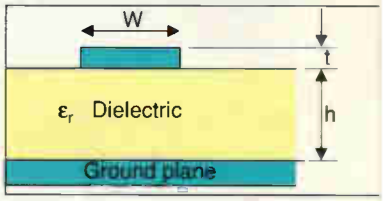
Cover – *Hashim Akib*

Regulars

Special offer

With three ranges, this sensitive frequency meter works to 2.5GHz yet costs under £100 – exclusively to EW readers. *See page 1010.*

This is the secret to coating disks and heads with diamond to stop hard disk crashes – *see page 978.*



There are complex procedures for designing high performance microstripline circuits, but a few shortcuts result in a technique that is quick and easy to implement yet more than adequate for most applications, *see page 1019.*

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At last – ISDN for all

A mass-market for ISDN was never in doubt; the only thing that blurred the vision was the cost of providing it. While large organisations found it easy to cost-justify the start-up charges, the cost of entry barred very effectively the wider army of hobbyists, home-based workers and small businesses from exploiting the undoubted benefits of ISDN.

Workarounds were promised to provide alternative routes onto the information superhighway 56kbit/s modems, XDSL techniques and cable modems but none of these have offered the total flexibility of real ISDN and users still waited for the cost of ISDN to fall.

And at last this fall is imminent, now that the key to providing affordable ISDN digital telephone service has been offered to UK network operators. British telecomms manufacturer GPT, which supplies the System X exchanges used by BT and many other providers, has come up with an easy-to-install package to enhance these switches. For the first time it offers a means of providing low-cost ISDN service for the mass market of small business and residential customers.

What is particularly elegant is that it achieves not one but two breakthroughs. First, the cost barrier which held back smaller users from exploiting the acknowledged benefits of digital data communication has been eliminated. In particular, home-based workers now have a cost-effective means of exchanging documents and sharing data with office-based colleagues, as well as gaining faster access to the Internet and a host of interactive information and entertainment services to come.

Small businesses requiring high-speed data transfer no longer need to install special private wires for the purpose.

The second advantage is the fact users can enjoy the benefits of ISDN without forfeiting their existing phones and numbers. Unlike previous implementations of ISDN, customers do not forfeit useful services such as call waiting, three-way calling, ring back when free, 1471 call return, Centrex and so on. Analogue caller display devices, which cannot operate on normal digital ISDN lines, still work under this new implementation.

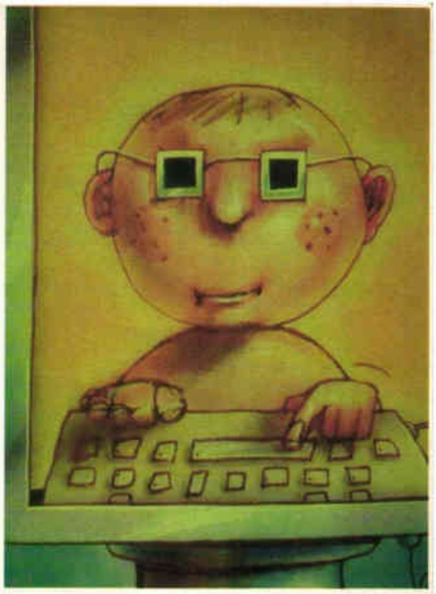
It adds up to a double bonus. For network operators it is a cost-effective delivery mechanism for providing twin-channel Basic Rate ISDN (ISDN2) service without the need for a second pair of wires. The customer benefits are equally significant; users gain the enhanced capability and 128kbit/s bandwidth of ISDN while retaining the convenience and established popular features of their existing analogue connection. Because the system can provide two simultaneous connections, users are spared the cost of having a second line installed.

The technical implications are relatively modest, involving an upgrade to System X exchanges and the provision of a compact plug-in unit at the customers end for connecting a customers existing telephone equipment with direct Plug-and-Play hook-up of a PC compatible. Installing this network terminator, which replaces the existing master phone socket, does not take long and setting up takes only a minute or two; the switchover from old-style analogue to ISDN service is effectively instantaneous and requires no access to customer premises.

The all-British technology is now being proposed to the European Telecommunications Standards Institute as a pan-European standard for narrowband multi-service delivery.

For network operators the new technique provides the opportunity to offer an extremely attractive ISDN package to a large and growing market. In the same way as high-speed modems soon became affordable and commonplace, it is reasonable to expect ISDN will make digital data communication the norm for small users as well as large.

For people working from home (the so-called teleworkers), low-cost ISDN will provide cost-effective means for exchanging documents and sharing data with office-based colleagues, as well as offering faster access to the Internet and a host of interactive information and entertainment services to come. Small businesses will be able to link up with one another electronically to create virtual enterprises, working together to provide a single concerted service to their clients. Faster connection and data transfer will give surfing the Internet a major new impetus, opening up new prospects of online shopping and multimedia information delivery from low-cost terminals alongside the domestic television set.



The entry cost of ISDN has barred very effectively the wider army of hobbyists, home-based workers and small businesses from exploiting its undoubted benefits.

Supporting this, the market for low-cost user products and solutions for ISDN is maturing rapidly, with a broad range of mini-PBXs, line cards and multi-function devices selling at truly affordable prices. One vendor offers a PC card combining ISDN file transfer at up to an effective 412kbit/s, V.34 modem emulation, Group 3 fax operation and a digital voice answering machine. It comes bundled with video conferencing software and an electronic secretary capable of storing and forwarding voice, data and fax messages. This sells for an extremely affordable 99 and with this kind of functionality on offer, users cannot fail to reconsider their attitudes to ISDN. ■

Andrew Emmerson

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Bright prospects for UK employment

Employers are forecasting more jobs within the UK's electronics industry, according to a recent survey conducted by Manpower.

Meanwhile, the September pay bulletin from the engineering employers' federation (EEF) has found that pay settlements within UK industry as a whole remain subdued and below the headline rate of inflation.

The average pay settlement for the quarter up to August was 3.17 per cent compared to 3.5 for inflation.

Manpower's latest Survey of Employment Prospects, conducted amongst 2,339 employers around the country, found 33 per cent of electronics employers expect to recruit during the fourth quarter of 1997, against seven per cent who expect to lay off staff, registering a balance of 26 per cent.

This quarter balance exceeds the national average of 23 per cent and the manufacturing sector average of 22 per cent.

Digital tv to miss World Cup

Digital Terrestrial Television (DTT) in the UK is set to be delayed by at least three months, following the deferment in the ITC's awarding of licences. Originally DTT was to be launched in July 1998, in time for the World Cup. Now October 1998 is the expected launch date.

"There has been a gentle slippage, but as is already known, the permission from ITC came out a bit later," said Jim Slater, project coordinator for DTT at the Digital TV Group.

Currently, the set-top box makers are waiting for the reference design from British Digital Broadcasting (BDB) for the manufacture of the first DTT receivers.

"We are currently drawing up the specification for manufacturers to build the boxes to," said a BDB spokesman.

New video camera has $8 price tag

Edinburgh-based c-mos camera specialist Vision has added a low cost sensor to its range. The 160 by 120 pixel digital cameras are aimed at the security, videophone, toys and games markets. Eight-bit resolution monochrome and colour cameras are available. The monochrome *VV5300* is priced below $8 in volume quantities. Vision also offers a $450 development system which includes PC software, capture card and lens.

Fairchild to decide on Temic by Christmas

Fairchild Semiconductor will make its decision whether or not to buy Temic by December. The company is also considering buying GEC-Plessey Semiconductors (GPS).

"We're still at the discovery stage – asking: 'Does Temic make sense?'" said Joe Martin, chief financial officer at Fairchild. "We'll make the decision on Temic in four to eight weeks."

"We've looked at 50 companies around the world since we spun-off from National on March 11th," added Martin. Citicorp Venture Capital, part of Citibank, financed the $550m Fairchild management buy out and gave it a brief to look for acquisitions.

"A lot of companies have a semiconductor subsidiary for strategic products. Some want to divest themselves of these companies and Fairchild provides them with an opportunity as a credible vehicle for consolidating these sort of operations. We are actively soliciting these companies as part of our acquisition strategy," said Kirk Pond, president and CEO of Fairchild.

Asked if Fairchild was interested in buying GPS. Martin replied: "We could be. It's only recently become available. We haven't had time to look at it yet, but it could be a possibility."

Fairchild is interested in the semiconductor side of Temic, not the automotive components side. The main attraction of the semiconductor side for Fairchild is the former Siliconix product line of power semiconductors, including small signal transistors and its *PowerMOS* line, explained Martin.

Flat and wide... NEC has put into production a 50-inch plasma display intended for the high definition television market. The Hi-Vision PlasmaX panel is capable of displaying one million pixels and will go on sale in Japan next February, in time says the company, to be used in high definition tv systems to be used during the Nagano Winter Olympic Games. But UK skiing enthusiasts – an exclusive bunch at the best of times – should be warned the selling price of the 50in flat format plasma televisions will be around ¥2.7m (£13000).

PCI bus is to get a refit

Power management and hot-plug capabilities are to be added to the PCI-bus, it was announced recently. Version 2.2 of the PCI specification is to be released early next year, said Donald Coffin of the PCI special interest group.

"Power management is the hardware side of Microsoft's ACPI or advanced configuration and power interface," explained Coffin.

ACPI allows parts of a PC to be powered down when not in use. Notebook PCs are already early adopters of the technology. By adding ACPI to the computer's local bus, peripherals will be more energy efficient.

Hot plug is aimed for use in servers, so that disk and Ethernet cards can be swapped without shutting down the system.

The last major update to the PCI specification, version 2.1, was released two years ago. This extended the bus from 32-bit at 33MHz to 64-bit at 66MHz. Bandwidth quadrupled to 528Mbyte/s.

"We are now seeing 64-bit, 66MHz systems, especially in servers," said Coffin.

The move to the higher rate will continue in servers and workstations over the next year to 18 months, claims Coffin.

Desktop PCs are likely to continue with the 32-bit, 33MHz version. This is because Intel's accelerated graphics port (AGP), which links the graphics system directly to the microprocessor, has greatly reduced the burden on the PCI-bus.

Awareness funding for year 2000 bug redirected

The Government is to stop funding Taskforce 2000, the body assigned to raise awareness about the Millennium bug. Instead, £1m will be given to Action 2000, a new organisation tasked with taking a more active approach to the problem.

Robin Guenier, head of Taskforce 2000, criticised the decision, saying it would waste valuable time. "We're going to lose weeks while people are appointed to [Action 2000]," he said. "Wouldn't it have been more sensible to continue with Taskforce 2000?"

A DTI spokesman countered: "Taskforce 2000's aim was to raise awareness and it did it very well. However, companies are not doing enough and a new impetus is required to turn awareness into action. We are showing our commitment, now it's up to industry to make its contribution."

Guenier insisted there is still a need to continue with his awareness programme, especially with regard to how Year 2000 affects embedded systems, to get the whole of industry and commerce tackling the problem rather than just helping the estimated 15 per cent that have already begun.

However, the DTI spokesman stressed the awareness campaign would continue as part of Action 2000's ambit.

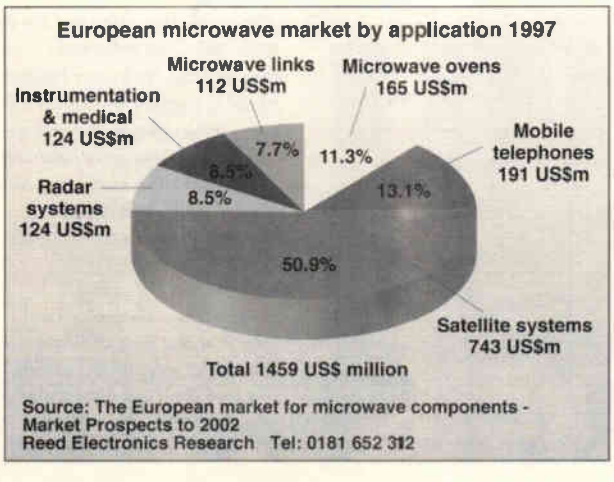
UK electronics industry warned

The UK electronics industry faces a great threat from Eastern European-based manufacturing, according to a paper from Loughborough University.

Professor David Williams, presenting the paper to the Cambridge Centre for International Manufacturing, warned that Eastern Europe threatens "unrooted" inward investments in the UK, and that even embedded foreign investment in Scotland is vulnerable. "Screwdriver inward investments will probably go to places like Poland and the Czech Republic," he said, pointing out that IBM had already transferred disk drive manufacturing capacity from Havant to the Republic.

Williams believes UK electronics companies need a greater awareness of their global role if they are to survive well into the next century. Success will mainly come for firms that couple investment in high value-added activities with sales into mature economies like the US. "Should we sell expensive products to rich Americans or cheap products to the Chinese?" he said.

But the UK must not pass up opportunities in the Far East for long term investment in areas of industrial growth. "We can't afford to miss another trick like we did with Singapore, where a low value region becomes a high value one," said Williams.



NEC details handheld Risc chip

NEC has released details of its latest microprocessor for the handheld computer (HPC) market.

The *VR4111* is a 64-bit Risc microprocessor based on the R4000 core from MIPS Technologies. It is an updated version of the VR4101 and 4102. The new devices are pin-compatible with the previous processors to allow easier design upgrades.

Performance reaches a claimed 130Mips at 100MHz by moving the device to a 0.25μm manufacturing process. Power consumption has been reduced from 250mW to 180mW by running the device at 2.5V.

To reduce memory requirements, the devices can execute a 16-bit instruction set, similar to the ARM *Thumb*.

NEC has integrated HPC-specific peripherals for LCDs, keyboards, infra-red and serial I/O, audio and touch screen interface.

Future bleak for NCs

Network computers (NCs) could be dead in the water as network managers decide to stick with existing network equipment.

This message emerged from a Fodor Wyllie survey commissioned from Pin Point Software, the network management tools supplier. According to the survey, 56 per cent of companies questioned did not believe the NC would make a significant impact on their organisations even though network management would be made easier.

"The momentum for network computers has been lost. We were amazed at the ambivalence towards the network computer. Clearly there is a belief that the NC can help but there's a lot of pessimism about its implementation," said Iain Franklin from Pin Point Software. "Although the NC saves a lot of time managing the network, companies prefer to stick to what they have rather than switch their whole networks to NCs."

According to Pin Point Software this is a surprising revelation as the market for network management devices is growing at a rate of 30 per cent per year and it has already reached the $1bn mark.

Another analyst, Bloor Research disagrees. "I don't believe the momentum for NCs is lost," said Rob Hailstone, chief analyst at Bloor Research. " I suspect until we see the first successful implementation of NCs people will think it's not going to happen."

Guidelines aimed at cutting keyfob interference

The chance of getting locked out of your car by EMC problems should be reduced if guidelines by RAKE, the Radio Activated Key Committee, are adopted.

Problems arise because EU legislation added automotive keyfobs to the list of legitimate users of 433.93MHz in the UK, which already included the MoD and radio amateurs. "We are in no position to ask the other user to get off the frequency – the radio amateurs have been using it for over 40 years," said Peter Brill, RAKE spokesman at the RAC.

This said, the guidelines contain the paragraph: 'As a first step, in the UK, the next two years will see the removal of some users of the 433MHz band to other parts of the spectrum'.

The guidelines are split into two parts: a guide to motorists and a guide to manufacturers for future products.

"One of the simple things we are encouraging manufacturers to do is ensure there is an alternative method of entry," said Brill.

Other guidelines suggest marking 'hot spots' on and in the vehicle where it is most sensitive to the keyfob being held, visible instructions to remind drivers what to do if the fob fails to operate the car and a reduction in receiver bandwidth.

Motorist Web update: http://www.rac.co.uk/html/fut/pr971008.htm

Initial products for FireWire announced

The first FireWire products are to start appearing on the market in the next six months, two years after the IEEE1394 -1995 standard was agreed.

"We had the standard two years ago. It's time to start generating products this and next year," said Hiro Tsutsui, marketing manager at Sony Image Sensing Products in Japan.

Intel will deliver a prototype motherboards with the FireWire-1995 digital interface early next year. The software by Microsoft will also cater for this type of linking then and Apple will introduce Power Macs with this interface.

Last week, Sony announced its DFW-V300 digital colour camera, one of the first FireWire peripherals, with this high-speed serial bus standard on board. In 1998, Sony hopes to launch another six products from its image sensing portfolio.

"The real benefit is that IEEE-1394 standard allows digital data to be imported directly into a PC or device, meaning the DFW-V300 camera provides a very quick and very scalable solution for transferring video data for processing by computer," said Steve Hearn, business manager at Sony Image Sensing Products.

The next FireWire specification, IEEE1394-1998 or IEEE1394-2000, is being discussed between the industry players and the standardisation bodies including the 1394.b Working Group that is working towards the 3.2Gbit/s specification.

In brief

New specs for DVD

The latest specification for the write-once DVD-Recordable (DVD-R) format has been published by the DVD Forum. The version 1.0 specification offers a 3.95Gbyte single-sided or 7.9Gbyte double-sided storage capacity that will be compatible with other DVD formats such as DVD-ROM and DVD-Video. The next generation 4.7Gbyte DVD-R is currently under discussion by the DVD-R Working Group. Meanwhile, Matsushita Electric Industrial expects first-year total worldwide sales of DVD players to be half of that first envisaged. Initial forecasts were for sales of up to two million units, however, the forecasts have been downgraded to below one million units.

Risc processor for hand-helds

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CO_2 will affect chip production

Global warming will affect the semiconductor industry, says a report from the Economic Strategy Institute of Washington DC.

The Institute warns that if proposals to cut CO_2 emissions proposed by the Kyoto global warming summit are adopted, then the global economy will slow down leading to an eight per cent reduction in the demand for semiconductors. ■

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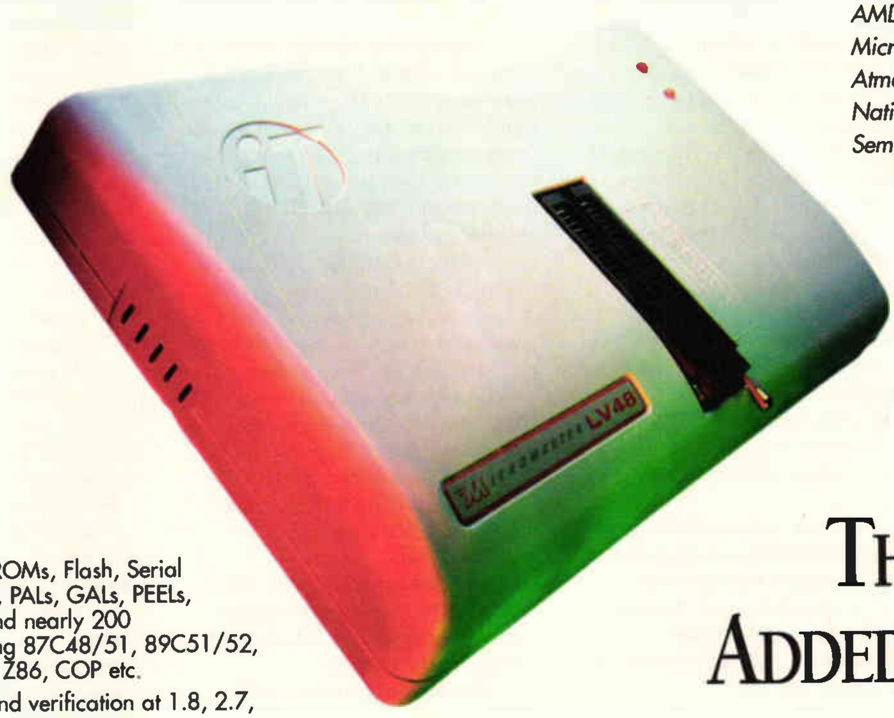

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RESEARCH NOTES

Jonathan Campbell

Diamond coating adds twinkle to disk storage

One of the major stumbling blocks in the design of ultra-high storage disk drives is that the read/write heads must be located ever nearer to the disc surface to be able to resolve the data. At these tiny distance, impacts are inevitable so the heads must be coated. But any protective coating must be super thin too, so as not to disturb magnetic performance.

Now, a way has been found to shield disks and sliders – reader heads – with ultra-thin 'overcoats' of diamond-like carbon that can survive repeated crash landings at 3600rev/min. Success is being welcomed as a breakthrough in progress to high density storage devices.

Towards more reliable hard disk drives – ionised plasma follows the curves of the magnetic field, but unwanted macroparticles just hit the wall.

The work has been carried out by Simone Anders of the Accelerator and Fusion Research Division at the Ernest Orlando Lawrence Berkeley National Laboratory and her colleagues from IBM and University of California, Berkeley.

IBM has already brought to market disks that store 4Mbytes/mm^2 of data, while densities almost twice that have been demonstrated, and researchers are aiming for 16Mbytes/mm^2 and more. To read a disk where magnetic domains are packed only 25nm apart, disk surface and slider will have to move so close to each other that it's almost a matter of semantics whether they will actually be touching.

Typical high-quality commercial overcoats now in use are made of sputtered-on, hydrogenated carbon 12 to 15nm thick. But higher data densities require reduced magnetic spacing between heads and disks – so disk coatings must be thinner and made of even harder material. Sputtering can't do the job, but a technique called cathodic arc deposition can.

Unlike sputtering, cathodic arcs produce a fully ionised plasma of whatever material, including carbon, is used for the cathode.

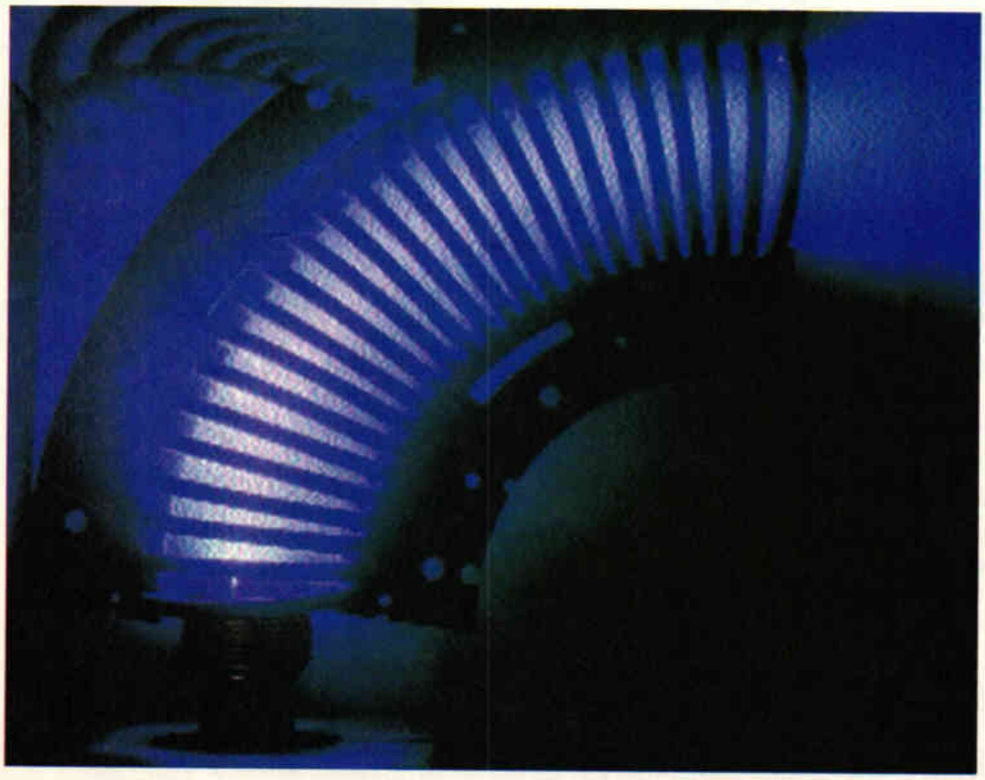
A fully ionised carbon plasma allows electrons and carbon nuclei to reassemble themselves as diamond, in a three-dimensional lattice in which each atom is bound to four others by electron pairs – a tetrahedral bond. In contrast, atoms in graphite are bound to only three other atoms, forming a much less stable configuration. By tuning the energy of the incoming carbon ions, the tetrahedral-bond content of the deposited film can be optimised, so films have been made that, while technically amorphous, are 85% diamond.

"Still, the method hasn't been practical for coating disks," says Anders, "because micron-sized chunks of the cathode boil off and contaminate the films."

For cathodic arc deposition to be useful in coating disks and sliders, a way must be found to completely filter out the macroparticles. What Anders and her team has achieved is to devise a filter so good that all the goals of thin, flat, hard, macroparticle-free carbon are fulfiled.

The secret is a magnetic coil that looks much like a Slinky toy, placed between the plasma source and the substrate to be coated. The fully ionised plasma is easily bent through this S-shaped magnetic field – effectively two fields at right angles – but the massive macroparticles of carbon can't turn easily; they fly right through the sides of the coil or pile up on its walls. A coil that has been used for some time is thickly coated with a dust of macroparticles near the plasma source, yet dust-free at the substrate end.

Tests show that disks coated with cathodic arc carbon have a coefficient of friction half that of those coated with hydrogenated carbon and cause 20 times less wear on the slider. In additional studies, when a silicon wafer coated with cathodic arc carbon has been examined at nanometre scale after repeated loading, it shows virtually no scratches.



Contact, Simone Anders, Accelerator and Fusion Research Division, Ernest Orlando Lawrence Berkeley National Laboratory, Berkeley, California, USA

Radar finds people behind locked doors

The radar torch can detect the presence of a human through a wall or door (Stanley Leary, Georgia Tech)

Lives could be saved by a prototype radar system, developed at Georgia Tech Research Institute, that can sense someone moving or breathing up to 3m behind a locked door. For police and hostage negotiators, the device, little bigger than a torch could provide vital information about where people are located inside a room. The sensor could also be used to find survivors in the rubble of accidents or earthquakes.

A narrow radar beam of about 15 to 20° and a specialised signal processor is used to detect the body movement generated by breathing.

Research that evolved into the radar torch began in the mid-1980s, with the patenting of a frequency modulated radar for remotely checking vital signs of battlefield wounded before sending in medical teams. This early technology was also tested for its ability to monitor vital signs of soldiers clothed in chemical or biological warfare suits, without requiring them to risk contamination by removing the protective gear. Now the technology has been refined for more general use.

Radar has some advantages over other technologies. The signal from the radar will penetrate clothes and detect respiration through a heavy jacket, and requires a body movement of only a few millimetres to detect human presence.

"Based on respiration signature alone, the radar flashlight allows us to detect a stationary individual behind a solid wooden door, or standing 1.2m behind a 200mm block wall," explains Gene Greneker, a principal research scientist at the GTRI.

Even so, the amount of electromagnetic radiation exposure from the device is still small – said to ten times less than the voluntary exposure leakage level for microwave ovens in the United States. Exposure is estimated to be about the same as a person receives when walking under an automatic door opener triggered by a microwave sensor.

For now, the signal processor is external to the radar sensor, and the respiration signature is displayed on a monitor driven by a computer-based radar signal processor. But Greneker plans to make everything small enough to fit inside the torch body by incorporating high-speed signal processing technology.

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Chill settles on car electronics

Daimler-Benz has constructed a rig specially designed for low temperature testing of microwave components.

Electronics are now almost as an important component of cars as the engines and bodies themselves. Over at the Daimler-Benz Research Centre in Ulm, scientists are putting the next generation of microwave circuitry through its paces, using a specially-constructed low-temperature measuring station allows researchers to investigate the characteristics of new microwave circuits at temperatures far below freezing point. In this way, any adverse effects on the performance of superconductor elements can be predicted.

At the measuring station, tests can be carried out at temperatures as low as –250°C, allowing the scientists to examine new components for high-power radar sensor units or for compact narrow-band receivers in communications technology, while detailing their behaviour at extreme sub-zero temperatures. On the basis of these results, the characteristics and capabilities of superconductor elements can be precisely predicted.

The use of superconductor materials in place of metals and the resulting new approaches to modelling in hybrid combination are seen as milestones along the road to new, highly sensitive receiver systems for the microwave band. The objective is to benefit from the advantages of various available components and unite them in a single overall system.

Automotive researchers currently regard this as the most cost-effective solution for the production of such components in small and medium volumes.

Organic computer comes closer...

Researchers at Yale University in the US have succeeded for the first time in measuring an electric current flowing through a single organic molecule sandwiched between metal electrodes. The feat could pave the way for a new generation of transistors so small that a beaker full would contain more transistors than exist in the world today, according to Yale electrical engineer Mark Reed, team leader.

The accomplishment is a fundamental step toward creating computers and sensors that are smaller, faster and cheaper than today's silicon-based computers. The next step is to design computer chips whose wires are made of self-assembling strings of organic molecules that grow in a beaker, since the wires would be far too small to produce any other way. The organic wires would adhere to metal electrodes, a revolutionary strategy for fabricating electronic devices for which Professor Reed and Yale hold a joint patent.

But Reed warns that it could take a decade to learn how to make useful devices out of quantum components made from organic compounds.

To capture the historic measurement of current across a single organic molecule, the researchers made a mechanically controllable break junction by gluing a notched gold wire to a flexible substrate, then fracturing the wire to make an adjustable gap. Next, they sandwiched a single molecule of benzene (a hexagonal ring made up of six carbon and six hydrogen atoms) flanked by two sticky sulphur atoms between the two gold electrodes. The process required self-assembly of benzene molecules onto the electrodes.

For the future, perhaps the greatest obstacle to be overcome in fabricating useful quantum devices is to find better, faster ways to make large quantities. Electron beam lithography is used at the moment and this is just not practical for volume production. The answer could be to find materials that will assemble themselves into quantum components.

Reed says his goal is to find organic chemicals that will combine to form a substrate of conducting molecules.

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...But can silicon wait?

Progress to ever smaller, faster and cheaper computers may soon come to a grinding halt because microscopic silicon chips are getting so small that eventually they will contain too few atoms to work, warn two University of Florida researchers.

By the year 2010 the limit will be reached, and microprocessors will be as small and as fast as they can get, says Kevin Jones, professor of materials science and engineering and co-director of UF's SoftWare and Analysis of Advanced Material Processing (the Swamp Center).

Jones and Swamp co-director Mark Law are concerned that the heart of the Pentium processor transistor, a layer that once was thousands of atoms thick, is getting so small that it soon will be only 50 atoms thick. They say the Pentium processor may eventually shrink itself out of function when it gets to be fewer than 10 atoms thick, in just over a decade. That means unless there is a revolutionary change in computer technology, the trend toward smaller, faster computers will have reached its limit.

Effectiveness of shrinking processors is also threatened by the impurities in microscopic silicon chips. Law and Jones are currently using 3d computer simulation to investigate the nature of the impurities in silicon which can cause current to flow where it is not supposed to.

Impurities can cause catastrophic failure when the impurities run into the crystal and cause electronic switches to short circuit.

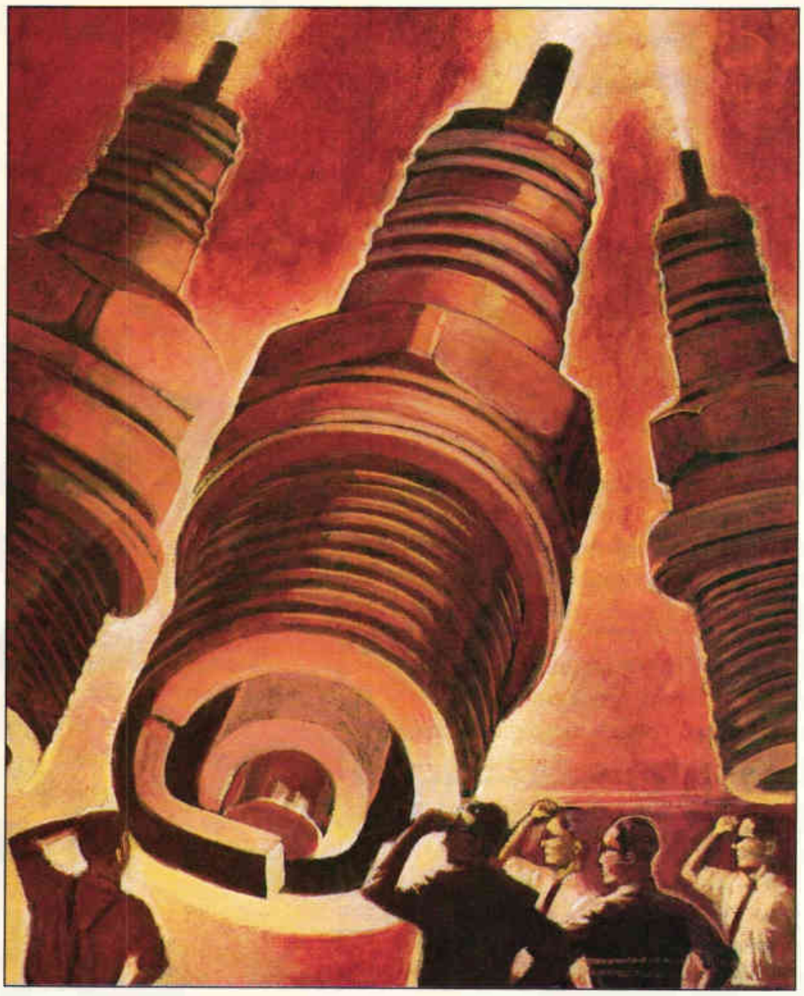
The researchers are currently making predictive models for the industry, primarily using computer simulation because testing with real materials is so expensive.

"We need to know how impurities diffuse, at what rate and at what temperature," says Law.

"We can then code that data and build predictive models that allow the industry to build better chips." ■

Contact: Kevin Jones, University of Florida, Gainesville, USA. Tel: 00 1 352 392 9872.

In petrol-engine efficiency terms, an increase of 5 to 10% is exceptional. Automotive engineer Michael Ward says that his ignition system can achieve this – and it can do so with the side benefits of cheaper emission-control systems together with simpler engine mechanics.



Bright sparks

The conventional inductive ignition system, based on the flyback circuit, has survived without significant challenge for almost a century. Invented and developed by Charles Kettering, circa 1910, it powers essentially all passenger cars and trucks today. It features simplicity, low cost, and reliability.

However, it masks several shortcomings – including low spark energy relative to what is required for best engine efficiency. These shortcomings are taking on a particular importance with the industry's call for one-coil-per-plug ignition and with the international call for improved fuel efficiency to counter ever growing carbon emissions worldwide – which equates to global warming.

Related concerns of high fuel costs in Europe and Japan, and failing Corporate Average Fuel Efficiency standards, CAFE, in the US further increase the concerns. The increased popularity of minivans, sport-utility vehicles, and light trucks in the US is leading to a reduction in vehicle gas mileage – the lowest since 1985 – and to growth in US oil imports to over 50%.

In the latter seventies, Robert Bosch and General Motors showed that for best engine efficiency, an ignition system needed to deliver 150 to 250 millijoules of spark energy. Yet today's inductive ignition systems deliver only a small fraction of that – typically 30 to 50mJ as measured by an 800V zener load, which is the industry standard.

Without a substantial increase in coil weight, from the state-of-the-art 200g to an unacceptable 500g, the required minimum spark energy of 150mJ cannot be achieved. Moreover, today's inductive ignition systems deliver a spark with a current of 50 milliamps. This equates to a spark power of just 20 watts. But to achieve the optimum combustion power in the initial propagating flame demands a spark of around a 100W. More importantly, such spark discharges are susceptible to break-up by the high mixture flows found in engine cylinders of modern engines.

Why a bigger spark?

From the 20 years of research conducted at CEI on ignition and early flame propagation, we have found that spark power more closely equal to that produced by the initial flame is better suited for ignition of dilute mixtures. This equates to lean or high exhaust-gas recirculation, or EGR, mixtures which increase engine efficiency and lower emissions.

Such higher-power spark discharges require spark currents above 200mA, which is the demarcation

Dr Michael Ward is President of Combustion Electromagnetics Inc., Arlington MA.

between a low current 'glow discharge' with its high electrode drop and low efficiency, and the 'arc discharge' with its low electrode drop and higher efficiency. The arc discharge has a much greater resistance to flow segmentation, allowing the spark to be located in a more central portion of the combustion chamber where it can interact with the flow to provide the greatest lean burn capability – as we have demonstrated to the Japanese and General Motors.

Existing ignition shortfalls

Meeting the combined goals of high energy, light coil weight, high spark current, and high spark flow resistance needs a different approach to inductive ignition. To this end, we have developed of a new type of hybrid inductive ignition, which we call HBI. It is based on improvements in insulated-gate bipolar transistor, or igbt, technology and innovations in circuit and component design.

The basic inductive ignition system is made up of the car battery of voltage V_b, the ignition coil T, the power switch S, the clamp D, and the load which is a spark gap G on the coil secondary winding, **Fig. 1**.

In operation, the system is fired to sequentially store magnetic energy in the core of the coil through current flow in the primary winding of inductance L_p to a maximum peak 'break' current I_p. This current produces a positive sense voltage in the sense resistor r which is measured by a sense and control circuit.

When switch S opens, the voltages on the primary and secondary windings rise. Initially, the primary voltage rises due to the coil leakage inductance L_{pe} energy charging the primary circuit capacitance and getting clipped and dissipated by a clamp D. Primary voltage then rises in phase with the secondary voltage V_s to charge the secondary capacitance C_s. It continues up to the spark gap breakdown voltage V_{br}, to deliver the remaining energy stored in the core to the spark gap.

Breakdown voltage V_{br} depends mainly on spark-gap size and engine load – as measured by the cylinder volumetric efficiency. It is typically 5 to 30 kilovolts.

Peak output voltage, V_{spk}, is controlled by the clamp D, which for a coil turns ratio N of 90, and assumed clamp voltage V_{cl} of 380V, limits the peak voltage to approximately 36kV as a result of transformer action. This value is greater than $N \times V_{cl}$ due to imperfect winding coupling.

The energy that can be stored in the core, E_p, must satisfy limitations on the peak core magnetic flux density B_p given below, where N_p is the primary winding turns and A_p is the core

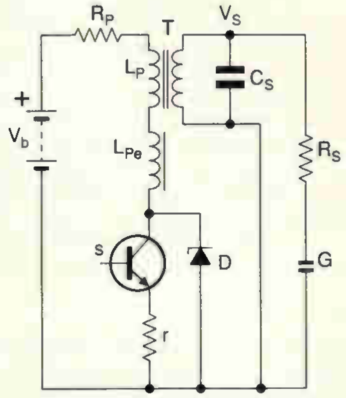


Fig. 1. In the basic ignition system, arcing at the load, i.e. the spark plug, is caused when the switch opens to transfer energy stored in the coil.

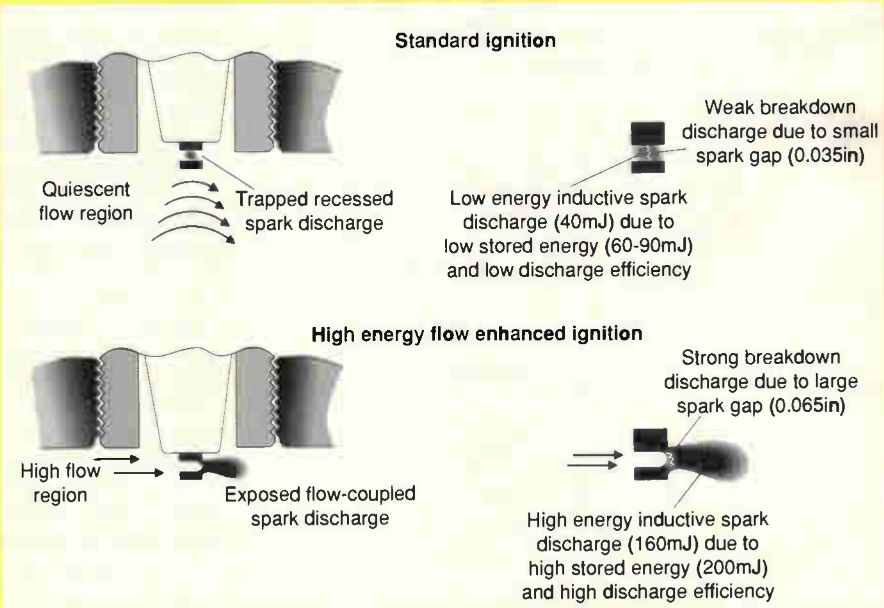
Advances in spark plug design

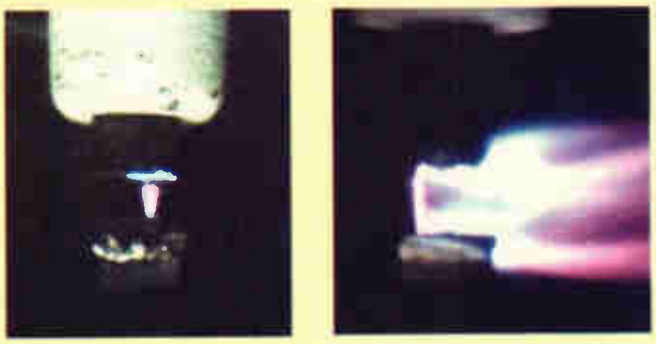
Significant work has been taking place recently in improving the design of spark plug tips to minimise their quenching effects on the ignition spark and on the early flame propagation, as well as to increase spark plug life.

Such plug designs are of particular interest in this application since the HBI system produces a spark with a higher spark current which is much less susceptible to being quenched or segmented, i.e. broken up by high in-cylinder flow velocities.

HBI would preferably use such spark plugs with a larger spark gap enabled by the higher output voltage of the ignition. The plug tip would be placed well within the combustion chamber within the flow stream to provide even greater capability for dilute combustion, i.e. lean or high EGR burn, to further improve engine efficiency.

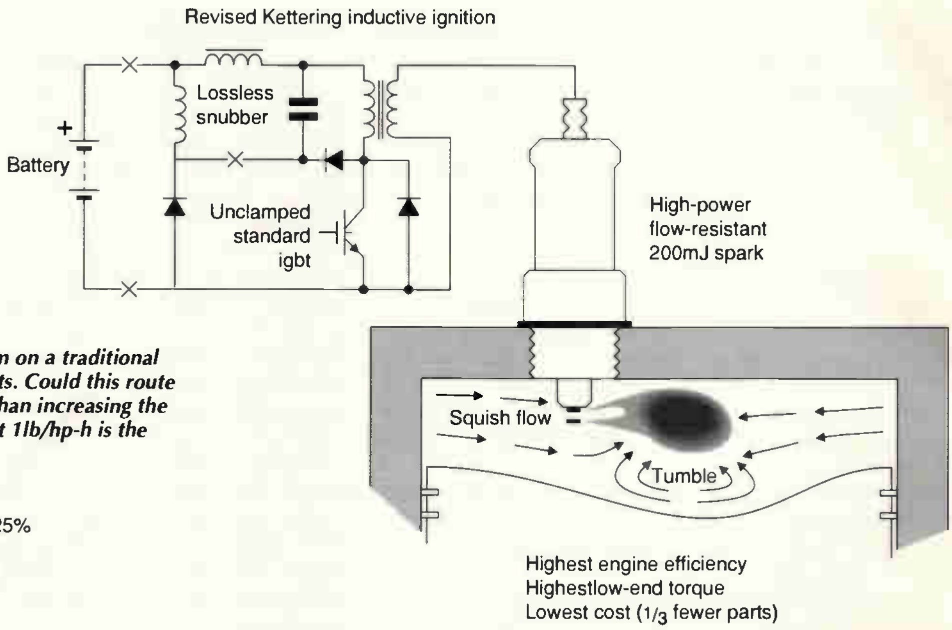
CEI has developed its own patented version of a spark plug, called halo-disc, which was successfully tested at one of the Big Three and gave an actual fuel economy improvement in and of itself. The spark plug features a thin, suspended, ground ring electrode to minimise electrode obstruction and quenching. The electrode is best made of erosion resistant material such as tungsten-nickel-iron, and has a recessed insulator to minimise fouling of the insulator end.



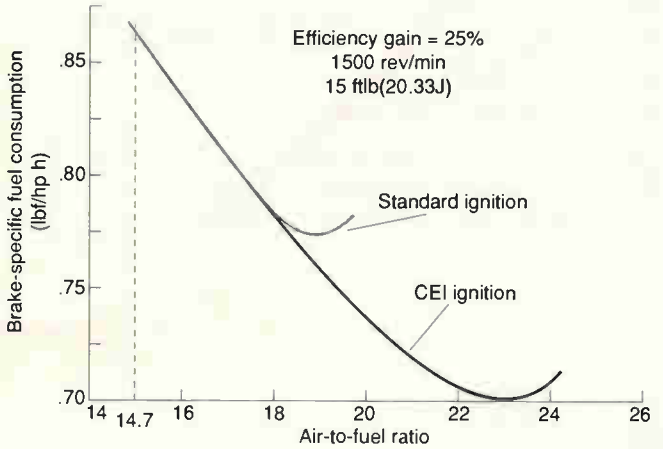


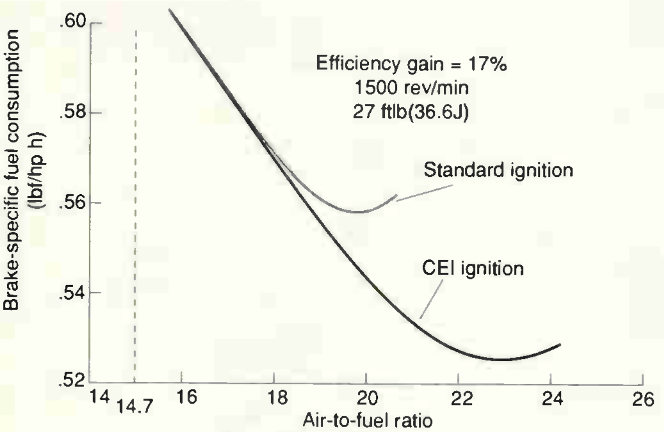
Being deeper in the cylinder and three times more powerful, the spark on the right acts with fuel/air turbulence to become a far more effective igniter than the traditional spark on the left.

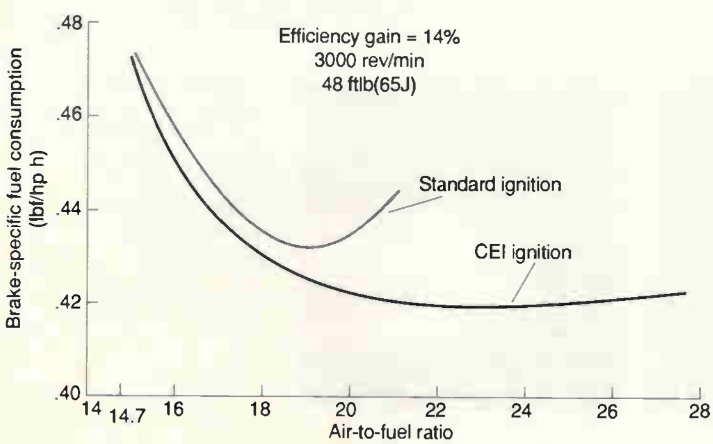
Having a large spark well down in the cylinder allows a leaner burn. When a car slows down with its clutch engaged, fuel is sucked into the engine and wasted. Driving around town, this fuel waste is significant. With existing ignition systems, shutting off the fuel during deceleration is problematical because of lean surges during the deceleration-acceleration transition. But the leaner burn of the new system should allow a more complete fuel shut off during deceleration as one of its several benefits.



Tests carried out using the new ignition system on a traditional twin-valve push-rod engine show clear benefits. Could this route to a cleaner, more efficient engine be better than increasing the number of valves and injecting fuel? Note that 1lb/hp-h is the equivalent of 608g/kW-h.







Might we be better off without 16 valves and fuel injection?

The most effective application of the HBI system is with engines that possess some degree of controlled in-cylinder air flow, such as swirl, tumble-flow, or squish flow. Squish flow is the most desirable as it provides high flows at low engine speeds, when combustion is normally poorest.

Unlike the four-valve-per-cylinder alternative, the twin-valve push-rod engine has natural squish flow, making it an ideal candidate. It has powered more cars and trucks than any other engine in the world. And as has been pointed out in the Wall Street Journal, the twin-valve push-rod engine has a third the parts of the four-valve overhead cam engine and it does not suffer the same friction losses in the valve train.

Traditionally, the spark gap has been located in a recessed, quiescent part of the cylinder for fear of being blown out, or broken-up. With HBI, the ignition spark is positioned in a more central part of the combustion chamber. Here, it is in the squish zone and can be directly coupled to the flow fields to spread combustion efficiently, without the fear of break-up.

In turn, the higher spark voltage associated with the flow coupling will maximise the ignition circuit efficiency. In this way, the maximum of the available, much higher, stored energy is delivered to the flow-coupled spark discharge, as in the drawing.

In tests using an early version of high energy flow resistant ignition on a Ford-Europe engine sponsored by Dr. Tony Jarrett – Group Director of Product Technology for Lucas Industries, Birmingham, England – the air-fuel ratio of the twin-valve push-rod engine tested was improved by three to four air-fuel ratios at low speeds of 1500rev/min and low flows. Gains in fuel economy were substantial, as the graphs illustrate.

At higher speeds, around 3000rev/min, and loads where the flows where much higher, but not sufficient to break up the higher flow-resistant spark, the air-fuel ratio was improved by a remarkable eight air-fuel ratios to provide the lowest possible N0x emissions with significant gains in fuel economy.

The new HBI system, and its equivalent capacitive discharge version which has comparably high spark energy and even greater resistance to flow-segmentation, can substantially improve the conventional twin-valve push-rod engine. Only a slight modification is needed to place the spark plug at the inside edge of the squish zone.

Such an approach would revitalise the twin-valve push-rod engine and make it competitive with the most advanced high-efficiency engine technologies – including the diesel engine and the recently developed gasoline direct injection (GDI) stratified charge engine. Added bonuses would be a much simpler design and lower cost than either alternative.

This would also be the case for two-stroke engines, which suffer from poor combustion at low speeds.

area. In addition, the coil windings must be designed to have acceptably low primary and secondary resistances R_p, R_s to limit the resistive heat dissipation during coil operation. This prevents the coil from overheating at high engine speeds.

For the same stored energy, current, and winding length, the winding resistances R_p and R_s increase inversely with the fourth power of the coil diameter, increasing by a factor of 16 times when the coil diameter is halved.

More revealing, the primary circuit dissipation during coil energising, E_{dp}, for fixed coil size and peak flux density B_p, is proportional to the cube of the stored energy E_p and inversely proportional to the power delivered to the primary winding. This means that it is inversely proportional to the product of the primary-winding voltage V_p and the current I_p through the winding.

Coil energising time, or 'dwell' time, T_p, is also given below assuming T_p is less than the primary circuit time constant L_p/R_p, recognising that the voltage V_p across the primary winding is less than the supply voltage by the switch voltage drop V_{ce}.

$$E_p = \frac{1}{2} L_p I_p^2$$

$$B_p = \frac{L_p I_p}{N_p A_p}$$

$$E_{dp} = \frac{C \times E_p^3}{V_p I_p}$$

$$T_p = \frac{L_p I_p}{V_p}$$

Higher output – more dissipation

Doubling the stored energy E_p increases the dissipation E_{dp} by a factor of eight. This makes it impracticable to maintain the higher spark energy for the same size coil, supply voltage and peak current, i.e. by increasing the inductance L_p, as is preferred with conventional ignition systems.

Only by increasing the primary power in proportion can the same dissipation be maintained. This is a criterion we choose to use for a properly designed ignition coil – assuming the base coil has been properly designed for heat dissipation.

For the case of a state-of-the-art inductive ignition system of Fig. 1 and ideal coil of weight 200g, the theoretical stored energy E_p is 60mJ, as given below. Assuming a high coil charging efficiency of 60% and spark energy transfer efficiency of 75%, overall efficiency is high, at 45%. Resulting spark energy E_s is 45mJ.

Typical parameters for such an inductive ignition system are given below, where a peak magnetic flux density B_p of 1.4 tesla is assumed as the practical maximum value for standard ignition. This assumes a magnetic core material of silicon iron.

The charge or dwell time T_p is calculated at full battery voltage of 13.7V, or V_p of 12.4V assuming a switch drop V_{ce} of 1.3V.

L_p=2.4mH
I_p=7A
A_p=1.2cm^2
N_p=100
N=90
B_p=1.4T
E_p=60mJ
E_s= 45mJ
T_p=1.4ms

Doubling the stored energy E_p to 120mJ and maintaining the same dissipation E_{dp} requires that the break current Ip be raised from 7A to 56A for the standard inductive ignition powered by the car battery. Even assuming state-of-the-art logic-level switching 400V igbts with low saturation voltages, this remains impracticable in view of the high switching losses.

Switching losses are proportional to the peak current I_p and increase further with temperature. The higher switching losses would be a source of further internal heating of the igbt which would not be tolerable in the high temperature environment of the engine compartment.

Losses due to switching can reach 60mJ – or half the stored energy – offering no improvement in energy delivery. Moreover, even if you could attain the spark energy of 90mJ, you still have a 60mJ energy shortfall relative to the minimum 150mJ requirement.

Peak spark current I_s, of 620mA, would also be higher than desired. Without some significant change in strategy for inductive ignition, it is not possible to meet the competing goals of one-coil-per-plug ignition with very small coils, and the need for high spark energy for improved fuel economy, within the additional constraints of acceptable costs and other preferred features.

A new approach to ignition

The equation for dissipation E_{dp} suggests the direction needed to be taken. It is necessary to make the system independent of the battery voltage – assumed to vary between 6 and 14V.

It is also necessary to raise the voltage V_p, or generalised supply voltage V_c to a higher value of approximately three times maximum battery voltage. A value of 42V is represents a good compromise between an even higher voltage V_c and a value required to limit the output voltage upon switch closure to prevent false firing of the spark plug.

By using the moderately high operating voltage of 42V, then a higher primary current I_p of approximately 30A can be used. A proviso is that a snubber circuit needs to be included to reduce the otherwise high switching losses at the higher current Ip.

Proper design of the snubber allows for a best trade-off between conduction losses and switching losses, to the point where conduction losses dominate to allow for the use of the lowest cost, standard speed igbt, versus more expensive fast or ultra-fast devices.

Table 1 shows measured values of switching and conduction losses of 600V standard speed and fast igbts under operating conditions, i.e. with a peak current I_p of 30A, a stored energy E_p of 200mJ, and a peak snubber capacitor voltage of 440V.

As is clear from the table, conduction losses dominate, suggesting use of the standard speed igbt. Among such devices, there are two choices. One is the present logic level, 400V, 15 or 20A voltage-clamped, n-channel ignition igbt. The second is the 600V, 20A, unclamped n-channel general purpose igbt.

In the case of the 400V ignition igbt, the coil turns ratio N and snubber capacitor would be adjusted to accommodate the peak voltage of 400V.

The 400V alternative has three main advantages. Firstly it has a low drive requirement of only 5V for a peak current I_p of 30A. At this level, collector-to-emitter voltage drop V_{ce} is only 2.5V. Secondly, the 400V igbt has a low V_{ce} of only 2.0V at 30A at the battery rated drive of 12V. And finally, its internal 400V clamp allows greater flexibility in use of the device including placing it at remote locations such as on the ignition coil.

The 600V general-purpose igbt also has several advantages. Its collector-to-emitter voltage V_{ces} is high, at 600V, and it has a low V_{ce} of 2.5V at 30A at the battery rated drive of 12V for

Table 1. Measured switching and conduction losses of standard and fast igbts for a peak current of 30A, stored energy of 200mJ and 440V peak snubber capacitor rating.

Device	Conduction loss	Switching loss	Total losses
Fast igbt	26.9mJ	10.2mJ	37.1mJ
Standard igbt	22.4mJ	12.4mJ	34.8mJ

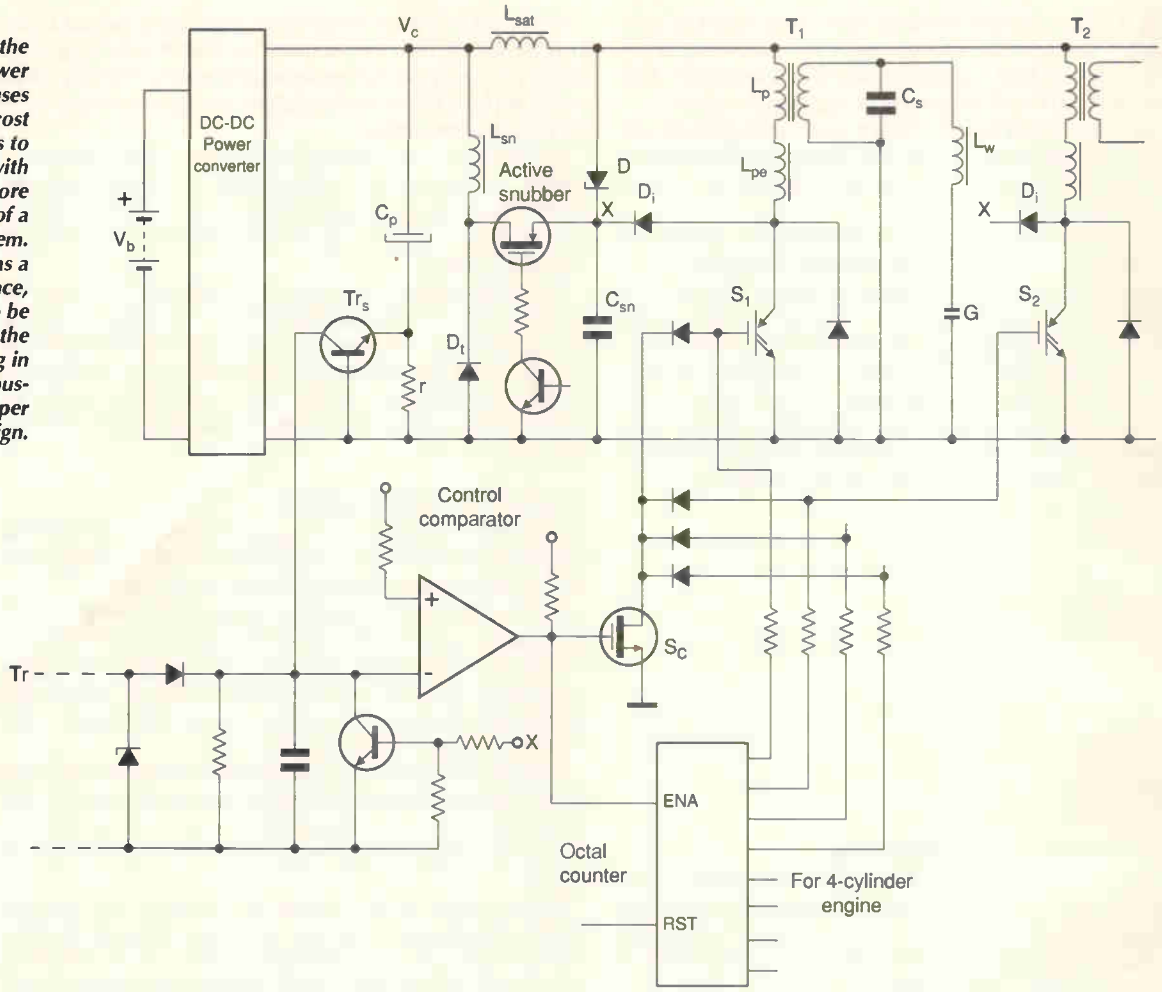


Fig. 2. Circuit of the new high-power ignition system uses standard low-cost components to produce a spark with over three times more energy than that of a conventional system. The spark also has a higher flow resistance, allowing it to be initiated deeper in the cylinder, resulting in more efficient combustion and a cheaper engine design.

fourth-generation devices. This igbt also exhibits fast switching, of 1 to 2μs, at high junction temperatures of 150°C. It has the lowest number of masks and few critical alignments in the manufacturing process, resulting in high yields and low cost.

No clamping necessary

For the application involved here, the internal clamp of the ignition igbt is not necessary. A snubber circuit is used with a single external clamp. Neither is the low switch drive required, as a result of the higher available voltage V_c under all conditions – including 6V engine cranking.

On the other hand, the 600V rating of the general purpose igbt allows for the use of lower coil turns ratio N of 75 instead of 90 and provides higher output voltage of 42 *versus* 36kV. Also, its switching time is acceptably fast as seen above.

The lower turns ratio reduces the peak output voltage upon switch closure and reduces the number of coil secondary winding turns. This allows a heavier, lower cost wire and an easier to wind secondary to be used. In addition, the cost of the general purpose 600V, 20A, standard switching speed igbt, quoted at under $1 a unit, is lower than that of the 400V igbt.

As a result, the general purpose igbt is the preferred candidate for this particular application. But note that in other ignition applications, the low drive and internal clamping of the 400V ignition igbt may make it the device of choice despite its higher cost. Such applications are the coil with integral switch, and a new ignition application developed by us that is yet to be announced.

With the general purpose 600V igbt, a saturating inductor and a low-loss active snubber, it is possible to achieve an energy storage two and a half times times the value of 60mJ. A further effective increase of 20% to three times the energy, or 180 mj, is possible as a result of certain parallel improvements in coil design arising from a new approach, discussed later.

Ignition circuit details

Analysis of the HBI system features, shown in **Fig. 2**, begins with the supply capacitor C_p. It is charged to voltage V_c by the power converter, and is typically made up of two or more high-efficiency, high-temperature electrolytic capacitors which store many times the coil energy.

Included in the capacitor is a current sensor advantageously located at its low end, with a simple form of current sensing system. This sensor is based on a resistor and n-p-n transistor Q_s whose base-emitter junction is across the sense resistor with grounded base. The base-emitter junction voltage V_{be} decreases with temperature to provide a preferred lower current and energy with temperature to conform to typical engine ignition requirements.

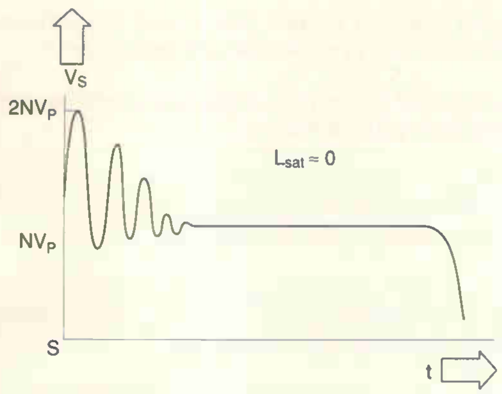
The next relevant component is the low-cost saturating inductor L_{sat}. In combination with the low turns ratio N and preferred large spark gap made possible by the higher output voltage, this inductor eliminates the problem of false firing of the spark plug upon switch closure. False firing is aggravated by the higher voltage V_c.

The inductor is designed to have an initial inductance approximately equal to the coil primary inductance L_p, and then made to drop rapidly with current to an inductance of one order of magnitude smaller. Its function is to reduce the high-frequency peak voltage $V_{s(0)}$ upon switch closure, equal to $2N \times V_p$, to one half its value, **Fig. 3**.

This high-frequency, voltage doubling effect was discovered and mathematically analysed by us in 1989 (see references).

Assuming a lower turns N of 75 through use of the 600V igbt, the output voltage reaches a peak value of only 3000V on switch closure. This is less than the minimum spark gap breakdown voltage.

Our HBI coil has a short energising or charge time T_p of



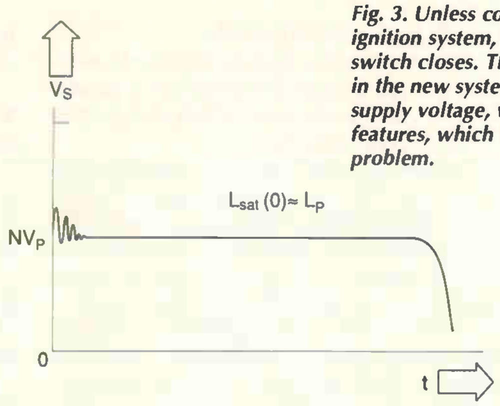


Fig. 3. Unless counter measures are taken, in an ignition system, false firing can occur when the switch closes. This problem would be aggravated in the new system due to the significantly higher supply voltage, were it not for L_{sat}, and other features, which completely eliminate the problem.

approximately 400μs versus 1.5 to 5ms for standard ignition. Because of this, the time of switch closure will occur later in the engine compression stroke at a higher cylinder pressure to further reduce the possibility of false firing.

The preferred larger spark gap of 0.060in made possible by the higher output voltage of the HBI system also helps reduce false firing – i.e. sparking upon switch closure. Moreover, energy associated with the saturating inductor at the end of the charge time is not wasted, but returned in large part to the power supply by the low loss active snubber discussed next.

Low-loss active snubber

The third key circuit feature is the low-loss active snubber already mentioned. It reduces switch-opening losses and provides other benefits.

The snubber is made up of isolating diode D_i, snubber capacitor C_{sn}, a p-type fet switch, snubber inductor L_{sn}, and return diode D_f. It serves two main functions. First, it smoothes and lowers the high-voltage transient that occurs when the switch opens. This transient is associated with the energy remaining in the saturating inductor and in the coil leakage inductance. It delivers the energy onto the snubber capacitor, minimising switching losses and electromagnetic emissions.

Secondly, the active snubber delivers most of the energy stored on the capacitor back to the power supply. This minimises clamp, switch, and resistive losses, as well as raising the ignition system efficiency, in turn reducing demands on the power converter to maximise system efficiency.

The snubber capacitor is selected to reach a peak voltage V_{sn} less than the clamp voltage V_{cl}, from transfer of the circuit leakage energy to the capacitor. Typical ignition parameters for the HBI system operating at the preferred 42V (V_c) are,

L_p=0.4mH
I_p=32A
A_p=1.35cm^2
N_p=60
N=75
B_p=1.55T
E_p=200mJ
E_s= 150mJ
T_p=0.35ms

A higher core area is assigned to the HBI system to compensate for its smaller coil size, **Fig. 4**. In addition, a higher flux density is assigned due to a better lamination design and the larger effective air-gap which allows for greater stressing of the core.

Peak current drawn from the battery is comparable to that of today's inductive ignitions, ranging typically between 5 and 10A. It has a very low ripple resulting from using a fast-switching power converter design.

Controlling the switch

To complete the HBI system, a particular simple and low cost form of switch controller and driver has been developed, Fig. 2.

Assuming multiple switches for one-coil-per-plug ignition, the controller holds the gates of ignition coil power switches low. It does this by means of an n-type fet switch S_c, whose drain interconnects the igbt gates through isolation diodes.

On ignition firing, i.e. the controller receiving a conditioned trigger signal at the inverting input of a control comparator, control switch S_c is turned off. This enables the gates of the igbts and sequences an octal counter.

The counter provides a turn-on signal through a resistor to one of the igbts to begin energising its associated coil. This continues until the set peak current I_p is reached. At this point, sense transistor Q_s is turned on, pulling the comparator inverting-input low and turning on control switch S_c.

Control device S_c disables the switches by rapidly discharging the igbt input capacitances C_{ies} for rapid turn-off of the energised switch. It also disables the remainder of the igbt switches.

Supply voltage for the octal counter is derived from the output voltage V_c. This is because there would not be enough voltage to drive the gate of the igbt at the lower battery voltages encountered in automotive systems during cranking.

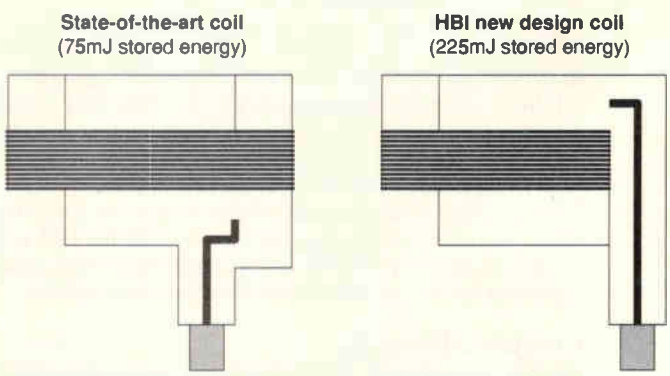


Fig. 4. State-of-the-art coil on the left has a two-piece welded core, 10000 turns of 42AWG secondary and a difficult-to-implement HT terminal. The new coil on the right has a one-piece E core, 5000 turns of thicker, easier to wind wire on its secondary and provides easier access to the HT end of the secondary. Benefits are,

- *One piece lamination requiring no welding and easy packaging.*
- *20% lighter weight coil due to the open E design.*
- *40% lower primary wire turns, i.e. 60 versus 100 turns, ideally suited for a two layer winding.*
- *50% lower secondary wire turns and use of 36 versus 42AWG wire which is much cheaper and easier to handle on a winding machine (less breakage) and takes half the time to wind.*
- *20% lower cost due to the above listed advantages despite the three times higher energy.*

Performance with low-volatility fuels

Once this new strategy for inductive ignition is realised, numerous advantages follow. A particularly interesting one is the ability to fire the ignition with closely spaced high-energy spark pulses under engine cranking conditions.

Long-duration spark firing during cold start is possible due to the rapid charge time T_p, which is an order of magnitude shorter than that of conventional inductive ignition. A much higher spark energy and longer spark duration are provided. This allows the ignition of lower volatility fuels, such as further reformulated gasolines, neat alcohol, JP5 fuel – used because of its reduced fire hazard – water-gasoline mixtures, and others.

Low volatility fuels can have a major impact on volatile organic compound, or VOC, emissions. As cars become progressively cleaner and tail-pipe emissions are reduced, evaporative VOCs become the principal hydrocarbon emission.

Evaporation of fuel upstream of the engine is projected to account for approximately 80% of the total VOCs from new cars sold in California in the year 2003. The remaining 20% will come from the tail-pipe. Ignition of low volatility fuel can be accomplished even at low battery voltages of 5 to 6V. Such voltages occur in extremely cold climates during engine cranking.

In addition, 4kV diodes may be used at the coil secondaries, allowing for even closer spaced spark firing pulses, further improving cold-start capability. Such diodes are cheaper to implement than the 8kV types.

To take full advantage of the low-cost, high voltage diodes, high inductance magnetic core spark plug wire is preferred to limit peak spark breakdown currents to under 10A. Use of such wire also reduces electromagnetic interference.

Other advantages of the new system

A particularly important advantage of the HBI system is the development of a simpler and lower cost coil resulting from the low coil primary inductance dictated by the new approach. The low inductance permits use of single open E-core coils versus the two piece welded design shown in Fig. 4.

When considering the benefits mentioned in Fig. 4, keep in mind that new coil delivers around 150mJ of spark energy versus approximately 50mJ for the current inductive ignition.

Table 2 compares ignition characteristics of the HBI system against various standard and state-of-the-art ignition systems. These were measured, in part, by a reputable car manufacturer conducting testing of the various ignition systems.

Where biasing magnets in the magnetic cores are needed to reduce size, this can be done more easily with the HBI coil because of its larger inherent air gap, retaining its three times higher energy density over the standard inductive ignition.

Unexpected benefits

There are other significant advantages than that of the ignition coil which follow from implementation of the HBI system. Most of these were not anticipated at the outset of the work to develop the HBI system and therefore only add to the motivation for use of this new system.

- Higher peak voltage to allow use of larger spark gap to further improve ignitability.
- Greater resistance to spark plug fouling due to the much lower inductance, faster rise times, and higher energy.
- Elimination of the need for computerised dwell – making the ignition universally retrofitable.
- Elimination of power switch (igbt) operation in the linear mode with its high dissipation and possibility for destructive oscillations due to unstable feedback loop in linear mode.
- Lower overall switch dissipation despite the much higher energy and peak currents as a result of the rapid charging of the coil and use of the low-loss snubber.
- Higher spark current characterised by higher spark power and greater resistance to spark segmentation by the high engine flows in modern high-output engines

Table 2. Comparison between standard, state-of-the-art and HBI ignition systems reveals high energy output and efficiency of the new system.

Ignition	Inductance	Resistance	Spark energy	Coil weight	Efficiency
HEI	6.6mH	0.4/9500Ω	40mJ	750g	<20%
Waste Spark	3.5mH	0.4/6000Ω	35mJ	260g	25%
Make 1	3.0mH	—	60mJ	265g	—
Make 2	2.5mH	0.6/—	50mJ	190g	35%
Make 3	2.8mH	0.7/7800Ω	50mJ	210g	35%
HBI	0.4mH	0.15/500Ω	150mJ	200g	50%

Three important questions remain regarding the new HBI system – what is the system cost, can it easily be implemeneted on a typical four-cylinder engine and what fuel efficiency gains can be expected?

Cost comparison

System cost can be analysed in a preliminary, general way by comparing costs of the three new critical systems of HBI against cost savings associated with the power switches and ignition coils. This assumes that biasing magnets are used in the coils of the conventional inductive ignition to bring its spark energy to 100mJ – a value which is still short of the 150mJ offered by the new system.

From data on switch costs, savings of $1 to $3 are achieved based on the prices quoted for the 400V, 20A ignition igbt which would be required for the improved standard ignition, versus that quoted for the 600V general purpose igbt. For the new ignition coils, 20% reduction results from the use of one piece lamination, half the number of secondary turns, 36AWG wire versus 42AWG, and removing the cost of biasing magnets.

The estimated overall cost savings are equal approximately to the cost of the additional components making up the HBI system. That includes the power converter with its input filtering capacitors, output capacitors, the transformer, the fet switch, and output diode. It also includes the saturating inductor, the active snubber and various other low-cost components.

Elements of the HBI system, i.e. power converter, saturating inductor, and the low-loss snubber, and the associated igbt switches can be mounted in the controller box. Preferably though, they would be mounted in a separate box located near the ignition coils, but sufficiently remote to avoid overheating.

On the other hand, predictions have been made by Kassakian (MIT), Wolf (Mercedes-Benz), Miller (Ford), and Hurton (G.M.) as reported in *IEEE Spectrum*, August 1996, that 42V will be made available in the passenger car engine compartment for operating motors and actuators. If this is the case, the HBI system can be fully integrated with an equal cost or cost savings despite its numerous advantages and benefits.

Fuel economy

On the issue of improved fuel economy, with high-energy ignition, extensive experience over a 20 year period – which has included testing at GM, Mazda Motors, Lucas Industries, and Chrysler Corporation – has shown fuel economy to be improved by up to 20%. This is based on dilute combustion, principally lean burn.

Typically, the fuel economy gains would be largest under light load where air-throttling losses are high, and minimal at high loads. A conservative estimate over the typical driving cycle is 6% assuming air dilution is used at the light loads where NOx is low, and EGR is used at the higher loads where NOx is high.

These figures assume a full implementation of the HBI system which includes, in addition the electronics described, a larger spark gap and better penetration of the spark plug tip in the combustion chamber. Placing the plug deeper in the cylinder allows the spark to interact more fully with the high in-cylinder air flows, resulting in higher output at cold start, etc.

A 6% improvement in fuel economy on an average of 25 miles per gallon would bring the mileage closer to the current CAFE standard of 27.5 miles per gallon – a target which car companies are currently missing. It translates to a 1.5 miles per gallon improvement, or a saving in CAFE penalties of $75 per vehicle – a far greater cost than any additional cost associated with the HBI system.

In terms of gasoline savings over the life of a car, assumed at 125 000 miles, this comes to a savings of 300 gallons of gasoline, which in the US would amount to $400 based on a fuel price of $1.33 per gallon and in Europe and Japan to $1000 based on a price of $3.33 per gallon. Even the smaller cost savings in the US would be well over ten times the projected additional small cost increase of the HBI system.

In summary

A new form of inductive ignition system has been developed which provides the required three times higher energy density of state-of-the-art inductive ignition systems and numerous other benefits not initially anticipated when this development program was begun some two years ago.

This new inductive ignition uses a higher, independently generated, constant voltage source of 42V – the value projected as best suited for other vehicle systems. It incorporates a standard speed, low-cost, 600V general purpose unclamped igbt as the switching element. It also uses a low cost saturating inductor to handle the switch closure false-firing problem and an active low loss snubber to minimise switching losses, rfi and system heating, as well as maximising system efficiency.

Complementing the system is a unique, low cost, 85% to 90% efficiency high-output power converter based on the use of state-of-the-art low cost 60V fets, a ferrite core transformer, a 150V ultra-fast output diode, and a simple form of controller made up of a dual comparator.

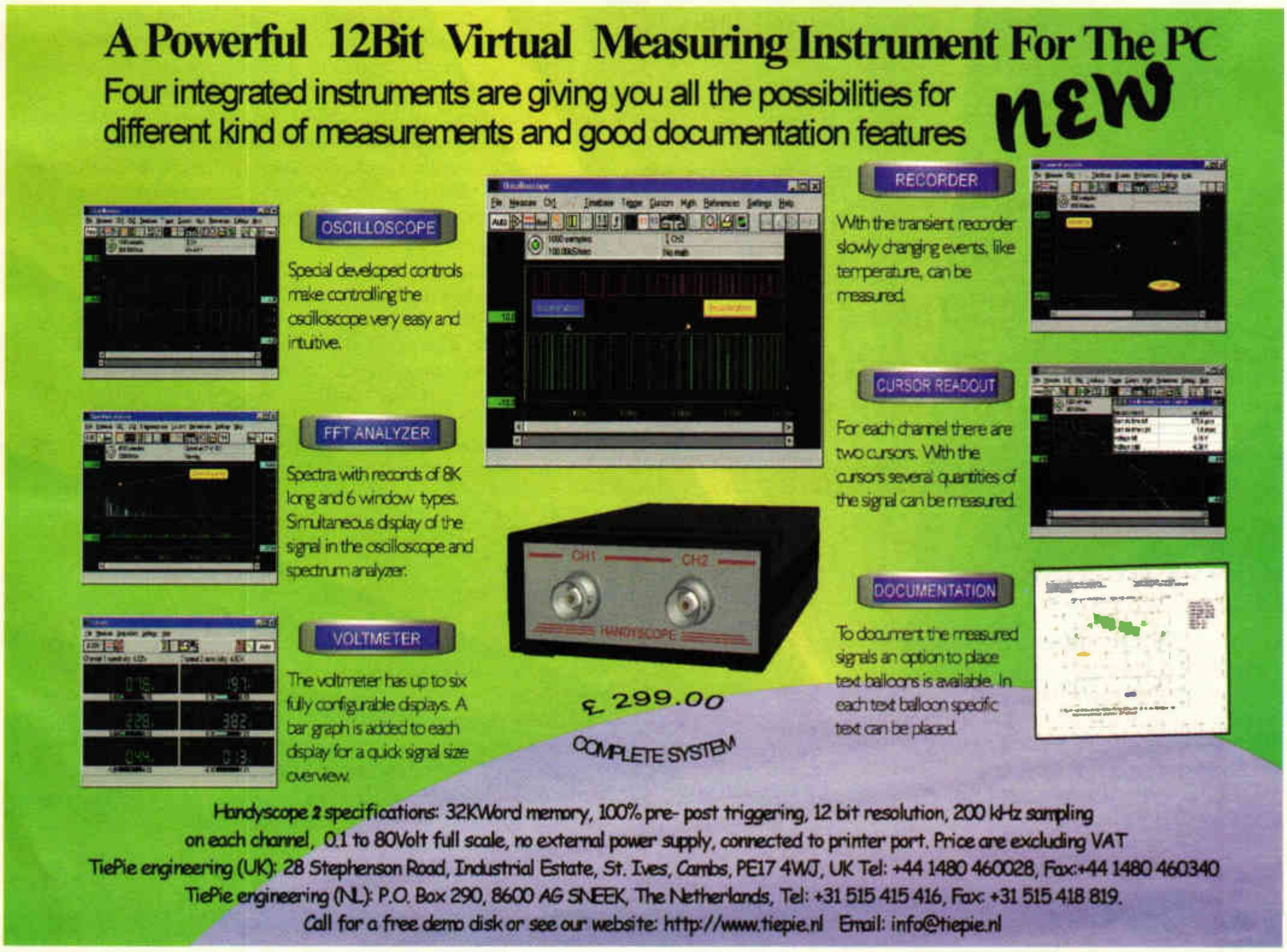
Applied to current engines, HBI provides the small, high-energy coils of the one-coil-per-plug ignition the industry seeks, and a range of other benefits. These include short dwell time, fast rise-time, resistance to plug fouling and high efficiency.

More importantly, given the current growing concerns over fuel efficiency and emissions, HBI has the potential for up to 10% improvement in fuel economy and large savings in capital investment to car manufacturers willing to revitalise their current power plants – such as their twin-valve push-rod engines – to make them cleaner and more efficient.

My thanks to Dr Tony Jarret of Lucas Industries and Joe Franklin, both of London, England, for their unwavering support of this work. ■

Reference

Ward, M, 'A new spark ignition system for lean mixtures based on a new approach to spark ignition,' Society of Automotive Engineers' Technical Paper Series, No 890475, Warrendal, PA.



Hands-on Internet

Cyril Bateman presents tips for speeding up your searching and downloading, and an update on circuit simulation software.

Internet provides designers with the ability to locate then download sources of needed information into a personal computer. Tracing through random links can sometimes be surprisingly successful, but methodical use of a search engine is more reliable.

Benefits of the net are maximised by quick search responses and fast download times for Web pages or files.

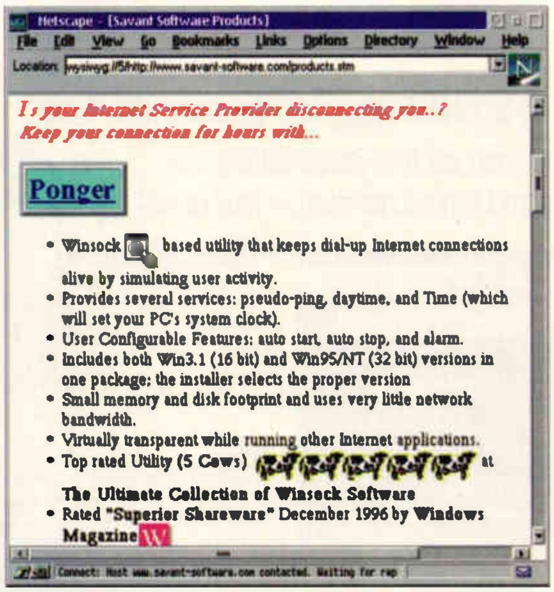


Fig. 1. 'Ponger' is designed to help maintain your Internet connection to prevent aborted file loads and the like. If moving your mouse or sending 'Pings' improves your connection, try this utility.

Internet access can be extremely slow during peak times, and of late, this slowness is extending well outside the peak periods for many popular sites.

Traditionally, one sent a 'Ping' to check that an address was 'on-line'. I now find sending regular 'Pings', say one a second, can help in maintaining my route. A utility called 'Ponger', which runs on Win3.1 or Win95, can be obtained from Tucows[1], the specialist Winsock archive, to automate this task, **Fig. 1**.

Two years ago relatively few Web pages existed, so today's popular search engines had not been devised. Archie, Gopher, Veronica and Wais[2] were the methods then used to locate sources. FTP was the only practical method available to download files from Internet.

One year ago, Java applets and the Netscape2 browser were introduced. Direct software downloads using a browser became commonplace, effectively supplanting use of FTP. As the numbers of Web pages increased, the AltaVista, Excite and Infoseek keyword search engines were developed to supplement the older directory searching methods.

There has since been an explosion of Web pages, to the extent that some companies now host 1000 Web pages on their site. This rapid expansion has resulted in major changes to all search engines – the largest now maintaining indexes covering more than 50 million Web pages and more than two-million URL site addresses.

This enormous increase of activity is linked to the trend towards larger and more glossy corporate Web pages,

Where to surf

1. The Ultimate Collection of Winsock Software. http://tucows.ip.pt
2. Surfing with intent. C.Bateman. *EW&WW* June 1995
3. Micro-Cap V version 2.0 http://www.spectrum-soft.com/down/demo.zip
4. Software Explorer http://www.windowscentral.com/reviews/internet/getright
5. Headlight Software. http://www.headlightsw.com
6. Tracing a Route Back http://noc.nwn.noaa.gov/traceback/traceback.html
7. Vital Signs Net.medic http://www.vitalsigns.com/product/netmedic_preview
8. HotBot Search Engine http://www.hotbot.com
9. LookSmart Search Engine http://www.looksmart.com
10. News.Com http://www.news.com/News
11. PCWeek Online. http://www8.zdnet.com/pcweek/news
12. Demo of Browser Security Hole http://neurosis.hungry.com/~ben/msie_bug
13. Number One Systems Ltd. http://www.numberone.com
14. Mental Automation Inc. http://www.mentala.com/SUPERCAD.HTM

large graphical images and extensive use of Java applets. My present Internet access speed is frequently no faster using a V34 modem than it was two years ago with slow hardware. A recent and growing tendency for links to work initially quite quickly then slow considerably – or worse, simply drop out – has become apparent when downloading Web pages or transferring files.

In the November issue, I reported on the new Micro-Cap V software. I abandoned several attempts to download the 1.7 Mbytes evaluation version[3] because my Netscape browser closed down the transfer before completion of the file. On one occasion this happened when the transfer was more than 90% complete.

One obvious solution is to use a traditional FTP client, but FTP from a Web page is only possible if that page provides an FTP address. Having only an http link, my FTP client could not be used to download the Micro-Cap V software file.

When a browser download slows or stalls, requesting a second download which you immediately cancel, often seems to 'wake-up' the route. Sending regular 'Pings', even continuously moving your mouse, sometimes helps maintain the route. Combining these techniques, I completed the download.

The Software Explorer page[4] recently reviewed a potentially better transfer method for browsers. The 'GetRight' utility is a mini FTP client add-on which can even permit resumption of a broken transfer. Windows 3.1 users should first read the special instructions on the Headlight Software[5] page.

Routeing

Internet's growth has resulted in two aspects contributing to these problems. You may have noticed that the time taken by your browser to look up Internet addresses has increased. These addresses are stored in increasingly large subnet tables on dedicated servers, used to determine the routeings used.

Transfer from the Web server to your computer does not happen in a direct fashion, but as a number of 'hops'. Each hop determines the next hop location. Of course Internet uses many supplementary routes, rather like having a number of bucket chains, with each handler passing its bucket to whichever chain is next ready to accept.

A secondary problem results from data packets queuing at these routers. When their buffer memory overflows, routers lose data packets. Packet loss is an increasing problem, thought to result in e-mails going missing and to contribute to file transfer problems. While not generally relevant, any data packet which requires 255 or more hops to reach its destination, is automatically killed.

'Ponger' and 'GetRight' are palliatives. A diagnostic approach is needed to identify whether problems are caused by Internet, your provider or your phone line or hardware. Most modern operating systems provide usable utilities. 'Ping' has already been mentioned; Traceroute, IPtrace and TCPdump, or variations on these names, can help.

Traceroute can identify your outgoing Internet routing, but the return route might be quite different. 'Traceback' is a more difficult problem when running from your own computer, but it can be run using software at the NCDC site.[6] My recent test indicated 18 hops were needed from Spectrum-Soft, several packets were lost and seven routeings delayed, **Fig. 2**.

NetMedic[7] is an excellent diagnostic utility and Internet tuning aid for Win95/NT. Its 'hop' monitor can be installed onto Netscape's toolbar, for continuous use, **Fig. 3**.

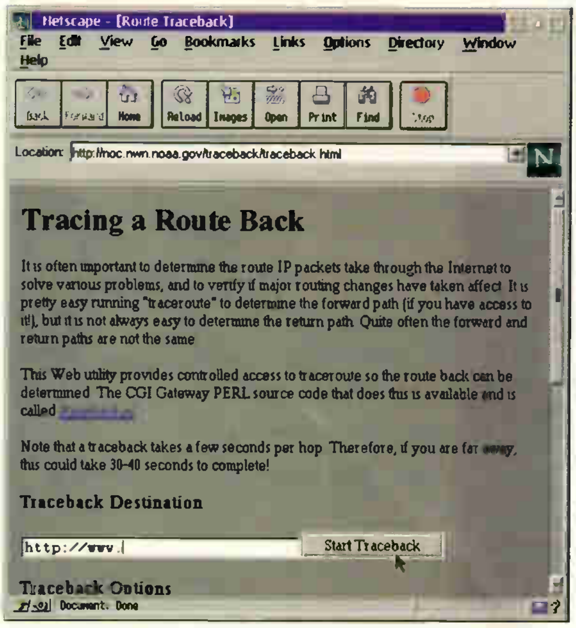


Fig. 2. If slow transfers bug you, route tracing can highlight problems. Find out how many lost packets and exactly where your delays occur.

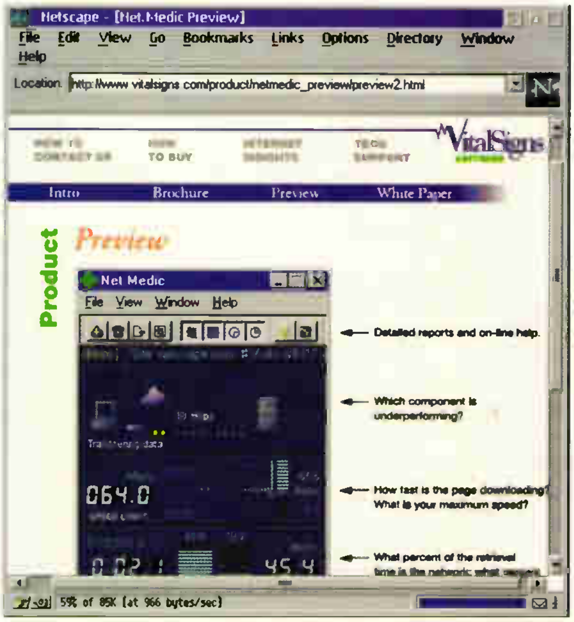


Fig. 3. 'NetMedic' provides short term and diagnostic long term logging of on-line problems. Using Win95 or NT, find where and when your Internet response problems occur.

Search engines

Launched in May 1996, HotBot is Wired's entry into the search engine market. It uses Inktomi technology and has more than 50 million Web pages indexed.[8] It provides an easy menu selection for your desired search parameters, **Fig. 4**.

LookSmart[9], launched by Reader's Digest in October 1996 using AltaVista technology, was redesigned in June this year and provides two search modes. Clicking on one of the border categories reveals more menu selections, thus combining the AltaVista keyword virtues with a simplified directory search.

More bugs

Following on from last months WinNuke problem, two bugs, affecting Microsoft or Netscape's browsers have been reported. *News.Com*[10] reports of 5 and 11 September detail the bug discovered by the Massachusetts Institute of Technology. This problem affects Microsoft's Explorer 3, and its beta 4 version, software. It results from the Java

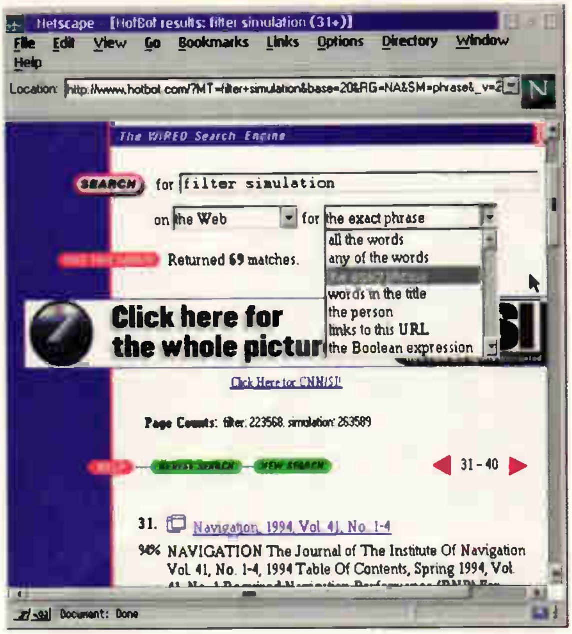


Fig. 4. HotBot – Wired's search engine – now indexes more than 50 million pages. Offers easy search control to finds results others miss.

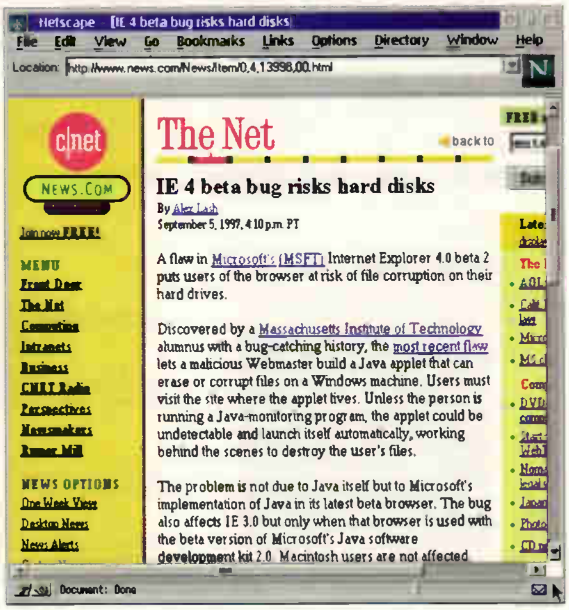


Fig. 5. Essential reading if you use IE3.0 or IE4 beta browsers. Macintosh users are not affected but Windows users should install the final Explorer4 version.

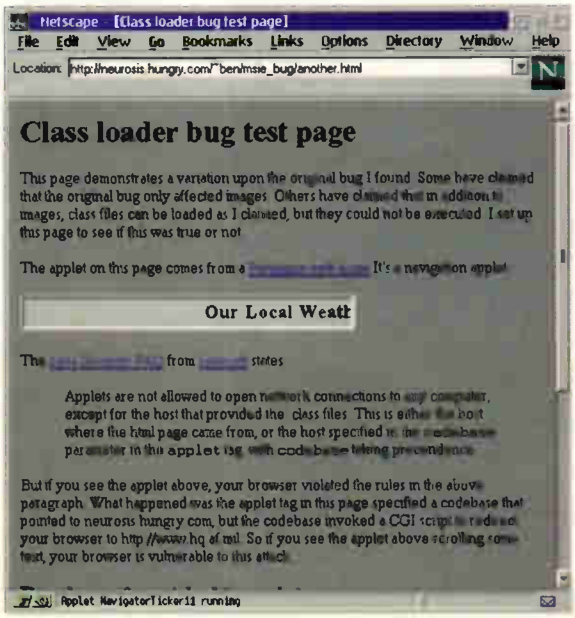


Fig. 6. Java applets cannot open connections to other hosts – or can they ? Prove or disprove your browser's Java security in the safety of this test site.

methods used within the browser, not a function of Java itself, **Fig . 5**.

A *PC Week* article[11] dated 12 September details that Microsoft is actively removing all Java applets from its Web sites. This is apparently part of a campaign to speed up the downloading rate for its pages. Where needed, Java applets will be replaced by HTML code or Javascript.

The second bug, discovered by Ben Mesander and reported by *News.Com* 8 and 12 August, affects both Microsoft and Netscape browsers. It can be demonstrated using Ben's[12] test page which houses two demos, demonstrating the two variations.

Ben first uses a harmless applet to redirect and download an image from Microsoft's pages. For security, Java applets are only permitted to communicate with their host page, so should not transfer this image. I tried my machine, which passed this test.

Ben's second test demonstrates the 'Class Loader Bug'. In this test, the applet tries to download a harmless text 'ticker' again by redirection, from http://www.hq.af.mil. I also tried this out on my system, running Netscape 2.02 under OS/2 Warp4, and failed this test. The running ticker can be clearly seen.

While Ben's tests were deliberately designed to be harmless, others could as easily use these techniques, to cause harm, **Fig. 6**.

Simulation and design software

Not all circuit simulation software is based on Spice. Many vendors specialise in providing frequency-domain-only simulation to extremely good effect.

One such system originates from Number One Systems[13] – a long established UK developer. My very first commercial simulator was their original *Analyser* of 1983, running on the *BBC model B* microcomputer.

Since then I have regularly relied on the company's Windows suite of integrated packages, for quick look-see simulation and final pcb design. Number One's unique *Layan* simulator is able to extract printed circuit layout traces as circuit elements. These combine with conventional elements for Analyser simulation. If desired the results can be exported seamlessly to their Smith-chart program.

If you are looking for a low-cost, easy-to-use integrated suite of Spice based simulator, schematic editor and pcb layout tool, which you can download and evaluate before purchase, *SuperCAD* from Mental Automation[14] is worth a look. The company's *SuperSpice* simulator displays results using a virtual oscilloscope like display, having volts/division adjustment for each of four traces. Mental claims its *SuperCAD* schematic editor is fast becoming an industry standard.

Circuit applications

The Mc Graw-Hill Encyclopedia of Electronic Circuits is a cd rom containing 1000 circuit designs taken from industry leaders.

Also developed by Mental Automation this rom is based on the popular six-volume book series from Mc Graw-Hill and the circuits used were selected by the original author of the series, Rudolf Graf. Each circuit can be viewed or printed out using the supplied schematic viewer or can be edited with the *SuperCAD* schematic editor. ■

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A window on analogue i/o

Colin Attenborough **outlines a 12-bit analogue input and output subsystem that runs via the pc's printer port. Colin is also making available his Visual Basic software for controlling the port via Windows.**

In the April 1997 edition of *Electronics World*, I showed how to read and write 16-bit digital words through a standard pc printer port, under the control of the Visual Basic language. I used a dynamically-linked library, written in C, to drive the hardware.

This article extends the earlier one to transfer 12-bit analogue data through the printer port – again under Visual Basic control. My software is available on disk; you need only Visual Basic and the disk to control the hardware. For interest, I have added the source code for the necessary dynamically-linked library on the disk, but you do not need a C compiler to use the software.

The circuit of the analogue input/output hardware shown here can be built on its own or added to the digital i/o subsystem previously described. Both clock and data lines use the same pins on the printer port in both cases. Analogue and digital functions available in Visual Basic when the analogue and digital circuitry of the present and previous article are combined are listed in the appendix.

Levels between 0 and 2.5V are accepted by the analogue input; the d-to-a converter output generates between 0V and 4.095V.

Analogue input

The analogue section of the *LTC1286* a-to-d

Subroutines and functions available to Visual Basic

These subroutines and functions are included in my software, which is available on 3.5in disk for £12.50 fully inclusive by writing to Department VB2, Electronics World Editorial, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. You need Visual Basic only to run this software.

With the two combined, you can control the analogue i/o port hardware shown via the graphical user interface shown in the screen shot here.

The address of the printer port is represented by, _port. This is usually 378_{16}, but watch the screen as the computer wakes up to check. Function are as follows.

`initialise(int _port)`
Sets the d-to-a converter chip enable, and clears all other lines on the printer port.

`shutdown(int _port)`
Clears all lines on the printer port.

`AdcOn(int _port,int _d)`
Sets the a-to-d converter power line to turn on the a-to-d converter, and takes some dummy readings to allow the supply voltage to settle. Variable _d is the delay parameter already mentioned.

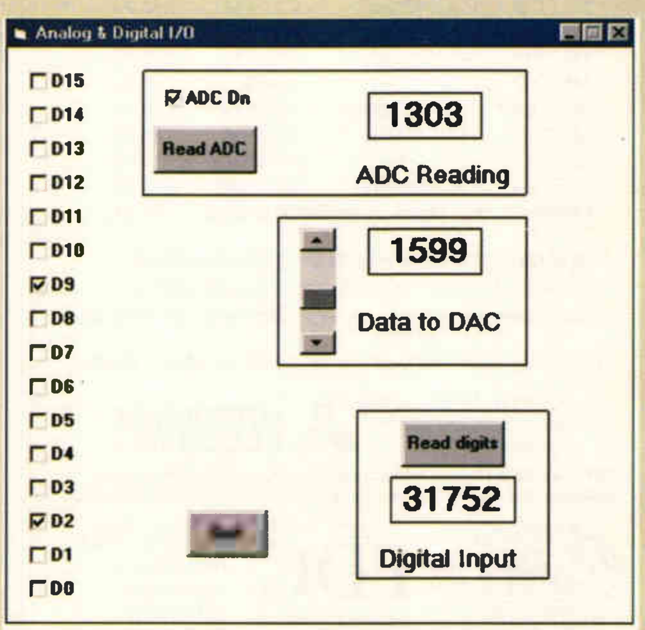
`AdcOff(int _port)`
Clears the a-to-d converter power line to turn off the a-to-d converter.

`in_A(int _port,int _d)`
Returns an integer representing the a-to-d converter output. Variable _d is the delay parameter.

`in_D(int _port)`
Returns a long representing the state of the digital inputs.

`out_A(int _port, int _data)`
Gives an output of `_data` millivolts from the d-to-a converter.

`out_D(int _port,long _data)`
Sets the digital outputs to `_data`.



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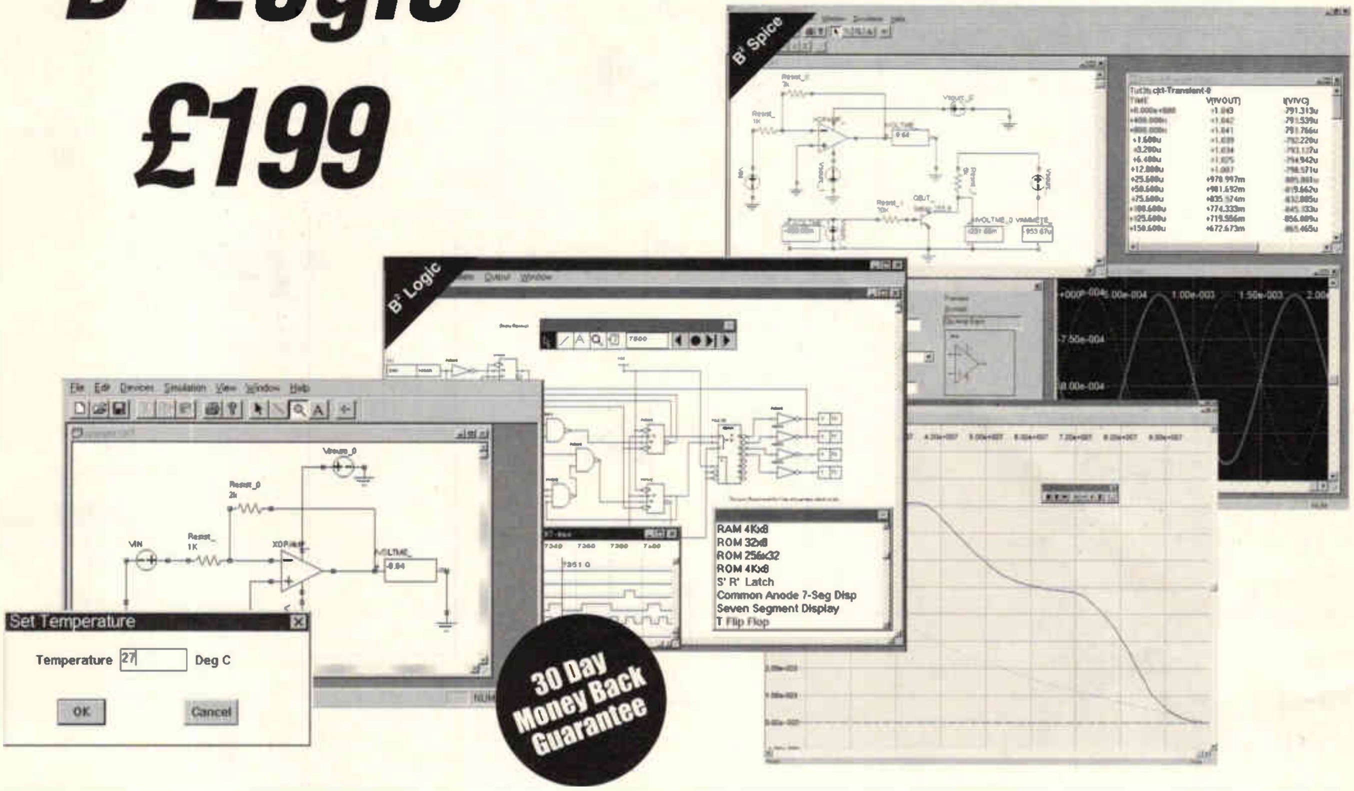

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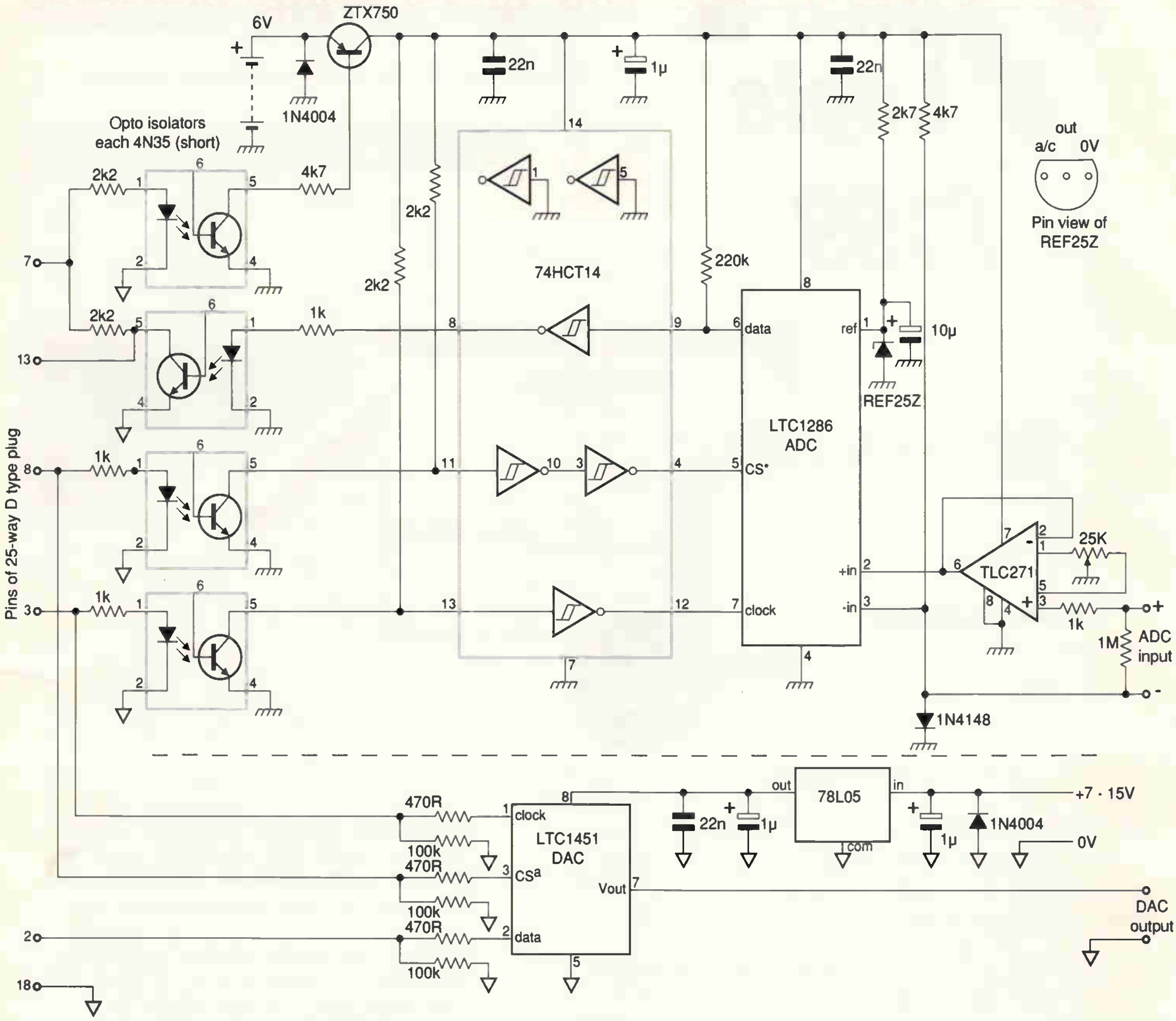


Fig. 1. Designed for battery-powered operation, this 12-bit accuracy analogue-to-digital and digital-to-analogue i/o subsystem is controlled via the pc's printer port. Optoisolators not only provide protection, but also allows floating measurements and removes ground-loop problems.

converter accepts a differential input voltage and a voltage reference; when appropriately clocked, it gives a serial digital output of,

$$code = \frac{4095 \times V_{in}}{V_{ref}}$$

By default, chip enable is at logic 1 to power down the a-to-d converter. To make a conversion, clear chip enable to logic 0 and apply fifteen clock pulses; read data at the rising edge of each clock pulse, most significant bit first, and then set chip enable to logic 1.

Opto isolators convey the chip select, clock and data signals; a fourth opto isolator switches power to the whole analogue input system. Since the system is powered by a battery – of typically four AA cells – power conservation is important.

I've chosen to isolate the analogue input section from the computer via opto isolators; this allows floating voltage measurements to be taken, and avoids earth loops with their attendant errors.

The a-to-d converter is most accurate when fed from a low source resistance, so a *TLC271* op-amp is inserted as an input buffer. This op-amp also has the virtue of buffering the voltage being measured from the capacitive switching input current spikes which occur during conversion.

To allow the op-amp some headroom (shouldn't that be footroom?) I have exploited the differential input of the a-to-d converter, putting the negative input a diode drop above the system 0V line. This means that the op-amp output is always comfortably above its negative supply.

The system needs a supply voltage of between 4.5 and 6V. While the optoisolators specified have a breakdown voltage of 2.5kV,

Did you read Colin's previous article?

Colin points out that there was an error in the listing in his article in the April 1997 issue, near the start of page 333. The line,

```
Declare i_o Lib "<full path
name>\IO8.dll" (ByVal ip As Integer,
ByVal op As Integer, ByVal Written)
As Integer
```

should have read,

```
Declare Function i_o Lib "<full path
name>\IO8.dll" (ByVal ip As Integer,
ByVal op As Integer, ByVal Written)
As Integer
```

This omission did not occur in the software on the disk that we offered. Apologies.

other factors may limit the voltage that can safely be applied. Layout of whatever pcb or other form of construction is used in the area of the optoisolators is an example. *In any case, consider your own safety when there are large voltages across the optoisolators.*

Converter setup and accuracy issues

I have deliberately run the optoisolators at low currents to extend battery life. The penalty is that the comparatively high load resistors mean that the rise times at the phototransistor outputs are quite long.

To guarantee correct clocking, I have included a delay parameter in the function which reads the a-to-d converter. This is just a counter for a delay loop; a large value gives a large delay. Start with a large value, of say 2000, read the a-to-d converter with a known input voltage, and reduce the delay as far as possible while reliably getting the same output reading.

The delay does *not* affect accuracy. With a 100MHz 486 machine, a delay parameter of 250 gave stable results; the conversion time was about 1.5ms. Faster machines will need a larger value for the same a-to-d converter clock pulse width.

The offset adjustment potentiometer should be adjusted so that with no input voltage the output code flickers between 0 and 1.

Input voltage is defined by,

$$V_{in} = \frac{code \times V_{ref}}{4095}$$

If you want to obtain best accuracy from the system, take into account that the reference generated by the *REF25Z* can be between 2.475 and 2.525V. If you have a sufficiently accurate voltmeter you can measure it and take into account its real value when you use the converter. Alternatively, you can calibrate it against a known voltage with a voltage source more accurate than the *REF25Z.*

Analogue output

I've used an *LTC1451* serial d-to-a converter. Rather like the a-to-d converter, chip enable, referred to as CS* in circuit diagram, is at logic 1 by default.

To convert a digital data stream – most significant bit first – to an analogue output voltage, clear chip enable to logic 0 and apply twelve clock pulses, with the bits of the data stream valid at the rising edge of the clock pulses, and then set chip enable to logic 1.

The d-to-a converter gives an output of,

$$V_{out} = 2V_{ref} \times \frac{code}{4096}$$

For the nominal reference voltage of 2.048, this simplifies to,

$$V_{out} = code \times 1\text{mV}.$$

There is a maximum offset error of ±12mV at room temperature; the nominally 2.048V reference has a maximum error of ±40mV.

I would like to thank Cambridge Consultants Ltd for permission to publish this article. ■

You may know which capacitor technology to choose for a given task, but do you know why it's the best choice? Having spent thirty years designing, specifying and troubleshooting capacitors, Cyril Bateman shares his expertise.

Understanding capacitors

To my surprise, recent review of my publications library revealed a scarcity of capacitor articles. Perhaps I should have anticipated this since on reflection, many text books and circuit simulators all consider capacitors as ideal, so why delve further. On the other hand, capacitors have also been called 'strange devices', i.e. *unfamiliar*, which this article seeks to remedy.

Thirty years of experience as a capacitor design and applications engineer has convinced me that over-stressed or mis-applied capacitors are involved either directly or indirectly in most circuit failures. All components wear out in time, but mis-applied capacitors can fail extremely quickly. Worse still, before the capacitor ultimately fails, it can directly contribute to failure of semiconductors, masking the prime failure mechanism.

The best way to understand unfamiliar components is to perform measurements on representative samples, then dismantle them to understand their differing constructions. But with capacitors, not all the significant details are immediately apparent.

Capacitor overview

The fundamental definition of capacitance relates to charge and voltage, measured statically. Descended from the Leyden jar, a capacitor is essentially an energy storage device. Since its stored energy can be charged or discharged extremely quickly, alternative common usage ac definitions have been derived, as discussed in the panel entitled 'Defining capacitors'.

Of these definitions, the most important relates a capacitor's impedance to frequency,

$$|Z| = \sqrt{R_s^2 + (Xc_s - Xl_s)^2} \quad (1)$$

Where,

$$Xl_s = 2\pi F L_s \quad (2)$$

and,

$$Xc_s = \frac{1}{2\pi F C_s} \quad (3)$$

Throughout this article, the series equivalent expressions are used, denoted as Xc_s, R_s etc., unless otherwise stated.

At any frequency, the term $Xc_s - Xl_s$ can be simplified to jX_s, giving the fundamental vector equation for impedance, $Z = R_s \pm jX_s$.

This vector equation leads to the common usage expressions for impedance magnitude $|Z|$,

$$|Z| = \sqrt{R_s^2 + X_s^2}$$

and phase angle,

$$\theta = \tan^{-1}\frac{X_s}{R_s}$$

This phase angle definition results in the second most important quantity for capacitors – loss factor,

$$\tan\delta = \text{abs}\frac{R_s}{X_s}$$

An ideal capacitor would have a phase angle, hence tanδ, that remained constant regardless of frequency. Since by definition, eqn 3, a capacitor's reactance, Xc_s, is totally dependent on frequency, so also must be the capacitor's series resistance, R_s, known as esr. This is clearly demonstrated by the results measured for a very high quality polystyrene capacitor, **Table 1**.

Table 1. Clear demonstration of how esr changes significantly with frequency. Measured results of a high quality 10nF polystyrene foil/film capacitor, made using a four terminal Wayne Kerr 6425 precision component analyser at a test voltage of 1V.

Frequency	Capacitance (nF)	Tanδ	'Q'	ESR (Ω)
100Hz	9.9982	0.00010	9000	17.0
1kHz	9.9988	0.00005	20 000	0.8
10kHz	9.9986	0.00015	6000	0.26
100kHz	10.0000	0.0005	3000	0.05

What is a capacitor ?

Any two conducting surfaces, separated by an insulator, exhibit capacitance. The value of this capacitance increases with surface area and reduces with separation.

The fundamental definition of capacitance assumes this insulator is a vacuum, thus directly relating to the permittivity of free space.

$$\text{Capacitance} = \text{Free space permittivity} \times \frac{\text{Area}}{\text{Separation}}$$

But area and separation alone are insignificant compared to the contribution provided by a change of insulation material. Each insulation material used is rated for dielectric constant, or

'K' value. This represents its permittivity relative to free space.

In practical materials used to make commercial capacitors, this 'K' ranges[1] from 1.00059 for air, 2 to 4 for plastic films, 8 for aluminium, 28 for tantalum electrolytics, and from 6 up to 12000 for differing ceramic formulations.

Given this wide range of dielectric constants, it is natural to loosely categorise all capacitors by the dielectric material used. While in general this is valid, the final capacitor's size and electrical properties also vary according to the construction methods used.

For a given dielectric, thickness – or rather thinness – is all important in determining the capacitance achieved in a given physical size. Usable dielectric thinness is limited by its ability to sustain the required voltage as well as surviving manufacturing methods. Common plastics which can be manufactured in micrometre thicknesses[2], withstand around 500V dc per micrometre, at room temperature, short term.

Since the dielectric for electrolytic capacitors is 'formed' in situ on the base foil, it doesn't need a minimum thickness to provide handling strength, as do plastic films. Together with certain multilayer and disc ceramic construction techniques, electrolytic dielectric thinness is limited only by the voltage that needs to be sustained.

To minimise the final size, the dielectric together with its conducting surfaces – the electrodes – may be compacted using winding, stacking, folding or layering techniques. These processes lead to supplementary descriptions including wound, stacked and multilayer.

Electrodes formations

The conducting surface for paper or plastic dielectric materials is produced by one of two common techniques. In one of them, an extremely thin, visually transparent coating of metal, generally aluminium, is vapour deposited on to the insulator in a vacuum chamber, resulting naturally in the description metallised capacitor. This method however cannot be used with polystyrene.

When the dielectric itself is not metallised, the alternative and electrically superior method uses thin flexible soft metal foils or discrete metallised plastic foils, as electrodes. These foils are interlaid with the dielectric during assembly.

The surface of ceramic dielectrics is generally made conductive by coating with suitable metal inks. After air drying, these are 'fired' at high temperature.

With aluminium electrolytics, the aluminium base anode foil has the dielectric oxide, i.e. Al_2O_3, pre-formed electrolytically on its surface prior to assembly. The anode foil acts as one conducting electrode.

This electrolytically formed dielectric's thickness is self regulating, attaining some 14Å for each volt applied in manufacture. At 10V working for example, this represents only 0.02μm dielectric thickness (1μm, or micron, is 0.00394in). The true second electrode is the electrolyte material with which the separating tissue paper is impregnated.

Assuming a polarised capacitor, connection to the electrolyte is made using a second, usually thinner, aluminium foil or cathode. While this cathode is not deliberately formed, it inevitably possesses a much thinner naturally occurring aluminium oxide, electrically equivalent to a few volts. In this way, a pair of back to back capacitors is created. One has the desired capacitance and voltage capability while the other has a much greater capacitance but only a few volts capability, applied in reverse.

With a non-polarised, or bi-polar capacitor, this unformed cathode is replaced by a second deliberately formed anode foil. This results in two capacitors back to back, usually having the same capacitance and voltage capability.

Depending on the desired life-length characteristics, the pre-forming voltage used can range from 1.2 times to more than double the working voltage, trading size and cost against leakage current, and hence endurance.

Properties of capacitors

Every capacitor needs conducting surfaces, or electrodes, which inevitably have some intrinsic resistance. Having a physical cross section and length equates automatically to inductance, called self-inductance. As a result, a capacitor must be represented as a series *CLR* combination circuit.

This *CLR* network results in a circuit whose resonant frequency depends on the capacitor's value and construction. At the resonant frequency, the capacitor represents a dc blocking low value resistance, and a dc-blocking but inductive reactance at frequencies above this resonance.

At any frequency, a perfect capacitor, having neither resistance nor inductance, would sustain a voltage in quadrature with the applied current. Analysed on a polar display, voltage would be –90° and current 0°, so the complementary angle delta would be zero. Having no resistive element, this perfect capacitor cannot lose or dissipate energy.

The self-inductance results in a voltage at +90°. This subtracts magnitude from the capacitive voltage, increasing the apparent capacitance value which would be measured.

With a near perfect capacitor, the above resistance, appearing in series with the capacitive reactance, degrades this –90° angle, resulting in the complementary angle delta increasing. This change in phase angle represents the resistive element which dissipates energy as heat in the capacitor, **Fig. 1**.

With capacitors, it is usual to refer only to this delta loss angle, which is generally described as tanδ.

Dielectrics

Unfortunately, all dielectrics other than vacuum contribute their own particular degradations from this near ideal capacitor. In general the more highly stressed the dielectric, in volts per micron, the greater the degradations.

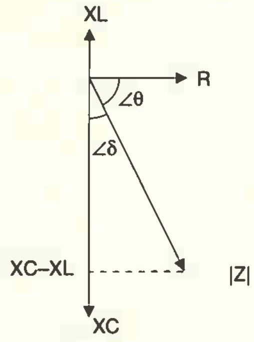


***Fig. 1.** This simple vector diagram shows the relationship between tanδ and cosθ. Both are essentially equal only with small loss angles. The interaction between the capacitance Xc and inductance Xl vectors is also visible.*

Since the characteristics of both vacuum and air capacitors depend principally on the particular insulators used to support their assembly, they are excluded from these discussions.

Depending on the symmetry of its molecular structure, each plastic dielectric can be described as having either polar or non-polar characteristics. A symmetrical-molecule plastic, or other non-polar dielectric, has electrical characteristics effectively constant with changing frequency, and exhibits minimal dielectric absorption effects.

If this polymer molecule is not symmetrical, it has a dipole moment resulting in increased dielectric constant. Similarly, ferroelectric ceramic crystal poles and domains can produce extremely high dielectric constants. Such types are known as high-K dielectrics.

Both result in polar characteristics, i.e. a capacitance that reduces and a tanδ that increasing with frequency. These are functions of the basic materials used – not to be confused with the constructional terms polarised and non-polarised or bi-polar, as applied to electrolytic capacitors.

While manufacturing techniques cannot change a polar dielectric into a non-polar, they can enhance the polar effect. Stressed with dc voltages, polar materials exhibit a reluctance to accept or release their full charge instantaneously – a behaviour called dielectric absorption.

What is dielectric absorption?

If a capacitor is fully charged, to say 10V, for a considerable time, briefly discharged using a short circuit, then left to recover, a voltage is found which develops with time. The ratio of this resultant voltage compared to the initial charge voltage, is described as dielectric absorption.

In past years dielectric absorption was only

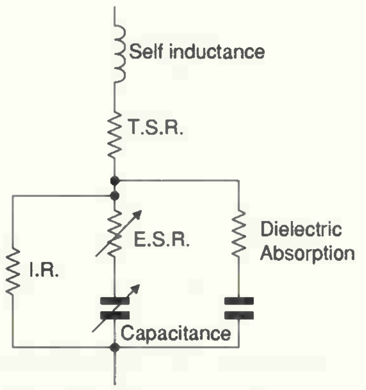


Fig. 2. Full equivalent circuit for a practical capacitor. In practice esr includes the tsr resistance, which is only shown separately to aid explanation.

considered relevant to high voltage capacitors, as a safety hazard. With modern techniques it becomes relevant at very low voltages.

Measured values of dielectric absorption vary with measurement technique, but more importantly with dielectric stress applied, in volts per micron. Typical figures range from 0.05% for polystyrene[2], 0.25% for polyethylene terephthalate, or PET, to 5% for aluminium.

Some dielectric materials also have a small degree of piezoelectric effect, which can result in a voltage when they are mechanically constrained, whether physically or from temperature changes while rigidly or surface mounted. Some care is needed to differentiate this effect from that of dielectric absorption.

Both effects are small and are irrelevant for most applications. However when working with sample/hold circuits, charge amplifiers, etc., capacitor choice should account for these effects.

According to manufacturing techniques used, every practical capacitor exhibits some dielectric absorption effect. However polystyrene or polypropylene film/foil and COG NPO dielectric ceramic capacitors, show minimal effects.

With the inclusion of this dielectric absorption mechanism, it is now possible to deduce the equivalent circuit for a practical capacitor, **Fig. 2**.

How temperature affects capacitors

Ceramic COG NPO type capacitors are restricted to ±30ppm/°C change of capacitance with temperature and are the most stable available. With the exception of these, all dielectrics show easily measured larger capacitance changes over their temperature range.

General-purpose ceramic capacitor dielectrics are categorised under the EIA classification scheme. The popular X7R material is thus restricted to a box envelope allowing ±15% change in capacitance over its working temperatures. However the exact profile within this envelope, differs with manufacturer.

Change in temperature also results in a change of measured tanδ for most common dielectric materials. Non-polar dielectrics show very small changes, but with a polar dielectric, the maker's data should be consulted.

The behaviour of plastic-film capacitors however tends to be consistent according to the materials used and less dependent on specific manufacturer.

Voltage effects

Many polar dielectrics have a capacitance which changes with applied ac or dc voltage. With voltage, capacitance tends to increase above that measured at 1V then declines. Since this behaviour depends on the precise dielectric chosen and manufacturing technique, makers data should be consulted.

Frequency effects

As shown in published data, almost all capacitors exhibit a frequency-dependent capacitance change. Less well known, the dielectric strength or voltage withstand of film dielectrics, can reduce[2] with increasing frequency. For many applications this is not important since the power or current rating constraints which should be applied, dominate.

With pulse waveforms having large peak-to-mean ratios, power constraints no longer dominate, so it is essential to consult makers' data when choosing a capacitor for pulse duty

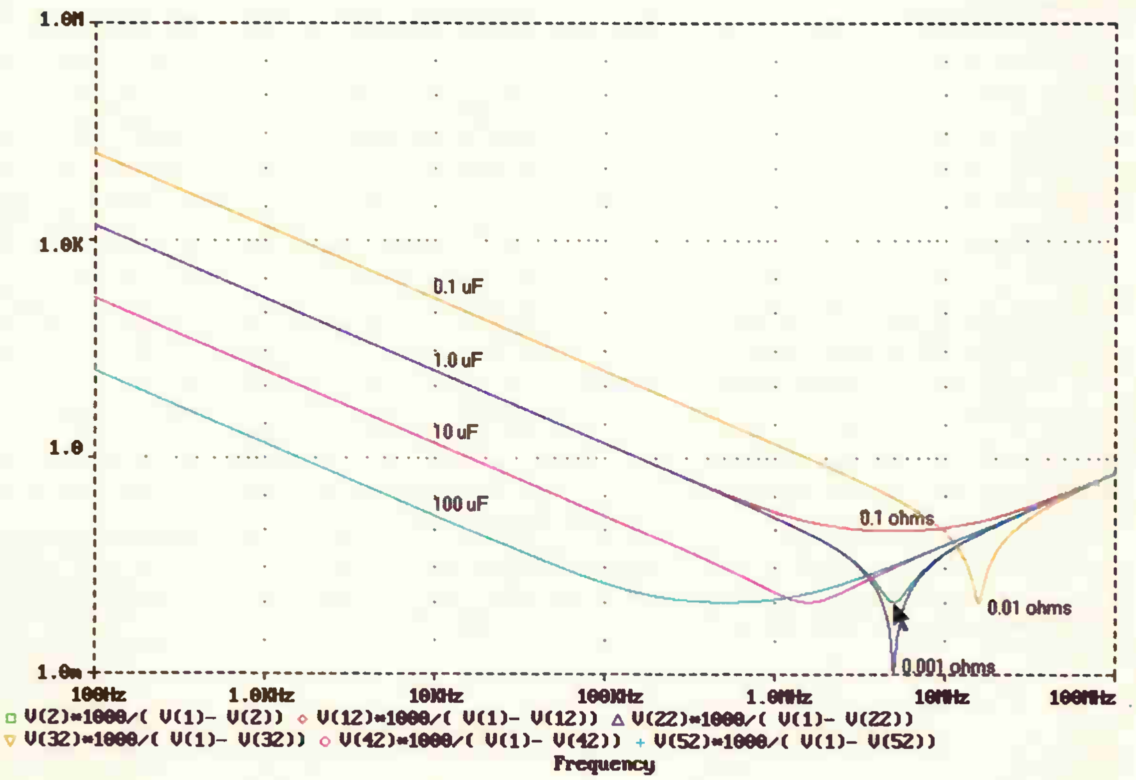
More significant for general applications, every capacitor regardless of dielectric has frequency dependent tanδ losses, which increase with frequency.

Power limitations

Having mentioned power ratings, surely since current and voltage are in quadrature, no power is dissipated in a capacitor?

I mentioned that a near ideal capacitor has

Fig. 3. This PSpice simulation shows the affect of 1nH inductance with 0.01Ω esr or tsr on four near ideal capacitors, 100µF through 0.1µF. The 1µF capacitor highlighted also shows comparison of increased esr of 0.1Ω and reduced esr of 0.001Ω, giving dramatically changed plot shape, with small values of esr.



conducting surfaces with a finite resistance. These electrodes must connect to the outside world using one of a variety of means, inevitably adding resistance. The sum of these metallic resistances, true series resistance, or tsr, is a very low value fixed resistance. For convenience, since it cannot be otherwise separately measured, it is sometimes viewed as being the minimum impedance, seen at resonance, on a conventional impedance and frequency plot, **Fig. 3**.

However other, usually much larger, loss resistances are found in practical capacitors. Every capacitor exhibits a leakage current, which is voltage and temperature dependent. A rule of thumb based on the Arrenhuis law, is to assume this current doubles for each 10°C increase in temperature. Obviously this current can be represented by a high value insulation resistance in parallel with the dielectric. More conveniently, following the parallel-to-series conversion rules, it equates to a very small fixed value series loss resistance.

With most capacitors, the dielectric's tanδ, frequency, voltage and temperature related losses dominate. These are also expressed as series resistance.

The sum of all these resistances, at any one frequency, is called the equivalent series resistance or esr. As can readily be seen, esr is frequency, voltage and temperature dependent, so when quoted in makers' data, it only applies to the particular conditions quoted.

To aid understanding, it is perhaps easier to consider esr as a combination of the above leakage resistance and tsr, together with a

Defining capacitors

The fundamental definition[1] of capacitance *C* in farads is the ratio of charge acquired in coulombs *Q* to the applied voltage *V*.

$$C=Q/V$$

For two flat conductive plates separated in vacuum, the fundamental capacitance formula becomes,

$$Capacitance = Free\ space\ permittivity \times \frac{Area}{Separation}$$

where free-space permittivity, ε0, is 8.854188×10^{-12}F/m

All other insulating materials have a dielectric constant, K, relative to this free space value,

$$Capacitance = K \times Free\ space\ permittivity \times \frac{Area}{Separation} \quad F/m$$

$$= 0.0885 \times K \times \frac{A}{S} \quad pF/cm$$

In essence any two conducting surfaces, separated by an insulator, exhibit capacitance, with a value which increases with surface area and reduces with separation. Subjected to an ac voltage, an ideal capacitor resists or impedes the passage of current according its reactance value,

$$Xc = \frac{1}{2\pi FC}$$

developing a voltage lagging the current by 90°.

In practice, every capacitor incorporates a resistive component. Expressed using the conventional series notation, this resistance further resists or impedes the passage of current,

$$Z = \sqrt{R^2 + Xc^2}$$

Each practical capacitor also incorporates an inductive element, which incorporated in the above provides the full equation for impedance,

$$Z = \sqrt{R^2 + (Xc - Xl)^2}$$

where $Xl=2\pi FL$.

At any given frequency, the term $Xc-Xl$ can be simplified into its series equivalent, j*Xs*, giving the vector equation, $Z=Rs\pm jXs$. This results in the magnitude/angle expressions,

$$|Z| = \sqrt{R_s^2 + X_s^2}$$

$$\angle\theta = \tan^{-1}\frac{X_s}{R_s}$$

In the above expressions the *R* term represents the equivalent series resistance of the capacitor, while the *X* term represents the series reactive component. When viewed as a vector diagram, a polar plot, or on a Smith chart, this *X* term has a negative value for capacitors, Fig. 1.

The commonly used expressions,

$$\tan\delta = \text{abs}\frac{R_s}{X_s} \quad \text{and} \quad C = \frac{-1}{2\pi FX_s}$$

also apply.

Series or parallel?

The impedance vector of a practical capacitor at any one given frequency can be represented using an equivalent circuit of the device with a resistor. The resistor, used to degrade the phase angle to that measured, can be either a high value in parallel with the device, or a low value in series with the device, leading to the term 'equivalent series resistance' or esr.

While the parallel equivalent values have use for certain calculations, the series equivalent values are more commonly used. Throughout this article, the series equivalents are used unless otherwise stated.

Take this practical example. An impedance vector, magnitude 100Ω and phase angle –84.3° at 1kHz, represents a capacitor having a tanδ of 0.1 and a Q of 10. This vector would result from a series combination of 9.95Ω resistive and –99.5Ω reactive, i.e. a 1.6μF capacitor or a parallel combination of 1005Ω and 1.584μF. A difference in equivalent capacitance value of 10%. The equivalent series resistance would be 9.95Ω.

Parallel impedances

Certain measuring instruments or mathematical calculations are more suited to the equivalent parallel expression, which can easily be converted to or from the series values.

$$R_p = \frac{R_s^2 + X_s^2}{R_s} \quad \text{and} \quad X_p = \frac{R_s^2 + X_s^2}{X_s}$$

Sometimes the measured results are needed as admittance rather than impedance; conversion from the parallel impedance expression is simple,

$$Y = \frac{1}{R_p \pm jX_p} = G_p \pm jB_p$$

$$G_p = \frac{R_s}{R_s^2 + X_s^2} \quad \text{and} \quad B_p = \frac{X_s}{R_s^2 + X_s^2}$$

The conversion from parallel impedance back to series impedance format, following,

$$R_s = \frac{R_p \times X_p^2}{R_p^2 + X_p^2} \quad \text{and} \quad X_s = \frac{R_p^2 \times X_p}{R_p^2 + X_p^2}$$

is equally simple.

number of variable contributions dependent on frequency, voltage or temperature. This esr, together with the alternating current passing through the capacitor, can be used to calculate the power dissipated in the capacitor.

When the capacitor is subjected to a sinewave, simple calculations suffice. Given a complex waveform, the only method[3,4] that ensures accurate results is to use Fourier transforms to characterise the waveform into its discrete frequency components.

Only when a capacitor is measured at its final working frequency, temperature and voltage can its esr be derived directly from bridge measurements of tanδ and capacitance But in most applications it is not practicable to make bridge measurement under such conditions. Consequently esr must be estimated taking account of each variable in turn.

I have stressed that esr is frequency dependent, but does it really change by a significant amount, or am I simply being pedantic?

Consider the esr of a high-quality 10nF polystyrene foil/film capacitor. I selected such a device as one of the standard capacitors when building my capacitance bridge. All measurements were taken using a four terminal Wayne Kerr *6425* precision component analyser, with a test voltage set to 1V, Table 1.

These results show clearly how esr values do change significantly with frequency, for this high quality capacitor. Many writers on this topic have confused these esr and tsr terms. Obviously they differ substantially, except at that frequency when the capacitor is self resonant.

Since correct understanding of esr is essential to avoid over-stressing capacitors, I make no apology for labouring the point.

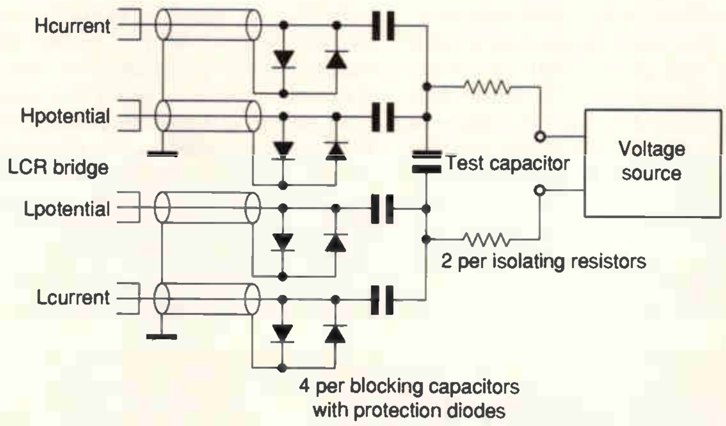


Fig. 4. This simple dc blocking buffer circuit allows measurement of capacitance change with applied dc volts. The isolating resistors should each present at least ten times greater impedance than the test capacitor. Likewise, the dc blocking capacitors must withstand the dc test voltage and have at least ten times the test capacitor's capacitance.

Implications of voltage ratings

Many years ago, when impregnated metallised paper capacitors were the standard workhorse, it was considered that a capacitor rated for 400V dc or above could be used on 250V ac mains. Since these capacitors were impregnated, this was just about feasible. Unfortunately this premise tends to continue even today.

When the then new unimpregnated metallised PET capacitors became commonly available thirty years ago, 400V dc parts were used for many of these 250V ac mains requirements. Result – misery. If you were lucky, the end terminations eroded, disconnecting the capacitor. If you were unlucky, the capacitor caught fire.

Even today, I have vivid recollections of this unhappy time, when my task was to withdraw from all 250V ac applications and re-rate the capacitors to 160V ac, on behalf of my employer, for this particular construction.

Why should this problem arise ?

Given an impregnated or otherwise solid, void free, capacitor construction, 250V ac and above causes no insuperable problems. However with non-impregnated non-solid constructions, air voids occur within the capacitor.

According to Paschens curve of ionisation, an air-filled void having optimum size and air pressure can exhibit ionisation inception at voltages as low as 185V ac. This is why 160V ac was adopted in the previously-mentioned application to provide a safety margin.

Once triggered, the ionisation current is self sustaining at lower voltages – in fact almost to zero volts. Thus once triggered, the resulting discharge continues for at least 50% of the periodic waveform.

This ionisation discharge is damaging to almost all dielectric materials, resulting ultimately in a short circuited capacitor.

From these experiences, international and national safety rules for class X capacitors, used across the 250V ac domestic mains, were developed. Two main capacitor styles emerged. These were a much updated resin impregnated metallised paper capacitor[5] and the two-in-series metallised polypropylene style, which worked since its two series capacitor elements shared the applied voltage.

Manufacturing measurements

National and international capacitor approvals require manufacturers' measurements to be 'true' values, i.e. traceable to nationally held standards. In general, this means that measurement equipment must have an inherent accuracy ten times better than any claimed component parameters. Measured values must be 'inset' sufficient to eliminate all known measuring equipment errors.

With low-loss or close-tolerance capacitors, these requirements are not easily attained. A test frequency of 1kHz is standard for capacitors of value greater than 1nF except for electrolytic types, which are generally tested at 100Hz. Resulting from their high impedance at 1kHz, capacitors equal to or less than 1nF are tested at 1MHz. In general, test voltages of 1V ac or lower are used.

Experimenters' measurements

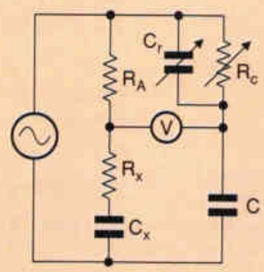
Commercially available capacitance test equipment can supply a DC polarising voltage to the component, but is generally restricted to a maximum of 20V dc.

However an adapter permitting much higher voltages can be simply made. Hewlett Packard application note 346 – a Guideline for Designing External DC Bias Circuits[6] – provides details of adapters for their ranges of precision meters. However these principles can easily be extended to any equipment

Wheatstone bridge

A capacitance bridge[7] that can be used at audio frequencies to directly measure capacitance and esr is shown. It is easily built using standard resistor and capacitor decade boxes together with one known high-quality standard capacitor.

With range switching resistors to replace R_A and three switchable standard capacitors, this circuit has been used to measure both capacitance and esr from a few picofarads to a hundred thousand microfarads. For best accuracy with the larger capacitors, four-terminal connections and switching of the low-impedance current paths is advised.



Power stress circuit

When designing switching supplies, much benefit derives from measuring a capacitor at its final working voltage and frequency. While low power measurements using a bridge are possible, other techniques are essential for higher powers or voltages.

These power or voltage limitations result because taking current in quadrature with voltage stresses any power amplifier. The amplifier's power dissipation increases rapidly when driving a capacitive load, leading to instability or failure.

If this capacitive load is resonated using a suitable inductor, the power amplifier is presented with a zero phase load. The inductor provides either the voltage needed to the capacitor, or the current. Since it also needs to supply only voltage or current and not both, amplifier power dissipation can be low, **Fig. 5**.

Using a series-resonant circuit, the amplifier supplies the required through current but at a much reduced voltage. With a parallel resonant circuit, the amplifier supplies the required voltage at a much reduced current. The inductor's stored energy supplies the missing voltage or current drive required by the capacitor. The amplifier only supplies the power needed to replace that lost due to capacitor dissipation, inductor resistive losses and the protection resistor.

Using a stable air cored inductor with a Q of ten or better, and a 100W mosfet audio power amplifier, capacitor voltages of 250V can be easily attained using the series circuit. An air cored inductance provides the stability needed for meaningful calculations.

Conversely the parallel configuration easily provides 5A. The schematic circuit suggests using a series resistor to protect the amplifier in the event of capacitor failure. Much higher levels are possible if this resistor value is reduced.

Capacitance change with frequency, voltage or current can be calculated from the circuit's change in resonant frequency, the voltage drop and through current measurements of the test capacitor.

which might need to be used, **Fig. 4**.

Commercially available capacitance testers can often measure using more than 1V ac test voltage, but they rarely extend above 20V ac. Higher test voltages are possible using Wheatstone bridge methods but care is needed not to overload either the test source or the bridge measurement arms – especially with increasing frequency.

A low accuracy alternative method which can measure to very high voltages and frequencies requires use of a power amplifier together with suitable high 'Q' inductors. Depending on whether higher voltage or greater current than can be sourced by the amplifier is needed, these inductors are used to either series or parallel resonate with the capacitor. Using these techniques, satisfactory measurements to 500V ac and several amps at frequencies to 1MHz have been performed, **Fig. 5**.

Capacitor life

Qualification testing requires capacitors to survive continuous operation at maximum ratings and maximum temperature for typically 1000 hours. In some instances the capacitors are required to be stressed in excess of claimed levels, or for much longer times. To understand the implied life-test claims, you need to read the specification.

Compared to actual end use 1000 – and even 10 000 – hours endurance is extremely short. But components in end use are not generally continuously stressed, certainly not to their maximum capability and temperature.

Arrenhuis law suggests that insulation resistances halve, alternately leakage currents double, for each 10°C increase in temperature. Consequently 1000 hours at 125°C can represent a useful life[4] under normal end use conditions, of 10 to 20 years – even assuming maximum applied voltage.

A secondary benefit results from reduced voltage. All capacitors, including electrolytics, exhibit prolonged life with reduction in operating voltage, even to zero dc, provided any applied ac does not otherwise contravene the capacitor specifications.

One common mis-statement – that electrolytics exhibit no capacitance with zero or reverse bias – is completely unfounded. The dielectric film is chemically robust and cannot in the short term be changed. Long term, assuming reverse voltage is within the permitted levels, leakage current increase can result in parametric failure.[4]

However if an electrolytic which has been operated for some time at reduced voltage, is then subjected to increased voltage, a temporary increase in leakage current results. The capacitor may then fail to meet its specification. Since the converse also applies, any capacitor subject to excess temperature, voltage or current or mechanical stress, will fail quite quickly.

One special aspect which certainly causes electrolytic failure, is unwitting excessive repetitive reverse bias. This can arise when a polarised capacitor is used to couple the drive waveform into a switching transistor base and at the same time block dc. This mis-use is especially common. I have personally experienced this failure mechanism many times, in both television and satellite receivers.

With switching power supplies, an early indication of an electrolytic being reverse biased is increased transformer noise and notable temperature increases. However with television line and frame timebase generators being less audible, the first indication can be either reduced drive amplitude or semiconductor failure.

Such abused capacitors usually show visible signs of overheating or electrolyte leakage, measurable loss of capacitance and increased esr and reverse voltage withstand. I hope to delve further into this topic in a subsequent article which explores and measures various capacitor constructions. ■

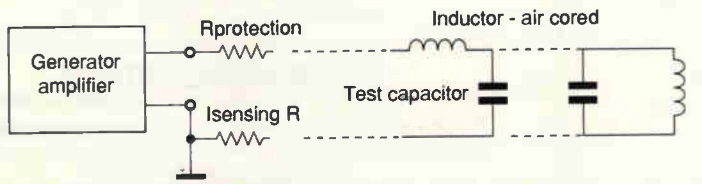


Fig. 5. Taking advantage of inductor Q to produce high test voltages on the capacitor while maintaining the amplifier load near unity power factor. Variations on this circuit enable much investigation into capacitor behaviour using only minimal equipment.

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T67110A-612M(7H2)-10M	595-640M	4.5	10.0M	(50)37	(50)32	40.0ns
T6796D 612M(7H2)-20M	590-640M	3.0	20.0M	(50)31	(50)25	23.0ns
T6751D 612M(7H2)-25M	595-640M	2.6	25.0M	(50)25	(50)18	20.0ns
T6862A 612M(7H3) 7M	590-640M	7.0	7.0M	(20)42	(20)31	75.0ns
T67112A 612M(7H3)-9M	610-640M	5.0	9.0M	(50)53	(50)50	67.0ns
T67104D-612M(7H3)-19M	590-640M	3.0	19.0M	(50)50	(50)38	37.0ns
T67108A-612M(7H3)-22M	595-640M	3.0	22.0M	(50)48	(50)35	32.0ns
T6848A-480M(7H2)-4.5M	460-515M	5.5	4.5M	(20)34	(20)26	76.0ns
T6746A 480M(7H2)-6M	460-515M	5.3	6.0M	(40)44	(40)43	72.0ns
T6793D-480M(7H2)-10M	460-515M	3.5	10.0M	(50)35	(50)28	38.0ns
T6746D-480M(7H2)-18M	460-515M	3.0	18.0M	(50)34	(50)24	26.0ns
T6747F-480M(7H2) 24M	460-515M	2.0	24.0M	(50)18	(50)19	19.0ns
T67159B-480M(7H2)-29M	460-515M	2.0	29.0M	(60)18	(60)60	14.0ns
T6849A-480M(7H3)-4M	460-515M	7.0	4.0M	(10)27	(10)19	92.0ns
T67101A 480M(7H3)-8M	460-515M	5.0	8.0M	(20)43	(20)32	72.0ns
T67101D-480M(7H3) 12M	460-515M	4.0	12.0M	(20)32	(20)24	51.0ns
T6774D-480M(7H3) 18M	460-515M	3.0	18.0M	(20)21	(20)14	35.0ns
T67158 480M(7H3) 23M	460-515M	2.0	23.0M	(60)38	(60)26	24.0ns
T6775B-480M(7H3) 30M	460-515M	2.0	30.0M	(60)38	(60)26	20.0ns
T6711A-410M/7H2)-7M	405-455M	4.5	7.0M	(40)40	(40)40	57.0ns
T6791D-410M(7H2)-10M	405-455M	3.0	10.0M	(20)20	(20)18	45.0ns
T6711F-410M(7H2) 13M	405-455M	3.0	13.0M	(50)38	(50)26	33.0ns
T6745F-410M(7H2) 18M	405-455M	2.0	18.0M	(50)35	(50)23	23.0ns
T67156B-410M(7H2 21M	405-455M	2.0	21.0M	(50)18	(50)28	20.0ns
T6799D-410M(7H3) 9M	405-455M	4.0	9.0M	(20)41	(20)30	66.0ns
T6772F-410M(7H3)-14M	405-455M	3.2	14.0M	(20)30	(20)20	46.0ns
T67157B-410M(7H3-17M	405-455M	3.0	17.0M	(50)44	(50)35	35.0ns
T6773F 410M(7H3) 23M	405-455M	2.0	23.0M	(50)50	(50)28	32.0ns
T6790A 360M(7H2) 5M	350-400M	4.5	5.0M	(40)38	(40)38	66.0ns
T6790D 360M(7H2)-7M	350-400M	3.5	7.0M	(50)46	(50)30	46.0ns
T6742F-360M(7H2) 10M	350-400M	3.0	10.0M	(50)40	(50)26	38.0ns
T6743F 360M(7H2) 20M	350-400M	2.0	20.0M	(50)27	(50)19	30.0ns
T6841A 360M(7H3) 4M	350-400M	8.0	4.0M	(10)35	(10)25	85.0ns
T6798D 360M(7H3)-7M	350-400M	5.7	7.0M	(10)23	(10)16	70.0ns
T6770F 360M(7H3)-10M	350-400M	5.0	10.0M	(20)33	(20)24	60.0ns
T6771F 360M(7H3)-20M	350-400M	3.0	20.0M	(20)17	(20)10	35.0ns
T6789D 325M(7H2) 6M	310-350M	3.6	6.0M	(10)16	(10)15	66.0ns
T6740A 325M(7H2) 8M	310-355M	3.5	8.0M	(30)45	(30)31	57.0ns
T6741B 325M(7H2)-14M	315-355M	2.0	14.0M	(50)26	(50)36	31.0ns
T6830B-325M(7H3)-4M	300-340M	8.5	4.0M	(8)30	(8)24	133.ns
T6797D-325M(7H3)-9M	300-340M	5.0	9.0M	(10)25	(10)18	80.0ns
T6769B-325M(7H3) 15M	310-355M	3.0	15.0M	(50)60	(50)46	35.0ns
T6831B-240M(7H2)-3M	220-245M	5.0	3.0M	(10)16	(10)25	94.0ns
T6701B 240M(7H2)-9M	220-245M	3.0	9.0M	(30)23	(30)35	50.0ns
T6735B 240M(7H2) 15M	225-250M	2.0	15.0M	(50)28	(50)38	34.0ns
T6832B 240M(7H3)-4M	200-245M	7.0	4.0M	(10)35	(10)38	160.ns
T6762B 240M(7H3)-9M	225-250M	3.0	9.0M	(20)39	(20)37	70.0ns
T6763B 240M(7H3) 15M	225-250M	2.5	15.0M	(50)68	(50)66	55.0ns
T6727D 140M(7H2) 5M	130-145M	3.6	5.0M	(10)14	(10)15	67.0ns
T6756B 140M(7H3) 2M	130-145M	8.0	2.0M	(8)33	(8)34	173.0ns
T6757B 140M(7H3) 4.5M	135-145M	6.0	4.5M	(8)26	(8)23	116.0ns
T6812A 110M(7H2) 3M	107-115M	4.5	3.0M	(20)30	(20)39	101.0ns
T6813A-110M(7H3)-3M	107-115M	6.0	3.0M	(10)40	(10)35	160.0ns
T67113D-89M(7H2) 3M	85-95M	5.0	3.0M	(10)27	(10)32	154.0ns
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Never mind the quality – feel the bandwidth

As MPEG video and audio compression becomes increasingly popular, there are signs that the quality is suffering. Here John Watkinson explains why compression artifacts occur and – more importantly – what can be done to reduce them.

John Watkinson, FAES, B.Sc., M.Sc.

In audio/visual program material, the advantages of compression are many – hence its popularity. Compressed material requires less bandwidth which is ideal for broadcasting, where the radio spectrum is under increasing pressure from other mobile services.

Compression also allows the cost of storage or recording to be reduced, although as recording economics improve year on year this may be a transient advantage.

Compression principles

Figure 1a) shows that in all real digital program material the bit rate is the product of the sampling rate and the word length. In practice the overall bit rate is made up of a varying mixture of unpredictable or novel material, known as entropy, and the remainder which could be deduced from first principles, known as redundancy.

An ideal compressor would separate the two perfectly so that only the entropy need be sent. An intelligent decoder would work out the redundancy for itself and reproduce the source signal without loss.

Entropy is a characteristic of the signal and varies. Figure **1b)** shows that if all of the entropy is not sent there is quality loss. The ideal is a variable rate channel which allows constant quality. If a fixed rate channel has to be used, the quality will vary.

In an MPEG-2 transport stream, several compressed signals can be statistically multiplexed together. It is unlikely that all will reach an entropy peak together, consequently a transport stream can be divided into a number of varying bit rate channels.

Provided that the overall bit rate remains constant, individual channels can demand more bandwidth when difficult material is encountered on the assumption that other channels are probably handling easier material at that time.

In the DVD – digital video disk, also known as digital versatile disk – a variable bit rate is supported in a single program stream simply by changing the rate of disk accesses.

Unfortunately the ideal coder of Fig. 1 is infinitely complex and has an infinite processing delay. Practical coders have to constrain both. When either of these constraints are applied, the bit rate has to go up to maintain quality, as Fig. **1c)** shows. **Figure 2** shows that for constant quality the bit rate will reduce as the latency increases.

Video compression

In MPEG-2, the temporal compression is obtained by sending motion compensated difference pictures, and further spatial compression is obtained by transform coding the differences.

Differential coding simply subtracts the previous picture from the current picture and sends the difference. When there is motion differences increase. This is handled by measuring the motion between pictures on a 16-by-16 pixel block, or macroblock, basis

Fig. 1. Entropy is always less than bit rate, a). To avoid quality loss, all entropy must be sent, b). The simpler or faster the coder, the more data that must be sent, c).

and transmitting a vector for each block. The decoder and the encoder both shift the previous picture using the vectors and only the difference between the shifted previous picture and the current picture need be sent.

Pure differential coding fails if there is a transmission error because that error propagates indefinitely. It also makes it hard for the viewer to change channel! In practice periodic whole pictures have to be sent to prevent error propagation and to create decoder entry points. These are known as intra-coded, or I, pictures because they make no reference to any other picture and are only spatially compressed. In between these I pictures differential coding is used.

Moving objects cause problems in differential coding because they reveal background at their trailing edge which is previously unknown.

This is overcome by using information from future pictures. **Figure 3** shows that in bidirectional coding a picture can be decoded using information from pictures before or after. The decoder does not need a crystal ball to obtain the future pictures: instead, pictures are sent out of sequence.

Figure 4 shows that after an I picture, a future picture is differentially coded in the forward direction only. This future P (predicted) picture is sent immediately and stored in the decoder.

Pictures between the I picture and the P picture can now be sent these B (bidirectional) pictures can be created by forward or backward motion compensated differences on an individual macroblock basis.

The pictures and difference pictures are spatially compressed. The process begins by performing a dct, or discrete cosine transform, which expresses an 8-by-8 pixel block as a set of 64 coefficients. In typical video material, many of the coefficients will have zero or negligible values so that only the significant ones need to be transmitted.

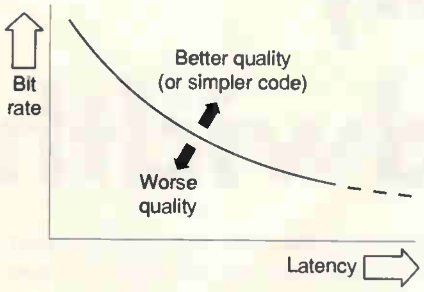


Fig. 2. Shorter latency needs a higher bit rate.

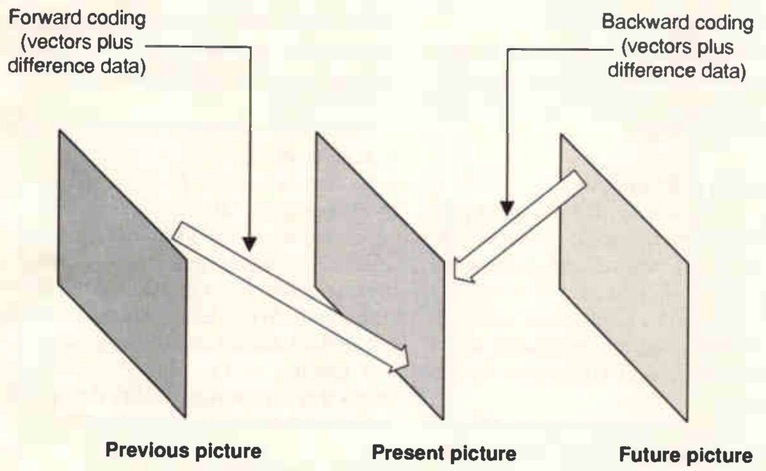


Fig. 3. Bidirectional coding uses information from both past and future picture frames.

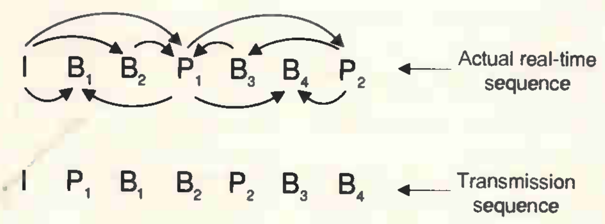


Fig. 4. Bidirectional coding requires pictures to be transmitted out of sequence.

Compression artifacts

Compressors are generally iterative and are driven by a bit-budget measurement. If the output bit rate is too high for the channel the dct coefficients will have to be expressed in fewer bits. In the case of large value coefficients, when low order bits are lost they become less accurate. In the case of small value coefficients, they may be truncated to zero.

This has a number of side effects. Coarse quantising of large value coefficients means that after the inverse dct at the decoder the eight-by-eight pixel block may have considerable errors in the sample values.

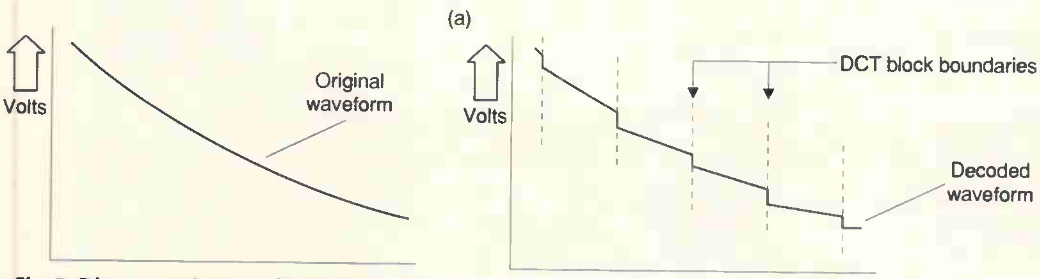
While these are not necessarily visible in themselves, the errors in adjacent blocks will mean that there is a discontinuity in the block boundaries so that the blocks become visible as shown in **Fig. 5a**). If high frequency coefficients have been truncated to zero the block will lack detail and resemble a tile as in **5b**).

This effect occurs in both the luminance and colour difference paths. In luminance the effect is called contouring whereas in colour the effect is called posterising, where gradual colour changes have been replaced by a limited colour set, as might be available in a box of poster paints. In MPEG the colour posterising can be quite obvious because the chroma blocks are the size of a macroblock and have four times the screen area.

Effects of truncating hf coefficients

Where high-frequency coefficients have been truncated to zero, the effect is to introduce ringing on edges. This is because an edge contains high frequencies and removing them is the equivalent of a sub-optimal low pass filter, hence the ringing. This is particularly noticeable on graphics and captions, less so on natural subjects.

When the prediction of the temporal coding fails, the data in the difference pictures will necessarily increase and this will force the compressor to quantise more heavily, raising the artifact level. This is particularly noticeable on B pictures since they are generally allocated only 10% of the data rate.



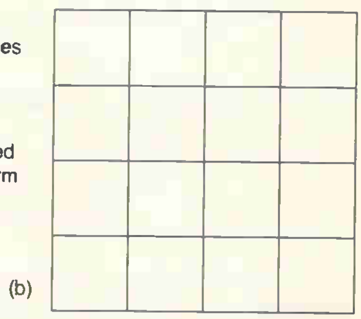


Fig. 5. Discrete cosine transfer blocks may not blend together if excessive truncation is used, a). Where high-frequency coefficients are also discarded, dct blocks become tiles, b).

Temporally difficult material, such as when frequent cuts are made, may overload an encoder. Cuts remove temporal redundancy and defeat bidirectional coding. Following a cut, several pictures may contain serious blocking artifacts.

The real MPEG killer material is video from a press conference where flashguns are firing. Each flash drives up every single pixel value for one picture, and then in the next picture the values come back to normal. This causes temporal chaos and most MPEG coders substitute a picture of the designer's bathroom wall under these conditions.

Pre-processing controls artifacts

The level of artifacts can be controlled by pre-processing. There are three levels at which a pre-processor can operate,

- By removing noise from the source material.
- By removing entropy from the source material.
- By aligning the I pictures in the coder output with the temporal entropy of the source.

Noise in a source pixel block creates more coefficients than a noise free source would. Thus all coefficients have to be truncated more aggressively to carry them, raising artifact level. Noise also increases data in difference pictures. Hence noise reduction will lower artifact levels by reducing spurious coefficients and reducing picture difference data.

If, after other steps, the artifact level is still too high, then the only approach is to restrict the entropy entering the coder. This is done by down-sampling the source images either spatially, so they contain less pixels, or temporally, so there are fewer pictures per unit time, or both.

In source material from telecine, the use of 2:2 and 3:2 pull-down creates what could be called false entropy, because in 2:2 frames are interlaced to make fields, giving a false doubling of picture rate.

The ratio 3:2 gets its name because 24Hz film frames are alternately converted to two and three fields to give a 60Hz output. One in five fields is redundant. Prior to MPEG coding telecine material has to have redundant fields discarded and remaining field pairs are de-interlaced to obtain the original frames.

The largest usage of data in MPEG is the I picture. This is because it does not use any previous information from the source. Consequently it makes no difference if the source I picture is radically different from the ones which went before. In contrast both P and B pictures will require significantly more difference data if there is a cut.

It follows that a significant reduction in artifacts can be obtained if I pictures are temporally aligned with source cuts. The only drawback of this approach is that to do it in real time a great deal of memory is needed to pipeline a stack of frames so that picture type decisions can be taken.

The alternative is to use a time coded source recording and use a two-pass encoding process. On the first pass the cuts are detected and used to design a picture type structure which is stored, and on the second pass the structure is implemented. ■

John is an independent consultant in digital audio, video and data technology and is the author of fifteen books on the subject, including Compression in Video and Audio. He is a Chartered Information Systems Practitioner, a Fellow of the Audio Engineering Society.

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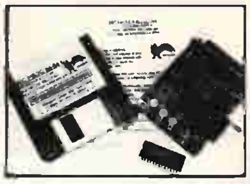

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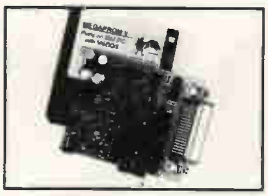

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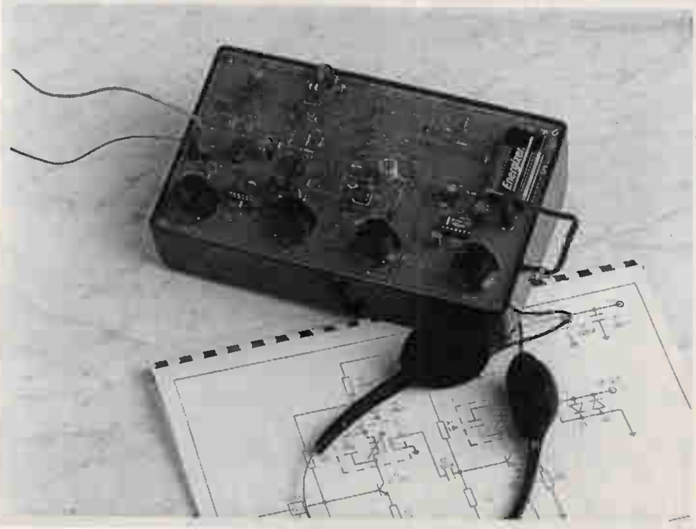

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Seeing through noise II

Patrick May extends his new noise theory to deal with surface diffusion as a 1/*f* noise source in thin metal films and explains how this work leads to a better understanding of low-frequency noise in thin-film resistors and semiconductors.

I show here that the impulses due to the emission of electrons from the surface of a continuous thin film of metal into the electron cloud at the surface produce a diffusion current in the metal. For the first time, I show that this effect in thin films equates to 1/*f* noise.

My mathematical analysis reveals that this gives rise to a time varying component of electron carrier density in a layer near the surface; the square of the Fourier transform of this time function has a 1/*f* characteristic which manifests itself as 1/*f* noise.

Calculations of this noise are made using a modified form of the Richardson-Dushman equation to estimate the rate of electron emission from the surface. These show that the estimated noise in copper thin films is equal to that measured and reported by others in the field. Also, there is good agreement with the strong temperature dependence, observed during experiments.

Current thinking on 1/*f* noise

In semiconductor devices, holes and electrons recombine in the surface under the auspices of the 'fast surface' states. It has been shown that the impulses produced by the ejecting carriers give rise to a diffusion current which produces a time varying component in carrier density in a layer near the surface[1].

It is assumed that a similar phenomenon occurs in metals. But in the case of metals it is the thermionic emission of electrons from the bulk to the 'electron cloud' at the surface which provides the impulses. The analysis for a semiconductor strip has already been presented[1,2]. A similar analysis for a thin metal film is presented here. Next, I present a study of the thermal emission of electrons from the surface of a metal into the 'electron cloud' surrounding the surface.

Electron emission

Figure 1 shows a thin film resistor on a substrate. The rate of electron emission per unit area from the bulk into the 'electron cloud' covering the surface of the metal is derived from the Richardson-Dushman[3] equation,

$$J = AT^2 e^{\frac{-\phi}{kT}} \qquad Am^{-2} \qquad (1)$$

where J is the current density, A a constant for metals, T the absolute temperature, k Boltzman's constant and ϕ the work function of the metal.

The current is due to electrons having an energy greater than the work function. The energy of electrons emitted from the surface into the 'electron cloud' is much lower. According to the image theory, there is no potential barrier at the surface, but the image theory does not hold for short, atomic, distances from the surface. If there is a small barrier potential ψ at the surface of the metal, equation 1 must be modified[3] as,

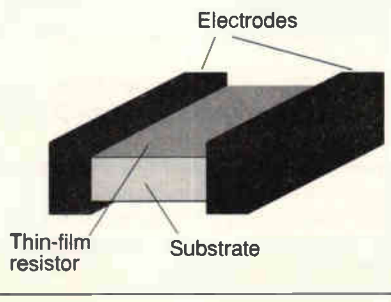
$$J = AT^2\left[e^{\frac{-\varphi}{kT}} + \frac{1}{4}e^{\frac{-2\varphi}{kT}} + \frac{1}{9}e^{\frac{-3\varphi}{kT}}\right]Am^{-2} \qquad (2$$

Electron emission per unit area, v, is given by J/e where e is the electron charge.

Metal thin-film 1/*f* noise – a thermal problem

One of the earliest manifestations of 1/*f* noise was in directly-heated thermionic valves, these had tungsten filaments which also formed the cathode. In the model I propose here, I show that 1/*f* noise in thin metal films is due to thermionic emission from the surface of the metal.

Aspects of the analysis are based on thermionic valve technology dating back to the early part of this century.



A practical application of this work is in precision thin metal film resistor design. The 1/*f* noise in the film can be estimated over a wide range of temperatures using constants that are readily available in physics text books.

The ideas presented here reinforce the surface-diffusion theory of 1/*f* noise in semiconductors.[1,2,14] They can be used to study 'fast surface states' in semiconductor devices, enabling optimisation of low-frequency noise performance.

The constants involved in the calculations are readily available in physics textbooks. There is no credible alternative model that predicts 1/*f* noise in thin metal films and no attempt has been made to explain its strong temperature dependence.[11]

Picture courtesy Alexander S Popov Museum, St Petersburg

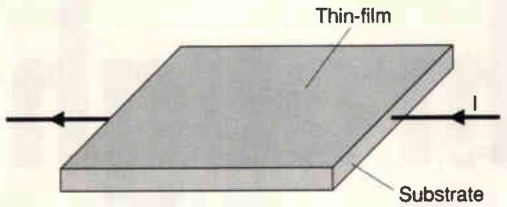


Fig. 1. In a thin-film resistor, electrons being emitted into the electron cloud due to current flow are a cause of low-frequency noise.

Diffusion current

The emission of electrons from the bulk gives rise to a diffusion current in the metal. The analysis appertaining to the diffusion process is based on the continuity equation,

$$\nabla \bullet J = -\frac{\partial \rho}{\partial t} \qquad (3)$$

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

The metal film can be visualised as a strip of metal on a substrate having a thickness less than 1μm, **Fig. 1**. Thermal emission of electrons from the main face produces a diffusion current in one dimension perpendicular to the surface. You can assume that the emission is evenly distributed but uncorrelated.

If current density, J, is assumed to be a diffusion n type current, equation 3 for currents in the x-direction perpendicular to the main faces becomes,

$$\frac{dn}{dt} = D_n \frac{d^2 n}{dx^2} \qquad (4)$$

where $n=n(t,x)$ is electron density and D_n the diffusion constant.

The solution to equation 4 for an impulse of unit strength is,

$$n = \frac{1}{(\pi D_n t)^{\frac{1}{2}}} \exp\left[-\left(\frac{x^2}{4D_n t}\right)\right] \qquad (5)$$

This equation expresses the probability of the absence of an electron at any point in the bulk after the ejection of an electron from the surface. For $t>0$ the boundary condition below must prevail,

$$\int_0^\infty n\,dx = 1$$

The square of the modulus of the Fourier transform of n is,

$$|N(\omega,x)|^2 = \frac{1}{D_n}.\frac{1}{\omega}\exp\left[-\sqrt{\frac{2}{D_n}\omega}\right]x \qquad (6)$$

Note that expression 6 gives the amplitude of the square of the fluctuating component of n^2 – the carrier density squared – in the frequency range ω to $\omega+\omega\delta$. The amplitude of these fluctuations are only significant for the condition,

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

when the expression 6 becomes,

$$|N(\omega,x)|^2 \cong \frac{1}{D_n \omega} \qquad (8)$$

It is evident from equation 7 that there is a cut off frequency for a given thickness of film. The time varying component of n^2 is many orders of magnitude lower than n^2 in the sample. It depends on the spectral expression given in equation 8, the rate of emission of electrons per unit area expressed in 2 and the thickness of the sample, L. It is given by,

$$S_n(\omega) = \frac{J}{eLD_n\omega} \qquad (9)$$

This can be expressed in hertz as,

$$S_n(f) = \frac{J}{eLD_n f} \qquad (10)$$

If $S_v(f)$ is the power spectrum when the mean square of the voltage across the sample is, $\overline{V}^2$.

$$\frac{S_v(f)}{\overline{V}^2} = \frac{S_n(f)}{n^2} = \frac{1}{n^2}.\frac{J}{eLDf} \qquad (11)$$

Estimating noise

The expression for $1/f$ noise in equation 11 shows a strong temperature dependence. The quotient J/D gives its temperature dependence. By making ψ in the expression for J, in equation 2, equal to 0.01eV, the temperature dependence of,

$$\frac{S_v(f)}{\overline{V}^2}$$

expressed in equation 11 follows that obtained experimentally for copper thin films[4]. The experimental value of,

$$\frac{S_v(f)}{\overline{V}^2}$$

has been measured[5] for a sample of thickness $L=10^{-7}$m at 330K as $6.4\times10^{-16}Hz^{-1}$ at 10Hz. This was repeated independently and confirmed[4].

The evaluation of,

$$\frac{S_v(f)}{\overline{V}^2}$$

from equation 11 for copper is obtained from the following physical constants: n is the effective number of free electrons, $m^{-3}=2.5\times10^{27}m^{-3}$ (from reference 9), $e=1.6\times10^{-19}$C,

$$J = 0.65\times10^6 T^2\left[e^{\frac{-\varphi}{kT}} + \frac{1}{4}e^{\frac{-2\varphi}{kT}} + \frac{1}{9}e^{\frac{-3\varphi}{kT}}\right]Am^{-2}$$

and $D=0.8\times10^{-4}m^2s^{-1}$. For T=330K, $\psi=0.01eV$ and $L=10^{-7}m$, the calculated value is $7.2\times10^{-16}Hz^{-1}$ at 10Hz. The result is very close to the value obtained experimentally by Voss.[5] This can also be shown to be true for silver films.

In summary

The basis of the diffusion current analysis lies in the equation for an impulse of unit strength, equation 5. This equation only applies in practice when a very large number of electrons is ejected – or injected – *evenly* from a plane surface. This insures a one-dimensional diffusion current in the bulk of the thin film perpendicular to the surface.

The absence of an electron per unit area in the bulk as a result of the ejection, or injection, of an electron from the surface is described by a density probability function. This function applies to every single ejected or injected electron. This means that in the transient case, when all the N electrons per unit area are ejected, the solution is given by the product of N and the expression in equation 5.

The result agrees with that of the Haynes-Shockley experiment[6]. Note that equation 5 holds when the electrons are ejected or injected *evenly* from a plane surface *irrespective* of whether they are ejected simultaneously as a transient or whether they are ejected continuously and randomly in time as a steady state process. This is the case in the diffusion-current analysis, and leads to the application of Carson's theorem[7].

The expression in equation 5 fulfils the necessary condition of absolute integrebility[7]. This results in equation 6 leading to the conclusion in equation 11.

From the proposed model it is clear that the origin of $1/f$ noise – like Johnson noise – is the kinetic energy of the electron carriers. Their kinetic energy is responsible for both the thermal emission at the surface and the consequent diffusion current in the bulk which is accompanied by a fluctuation in electron concentration.

Under equilibrium conditions, the diffusion current is opposed by an equal drift current from the 'electron cloud' resulting in a zero average current. The average carrier concentration is that normally quoted for the metal but it fluctuates about this value.

The fluctuations are not affected by the bias drift current because it makes a negligible contribution to the electron thermal velocities responsible for the diffusion process. The fluctuations involve the whole of the carrier population under equilibrium conditions. Hence the argument by Weissman[8], that the 'tied individual electrons must remain in the sample' for periods well in excess of the actual time they spend in transit does not apply to my model.

On the other hand, Weissman's argument does invalidate some theories on *1/f* noise; it shows that the *N* term – total carrier population – in the empirical Hooge formula[9] is unacceptable. Accordingly, equation 11 does not contain *N*. But it shows that *1/f* noise is inversely proportional to the thickness of the film as has been demonstrated experimentally[8,10].

Hooge's formula[9] does not predict temperature dependence. Nor does it allow for the wide scatter in experimental results.[11-13] My model predicts the noise level and is the only proposed model to predict temperature dependence.

The wide scatter in experimental results is also accounted for; thermionic emission is very sensitive to surface treatment. There is, however, a component of *1/f* noise in continuous thin films which is not temperature dependent[13] and may be associated with the substrate.

My thanks to University of Exeter's School of Engineering for providing library facilities and to Drs MR Belmont, J Baker and MK Horwood for their encouragement and advice. ■

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SPEAKERS' CORNER

John Watkinson looks at the problems involved in specifying a loudspeaker.

There is no point in designing a loudspeaker without a clear idea of the technical specification it should meet. Without a technical specification, objective testing is impossible and development proceeds in the dark.

As the loudspeaker is only for human indulgence, the technical specification must be derived from psychoacoustic considerations. A good general specification for a loudspeaker is that over the audible frequency range it should reproduce the original electrical waveform as an acoustic waveform. It should do this both on-axis and a within reasonable angle off axis at a realistic sound pressure level.

Microphones, recorders and amplifiers have to do this and so it seems a reasonable goal for a loudspeaker. This goal implies a flat frequency response which is phase linear and a transfer function which is also linear. True phase linearity is virtually impossible to achieve in a transducer.

Fortunately this also includes the human ear. Consequently an acceptable goal is a minimum phase characteristic. Rapid phase changes, particularly within a critical band, should be avoided which requires a greater degree of discipline in crossover design.

The audible frequency range requires some defining. At the top end, we usually pay lip service to a 20kHz bandwidth in recorders and circuitry even though many of us cannot hear such a frequency. It does no harm to specify such a bandwidth in the hope that a good transient response will result, but I am not convinced that a ruler-flat response to that frequency is necessary. A mild but monotonic roll-off is quite acceptable provided it is truly monotonic.

Interestingly enough the same argument applies at low frequencies. The lowest frequency to be reproduced is debatable and depends upon the material to be reproduced. If we want to be able to reproduce all musical instruments, response has to be maintained to around 20Hz. Low frequency roll-off is unavoidable but it must be monotonic and preferably have a slope of no more than 12 dB/octave.

Tuned speakers having ports or auxiliary bass radiators use resonance to extend the low-frequency response, but the result is that the ultimate roll-off is much steeper, leading to a now-you-hear-it, now-you-don't effect. In any case such tuning techniques are undesirable because they introduce linear distortion, audible as hangover.

While porting allows more sound-pressure level, or lower cost, accurate low-frequency requires phase linearity and only an unported unit can achieve that.

The ported speaker is so common that many people think of it as the norm. The trouble is that it doesn't bear comparison with the original. A clear example is the offset – i.e. the opposite of onset – transient which occurs when the air supply to an organ pipe is cut off. On a linear-phase woofer this is audible whereas on a tuned woofer it is masked by hangover.

Music editors prefer phase linear woofers because if the low and high-frequency components of a transient do not arrive time aligned there is some ambiguity over where the edit point should be.

An unported woofer still has a fundamental resonance due to the moving mass and the compliance it sees, but in a correctly designed unit this resonance is damped by the negligible output impedance of the amplifier. The result is that the damped speaker acts as a high-pass filter, rolling off monotonically at 12dB/octave.

When considering the power of a speaker, quoting the input power is a waste of time. What matters is what comes out. If realism is the goal, the sound-pressure level must be the same as the original.

Musical instruments – especially the piano – and the human voice change timbre as they get louder. With a good recording there is only one level at which it sounds right.

The threshold of hearing is irregular and rises at low frequencies. While active techniques can extend low frequency response almost arbitrarily, there is no point in doing this if sufficient sound-pressure level is not available as it simply won't be heard.

Non-linear distortion is the generation of harmonics due to the transfer function not being straight. This is critical in stereo because the creation of multiple images assumes linear superposition of the pairs of signals belonging to each image. Non-linearity causes intermodulation which results in phantom sound objects in the stereo image.

At high frequencies cabinet diffraction must be carefully controlled otherwise it causes multiple re-radiation which puts ripples in the on-axis response and makes the polar diagram extremely uneven. This causes colouration in the reverberant field which contributes to listening fatigue.

While these are strict requirements, they should not discourage. The market for mediocre loudspeakers is well supplied by traditional products. The only hope for a newcomer is to create a new market by attempting the impossible. Sometimes, however, the barriers are self-imposed rather than technological and the impossible becomes a product. ■

Speaker criteria

- Frequency response accurate enough to avoid timbral change.
- Linear distortion or phase linearity accurate enough to reproduce transients.
- Non-linear distortion low enough to eliminate false sound objects in stereo.
- No resonant or tuned behaviour to prevent hangover.
- Realistic sound-pressure level over whole frequency range.
- Enclosure diffraction controlled to give clean polar diagram.
- Wide dispersion so that reverb has same frequency response as direct sound.
- Fatigue free to allow extended listening.

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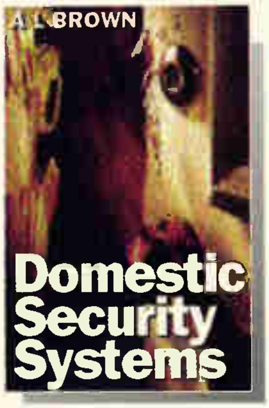

House break-ins have increased threefold in the UK over the last 20 years. Few have not been touched by the affects, even if only though the experience of family and friends who have suffered a burglary. There is a way to reduce significantly the chances of being targeted by thieves: fit an alarm. But isn't that expensive and complicated? Not if you build your own system. This book shows you how, with common sense and basic DIY skills, you can protect your home.

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Specification, installation and maintenance. Gerard Honey

This book covers Intruder Alarm Systems (C+G 1851 syllabus) as well as providing the underlying knowledge required to achieve a level 2 NVQ (National Vocational Qualification) in Understanding, Specifying, Installing and Maintaining Intruder Alarms (C+G 1863, 1864 and 1865). Familiarity with the contents of this book are required before an award will be made.

Gerard Honey is an experienced installer and writer and has used that experience to produce a book that not only provides essential information in a way that is easy to follow and learn, but also makes the book a fine practical source of advice. Each chapter contains summaries, self-tests and other features designed to help the student to understand and gain knowledge easily. Intruder Alarms has been published with the help of SITO, the Security Industry Training Organisation, who design courses and organise training for security installers and professionals.

*Only course book written for syllabus for Security NVQ
*Comprehensive study of intruder alarms
*Author is a practising international security systems expert

CONTENTS: Intruder alarm systems; Circuitry; Detection devices; Power supplies; Control equipment; Signalling systems; Wiring systems; Inspection of the mains supply; Commissioning, maintenance and fault finding; Index

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Microstrip made easy

Nick Wheeler explains that although there are complex procedures for designing high performance microstripline circuits, a few shortcuts result in a technique that is quick and easy to implement yet more than adequate for most applications.

Micro stripline consists of conductive traces of defined width on low-loss dielectric superimposed on a conducting ground plane. It can be created by etching one side of ordinary double-sided printed circuit board material.

Stripline on the other hand consists of flat, thin conductors sandwiched between two ground planes and usually embedded in dielectric. It has many useful properties such as inherently good screening, but it is difficult to use and is not further discussed here.

For micro-stripline, glass-reinforced epoxy is the preferred dielectric. The cheaper phenolic type is too lossy, and PTFE or ceramic substrates are both much more costly – PTFE boards cost in the region of £100 a square foot – and are difficult to work with. On the other hand, PTFE boards are usable far into the microwave region. In this article I refer only to G-10 and FR-4 substrates. Apart from being fire-resistant, FR-4 is very similar to G-10.

This article is directed towards those of you wanting to make high-performance equipment with relatively limited resources. I have interpreted this as meaning that trace widths are defined to an accuracy of no better than 0.01in. This means, roughly, that characteristic impedance Z_0 of a typical trace will be accurate to one or at worst a few percent.

A down-to-earth attitude is adopted in the ARRL Handbook.[1] What I have done below is to explain how their results can be replicated, while referring to more rigorous treatments of the subject to reassure readers that readily available techniques are virtually indistinguishable from the very best that can be done.

Microstripline principles

Most microstripline calculations refer to traces of width *W* on an infinite layer of dielectric whose dielectric constant is ε_r and of thickness *h* backed by an infinite ground plane, **Fig. 1**.

In practice – and more so with increasing frequency – the electric field is concentrated in the volume of dielectric lying directly under the trace. However, some of the field is in air, leading to the concept of ε_{eff}, the effective dielectric constant, which is lower than that of the dielectric.

The greater the width of the trace the more ε_{eff} tends towards ε_r. If the ground plane and dielectric are truncated, as in **Fig. 2**, there is little effect upon Z_0 if $T>2W$. If only the dielectric is truncated[2], as in **Fig. 3**, Z_0 is raised by less than 0.5% when T/W is greater than 0.5.

Because W/h is dimensionless, if the dielectric thickness is halved, the same Z_0 is achieved with half the trace width. There are many published formulae for accurately calculating Z_0. These take into account the thickness *t* of the trace and even whether the etching process has resulted in a trace of

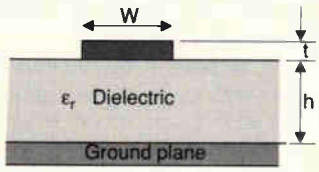

Fig. 1. Most microstrip calculations refer to traces of width W on an infinite layer of dielectric with constant ε_r, and of thickness h backed by an infinite backplane.

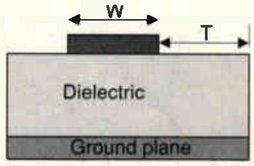

Fig. 2. In this microstripline example, both ground plane and dielectric are truncated.

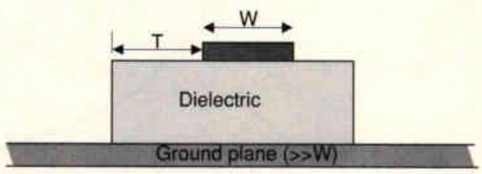

Fig. 3. If only the dielectric is truncated, Z_0 is raised by less than 0.5% for when T/W is more than 0.5.

trapezoidal rather than rectangular cross section.

For traces produced by etching ordinary 1oz circuit board material, with a thickness t of 0.0014in, and for typically useful values of Z_0, these considerations are very much second-order effects. The effect of varying t from 0.0014in to 0.0056in on ordinary G-10 board of 0.0625in thickness is to reduce the trace width for Z_0=50Ω from 0.11984in to 0.11691in. Incidentally, the trace width for 50Ω on this board is commonly specified as 0.1in but 0.12in is clearly a better approximation. This can be achieved by carrying out the following.

Evaluating complex formulae is tedious and there are many published tables relating W/h, ε_r and Z_0. The graph of **Fig. 4**, derived from several sources, gives trace widths for a useful range of Z_0 using commonly available double-sided glass-epoxy pcb material It should be accurate enough for most purposes. This is particularly so since G-10 and FR-4 are supplied by various manufacturers with ε_r varying over a range from 4.22 to 4.9. The most commonly quoted range though is 4.3-4.5.

Applying microstripline

There are in print several well-documented project descriptions.[1] These give a good overview of what can be done using this technique. Quarter wavelength, or $\lambda/4$, transformers for frequencies around 1GHz are particularly easily implemented, but bear in mind that the electrical length of microstripline is the reciprocal of $\sqrt{\varepsilon_{eff}}$ multiplied by the physical length. The term λ_g is commonly used to describe the on-board wavelength.

Perhaps one of the most useful applications is in conjunction with monolithic microwave ICs, or mmics, as these are 50Ω devices in most cases. On the other hand, many other devices can be matched.

It is important to bear in mind that the cross-section of the trace of a typical microstripline is very small. If it is operated at significant power levels in a mismatched condition, destructive current antinodes may lead to failure.

Some general-purpose microstripline applications are dealt with below.

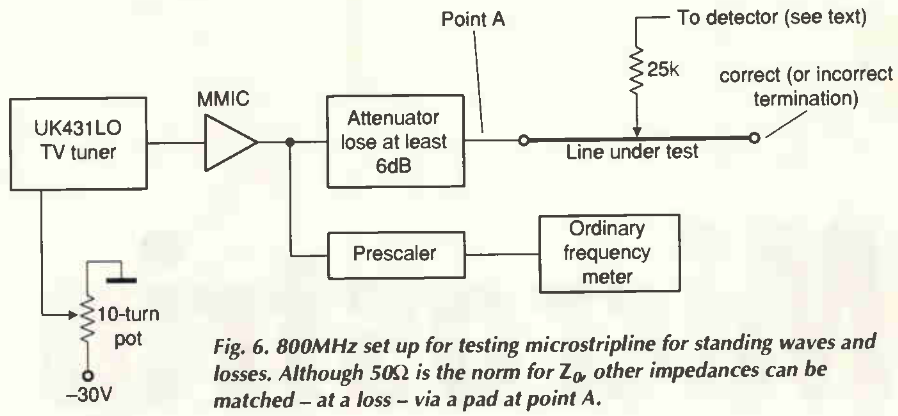

Fig. 6. 800MHz set up for testing microstripline for standing waves and losses. Although 50Ω is the norm for Z_0, other impedances can be matched – at a loss – via a pad at point A.

Applying microstrip

As reference 1 suggests, microstripline can be implemented by attaching, with superglue, 0.1in strips, or other widths, of thin copper foil to the plain side of single-sided pcb material. This works quite well, but I will now describe a reproducible photo-etch approach.

I use Windows 95 with one of the many available graphic design programs, in my case *Serif Draw Plus*, an *HP Deskjet 693C*, and the appropriate Premium Transparency Film. This combination I can vouch for, but there are doubtless other suitable films and many other software packages and printers that will work too. Do not attempt to use films designed to take spirit-based felt pens or films made for overhead projector transparency work.

The transparencies produced are fine as masters for conventional photo-etching. It is important to remember that printer ink takes a long time to dry on transparency film. I recommend leaving for at least half an hour. I will not describe this well-known technique, except to point out that the transparency should be flipped so that the pattern ends up in face down, in direct contact with the pcb material before exposure.

Serif Draw Plus V2.0 has the option of selecting a grid based on 0.1in, with a spatial increment of 0.01in. You will suffer serious problems if you try to create circuit boards on a metric grid of any coarser resolution than this. I've tried it.

Graphic programs define line widths in 'points.' As far as *Serif Draw* is concerned, 7 points closely approximate to 0.1in*. This makes 8 points a good line width for 50Ω on 0.0625 G-10 board.

Most transmission-line configurations can readily be implemented in microstripline. References 1-3 give many examples. The attraction here is that any Z_0, within a range generally quoted as being from 16Ω to 125Ω can be achieved without resorting to tedious methods such as replacing the inner conductor of co-axial cable with wire of a different diameter.

A few practical points. Parallel traces, separated by a distance of W or less couple and can be used to effect directional couplers. Undesired coupling can be virtually eliminated by making the spacing >2W. The design parameters for couplers are highly interactive, and the empirical approach – i.e. suck it and see – may be the least difficult.

Bends can be radiused or mitred. Bends through any angle with a radius, to the centre line of the trace, >4W do not have a significant effect on standing-wave ratio. Mitred bends, which occupy less board space, are theoretically complicated. However, using the approach of **Fig. 5** seems to yield generally satisfactory results.

The advantage of using the computer-graph-

Fig. 5. Not an ideal microstripline bend shape, but it does the job for most applications.

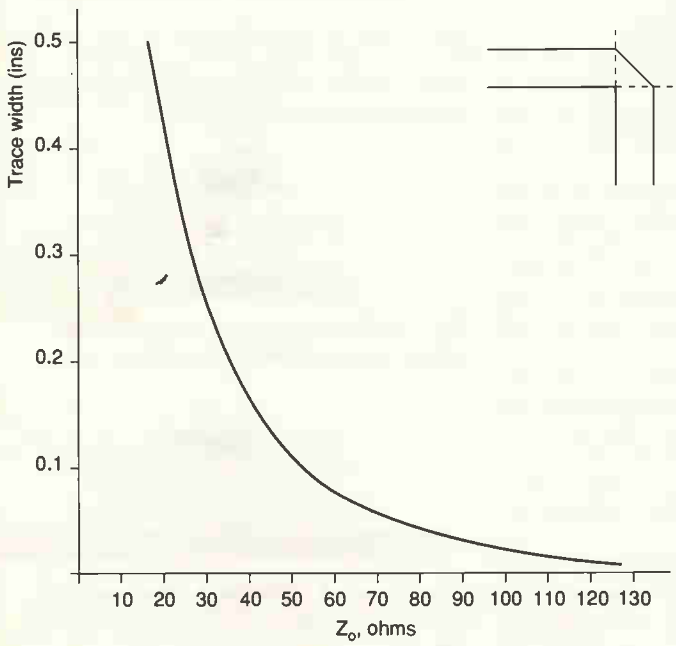

Fig. 4. Derived from several sources, this graph gives trace widths for a useful range of Z_0, assuming common glass-epoxy pcb.

*A point is 1/72 of an inch – Ed.

ic approach is that your artwork is stored and can easily be altered before committing to transparency. Good results – within the limitations of the transfer sheets available – can be obtained much more directly by the use of etch-resistant pcb transfers. Note that not all transfers are etch-resistant, though they can save a lot of time in producing artwork for photo-etching.

Making measurements

In many cases, the success of a microstripline circuit design can be estimated by driving it from a source of the right impedance, terminating it correctly and looking for standing waves or excessive losses.

I made a line 210mm long on G-10 board. One source quotes ε_{eff} for a 50Ω line as about 3.4, when ε_r is 4.4. Using a test frequency of 800MHz, this accommodates just over a wavelength. An ε_r of 4.5 is favoured in reference 1.

Figure 6 shows a test setup. The signal source is a *UK431LO* television tuner, which has a local-oscillator output ranging from 431 to 900MHz. Many other tuners have local-oscillator outputs. This signal is buffered by a suitable mmic. Almost any of those currently available will suffice.

Frequency is measured using the gigahertz prescaler of reference 4. The detector probe is connected to the 75Ω input of the low-cost spectrum analyser described in reference 5.

A 25kΩ series resistor at the probe tip ensures negligible loading effects when applied to the line under test. Measurements are made by the null method of varying the attenuator to produce the same outputs at the points being tested.

As the spectrum analyser uses a television tuner there is obviously plenty of scope for the use of other tuners. Although I have not tried it, the raw intermediate frequency from any tuner could be amplified and inspected on a modest oscilloscope.

In the case of my test line, probing it when properly terminated with 50Ω disclosed no perceptible standing wave pattern. Operating with no termination produced deep voltage nodes separated by 100mm – almost exactly – on the board. These are, of course, $\lambda_g/2$ apart.

Working backwards through the relationships outlined above, you can deduce that in this case the apparent ε_{eff} is 3.5. This degree of agreement lies well within the limits which might be expected.

An error of 1% in the measurement of the distance between the nodes could account for more than half the difference. Because of the element of empiricism inherent in all microstripline calculations this seems to be a very good first attempt, and easily good enough for almost all applications.

In summary

I have shown that microstripline is a versatile technique, easily implemented to useful accuracy for many uhf applications. I recommend a computer-graphic approach for the generation of the artwork, since this gives quite precise control over trace widths.

The rigorous treatments of references 2 and 3 are available to those of you wanting to use more precise techniques. This will not normally be necessary unless high power levels are to be involved. ■

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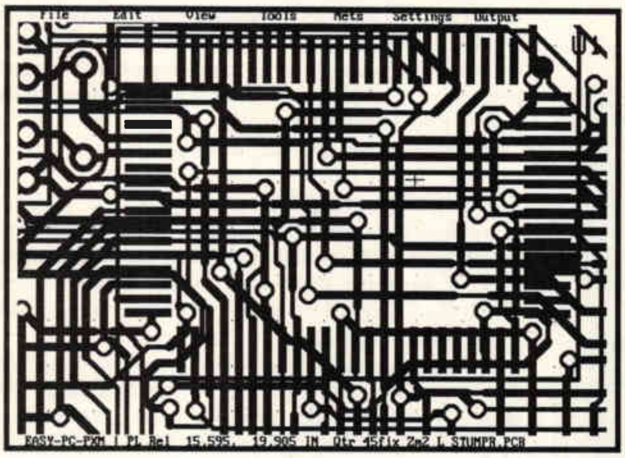
1. ARRL Handbook.
2. Waddell, BC, Transmission Line Design Handbook, Artech House 1991.
3. Edwards, TC, 'Foundations for Microstrip Circuit Design', Wiley 1981
4. Wheeler, NPE, Gigahertz prescaler. *Electronics World*, Sep. 1996
5. Wheeler, NPE, 'Spectrum Analyser on the Cheap,' *Electronics World*, Mar 1992.



Radio reflections from Russia

St Petersburg's Alexander Popov Museum of Communications – one of the world's oldest technical museums – houses a wealth of early British wireless apparatus – as Khatskel Ioffe reveals.

Founded in 1872 as a telegraph museum, the Alexander S. Popov Central Museum of Communications in St. Petersburg has a collection of British equipment dating from the earliest stage of the development of wireless. Building on last month's article, which described artefacts ranging from Flemming's diode to a Marconi direction finder, this article cover our Marconi 250-1300m receiver, 500W transmitter and a wavemeter – Fleming's cymometer.

A standard for 250 to 1300m

A receiver for 250 to 1300m from our collection is shown in **Photo 1**. An identical receiver is depicted in a magazine printed in 1915 , where it makes part of a field wireless outfit.[1] At around the same time, a receiver for this range also made part of the Marconi Company's 0.5, 1.5 and 5kW ship and coastal wireless stations.

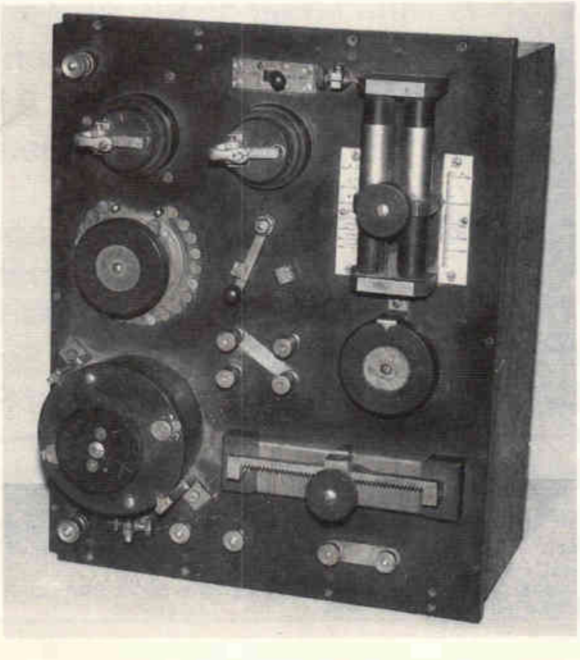


Photo 1. This Marconi Company receiver from circa 1915 was designed for reception between 250 to 1300m. It represented a standard in coastal and marine communications.

We consider this receiver to be Marconi's standard apparatus for 250 to 1300m during WWI. The receiver is complete with two carborundum contact detectors, one of which was a spare.

The receiver also has terminals for connecting a magnetic or another external detector. This allowed reception of strongly damped oscillations in addition to the undamped or feebly damped oscillations then coming into use.

The receiver has two subranges, 250–800m, and 500–1300m, selected by placing an aerial tuning capacitor in parallel or in series with an aerial tuning inductance coil, respectively. Placed into the aerial circuit, in addition to the variable inductance coil, is a moving detector-coupling coil.

Along with the Billi condenser, the detector and the telephone, this coil makes up the detector circuit. Connected to the receiver is an external storage battery which secures a required mode of operation of the receiver using a potentiometer. Provided along the periphery of the top panel are holes for screws fastening the receiver at its location on the frame of the transmitter-receiver station.

Our museum has two such receivers.

500W field transmitter

A wireless transmitter with a rotary spark discharger for a 0.5kW field wireless station mounted on a cart is shown in **Photo 2**. In the late 1900s and early 1910s wireless telegraph with a shock-excited transmitter generating a musical spark were becoming popular.

The Marconi Company obtained this effect by means of a rotary, toothed, spark discharger wherein a spark arose as each tooth passed under fixed high-voltage electrodes. The use of shock excitation meant a transition to feebly damped oscillations, which increased the energy sent by the station and enhanced its efficiency.

Installed on a two-wheel cart, the Marconi Company's 0.5kW wireless station was used in the Russian Army during WWI. In our set, a 7hp air-cooled two-cylinder Douglas petrol motor and an electrical generator are installed on a cast-iron base. The shafts of the motor and generator are connected by a bilateral coupling. A toothed spark discharger is fixed on the shaft of the electrical generator.

The spark discharger is separated from the electrical generator by a chamber made of an aluminium alloy. The spark discharger has a copper disc, 7mm thick, fixed on a hard-rubber hub, through which a shaft supported in ball bearings passes.

Mounted in the chamber case are two fixed electrodes constituted by two copper rods, 4mm in diameter, placed inside hard-rubber insulators, the disc tooth-to-electrode spacing being set at 0.15-0.2mm.

The electrical generator has a stator with 12 poles arranged as 6 pairs. In the magnetic field formed by these poles, an anchor rotated with dc and ac windings. The current of the dc winding was fed into the collector and used in the excitation circuit while the current of the ac winding was fed into the rings. The ac voltage, transformed via an external high-voltage transformer, was fed to the electrodes of the spark discharger.

The spark discharger is of the synchronous type. It has the same rotational speed as the electrical generator and discharges at each peak of the ac voltage, ie at each alternation. Running at 1900rev/min, the electrical generator produced a 190Hz alternating current and the spark discharger produced 380 discharges per second. This corresponds to a relatively good tone – the musical spark effect.

On air in the air

The transmitter of an aeroplane wireless is shown in **Photo 3**. This is one of the first transmitters designed for the purpose.

It dates back to the times of WWI. It is a low-power spark transmitter consisting of an induction coil with a mechanical interrupter, an adjustable spark discharger in the form of two cylindrical aluminium electrodes, and an oscillatory circuit. The latter is comprised of a fixed capacitor and a variometer whose fixed coil is placed in the aerial circuit.

Using the variometer, the 230–430m wavelength range is covered. The induction coil was operated from an external dc power supply. A telegraph key was hooked into the power supply circuit. At the front of the transmitter case, access is provided to the spark discharger. Also provided is an inspection window to inspect the spark discharger.

Fleming's cymometer

The earliest type of the Marconi Company's wavemeter was Fleming's cymometer. I believe that the instrument that we have dates from 1906–1907.

This cymometer opened up the use of scientific methods for tuning transmitters and receivers. Earlier methods involved trial and error. It comprises a variable inductor and a variable capacitor. In the inductor, a coil of bare copper wire is wound around an ebonite tube. The self-inductance is variable by means of a slide contact.

The variable capacitor – a Billi condenser – consists of two thin-wall brass tubes separated by a hard-rubber bushing. The capacitance is varied by moving the outside tube using a handle. As the tube moves, the slide contact of the coil also moves so that the values of both elements of the measurement circuit are varied at the same time and in the same direction.

In order to find the instant that the measurement circuit is at resonance with the transmitter frequency, a neon tube was connected in parallel to the capacitor.

The scales of the instrument are calibrated in the values of an 'oscillation constant' – the term used by Fleming to denote the expression $\sqrt{LC}$. There are also scales for wavelength in metres, wavelength in feet, and the numbers of oscillations per 1/1,000,000th of a second.

It is known that the Marconi Company produced four types of cymometers to cover the ranges 33–700m, up to 1400m, up to 2000m and up to 3000m.

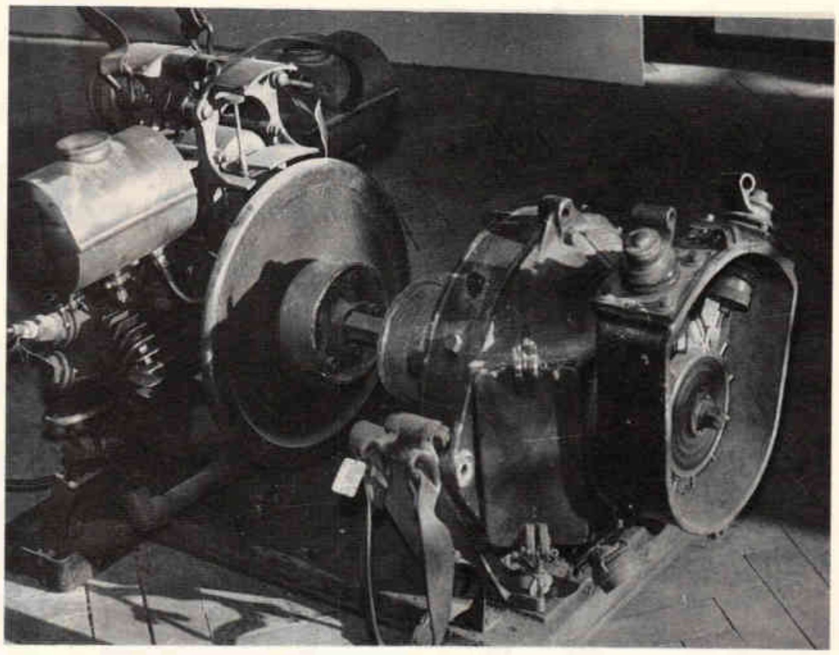


Photo 2. Transmitter apparatus with rotary discharger for a 0.5kW field wireless station from circa 1916.

Our cymometer is for 33–700m. To determine the wavelength of a transmitter, the cymometer was installed parallel to an antenna section. The operator moved the handle of the instrument in either direction so as to attain the brightest glow of the neon tube. The scale indicator, rigidly bound with the handle lever, ensured direct readings of the wavelength and frequency from scales.

To measure capacitance and inductance, the operator connected these components to a standard self-inductance or standard capacitance, respectively. The components excited oscillation via an ancillary inductor with a spark gap. The cymometer was used to determine $\sqrt{LC}$, whereafter the unknown value of C or L could be calculated.

Rounding up

All of the above-described apparatus were received by our museum in the 1920s from Russia's various communications establishments and educational institutions.

The British equipment presented in this article along with the apparatus of the Popov-Ducretet, Slaby-Arco and Telefunken systems of the early 20th century, also in the keeping of our museum, preserves the memory of the first steps in the development of wireless worldwide.

I cannot guarantee the accuracy of the dating and operating descriptions of the instruments described since my collection of literature is limited. Any corrections will be gratefully received. ■

Translated from the Russian by L.N. Kryzhanovsky. Khatskel Ioffe is with the A.S. Popov Central Museum of Communications, St. Petersburg.

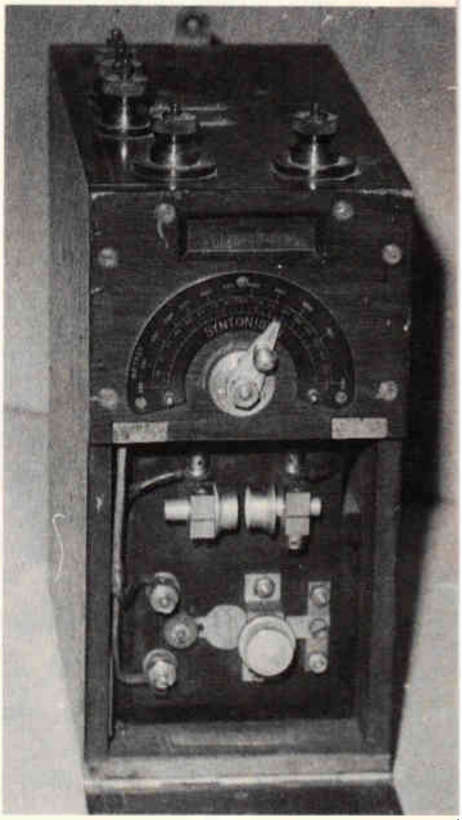


Photo 3. Transmitter for aeroplane wireless station from circa 1915 – one of the first transmitters designed specifically for aviation.

Reference

1. 'Wireless Telegraphy in the War', *The Wireless World*, August 1915, p. 299.

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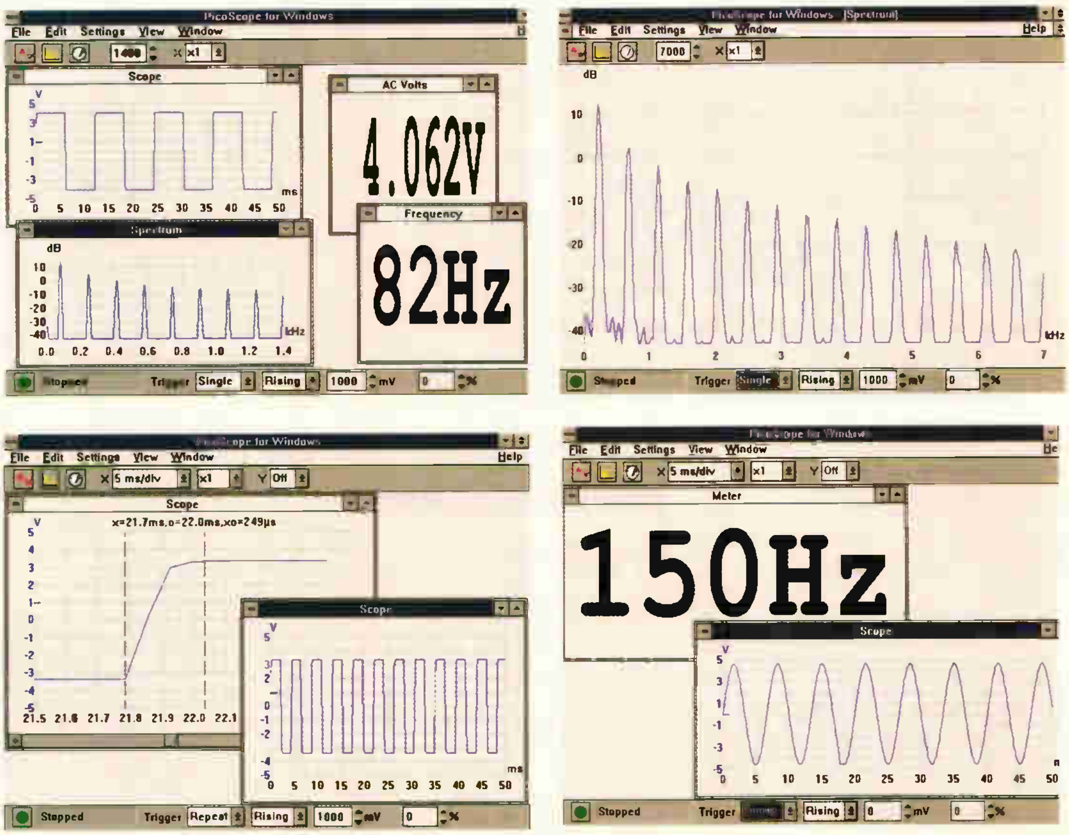

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Trickier keypad lock

While small keypad locks are simple and quick to make, their four-digit code offers limited security; it would be possible, if tedious, to plod through them all. This circuit puts a further barrier in the way in that two buttons must be pressed at the same time, and the hopeful tealeaf does not even know to do that, let alone which two. In the circuit shown, the entry code is 4, 1, 7, 92; the pair of buttons are best well separated to avoid accidental operation.

Keys on the pad I used had one side common. Cutting one track and wiring the switch in series gave the circuit shown. An *LS7220* does the job, but others would perform in the same way.

Neville Frewin
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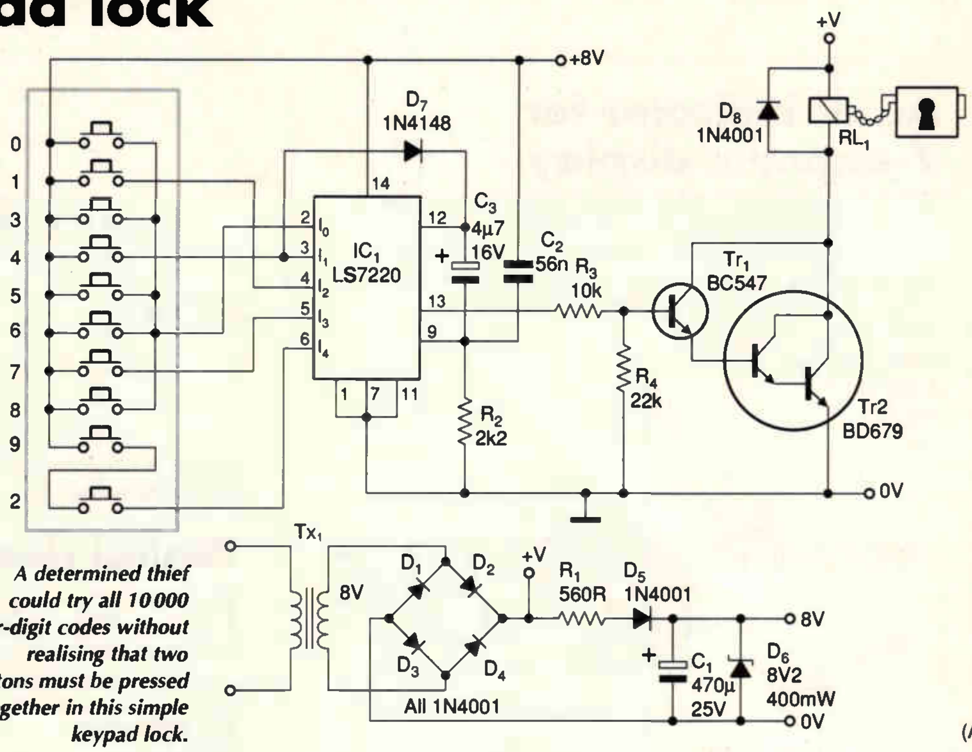
(A31)



A determined thief could try all 10 000 four-digit codes without realising that two buttons must be pressed together in this simple keypad lock.

(A31)

5MHz triangle generator

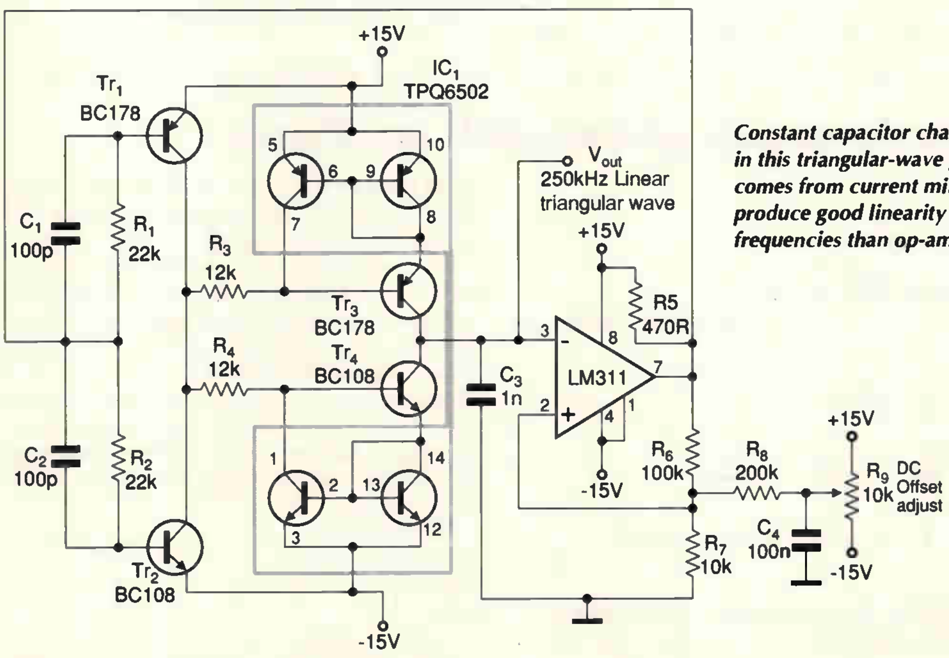
Generating linear, high-frequency triangular waves has its problems. Although there are dedicated ics, some are fairly expensive and others put out a linear waveform, but at the expense of bandwidth. This circuit avoids these problems.

Instead of the usual op-amp *RC* integrator, the constant charging current is provided by a pair of current mirrors contained in a *TPQ6502* array to give a good match. In response to the output waveform, an *LM311* comparator switches the current mirrors on and off via $Tr_{1,2}$, slopes being independently determined by the values of $R_{3,4}$, which may be replaced by variable resistors. Frequency is set by R_7 which may also be made variable, and R_9 adjusts the dc level of the output.

Board layout demands care: supply lines to $Tr_{1,2}$ and those to the comparator must all be kept separate from the supply to the mirrors. A single ground and good decoupling are essential for a good performance at high frequencies.

Higher frequencies than 5MHz are possible, while retaining good linearity, if the *LM311* is replaced by a faster type.

K P Cummings
University of Nottingham
(A30)



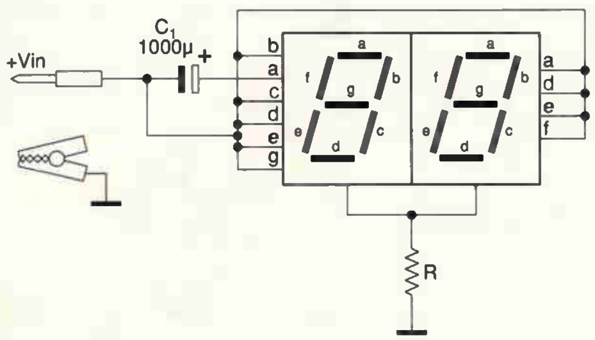
Constant capacitor charging current in this triangular-wave generator comes from current mirrors, which produce good linearity at higher frequencies than op-amp integrators.

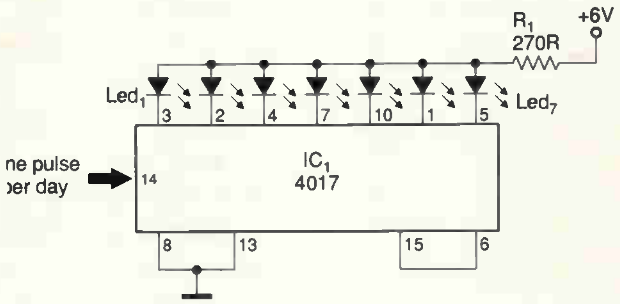
ADC-42 WINNER

Ac/dc indicator for 7-segment display

This is possibly the simplest idea yet seen in this column and is ingenious. It indicates whether the input is ac or dc by virtue of the fact that, if it is dc, the “a” element does not light and the display reads lower-case “dc”. If ac, the element lights and the display shows “ac”. Resistor R should be V_{in}–1.7/0.01Ω.

Raj K Gorkhali
Kathmandu
Nepal
(A38a)





Counter and leds indicate whether it is Sunday or Monday and you have to get up.

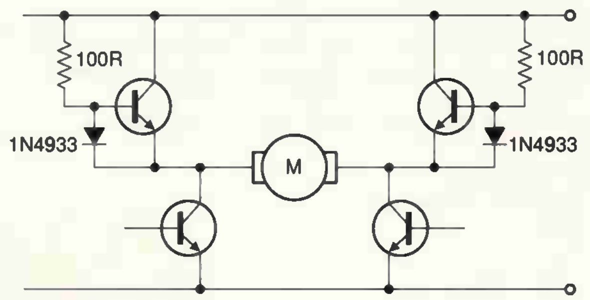
Digital clock day indicator

For those who, on waking, are often unsure not only of where and possibly who they are, but also what day it is, this little addition to a digital clock will reassure them. If the clock has an am/pm signal, applying it to the counter steps it on once per day, the relevant led being illuminated.

Raj K Gorkhali
Kathmandu
Nepal
(A38b)

Small motor drive

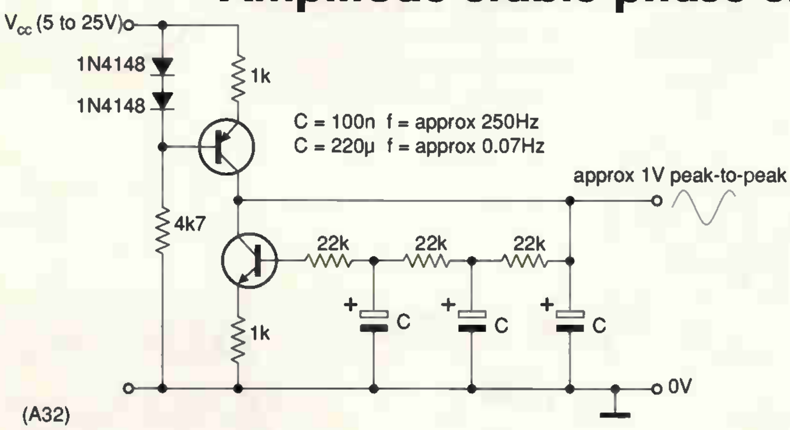
Switched motor drive for 12V dc motors, driven by a pwm signal.



This drive for small dc motors is a variation on those described by Peter Hale in his article ‘Driving and controlling small motors’ in *EW*, May 1997, p.397. It uses a single supply rail and consists of two drive transistors, the two upper devices acting as voltage-dependent resistors. At low voltage, resistance is low, the reverse applying, so that the bridge halves behave as switches.

Ray Stead
Hampton,
Middlesex
(A35)

Amplitude-stable phase-shift oscillator



(A32)

Variation on the classic phase-shift oscillator, having a constant-current collector load to avoid the use of thermistors or an agc circuit.

This low-frequency oscillator runs from an unregulated supply and is meant for use where precise frequency setting is not a requirement.

It is based on the classic design of *RC* sinewave oscillator, each *RC* contributing 60° towards the necessary 180° phase shift from collector to base. Unlike the standard design, amplitude stability comes from the p-n-p current source instead of a thermistor or an agc circuit.

Capacitors are grounded to allow the use of electrolytics. Components are not critical, except that the n-p-n transistor should have an h_{FE} of at least 180. Amplitude and frequency remain within about 10% with double the supply voltage.

L S Whitlock
Taunton
Somerset
(A32)

Analogue frequency doubler

Two comparators may be connected to either a difference amplifier or to a summer to produce a frequency doubler or a variable-width output pulse.

Basic comparator operation in **Fig. 1** shows the effect of reversing inputs, the arrangement in **Fig. 2** producing, in response to a sawtooth input waveform, positive pulses at pins 1 and 7, the width of which depend on the rectangular $V_{ref1,\,ref2}$ waveforms and V_{in}. Subtracting one from the other in the *LM301* difference amplifier gives the output shown in **Fig. 3**.

Alternatively, connecting the comparator outputs to a summer combines the comparator outputs to give a pulse proportional to the input and reference voltages, variable from zero to 100%. Output spectrum is variable in this way, so that given harmonics may be eliminated or created by varying the reference voltages.

Kamil Kraus
Rokycany
Czech Republic
(A40a)

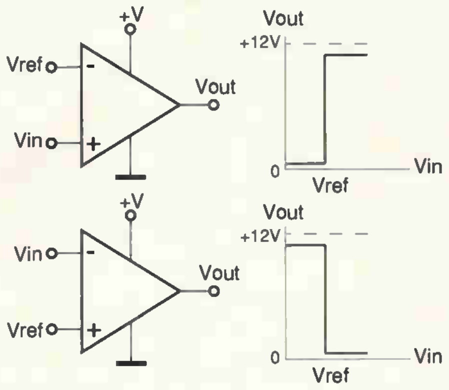


Fig. 1. Comparator output with rectangular wave applied to either of the inputs.

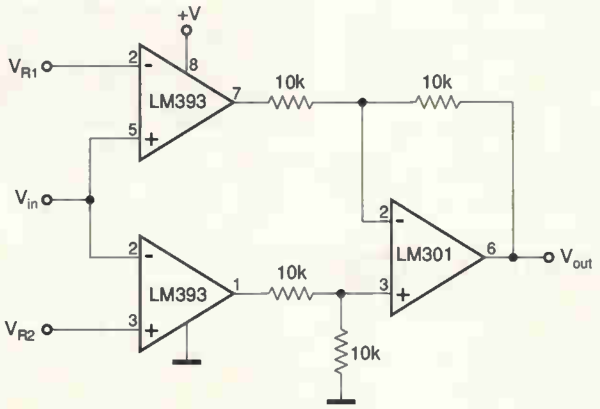


Fig. 2. Full circuit of frequency doubler. Replacing difference amplifier with summing amplifier produces a pulse variable in width from zero to 100% of the input.

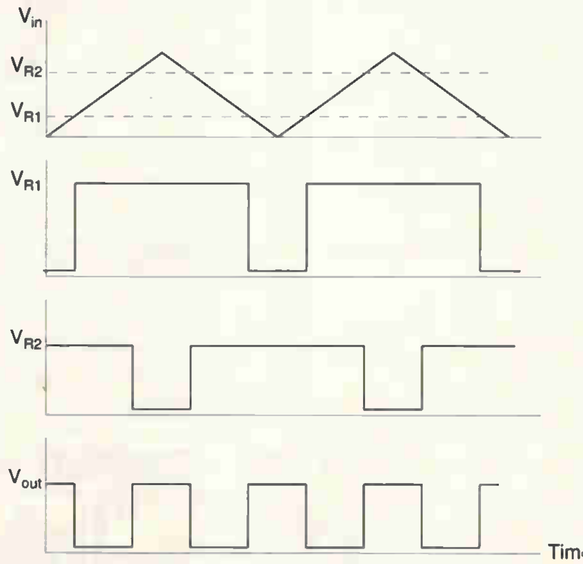


Fig. 3. Difference amplifier subtracts one comparator output from the other to produce a frequency twice that of the input triangular wave.

Input signal	Comment	LEDs			
		Red	Green	Orange	Yellow
	High	On	Off	N/A	N/A
	Low	Off	On	N/A	N/A
	Transition low to high	On	Off	On	Off
	Transition high to low	Off	On	Off	On
	Single pulse	Off	On	On	On
	Single pulse	On	Off	On	On
	Pulsing	Blinking	Blinking	On	On

Fast pulse capture/analysis

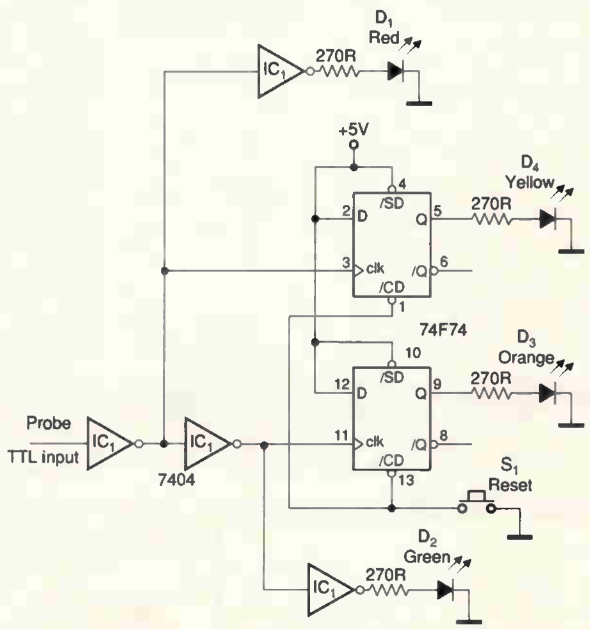
Needing to detect the presence of a single 1μs pulse, which was too fast for a logic analyser to see, I made the circuit shown here, which will show the presence of a pulse down to 10ns in width and also indicates the activity on a signal line.

Indication is by a series of leds. Logic high at the input causes the red led to illuminate, a low lighting the green one. Inverted and non-inverted versions of the input form clock signals for the two edge-triggered flip-flops; a low-to-high transition triggers the Q output at pin 5 to light the orange led, a high-to-low triggering the pin 9 Q output to illuminate the yellow led; a single pulse turns on both orange and yellow leds, the red and green indicating whether it is negative or positive going, as shown by the table.

Switch S_1 resets both flip-flops. Continuous activity on the line causes the red and green leds to flash.

Ken Yang
La Jolla
California
USA
A43

Using only two ics, this logic-level indicator will capture a single pulse of around 10ns in width and analyses the state of a signal line.



Precise mains voltage measurement

Having a need to display high-resolution mains voltage readings in the 190-250V ac range at the input of an uninterruptible power supply which has an 8-bit a-to-d converter, I had to convert the range into a 0.5-4.5V dc swing, that being the input range of the converter. In addition, I needed overvoltage and undervoltage warnings on a seven-segment display.

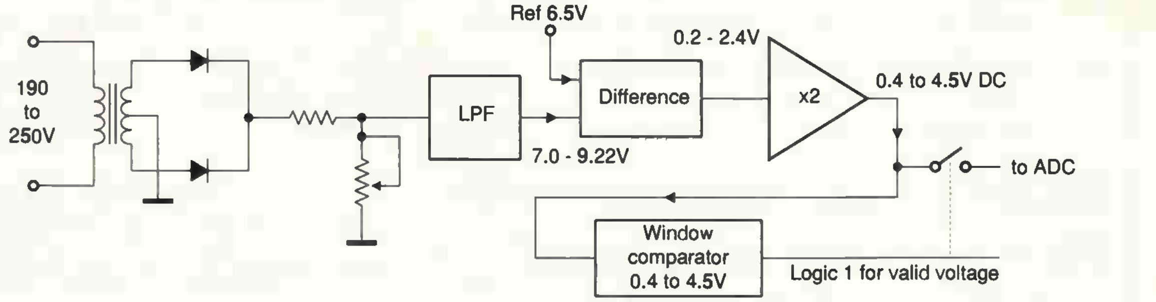
Ups output is transformed, full-wave rectified (the 6-0-6V transformer was handy) and passed through two second-order 15Hz filters. Output is adjusted to give 7V dc for a 190V ac input, the swing being 7-9.22V dc for the full 190-250 ac mains excursion.

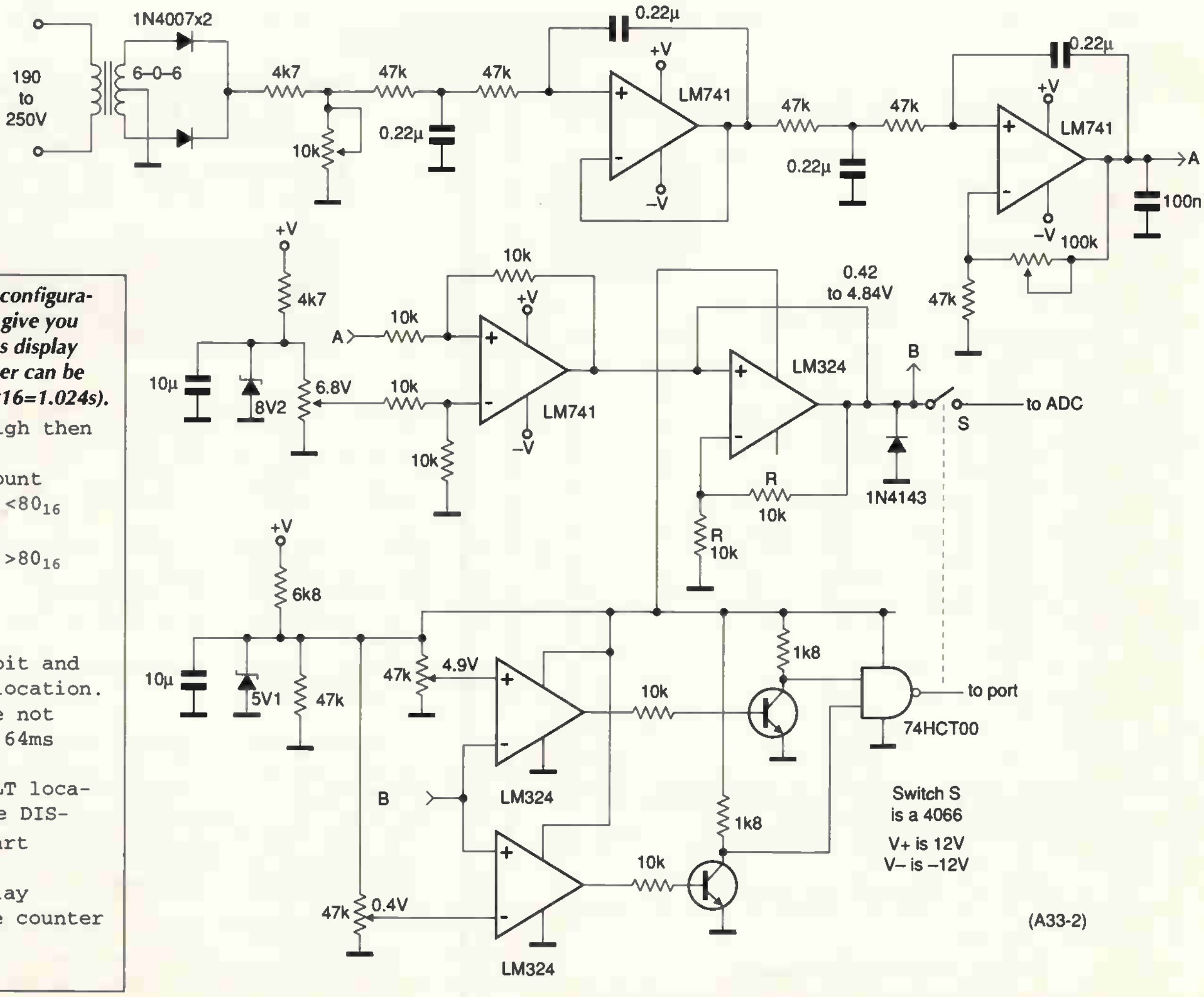
This goes to a difference amplifier, of which the other input is 6.8V, so that the 0.2-2.4V swing, having been amplified by a factor of two, is within the limits of the a-to-d converter at B.

An out-of-range signal results from the action of the window comparator, which takes its input from B and has comparison levels of 4.9V and 0.4V, producing a logic 1 at the output when the input is within range.

Jayant Kathe
Bombay
India
(A33)

High-resolution mains voltage measurement in the 190-250V ac range, requiring a 0-5V dc output to an 8-bit a-to-d converter. Rest of the circuit is an in range/out of range warning.





Software depends on the configuration, but this flow should give you some ideas. Assuming a 1s display update, an a-to-d converter can be taken every 64ms (64ms×16=1.024s).

```
1 If port pin is high then
goto 6
2 Get stored ADC count
3 If I/P ADC count <80₁₆
Display
4 If I/P ADC count >80₁₆
Display
5 goto 1
6 Get ADC count
7 Right shift one bit and
ADD to MAINS-VOLT location.
8 If 16 samples are not
over than wait for 64ms
goto 1.
9 Add the MAINS-VOLT loca-
tion ₁₆ value to the DIS-
PLAY-DATA-TABLE start
address.
10 Update the display
11 Reset the sample counter
to 0 goto 1
```

50/60Hz sine oscillator

A Wien-bridge oscillator is the obvious type of circuit to use for a vlf oscillator, providing a low-power, stable output while using low-voltage, low-current supplies – particularly if op-amps providing rail-to-rail output are used. However, the circuit shown offers good initial accuracy and low-frequency drift by virtue of the crystal clock oscillator, and amplitude stability. No adjustments or high-value precision capacitors are needed.

The *R,2R* ladder is driven by the logic-level i/o lines from a *PIC* microcontroller, data coming from a software look-up table. Resolution is to seven bits and amplitude 0.02V-2.61V. A 3-bit counter output is also available. High ladder resistance provides a light load for the logic outputs, although they are capable of much more current. The circuit takes 52μA as shown, and may be reduced by scaling up resistor values at the risk of increased emi.

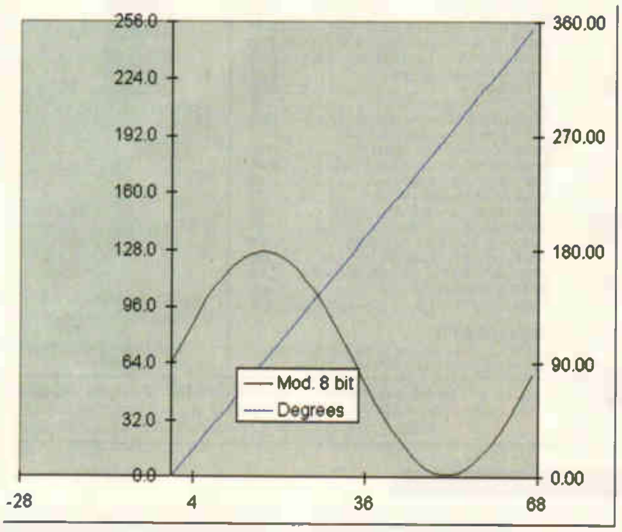
At a standard clock frequency of 32.768kHz, output frequencies are 49.951Hz and 60.235Hz; to make these exact, the clock should be 32kHz. You can do this, at the expense of a higher quiescent current and a greater number of components, by using a 4.096MHz crystal and dividing it in, say, a *74HC4060* binary counter/divider to give exactly 32kHz.

MAX951 contains an op-amp, a comparator and band-gap reference, the op-amp and reference providing a 2.63V supply for the microcontroller and the comparator converting the microcontroller's msb to a 'phase' signal that goes high during positive half-cycles of the output.

As well as the look-up table for both frequencies, the

PIC listing on disk

Because the assembly language routine for the PIC would have taken up around four magazine pages we were unable to publish it. If you want a copy of this fully-annotated listing and the Excel spreadsheet shown, on 3.5in disk, send £10 to cover administration and copying to Electronics World's editorial offices, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS, with a clear note of what you are requesting and your address.



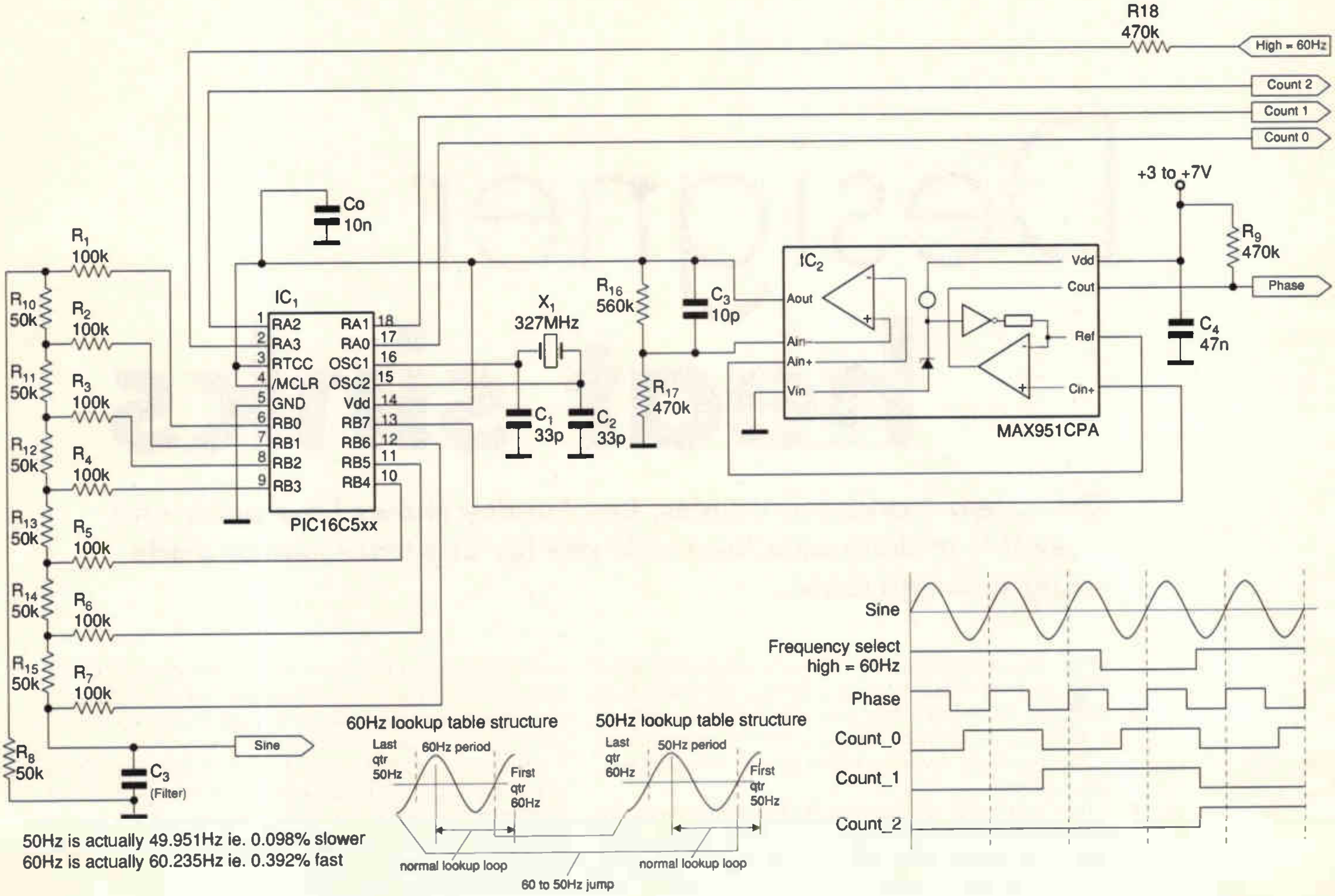
Begin and end of the Excel file, below, used to determine values in the PIC program illustrates how a spreadsheet can break the back of a digital design.

File Edit View Insert Format Tools Data Window Help

	A	B	C	D	E	F	G	H	I	J	K	L	M	N	O
1	Steps:	68.00				60 Hz									
2	2pi/steps:	0.09													
3	Address	Degrees	Angle	Mod. 8 bit	Rounded	PIC Instruction								Replace Q with \newline	
4	0	0.00	0.000	64.0	64	movlw 0x	C0		0	at 60Hz is	64.0	rounded to	64	Q movwf port_b ;	
5	1	5.29	0.092	69.8	70	movlw 0x	C6		1	at 60Hz is	69.8	rounded to	70	Q movwf port_b ;	
6	2	10.59	0.184	75.6	76	movlw 0x	CC		2	at 60Hz is	75.8	rounded to	76	Q movwf port_b ;	
7	3	15.88	0.274	81.2	81	movlw 0x	D1		3	at 60Hz is	81.2	rounded to	81	Q movwf port_b ;	
8	4	21.18	0.361	86.8	87	movlw 0x	D7		4	at 60Hz is	86.8	rounded to	87	Q movwf port_b ;	
9	5	26.47	0.446	92.1	92	movlw 0x	DC		5	at 60Hz is	92.1	rounded to	92	Q movwf port_b ;	
10	6	31.76	0.526	97.2	97	movlw 0x	E1		6	at 60Hz is	97.2	rounded to	97	Q movwf port_b ;	
11	7	37.06	0.603	102.0	102	movlw 0x	E6		7	at 60Hz is	102.0	rounded to	102	Q movwf port_b ;	
12	8	42.35	0.674	106.4	106	movlw 0x	EA		8	at 60Hz is	106.4	rounded to	106	Q movwf port_b ;	
13	8	47.65	0.739	110.6	111	movlw 0x	EF		9	at 60Hz is	110.6	rounded to	111	Q movwf port_b ;	
14	10	52.94	0.798	114.3	114	movlw 0x	F2		10	at 60Hz is	114.3	rounded to	114	Q movwf port_b ;	
15	11	58.24	0.850	117.6	118	movlw 0x	F6		11	at 60Hz is	117.6	rounded to	118	Q movwf port_b ;	
16	12	63.53	0.895	120.4	120	movlw 0x	F8		12	at 60Hz is	120.4	rounded to	120	Q movwf port_b ;	
17	13	68.82	0.932	122.7	123	movlw 0x	FB		13	at 60Hz is	122.7	rounded to	123	Q movwf port_b ;	
18	14	74.12	0.962	124.6	125	movlw 0x	FD		14	at 60Hz is	124.6	rounded to	125	Q movwf port_b ;	
19	15	79.41	0.983	125.9	126	movlw 0x	FE		15	at 60Hz is	125.9	rounded to	126	Q movwf port_b ;	
20	16	84.71	0.996	126.7	127	movlw 0x	FF		16	at 60Hz is	126.7	rounded to	127	Q movwf port_b ;	

File Edit View Insert Format Tools Data Window Help

	A	B	C	D	E	F	G	H	I	J	K	L	M	N	O
1	Steps:	68.00				60 Hz									
2	2pi/steps:	0.09													
3	Address	Degrees	Angle	Mod. 8 bit	Rounded	PIC Instruction								Replace Q with \newline	
4	0	0.00	0.000	64.0	64	movlw 0x	C0		0	at 60Hz is	64.0	rounded to	64	Q movwf port_b ;	
5	1	5.29	0.092	69.8	70	movlw 0x	C6		1	at 60Hz is	69.8	rounded to	70	Q movwf port_b ;	
6	2	10.58	0.184	75.6	76	movlw 0x	CC		2	at 60Hz is	75.6	rounded to	76	Q movwf port_b ;	
7	3	15.88	0.274	81.2	81	movlw 0x	D1		3	at 60Hz is	81.2	rounded to	81	Q movwf port_b ;	
8	4	21.18	0.361	86.8	87	movlw 0x	D7		4	at 60Hz is	86.8	rounded to	87	Q movwf port_b ;	
9	5	26.47	0.446	92.1	92	movlw 0x	DC		5	at 60Hz is	92.1	rounded to	92	Q movwf port_b ;	
10	6	31.76	0.526	97.2	97	movlw 0x	E1		6	at 60Hz is	97.2	rounded to	97	Q movwf port_b ;	
11	7	37.06	0.603	102.0	102	movlw 0x	E6		7	at 60Hz is	102.0	rounded to	102	Q movwf port_b ;	
12	8	42.35	0.674	106.4	106	movlw 0x	EA		8	at 60Hz is	106.4	rounded to	106	Q movwf port_b ;	
13	9	47.65	0.739	110.6	111	movlw 0x	EF		9	at 60Hz is	110.6	rounded to	111	Q movwf port_b ;	
14	10	52.94	0.798	114.3	114	movlw 0x	F2		10	at 60Hz is	114.3	rounded to	114	Q movwf port_b ;	
15	11	58.24	0.850	117.6	118	movlw 0x	F6		11	at 60Hz is	117.6	rounded to	118	Q movwf port_b ;	
16	12	63.53	0.895	120.4	120	movlw 0x	F8		12	at 60Hz is	120.4	rounded to	120	Q movwf port_b ;	
17	13	68.82	0.932	122.7	123	movlw 0x	FB		13	at 60Hz is	122.7	rounded to	123	Q movwf port_b ;	
18	14	74.12	0.962	124.6	125	movlw 0x	FD		14	at 60Hz is	124.6	rounded to	125	Q movwf port_b ;	
19	15	79.41	0.983	125.9	126	movlw 0x	FE		15	at 60Hz is	125.9	rounded to	126	Q movwf port_b ;	



Microcontroller drives resistive ladder from a look-up table to produce an accurate sine output at selectable 50Hz or 60Hz frequencies, with a 3-bit period timer output and a sync. signal.

software produces a continuous wave train by repeatedly jumping from the end of the table to the start at the positive peak, ensuring a smooth 50-60Hz change by having a sequence of the last quarter-period of the other frequency. Changes occur at rising zero crossings, as happens in frequency-shift keying.

Tim Herklots
Maxim Integrated Products
Reading
Berkshire
(A42)

Lead-acid battery saver

When a lead-acid battery is flattened completely, it is little further use. This circuit, which is small enough to use in a battery pack, prevents such abuse, automatically disconnecting the battery if its voltage drops below 10V. It allows the occasional voltage drop caused by equipment drawing a high inrush current and also automatically connects the battery to a correctly applied charging voltage.

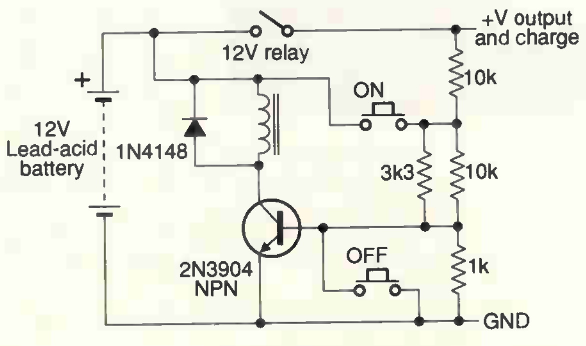
Starting from the condition when the relay is on, the battery voltage appears at the output terminals and the potential divider applies 0.81V to the transistor base, assuming battery voltage is more than 10V, the resulting base current turning it on and maintaining the relay.

Falling battery voltage reduces base current and eventually, at about 9.5V, the transistor turns off and the relay drops out, disconnecting the battery; the same result is brought about by the off button.

Starting from the odd condition, pressing the on button momentarily injects current into the transistor base and turns it on.

Daniel Greenspan
Jerusalem
(A44)

This protective circuit prevents lead-acid batteries completely discharging and is inexpensive and small enough to be used in the battery pack itself.

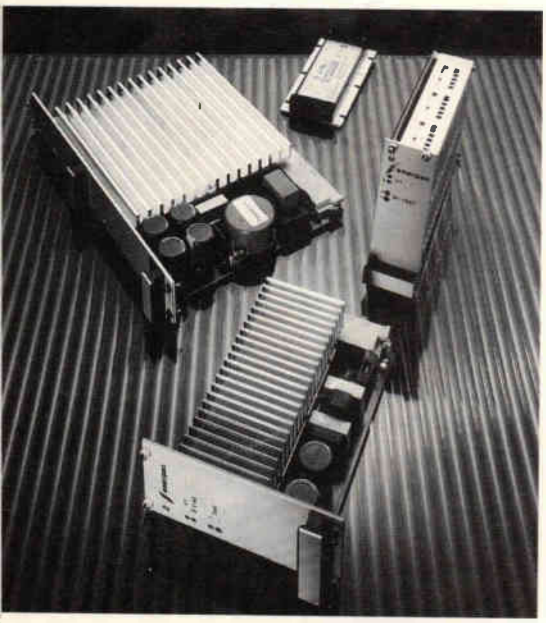


Designer heat sinks

Using two worked examples, Ray Fautley shows how quick and easy it is to determine heat sink size for any transistor or diode using a design table.

Semiconductor devices need to be mounted on metal panels of sufficient size and thermal conductivity to be able to conduct excessive heat away from the junction of the transistor or diode. This is necessary to prevent the temperature of the device's junction from exceeding its maximum rating.

The total thermal path between the device junction and the ambient temperature existing inside the case of the equipment is measured as a thermal resistance. It is referred to as $R_{th(j\text{-}amb)}$, or as $\theta_{(j\text{-}a)}$. This term that is easy to assess. You only need to know three things. One is the maximum allowable junction temperature – which for most discrete silicon devices may be assumed to be about 200°C. Secondly you need to know the ambient temperature, which, inside electronic equipment is usually about 50°C, and finally you need to know the power dissipated in the semiconductor.



This total thermal path $R_{th(j\text{-}amb)}$ comprises three parts,

- The thermal path from device function to its case, $R_{th(j\text{-}c)}$, also called $\theta_{th(j\text{-}c)}$ or $R_{th(j\text{-}mb)}$.
- The thermal path between the case of the device and the heat-sink to which it is fitted, $R_{th(c\text{-}hs)}$, also called $\theta_{th(c\text{-}hs)}$.
- The thermal path between the heat-sink and the ambient temperature inside the equipment, $R_{th(hs\text{-}amb)}$, or $\theta_{(h\text{-}a)}$.

All of these three paths – and of course, the total path $R_{th(j\text{-}amb)}$ are measured in °C/W. This makes the total thermal path $R_{th(j\text{-}amb)}=R_{th(j\text{-}c)}+R_{th(c\text{-}hs)}+R_{th(hs\text{-}amb)}$.

The first of the three terms, $R_{th(j\text{-}c)}$, can be found in the data sheets provided by the manufacturer of the semiconductor device.

Path $R_{th(c\text{-}hs)}$ depends on the method used to mount the transistor or diode to the heat-sink. In some cases the device may be mounted directly on to the heat-sink, but in others it may be necessary to provide some form of electrical insulation between the device and the heat-sink.

Ray Fautley, C Eng, MIEE

Unfortunately, electrical insulation also means some degree of heat insulation so a compromise is necessary. A thin mica washer can provide the electrical insulation. As the semiconductor's working voltage does not exceed a few tens of volts, a thick insulator is unnecessary.

Heat conduction is assisted by the use of silicon grease between the surfaces. Path $R_{th(c\text{-}hs)}$ can be considered to be between 0.1 and 0.7°C/W. In practice, a value of 0.4°C/W is usually acceptable for this term. This leaves the third term, $R_{th(hs\text{-}amb)}$, which needs to be evaluated.

Heat sink design steps

1. From the semiconductor data sheet, find $R_{th(j\text{-}c)}$ in °C/W.
2. Determine $R_{th(c\text{-}hs)}$ in °C/W. This value depends on the effectiveness of the thermal contact between the case of the semiconductor and the surface of the heat-sink. A value of 0.4°C/W will be a good enough approximation if the actual value is unknown.
3. Determine the power dissipated in the device,

   (a) For power-amplifier transistors the power dissipated is,

   $P_D=P_{in}-P_{out}$.
   $=V_{ce}\times I_c-P_{ac}$

   (b) For rectifier diodes,

   $P_D=V_D-I_{ave}$.
4. Find,

$$R_{th(j-amb)}=\frac{T_j-T_{amb}}{P_D}\text{ in °C/W}$$

For silicon devices, maximum junction temperature T_j is usually 200°C, but to be safe it is better to design around 150°C. In some cases T_j may be specified lower than 200°C, so always check with the manufacturer's data before proceeding with the heat-sink design. Inside the equipment, the ambient temperature T_{amb} can be assumed to be 50°C.

5. Determine,

   $R_{th(hs\text{-}amb)}=R_{th(j\text{-}amb)}-R_{th(j\text{-}c)}-R_{th(c\text{-}hs)}$

6. The surface area of the heat-sink required to provide the necessary $R_{th(hs\text{-}amb)}$ can be found from **Table 1**.

A variation of the thickness of the heat-sink between 1mm and 5mm does not have very much effect. It is surface area that counts. These figures should provide adequate design margins.

Note that the surface areas in the table are for one side of the heat-sink only, the total surface area is thus twice the table figure – plus twice the thickness of the metal.

All heat-sink surfaces are assumed to be painted matt-black and mounted horizontally. Vertical mounting increases the cooling effect and adds a further margin of safety. But bright metal surfaces will need the figures of Table 1 to be increased by some 33%.

Note that the figures were arrived by experimentation. After comparing various results using different methods for arriving at the necessary surface area for the same value of $R_{th(hs\text{-}amb)}$, agreement was not 100%.

Design examples

These two worked examples should help clarify the design procedure for you.

Example 1. A rectifier diode, for which the manufacturer's data gives $R_{th(j\text{-}c)}$=5°C/W, has a forward voltage drop 1.0V when the average current through it – or dc load – is 5A. The maximum junction temperature T_j of the diode is given as 100°C. Inside the equipment that the rectifier is to be used in, the ambient temperature will be assumed to be 50°C.

Design procedure

1. $R_{th(j\text{-}c)}$ is known to be 5°C/W.
2. Assume that $R_{th(c\text{-}hs)}$ will be 0.4°C/W.
3. Power dissipated in the diode is,

   $P_d=V_d \times I_{ave}$

   $=1.0\times5=5$W for a T_j of 100°C

4. $$R_{th(j-amb)}=\frac{T_j-T_{amb}}{P_D}$$

   $$=\frac{100-50}{5}=\frac{50}{5}=10°\text{C/W}$$

5. $R_{th(hs\text{-}amb)}=R_{th(j\text{-}amb)}-R_{th(j\text{-}c)}-R_{th(c\text{-}hs)}$

   =10–5–0.4=4.6°C/W

6. Referring to Table 1 gives the following dimensions for a suitable heat-sink.

   For copper, 84cm^2 or 9.2cm×9.2cm
   For aluminium, 100cm^2 or 10cm×10cm
   For brass, 113cm^2 or 10.6cm×10.6cm
   For steel, 128cm^2 or 11.3cm×11.3cm

Example 2.
A *BLW81* transistor is to be used as a vhf power amplifier to provide a cw output of 10W.

Design procedure

1. The thermal resistance from junction to case $R_{th(j\text{-}c)}$ is given by its manufacturer as 4.3°C/W.

2. Say R_{th} is 0.4°C/W.

3. Efficiency of the *BLW81* is given as about 60% as a cw amplifier, so the input power for 10W output will be,

   $$P_{in}=\frac{P_{out}}{0.6}=\frac{10}{0.6}=16.7\text{W}$$

   So,

   $P_D=P_{in}-P_{out}$=16.7–10=6.7W.

4. $$R_{th(j-amb)}=\frac{T_j-T_{amb}}{P_D}$$

   $$=\frac{150-50}{6.7}=14.9°\text{C/W}$$

5. $R_{th(hs\text{-}amb)}=R_{th(j\text{-}amb)}-R_{th(j\text{-}c)}-R_{th(c\text{-}hs)}$

   =14.9–4.3–0.4=10.2°C/W

6. From Table 1, suitable heat-sinks are,

   For copper, 29cm^2 or 5.4cm×5.4cm
   For aluminium, 35cm^2 or 5.9cm×5.9cm
   For brass, 37cm^2 or 6.4cm×6.4cm
   For steel, 130cm^2 or 11.4cm×11.4cm

My thanks to Anglia Microwaves of Billericay, Essex for their help in providing semiconductor data. ■

Heat sink terms

Term	Description	Units
I_{ave}	Average current through diode	A dc
I_c	Transistor collector current	A dc
P_D	Power dissipated in transistor	W
P_{in}	Power input to device, dc	W
P_{out}	Output power – rf or af from transistor	W
$R_{th(c\text{-}hs)}$	Thermal resistance between device case and heat-sink	°C/W
$R_{th(hs\text{-}amb)}$	Thermal resistance between heat-sink and T_{amb}	°C/W
$R_{th(j\text{-}amb)}$	Thermal resistance between device junction and T_{amb}	°C/W
$R_{th(j\text{-}c)}$	Thermal resistance between device junction and case	°C/W
T_{amb}	Ambient temperature inside equipment	°C
T_j	Maximum temperature of device junction	°C
V_{ce}	Potential between collector and emitter of device	V
V_d	Voltage drop across diode at stated current	V dc

Table 1. These figures give you the surface area in cm^2 for one side of the heat sink for a known value of $R_{th(hs\text{-}amb)}$.

$R_{th(hs\text{-}amb)}$ °C/W	Material Copper	Aluminium	Brass	Steel
1	734	—	—	—
1.5	342	—	—	—
2	220	376	477	—
2.5	175	277	302	711
3	135	187	210	338
3.5	116	149	175	222
4	94	120	150	165
4.5	85	102	115	132
5	70	83	90	109
6	53	68	72	82
7	41	61	63	66
8	37	55	58	61
9	33	39	44	47
10	29	35	37	41
11	26	30	33	36
12	24	29	30	32
13	22	27	29	30
14	20	25	26	27
20	14	17	18	20

NEW PRODUCTS CLASSIFIED

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ACTIVE

Discrete active devices

15W rf power. Ericsson's surface-mounted, *PTB20220* rf power transistor provides 15W of rf power in the 915-960MHz cellular radio band. It is a Class AB n-p-n, common-emitter device for pep and cw use, efficiency being 54% and power gain a typical 10dB at 15W. Ericsson Components AB. Tel., 01793 488300; fax, 01793 488301.

Efficient small-signal mosfets. Using the company's high cell-density mos process, Motorola's new range of SOT 23-packaged, low-$r_{DS(on)}$ mosfets show a 50% power-loss reduction over earlier types. On resistances vary from 5Ω at 10V in the 60V *MMBF170LT1* down to 0.085Ω at 10V in the 20V *MGSF1N02LT1*. Four of the six new devices are n-channel types; two 20V p-channel devices give 0.1Ω and 3.5Ω at 10V. Motorola. 001 602 244 6108; fax, 001 602 244 4597.

Linear integrated circuits

High-power op-amp. *OPA547* by Burr-Brown is designed to drive a variety of heavy loads, such as motors and actuators, for use in power supplies or in audio amplifiers. It will take a single 8-60V supply or dual ±4V to ±30V lines and produces a continuous current output of 500mA. It is protected against excess temperature and current overload and there is provision for an accurate, user-set current limit of up to 750mA, not using a power resistor but by means of an indirectly sensed current, a limit pin accepting input from a potentiometer or d-to-a converter. Thermal shutdown is indicated by a status pin. Burr-Brown International. Tel., 01923 233837; fax, 01923 233979.

Hybrid 4kW pwm amplifiers. Apex Microtechnology *SA03/04* pwm amplifiers are for use in applications from motor control to low-frequency sonar, providing continuous current output up to 30A and handling an input voltage range of 16-200V; all circuitry is contained in hermetically sealed MO-127 power packages rated at temperatures in the –55°C to 125°C range. These devices accept analogue or digital input and switching frequency is 22.5kHz or, for lower frequencies, external oscillator input is provided. Ashwell Electronics Ltd. Tel., 01438 364194; fax, 01438 313461.

Memory chips

4Kb SPI eeprom. *25C040* from Microchip is the newest member of the 8, 16 and 32Kbit Serial Peripheral Interface eeprom family, designed to interface directly with the SPI port of many microcontrollers, including PIC16CXX and PIC17CXX types. This 3MHz device has several security features, including user-selected write protection and up to 10 million erase/write cycles are guaranteed. A serial eeprom designers' kit is available. Arizona Microchip Technology Ltd. Tel., 01628 851077; fax, 01628 850259.

Non-volatile sram. In a single ic, the *STK 14C88* from Simtek, there is a fast 32K by 8-bit static ram with an access time of 25ns in the fastest version, backed up by a eeprom. By means of a function called Autostore, the device monitors the 5V supply and starts a backup cycle if power is switched off or low. If one cannot be sure that the supply will stay above 3.6V during the 10ms backup time, a 100µF capacitor can be used to hold it up. On reappearance of the power supply, the data is automatically copied back to the sram, this being possible at any time under the user's control. Pronto Electronic Systems Ltd. Tel., 0181-554-5700; fax, 0181-554-6222.

Motors and drivers

Ac motor power conversion. International Rectifier's series of *PoweRtrain* power conversion systems for ac motors is increased to handle 1.1kW (*IRPT2055*) and 5.6kW (*IRPT4052*) motors, each device containing all necessary circuitry to make motor drives and controls, with or without braking, in packages 75% smaller than competing types. Both types take 460V ac inputs, the former being in a chip-and-wire module and the latter a surface-mounted assembly on a metal substrate. Both include a board containing a driver with fault protection and shutdown facilities. A new brochure on PoweRtrain devices is available. International Rectifier. Tel., 01883 732020; fax, 01883 733410.

Optical devices

Dual-wavelength wdm transceiver. Using the company's active silicon integrated optical circuits technique (ASOC), Bookham Technology's *BKM 2400* dual-wavelength transceiver, provides full duplex, single-mode, single-fibre, dual-wavelength operation on 1310nm and 1550nm at data rates of 155Mb/s or faster. It is meant for use in bidirectional networks such as point-to-point and passive optical networks for fibre-to-the kerb, fibre-to-the building and fibre-to-the-home. Transmission is on 1310nm and reception on 1550nm, a wavelength-dependent waveguide structure routeing the signals to the appropriate circuitry. Bookham Technology. Tel., 01235 827200; fax, 01235 827201.

PASSIVE

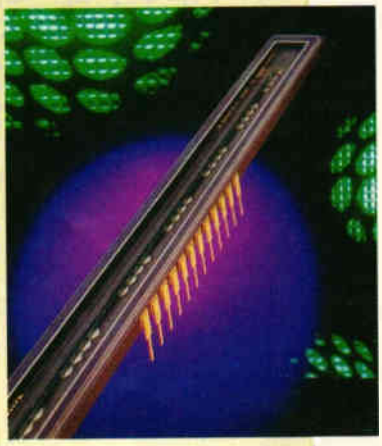
Cameras

Camera chipset. Sony's *SS-1* three-chip set, together with a single colour ccd, forms a low-cost, light PAL/NTSC camera meant for use in security and multimedia. It carries out all ccd driving and signal-processing functions, working with Sony ccds from 0.2in, 180k-pixel types to 0.5in, 380k-pixel versions. A *CXD2163* signal processor has an 8-bit a-to-d converter, or a 9-bit or 10-bit converter may be used externally for better resolution, generates sync. and luminance/chrominance processing for both standards. Both analogue and CCIR601-compatible digital output is available. Silicon Concepts Ltd. Tel., 01428 751617; fax, 01428 751603.

Passive components

Low-esr electrolytics. Sanyo *Poscap* surface-mounted electrolytics are for use in high-frequency power supplies, offering low equivalent series resistance and impedance; they use a conductive organic polymer for electrolyte. There are two versions: as examples, the AP (aluminium) 16V, 2.2µF capacitor has an esr of 120mΩ and 800mA rms ripple, while the tantalum TP 4V, 220µF handles ripple current of 1.2A rms and has an esr of 80mΩ. Semicom UK Ltd. Tel., 01279 422224; fax, 01279 433339.

Membrane pots. *M-POT 890* is part of a range of membrane potentiometers developed by CTS Corp., in which the resistive layer is formed by a carbon deposit on a substrate providing a resistive gradient between the ends. A slider on the top surface connects conductive and resistive layers to vary the resistance from zero to over 500kΩ. These devices are immune to the effects of humidity and static discharge to 18kV. Thickness is 2mm. Quiller Switches Ltd. Tel., 01202 436777; fax, 01202 421255.



Linear document reader. A colour ccd linear sensor for A4 image scanners is announced by Sony. *ILX734K* is a reduction-type sensor having a shutter for each colour and, with 8µ by 8µ pixel size, has 31500 effective pixels to read an A4 document at 1200dots/in, anti-reflection coating on the lower glass reducing unwanted signals. The sensor is contained in a 24-pin dip, requires 12V and a 5V clock pulse. Sony Semiconductor Europe. Tel., 01256 478771; fax, 01256 818194.

Audio products

MPEG2 decoders. Philips has announced a family of multi-channel audio decoders to the MPEG2 standard for use with DVD-Video players, set-top boxes, multimedia pcs and the rest. The first to emerge is the *SAA 2503* decoder, which provides a down-mix of the datastream and so enables 5+1 (5 main speakers and an lf one) or 7+1 (7 and the lf) to be preserved when heard as a stereo output; this would not be the case without this chip, as the additional audio information in MPEG2 would be lost and the MPEG1 data only heard. For this reason, MPEG standards are reverse

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and forward compatible. Further chips in the family, to the AC-3 standard for NTSC, will appear later, as will dual-standard devices. Philips Semiconductors (Eindhoven). Tel., 00 31 40 2722091; fax, 00 31 40 2724825.

Connectors and cabling

Thin s-m connectors. Up to 30 contacts are possible within a mounted height of only 2mm in the JAE *LZ Series* of wire/board connectors. Pitch is 1.25mm and there is a polarising key to prevent misalignment. Current rating is 1A and dielectric strength 500V ac. Flint Distribution. Tel., 01530 510333; fax, 01530 510275.

Surface-mounted coax. connectors. Transradio's *MMT* series of s-m coaxial connectors provide very low vswr, low insertion loss and good shielding. There are models for use at frequencies from zero to 6GHz and a designers' kit and an extraction tool are available. MMT connectors consist of a s-m receptacle, mating with a right-angled plug, the receptacle having a male centre contact. The plug is nickel-plated, die-cast, and has a crimp ferrule, being suitable for cable types including RD316 double-screened 2.6mm diameter cable, RG174 and RG178 types. Receptacles are stamped and formed and has a gold-plated body and centre contact, standing 6.8mm off the board. Transradio Ltd. Tel., 0181-997-8880; fax, 0181-997-0116.

Vibrators. Vibrator motors by Copal, intended for use in mobile telephones and pagers, use gold commutators to avoid the need for grease and to prolong their life – around 1000h continuous. *LA series* motors are 6mm in diameter, from 19.1mm to 22.8mm long and run at speeds from 5700rev/min to 7700rev/min, all being rated at 1.3V. Eccentric tungsten weights provide the vibration. A recent introduction is a motor for 5mm card pagers, measuring 5mm by 4.2mm by 22mm. electronic Ltd. Tel., 01993 778000; fax, 01993 772512.

Displays

RGB dot-matrix led module. The American company Lumex announces a 5-by-7 dot-matrix RGB array of leds, 2.09in high. There are three leds per dot, giving a minimum of 300mcd at 10mA in blue and a viewing angle of 100°. Each colour of every dot may be addressed either individually or collectively and response time is short enough for fast animation. Dots are 0.3in diameter on 0.3in centres, two rows of 11 pins on the back, in 0.1in pitch being suitable for pcb mounting. Non-standard sizes or patterns can be made to order. Lumex Opto/Components Inc. Tel., 001 847 359-2970; fax, 001 847 359-8904

Filters

Drive-bay signal conditioners. Kemo has a range of active filters and signal conditioners that fit into the 5.25in floppy drive bay, which is now usually little used. Benefits of this type of mount are that screening is much better than when mounted in a card slot, and that the larger space allows better layout; filtering, too, is independent of the computer type and operating system. The *Model 100* range has front-panel BNC connectors for inputs and cabling at the rear to connect the relevant data acquisition card. Signal conditioning may be computer-controlled or manual, the latter having front-panel switches to set functions. Kemo Ltd. Tel., 0181-658-3838; fax, 0181-658-4084.

Hardware

Desiccators. Brownell has a range of desiccators designed to eliminate humidity and condensation inside equipment housings. They are small polycarbonate mouldings in two types: one uses silica gel for general use, the other having molecular sieves for use with optical and laser equipment. Both types protect enclosures with volumes up to 100l, the type selected depending on volume and IP rating or equivalent. Each unit has a saturation indicator to show the need for reactivation or replacement. Brownell Ltd. Tel., 0181-965-9281; fax, 0181-965-3229.

No-tooling enclosures. *K Box*, a new division of Boss, has a new method of making low-cost plastic enclosures that need no tooling and are designed to order. They are made entirely of flat sheet material, so that no moulding or forming is needed and prototypes are available in a few days, modifications being easily accommodated. The cases can be produced in several materials in 2-10mm thickness, may be internally coated for shielding and designed to take rough handling. Boss Industrial Mouldings Ltd. Tel., 01638 716101; fax, 01638 716554.

400MHz dso. The Tek *TDS380* digital storage oscilloscope is the fastest in the *TDS300* series, sampling at 2Gsample/s and having a bandwidth of 400MHz. It is able to collect enough samples to show waveforms at full bandwidth to make single-shot acquisitions at timebase speeds to the maximum of 1ns/div, the high-speed sampling also reducing aliasing. There are 21 automatic waveform measurements and the four acquisition modes of sampling, peak detection, envelope and average with, in addition, live FFT analysis. A package called Option 14 provides GPIB and RS232-C interfaces, a VGA monitor and a Centronics port, all for remote control of the instrument from a pc. Waveforms displayed may be saved in TIF, PCX, BMP and EPS file formats. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Access to bgas. Difficulties in attaching probes to ball-grid arrays for test purposes are eliminated by the adaptors now produced by Winslow International. They consist of a solder-down module to suit any device with a square matrix down to 1mm pitch (interstitial), using a hard-ball process with pins coming from the top, for which any interface arrangement can be supplied. Winslow International Ltd. Tel., 01874 625555; fax, 01874 625500.

Test and measurement

Analogue oscilloscope. Hitachi's *V-252* 20MHz, dual-channel oscilloscope is claimed to be the best value for money and, since the information supplied says it costs £0000, there is no arguing with that. Unfortunately, further enquiry reveals a price of £435, which is still pretty good; its warranty lasts three years. Y sensitivity is 1mV/div using x5 magnification and timebase speed a maximum of 0.1µs/div with a 10x magnifier. All the usual X and Y facilities are provided, as are accessories including a viewing hood, although an electrostatic charge repellent blue screen filter helps with high ambient light levels. Hitachi Denshi (UK) Ltd. Tel., 0181-202-4311; fax, 0181-202-2451.

Accessible oscilloscopes. While sacrificing nothing in the way of performance, Hewlett-Packard has reduced the resemblance of digital storage oscilloscopes to 747 flight decks, so that the new *Infinium* family of instruments looks rather more like a 'real' oscilloscope, with controls like the older, purely analogue types. Even so, with bandwidths to 1.5GHz, sampling rates to 8Gsample/s and memory depths up to 64K/channel, these are top-end instruments. A Windows '95-based graphical user interface eases access to advanced features via dialogue boxes and icons and an information system gives a guide to measurements of noise and jitter, to setting up advanced facilities such as FFTs and other, even basic procedures. Hewlett-Packard Ltd. Tel., 01344 366666; fax, 01344 362269.

Emissions testing. Laplace has added two new facilities to its 1.1GHz emc emission test *SA1000* hardware and software, addressing problems of test site anomalies and widespread unfamiliarity with the subject. Ideally, test sites should be open, free of reflections and using a 4m mast, a set of conditions that is rarely met. The new software automatically corrects for test site errors. To take account of unfamiliarity with the process, the software now includes the

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TestDirector mode, essentially an expert system that guides the user in the procedure. Laplace Instruments Ltd. Tel., 01692 500777; fax, 01692 406177.

Oscilloscope/recorder. Yokogawa's *DL708* is the first in a series of instruments that combine the capabilities of oscilloscopes and recorders, costing less than either of similar performance. Display is on a 10.4in colour tft screen and there is a built-in thermal printer for hard copy. Data can be saved on the 1.2Gbyte hard disk or on a 3.5in floppy, the drive also being built in. There is a number of signal-conditioning modules, with bandwidth to 2MHz and 16-bit resolution, and up to eight channels may be plugged into the mainframe for simultaneous capture or recording. Analysis comprises voltage measurement, timing and FFT and waveforms can be analysed in terms of overshoot, period, standard deviation and rise time. Martron Instruments Ltd. Tel., 01494 459200; fax, 01494 535002.

Recorder with memory. The Hioki *Model 8806* 400ksample/s memory recorder is a battery-powered, two-channel recorder that uses pc sram cards to extend its capacity and to allow data to be analysed on the pc, operating in both recorder and memory modes. It will handle two analogue inputs and two logic signals, which are displayed on a 5in screen and printed immediately or saved for later printing. Telonic Instruments Ltd. Tel., 01734 786911; fax, 01734 792338.

Inrush protection. *Joyin JNR* thermistors are ntc metal-oxide ceramic types capable of suppressing inrush currents up to 12A in power supplies, lighting and in motor starting circuitry. They are available in cold resistances of 0.7Ω to 220Ω, maximum hot resistance being down to 32mΩ. Five body sizes and characteristics can be had, from the 10mm diameter *JNR08* to the 25mm *JNR20*, with time constants of 36s to 115s. Anglia. Tel., 01945 474747; fax, 01945 474849.

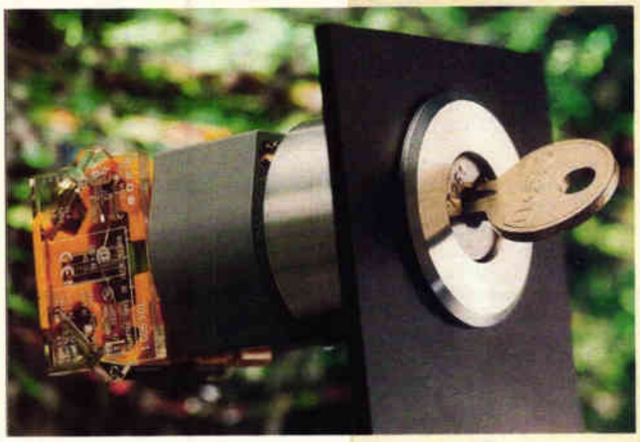
Surface resistance tester. *SRM100-3/B* from Practical Power checks surface resistivity in the 10-100V range and displays the result on an analogue panel meter. It uses a single 2.5kg weight with a square probe and conforms to BS EN100015, has CE marking and EMC approval. Practical Power Ltd. Tel., 0118 9699170; fax, 0118 9699171.

Three data recorders. Gould has introduced three new recorders. *DL250* performs the functions of data logging, multipoint recording, alarm annunciation and digital indication and is available in 16-64 input form, any combination of thermocouples, rtds, current and voltage being simultaneously monitored. RS232 or RS485 communication is provided. With up to 15 channels, the *DL100* is a 100mm type having four direct inputs for mV, V and mA input and a variety of thermocouples and rtds, inputs being scalable for linear, square root and/or log. characteristics. Finally, the *Digigraf* is a paperless recorder for six channels, a 16-bit a-to-d converter giving good resolution; data points are stored in a buffer for display, alarm, calculation or archive. Many maths functions are available and results are displayed on a 5in colour lcd with touchscreen menus. Gould Instrument Systems Ltd. Tel., 0181-500-1000; fax, 0181-501-0116.

Curve tracer. Hameg's *HM6042* semiconductor curve tracer is simple to use, is suitable for the testing of two and three terminal devices and gives an on-screen display of five curves, with additional information on an lcd. The five displayed curves indicate base voltage, base current, collector current, collector voltage and beta, characteristics of the device under test being entered by means of a front-panel keypad. A power limiter prevents damage to the device. Sets of curves may be stored in memory to ease comparison with other devices. Feedback Test and Measurement. Tel., 01892 653322; fax, 01892 663719.

Interfaces

Eight-channel driver ic. Allegro offers the *UDN2987LW*, designed as an interface between standard low-level logic and higher-power devices such as relays, motors and lamps. It is in a 20-pin, wide-body soic and has thermal shutdown and output transient protection with clamp diodes for use with sustaining voltages up to 35V. Each channel has a latch to turn the channel off in the presence of excessive current, all channels being turned off in thermal shutdown, either condition being indicated by a common fault output. All outputs produce over 100mA continuously and inputs handle 5V and 12V logic, including ttl, Schottky ttl, dtl, pmos and cmos. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.



Keylock switches. New two and three-position, flush-mounted keylock switches in EAO-Highland's 04 family for use in control gear are tamper-proof and easy to clean. They fit a 30.5mm diameter hole and protrude a mere 2mm from the panel. Switches are sealed to IP65 and are rated at 500V, 10A, each being fitted with up to four snap-action, tactile contact blocks with plug-in screw terminals and a choice of silver or gold-on-silver contacts. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Literature

Burr-Brown CD-rom catalogue. On one CD-rom, Burr-Brown has collected the data from two 1300-page data books of analogue and mixed-signal devices and an applications handbook, together with search facilities. It is free and can be obtained by e-mail on cd-rom@burr-brown.com, from the website on http://www.burr-brown.com/ or from Burr-Brown International. Tel., 01923 233837; fax, 01923 233979.

Rf transistors. Two catalogues from Ericsson on rf power transistors describe around 100 devices, ranging from 0.25W to 175W at frequencies from 300MHz to 2.2GHz, in both ldmos and bipolar types and contained in flange and surface-mounting packages. One of the publications is a short listing with a selection guide, while the other is a 350-page data book. Ericsson Components AB. Tel., 01793 488300; fax, 01793 488301.

DC-to-dc converters. Power Convertibles offers a six-page selector, giving details of the company's modules in the 0.75W to 100W range, including the new, low-cost 2W and 3W types. Those models in the 0.75-5W area are isolated-output types in s-m and sip/dip packages, providing 5V, 9V, 12V and 15V, while the *WFC02R* and *WFC03R* 2W and 3W models are lower in cost and are 81% efficient, in 32mm by 20mm by 10mm packages and 24-pin dils. There are also 8000V isolation types with output noise down to 1mV pk-pk. Power Convertibles Ireland Ltd. Tel., 00353 61 474133; fax, 00353 61 474141.

Power supplies

100W power supply. Weir Lambda's new *Excel 100* is a 100W, four-output, convection-cooled supply for general-purpose use. It meets EN61000-3-2 but is available without power-factor correction for those applications in which it is not needed, in which case it costs less. It stands 42mm high without its cover (43mm with) and fits a 1U rack. As standard, outputs are three positive and a common zero, in a variety of arrangements. Input is 85-127V and 170-254V. Weir Electronics Ltd. Tel., 01243 865991; fax, 01243 868613.

60W dc-to-dc converters. The *PKG 2000 I* family of 11mm high converters by Ericsson have a power density of 20W/cm^3, working at an efficiency of more than 84%. The units are intended for use in decentralised 24V dc systems and accept inputs of 18-36V, turning off at 16V. *PKG 2611* provides one output of 5V at 12A, the

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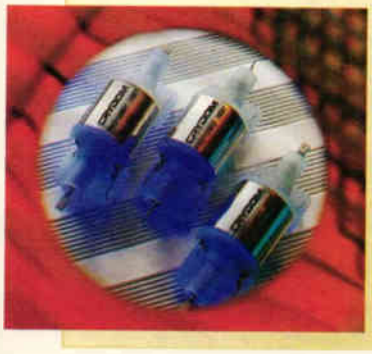
PKG 2623 giving two of 12V at 4A, and all can be connected in parallel or series. If forced convection cooling is applied, the units need no heat sink up to 60°C. Acal Electronics Ltd. Tel., 01344 727272; fax, 01344 424262.

Radio communications products

Full duplex wlan chipset. Harris's *Prism* 8-chip set forms a full duplex, heterodyne radio transceiver with agc and carrying voice and data for use in wireless lans, point-to-point microwave, hand-held data transceivers and personal communications. It operates in the 1.7-2.7GHz band and carries out all necessary functions of amplification, up/down conversion and modulation/demodulation. Receiver sensitivity is –11dBm and agc range 90dB. Harris Semiconductor UK. Tel., 01276 686886; fax, 01276 682323.

Pll synthesisers. Vari-L *PLL300/400* series of phase-locked loop synthesisers for mobile communications, industrial use and wireless modems use a patented single-ended voltage-controlled oscillator for very low phase noise, low harmonic content and stability. The devices contain all necessary circuitry, including programmable dividers, phase detectors and loop filter to make a complete circuit. There are frequency-agile devices, requiring serial programming and fixed-frequency units, with or without programming. In 13 variants, centre-frequency coverage is 755-992MHz, swinging ±15MHz, and two units of 2150MHz and 2450MHz centre frequency swinging ±50MHz. Acal Electronics Ltd. Tel., 01344 727272; fax, 01344 424262.

Reed relays. Reed relays in Crydom's *FRD 13000 Series*, which are intended for rf work, are fully screened in a metal encapsulation and mounted on an industry-standard frame. Applications are to be found in vehicle and base station communications, antenna filters and tuners, rf test gear and many other rf applications. Screening reduces interference between coil and switch which reduces heat dissipation and gives low transmission losses of less than 50mΩ at 30MHz. Both normally open and normally closed types are available. Isolation voltage is up to 9kV dc and the switches carry 6A at 30MHz. Crydom. Tel., 01202 897969; fax, 01202 891918.



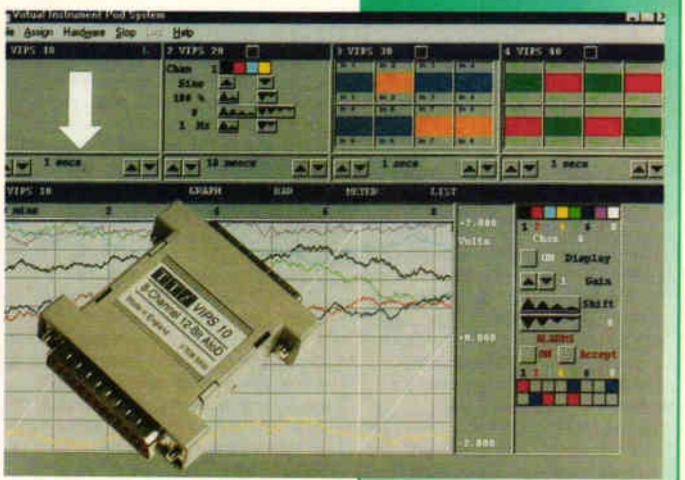
Protection devices

600W transient suppressors. Surface-mounted components from Semitron, using the glass passivated TVS diode technique as the axial-lead types, are now available. *SMB Series* transient suppressors have large contacts to handle fast rising pulses caused by lightning, inductive switching and esd and are rated at 600W during a 10x1000µs double exponential waveform. Fast turn on, a low clamping factor and low on impedance make them suitable for the protection of cmos ics, the surface-mounted form with its lower lead inductance leading to an application on high-speed, low-voltage data lines. Voltage stand-off range is 5-170V, clamping to 9V taking place in under 1ps. The devices are unidirectional or bidirectional; in the forward direction the SMB series sustains 100A for 8.3ms. Surtech Distribution Ltd. Tel., 01256 840055; fax, 01256 479785.

Fused resistor networks. Ericsson offers the *PBR53001/1* and *PBR53201/1*, which protect telecommunications line cards against induced voltages or direct shorts from underground power lines; they have integral thermal fuses and open safely when temperature in the device is greater than 240°C for 20s, protecting the pcb from melting or charring and reducing the risk of fire. Sustained overvoltage causes the resistive elements to crack and open, the fuses taking care of slow-burn conditions caused as above. Resistances are 50Ω and 20Ω respectively, matched to within ±0.5%. Ericsson Components AB. Tel., 01793 488300; fax, 01793 488301.

Switches and relays

Pcb switches. A new range of tactile, sealed switches for board mounting are 12.5mm square and allow keyboard layouts to be designed on a 12.7mm grid. All models in the Secme *Cosmos* range are proof against dust, splashing, wave soldering and cleaning solvent and are provided with removable covers in various shapes and colours. Silver or gold-plated dome contacts are used and the switches can have a red, green or yellow led. Ratings are 24V, 125mA, the gold ones switching down to 500µA. Insulation at 100V is over 1GΩ. Hawnt Electronics Ltd. Tel., 0121-784-3355; fax, 0121-783-1657.



Transducers and sensors

Digital thermometer. *DS1624* from Dallas is a direct-to-digital temperature sensor having 256byte of eeprom on-chip to store temperature-related compensation information, eliminating the need for an external eeprom or the use of a microcontroller's memory; no external components at all are needed. It is chiefly intended for use in crystal compensation, in which the eeprom contains data on the frequency/temperature curve of the crystal. Temperature information is in the form of a 13-bit, twos-complement word, equivalent temperature measurement being in the –55°C to 125°C in 0.03125° steps, which Dallas says is the smallest available. Supply is 2.7-5.5V and the device uses the two-wire bus with three-bit addressing for up to eight chips on the bus. Sensors come already calibrated. Dallas Semiconductor. Tel., 0121-7822959; fax, 0121-7822156.

Vision systems

Video conference kit. Premier's H.324-compliant kit consists of a PCI card, a colour camera and cable, the software providing facilities for video conferences, moving picture storage and frame grabbing, with set-up assistance for user details, e-mail address, resolution and picture quality. PAL, SECAM and NTSC formats are available and the system copes with communication between computers or connection to the internet over standard lines, ISDN or lan. The software allows storage of motion video direct to disk in an AVI file. Requirements are a 33MHz 486, 16Mbytes, 5Mbyte hard disk space and Windows '95. Premier Electronics Ltd. Tel., 01922 634652; fax, 01922 634616.

Data acquisition

Data acquisition pod for pcs. From Thurlby-Thandar, the *VIPS 10* is a software/hardware package consisting of an 8-channel analogue/digital converter pod that fits a pc's parallel port to form a multi-channel system for applications from low-speed data logging to fast, 12-bit data acquisition, up to four such pods working from one printer port, from which the pods take power. VIPS Windows software ('95 or 3.1) gives display and control for up to four pods and supports direct data exchange between VIPS and other Windows applications. There are five types of pod for input/output in analogue/digital form, operating in any combination. And there are more to come. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

COMPUTER

Computers

Hardy palmtop. Fully proofed against most of the hazards of this world, *Ultimax* palmtop computers are 386SX-based (*Ultimax 2000*) and 486DX2 66MHz-based (*Ultimax 4000*) and operate for long periods on four AA batteries; they are made by ST Research Corp. Both models are provided with two PCMCIA slots, memory upgrade from 4Mbyte to 16Mbyte in the 2000 and from 8Mbyte to 32Mbyte in the 4000, a backlit display and keyboard, which is a sealed membrane, tactile type. The *2000* is dos-based and has a CGA lcd, while the *4000* runs Windows or dos software, has two serial ports, one external configurable connector,

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

a parallel port and an external keyboard connector, with a full-colour VGA lcd. Workstation Source Ltd. Tel., 0118 9227888; fax, 0118 9227820.

Development and evaluation

78K/IV development. New from NEC, a software system simulator for the 16-bit, 32MHz *78K/IV* microcontroller family. The simulator is Windows-compatible and simulates all peripheral functions of family devices, including timers, interrupt controller and i/o ports. allowing the simulation of waveform output, pull-up/down resistors, leds, keypads and external interrupt sources. In program development, there is single-step operation with a trace and access to source-level debugging. Cost of the simulator is £85. Sunrise Electronics Ltd. Tel., 01908 263999; fax, 01908 263003.

PIC emulator. RF Solutions' new single-board, in-circuit emulator for PIC microcontrollers provides real-time, non-intrusive development to 20MHz and costs less than £300. *ICEPIC Junior 5X* rapidly identifies system bugs in all members of the Microchip PIC 12C5XX and 16C5X families. Source-level debugging is in either assembler or C, running in Windows; the system can therefore set an unlimited number of hardware breakpoints and arrange real-time emulation or single, multiple and procedure step execution. The emulator contains a Microchip emulation ic, 8K of emulation memory and an RS232 interface for data transfer at 115kb/s. RF Solutions Ltd. Tel., 01273 488880; fax, 01273 480661.

Mass storage systems

Rewritable disk/cd-rom drive. Panasonic's *LF-1196* is a removable, rewritable optical disk drive, combining an IDE ATAPI interface and Seagate software to provide 650Mbyte of storage on optical disk or to allow reading of 80-120mm cd-roms at up to 8x. There is no need to install an additional SCSI board, and Windows 3.1 and '96 versions of Seagate Backup Exec simplify backup and restore facilities by means of its easy Wizard interface for simplified backup and restore and an advanced interface for customised work. The drive is a half-height device, mounted vertically or horizontally, with seek times of 92ms for both optical disks and CD-roms, data transfer proceeding at 1141Kb/s and 1200Kb/s respectively. Panasonic Industrial (Europe) Ltd. Tel., 01344 853157; fax, 01344 853081

Software

WorkBench PC v.3.0. Strawberry Tree's *WorkBench PC for Windows* has undergone a major upgrading to produce version 3.0 of this graphical data acquisition package, which allows the collection, analysis, display, logging of data and the control of external systems, the only programming needed being the shuffling of icons. A choice of modules provides for the representation of data inputs and outputs, controls, displays and functions for data reduction, signal analysis, maths and statistical operations, the modules being selected from a pull-down menu and "connected" on screen. The new version has a cut-and-paste facility for the selection of modules by mouse for copying to other parts of the worksheet or to a macro box. An optional tool caters for up to 20 pages to be designed either in this program or to be imported as .BMP files and virtual instruments may be integrated into the design. Five new modules include a counter/timer, a formula interpreter, an RS232 output, trigger modules and a min/max module. Adept Scientific Micro Systems Ltd. Tel., 01462 480055; fax, 01462 480213.

Bar-code labels. *BAR-ONE Platinum v.4.0* from Zebra Technologies is the company's newest bar-code label design and print software. which allows users to access, process and combine variable information, including text and graphics, from many sources, this latest version having improved networking capability. It meets Microsoft's Open DataBase Connectivity standard and runs under Windows, so that a preview feature in the print queue allows a sight of the label before it is loaded for printing. Zebra Technologies Europe Ltd. Tel., 01494 472872; fax, 01494 450103. ■

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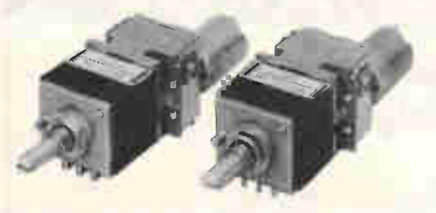

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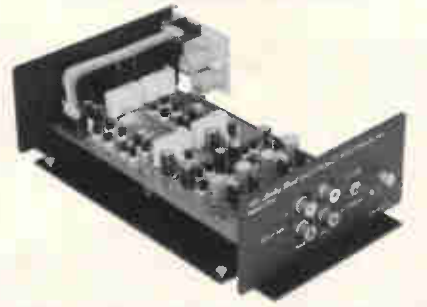

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A battery of chargers

Battery design being what it is, battery chargers are not the simple devices they used to be. They even talk to computers, as Philip Darrington explains.

Ever since people decided they needed to carry computers, tape players, radios, telephones and all those life-enhancing games around with them, and the parking-meter reader stopped writing the numbers in a little book, the battery makers have had an extended birthday. But the one-discharge type of battery came a bit expensive – and a bit too big and heavy – for some of the current-hungry devices, so rechargeable types were needed.

Cue for new chemistry: you can now be faced with nickel-cadmium, nickel-metal hydride, rechargeable alkaline and lithium ion batteries all in the same household and all needing chargers with various protective limiting facilities. Maxim makes ics for the purpose, described in the *Maxim Engineering Journal* vol.27, from which this is an extract.

NiCd *versus* NiMH

If you need a high current in short bursts, nickel-cadmium (NiCd) batteries are the choice; they have a lower capacity than lithium ion (Li+) or nickel-metal hydride (NiMH) types, but possess low impedance.

NiMH batteries are in some ways similar to NiCd ones, but have more capacity, although self-discharge rate is about double at 1% of capacity per day and they do not, therefore, hold a charge for long periods. Both types can be fast-charged in about an hour by a current equal to the capacity in amp-hours, although this process will not give a full charge because of losses; to get a full charge, charge for longer or with a higher current.

Charge termination is slightly different for the two types: charging a NiCd battery should stop when its terminal voltage begins to decrease, because the thing will probably explode if provoked for much longer; an NiMH charge should be stopped when the voltage peaks.

You can trickle-charge both types without worrying about stopping the charge or checking the voltage because temperature rise is not nearly enough to cause any trouble. Trickle current should be around (capacity in Ah)/15.

Lithium-ion batteries

Li+ batteries have the edge over other rechargeables as regards capacity. Compared with NiMH types, energy per unit volume is 10%-30% better, but energy per unit mass is about double, since they are lighter.

But they are not perfect; current and voltage both need watching in both charge and discharge regimes (if you keep discharging the battery to a low voltage, it loses capacity). For these reasons, Li+ packs usually have a fuse to prevent overcurrent and a switch to disconnect the battery if it looks like venting or, to put it another way, exploding.

Li+ battery packs usually use mosfets to disconnect the battery if it is being subjected to under or overvoltage. These mosfets also make a different charging method possible, in which a constant-current charge without voltage limit is applied, the mosfets turning on and off as needed to maintain the battery voltage. Battery capacitance slows

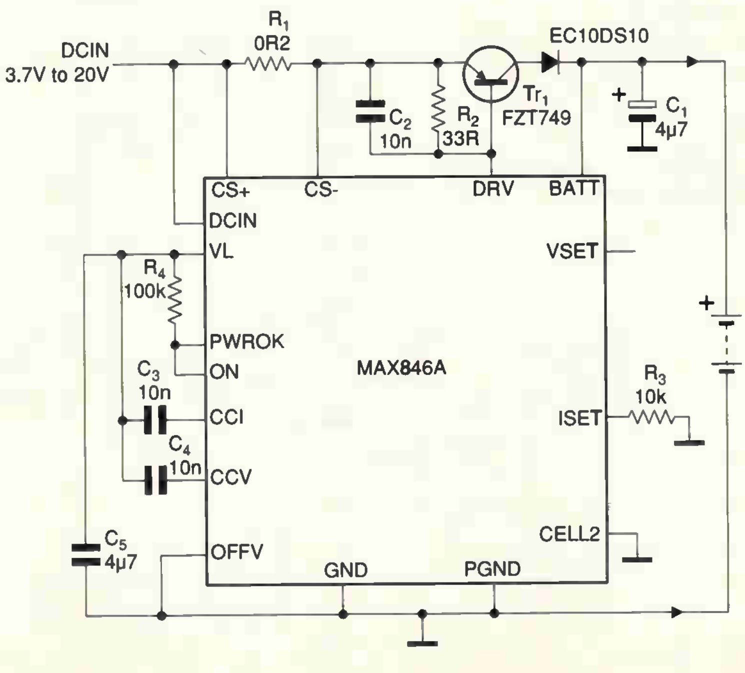


Fig. 1. Simple linear charger for one lithium ion cell. Heat generated by this type confines its use to the cradle type of separate charger.

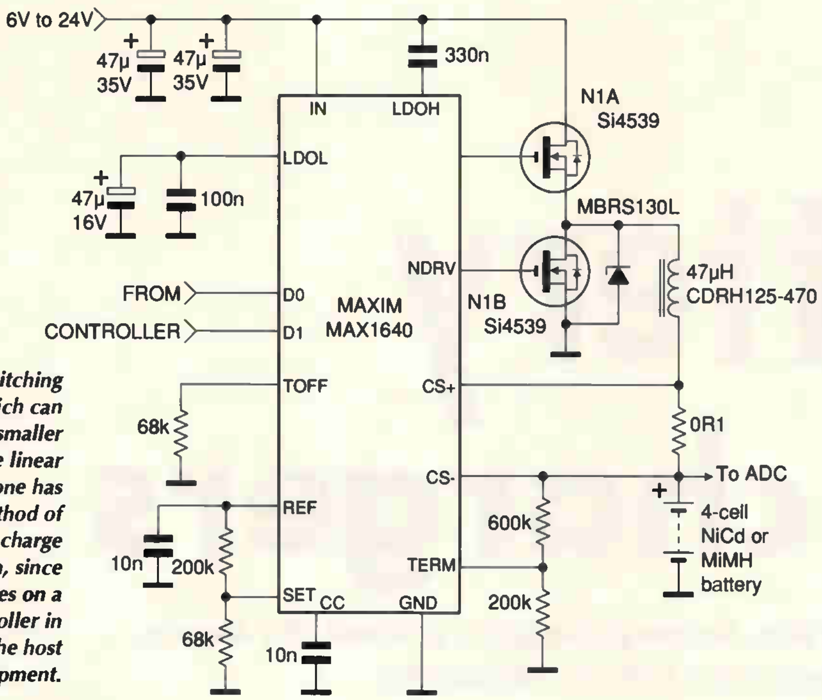
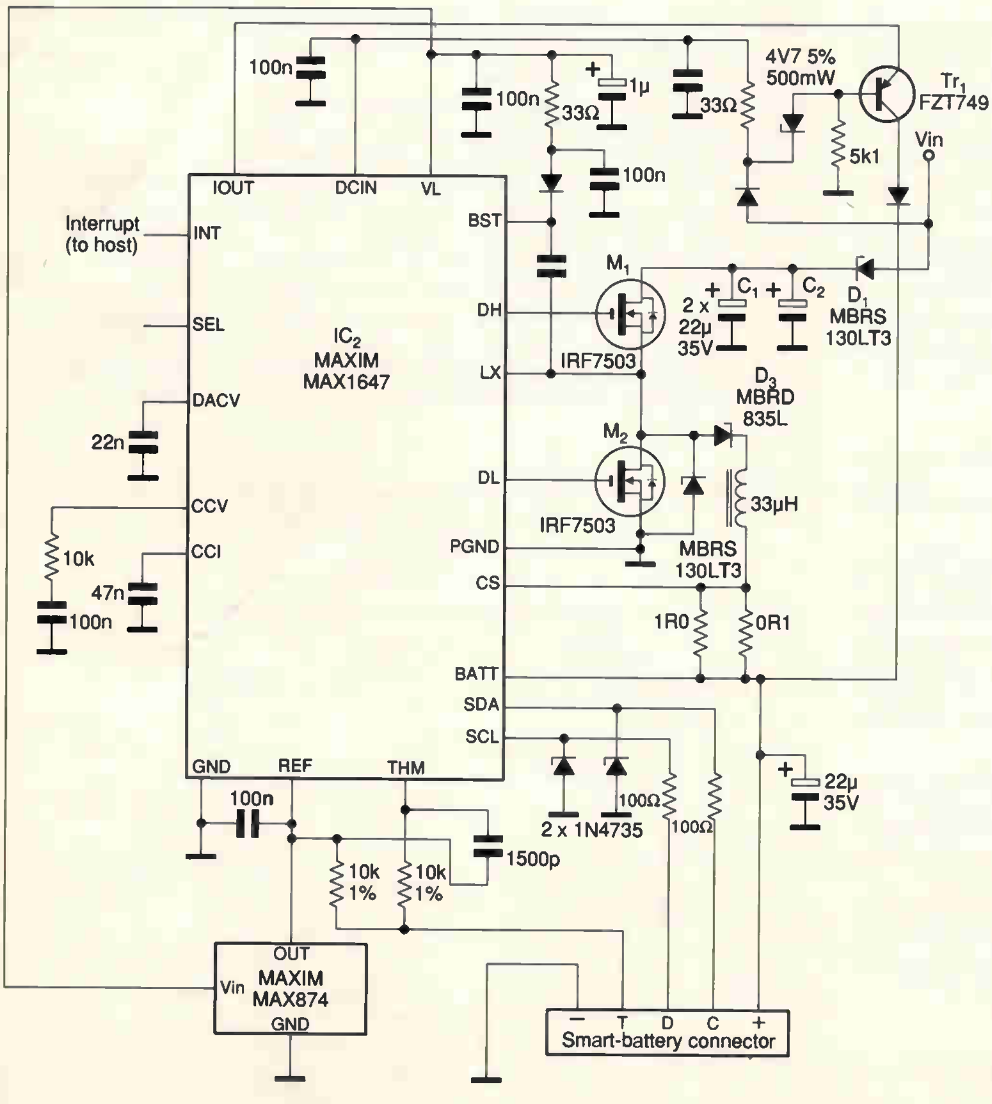


Fig. 2. Switching charger, which can be made smaller than the linear type. This one has no method of charge termination, since it relies on a microcontroller in the host equipment.

Fig. 3. Charger for smart battery packs, using the SMbus interface. This one is 89% efficient. The MAX874 is a voltage reference.



down the voltage rise, but care is needed because the capacitance varies with frequency and with battery type.

In some kinds of equipment powered by, say, alkaline AA batteries, an intermittent load calls for more power than the battery can supply, for example during transmission in a two-way pager. In such a case, the answer is to have an extra, rechargeable type to answer the call for action and then to relapse into relative inactivity while it is recharged by the main alkaline battery ready for the next assault on its composure.

Chargers

Separate chargers, including the cradle type into which the cellphone, for example, or its battery pack fits do not have to be extravagantly efficient, since any heat generated, within reason, will not affect the equipment.

Figure 1 shows the simplest type of charger for this purpose– a linear regulator designed to charge one LI+ cell at 1A. An external power transistor drops the input voltage to 2V, the transistor dissipating most of the power in the circuit, the *MAX846A* charger ic, with its internal reference, staying relatively cool and therefore stable. Series input current-sensing resistor R_1 and the ISET resistor R_3 set the current regulation level and ISET the current limit, in this case 1A. Terminal VSET allows for voltage limit adjustment.

If the series-pass transistor is contained in a small volume without cooling, heat might still be a problem. For four NiCd cells charging at 1A, total power dissipation could be 8W, 4.4W of that being in the transistor.

Built-in chargers are another matter; the life of a heated battery is shorter, but not necessarily sweeter. Linear regulators have to give way to switching regulators, which dissipate little power, regardless of the input/output difference, and which are smaller than the linear type. There is the disadvantage of the necessary filter, of course, although the battery capacitance can form part of the filter, if *C* is suitable for the filter frequency.

Noise can be a nuisance, but some switchers can be synchronised to an external signal to move the noise away from sensitive frequency bands. Good layout and shielding will avoid most of the trouble.

A four-cell NiCd/NiMH/ charger of the switching type is shown in **Fig. 2**. This the kind that works with a microcontroller to terminate the charge, having no provision for that in itself. If the equipment has no such controller, it only needs to be an inexpensive type with an a-to-d converter. The *MAX1640* chops the input by means of the switching transistor N_{1A} and synchronous rectifier N_{1B}, the chopped waveform being imposed across the inductor to form a current source, the diode D_2 preventing the battery reversing the process and supplying current to the source. In response to controller signals on pins D0 and D1, there are three modes of working: fast charge, pulse-trickle charge and top-off charge. Fast charging produces a current of 1.5A, in top-off the current is about a quarter of the fast-charge current, or 381mA, and in trickle charge it is the same but pulsed with a 12.5% duty cycle. All these current levels are produced by the voltages on SET and REF pins and the value of the current-sense resistor, in this case 0.1Ω, all selected by the microcontroller input signals.

If everything goes according to plan, the circuit stops the charge when the battery voltage rise becomes zero or negative, depending on whether the battery is an NiMH or NiCd type. If it doesn't, the potential divider $R_{6,7}$ are there to set a limit to which the battery voltage may rise, in this case 8V.

Smart-battery chargers

One has the thought that battery design might perhaps be getting a little out of hand when even battery packs have microcontrollers. Nevertheless, they do and are useful in that one type of charger is able to cope with any kind of battery that conforms to the smart-battery standard.

You can also replace a battery with any other that conforms to the standard, which is concerned with the manner in which the battery pack connects to the equipment it powers and the way in which it communicates with it using the Intel SMBus, itself derived from the I²C protocol. There are many I²C-compliant ics around already. **Figure 3** shows a charger with an SMBus interface.

Currents under 31mA, which corresponds to the five least significant bits from the a-to-d converter in the host controller, come from an internal linear current source, since the switching regulator and its low-value current-sense resistor cannot handle the 1mA resolution needed. Currents over 31mA are provided by a switching regulator to maintain an efficiency of 89%, but the linear source remains active so that monotonicity at the a-to-d doesn't suffer whatever the value of the sense resistor or current-sense amplifier offset.

If the input voltage is much greater than the battery voltage, transistor Q_1 helps to relieve the power dissipation in the internal linear regulator; inputs up to 28V are acceptable and outputs are selectable at 1A, 2A and 4A. Switching is at 250kHz, the size of the inductor reflecting the fact. ■

Technical support

Maxim, Unit 3, Theale Road, Theale, Berkshire RG7 4XX. Tel., 01734 303388; fax, 01734 305577.

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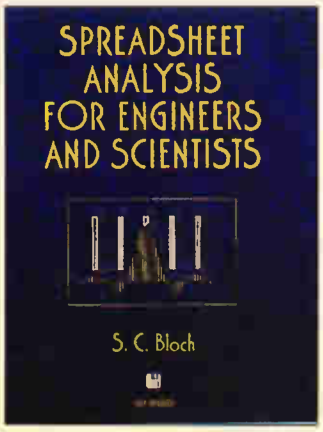

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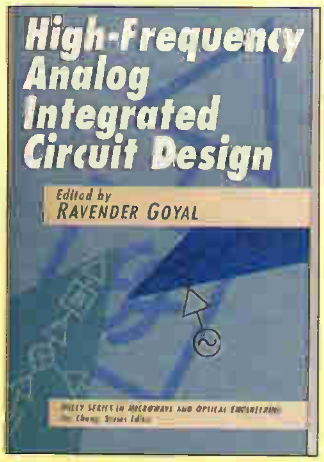

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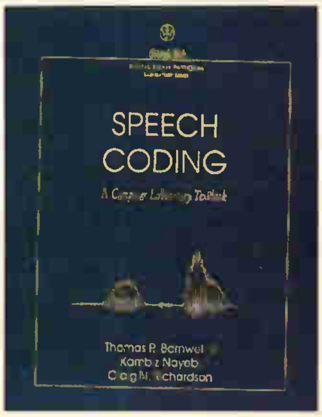

perform a wide variety of speech coding and speech processing experiments. The text presents and explains a set of basic speech coders analytically and in terms of the specific parameters controlling each coder. The manual leads the student through the experimental process of understanding how speech 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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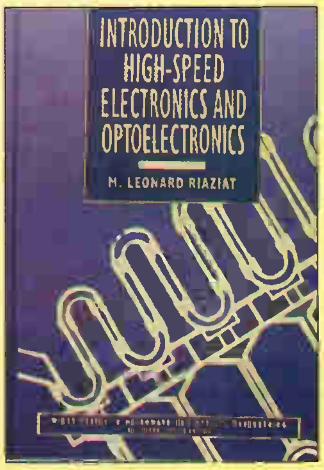

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LETTERS

Letters to "Electronics World" Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

Why Wien?

I was interested to read Bryan Hart's article on the design of Wien bridge oscillators in October issue of *Electronics World*. But why would anyone nowadays want to build a Wien bridge oscillator? I find a double-integrator oscillator – such as described on pages 53-57 of the 'Analog Circuits Cookbook' by Ian Hickman – much more reliable and predictable. Moreover, its effective filtering action makes it possible to use diode limiting in the feedback path for level control, and still obtain a low level of harmonics.

True, this alternative requires two more op-amps, but as you can buy four high quality op-amps in one package for about £1 retail, it is a price well worth paying.

J S Linfoot
Oxford

Flat-panel speakers

The article on solid-diaphragm speakers, 'Flat but not flat' in the August issue, rather bypasses the matter of woofers with polystyrene diaphragms of maximum strength.

Two or three decades ago you published two remarkable articles I cannot trace, by the same author. One showed that acoustic gramophone records could give a useful audio power at reasonable fidelity, provided the grooves were cut not with a diamond profile, but shaped with a solid of revolution so as to give line contact with a suitably shaped stylus.

The other showed that a very rigid diaphragm for a loudspeaker could be made from polystyrene foam with stretched aluminium foil glued to each side, and possibly because of this the Kefkit 3 I used to listen to mono on a woofer of solid polystyrene – without the foil. It has the advantage of zero colouration. It does not sound like a loudspeaker.

I see a 1949 book 'The Why and How of Good Reproduction' By G H Briggs is still on offer from one of your advertisers and wonder if the people who worry about cabinet resonances would not be happier following his suggestion for students of using a drainpipe on a cabinet with a speaker set on top of it and locks spacing it off the floor; in such configurations one can set an inverted cone over the centre of the speaker to disperse the higher frequencies and of course add a tweeter, though I think he was probably using Wharfdale speaker (his own firm) intended for all frequencies. To make more of a port at the bottom a rectangular-section port could be built with blocks taking its plan view up to the size of the pipe. Alternatively owners of rectangular speaker cabinets could glue in large numbers of light wooden spars going in all directions but favouring the centres of the panels, damping the spars with felt.

With regard to the letters on speaker design suggesting that speaker cabinets can be tuned to match the room, I very much doubt the room affects what happens at the speaker front panel, and also I've been told that in judging levels people use only the direct speaker sound. On the other hand Cecil B Watts used to say that solo instruments should be recorded without reverberation as the reverberation in the listening room did the trick. Who is right?

Sometimes I wish we could stop worrying about speakers, but we don't.

Name not available
London

The attraction of reed relays

Specifications of a reed-relay can be made to change drastically with the aid of a permanent magnet. With increasing magnetising influence, depending on the magnet's position relative to the axis of the relay and its 'strength', the permanent magnet can,

- make the relay pull in and release at smaller currents than without the permanent magnet (normally open),
- make the pull-in current still smaller, latching and opening only if the current is reversed (memory), and
- make the relay normally closed, opening only with the current reversed. And closing again at further increase.

The use of the permanent magnet affects both speed and reliability and increase the pull in to release current ratio. Less than 1cm of the magnetic PVC strip that kept the door of your old refrigerator closed will do the job. Just cut it with scissors and glue it to the top of your reed-relay.

Scott Arnesen
Oslo
Norway

Stability in audio amplifiers

Electronic stability. Audio amplifiers must remain electronically stable during normal – and sometimes abnormal – operation, if only to avoid writing off output transistors and possibly speakers.

A solution adopted by many designers including Doug Self,[1] is to connect a capacitor of say 100pF across the voltage amplifying stage (VAS). Ed Cherry recommends that this capacitor be connected to include the output devices so as to gain a useful reduction in crossover distortion. There is perhaps understandably some reluctance to try this, but I can concur with Ed that it can indeed work and result in stable amplifiers.

I discovered this after reading Ed's comment about instability in driver stages. Firstly, I connected 39pF capacitors between base and collector of the *BD139/BD140* drivers of my type II emitter follower output stage. A clearly discernible improvement in sound quality was noted.

I do mount my drivers on short leads, but clearly the pcb track inductance is sufficient to cause instability in this stage when using transistors having frequency responses up to100MHz or so. The output devices are *BD911/BD912*. Many thanks Ed.

The C_{dom} capacitor of 100pF was connected between the output stage and the input base of the VAS stage. A 1kHz square wave response showed only a small amount of overshoot with a reactive load (8Ω in parallel with 2µF). Response was optimised by means of a 22pF capacitor in series with 150kΩ, connected between VAS collector and the negative feedback point on the input pair.

Final response and stability are good. Audible results are very satisfactory. Again, thanks Ed.

The amplifier circuit includes a cascode stage and three transistor current mirror on the input pair, and a cascode VAS stage. Drivers and output devices are operated on ±35V rails. Input and VAS stages are fed from similar, but voltage stabilised rails. Input impedance of the amplifiers are around 35 kΩ and voltage gain is 20.

I always check and do initial setting up of amplifiers with drivers and outputs stages supplied via ±15V 1 A current limited supplies – just in case

Thermal stability. It can be difficult to ensure that optimum quiescent current is maintained during operation of class B emitter follower driver/output stages. Common practice is to mount drivers and output devices on a common heat sink and to use a single – or perhaps double – transistor V_{be} multiplier, also mounted on the heat sink – or possibly on one of the output devices. This is far from ideal as the drivers and output devices invariably operate at different temperatures.

A better solution is to use a conventional V_{be} multiplier connected in series with two forward biased diodes. The V_{be} multiplier transistor should be mounted as closely as possible to one of the output devices and the diodes should be mounted in contact with each of the driver devices. This arrangement has provided me with very good thermal stability in type II emitter follower circuits. An advantage of the arrangement is that driver devices can be located more conveniently, possibly away from the main cooling fin.

I claim no credit for the use of diodes to control quiescent current. This was done by B J Codd[3].

Ken Hough
Amersham,
South Bucks

1. Self, D, 'Distortion in Power Amplifiers,' *Electronics World*, Feb. 1994 p137.
2. Cherry, E., 'Ironing out distortion,' *Electronics World*, July 1997, p577.
3. Codd, BJ, 'Low distortion amplification,' *Wireless World*, Oct.1979, p81.

1.3V mystery

I have been an electronics engineer for 30 years and I have just realise that I do not know how a quartz watch can work so efficiently from only one 1.3V cell. The same goes for calculators and thermometers with liquid-crystal displays.

So what is the secret and why don't many other things use the same technology? One single cell is so convenient and contains size for size, much more energy than say a PP3.

You never see a circuit diagram of one of these single cell devices. Is there a conspiracy by the battery makers? Is there a parallel universe that uses a semiconductor that only needs a bias voltage of a millivolt?

While we are on the subject of the unknown, how do hearing aid makers squash all the electronics into such a small space, and without feedback? Do they use normal smt or again is there another Flint in this parallel universe that supplies $^1/_{100}$ size components?

Ken Hawes
London SW17

On beating frequencies

In the September 1997 issue on page 786, Chris Bulman raised a question about beating frequencies. This query prompted an unusually high response. These are the last few responses that we are able to publish on the subject, but many thanks to everyone who replied. Ed

Chris Bulman is correct in thinking that a sinewave signal very close to the Nyquist frequency will, when sampled, produce a set of data points which encode a "wuhwuhwuhwuh" (sic).

Fourier analysis of the wuh (forgive the abbreviation) would show in his example at 19.999kHz and one at 20.001kHz. There are also pairs of tones at 59.999k/60.001k and all subsequent odd multiples of the Nyquist frequency, with gradually reducing amplitude.

If you fed the data to a d-to-a converter and listened to the output, I'm sure those with a reliable aural response to 20kHz would hear this pulsating effect.

But this is not what happens in a proper digital audio system. A proper system uses an anti-imaging filter whose primary role is to remove the higher of the wuh tones, leaving the lower one pristine, smooth and un-wuhed.

This filter would indeed have a delay of many seconds if it had to resolve the 2Hz frequency difference between the wuh tones, which is why a 'guard band' is used between the highest encoded frequency and the Nyquist frequency. It makes the anti-imaging – as well as the anti-aliasing – filters economically realisable, whether they are 'traditional' analogue or computational exercises in silicon.

***Kendall Castor-Perry**, Kemo Ltd, Kent*

Chris Bulman asks about the behaviour of a digital sampling system, and the effect on it of signals at or close to the Nyquist frequency – i.e. half the sampling frequency.

In his example, Chris Bulman explains that a signal at 19.999kHz passed through a system that is sampling at 40kHz, and has a brick-wall filter that cuts off at 20kHz, will result in the d-to-a converter producing a 20kHz signal that is modulated at 1Hz, reversing in phase every half cycle. He believes his analysis of the signal, but does not believe that this is what really happens, and would like a non-mathematical explanation of the paradox.

A simple argument is to observe that Chris's waveform is a 20kHz signal modulated at 1Hz. Since modulation produces sidebands, this signal is impossible, as the brick wall filter removes sidebands above 20kHz.

A more complete answer to the paradox (but unfortunately requiring a small amount of maths) is as follows:

Chris's description of the signal is in fact a correct one if the brick wall filter is ignored. Ignoring the brick wall filter has no effect on Chris's argument, except that signals appear around a series of harmonics of the 40kHz sampling square wave.

The signal as described, viz. a 20kHz signal modulated at 1Hz, reversing in phase every half cycle, is actually a suppressed-cartier, double-sideband signal. In this case, the sidebands are 1Hz either side of the suppressed 20kHz cartier at 19.999kHz and 20.001Hz.

A way of understanding this is to look at it the other way round: there are two sidebands, and they are beating together. To find out what happens when two sinusoidal signals are added together, consult an A-level maths book, where you will find the identity,

$$\cos A + \cos B = 2\left[\cos\left(\frac{A+B}{2}\right)\cos\left(\frac{A-B}{2}\right)\right]$$

which is also known as 'two cos semi-sum cos semi-diff.'

This formula shows that sinusoidal waves of 19.999kHz and 20.001kHz will beat together to give a result that is a 20kHz sinusoidal wave (the cos(A+B) term) multiplied by a 1Hz sinusoidal wave (the cos(A-B) term), which is the waveform originally described by Chris.

Once you realise that C.B.'s fluctuating signal is actually a pair of frequencies, then all you have to do is put the brick-wall filter in, and, since it removes everything above 20kHz, what is left is 19.999kHz, as per Nyquist's theory. The 20.001kHz signal that has been removed is known as an alias signal.

Incidentally, a true brick wall filter is impossible to realise. The more brick-wall-like a filter is, the more delay the filter have, and an infinitely steep brick-wall response would require an infinite delay. However, replacing the brick-wall filter with a real filter, and leaving a guard band between the highest signal frequency and the Nyquist frequency, as is done in practice, does not in any way alter the paradox as described by Chris, and its explanation as given above, since the delay in the filter – ideally – affects all frequencies by the same amount.

Brian Pollard
Maidstone
Kent

The 'paradox' is easily resolved if one bears in mind what the spectrum – or Fourier transform – of a sampled signal looks like, and the practicalities of designing an appropriate reconstruction filter.

Shannon's sampling theorem[1] states that if a signal is bandlimited to WHz then it may be reconstructed from its values taken at $1/2W$ (s) intervals.

Consider the process of sampling and reconstruction in the time and frequency domains. In the frequency domain the act of sampling causes the spectrum to be made periodic, in a way that is best represented by the diagram, and the reconstruction process consists in filtering out the copies. If the copies overlap the resulting reconstruction is not the original signal and an aliasing error has occurred. In the time domain the reconstruction involves a convolution – an averaging process in which a 'window' is slid along the data and the weighted average of the samples in the window is computed.

The form of the weights determines the frequency characteristics of the filter. To pass frequency f_1Hz and reject f_2Hz requires a window of at least $1/|f_1-f_2|$ seconds in duration: over a shorter period, $\cos(2\pi f_1 t)$ and $\cos(2\pi f_2 t)$ are not significantly different.

The Shannon reconstruction filter has a vertical cut-off at W, and therefore requires a window of infinite time extent.

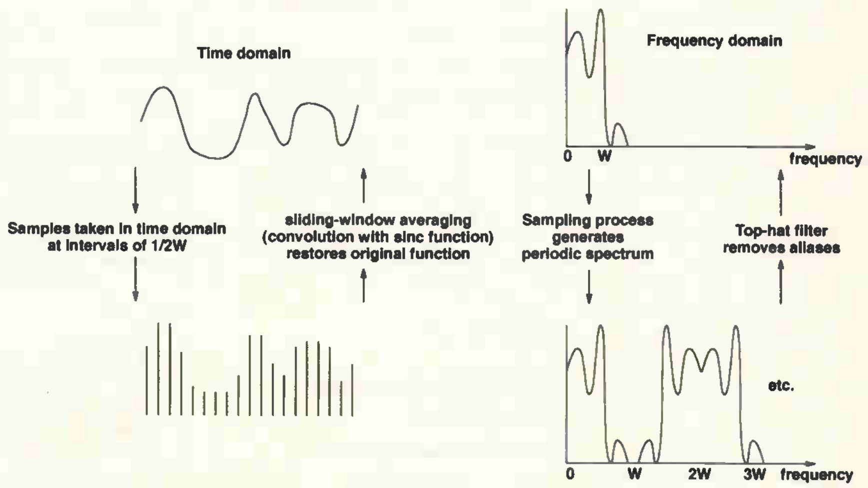
Considering Mr Bulman's example, when a 19.999kHz sinewave is sampled at 40kHz, the spectrum of the samples contains components at 19.999kHz, 20.001kHz, 59.999kHz, 60.001kHz, and so on. Exact reconstruction requires a filter that passes all frequencies up to 20kHz and rejects higher ones: in particular, it must pass 19.999kHz and reject 20.001kHz.

Can this be done in practice? In practice the window length is restricted to, perhaps, 100ms. This gives a selectivity of 0.01kHz, and so if 19.999kHz is to be passed then 20.001kHz will come through as well.

Having established that a practical reconstruction contains both the desired sinusoid and its first alias, it is easy to see what happens: the two frequencies beat together in the way that Mr Bulman describes. Note that the beat frequency does not actually occur in the spectrum – though if the signal were nonlinearly distorted it would. The problem will not arise if the difference between the signal and Nyquist frequencies is reasonably large – perhaps 1kHz – because then the signal's aliases can be removed very effectively.

It is therefore a good idea, once the sampling rate has been selected as 2W, to remove all frequencies above rW (for some r less than 1) before sampling. This creates a guard band between the highest frequency present, rW, and the Nyquist frequency, W, and makes accurate reconstruction a much more practical proposition. Helms and Thomas were the first to discuss this in a classic paper[2] on rapidly-convergent sampling expansions; since the sampling expansion gives the filter weights, their work is of immediate relevance in the design of reconstruction filters.

Richard Martin
GEC Marconi Research



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Peter Starling - Ref: PS190EWo

DSP Engineers

Continuing European growth has led to the immediate requirement for a couple of good allround DSP Engineers to join this small interactive team. You will be responsible for Hardware and Software Design (algorithm design, coding and testing) used in the development of new GSM and PMR comms products. Experience, allied to a degree level qualification should include real-time embedded software experience in C/C++ and Assembler with solid exposure to Texas DSPs.

Frazer Martin - Ref: FM/EWo70

Software Test

Testing times indeed! We can't find enough good experienced Software Test Engineers, therefore we have numerous vacancies from Junior to Team Leader level with clients in Telecomms, Datacomms and Consulting industries. Their requirements are similar, in that they all seek qualified, well trained Engineers to take responsibility for this critical production area, whether you have just one years experience or whether you develop test strategy, enhancements, procedures and seek a move into Consultancy - we have probably got a good move for you.

Frazer Martin - Ref: FM/EWo753

Fraud/Security Systems

As part of this major network provider/operator you will be responsible for the design/ integration of security and fraud systems into the existing comms infrastructure. You should possess a good knowledge of systems and protocols, encryption and authentication techniques. Additional exposure to GSM or RF would be advantageous.

Frazer Martin - Ref: FM/EWo40

Software Tools/Billing/CTI

The UK's most popular Mobile Telecomms company require talented qualified Engineers to join them in developing software tools for their next generation GSM cellular systems, CTI, business critical software, designing subscriber admin/billing systems software or simply planning/optimisation tools for their new and current networks. You will have a good Degree plus experience of development in C, C++, OOA/D, RDBMS.

Frazer Martin - Ref: FM/EWo63

Development Manager

Our client is a leading blue chip company who require an experienced Development Manager. Based in Hampshire you will be developing a wide range of satellite based (GPS) positioning and using a range of communication systems including RF tagging products. You will also be involved in taking the designs through the complete product lifecycle and have some contact with customers and supplier. You will be qualified to Degree level in Electronics and/or will have a background in analogue and digital design and some knowledge of embedded microcontroller firmware design, a good allround package is offered to the ideal candidate.

David Tillyer - Ref: DT115/EWo

Midlands

£16-28K

Various companies in the East and West Midlands have a number of openings for Hardware Engineers with digital and analogue skills in the following fields:

- ★ Medical
- ★ Instrumentation
- ★ Power
- ★ Automotive
- ★ Avionics
- ★ ASIC/IC Design

Roma Das Gupta - Ref: RDG29EWo

Cash

World leader in the provision of transaction products and services based in the South of England are looking for Digital and Analogue Electronic Engineers at all levels. Digital position requires electronic design experience with 16/32 bit embedded microcontrollers, analogue position requires experience with low noise circuits, analogue circuit simulation and some knowledge of sensors. All positions require a good Degree and knowledge of EMC with software experience in assembler and C. Excellent company with benefits to match

Roma Das Gupta - Ref: RDG20EWo

Design Engineer

East **£18-27K**

A leader in the field of medical instrumentation is looking for a Hardware Design Engineer to develop new ideas. The company who are based in the East of England have just received a SMART grant to produce state-of-the-art medical equipment. The successful candidate must be prepared to work in a hands-on role, individually and as part of a team. We'd be keen to talk to Engineers with experience of digital and/or analogue hardware in addition instrumentation/laboratory equipment experience would be an advantage.

Roma Das Gupta - Ref: RDG25EW

Please call quoting the appropriate Ref. No.on 01442 212555 or fax/email/post your CV to:

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2nd Floor,
89 The Marlowes,
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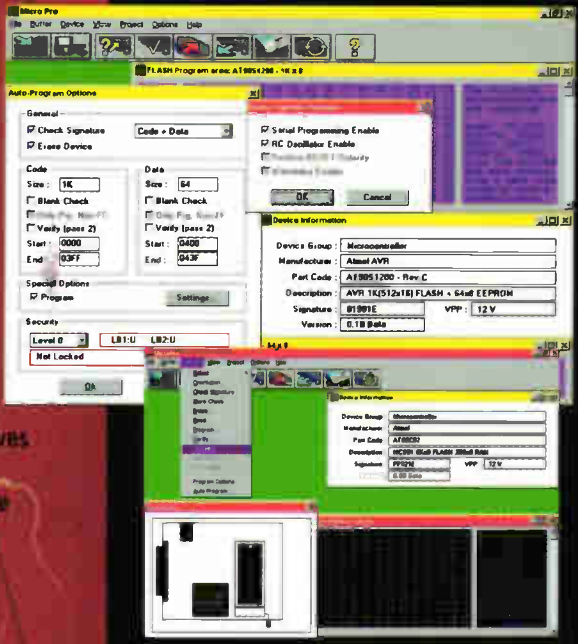

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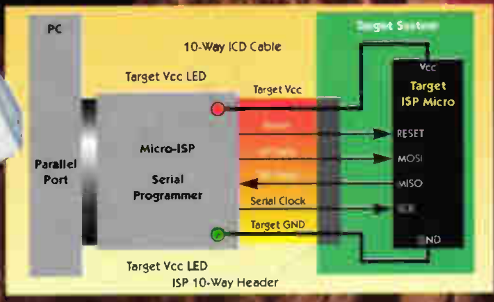

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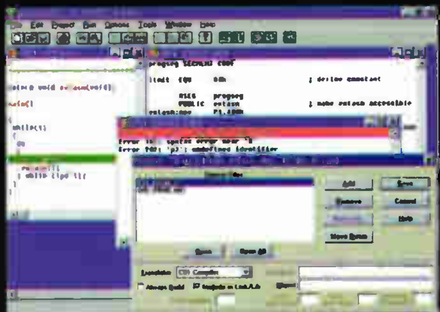

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