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The electron's centenary

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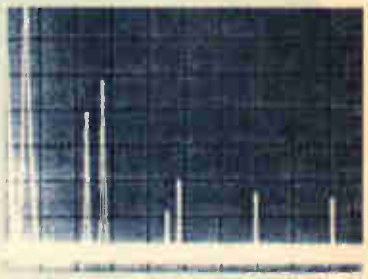

Interested in reducing phase noise in rf oscillators? See page 642.

Exactly a century ago, this man broke the myth that the atom was the smallest particle – see page 682.

A team of Australian researchers has made a biosensor so sensitive it could find uses in everything from pesticide monitoring to drug testing, see page 626.

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Digital deal spells danger

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NEWSTRADE ENQUIRIES
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The new Grand Alliance of BT, BSkyB, Matsushita and others to introduce interactive digital services sounds extremely impressive. All the same, big budgets, major players and grand plans alone do not guarantee success, and British Interactive Broacasting's launch plans show all the signs of a fiasco in the making.

For proof that big names can make big mistakes one needs only to recall the brief and poignant British Satellite Broadcasting episode. It's smart to be square – remember? Those with even longer memories may recall the ill-fated and equally short-lived Unisat alliance of British Aerospace and British Telecom, which also proved that flawed schemes have no God-given right to success.

While the plans of the new British Interactive Broadcasting (BIB) venture display an undoubted technical elegance, there is less evidence of guaranteed market success, which disturbs some observers. The saving grace is that the investors will not be hit too hard if their eggs don't hatch as hoped.

This may of course sound like heresy. BT and BSkyB are both shrewd and successful operators in their own field, and both bring into the alliance valuable talents and expertise. BT has already expended a rumoured £100 million – significantly more than its stake in BIB – in trialling multimedia and video-on-demand in East Anglia last year. And the company now possesses market insights that are both unique and of great commercial value. BSkyB for its part has displayed consummate skill in packaging and delivering entertainment programming with a mass-market appeal.

Of all the potential delivery mechanisms for digital interactive services, the chosen marriage of satellite delivery and telephone line return path is well-chosen and the sole problem (sorry, opportunity) is to judge whether there is a market. By sharing the financial risk, this can now be put into practice.

The risk is high nonetheless. On a technical plane it will undoubtedly work but as a mechanism for kick-starting interactive services it will probably disappoint its investors. BT's trials have demonstrated that there is a market for new interactive services – but not at any cost. So long as punters can rent video tapes for £1.50 a night from the corner shop and call by their high street bank branch, they are unlikely to pay a premium for alternative solutions.

Moreover, most new technologies take off when consumers feel comfortable with them, not when they are first introduced. Remember the slow initial take-up of compact discs and mobile phones? Only when the shock of the new wears off and prices fall to consumer levels does the critical mass appear.

It was smart to be square. Will it be intelligent to be interactive?

The optimism of the current protagonists reflects that of the UK government's enthusiasm for nation-wide cable television back in 1983. Kenneth Baker mistakenly thought he could fund a new communications infrastructure on the appeal of multi-channel entertainment alone. It didn't work then and it won't work now – people are not manipulated like that.

To take advantage of the new interactive services, it is said subscribers must sign up for the complete entertainment package as well, which displays a fundamental misunderstanding of the target audience. There must indeed be some people to whom 200 channels of entertainment will appeal but statistics show the numerical majority of UK households are perfectly content with the basic terrestrial channels.

Its equally obvious that the sort of people who would make significant use of interactive services – and could afford to use them – are not the sort of people who enjoy wall-to-wall low-grade films and television. The Department of Employment says it will post job advertisements on the new interactive medium – but can the unemployed afford the high price of this premium package?

In short, the introduction of interactive services combined with wall-to-wall television is a laudable, perhaps even desirable, notion and the BIB delivery mechanism is the most elegant yet put forward. Too bad that this may be compromised by a launch package which displays a fundamental misunderstanding of human nature and market economics. ■

Andrew Emmerson

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DAB products set for release

The first consumer units for the Digital Audio Broadcasting (DAB) standard could become available as early as this year.

Set makers are already developing in-car and portable radio sets that can be connected to cellular phones to provide return paths for interactive DAB services. These can also be linked to in-car navigation units. Home pcs will gain cards that will be able to receive DAB services and even receive DAB information overnight and store it for later listening/viewing.

"The possibilities are limitless," said Dominic Riley, DAB marketing manager at the BBC.

Over ten set makers are expected to announce the first consumer DAB radios at the IFA show in Berlin this year. "We hope to launch music and multimedia DAB services but the dates for these launches depend on when the sets appear in the shops," said Riley.

Solar energy will be viable alternative

Solar energy will be a viable alternative to oil and gas within ten years, according to BP's chief executive, John Browne.

Browne, whose company develops and manufactures solar power products, insists it will be a complementary energy source. "I am convinced we can make solar competitive in supplying peak electricity demand within ten years," he said.

The company hopes to expand BP Solar, its division which holds ten per cent of the world's solar market, to achieve annual sales of $1bn over the next decade.

Mark Hammonds, of BP Solar's UK research facility, explained that while photovoltaic cells, from which solar devices are made, are essentially silicon p-n junctions, they are relatively expensive.

BP Solar is working on several approaches to improve the efficiencies of solar cells, and hence their cost.

One is the use of a laser to cut a grid into the silicon before it is plated. Burying the grid deep within the silicon increases the number of electrons held, thereby raising the cell's efficiency. The resulting proportion of sunlight converted to electricity is 17 per cent. This compares to the typically achieved efficiencies of 12 per cent.

Another approach involves the use of lenses to concentrate the light hitting the cells. "By focusing light you get more energy per unit surface area," said Hammonds.

Cost can be further reduced by laying a thin film of cells, made from cadmium telluride, onto a steel substrate.

BP Solar is commissioning a factory in California, adding to its existing manufacturing facilities around the world.

Japan focusses on pdp displays

The cream of corporate high-tech Japan is gearing up for an assault on a product area where it has a marked technology lead - gas plasma displays (pdps).

Matsushita, NEC, Hitachi, Mitsubishi and Pioneer are all preparing investment plans to build a capability in pdps that will match, or overtake, pdp market leader Fujitsu.

"We are making 5000 42in panels a month," said Takamitsu Tsuchimoto, chief executive officer of Fujitsu's Electronic Devices Division.

In April, Mitsubishi Electric started production of a factory capable of 6000 42in panels a month. Matsushita has a production capacity of 1000 to 2000 42in panels a month and will start selling tvs using pdps later this month.

NEC is already selling 42in pdp-based tvs – costing $12 000 – and can make 1000 panels a month. NEC is building a production plant in Kagoshima on Kyusku Island which will start producing 10 000 panels a month in spring 1998.

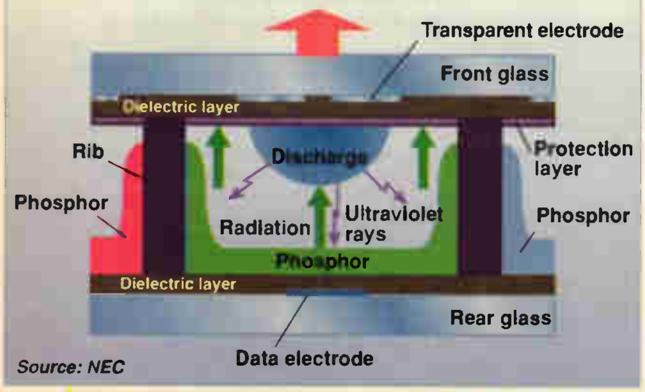
Pioneer has a 1000 to 2000 unit-a-month pdp production line, while Hitachi starts selling a smaller, 25in pdp display later this year, and is currently developing 42in panels.

According to Michihiro Ohta, general manager of NEC's pdp division, the company's target is to reduce the cost of a 42in pdp-based tv to $4200 by the year 2000 – "$100 per inch", he says.

Panel yields are running at 50 per cent, said Ohta, and with the improvements being made he expects 80 per cent yields by the year end.

The cost of making the panels is half the cost of making the assembled pdp module including the electronics to work the panel, said Ohta. And the cost of the module represents one third to one half of the final production cost of a pdp-based television.

Colour plasma display panel principle of operation



Source: NEC

Bright future... When addressed, a plasma is formed within a display cell which emits ultraviolet light. The light, in turn, excites the phosphor within the cell to emit its native visual colour.

Why aren't stolen phones barred?

UK mobile phone operators are not barring stolen mobile phones from their networks despite having the technology to do so.

Although operators can access every phone's IMEI (international mobile equipment identity) number – a supposedly unique identification code – they will not use it to remove stolen phones from networks for fear of switching off legitimate owners.

A Cellnet spokesman said that criminals can use a phone's IMEI number in hundreds of others before selling them on to unsuspecting customers.

"In Scandinavia there was a scam where phones were sold using the same number," he said. "One customer, who'd bought a phone in good faith, had it stolen. The operator switched it off, along with hundreds of others which had the same number."

An approach adopted in Australia offers a more promising solution. This is based on compiling a register of IMEI numbers to help police track geographical concentrations of phones using the same number. Police use the resulting intelligence to launch raids on criminals.

Cellnet, along with other operators, is currently compiling an IMEI number register for PCN and GSM phones.

Telemetry device for body signs

Oak Ridge National Laboratory (ORNL) in Tennessee is developing 'medical telesensors', chips that measure vital body signs, before transmitting the data to a remote receiver.

Parameters like temperature, blood pressure, oxygen level and pulse rate will be covered in the military funded project. So far a group has built a temperature sensing IC that can be attached to a finger or placed in an ear.

Non-military applications are also foreseen. "Wireless monitors attached to the skin," said an ORNL spokesman, "could provide valuable information on the physiological condition of police and firefighters for example."

Fastest JPEG codec IC?

Pixel Magic says it has the fastest JPEG codec chip available, able to process image data at up to 70Mbyte/s. The *PM-35* is being targeted at such uses as digital copiers, printers, scanners, fax machines and digital cameras. Although the market for JPEG codec chips is currently small, Pixel Magic is betting on increased demand as colour image processing takes off in mainstream applications.

GPT to create 400 jobs

GPT, manufacturer of telecommunications equipment, is looking to recruit up to 400 staff.

GPT is seeking experienced engineers and recent graduates for positions including system designers, test analysts, system integration and proving engineers, Asic, hardware and software designers.

"GPT is looking to recruit high calibre graduates and engineers. Successful individuals can expect very competitive salaries and benefits packages and an opportunity to make a difference in the industry," said Tony Lewis, GPT Public Networks Group's recruitment co-ordination manager.

The recruitment programme follows the success of GPT's synchronous digital hierarchy products. The vacancies are at facilities in Coventry, Liverpool, Poole and Beeston.

56k modem standard is uncertain

Both Rockwell's K56Flex and US Robotic's x2 modems are highly unlikely to be adopted for the 56kbit/s international modem standard. "It's very unlikely we would choose either of them. In fact, I can almost guarantee we won't," said John Magill, of the International Telecommunication Union. Instead, a hybrid of the two modems is expected when the ITU's Expert Group delivers its recommendation in the autumn. US Robotics guarantees x2 users a free upgrade to the eventual standard, while Tornado, a UK manufacturer of K56Flex, is offering a similar service.

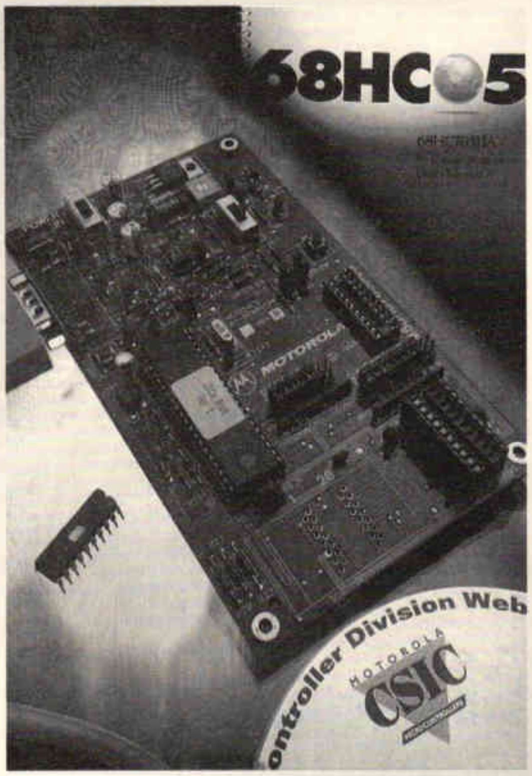
Motorway tolling trials due for review

GEC-Marconi's motorway tolling system, based on 5.8GHz microwave technology, is still being tested at the Transport Research Laboratory's test track. The trial is due to finish within weeks, after which a report will be submitted to Labour's transport minister.

The Department of Transport (DoT) could not comment on the future of motorway tolling. "We honestly don't know. We have to ask the new ministers," said a DoT spokesperson.

FEI speaks out on spectrum pricing

The pricing of the radio spectrum must be managed effectively to provide improved availability with fair pricing to users. So argues the Federation of the Electronics Industry, which wants checks put in place, and is concerned about the proposed 'allocation through auction' system, which is expected to raise £1bn.



A development kit for Motorola's eight-bit *MC68HC(7)05J1A* microcontroller is available at a cost of around £30. Called *JICS*, the kit is being made available by Motorola via all its distributors. It includes in-circuit simulator, cabling, power supply and sample device. As part of the 68HC05 family the microcontroller includes on chip, a 15-stage timer, 1.2k rom, 64 byte ram, keyboard scan and mask-selectable pull-downs.

Change predicted in semi industry product structure

The product structure of the semiconductor industry is changing, says Jean-Philippe Dauvin, president of World Semiconductor Trade Statistics, WSTS.

Instead of the industry's growth being led by dynamic ram and microprocessors, it will be led by what Dauvin calls 'differentiated ICs' such as digital signal processors.

That's because the performance curve for pcss is slowing: "133MHz and 32Mbytes is sufficient for most people," says Dauvin.

So the growth of microprocessors and dynamic ram will slow from the 40 per cent per year level of the last five to six years to around 20 per cent for micros and five per cent for dynamic ram over the coming five to six years.

The dynamic ram market will be "red and wet for a long time," says Dauvin. He reckons that continued overcapacity will drive down prices for some time to come.

Overall, the world semiconductor industry will grow at 21 per cent between 1997 and 2002, driven by the move to mobile products and digital consumer products.

When the dip of 1996/7 is factored in, the current decade's growth rate is restored to the industry's traditional 17.5 per cent per annum.

New LAN boasts nearly half a Tbit/s

BT Laboratories has a prototype local area network, or LAN, operating at 40Gbit/s – thought to be the fastest LAN yet reported.

"The work is looking at requirements five to ten years in the future," said Julian Lucek, head of the development team. "Currently it is the sort of network that could be used to connect supercomputers within a building, but today's supercomputer is tomorrow's server or desktop and it could be applied to a future server cluster, processing something like weather data or real-time ultra high definition images."

Called SynchroLan, it is based on optical time division multiple accesS, OR TDMA, and provides 16 simultaneous 2.5Gbit/s channels.

The system is timed by a clock generator which feeds a polarised clock signal of 4ps pulses separated by 400ps gaps into one end of a polarisation-maintaining fibre. This runs through the write-side of all the nodes in the system and these add their data bits onto the fibre, one bit per node per 400ps slot. The data is sent polarised at a different angle to the clock thereby avoiding interaction.

The fibre returns through the read-sides of all the nodes, where the data is extracted at each node. This is also timed by the polarisation-separated clock. Control is through an Ethernet bus. This could however be done over the main bus.

Lucek said: "Using TDMA guarantees the 2.5Gbit/s channels. Currently, the prototype is communicating over 50m in total. I think 1km is achievable without significant modification, and further still with dispersion compensation."

White-light diodes – now

While others are developing white leds, Nichia of Japan is manufacturing them. Nichia's led is a blue emitting chip with a layer of phosphor on top that converts some of the blue to the other colours making white. Luminous efficiency is 7.5 lm/W – a bit less than a normal light bulb. This is similar technology to that being developed by the Fraunhofer Institute in Germany, which should emerge as Siemens' white Lucoleds. Find Nichia on:
http://www1a.meshnet.or.jp/nichia/wled-e.htm
Just in case you want to play with them, a US company at http://www.photonlight.com/ sells key fob torches that include one and will ship to the UK.

Help with year 2000 date problems

Companies concerned about the possible effect of the Year 2000 problem on their embedded systems should consult the Institution of Electrical Engineers' (IEE) Web page set up to help address the problem. Unlike other Year 2000 sites, the page has embedded systems as its primary concern. The IEE advises that companies should include a consideration of embedded systems in their Millennium action plans. So if embedded systems are your bag, you had better check it out.
http://www.iee.org.uk/2000risk.

Japanese semi firms suffer due to d-ram price fall

NEC's semiconductor operation was the only one of Japan's big five semiconductor companies to report a profit for last year.

All five companies reported their annual results last week. Toshiba, Hitachi, Fujitsu and Mitsubishi Electric revealed their semiconductor divisions made losses following the collapse in memory prices. NEC made a profit. All five expect a better 1997.

Lack of pcb competition not healthy

A pcb industry executive has warned about a lack of competition in the UK market following recent mergers. David Douglas, finance director of Devon-based manufacturer Eurotech, is concerned about the limited choice in the market and the effect this could have on companies in the long term. "With the amalgamation of Forward, Exacta and now ISL, pcb buyers have a restricted choice of volume suppliers," said Douglas. "I have serious reservations about the long term effect this will have on the industry." ■

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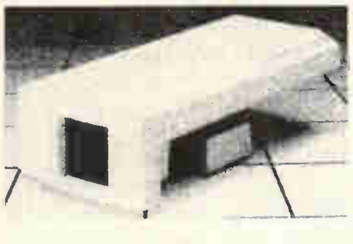

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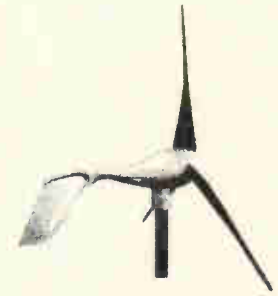

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RESEARCH NOTES

Jonathan Campbell

Ion drives provide thrust for UK spacecraft

UK space researchers are currently in the final stages of testing an electric propulsion system that could form the basis for the next generation of smaller, lighter – but much more efficient – spacecraft engines. The ion drive being developed in the Space Department at the Defence Evaluation and Research Agency (DERA), Farnborough, uses an electric field to accelerate ionised xenon propellant, producing an extremely energetic exhaust that offers much greater efficiency levels than are possible with conventional engines.

The original concept for the ion drive has been around for some years, and many engines have been built and tested around the world. But the UK design ("Spacecraft ion propulsion development in the UK," DG Fearn, Proc Instn Mech Enginrs, Vol 211, Part G, pp. 103-112) is generally accepted to have one of the greatest thruster efficiency yet achieved.

Ion drives have several advantages over chemical motors. In a chemical drive, the velocity achievable is limited by the energy liberated in a chemical reaction – currently less than 5km/s for the most energetic systems. But in an ion drive the velocity is limited only by the power available and, ultimately, by the speed of light.

The highly ionised plasma drifts towards a set of closely-spaced grids at the downstream end of the discharge chamber, where the positive ions are extracted and accelerated to a very high velocity–typically 30-50km/s – by electric potentials of between 1000-2500V. The positive space charge is neutralised by electrons emitted by an external cathode.

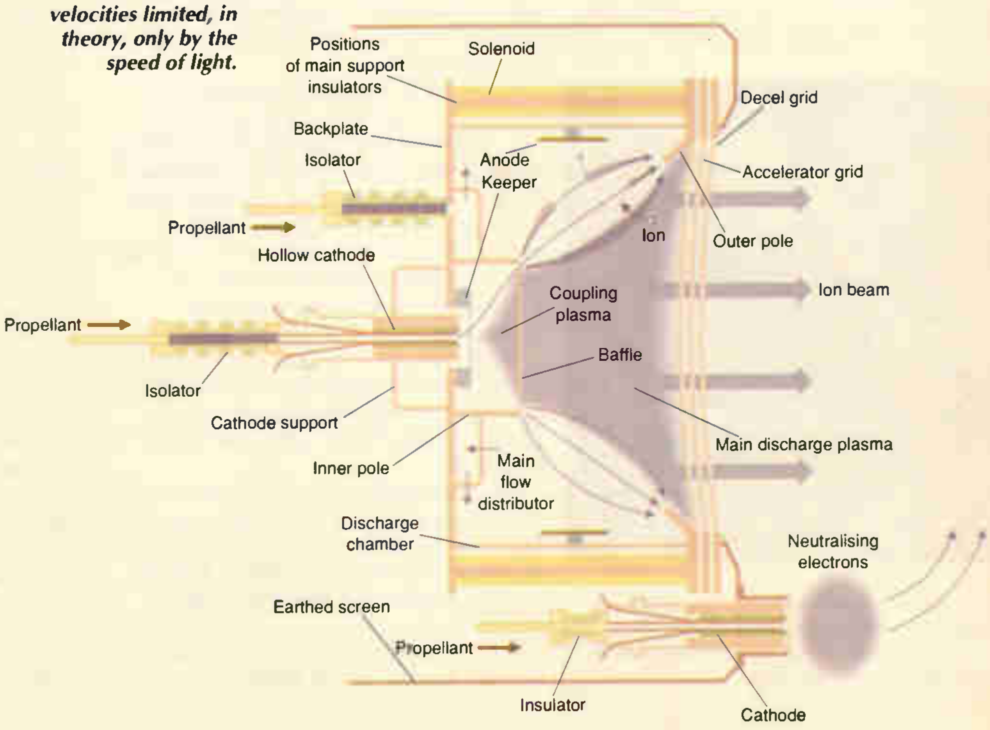
Three features of the engine in particular help give it a performance lift over other designs. The grids dish inwards giving preferable dimensional characteristics; solenoids are used rather than permanent magnets allowing the plasma conditions in the chamber to be modulated; and separate propellant streams flow to the cathode, discharge chamber and neutraliser – again aiding control of the engine.

Final trials and life testing are still to be carried out at dra and Matra Marconi Space at Portsmouth.

Unfortunately, the limiting factor on when the engine is going to be used in space is likely to be availability of a platform rather than perfection of the technology. Originally, the ion drive was to be used on the European Space Agency's (esa's) Artemis communication satellite. But this has now been substantially delayed until the year 2000 and the team are currently looking around for another opportunity.

Contact Dave Fearn, at the Space Department, Defence Evaluation and Research Agency (DERA), Farnborough, Hampshire.

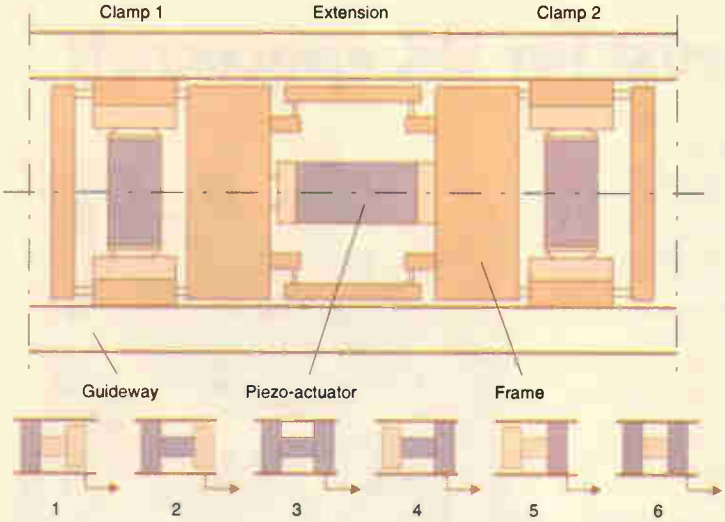
Ion drives could offer velocities limited, in theory, only by the speed of light.



Linear motor makes grinding noises

Three actuators located in a rigid frame add up to a workable piezoelectric linear positional motor. Six actuations are needed to move the motor forward one step.

Tolerances on precision grinding work are becoming ever tighter as the drive for higher quality continues across manufacturing industry. Now two researchers in the Precision Manufacturing Center at the University of Connecticut have developed a linear piezomotor system that provides nanometre positional accuracy for a machine tool, over unlimited length of travel – in a much cheaper solution than is provided by other approaches.



Leadscrews are still the predominant method for positioning elements in precision machining applications, and the problems of backlash and stick-slip must be taken into account. Alternative positioning approaches have suffered from limited stiffness, small positioning range, inadequate forces generated or high cost.

Piezoelectric systems can offer several benefits in these areas and other piezoelectric positioning systems have been tried. But only this latest development seems to earn good marks in all areas.

Designed by Bi Zhang and Zhenqi Zhu ("Developing a linear piezomotor with nanometre resolution and high stiffness,' IEEE/ASME Trans on Mechatronics, Vol 2, No1, pp. 23-29) the system is built of three piezoelectric actuators designed in an 'I' formation, moving backward and forward in a rigid guideway frame.

The front and back actuators lie between the guides and, when activated, expand to clamp against the guides. The third actuator lies in the plane of the guideways and provides the forward and backward movement.

To achieve movement takes six actuations. The first 'clamp' actuator locks into position then the central device expands, pushing against the clamp and propelling the front clamp forward. This then locks in place, so that when the central actuator is contracted, the rear clamp moves forward. In this way the linear motor progress along the guideways.

Tests have shown that the motor has a positioning resolution of 5nm, a high stiffness of 90N/μm and an output force of 200N. It can also travel at 6mm/s.

Biosensor has electronic output

A team of Australian researchers has fabricated a biosensor – a combination of biology and physics – that can detect substances with unprecedented sensitivity. It could find uses in everything from pesticide monitoring to drug testing in athletes.

Australian art: A fully functioning nanomachine composed of molecular chains whose special properties act together to form a minutely and specifically sensitive biosensor

Its central component is a tiny electrical switch, an ion-channel, 1.5nm in size, but the biosensor itself operates by blending the ability of biology to identify individual types of molecule in complex mixtures, with the speed, convenience and low cost of microelectronics

It consists of a synthetic membrane, chemically tethered to a thin metal film coated onto plastic. This membrane behaves like the outer skin of the cells of the human body in its ability to sense other molecules.

Human cell signalling and sensory systems depend on the behaviour of sodium and potassium in the body and the ion currents that flow across certain cell membranes.

When the membrane detects its target molecule, it turns these currents on or off by opening or closing molecular channels passing through the otherwise-insulating membrane.

"We have made a synthetic version of this mechanism that is stable, inexpensive and convenient to use as a molecular detector," says Bruce Cornell of Australia's CSIRO, who has led the team of designed the device.

Being a biological mimic, the device even operates successfully in whole blood.

The team claims to have demonstrated a sensitivity 1000 times better than previous biosensors, and a direct electrical rather than the usual optical, output ensures compatibility with microelectronics, allowing the device to produce a simple, quantitative, readout.

Commercial partner Ambri says it expects biosensors to find a huge range of potential uses, especially in medicine, for detecting drugs, hormones, viruses, pesticides and to identify gene sequences for diagnosing genetic disorders. In the pharmaceuticals industry the device may also be used to screen for potential new drugs and medically-active compounds.

Low cost, very high sensitivity and ease of use, means the devices could also find particular application in on-site measurements, such as ensuring food safety and quality, in environmental monitoring and drug detection in athletes.

Cornell describes the biosensor's sub-picomolar sensitivity as "equivalent to detecting the increase of the sugar content of Sydney Harbour after throwing a sugar cube from a ferry".

Contact: Keith Daniel, Ambri, +61 2 9422 3003 and Dr Bruce Cornell, CSIRO on +61 2 9422 3195.

Innovative lcd research has a pattern...

An elegantly simple method of anchoring liquid crystals to make patterns that are functional as well as visually pleasing has been developed by two researchers at the University of California, Davis.

Nicholas Abbott and Vinay Gupta of the chemical engineering and materials science department have combined recent advances in engineering to develop a novel way of organising liquid crystals. They are using molecular anchors made of simple hydrogen- and carbon-based compounds to provide a stable mooring for liquid crystals in patterns. Anchoring in this way will allow the unique qualities of liquid crystals to be exploited for practical uses on both flat and curved surfaces – in diffraction gratings and curved viewing screens for example. These anchored liquid crystals also promise to be a powerful tool in the study of the surfaces of materials.

Liquid crystals constitute a state of matter, just like the more widely known states of liquid, solid and gas. In liquid crystal form, molecules are free to move about in space, but are ordered in such a way that the molecules, on average, point in the same direction.

Despite the high-tech nature of many liquid crystal uses, the liquid crystals themselves are not exotic chemicals. They retain many of the optical qualities – such as the ability to bend light and change its colour – as their more stationary solid crystal counterparts. But they also have the added advantage of a liquid's mobility, so they are easily moved around in an electric field.

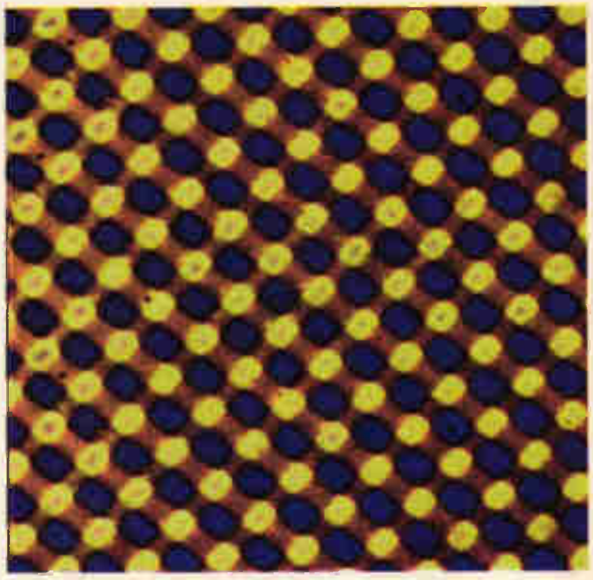
In a standard application, such as a computer screen, the surface is coated with a long-chain molecule, then rubbed with cloth so that the crystals face in a preferred direction.

But Abbott and Gupta have found another way to make the liquid crystals line up, by building a sandwich of two glass plates, one with a uniform orientation of anchors, and the other with a maze of anchoring docks. Rather than a physical connection like a boat's anchor, these molecular anchors function as berths to orient the liquid crystal molecules.

Metals, such as gold, have many free electrons flowing along their surfaces that cause electrically-sensitive liquid crystal molecules to orient themselves relative to the metal surface. Using layer of gold 30 to 50 molecules thick, adhered to a glass plate, the researchers added another single layer of molecules called alkanethiols, ranging from four to 18 carbons long. Very few things stick strongly to the surface of gold except alkanethiols. Once the layer has formed,which happens spontaneously, it directs the orientation of liquid crystals applied on top of it.

To make the more intricate patterns required of a diffraction grating, the researchers applied the thiol in a "rubber stamp" fashion. By using different thiols, the researchers created grids that caused sections of liquid crystal flowing above the thiols to orient differently from neighbouring sections.

The technology is unlikely to have an impact on the well-established methods of preparing thin layers of liquid crystal for portable computer screens or digital watches. But where it might find a use in speciality devices using a curved surface, such as optical lenses and screens visible from many different angles.



Liquid crystal patterns viewed under light microscopy with crossed polarisers. Anchoring liquid crystals into patterns opens up their use in speciality devices such as diffraction gratings.

...and could lead to faster devices

A new type of thin-film transistor developed at the University of Illinois could improve the resolution of flat-panel, liquid-crystal displays used in laptop computers. The transistor has a novel 'buried channel' allowing electrons to move faster and achieving much faster switching speeds.

In conventional thin-film transistors, electrons travel near the semiconductor-insulator interface, where the silicon is strained and is of poor quality. By creating a buried conducting channel, recessed about 500nm away from the interface, the researchers say they can increase the speed of the electrons and significantly enhance performance.

Flat-panel displays consist of hundreds of thousands of pixels, each controlled by a thin-film transistor. While the performance of these displays is adequate at present, future applications that require higher resolution (such as high-definition television or enhanced computer displays) will be limited by the speed at which the transistors can switch.

But, because the buried channel is placed in a region of higher-quality material, electrons in the new device can move nearly twice as fast as they normally would. So it takes significantly less time to deliver the necessary charge to turn a pixel on or off. Or many more pixels could be addressed in the same amount of time.

To fabricate the buried channel thin-film transistors, John Abelson, a professor of materials science and engineering has been using a technique called reactive magnetron sputtering. This method – which uses a plasma to erode a silicon target and deposit a film – provides precise control over layer composition and electronic properties.

To create a step in the conduction band – and a buried channel – involves varying the amount of hydrogen gas injected into the plasma as the amorphous silicon layer is deposited.

The sputtering technique also allows films to be deposited at much lower temperatures than currently possible with the plasma-enhanced chemical-vapour-deposition process used by industry. Transistors have so far been produced at a processing temperature of 125°C, opening up the possibility of using lightweight and impact-resistant plastic substrates in place of current glass substrates. ■

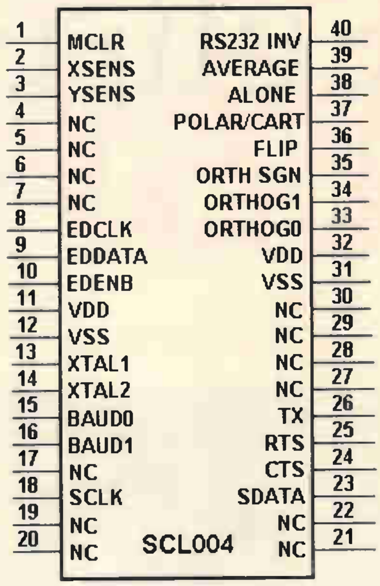
Contact: John Abelson, University of Illinois at Urbana-Champaign, 807 S. Wright St., Suite 520 East Champaign, IL 61820-6219, USA.

Exclusive EW reader offer

Two-axis sensor and the SCL004 signal processor at 33% discount.

Both the *FGM-2* two-axis magnetic sensor and *SCL004* signal processing chip are available with an exclusive *EW* reader discount of 33.3% until 30 September. These parts with a handful of other components form a precise electronic compass, with RS232 output, as detailed in the article on page 630.

Normally the set costs £45.83, but you can obtain them for £30.55, fully inclusive of VAT and UK postage, directly from their UK manufacturer Speake and Co, tel. 01873 811281, fax 01873 810958. Overseas readers should contact Speake and Co for offer details.



SCL004 pin details

Pin 1 is reset, which clears internal registers and restarts the autocalibration process. Tie directly to Vdd or via a 10k to 47kΩ resistor to V_{dd} for external control.

Pins 2, 3 connect directly to the output of an FGM series sensor.

Pins 4-7, 17, 19-22, 27-30 must be left open circuit.

Pin 8 is the clock for the serial link to a digital display. Suitable displays are *TSM6055*, *TSM6255* and *TSM6755* from Three-Five Systems.

Pin 9 is data input for the serial link to digital display.

Pin 10 enables data for the serial link to digital display.

Pins 11, 32 connect to +5V supply. Use 47nF decoupling to V_{ss}, close to the pin connection.

Pins 12, 31 are V_{ss} or GND connection pins.

Pins 13, 14 connect to an external crystal or ceramic resonator for 4MHz clock.

Pins 15, 16 control the RS232 bit-rate. Pull high via 10-47kΩ for external control or tie to V_{dd}.
00=1200 baud, 01=2400 baud,
10=4800 baud, 11=9600 baud.

Pin 18 is clock drive for I^2C interface to eeprom used to store autocalibration parameters. Leave open circuit if eeprom not used.

Pin 23 is the eeprom I^2C data line.

Pin 24 inputs the RS232 CTS signal. Must be taken high via a 10-47kΩ resistor if unused.

Pin 25 is RS232 RTS output. Leave open circuit if unused.

Pin 26 is TxD, i.e. outputs RS232 data. Leave open circuit if unused.

Pins 33, 34 carry ORTHOG0, ORTHOG1 signals for correcting non-orthogonality error in sensor assembly. Switch to V_{dd} or V_{ss} for external control or leave open circuit. Corrections are programmed in binary,
00=no correction, 01=0.5°, 10=1°, 11=2°

Pin 35 determines the sign of the non-orthogonality correction. Pull-up not required.

Pin 36 changes the direction of increasing angle around the circle. High for clockwise, with the sensor mounted pins down. No pull-up needed. This permits a normal direction of rotation to be obtained with the sensor mounted upside down.

Pin 37 controls the type of data sent by the RS232 output. No pull-up required. One form of output is an angular reading in degrees in the format 00*nnn*(CR), where *nnn* is the three digits of the 360° heading, preceded by two zeros and followed by a carriage return. This is obtained with a high input to the pin. The other is Cartesian, giving corrected *X* and *Y* values of the sensor field components, in the format ±0*xxxx*,±0*yyyy*(CR) and is obtained when the input to the pin is low.

Pin 38 should be left open circuit if no eeprom is used for parameter storage or taken low if one is used.

Pin 39 if this pin is left open, the long term average of a number of autocalibrations is used to set the sensor parameters. If low, it forces the most recent autocalibration to be put into effect.

Pin 40 taken high inverts the TX, RTS and CTS signal lines for use with inverting RS232 drivers.

Stop press

Indications that many potential users do not intend to use this type of chip in bolt-down mode, have resulted in a minor variant designated SCL004A, which we feel is more suited for use in gimballed systems.

The SCL004A is identical in all respects to the standard version described in the article, except that the 'stand-alone input' on pin 38 is replaced by a 'calibrate-on-demand input' function. This optionally permits a later upgrade of the system to take advantage of an electronic gimbal, an option not available to the bolt-down SCL004 version.

Note that the A variant **must** have a supporting eeprom chip. Both versions are identically priced. Please indicate which you require when ordering.

Self-calibrating compass

Featuring its debut in *Electronics World* is a new chip that forms the core of a precise, self-calibrating electronic compass, described by Richard Noble.

With only a few additional components, the new *SCL004* chip provides all the functions needed to make a precision compass, when used in conjunction with an *FGM-2* sensor. A feature of the compass system is its one degree resolution capability.

The chip has been programmed with algorithms described in the March 1997 issue of *Electronics World*[1]. These algorithms allow the chip to continuously remove the effect of the zero offsets and sensitivity variations of the sensing element.

To permit correction for the small amount by which the individual sensors may not be truly at right angles to one another, the chip also allows for externally programmable orthogonality adjustment.

Used properly, these features will usually achieve a one degree accuracy although the maker only claims a two degree overall accuracy for the product. Other factors can give rise to greater errors and it is important to pay attention to them in setting up a system. They are detailed later.

How the chip operates

The direction of increasing degree count can be set either clockwise or anti-clockwise, allowing the sensor to be mounted upside-down if this is more convenient.

Normally, after power-up, the compass needs to be rotated in order to trigger an auto calibration before readings are available. To circumvent this potential nuisance, a system of storing the correction factors in an external eeprom is provided. The auto calibration data, derived from a previous period of use, provides immediate readings during the wait for an initial calibration calculation.

Another selectable feature permits the use of the long term average of the calibration factors, rather than the most recently determined set. The averages are also stored in the eeprom if this feature is implemented.

Output from the chip is via an RS232 compatible serial port. The levels are ttl or 5V c-mos compatible. Over short distances the output signal can usually be used directly. If longer distances are involved, it is better to feed them to a standard RS232 interface chip. An external control permits inverting the outputs to make interfacing easier. Bit rates from 1200 to 9600 are pin programmable.

Two different types of output are available, selectable by DIP switch or signal line from a computer or microcontroller. One is an angular reading in degrees in the format 00*nnn*(CR), where *nnn* are the three digits of the 360° heading, preceded by two zeros and followed by a carriage return. The other is in Cartesian form, giving corrected X and Y values of the sensor field components, in the format ±0*xxxx*,±0*yyyy*(CR)

The Cartesian format allows you to carry out your own alternative processing of the basic corrected sensor data, rather than using the chip's internal trigonometric routines. This means that you can add linearity corrections, etc, or even use the data for some other purpose entirely.

An additional type of output is available, for those who do not wish to use a computer or other RS232 link. This is a serial feed to a direct digital display. The serial stream and the RS232 link can be used simultaneously.

Application to compass design

Originally, the chip was designed for use in bolt-down orientation applications, such as rotating aerials, etc. But if it is mounted in a suitable gimballing system, designed to maintain a mean level attitude, the chip can be used for conventional compass designs.

Because tilt from a level attitude has a marked effect on such sensor systems, some information on this is provided in what follows, both as regards bolt-down and gimballed systems.

The influence of the autocalibration algorithms in the presence of tilt is also examined.

Basic tilt effect

In most regions of the Earth's surface, the magnetic field vector is at an angle to the local horizontal plane. This tilt is a source of problems in compass sensors.

In the temperate latitudes, this dip angle is commonly very large. In the United Kingdom, for example it is around 67° to the horizontal.

It also points acutely downwards to the north, leading to rather awkward looking three dimensional diagrams. For an open axis view point the XY plane should be viewed from below and projections to the horizontal plane go upwards, leading to an unfamiliar look to the vector diagrams.

For the sake of more familiar visualisation, all subsequent diagrams are shown as for a southern hemisphere location of equivalent latitude, with an upward pointing field vector to the north.

Other simple geometric tricks are also used where appropriate to ease the visualisation process, such as always drawing the sensor axes as horizontal ones. For tilt, the world is tilted in the opposite direction instead. In place of tilting the sensor the field vector is given an opposing tilt. The geometry is relative and the results are the same, but some simplification of the diagrams is usually possible.

A compass sensor which is assumed to lie in the horizontal plane will see quite large changes, therefore, in its projected horizontal components if it moves out of that plane. This is because the components are quite small to start with.

Not all of these changes have an influence on the final result however. Assume that an orthogonal pair of sensors is aligned horizontally so that one of them, *Y* for example, lies in the north/south direction and the other, *X*, in the east/west direction. In this case, a tilt along the north/south line, ie a rotation about the east/west axis, has no effect on the compass bearing calculated from the sensor readings.

The reason for this is that the *Y* component is finite and varies with the tilt, but the *X* component is zero. A bearing computed from the arctangent of *X*/*Y* is always zero degrees regardless of the value of *Y*, as long as it does not itself go to zero. So no error occurs even for large tilts.

Tilt along the east/west line however, ie a rotation about the north/south axis, produces a finite value of the *X* component and a finite bearing error. Worse still, a small tilt gives quite a large value to the *X* component if the dip angle is large, so the bearing error is much larger than the tilt angle. This is illustrated in **Fig. 1**, in which the sensor is shown horizontal and the world rotated around the

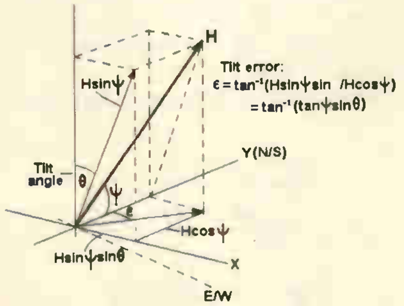


Fig. 1. This source of compass error – tilt along the east-west line – produces a finite value of the X component, thus a finite bearing error.

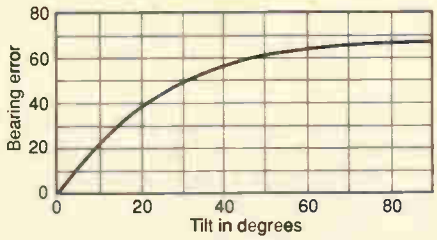


Fig. 2. Bearing error versus tilt angle shows how error rate increase with tilt gets smaller as the tilt increases.

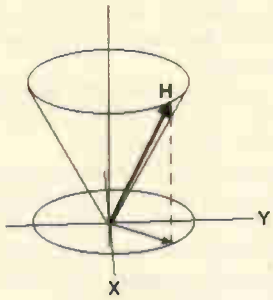


Fig. 3. Visualising the world rotating the around the sensor may help you understand error due to having the sensor parallel to a rotating tilt.

north/south axis.

For this sensor alignment, the bearing error, ε, is,

$$\varepsilon=\arctan(\tan\psi\sin\theta)$$

where ψ is the dip angle and θ is the tilt angle. For the previously quoted dip angle of 67° this gives a bearing error of 2.35° for a tilt of 1°.

The rate at which the bearing error increases with tilt gets smaller as the tilt increases up to a maximum error of 67° when the tilt is 90° as illustrated in **Fig. 2**.

For tilt along a line between these two extremes, a similar pattern of error occurs but with a smaller overall scale. Along a line close to the north/south direction tilt causes small errors. Along a a line close to the east/west direction tilt causes much larger errors and there is a smooth transition from large errors to small ones as the line of the tilt moves around the horizontal plane.

In fact it is really only the east/west tilt which causes the problem. The easiest way to deal with this is to resolve any tilt, like a vector into two components, one along the north/south direction and the other along the east/west direction. The north/south component can the be ignored, since we know it has no effect, the east west component being used on its own to determine the bearing error.

The tilt angle cannot be resolved directly. What is resolvable as a vector is the slope resulting from the tilt. But since this is just tan θ, where θ is the tilt angle, the trick is to take the tangent, do the resolution and then take the arctangent of the components to get the resolved angles.

So, for example, if the maximum tilt line is along a line at an angle to the north/south direction, the east/west component of tilt is

$$\theta_{ew}=\arctan(\sin\phi\tan\theta)$$

and the north/south component is,

$$\theta_{ns}=\arctan(\cos\phi\tan\theta)$$

So far, the analysis has been in terms of a sensor system in which the *Y* axis sensor is aligned north/south. In fact it does not matter which direction the sensors are oriented, the result is the same and only depends on the tilt in the magnetic east/west direction.

Bolt-down applications

In bolt-down applications there are two different basic kinds of tilt, as follows. Firstly, there are those in which the sensor plane is truly parallel to the rotation plane but the latter is not truly horizontal. Secondly, there are tilts in which the rotation plane is truly horizontal but the sensor plane is tilted with respect to the rotation plane.

Sensor parallel to a tilted rotation plane. This is equivalent to the situation illustrated previously, in which the sensor plane is horizontal, the rotation axis vertical but the world and field vector tilted sideways giving the previous calculated bearing error.

A little consideration will also show that it is no different from the ideal situation of a horizontal sensor and vertical rotation axis, except that the field vector now has a slightly smaller dip angle and no longer points precisely along the north/south line.

Another visualisation trick is to rotate the world around the sensor rather than rotating the sensor in considering the compass behaviour. The field vector traces out a cone in space, **Fig. 3**.

Projection of the cone on to the XY plane of the sensor is a perfect circle. This implies that the indicated compass angles will be correctly and evenly spaced, but will show a fixed error, related to the tilt, around the whole 360° of rotation. This is not a disaster since the sensor will normally be aligned in a direction which gives the desired reading during initial installation. It may be set up, for example, to read true north rather than magnetic north for the location in question.

Also of interest is the effect all of this has on the operation of the autocalibration algorithms used by the chip software. Since the situation described above is little different from normal compass operation, the algorithms continue to work correctly in eliminating any zero offsets and scale errors encountered, as accurately as they would in ideal circumstances. However they have no way of becoming aware of the bearing error and make no attempt to correct it.

Sensor tilted relative to horizontal rotation. This is slightly more difficult to visualise. Using the same technique as before, if the sensor is considered to be stationary, the world is rotated about a vertical axis and the sensor

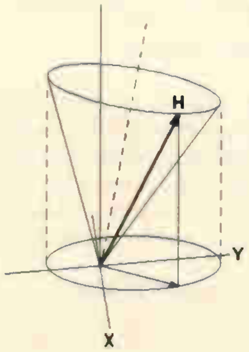


Fig. 4. When the sensor is tilted in relation to horizontal rotation, unlike the case in Fig. 3, the cone is tilted.

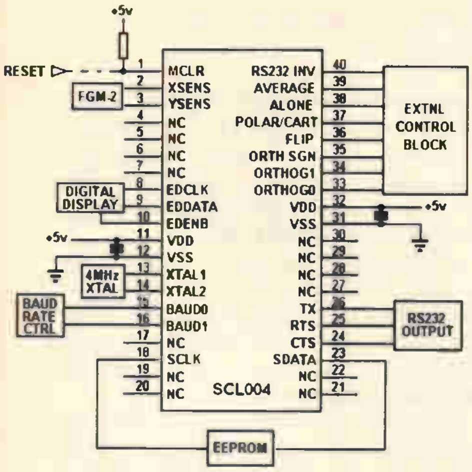


Fig. 5. Building blocks for complete electronic compass using the SCL004 signal processing chip.

plane is then tilted down to be horizontal, a little consideration shows that it is the same as the previous case, but with the cone tilted with respect to the sensor, **Fig. 4**.

The projection of the field vector tip path on to the sensor plane can be seen to be an off-centred ellipse. This is exactly what the chip sees coming from a raw, uncorrected sensor with its massive zero offsets and unbalanced scale factors, except that the errors arising from tilt are much smaller. Consequently, these smaller errors are swallowed up by the autocalibration algorithm and eliminated along with the raw sensor errors.

Effectively, the projection on to the XY plane is reduced to a perfect circle and the compass behaves as if the tilt did not exist. The autocalibration not only functions correctly in the presence of tilt, it also removes the effect of tilt.

This can be experimentally demonstrated to occur even for very large tilts giving even larger bearing errors initially.

As a consequence of this property of the autocalibration procedure, it is not necessary to take great care in the level mounting of the sensor in bolt-down type applications. An initial north south alignment is all that is required during the setting up procedure.

Applying a gimbal

The above discussion only applies to the bolt-down type of sensor mounting, where the sensor is constrained to rotate in a fixed plane, albeit not necessarily horizontal. This is necessary for the autocalibration to work.

An alternative to the bolt-down arrangement, to permit some tilting to the carrying platform or vehicle, is a gimballed or pendulous sensor. Provided that the axis of rotation remains vertical, which it does in such arrangements, the system is the equivalent of having the sensor tilted relative to horizontal rotation as discussed above.

The sensor may not be truly horizontal, but the effects of that tilt will be removed as before. It is not therefore necessary to take great care that the sensor hangs level as it is in most normal gimballed systems, again an advantage in ease of setting up.

Since in platform motion the rotation axis may not always be absolutely vertical, individual calibrations may suffer from momentary errors. However since the mean position over a longer period will approximate the vertical better than any instantaneous position, it is best to use the chip mode which makes its corrections by using the long term averages of the parameters, for gimballed systems.

While not perfect, this approach should improve the accuracy of the final output.

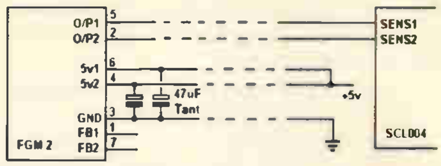


Fig. 6. Sensor connections are simple, but make sure there is adequate decoupling close to the sensor pins.

Magnetic material disturbance effects

My earlier article described an unusual property of the algorithms which suggested that autocalibration would remove the effects of magnetic disturbance caused by the hardware on which the compass system was mounted.

Experimental verification of this effect has been obtained and shows the cross-axis effects to be minor as anticipated, giving the overall sensor/chip combination a further useful protection against presenting erroneous readings.

Typical configuration

Figure 5 outlines a completely configurable system. It is not necessary to make use of all the configurable features at any one time. If a set of fixed conditions is all that is required then the control pins can be directly wired high or low as needed to set up those conditions. This reduces the external component count and simplifies the layout.

The reset input can be taken directly to +5V, or via a resistor as shown. Reset occurs on power-up. It can be initiated subsequently by an incoming ttl or c-mos signal going temporarily to ground, or by a manual push button to ground.

Sensor interface. Common supply-line impedance can give rise to a tendency for the individual sensors to lock together over short bands of their range. This can be eliminated by decoupling capacitors fitted close to the sensor pins, **Fig. 6**.

Separate supply lines are provided for each

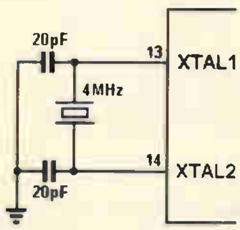
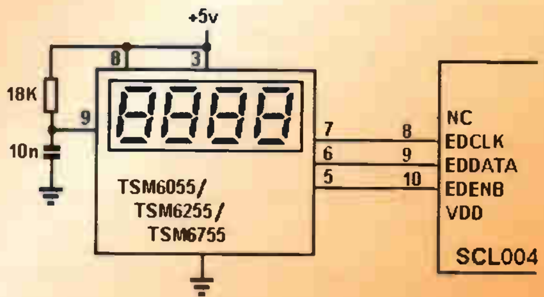


Fig. 7. Connection of a crystal or resonator for the 4MHz clock.

Connecting a display

The SCL004 drives a suitable external digital display directly. An interface is provided for immediate display without the use of a pc or microcontroller system in the shape of a serial data stream of seven-segment data for appropriate displays. Suitable types are Three-Five Systems Inc. modules TMS6055, TMS6255 and TMS6755.



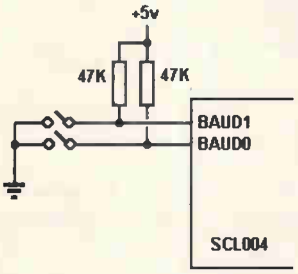


Fig. 8. Two pins program one of four RS232 bit rates, with 00 being the lowest at 1200baud, and 11 the highest, at 9600baud.

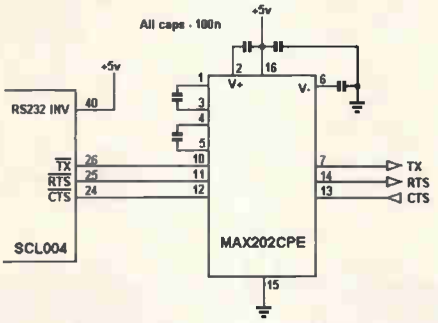


Fig. 9. For short runs, the SCL004 logic-level i/o can be used directly but if the microcontroller or computer is some distance from the sensor, a proper RS232 interface is needed.

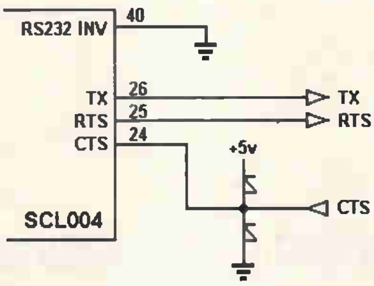


Fig. 10. When only a short RS232 link is needed, the full RS232 interface is not essential, provided that the other end of the link is also operating at ordinary ttl or c-mos logic levels.

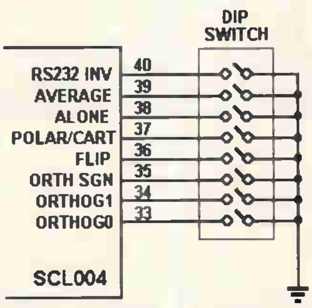


Fig. 11. Control signals can be hard wired, set via dip switches, or determined via external logic.

+5V rail. If long leads are used to the sensor, more effective decoupling is obtained by running these lines separately back to the supply source.

External crystal circuit. A crystal or ceramic resonator oscillating at 4MHz is required, **Fig. 7**.

Bit-rate control interface. A switchable rate system is illustrated in **Fig. 8**. The rate can also be controlled by ttl or c-mos input from an external source. If only one fixed bit rate is required, then the pins may be directly wired to +5V or ground as appropriate.

RS232 interface. A full voltage specification interface is shown in **Fig. 9**, using an inverting RS232 driver chip.

This type of interface is recommended where long lines will be used to connect the serial link. It is also necessary to take pin 40, RS232 INV high on the *SCL004* chip if inverting drivers are to be used.

Over modest length signal lines, to a pc or microcontroller, a simpler interface will suffice, **Fig. 10**. In this configuration, pin 40 on the *SCL* chip must be grounded. It is not necessary to use the RS232 interface if the direct digital display is used, but in this case, CTS on pin 24 must be connected high to avoid the chip stalling.

External control block. This block of controls affects the way in which various functions of the chip are employed and can be set by external ttl or c-mos levels from other equipment, dual-in-line switch or hard wired connections if fixed conditions are acceptable, **Fig. 11**.

Signal RS232 INV inverts the RS232 signals, where necessary. A high level on the AVERAGE pin forces the chip to use long term averages of both zero offset and scale factor to be used in the computation of angles, rather than the values which have come from the most recent re-calibration. This is useful for systems in which the sensor is gimballed to overcome the effects of tilt and acceleration in the vehicle or vessel in which the sensor is installed.

Momentary displacements in tilt are inevitable in such set-ups but the longer term averages of sensor position more nearly correspond to a level condition. This may be helpful in reducing the instantaneous tilt errors which will otherwise occur continuously, disturbing the proper operation of the autocalibration algorithms.

The *SCL004* is essentially intended as a bolt-down orientation device, but if it is gimballed it may still provide a reasonable performance in many cases.

A high level on the ALONE pin sets the chip into stand alone mode in which it will not provide data until it has carried out an autocalibration. This calls for some rotation in orientation before readings are available, usually not a problem in motor vehicles, for example, which tend to manoeuvre significantly immediately after start-up.

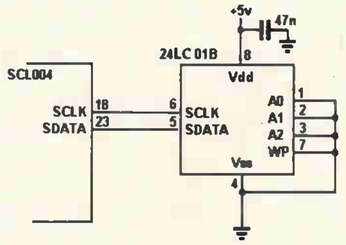


Fig. 12. Connecting an I²C interfacing eeprom to hold the configuration values during power down is simple.

If readings are required immediately after power up, then a low level on this pin will cause the chip to seek a set of parameters from an eeprom in which the last calibration has been stored. If an eeprom is in use the chip always stores its most recent calibration for subsequent use and this data is not lost during power-down.

A high level on POLAR/CART pin gives the angular reading, a low giving the X,Y values of the normalised field. If the FLIP pin is high, rotation is clockwise to increasing angle when the sensor is mounted with its pins facing down. A low level reverses this direction.

Control of the direction in which the orthogonality correction is applied, clockwise or anticlockwise, is carried out via the ORTHOG SGN bit.

Bits ORTOG0, ORTHOG1 can be programmed to give the magnitude of the orthogonality correction as described earlier. The size and direction of these corrections can only be found experimentally.

A simple method is to compare the north/south readings with the east/west readings. These should show an average difference of 90°.

The extent to which they do not is a measure of the direction and magnitude of the orthogonality error. The control pins can then be programmed to minimise this error, preferably spreading the error evenly between the four cardinal points of the compass.

EEPROM Interface. A suitable arrangement for storing the last calibration parameters is shown in **Fig. 12**. The process is automatic, no other action being required apart from setting the ALONE bit low. ■

References

1. D Risk and R Noble, 'True orientation', *Electronics World*, March 1997, pp 197-200

Looking into impedance

Cyril Bateman provides solutions for measuring a component's impedance – regardless of frequency.

Impedance is the ability of any circuit to oppose or limit the flow of current when stimulated by a suitable voltage source[1]. At the simplest level, it is the dc resistance value of a typical resistor, calculated from measurements of voltage drop and through current, using Ohm's law.

With an ideal resistor, circuit current and applied alternating voltage would be exactly in phase with each other, so measured using dc then ac stimulus, both values would be identical. Accurate measurements being made simply by measuring voltage and current magnitude.

In practice, such an ideal component cannot exist. Practical resistors possess both capacitive and inductive parasitic elements, so magnitude only measurements at ac or dc will differ. Parasitic elements cause the phase of the resistor's current to differ from the applied voltage. So apart from those special frequencies when the component is self resonant, true measurement of the ac impedance of components[1] requires that phase difference as well as impedance magnitude be measured.

Consider the measurement of capacitance and inductance. Neither parameter is directly measurable using ac test frequencies. The property actually measured is impedance, from which the effective capacitance or inductance and resistance values are calculated.

True impedance measurement of components at the relevant frequency is important for two reasons. Many circuit applications or simulations, require that component parameters be known at the end use frequency and secondly to permit correlation with the component manufacture end of line testing.

Terms used

Using Ohm's law at ac, it is usual to replace the resistance term with the magnitude of impedance vector **Z**, hence $|Z|=|V|/|I|$. Including phase angle, the description for impedance in polar format becomes $|Z|\angle\theta=(|V|/|I|)\angle\theta$.

Measuring phase

The phase difference of two waveforms, can be calculated by simply measuring the time difference at their zero crossing, dividing this time difference by the periodic time and multiplying by 360. This principle is used in many dedicated phase meters.

This time difference can also be measured using a double beam oscilloscope with an X-trace readout like the PicoScope virtual dual channel oscilloscope that I have used for the measurements shown in this article.

Audio Precision's System One is commonly used for audio measurements. It can be fitted with two voltage measurement channels and phase difference meter as an option. Consequently, this test set could also make reasonably accurate impedance measurements, subject to the limitations of its 100kΩ input impedance.

With electronic components, impedance values are usually expressed[2] in the rectangular format, $|Z|=R\pm jX$, in which R represents the in-phase resistive component, commonly called esr, or equivalent series resistance. Value X is the out-of-phase reactive component, i.e. capacitive or inductive reactance, of the measured vector.

Fortunately most pocket calculators include functions to easily convert impedances between polar and rectangular formats. From this basic rectangular format, many other mathematical transpositions are derived.
Further details are in the panel entitled Impedance transpositions.

Sources of error

Measurements of high impedance, at ac or dc, can be influenced by the shunting effect of the input impedance of the voltmeter used.

When measuring low impedances, the resistance, capacitance and inductance of any test leads used, can easily exceed the value to be measured. Consider the effects of the test leads when measuring dc resistance.

Conventional dc measurements. Resistance values greater than 100Ω, can be measured using almost any multimeter, with sufficient accuracy for most design needs. For values below 10Ω however, the measured value will be overstated, since the resistance of the test leads and contacts used, are included in the displayed result.

Using four terminals. When measuring resistance below 1Ω, lead resistance can exceed the value being measured. These errors resulting from test-lead resistance, can be avoided[1] by using the four-terminal or Kelvin techniques. This ensures that current and voltage are measured using separate pairs of leads. Of course the resistance being measured need not be a discreet component. It might be a track on a printed circuit board, a length of wire or the dc resistance of an inductor.

Four terminals in practice. For low values of resistance, the simplest four terminal measurement is the device-under-test voltage drop when subject to a known constant current. Using a 1A current with a 200mV digital meter, values of a few milliohms can be easily measured. More details are given in the panel entitled four-terminal measurements.

Some years ago, I was faced with the task of measuring resistances from 10mΩ to 100Ω accurately and quickly in an industrial environment. Commercial test meters I looked at were unable to provide a satisfactory solution – regardless of their cost. The components involved could only dissipate a few milliwatts so would not support any significant constant test current. The electrically noisy environment inhibited reliable measurements at the low currents used by commercial meters.

With a test current of less than 1A, power dissipated became excessive at resistance values above 0.1Ω, so a technique that would restrict the power dissipated in the test piece without sacrificing measurement accuracy was needed.

A test current which decreased with increasing device-under-test resistance was arranged using a voltage source and suitable series resistance. Current and voltage drop at the device under test were measured and resistance computed. This resulted in accurate resistance measurements while dissipating only few milliwatts in the test piece.

The final circuit – a trickle charged nickel-cadmium 'D' cell with a 5.6Ω series resistor – provided a short circuit test current less than 250mA, reducing with increasing resistance. This ensured a low, near constant device under test power dissipation. More information is given in the panel entitled Four-terminal measurements.

Four terminals and ac. Alternating-current measurements of low impedance circuits use similar four terminal techniques. They can involve either a known constant current, or a constant voltage source with series resistors.

Voltage, current and phase difference[3] measured using synchronous detectors, provides the basis for many high performance commercial *LCR* test meters. See the panel entitled Phase measurement methods.

While these methods are common in modem automatic *LCR* measuring equipment, they implicitly require meters able to accurately measure currents, voltages and phase angles, at the test frequency.

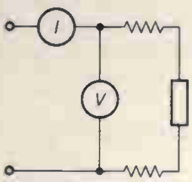
Wheatstone bridge. Long before suitable, easy to use meters were available, a common technique for the dc measurement of resistance, based on ratio arm techniques, was used to compare the unknown, with known resistances.[4] This circuit became known as the Wheatstone bridge.

Subsequently, the Wheatstone bridge was adapted to measure ac characteristics of capacitors and inductors, by adding known capacitor or inductor 'standards' in one arm of the bridge. These were compared with the

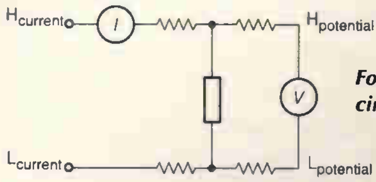
Making four-terminal measurements

Conventional two-terminal measurement of impedance uses the same test lead to both pass the measurement current and measure the voltage drop at the device under test. This results in the measured voltage drop being overstated according to the voltage drop along the test lead used and any contact resistances present.

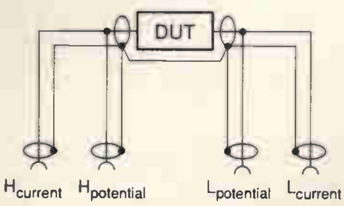
Supplying the test current along one pair of leads while measuring the voltage dropped at the device under test using a second pair of leads – taking care to avoid mutually induced lead voltages – largely overcomes these errors. Use of four coaxial test leads[1] to replace the two pairs of leads, using the screens to carry any return currents together with 'Kelvin' contacts to the device under test, eliminates almost all test lead errors.



Conventional two-terminal circuit for measuring impedance.



Four-terminal circuit.



Measurement configuration.

Occasionally a 'guard' technique, or three-terminal measurement, is used to isolate the device under test from other circuit parasitics. This technique can be upgraded to benefit from the four terminal concepts, when it effectively becomes a six terminal measurement.

The four terminal diagram shown, complete with Kelvin contacts to the device under test, describes the arrangement used for the *HP16047A* component test jig.

Transposing impedance

Series impedances. Any impedance value of the form $|Z|\angle\theta=(|V|/|I|)\angle\theta$ can be transposed from these polar co-ordinates to the rectangular series equivalents of $|Z|=R_s \pm jX_s$ most easily, by using a pocket calculator. Hence also the reverse calculation,

$$|Z| = \sqrt{R_s^2 + X_s^2} \qquad \theta = \tan^{-1}\frac{X_s}{R_s}$$

In the above expressions the R term represents the equivalent series resistance of the measured device, while the X term represents the reactive component. When viewed as a vector diagram, a polar plot or on a Smith chart, this X term has a positive value for inductors, and a negative value for capacitors.

With capacitors, the commonly used expressions are,

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

$$Capacitance = \frac{-1}{\omega X_s}$$

And the equivalent expression for inductors are,

$$Q = \frac{X_s}{R_s}$$

$$Inductance = \frac{X_s}{\omega}$$

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,

$$R_p = \frac{R_s^2 + X_s^2}{R_s} \qquad X_p = \frac{R_s^2 + X_s^2}{X_s}$$

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

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

$$G_p = \frac{R_s}{R_s^2 + X_s^2} \qquad B_p = \frac{X_s}{R_s^2 + X_s^2}$$

Conversion from parallel impedance to series impedance form is equally simple,

$$R_s = \frac{R_p \times X_p^2}{R_p^2 + X_p^2} \qquad X_s = \frac{R_p^2 \times X_p}{R_p^2 + X_p^2}$$

For more information, see ref 2.

unknown impedance, using ratio arm techniques.

With time, as measurements of more difficult unknown impedances became needed, the original Wheatstone bridge became much modified, with different versions able to measure capacitors or inductors in the series and parallel equivalent mode circuits.

While most Wheatstone bridge circuits suffer from interaction between their resistive and reactive balance controls, certain configurations eliminate this interaction, yielding bridges able to accurately and independently balance either or both terms. A particularly useful publication from Rohde & Schwarz[5], analyses the number of interactive adjustments needed to attain bridge balance, for ten different bridge configurations. One was used to successfully measure the esr of aluminium electrolytic capacitors cooled to –55°C – a most difficult measurement. For more on this topic, see the panel Wheatstone bridges.

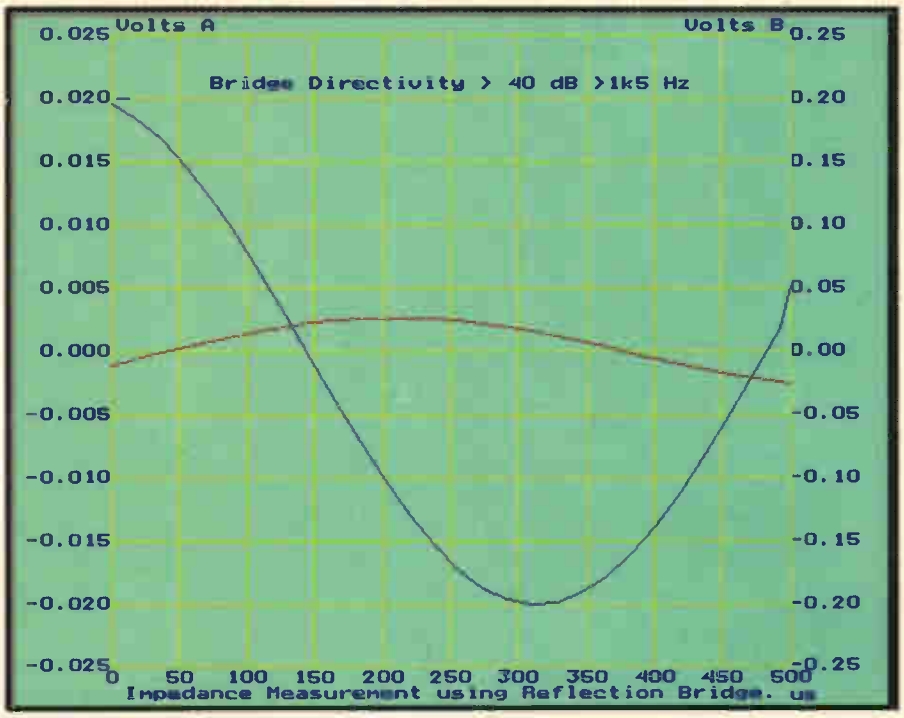


Fig. 1. Using Pico Virtual oscilloscope shows reflection coefficient of bridge loaded with 50Ω. Ideally this reflected voltage would be zero. In practice bridge directivity exceeds 40dB used with this termination at 1500Hz. At higher frequencies directivity improves due to increased impedance of balun used.

What are serial and parallel modes?

The impedance vector of a practical capacitor or inductor 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[1] to that measured, can be either a high value in parallel with the device, or alternatively a low value in series with the device, leading to the term 'equivalent series resistance' or esr.

This esr does not have a finite value. While it is principally frequency dependent, component temperature and applied voltage also have their effects on the measured values.

Have a look at the following practical example. An impedance vector, magnitude 100Ω, phase angle –84.3° at 1kHz, represents a capacitor having a tanδ of 0.1 and 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. The difference in equivalent capacitance is thus 10% and the equivalent series resistance 9.95Ω.

With practical capacitors, a secondary parasitic, self inductance, will be included in the measured result. Similarly for inductors, a self capacitance value will be included.

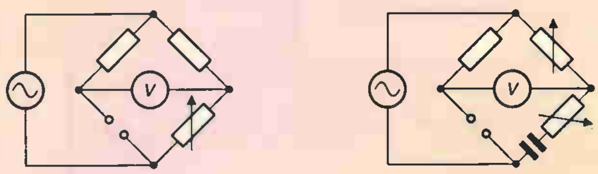
Self inductance and self capacitance

But do these secondary parasitics have any affect on measured values?

You will recall that an inductive reactance has a phase of +90° while a capacitive reactance has a phase of –90°. A simple vector drawing clearly shows that whichever is the minor term, subtracts magnitude from the corresponding major vector component.

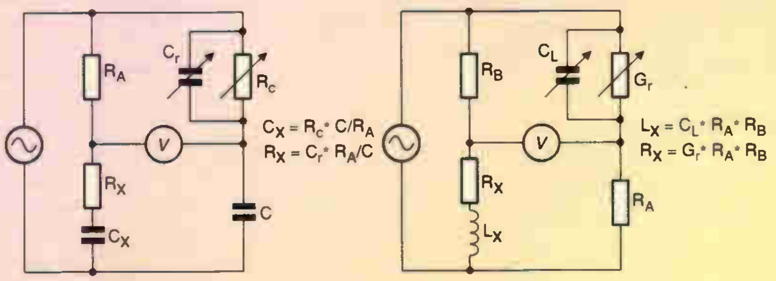
Wheatstone bridges

The conventional Wheatstone bridge used to measure ac impedance of capacitors or inductors, uses a fixed known capacitance standard, together with two calibrated variable resistances. It measures the unknown impedance as a ratio of the known standard, by balancing out the detector voltage to zero. This configuration is used commercially, but suffers from interaction of the two balance controls when measuring low Q or high tanδ components, needs repeated balancing, only slowly approaching true balance.



AC or dc resistance bridge, left, and ac capacitance bridge.

Other configurations are possible which do not suffer from this interaction and can balance resistive and reactive terms almost independently one from the other. A full analysis[5] of ten optional bridge configurations with estimates of the time to balance, shows improved results when using a variable capacitance to replace one variable resistance. The capacitance bridge shown, built using capacitor and resistor decade boxes, has been used to measure very high esr electrolytic capacitors, when two expensive commercial bridges failed to balance.



AC capacitance bridge, left, and ac inductance bridge

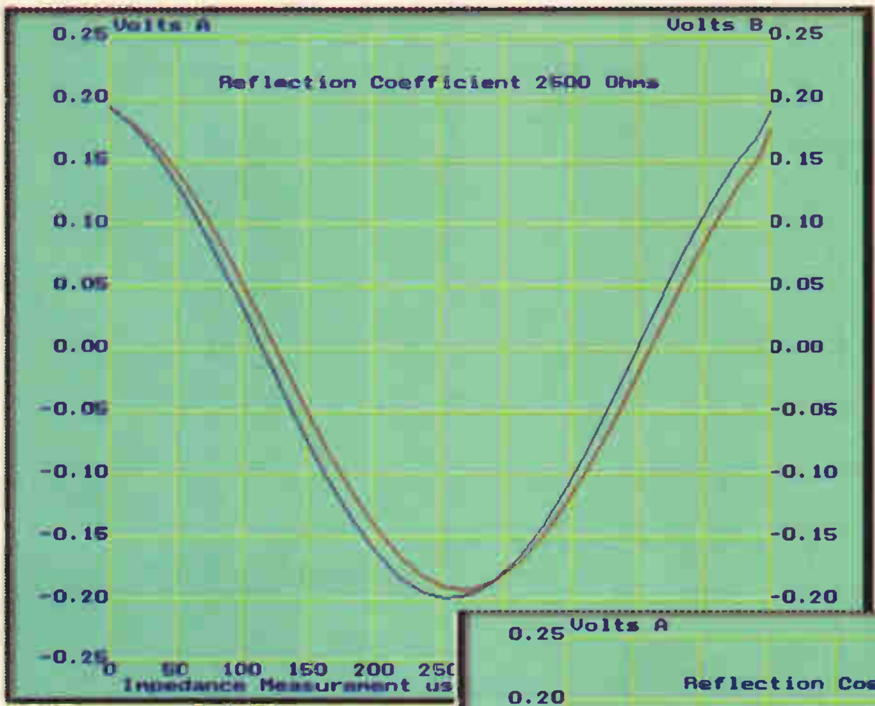


Fig. 2. With the bridge calibrated for equality with open load, this plot shows errors increase as reflection coefficient approaches unity. Measured values – reflection coefficient 0.9586, giving R=2349.8Ω. No error reduction was used. This plot also clearly shows Pico virtual oscilloscope has 10μs delay between channels – channel sampling.

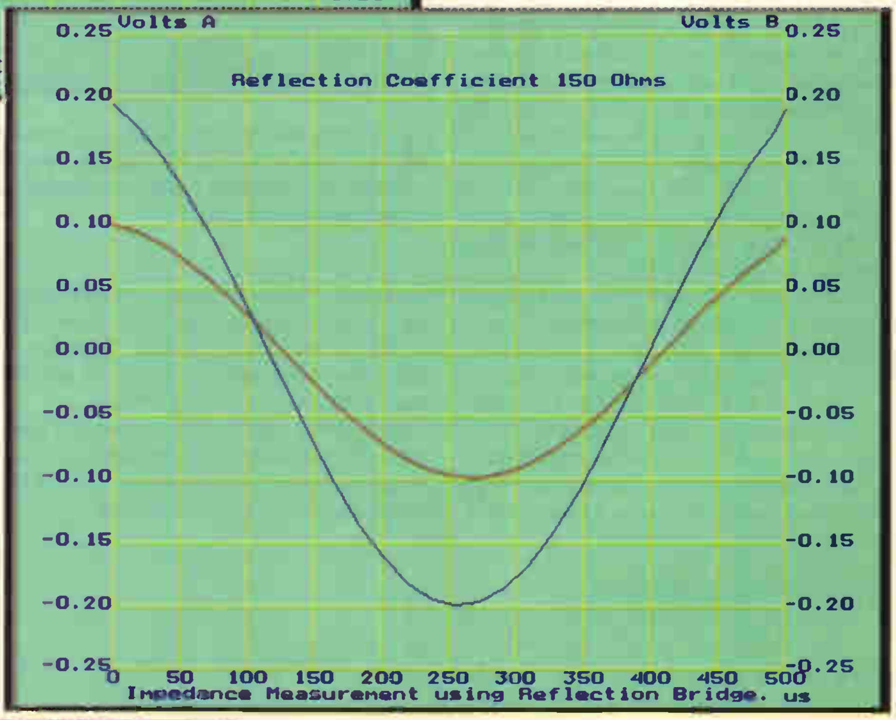


Fig. 3. With 150Ω load, in-phase reflection coefficient is visibly 0.500, giving R=150Ω. Shows this technique can be very accurate with impedances nearer the 50Ω nominal Z_0. Only open calibration used – no error reduction.

A smaller vector magnitude[1] represents a reduced capacitive reactance. When inductance is added in series with a capacitor, the equivalent or measured capacitance is increased. Conversely self capacitance added in parallel with an inductor decreases its measured or equivalent values. Consequently both capacitance and inductance measured values are frequency dependent.

These impedance measurement methods can be used from audio frequencies to a few megahertz. Reliability of readings at the higher end depends on the parasitics involved in connecting the measurement meters to the test device and the phase meter's upper frequency limit.

At low frequencies, the traditional Wheatstone bridge measurement performs extremely well, but due to internal parasitics in the bridge, is usually restricted to less than 1MHz. Additionally at 1MHz, 1m of test coaxial cable gives a phase shift of 1.8°. Since this cannot be fully compensated, makes tanδ or Q measurements of low-loss parts at 1MHz, extremely difficult.

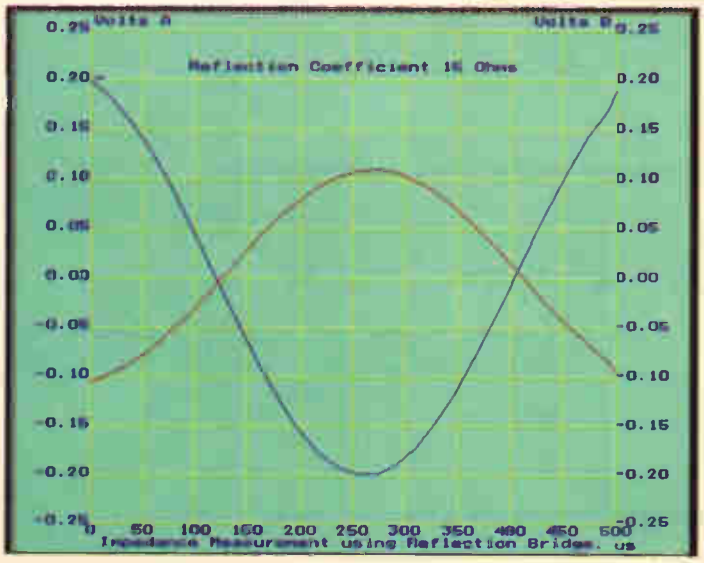


Figure 4. With impedances less than Z_0 the reflection coefficient is clearly phase reversed. Measured value of reflection coefficient = –0.54166 giving R=14.865Ω. All test resistors from Muirhead precision decade resistor box. Only open calibration used – no error reduction.

And measurements above 1MHz?

Impedance measurements at higher frequency require either expensive dedicated measurement systems, or the use of the vector network analyser techniques.

A vector network analyser overcomes the problem of test device connection by measurement using either a reflection bridge[6] or directional coupler. This is connected to the test device using low loss coaxial cable. Intermediate frequency superhet methods are used to down convert the test frequency, to overcome phase meter frequency limits.

Directional couplers must be electrically long and hence can operate only over limited bandwidths – typically one frequency decade. In practice, they are used only at the higher frequencies. A reflection bridge, made using a carefully wound balun is physically small, and can easily operate over four decades and at low or high frequency.

Traditional Wheatstone bridges are usually manually balanced for zero detector output. But Wheatstone bridges can be used over a limited impedance range without balancing, the resultant error voltage being measured and evaluated.

Designed for use in a 50Ω system and connected to the device under test using a low-loss coaxial cable, such a bridge can measure to high frequencies. Essentially this bridge reduces to three 50Ω precision, non-inductive, resistors together with a wideband 1:1 balun[7].

By changing the balun, this bridge configuration is usable to more than 1GHz or down to audio frequencies. There is more on this in the panel entitled Reflection bridges.

Reflections on transmission lines

Fundamental to a transmission line of any length, terminated by its characteristic impedance, is the complete absorption of all signals travelling towards this termination. Nothing is reflected, **Fig. 1**.

However if terminated other than by this characteristic impedance, regardless of line length[6], a 'reflected' signal is returned back along the line to the source. This signal is called the reflection coefficient since its amplitude cannot exceed that of the signal incident at the line end.

Reflection-coefficient magnitude, ρ, is measured relative to the incident signal and can vary in magnitude from 0 to 1. It is used in the polar notation together with an angle, as $\rho\angle\theta$.

When unterminated, i.e. terminated with an open circuit, the reflected signal's magnitude is identical to that of the incident signal. It is also identical in phase i.e. reflection coefficient is $1\angle0°$, **Fig. 2**.

When terminated with a short circuit, the reflected signal's magnitude is identical to that of the incident signal, with inverted phase so the reflection coefficient is $1\angle180°$.

If the termination impedance is resistive and exceeds the line's characteristic impedance, the reflected signal is identical in phase at the end of the line. The signal's magnitude depends on the termination mismatch, **Fig. 3**.

Should this resistive termination impedance be smaller than the line's characteristic impedance, the reflected signal is of opposite polarity, also with a magnitude dependant on the termination mismatch, **Fig. 4**.

Reflection bridges

A reflection bridge, or resistive coupler, resembles the familiar Wheatstone bridge. But while the Wheatstone bridge is normally balanced, the reflection bridge is not adjusted at all. The detector voltage is measured and evaluated instead.

A well designed bridge is usable over four decades of frequency. The detector voltage is described by the vector equation[7],

$$V = 0.125 \times V_o \times \frac{Z_x - Z_o}{Z_x + Z_o}$$

The detector voltage of the conventional Wheatstone bridge is floating and requires a balanced measurement, which is inconvenient at high frequencies. A reflection bridge overcomes this by using a balun transformer to convert detector voltage to an un-balanced output, measurable using conventional meters with coaxial cables.

Most commercial bridges are specified to have at least 40dB directivity over their usable frequency range, i.e. the ability to discriminate to 1% between the incident and reflected signals.

A reflection bridge resolves essentially to three precision resistors together with a precision wound 1:1 balun transformer of primary inductance. This inductance ideally gives 100 times greater impedance, at the required frequency, than the system characteristic impedance. The resistors have a value equal to the required characteristic impedance.

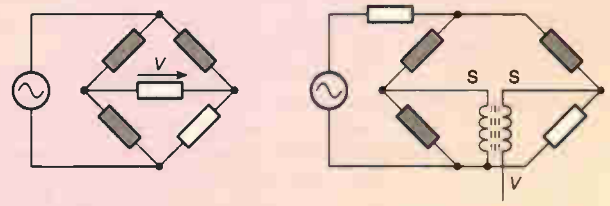
Within these limitations, bridges can be made for use at any desired impedance by amending the resistor values and frequency range. For audio frequencies for example, the balun could be selected for 4, 8 or 50Ω impedance.

The bridge used for the measurements described here was used without applying error correction techniques. It was designed to cover 1kHz to 1MHz frequency range at 50Ω impedance.

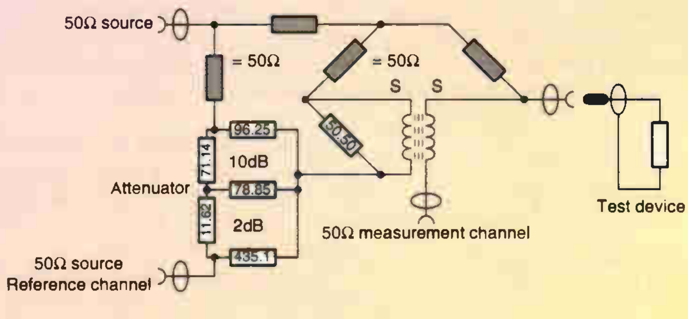
Resistor values used were the same as those of the Hewlett Packard *HP8721A* reflection bridge. The balun was wound with 125 turns of pre-twisted 32SWG wire on a very high permeability ferrite toroid measuring 35 by 22 by 15mm, giving a primary inductance of 240mH. The 2dB attenuator section was adjusted on test, to equalise the reference and measure channel amplitudes during open calibration.

For best accuracy of measurement[8], error reduction techniques can be used to compensate for bridge directivity, source and load matching. This requires the bridge be characterised at the measurement frequency by first measuring known open circuit, short circuit, and 50Ω impedances. These measurements resulting in the vector error factors for effective source match, effective load match and effective directivity, which are used to solve this vector equation, for each measurement and frequency,

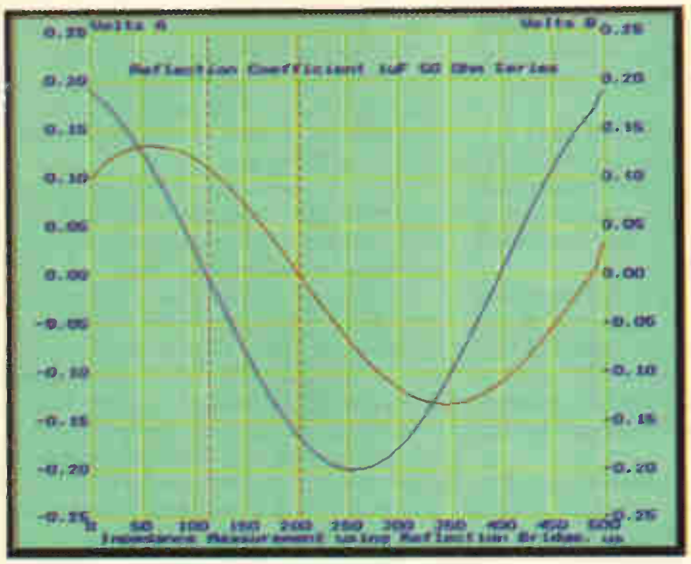
$$S_{11a} = \frac{S_{11M} - E_{DF}}{E_{SF} \times (S_{11M} - E_{DF}) + E_{LF}}$$



Wheatstone bridge, left, falls down at higher frequencies but the reflection bridge, right, will produce useful readings at rf.



Reflection bridge configuration used to produce the curves shown in this article.



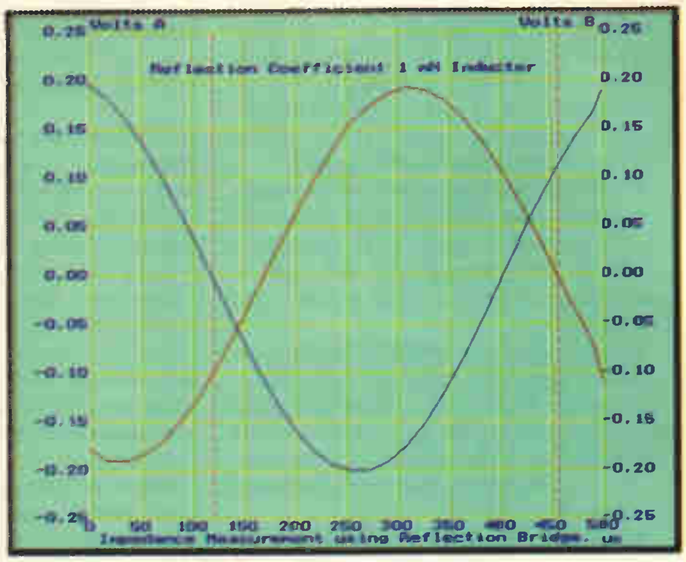


Fig. 5. Worked example 1, for 1μF polypropylene capacitor in series with 50Ω.
Reflection coefficient is 0.6665∠–48.919.
Polar to rectangular conversion Γ*x*=0.43797, Γ*y*=–0.502395.
Solve for *R*±j*X*, *R*=48.9Ω, *X*=–88.406Ω.
Solving for component value at test frequency of 1786Hz, equivalent series resistance=48.9Ω and capacitance=1.008μF

Fig. 6. Worked example 2, for a 1mH ferrite-core inductor with no added resistance.
Reflection coefficient is 0.955∠–205.087
Polar to rectangular conversion Γ*x*=–0.86491, Γ*y*=0.40491.
Solve for R±jX, R=1.208Ω, *X*=11.118Ω.
Solving for component value at test frequency of 1786Hz, equivalent series resistance=1.208Ω and inductance=0.9908mH.

Terminated with a pure capacitive or inductive load, the reflected phase is load dependent, and the magnitude equals the incident wave. Now the reflection coefficient is 1∠θ.

Phase angles calculated from time difference measurements, as in these examples, are read as being of negative phase, compared with the reference signal. Reflection Coefficient can be readily plotted on a Smith chart. But since the Smith chart uses negative angles to represent capacitive reflection coefficients and positive angles to represent inductive reflection coefficients, measured angles exceeding –180° should be normalised by adding +360°. Once plotted on a Smith chart[6], converted *R*±j*X* values are immediately available.

In rectangular notation the reflection coefficient is given the symbol Γ, having both a magnitude and phase value in the form, ±Γ*x*±Γ*y*.

Reflections to component values

In rectangular notation, the reflection coefficient can be converted into conventional component parameters of *R*±j*X*, simply by substitution in two standard equations. If measured using polar notation, the reflection coefficient must first be translated into rectangular notation[2], using a pocket calculator P→R function,

$$R = Z_0 \times \frac{1-(\Gamma x^2 + \Gamma y^2)}{1-2\Gamma x + \Gamma x^2 + \Gamma y^2}$$

$$jX = Z_0 \times \frac{2\Gamma y}{1-2\Gamma x + \Gamma x^2 + \Gamma y^2}$$

Having converted the measured reflection coefficient into *R*±j*X* format, any other desired impedance transpositions are simply performed. For more on this, see the panel Impedance transpositions and **Figs 5, 6.**

Measurement accuracy. At these frequencies, output from the test generator may be maintained constant. However, having passed through connecting cables and the reflection bridge, the test signal amplitude will suffer from variations, especially with test device loading.

These amplitude variations are compensated by performing all measurements relative to a portion of the generator output tapped off, using a two way splitter, for measurement by the reference channel.

Connecting this bridge to the test device at high frequencies is made easy by its inherent ability to discriminate between two signals travelling in opposite directions. Consequently the cable used to supply the test stimulus to the device under test also conveys the reflected error signal which is measured.

The whole system, including connecting cables, is calibrated using known open circuit, short circuit and matched 50Ω termination. These must be connected at the exact point where the device-under-test is to be connected – known as the reference plane – using the frequencies to be measured.

Accuracy enhancement. Obviously, this reflection bridge has limited discrimination and isolation of the forward and reflected signals. Typically, its directivity exceeds 40dB, making it better than 100:1.

The calibration routines using known open, short and 50Ω loads also allow the system errors to be measured. As a result, mathematical error reduction techniques can be used to correct for these deficiencies[8]. There is more on this in the panel Reflection bridges.

In addition simple techniques also exist to measure then correct for[2] and remove errors introduced by any test jig used to house the test piece.

Measuring in practice

Using these techniques together with the Hewlett Packard *8753* vector network analyser and coaxial test jig *HP16091A*, I have made many measurements of surface-mount miniature high-Q ceramic chip capacitors. These measurements have been made from 1MHz to 3GHz, with excellent repeatability between results.

Such high frequency measurements complement the lower frequency measurements, made using precision *LCR* meters. ■

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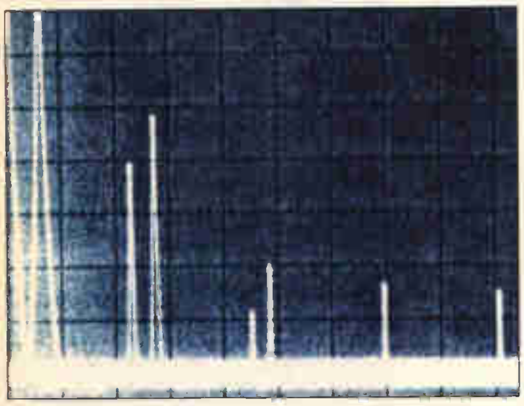
Understanding phase noise



Oscillator purity is an increasingly important factor for designers of communications equipment. Ian Hickman has been investigating phase noise in rf oscillators and has discovered a problem that seems to have gone unnoticed.

Modern wireless communication often uses one or other of the various types of digital modulation.

The earlier, simpler forms, such as basic dpsk, or binary phase shift keying, are relatively robust, requiring only a modest signal-to-noise ratio at the receiver to guarantee successful reception. But shortage of spectrum space spurred the search for greater bandwidth efficiency. This lead first to the development of variations on the theme of qpsk – quadrature phase shift keying – which conveys two bits of information per signal element or 'symbol'.

Later, more exotic forms, such as 16psk, 64apk and even 256apk appeared, carrying four, six and eight bits per symbol respectively.

At the receiver, the demodulator must effectively measure the phase difference between successive symbols. At the transmitter, this starts out as 0 or 180° – in the case of asymmetrical dpsk – or only + or –90° in the symmetrical form. But by the time the signal is received, noise and interference will have eroded the available phase margin, possibly leading to bit errors.

With asymmetrical form qpsk, the phase change between symbols is 0, + or –90 or 180°, or +45, +135, –45 or –135° in the symmetrical case, sometimes called as '$\pi/4$ qpsk'. So a higher signal to noise ratio at the receiver is required for the same bit error rate.

With the advanced forms of modulation mentioned earlier, the phase change from one symbol to the next may be only $22^1/_2$ degrees or even less, so clearly an even greater signal-to-noise ratio is required for an acceptable bit error rate.

Noise in the receiver.

Atmospheric noise and interference are not the only problems a digital data receiver faces. While an hf receiver with a reasonably efficient aerial is likely to be 'externally noise limited', at vhf and even more so at uhf and microwaves, external noise is so low that reception will usually be limited by the receiver's own noise.

In this context, one usually thinks of input stage noise. But in the reception of digital phase modulation, an important contribution to the factors eroding the essential phase discrimination, on which a low bit error rate depends, is the phase noise of the local oscillator.

Ideally, an oscillator produces an isolated spectral line, with zero energy output at any other frequency. Of course, there will be some harmonic content, but this is usually unimportant in a well designed receiver. Much more troublesome is energy at frequencies immediately adjacent to the oscillator output. This takes the form of noise sidebands, which can be quite large at very small offsets from the oscillator frequency, falling off at greater offsets, until at frequencies well removed from the carrier, their level bottoms out at the oscillator's far-out noise floor.

Why is phase noise important?

The sidebands consist of a mixture of amplitude-noise and phase-noise. In a receiver local-oscillator application, the amplitude noise sidebands are usually unimportant, since the local oscillator output is applied to the mixer at a high level: the local-oscillator input of the mixer thus operates in a heavily compressed mode. So minor level changes – even of a decibel or so – would have negligible effect. But the local-oscillator phase noise is quite a different story.

The intermediate-frequency signal reflects the phase difference between the rf signal input and the local-oscillator drive waveform. Thus local-oscillator phase noise adds linearly to phase disturbances of the wanted signal. These include noise, interference and multipath suffered in the over-the-air path, and front-end noise due to a marginal signal level.

The over-the-air path is outside the receiver designer's control; he can only concentrate on the other factors, of which – in a digital data receiver – oscillator phase noise is a major component.

Phase noise from the local-oscillator

In special cases of fixed frequency operation, a receiver's local oscillator may be a crystal oscillator. Such an oscillator is characterised by extremely low levels of sideband noise – which is usually denoted by $\mathcal{L}(f_m)$ and defined as the noise power in a 1Hz bandwidth at an offset of f_m. But usually, the local oscillator will be an *LC* type, and these exhibit a higher level of sideband noise, extending out much further on either side.

To highlight the difference, note that a good crystal oscillator may show a level of sideband noise, $\mathcal{L}(f_m)$, which is already down to –140dBc at only 10Hz offset from the carrier. By contrast, a commercially advertised varactor-tuned voltage-controlled oscillator module, covering the range 100 to 200MHz, claims a typical $\mathcal{L}(f_m)$ of –105dBc at 10kHz offset, and around –120dBc at 100kHz offset.

Where the *LC* oscillator forms the voltage-controlled oscillator in a phase-locked loop, its sideband phase noise within the loop bandwidth will be reduced by the loop negative feedback. Outside the loop bandwidth it returns to the level it would be were the voltage-controlled oscillator running open loop.

Clearly, even given a degree of phase-noise clean-up by the loop, one is better off starting out with a low phase-noise oscillator in the first place. A facility for measuring the phase noise of an oscillator is therefore an important item in any rf development laboratory, and can involve some very expensive equipment. For this reason, I was interested in an article which described such a measurement system using only standard laboratory instruments plus some inexpensive rf instrumentation[1]. The basic arrangement is shown in **Fig. 1**.

B is for bottom

I wanted to try and measure the phase noise of an oscillator, in order to settle a question which has interested me for some time. This question is, "is there an advantage in designing an *LC* oscillator in such a way that the transistor does not bottom at the negative-going peaks of the waveform?"

In fact, many *LC* oscillator designs do result in the transistor bottoming. This can be quite difficult to avoid in an oscillator with a wide tuning range, such as a three-to-one frequency ratio, given production spreads in transistor characteristics. The effects of bottoming in an rf oscillator had been explored in an earlier article[2], but equipment to measure phase noise was not available to me at that time.

I built an *LC* oscillator, operating at around 10MHz. Relative to vhf, this frequency is easier to measure. This, together with the other items needed for the Fig. 1 type set-up, is shown in detail in **Fig. 2**.

Tank circuit inductor L_1 was a Coilcraft

Working out sideband noise

In the lower trace of Fig. 5b), the measured level of sideband noise at 2.5kHz offset from carrier, with the circuit of Fig. 2, is –108dBV in a 30Hz measurement bandwidth. To work out $\mathcal{L}(f_m)$, the value in a 1Hz bandwidth is needed.

The analyser's intermediate-frequency filters consist of five synchronously tuned crystal filter stages, providing a Gaussian response. This characteristic is optimum for rapid settling to the true value of a swept signal. The noise bandwidth of such a filter is 12% greater than the actual –3dB bandwidth. The nominal 30Hz bandwidth is subject to a ±15% tolerance, so the actual –3dB bandwidth was measured, using the 1dB/division scale. It turned out to be 27Hz, giving a noise bandwidth of 30Hz, as near as makes no odds. Thus the level of –108dBV in 30Hz translates to –123dB in a 1Hz bandwidth. This represents the sum of the noise energy in both upper and lower sidebands, giving a figure of –126dBV or 0.5μV for the single sideband noise.

Given the measured sensitivity of the frequency discriminator of 6.1μV/Hz (see below), the rms frequency deviation f_d is 0.082Hz. For sinewave modulation at a frequency f_m, the modulation index $m=f_d/f_m$ equals the peak phase deviation in radians. Now $0.082/2500=3.3.10^{-5}$ radians, and for such a small phase deviation, only the first order fm sidebands are significant.

So if the modulating frequency f_m were a 2.5kHz sinewave rather than narrow band noise, the first order sidebands would each be $(3.3.10^{-5})/2$ in amplitude relative to the carrier, since for small angles, arctanθ=tanθ=sinθ=θ, with negligible error. So the sinewave single sideband amplitude would be simply $20\log(1.65.10^{-5})$ relative to the carrier, or –96dBc. This may be taken as a first order approximation to the value of $\mathcal{L}(f_m)$ at 2.5kHz offset, for the circuit of Fig. 2.

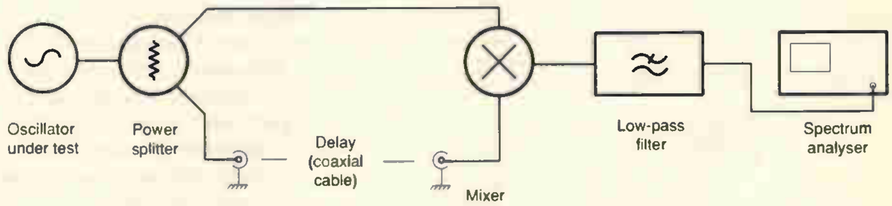


Fig. 1. Block diagram of a set-up to investigate oscillator phase noise.

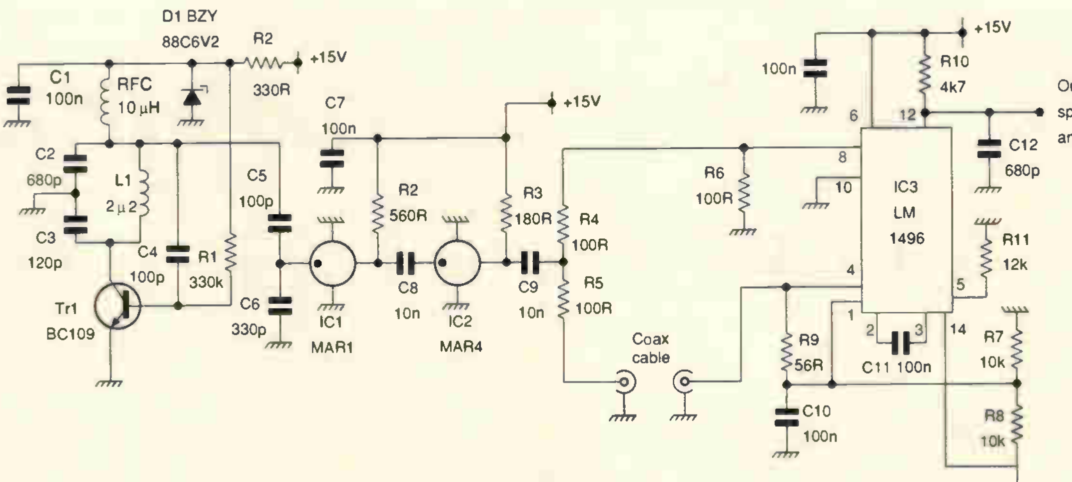


Fig. 2. Circuit diagram of an experimental set-up to measure oscillator phase noise.

SLOT-TEN-1-03 unshielded inductor with a carbonyl E core, having a quoted nominal inductance of 2.2μH and Q of 56 at 7.9MHz. I chose a Colpitts oscillator, as the inductor was untapped, arranged so that the transistor could be operated with the emitter connected directly to circuit ground.

To minimise loading and to maintain a reasonably high working Q, the output was taken from the base end of the tank circuit. This is a much lower impedance point than the collector end, and loading was further reduced by the using a capacitive divider, C_5 and C_6, to buffer the 50Ω input of IC_1. Together with IC_2, IC_1 provides a total gain of 26.7dB nominal, providing a level of –8dBm into 50Ω at the coaxial socket connected to R_5.

Frequency discriminator

The output of IC_2, which sees approximately 50Ω loading, is applied to the local-oscillator port of an active double balanced mixer, IC_3, an *LM1496*. **Figure 3a)** shows the internal circuit of IC_3. The 'carrier' or local oscillator is applied between pins 8 and 10, to four transistors connected in an arrangement often referred to as a Gilbert Cell. The signal input is applied between pins 1 and 4, the signal being steered in phase or in antiphase to the outputs at pins 6 and 12 (note the pin numbers quoted refer to the DIP packaged version of the *LM1496*). The transconductance of the signal long-tailed pair is set by the value of a resistor connected between pins 2 and 3. The magnitude of the tail currents is set by the current injected into the bias port, pin 5.

Figure **3b)** shows how the output at pin 12 is at its maximum positive level if the local oscillator and signal are in phase, is at zero (relative to its level in the absence of a signal input) when they are in quadrature, and at maximum negative level when in antiphase.

If pins 2 and 3 are shorted, so that both signal and local oscillator ports are overdriven – equivalent to squarewave drive in each case – the input phase to output voltage characteristic is linear, as in Fig. 3b), right hand side. If the signal port is operated in a linear manner, the characteristic is cosinusoidal, also shown in Fig. 3b).

In Fig. 2, the signal is applied to the signal input port via R_5 and a length of coaxial cable. The latter provides a fixed time delay, independent of frequency. Therefore if the oscillator frequency is varied, the electrical length of the cable varies, and so the phase of the signal applied to pin 4 of IC_3 will vary. So although IC_3 is a phase sensitive detector, in conjunction with the delay cable it forms a frequency discriminator.

The delay was provided by a reel of miniature polythene insulated coaxial cable, unearthed from my stock of handy bits and pieces. This cable had a silver on copper on steel inner, and could well have been UR94.

Monitoring pin 4 of IC_3 with one 'scope probe and the junction of R_4 and R_5 with the other. the waveforms were found to be in quadrature at 10.377MHz and in antiphase at 9.726MHz.

From these results, and assuming the velocity of propagation in the cable is two thirds that in free space, some simple algebra gives the length of coaxial cable as 14.95 quarters of a wavelength at 10.377MHz, say $3^3/_4$ wavelengths, allowing for experimental error. Thus T_d is 361ns and the physical length of the cable turns out to be 108.4m. I took this figure on trust, rather than unreeling the cable to find out.

Frequency discriminator sensitivity

Maximum sensitivity is ensured by C_{11} which provides an ac short between pins 2 and 3. As the dc resistance between these pins is infinite, in the absence of a signal input, the output sits at the midpoint of the characteristic, despite any small input offset voltage that there might be between pins 1 and 4.

By varying the tuning with the core of L_1, measuring the frequency with a digital frequency meter and the output level at pin 12 of IC_3 with a digital voltmeter, the frequency discriminator characteristic was measured. This is shown plotted in Fig. **3c)**. Due to the limited available tuning range, for the most part, only one side of the characteristic could be plotted, as shown.

The considerable length of coaxial cable used achieved a high sensitivity in the frequency discriminator, but introduced some inevitable attenuation. Consequently the signal voltage swing available at pin 4 was less than the local-oscillator input at pin 8, the attenuation in the cable being some 7dB. The result is that the discriminator characteristic is intermediate between those shown in Fig. 3b).

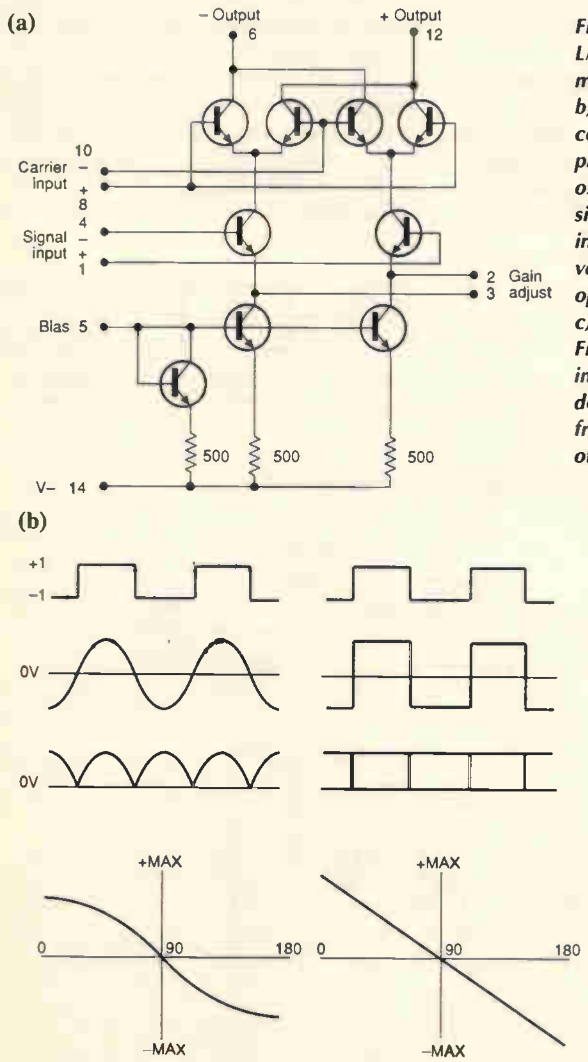
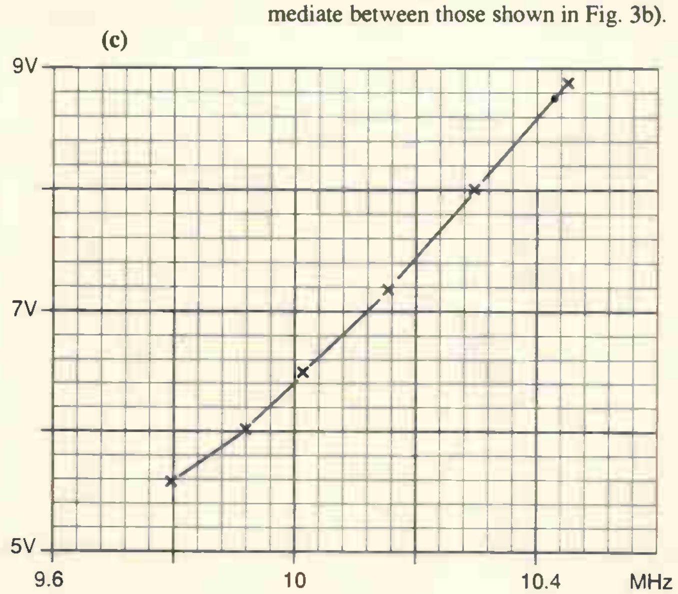


Fig. 3a) Internal circuitry of the LM1496 active double balanced mixer.
b) Showing response of the dc component of output voltage to phase changes between local oscillator and signal inputs, for sine and square wave signal inputs of equal peak-to-peak voltage, assuming linear operation of the signal port.
c) Showing the response of the Fig. 1 frequency discriminator, as implemented in Fig. 2. The black dot shows the measured centre frequency response, crosses show other measured points.



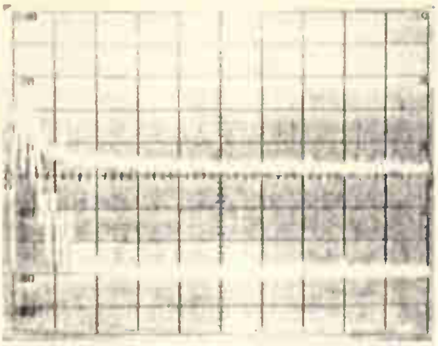
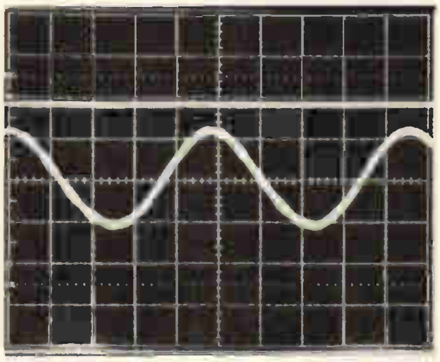


Fig. 4a) Spectrum analyser sweep, 0 to 5kHz, reference level (top of screen) –60dB, 10dB/division vertical, intermediate-frequency bandwidth 30Hz, smoothing maximum, 100s/div sweep speed. Lower trace, with + and –15V supplies off. Upper trace, supplies on, circuit as in Fig. 2.



b) Oscilloscope traces; horizontal, 20ns/division. Upper trace, IC_3 pin 12, 5V/division, 0V at centre line, with coaxial cable disconnected. Lower trace, IC_3 pin 4, 50mV/division ac coupled, coaxial cable connected.

Over the central linear portion, the characteristic sensitivity is 164kHz/V or 6.09μV/Hz.

Measured results

Output of the frequency discriminator, at pin 12 of IC_3, was connected to an *HP3580A* lf spectrum analyser, via the low pass-filter shown in Fig. 1. Figure 2 shows that the filter consisted simply of the 4.7kΩ phase detector output resistor R_{10}, in conjunction with some 800pF or so. This consisted of C_{12} plus about 100pF due to a screened input lead and the analyser's input capacitance. The cut-off frequency of this filter is a little over 40kHz, well clear of my range of interest, which was in noise sidebands up to 5kHz.

First of all, to establish a measurement noise floor, a spectrum analyser sweep from 0 to 5kHz was recorded with the power supplies switched off, **Fig. 4a)**, lower trace. This shows a measurement noise floor of about 80dB below a top-of-screen reference level of –60dBV, or some –140dBV. At this level, it is difficult to avoid some response from supply rail residual hum, visible as 100Hz and its harmonics at the left hand side of the trace.

Next, the circuit was powered up, but with the coaxial cable disconnected. Figure **4b)** upper trace 5V/division. 0V at centre line, shows the standing voltage at the frequency discriminator output, IC_3 pin 12. This was +8.75V, corresponding to the discriminator centre frequency.

Next, the coaxial cable was reconnected and the lower trace (50mV/division, 20ns/division ac coupled) shows the delayed signal applied to IC_3 pin 4. Some modulation of the trace is visible, but this was still there when the supplies were turned off – it turned out to be pick-up of the local fm radio station. As the frequency is unrelated to the local-oscillator waveform at IC_3 pin 8, it will not affect the result and can be safely ignored.

With the coaxial cable reconnected, the frequency was adjusted to 10.377MHz, by means of the core in L_1. At this frequency the signal input at pin 4 of IC_3 was in quadrature to the local-oscillator input at pin 8, corresponding to zero deviation from the discriminator's centre frequency. This is indicated by the black dot in Fig. 3c).

The oscillator's phase-noise sidebands (on both sides of the carrier) are translated by the frequency discriminator to baseband – from zero hertz upwards. The result is displayed in Fig. 4a), upper trace. This is over 30dB clear of the measurement noise floor, due to the high system sensitivity ensured by the generous length of coaxial cable used.

The corresponding value of $\mathcal{L}(f_m)$ at 2.5kHz offset was calculated as shown in the panel. The result seems plausible, even if only an approximation. However, for the purposes of comparing phase noise with the transistor bottoming, or not bottoming, comparative measurements suffice, and proved revealing.

I needed to know whether the oscillator was bottoming or not. An *HP8558B* spectrum analyser was used to sample the output at the base end of L_1. To avoid excessive loading of the circuit, the 50Ω coaxial lead to the spectrum analyser was connected via a 4.7kΩ resistor.

Figure 5a) shows the spectrum of the oscillator, with settings of 10dB/division vertical, reference level –10dBm, 5MHz/division horizontal, 30kHz intermediate-frequency bandwidth, video filter on maximum. The illustration is a double exposure, showing the output of the circuit as in Fig. 2 (0Hz marker at extreme left), with the fundamental at just over 10MHz and its second harmonic nearly 30dB down. The second trace, with increased Tr_1 base current (offset half a division to the right), shows a larger fundamental and prominent third and fourth harmonics in addition to the second.

The second trace is the result of connecting a 56kΩ resistor in parallel with R_1. Thus the base current was increased by a factor of over six, while the output amplitude increased only by some 8dB or times 2.5. This, together with the marked level of higher harmonics, shows that with the additional base current the circuit was bottoming, but without it was not.

Figure **5b)** shows (upper trace) the 0 to 5kHz baseband spectrum, with the increased base current, resulting in the transistor bottoming. The lower trace is a repeat of the upper trace in Fig. 4a), for comparison.

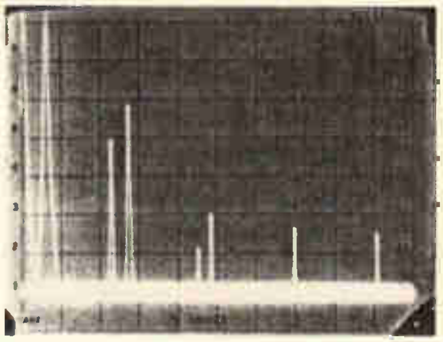
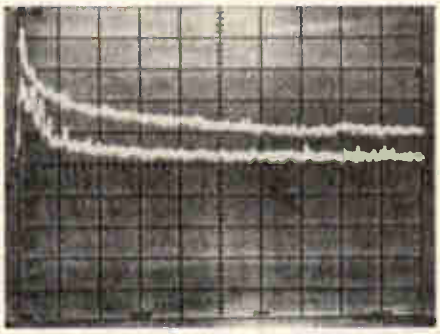


Fig. 5. a) Spectrum of the oscillator, with settings of 10dB/division vertical, reference level –10dBm, 5MHz/ division horizontal, 30kHz intermediate frequency bandwidth, video filter on maximum. Double exposure. Circuit as in Fig. 2 (0Hz marker at extreme left) shows the fundamental at just over 10MHz and its second harmonic nearly 30dB down. Trace with increased Tr_1 base current (offset half a division to the right) shows larger fundamental and prominent third and fourth harmonics.



b) Upper trace, 0 to 5kHz baseband spectrum, with increased Tr_1 base current, transistor bottoming. Lower trace, repeat of the upper trace in Fig. 4a), for comparison. Both use same settings as Fig. 4a).

Both traces were recorded with the same settings as Fig. 4a). For this test, care was taken that the signal applied to the frequency discriminator was the same as it was without the increased base current. To this end, after adding the 56kΩ resistor in parallel with R_1, the 100pF capacitor C_5 was replaced by a 5 to 65pF trimmer. This was adjusted to give the same amplitude inputs at the local oscillator and signal ports of IC_3 as previously.

The resultant small shift in oscillator frequency, due to the slightly reduced loading on the tank circuit, was removed by readjusting the core of L_1.

Findings

You can see from Fig. 5b) that in the range above 2.5kHz offset, the magnitude of the phase noise relative to the carrier is nearly 10dB lower when the transistor is not bottoming, than when it is. Note particularly. that the gap widens at lower offsets. This is presumably because bottoming involves higher

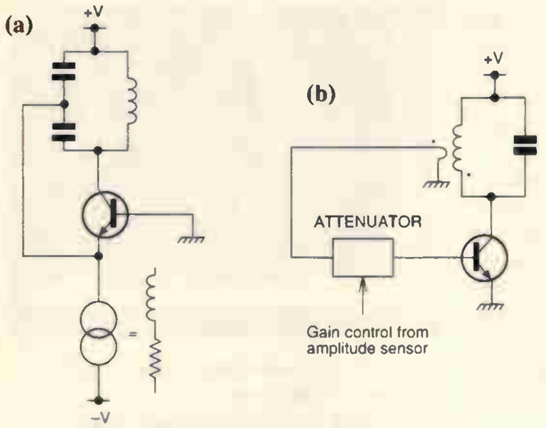


Fig. 6a). Defining the transistor's collector current. By means of a long tail as here is just one of many ways. The resistor may be replaced by the output of a d-to-a converter, permitting adjustment of the tail current under program control.
b) Separating the amplitude control mechanism from the oscillator should permit operation of the transistor in a linear regime. This should result in much reduced phase noise sidebands, by preventing the transistor's $^1/_f$ noise cross-modulating onto the carrier.

order non-linearities, resulting in the transistor's $^1/_f$ noise, cross modulated onto the carrier, effectively extending further out into each sideband.

By 5kHz, the noise, as measured with a frequency discriminator, has clearly flattened out. This corresponds to phase noise falling at 6dB/octave of offset frequency, or the f^{-1} region of phase noise, which continues until the far out noise floor is reached.

At smaller and smaller offsets, the slope becomes greater, f^{-2}, f^{-3} and at very small offsets f^{-4}. This tendency is visible in both traces in Fig. 4a), though setting in at a higher frequency when the transistor is bottoming. As the offset reduces to zero, the amplitude increases, up to the value of the carrier output. The trace in Fig. 4a) does not show this below 5Hz, as this is the low frequency limit of the *HP3580A* spectrum analyser. In any case, the output due to the carrier itself is (near) zero, since the local oscillator and signal inputs are in quadrature.

So when an oscillator with low phase noise is required, a circuit design should be selected which avoids bottoming of the collector. This can be achieved in a number of ways, for instance using a 'long tail' to define the emitter current, **Fig. 6a)**.

Where a large tuning range is involved, it may be advantageous to vary the tail current. Assuming capacitive tuning, the dynamic resistance of the tank circuit will increase with frequency. So to maintain a constant amplitude of oscillation, the tail current should be varied inversely as the oscillator frequency.

Of course, even when not bottoming, the transistor is still operating non-linearly, the collector current being cut off for part of each cycle. If amplitude control could be implemented independently of the transistor, as indicated in Fig. **6b)**, it should be possible to operate the transistor entirely in a linear mode, preventing the cross modulation of its $^1/_f$ noise onto the carrier output. This is an interesting possibility and one which I intend to pursue. Doubtless this has been done many times already, but I don't recall having seen any published results.

An alternative to Fig. 6b) would be to use a variable gain amplifier as the maintaining amplifier. A suitable candidate would seem to be the recently announced *CLC5523* from National Semiconductor, which I am trying to obtain a sample of. It has a 250MHz bandwidth at 135mW power consumption. ■

References

1. Suter, W. A., 'Phase Noise Measurement for Under $250' *RF Design*, Sep. 1995, pp. 60-69
2. Hickman, I., 'The ins and outs of oscillators' *Electronics World*, July 1994, pp 586-589.

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Hands-on Internet

If you have ever done a search on the word filter and found references to oil, air water and coffee, find out about new and more focussed search solutions. Cyril Bateman has also uncovered yet more circuit design tools worth investigating.

A traditional search engine can respond with huge numbers of references needing careful sifting to reveal pertinent resources. When seeking specialised engineering data, a limited-area search engine can be much more useful. This month, I draw attention to two development programs based in UK universities that are actively building up Internet systems. The projects are funded as part of the £15 million grant for the 'Electronic Libraries access to networked resource program'.

The Edinburgh Engineering Virtual Library[1] at Heriot-Watt, claims to be *the* UK gateway to engineering information on Internet. It provides a database of UK engineering resources and a speedy search engine for these UK Web sites together with all their Internet 'links'. A recent search on 'emc' and 'filter', quickly identified many Web references with a synopsis of their page content. Each one is topical and with not one irrelevant link to oil, air, water or coffee filters as found with traditional search engines.

Ukoln, based in the main library at the University of Bath[2], hosts the 'eLib' electronic libraries programme. It also provides a mirror of the 'Dlib' magazine, which reports the progress of the digital library project. Full details of the programs and contributors can be found in the information pack, which can be also be requested by telephone from the University of Warwick,[3] **Fig. 1**.

The ukoln site hosts the Web version of Ariadne,[4] available also as a printed magazine. Issue eight includes an Internet engineering resources overview, written by Roddy MacLeod, project manager information for the EEVL eLib project at Heriot-Watt. This section alone justifies downloading Ariadne.

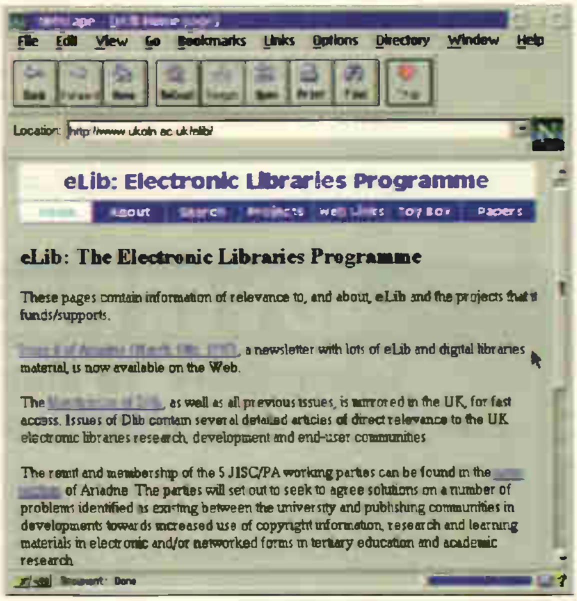


Fig. 1. The UK Electronic Libraries Project, funded by the Joint Information Systems Committee.

Where to look

1. Edinburgh Engineering Virtual Library, http://www.eevl.ac.uk
2. eLib Electronic Libraries Program, http://www.ukoln.ac.uk/elib
3. eLib Information Pack, Yanna Dandolos, 01203 538137.
4. Ariadne the Web version, http://www.ukoln.ac.uk/ariadne/issue8
5. Centre for Electronic Texts in Humanities, http://www.ceth.rutgers.edu/newhome.html
6. The Oxford Text Archive, http://sable.ox.ac.uk/ota
7. Communicata Electronics, http://www.communicata.co.uk
8. Nexor Ltd, http://www.nexor.co.uk/public/aliweb
9. The Institution of Electrical Engineers, http://www.iee.org.uk
10. TDS-NET Components Database, http://www.tds-net.com
11. EMCnet. http://www.emcnet.com/serv06.htm
12. Spice2: A Computer Program to simulate Semiconductor Circuits, ERL-M250 9May1975 UCA Berkeley CA94720
13. Helsinki University of Technology, http://www.aplac.hut.fi/main.html
14. F.Barbosa Project, http://www.inaoep.mx/~fbarbosa/project/macros
15. Teccor Electronics Inc, http://www.teccor.com/spice.htm
16. Intusoft, http://www.intusoft.com

While the core operation of Internet has traditionally been supported by computing students, the electronic text provision has been provided by the humanities. Inevitably they had favoured classic literature, but this is now slowly changing. The US centre for electronic text in humanities, based at Rutgers[5], offers links to many sites. One popular UK site, the Oxford Text Archive[6], has been actively collecting electronic text archives for over twenty years.

Communicata.co.uk[7] is a forum for the latest electronic design and development information. It references Web publications of current topics and welcomes both experienced and novice design engineers. Unlike many sites, Communicata does not copy existing Web information. Instead it sets out to provide easy access to information by targeting only electronics and mechatronics interests. Its 'lookover' directory and product locator tools will prove especially useful as they develop further.

Nexor.co.uk,[8] based at Nottingham, has a more conventional approach. Its Aliweb database was founded in 1993 as an early attempt to overcome difficulty in finding information on the rapidly growing Web. To avoid using up Web resources which would be needed to visit every page just to gather information, Aliweb proposed that Web page providers fill in a standardised description form which could be automatically incorporated into its database. This database, now mirrored on many sites, can be quickly and easily searched.

The 'Inspec' database located at the Institution of Electrical Engineers[9] is a vast resource, more than five million references, which can be searched in minutes. It claims to be the leading English-language service providing access to scientific and technical literature. The IEE site also has some 1000 conventional Web pages, many freely searchable. Access to the Inspec database is chargeable, costing some £20-£40 for a typical search.

Public access is possible via a CompuServe account. Simply 'go-Iquest' or 'go-KI', when the fees incurred will be charged to your account. A number of Lotus Screen Cam movies demonstrating the Inspec database, can be downloaded from IEE free of charge, **Fig. 2**.

The TDS-NET[10] semiconductor database is updated weekly. It contains more than 186 000 electronic component references from 32 manufacturers and more than 96 000 pages of data in PDF format from 119 data books. This $1 a week subscription service allows you to search for both discrete devices or integrated circuits, by name, function, or electrical characteristics. Application notes can be similarly searched by subject, component reference or manufacturer. Perhaps most useful of all, it can be searched to find replacement parts for obsolete discrete semiconductors, **Fig. 3**.

Currently very topical, the EMCnet[11] site was accessed following a search of the EEVL database at Heriot-Watt. EMCnet provides two functions, a limited-area search tool also links into simulation and computations. It has a fine index of relevant books for all aspects of EMC compatibility, including use of Finite Element Analysis simulation methods to solve EMC problems, **Fig. 4**.

Simulation software

The public domain 'Spice2' software, developed for designing integrated circuits, was developed in 1975 at the University of California at Berkeley by Laurence Nagel[12] for his doctoral thesis. While many commercial circuit simulators for pcs have been derived from his work, other viable simulation engines exist.

In 1972 Professor Valtonen started work on a simulation project called APLAC at the Helsinki University[13] of Technology. This work, since 1988, has been supported by the resources at Nokia Research Centre and Nokia Mobile Phones.

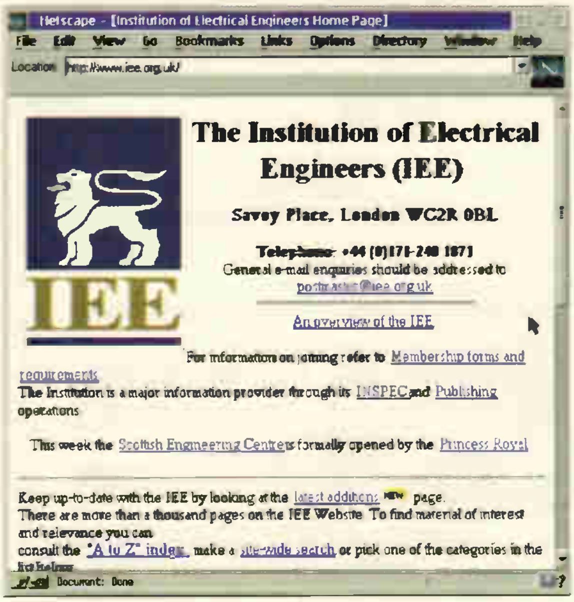


Fig. 2. The Institution of Electrical Engineers vast Internet site. Home of the IEE's Inspec database and much more.

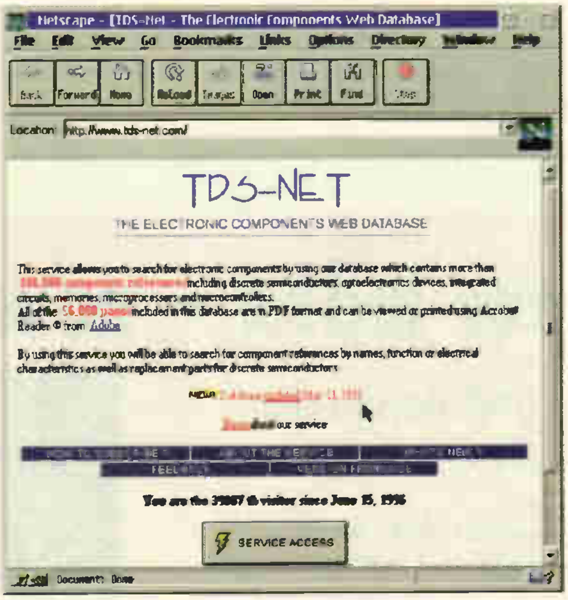


Fig. 3. Possibly the largest semiconductor archive at one location. Subscribe to download your data and application notes in PDF format.

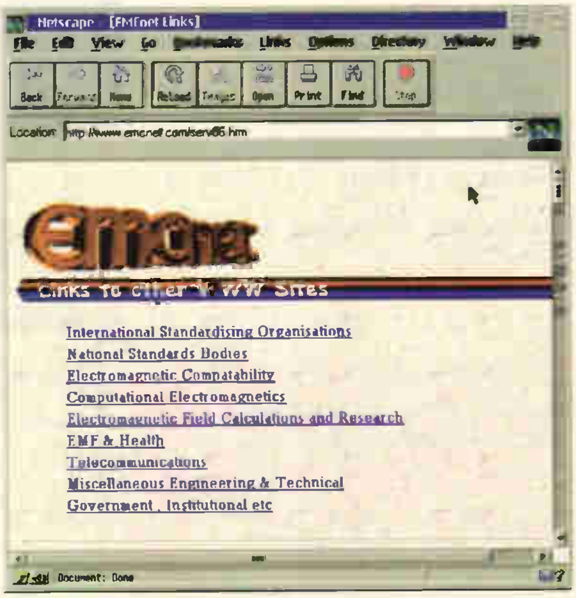


Fig. 4. All emc topics under one roof. EMCnet is dedicated to this topical and troublesome topic.

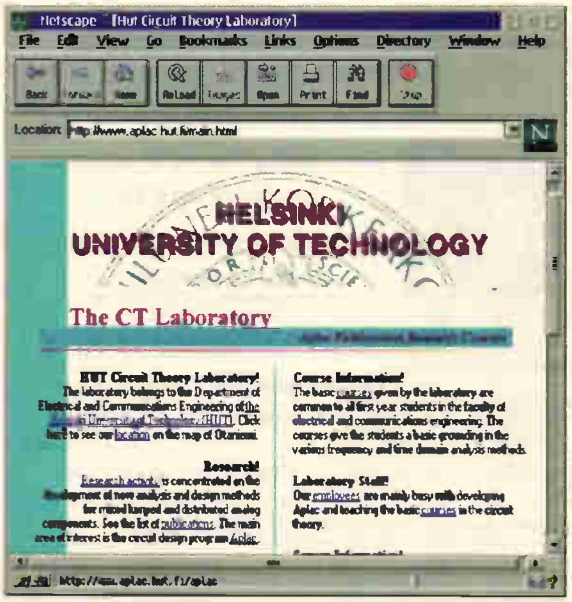


Fig. 5. The HUT circuit-theory laboratory – Helsinki University. Probably the most comprehensive circuit simulation tool available for the pc.

APLAC is a unique circuit simulator, having four major enhancements not found in Spice based engines. But it can also use Spice macromodels as well as its own much enhanced versions. It seeks to give the user full freedom in solving design problems, without the restrictions of conventional simulation tools.

This simulator provides both conventional electronic simulation with optimisation and simultaneous electrothermal simulation, or self heating, capabilities for both semiconductors and passive components. Its transient analysis uses convolution techniques to provide the correct treatment of components having frequency-dependent characteristics.

Finally it incorporates a finite difference time domain, simulator, or fdtd, for solving electromagnetic 3D field problems either independently or as part of the circuit design. All this added capability makes APLAC a uniquely competent, comprehensive yet user friendly circuit design tool, valid from audio to microwave frequencies.

APLAC's companion schematic editor, NASSE, makes net list generation for this large and unique simulator extremely simple. It also has the best method yet seen for 'wiring-up' components. A fully working but size limited, demo version for Windows3x, which is complete with full user manuals, is available in an 8Mbyte download, **Fig. 5**.

Before any simulation can be performed, macromodels for the desired component parts must be obtained. A unique, on-the-page, automatic macromodel generator program can be found on the F. Barbosa page[14]. Input your desired parameters into the boxes provided to obtain a Spice compatible macromodel.

At present this facility is restricted to providing macromodels for operational amplifiers, continuous time domain transfer functions, two to sixth-order filters and voltage controlled oscillators.

Finally, Teccor Electronics[15] Inc offers spice models for its range of specialist silicon controlled rectifiers and triacs. These can be downloaded from their home page, or from CompuServe on the CADDVEN forum and the Intusoft[16] Web page. ■

Designer power-supplies

Walter Gray's linear power supply designer is simple and written in Qbasic. Nevertheless, it takes into account transformer winding resistance and produces a comprehensive list of ratings.

Most electronics textbooks manage to discuss the basic mains transformer-rectifier-capacitor power supply without actually telling you how to calculate the output voltage. If you want to calculate the psu output before building it then the methods available are either graphical, such as the one referred to at the end of this article, or a simulation package, assuming you can find all the required model parameters.

My short computer program, described here, allows you to calculate pretty well everything you will need for a practical power supply design. It is written in Qbasic, making it accessible to anyone with a reasonably modern pc.

Structurally, the program is extremely simple, and it will be easy to convert it to another language. I have even shoehorned it into a Casio pocket computer, though the run time was about 30 seconds. On a 50MHz 486 pc however the run-time is a fraction of a second.

How the program works

The core of the program is a calculation of the charging and discharging curves of the reservoir capacitor, **Fig. 1**. The charging period from time t_1 to t_2 is described by a cumbersome expression. This expression is the solution of the differential equation for charging from a sinusoidal source, **Fig. 2**. Also shown in this diagram is the discharge equation, for the period t_2 to t_1.

Initially, the program assumes that the capacitor is fully charged. It then iteratively discharges and recharges it until the ripple voltage reaches a steady value. Times t_1 and t_2 are re-estimated at each iteration.

This is a crude procedure, but it is justifiable as the iteration seems to be strongly convergent. The convergence error and computed voltages are printed by the program, so you can check that it is producing reasonable answers.

Peak ripple and rectifier currents are estimated by assuming the capacitor is at its mean voltage and is charged from the peak supply voltage. This sometimes gives an overestimate, but agrees with the practical observation that peak ripple current is essentially independent of the capacitor value for reasonable designs.

The main simplifying assumption in this program is that the rectifiers are treated as ideal switches. In practice they actually turn on and off quite slowly, as can be seen with an oscilloscope. However, this does not appear to be a major source of error. Equivalent series resistance of the capacitor is also ignored. This tends to increase the ripple crest voltage slightly, but for most practical designs the effect is not important.

How to use the program

The first step is to specify the load current and/or the load resistance. Nowadays the majority of new designs will include a voltage regulator, so an unvarying load will appear as a constant current load.

You can specify a load current by setting

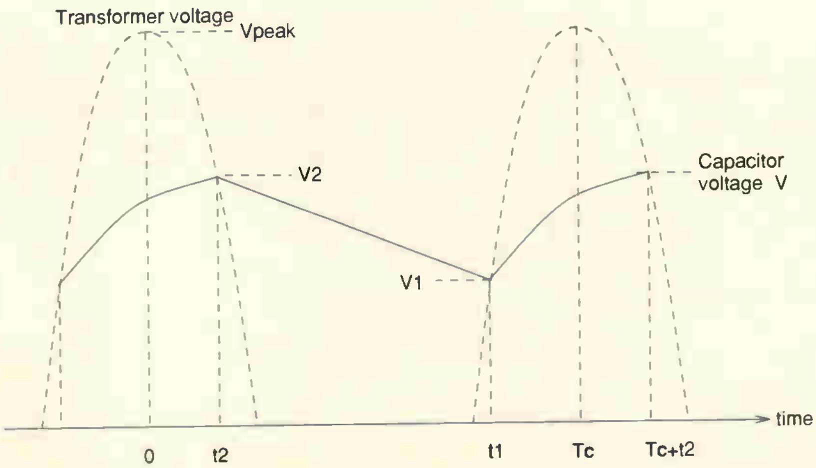


Fig. 1. Voltage waveforms typical of a linear power supply.

resistance to 1MΩ, or you can specify load resistance by setting current to zero. Alternatively, a combination of the two can be used.

For simplicity, a constant current load is assumed to be compliant right down to zero volts, though in practice you will have to specify a minimum value for the ripple trough voltage.

You next need to measure three parameters for the mains transformer. These parameters are listed in comments at the head of the program and are measured off-load. For all but a high current supply, a digital multimeter will suffice for this job. Where high currents are involved, a low-ohm meter with four-terminal measurement will be needed to check the secondary winding resistance. For a centre-tapped secondary, use the averages of the two secondary resistances and turns-ratios.

These three parameters *must* be measured as the manufacturer's figures are often approximate or non-existent. If you do not yet have the transformer to hand, you can use the program to estimate the allowable limits for winding resistances.

There may be other resistances to budget for too. A fuse-holder in the secondary circuit for example may add as much as 0.2Ω, which should be included as part of the transformer secondary resistance.

Table. Values calculated using the psu design program versus measured values demonstrate the usefulness of the program.

	Calculated	Measured
Mean capacitor voltage	37.35295V	37.4V
Ripple voltage	1.301182V pk-pk	1.28V pk-pk
Peak ripple current	3.340345A	3.4A

Figures of merit

The next step is to choose a value for the reservoir capacitor. Schade used a dimensionless figure-of-merit. This is $2\pi fCR$, where f is the mains frequency, C is the capacitance and R is the equivalent load resistance ($R=V_{op}/I_{op}$).

Capacitance C is chosen so that the figure-of-merit is around 10-15 for full-wave rectification or 20-30 for half-wave. The figure is not critical, so there is no need to go hunting for strange-valued capacitors. In any case the tolerances on electrolytic capacitors are very loose and the actual value has very little effect on the mean output voltage. If the capacitance is too low then the ripple voltage increases disproportionately and regulation will suffer. There is little to be gained by setting it too high.

The figure-of-merit is equivalent to saying that the time constant CR should be about five charging cycles. Easier still, you can try a few different values and let the computer do the work for you.

Finally, use 'wave' and Nrec to define the circuit topology. Wave is 0.5 for a half-wave circuit and 1.0 for full-wave. Variable Nrec will be 1 for half-wave or centre-tapped full-wave circuits and 2 for full-wave bridge circuits. Vrec should be acceptable if left at 0.7 V, though modern designs may well use Schottky rectifiers where Vrec will be about 0.5 V.

Output values printed at the end of the run

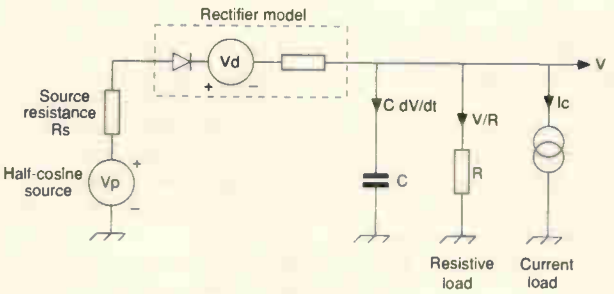


Fig. 2. Circuit model for charging and discharging of the linear supply's smoothing capacitor.

Charging equation

$$C\frac{dV}{dt}+\frac{V}{R}+I_c=\frac{V_p\cos(2\pi f(t-T_C))-V-V_d}{R_s}$$

Discharging equation

$$C\frac{dV}{dt}+\frac{V}{R}+I_c=0$$

Qbasic listing for designing power supplies. Known data is entered on the DATA line. Values currently following the DATA statement are for the example mentioned in the text.

```
REM Program psu6.bas for mains power-sup-
ply calculations.
REM For transformer-rectifier-capacitor circuits
with sine-wave
REM mains supply and load specified as current
and/or resistance.

REM By W. Gray, started 4-Aug-95, this version 6-
Jan-97

DECLARE FUNCTION arcos (A)

DATA 1, 1e6, 237.3, 50, .1354,33.3, 0.88, 5000e-
6,0.7, 1, 2
READ Ic, Rload, Vsup, freq, Trat, Rpri, Rsec,
cap, Vrec, wave, Nrec

REM load group
REM Ic, load current, A
REM Rload,load resistance, ohms

REM mains supply group
REM Vsup, mains supply to transformer primary,
volts rms
REM freq, supply frequency, Hz

REM components group
REM Trat, transformer turns ratio, secondary/pri-
mary
REM Rpri, transformer primary resistance, ohms
REM Rsec, transformer secondary resistance, ohms
REM cap, reservoir capacitor, farads
REM Vrec, forward voltage drop of single rectifi-
er

REM configuration group
REM wave, 0.5=half-wave, 1.0=full wave rectifica-
tion
REM Nrec, total number of rectifiers in series

CLS : REM initialise
Vsec = Vsup * Trat
Vpeak = Vsec * SQR(2)
Rsec = Rsec + Rpri * Trat * Trat
REM approximate allowance for dynamic resistance
of rectifiers
Iop = Ic + Vsec / Rload: Rsou = Rsec + .025 *
Nrec / Iop
Vrec = Nrec * Vrec
Tcyc = .5 / wave / freq
pi2f = freq * 6.2831853#
rc = cap * Rload
cr = cap * Rsou
A = (1 + Rsou / Rload) / cr
b = Vpeak / cr
c = b / (A * A + pi2f * pi2f)
d = (Vrec + Ic * Rsou) / cr
e = d / A
f = Ic * Rload
t1 = Tcyc
t2 = 0
V2 = Vpeak
```

are mostly self-explanatory. Various peak currents are needed to specify the capacitor, rectifiers and fuses. When using a regulator, the crest and trough voltages have to be known so that the regulator limits are not exceeded. These printouts only solve part of the overall psu design problem. However, you should need no further assistance to derive power dissipation, regulation factor, rectifier peak-inverse voltage, etc.

You can also vary data values to explore sensitivity to component tolerances and worst-case load and mains variations or to adjust a design to meet particular constraints.

The data statement at the head of the program lists values for a circuit that I constructed and measured. The exact calculated results are listed in the **Table**, so you can check that the program is working correctly. Measured values are also listed so you can see the sort of errors you might expect. The circuit in question had a high figure-of-merit. In general the calculated values become less accurate as figure-of-merit reduces.

Conclusion

Now you can throw away those tattered old photocopies of Schade's graphs. The program described here should give all that is needed to design the basic mains power supply, though at the minor cost of having to do some simple component measurements. To go further it would be necessary to use a simulation package, though the accuracy of the results at ordinary mains frequencies would probably not be much better. ■

Reference

Schade,O H, 'Analysis of rectifier operation', Proc IRE, 1943, Vol. 31, No 7.

Output from the power supply design program. Note that formatting of the data has been edited slightly to increase clarity.

n	Eror	Vop	Vrip
1	45439.35	43.40221	-7.395554E-02
2	.1363602	41.60973	6.240463E-02
3	.3032188	40.24191	.3656235
4	.2762871	39.2697	.6419106
5	.2124672	38.60608	.8543777
6	.1509972	38.16445	1.005375
7	.102787	37.87523	1.108162
8	6.821442E-02	37.68774	1.176376
9	4.458618E-02	37.56697	1.220963
10	2.885818E-02	37.4895	1.249821
11	1.856995E-02	37.43993	1.268391
12	1.190567E-02	37.40827	1.280296
13	7.614136E-03	37.38807	1.28791
14	4.859924E-03	37.37518	1.29277
15	3.105164E-03	37.36697	1.295876
16	1.979828E-03	37.36174	1.297855
17	1.25885E-03	37.35841	1.299114
18	8.010864E-04	37.35628	1.299915
19	5.149841E-04	37.35493	1.30043
20	3.242493E-04	37.35406	1.300755
21	2.098083E-04	37.35352	1.300964
22	1.296997E-04	37.35317	1.301094
23	8.773804E-05	37.35295	1.301182

constant current load= 1 A
resistive load= 1000000 ohms
supply voltage= 237.3 Vrms at 50 Hz
peak secondary voltage= 45.43928 V
effective source resistance= 1.540493 ohms
reservoir capacitor= 5000 microfarads

mean output= 37.35295 V, current= 1.000037 A
ripple= 1.301182 VPTP
ripple crest= 38.00354 V, trough= 36.70236 V
peak capacitor ripple current= 3.340345 A
peak repetitive rectifier current= 4.340383 A
peak non-repetitive (inrush) current= 28.58779 A
approx inrush duration= 7.702462 milliseconds
psu figure-of-merit= 58.67169

```
Vprev = 1000 * Vpeak
Eror = Vprev
n = 0

REM main loop
DO WHILE Eror > .0001 AND n < 200
n = n + 1

REM discharge capacitor from t2 to t1
V1 = (V2 + f) * EXP((t2 - t1) / rc) - f
t1 = Tcyc - arcos((V1 + Vrec) / Vpeak) / pi2f

REM charge capacitor from t1 to t2
temp = A * COS(pi2f * (t1 - Tcyc)) + pi2f *
SIN(pi2f * (t1 - Tcyc))
Q = V1 + e - c * temp
Q = Q * EXP(A * (t1 - t2 - Tcyc))
temp = A * COS(pi2f * t2) + pi2f * SIN(pi2f * t2)
V2 = Q - e + c * temp
t2 = arcos((V2 + Vrec) / Vpeak) / pi2f

REM calculate convergence error
Vrip = V2 - V1
Eror = ABS(Vrip - Vprev)
Vprev = Vrip
Vop = (V2 + V1) / 2
PRINT n; Eror; Vop; Vrip

LOOP: REM go back and do it again

REM estimate peak ripple and rectifier currents
Iop = Ic + Vop / Rload
Irec = (Vpeak - Vrec - Vop) / Rsou
Irip = Irec - Iop
Inrush = (Vpeak - Vrec) / Rsou

REM list data values
PRINT : PRINT "constant current load="; Ic; "A"
PRINT "resistive load="; Rload; "ohms"
PRINT "supply voltage="; Vsup; "VRMS at"; freq;
"Hz"
PRINT "peak secondary voltage="; Vpeak; "V"
PRINT "effective source resistance="; Rsou;
"ohms"
PRINT "reservoir capacitor="; cap * 1000000!;
"microfarads"

REM list results
PRINT : PRINT "mean output="; Vop; "V, current=";
Iop; "A"
PRINT "ripple="; Vrip; "VPTP"
PRINT "ripple crest="; V2; "V, trough="; V1; "V"
PRINT "peak capacitor ripple current="; Irip; "A"
PRINT "peak repetitive rectifier current="; Irec;
"A"
PRINT "peak non-repetitive (inrush) current=";
Inrush; "A"
PRINT "approx inrush duration="; cr * 1000!;
"milliseconds"
PRINT "psu figure of merit="; pi2f * cap * Vop /
Iop
END

FUNCTION arcos (A)
REM no arc-cosine function provided in MS basic
arcos = ATN(SQR(1 / A / A - 1))
END FUNCTION
```

Driving on the high side

Ian Hegglun

demonstrates how high-voltage high-side switching, ac switching and half-bridge switching can all be done reasonably fast using low-cost logic gates and small-signal diodes.

Half-bridge mosfet drivers usually use two n-channel mosfets because n-mosfets offer lower on resistance. But this complicates the gate drive for the high side mosfet. Integrated circuits for driving half bridges are now available. They reduce the component count but for some applications, they are either too expensive or do not allow control of the dead time.

A relatively simple discrete high side drive can be made using a voltage doubler, **Fig. 1**[1]. Here, a high-frequency oscillator drives a voltage multiplier which charges the mosfet's gate capacitance. When the oscillator signal is gated off the gate resistor R_g discharges the mosfet gate capacitance. This is simple, inexpensive but relatively slow. A number of methods described here can improve the speed and usefulness of this method.

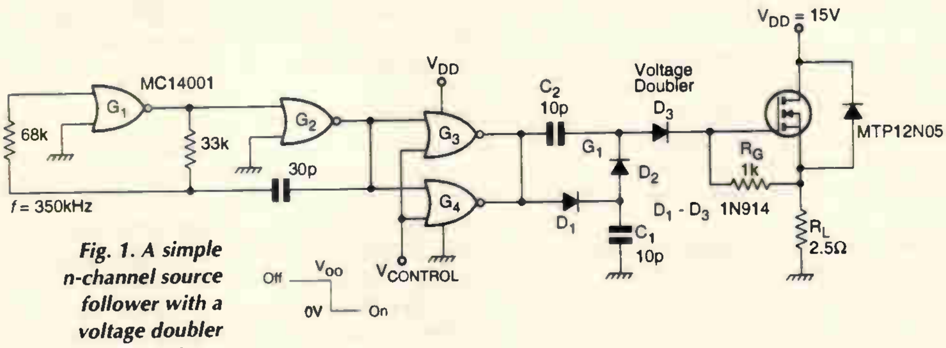


Fig. 1. A simple n-channel source follower with a voltage doubler drive.

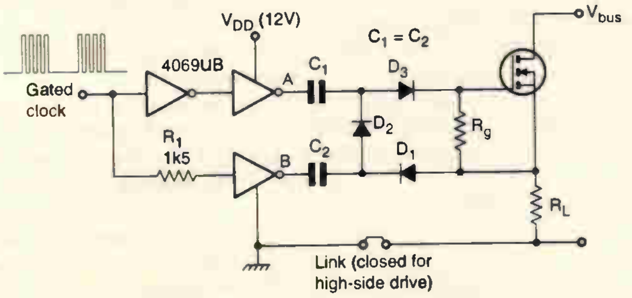


Fig. 2. The improved voltage doubler allows the mosfet to be turned on with the link open.

Improving on the doubler

Figure 2[2] uses the improved voltage doubler which allows the mosfet source voltage to rise to a higher voltage than in Fig. 1. Reference 2 includes a mains mosfet driver using this doubler technique. Analysis of the high side version (link closed) can be broken into three stages. First, with the link open and no voltage for supplying the load, then with supply voltage added and finally with the link closed.

With the link open and no load supply, start with no charge on the coupling capacitors $C_{1,2}$. When the inverters are powered up one output will be high and the other low. If output A is high then current will flow through C_1 and D_3 then through the mosfet's gate-source capacitance and back through D_1 and C_2. This causes C_1 and C_2 to be charged to nearly half V_{DD} while placing some charge on the mosfet's gate-source capacitance.

During the next phase, current flows through C_2, D_2 and C_1, which re-polarises the capacitors to around half V_{DD}. The next phase is a repeat of the first, except that C_1 and C_2 now have more than zero volts across them which allows the gate voltage to rise further.

Eventually, the gate-source voltage reaches a limit. If R_g is removed, the limit is $2V_{DD}$ less two diode voltage drops. When running c-mos gates on 12V the doubler output can reach 22V off load.

When R_g is added the final voltage is reduced because R_g forms a voltage divider with the output resistance of the doubler R_{out}. With a 12V supply the value of R_g should not be less than R_{out} because the gate voltage is pulled down to half of 22V when R_g=R_{out} or 11V which is usually sufficient for standard mosfets.

When R_g equals R_{out}, gate charging time constant $T=RC$ can be found. The total capacitance is $C_{iss}+2C_1$ and a net resistance of $R_g//R_{out}$ when turning on the mosfet. For turn off there is $C_{iss}+2C_1$ but only R_g for discharging. This makes turn off typically twice as long as the charging time constant if R_g equals R_{out}.

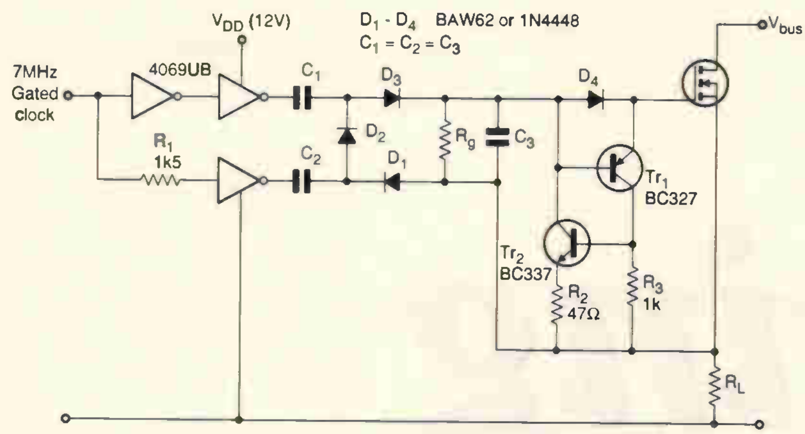


Fig. 3. Adding a regenerative clamp reduces turn off time to 1μs and allows higher frequency operation.

Next, with the link still open and a floating supply to drive the load the rise and fall times for the gate voltage increases due to the Miller effect. It is a rapidly changing drain-source voltage V_{ds} that causes current to flow through the mosfet's drain-gate capacitance C_{dg} which opposes gate current used to switch the mosfet. This current typically doubles the turn on and turn off transition times and the effect can be observed on the gate voltage waveform as a plateau region, similar to those shown in most mosfet data sheets. More on this is given in the panel entitled 'Mosfet input capacitance'.

For high-side switching, the link is closed. The circuit continues to operate as before except that C_1 and C_2 must be charged ultimately to the load voltage. This requires current flow through the link. When the mosfet turns on, this is not a problem but when switching a high voltage, the turn-off time is increased considerably. When switching high voltages such as 300V from the mains, the turn-off time is 10 to 20 times longer than the turn-on time. This restricts the switching frequency to only low switching rates.

The speed difference can be explained. When the mosfet begins to turn on, there is a path for current from the source through D_1 and C_2 to rapidly charge C_1 and C_2. This is a form of boot strapping where the mosfet is acting with near unity voltage gain with an input capacitance much less than C_{iss}. But when the drive to the mosfet is stopped Rg must discharge C_1 and C_2 starting from the load voltage down to zero.

Current through R_g holds up the gate-source voltage above V_t until C_1 and C_2 have discharged sufficiently. The effect is observed as an extended plateau when turning off.

Achieving faster turn off...

To speed up the high side driver a turn-off clamp was developed. The standard single-transistor p-n-p clamp, as used in PIV isolators[3], did not reduce the turn off time of the high side drive although it did assist turn off when the link was open.

A regenerative scr type circuit, **Fig. 3**, is needed to speed up the clamp the gate voltage when the gate begins to be discharged by R_g. When drive to the doubler is stopped, the mosfet gate discharges through Tr_1 which turns on the second transistor Tr_2, triggering the latch. This discharges the mosfet gate capacitance rapidly as well as the coupling capacitors through the link and load. Resistor R_2 limits the discharge current through C_1 and C_2 and drivers.

Capacitor C_3 prevents ripple from the doubler output triggering the scr. A full-wave doubler is helpful here to reduce the ripple and allow C_3 to be a small value ($0.1C_1$) which increases the rise time by up to 25% but requires two extra capacitors – all capacitors are halved – plus three diodes.

With these extra components, the turn off time can be reduced to much less than the turn on time. Fall time can be 1μs with 300V even with large gate capacitance. Turn on time is now the limiting factor.

...and faster turn on

Mosfet turn on time in any mosfet drive is limited by the magnitude of current that can be delivered to the mosfet to charge the gate-source capacitance. In a multiplier drive this current is limited by the coupling capacitors reactance and the driver's source resistance.

There is little point increasing C_1 more than half the mosfet's gate capacitance because more driver energy is wasted charging C_1 and C_2 than is delivered to the mosfet. This places a practical limit on C_1 and C_2 in the range of 1nF to 4nF (the value of C_{iss} for the mosfet used).

Raising the clock frequency to the highest possible frequency for full output swing is the best way maximise turn on time and minimise the coupling capacitance.

From tests the effective capacitive reactance seen at the output was found to be about seven times the capacitive reactance for a sinewave. Reactances of C_1 and C_2 appear in series and

Mosfet input capacitance

Current is required to charge the input capacitance of a mosfet to turn it on. Input capacitance is mostly gate-to-source metallisation and gate to channel capacitance plus some parallel gate to drain capacitance. Capacitance C_{iss} is the common source input capacitance and is the sum of C_{gs} and C_{dg} (on data sheets C_{gd} appears as C_{rss}.

Both C_{gs} and C_{dg} are voltage dependent and vary considerably with V_{ds} when V_{gs} is a few volts below or above V_{ds}, or V_{dg} is approximately zero. Calculation of gate voltage rise time based on data sheet C_{iss} value at V_{ds}=25V will give a value that is 200 to 400% too fast.

A better method to find the gate voltage rise time is from the effective C_{iss} value derived from the gate charge graph. The effective input capacitance is calculated from $C={}^{Q}/_{V}$ where V is the gate voltage rise required to turn the mosfet on and Q is the charge required to reach that voltage. The rise time can be estimated from 2.5 RC time constants – assuming it is supplied by the same V used above – where R is the driver source resistance in parallel with the gate discharge resistor, if used.

The plateau region is caused by the Miller effect while V_{ds} is changing[6]. This means rise and fall times can be found from Q=i×t, where Q is the charge required to move through the plateau region and i is drive current at this point where i is constant here because V_{gs} is constant. Hence i=$(V_s-V_{pl})/R_s$. Voltage V_{pl} is typically 3-4V for non-logic mosfets – a volt or two above V_{TH} – or can be found from the V_{gs} versus I_D plot from the load current.

Simple rules of thumb can be used for rough estimates. The effective input capacitance is around three times the data sheet C_{iss} value, at V_{ds}=25V, when charging of the gate to 10V. Rise and fall times for the load voltage will be around half the gate voltage rise time found from the C_{iss} estimate and gate driver source resistance. can be measured on a capacitance meter by shorting the drain to source together. This give C_{iss} at V_{ds}=0 and the effective C_{iss} value is around 1.5 times higher for 10V gate driving.

There are undoubtedly exceptions so use the gate charge method when ever possible.

likewise the driver resistances but these must be referred to the output like a transformer with a 2:1 step up ratio and is seen as $8R_{on}$ at the output. Hence we have $R_{out}=14X_c+8R_{on}$, where X_c is $^1/_{2\pi fC_1}$.

Equating $14X_c=8R_{on}$ gives,

$$C_1 < \frac{1}{3.6 f_{max} R_{on}}$$

and $R_g>16R_{on}$. **Table 1** shows measured values R_{on}, f_{max} for various drivers and calculated values for C_{1max} and R_{gmax} for use in Fig. 3.

Rise-time values are calculated using $t_r=2.5(R_g/2)(3C_1+C_{iss})$ where C_{iss} is 1.5nF – the effective value for the *BUK453-100A*. While the 74xx04 inverters offer low output resistance they are not suited to driving standard mosfets because of their 6V maximum supply. They are still useful for driving logic level mosfets. A tripler can be used if a 12V supply is not available, but loading on the drivers is doubled and the rise time increased.

The 4041B is a quad buffer with complementary outputs and internal equalisation delay. It offers the best performance but at higher cost. Farnell's list its price at about four times that of the 4049B. Complementary outputs for standard inverters can be obtained by the method shown in Fig. 3 and explained in the 'Speed trial' panel.

Table 2 shows measured rise and fall times for Fig. 3 driving a *BUK453-100A* mosfet (10A, 100V) with 22V into a 50Ω load. Rise time is not affected significantly when driving a 5Ω load or higher voltages although the fall times do increase with higher voltage. The fast discharge circuit allows the fall time can be kept small at higher voltages.

Table 1. Test results for c-mos inverters operating at 12V predict reasonably gate voltage rise times in the range of 1.5-10µs or 1-2 orders of magnitude faster than PIV relays.

		Measured		Calculated		
	V_{DDmax}	R_{on} (Ω)	f_{max} (MHz)	C_{1max} (pF)	R_{gmax} (Ω)	t_{rmin} (µs)
4069UB	15	270	7	150	4k3	10.5
4049UB	15	150	11	170	2k4	6.0
4049B	15	150	12	150	2k4	5.9
4041B	15	60	10	460	960	3.5
74HCU04	6	50	16	350	800	2.5
74HC04	6	50	17	330	800	2.5
74VHCU04	6	50	20	280	800	2.3
74AC04	6	20	18	770	320	1.5

Table 2. Measured rise and fall times for various c-mos drivers using the values shown when driving a BUK453-100A as in Fig.3. Only one pair of inverters are used as drivers in all cases.

	Test values			Gate voltage		Load voltage	
	C_1 (pF)	R_g (Ω)	f (MHz)	t_r (µs)	t_f (µs)	t_r (µs)	t_f (µs)
4069UB	220	10k	7	6	1	2	0.5
4069UB	150	10k	7.5	6	0.5	1.5	0.2
4049B	220	4k7	12	2.5	0.5	1	0.3
4041B	470	4k7	10	1.5	1	0.5	0.5

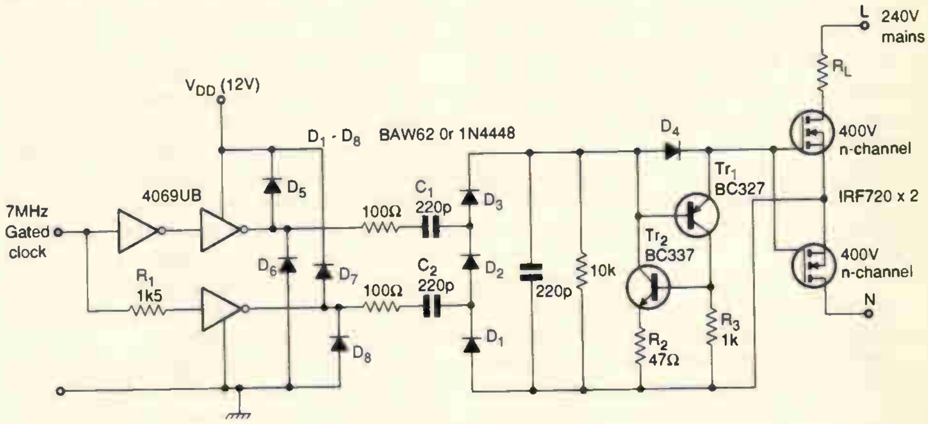
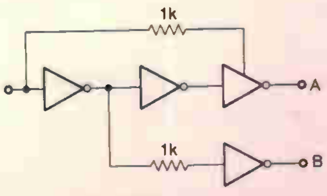


Fig. 4. With two mosfets ac can be controlled from ground potential but protection is needed to prevent high current spikes coupled through $C_{1,2}$ from damaging the c-mos drivers.

Speed trials

The propagation oscillator is a type of phase-shift oscillator which uses the phase shift of three stages to set the operating frequency. Overall phase shift required is 180° or 60° for each stage if they are identical.

The 4069UB, 4041B and 4049B give close to full output swing with a 1kΩ resistor between one stage to lower the operating frequency slightly for full output swing. The *74XX04* types require an additional additional 1kΩ resistor between stages plus a 10pF capacitor to ground on the inputs supplied by the resistors. Complementary outputs can be obtained by adding a fourth inverter with a 1kΩ delay equalisation resistor as shown.



Propagation oscillator for c-mos inverters, as used in the speed trials.

Note t_r for the gate voltage in this table is not the usual 10% to 90% measurement, but rather the time for the gate voltage to reach 10V from 0V. With R_g increased, the final voltage is closer to 15V.

The fast discharge circuit allows R_g to be higher than the minimum in Table 1 without affecting the discharge time. Faster gate voltage rise times recorded in Table 2 can be partly attributed to the higher R_g values. Note that if R_g is too large, the gate voltage may exceed the maximum allowable so a 15V zener diode should be placed across R_g for protection.

I was intending to apply this in switched capacitor converters[4]. Because electrolytic capacitors are used in these converters for charge pumping, the best operating frequency is around 30kHz. The drivers described here are more than adequate for this. The 4041B with all four buffers in parallel can drive a 4nF mosfet for 0.5µs transition times provided the coupling capacitance is also increased.

Solid state ac switch

Figure 4 shows a circuit that was used to chop 240V mains at several kilohertz. The driver should be ground referenced for safety but can be connected to any other potential if required.

Protection diodes and resistors are needed on the inverter outputs to divert current spikes caused by abrupt common mode voltage changes. The values shown protect the 4069 against mains spikes. To demonstrate this the driver common can be directly connected to phase while running – of course in this test the driver supply must be floating.

The 4049 or 4041 provides an extra margin of reliability for mains operation, although for harsh, noisy environments it would be wiser to use a high common mode dv/dt optocoupler instead of this method.

This ac switch was intended for use in a flying capacitor circuit[5], to allow the output of a bridge rectifier to be connected back to ground for safety, **Fig. 5**. This overcomes the need for a transformer. It operates as follows. Mosfets A_1 and A_2 charge the flying capacitor while B_1

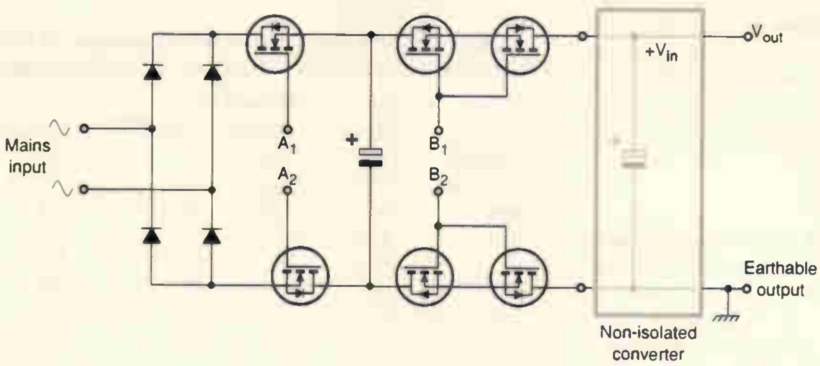


Fig. 5. The flying capacitor technique allows the normally floating output of a bridge rectifier to be connected to earth without damaging the circuit.

and B_2 are off. Fets A_1 and A_2 turn off before B_1 and B_2 turn on. When B_1 and B_2 turn on charge is transferred to the converter. Non-overlapping switching is essential when the output is connected to earth.

Note that this method does not allow the output to 'float' like a transformer because the mosfets must withstand mains voltage plus any common mode difference plus a safety margin.

A crowbar circuit and earth leakage protection should be used to protect against excessive common-mode voltage, which can destroy the mosfets and place mains on the output. With 500V mosfets and 230V mains, only 100V common mode difference is allowable.

Higher margins are offered by igbts at reasonable cost, namely 500V margin with 1kV devices, but tail losses limits the frequency to under 10kHz.

All four mosfets require high side drivers although *A* and *B* mosfets can share the same driving gates and control circuit. Regulations in some countries place a limit on the amount of coupling capacitance between line and earth of 0.005μF which allows four drivers each with up to 470pF coupling capacitors. This amounts to a reasonably safe current of 300μA from 230V.

Half-bridge drive

I have tested the half-bridge circuit of **Fig. 6**, and found that it can operate at 100kHz using the humble 4069. Three outputs are paralleled to drive a two diode charge pump, $D_{5,6}$, plus a doubler. The charge pump provides twice the current of the doubler in the plateau region around 3V and accelerates gate charging and reduces the turn on time. Once above this region the doubler takes the voltage to 10V.

The propagation oscillator, see the 'Speed trial' panel, is a feature that allows the drivers to run at the maximum frequency to suit the particular device used – 7MHz for the 4069 used here. Dead time is adjustable by delaying the turn on signal to each mosfet. The ability to reduce dead time improves the form factor and efficiency when operating at high frequencies. I have found the inability to adjust dead time on the *IR2151* to be a limitation when operating above 50kHz.

In summary, a wide range of applications can make use of this method of driving mosfets where the source voltage is different from the driving circuit common. High voltage high side switching, ac switching and half bridge switching can all be done reasonably fast using low cost logic gates and small signal diodes. No high-voltage transistors or opto-isolators are required.

My thanks to National Semiconductor for supplying complementary samples of the *74VHCU04*. ■

References

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2. SGS-Thompson, Power Supply Application Manual, July 1985, AN275, p277-280.
3. Moore, D. W., 'Using Photovoltaic relays in multiplexers', *Electronics & Wireless World*, July 89, pp. 725-8. (or International Rectifier note GBAN-PVI-1).
4. Hegglun, I., 'Stepping out', *Electronics World*, October 1996, pp. 759-806.
5. 'Novel power supply', *Electronics & Wireless World*, September 1988, p. 885.
6. International Rectifier, AN947, 'Understanding HEXFET® Switching Performance'.

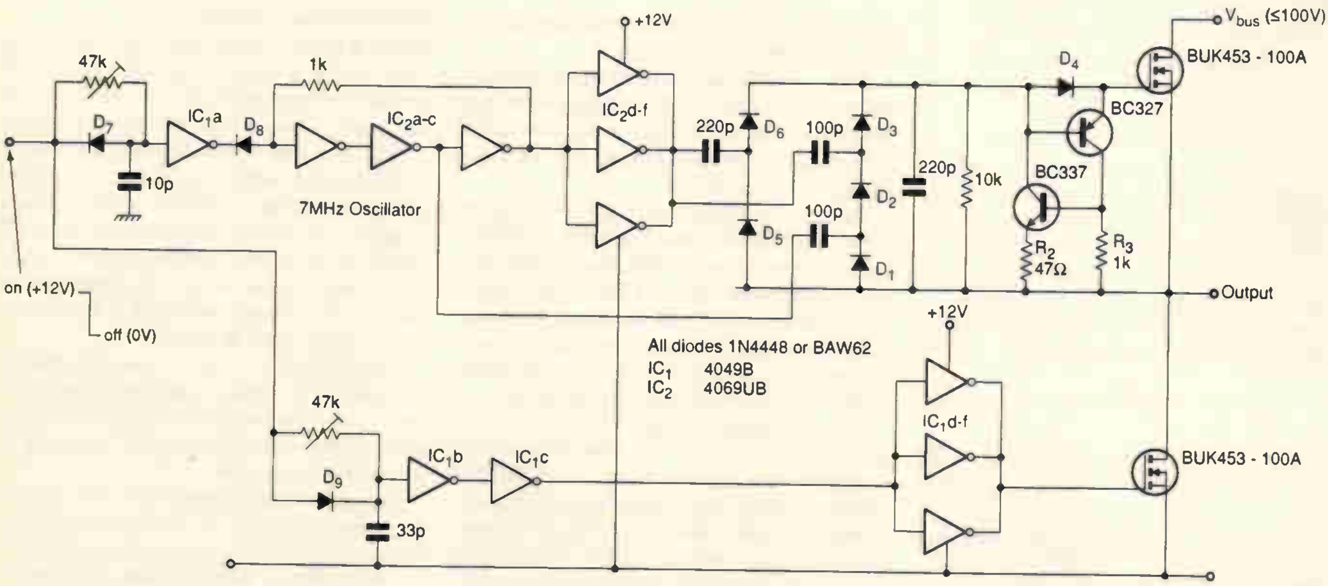

Fig. 6. This half bridge uses two low cost c-mos drivers, can operate up to 100kHz and the dead time can be varied.

Micropower crystal oven

Commercial crystal ovens are notoriously power hungry, being designed to work way above room temperature. Operating at a user-defined temperature just above the maximum anticipated ambient, Rae Perälä's oven can be run from a battery.

The circuit presented here can be used as a battery operated thermostat oven. It consists of two *7555* timer circuits, but it could incorporate a dual timer circuit *7556*. Both the *7555* and *7556* are c-mos types, which guarantees a very low power consumption.

Figure 1 shows the complete circuit. Timer IC_1 acts as an oscillator, and IC_2 as a heating control circuit. **Figure** 2 shows the IC output voltages. Output voltage of IC_1 lies mostly at the positive battery voltage V_B, giving only short 23μs zero voltage pulses, represented by T_2, in the intervals of 3.5ms, shown as T_1.

Output pulses of IC_1 trigger IC_2, which is connected as a monostable multivibrator. Its produces a positive-going 2.5ms pulse, represented by T_3, every time its trigger input in pin 2 goes below the one third of the battery voltage.

Timer IC_2 output controls an n-type enhancement mos transistor, used for driving current pulses to the heating resistor.

The triggering voltage of IC_2 is formed from the output pulses of IC_1 by a voltage divider. One part of the potential divider is a potentiometer used to adjust the temperature. The other is a 100kΩ negative temperature coefficient thermistor mounted inside the oven.

The advantage of using a voltage divider is that changes in the battery voltage have no influence on the circuit operation. The circuit is adjusted so, that the resistance value of P is exactly one half of the resistance value of the thermistor at the temperature required in the oven.

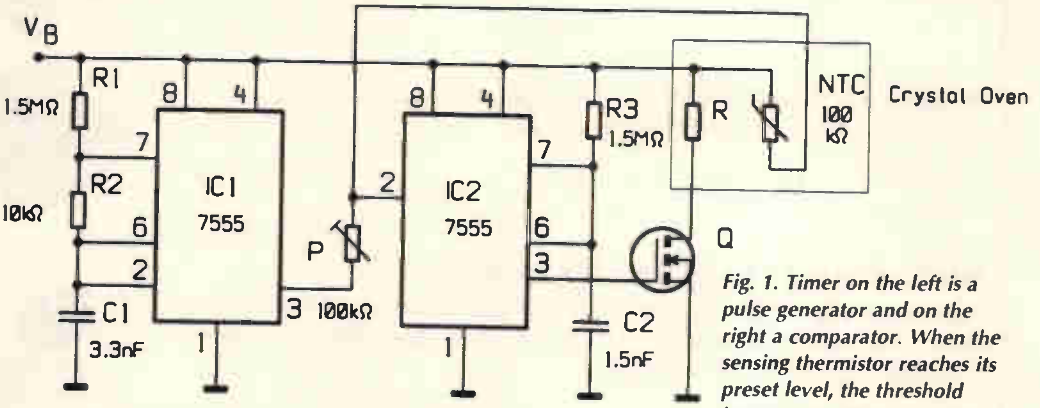

Fig. 1. Timer on the left is a pulse generator and on the right a comparator. When the sensing thermistor reaches its preset level, the threshold becomes too high to let the comparator trigger and heater resistor R receives no heating pulses.

How it works

While the oven temperature is below the preset value. the trigger input voltage of IC_2 at pin 2 goes below the triggering limit $^{V_B}/_3$ during every zero pulse from IC_1. This means that IC_2 is triggered, producing heating current pulses to the oven.

When the oven temperature reaches the value attained, the thermistor's resistance value decreases. This prevents the trigger pulses for IC_2 from going below one third of the battery voltage. Now, the heater no longer turns on.

From Fig. 2, you can see that the positive output pulses from IC_2 begin exactly at the rising edge of triggering pulses from IC_1. In this illustration, the first two triggering pulses trigger IC_2, as would be the case if the oven temperature were too low. During the third pulse the oven temperature level is assumed to have increased so. Triggering does not take place and the heater remains off.

The oven temperature should be adjusted a little higher than the maximum expected ambient temperature, via potentiometer P. An unnecessary high oven temperature increases the heating power demand.

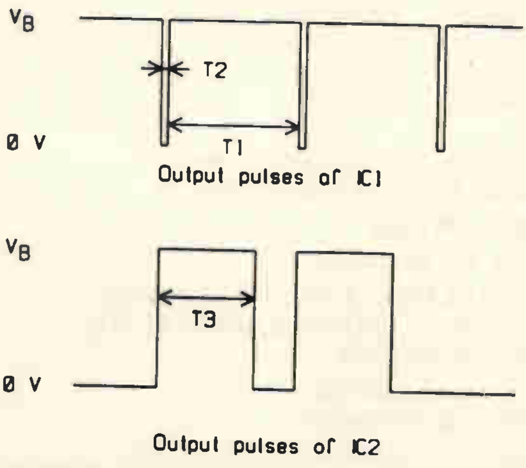

Fig. 2. Timing of the oven circuit, showing two trigger pulses causing heating and a final pulse that is not passed through to the output since the heater is up to temperature.

The circuit is a micro power one. While the heater is off battery current of 170μA will be sufficient for the whole circuit. A small crystal oven can be heated by a current of few milliamps. The heating resistor and the type of the mos transistor can be selected according to the amount of heat needed. Heater current can be minimised by making sure that the oven is well insulated.

The circuit operates with battery voltages between 8 V to 15 V. Theoretically, the circuit is not sensitive to battery voltage because operation depends on the resistance ratio of the thermistor and potentiometer P. In practice, I have found that the oven temperature alters a little when the battery voltage changes. This problem can be avoided by regulating the voltage connected to the ICs using a micropower regulator. Voltage supplying the heating resistor needs no regulation, so adding such a regulator does not have a significant affect on consumption. ■

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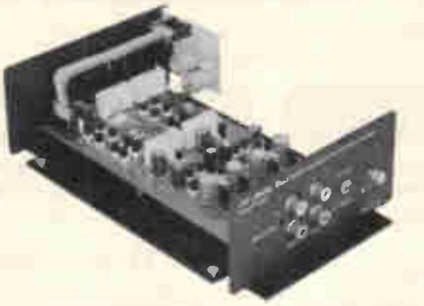

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I/O port for pcs

With tools like Visual Basic, you don't have to be a computer nerd to write a Windows application. But if you want to interface that application with the real world and other Windows programs, you do need an understanding of dynamic-link libraries. Pei An explains.

Last month, I described a data logging subsystem for the pc that has two analogue inputs, seven digital inputs and seven digital output lines. It interfaces to the pc via an RS232 link.

I wrote the software driver of the serial data acquisition and control system in Visual Basic version 3. The associated dynamic link libraries, or dlls, are written in Turbo Pascal for Windows.

Although the software provides useful features, such as automatic data acquisition and programmable digital outputs, it is primarily intended for exercising the functions of the system. However, the description of the software that follows demonstrates the basics of Windows Visual Basic programming. It will help you to develop more adventurous virtual instrumentation programs.

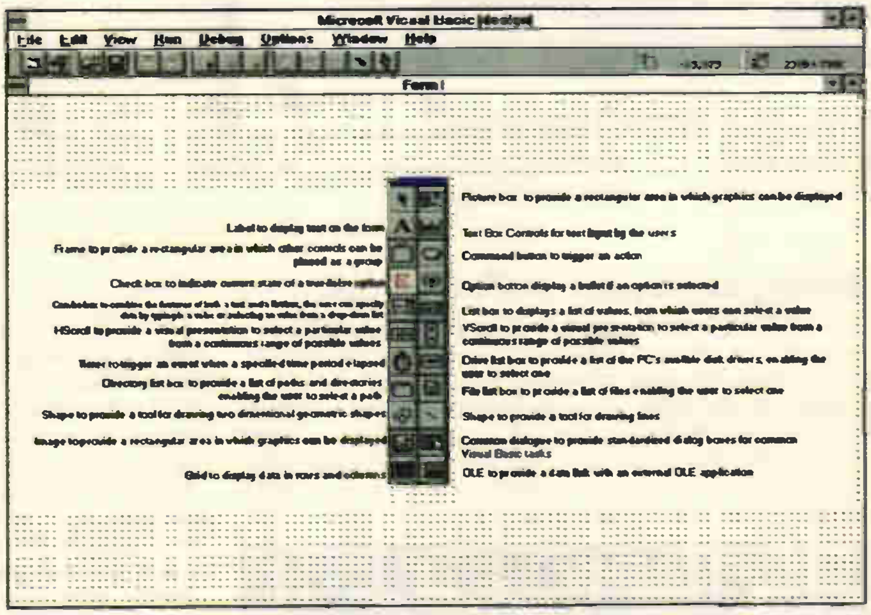


Fig. 1. Part of the task of writing a Visual Basic program involves selecting tools from the tool box and placing them on a 'Form'.

Visual programming language

Visual Basic was released in mid-1991 by Microsoft. Version 2.0 came a year later with Version 3.0 following in mid-1993. The latest version is Visual Basic 4. Visual Basic allows

What is a dynamically-linked library?

Although Visual Basic offers a wide range of program support, there are some functions that are not available. Hardware accesses to ports or to memory locations are two examples that Visual Basic does not support. In this case, specialised procedures written by other programming languages should be provided. They can be called by a Visual Basic programs anytime. These procedures are stored in dynamic link libraries, or dlls.

A dll is an executable program but it is not loaded into memory until at least one application needs a procedure contained in the library. Windows knows when a particular dll is needed by an application by examining its references to dll procedures. This is why these libraries are called 'dynamic'.

Once Windows loads the application program and its dlls, a process known as 'linking' occurs. This process connects the application program to the dll. As a result, the application runs as though the dll procedures were part of it.

A dll contains functions and procedures. There are other benefits from using dll files. First, dynamic-link libraries are available to all applications running under Windows. As a result, only one copy of a dll procedure is needed by Windows for programs which might use it.

Individual applications do not have to store the dll procedures. Therefore, each application needs less space. Second, modifications or performance enhancements to applications are easily made by changing the DLLs. The applications themselves don't need any modifications.

Table 1. Examples of 'events' that cause Visual Basic to execute instructions.

Event sources	Examples
User	Clicking the mouse, pressing a key
Computer	A specified time period elapses
Program	A program instruction activates an event
Another program	Another application request data interchange

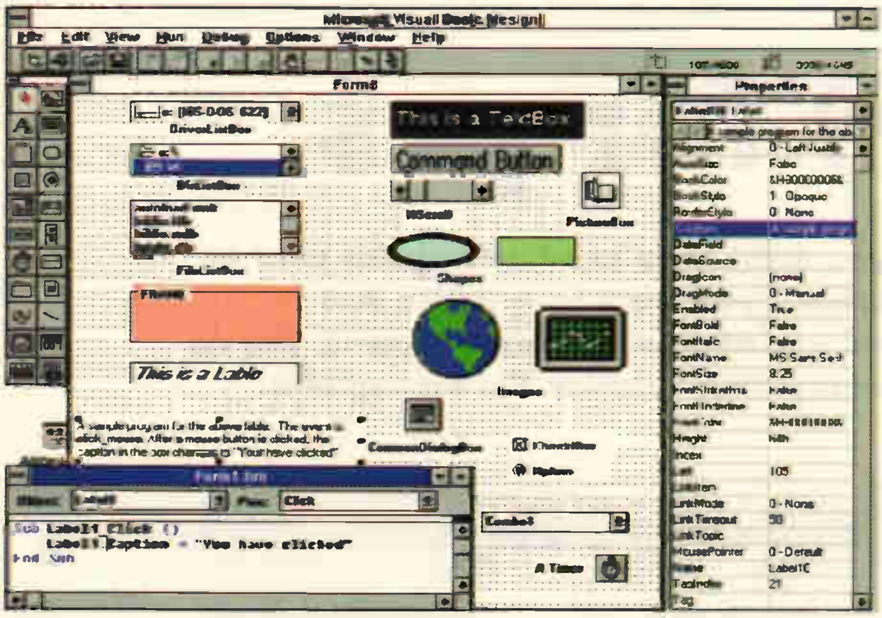


Fig. 2. Example of of a Visual Basic Form with tools placed on it. At this stage you can specify properties for each object as on the right of the display.

you to write Windows applications with ease.

Visual Basic is revolutionary. It is a visual programming language, it is event driven and it is object-oriented. It allows you to draw your own graphic interface directly on the screen without using any program instructions. This is just like drawing objects on the screen using Windows paint or other drawing packages.

You select a control from the Tool Box and place it on a form. **Figure 1** shows all the controls in the tool box. **Figure 2** is a screen dump showing that some controls are placed on Form 1. You can specify various properties for each object in the Properties windows. These properties include size, colour and caption.

What does event driven mean?

One of the major differences between Visual Basic and other 'non visual' programming languages such as QBasic, Turbo Pascal and C for dos is that Visual Basic is an event-driven language and the others are procedural languages.

The conceptual difference between these two programming languages is that in procedural languages, the program itself is in control, detecting the logic flow procedurally through the program from the beginning of the program to the end. An application is executed by proceeding logically through the program one line at a time.

Event-driven languages are completely dif-

Driving the logger in Visual Basic

To write Visual Basic programs, you need to be more of a graphics designer than a software programmer. You create the user interface for applications by simply drawing objects on the screen.

Objects include various controls and graphic objects on a form. At this stage, programming is more like using a Windows drawing package. After drawing an object, you can change properties of the object. These properties include positions of the object, colour of the background, font and size and colour of characters, etc.

Fig. 3. Shot of virtual instrumentation display produced using software developed by the author for controlling the interface.

So far, almost no programming codes are required. This is a remarkable advantage over QBasic and other non-visual programming languages. Programming graphical objects in the conventional way is code-intensive and error-prone. Only after the graphic user interface is completed, you start to write program code for each object drawn on the screen.

Figure 3 shows the visual instrumentation panel of the RS232 data logger and controller. Space does not allow detail about the Visual Basic software that I have developed for the data logger, but the code and its listings are available from me.

Briefly, the panel shown here provides a number of functions. You can configure the analogue input mode of the a-to-d converter for single-ended input or differential input mode by clicking the option button. Clicking the 'Manual logging' button results in the voltages measured by the a-to-d converter being shown on the screen. On the photograph, this is the two small screens showing 0.0000V. The display mode of numbers is configured by clicking these two small screens.

The configuration allows you to display the actual physical parameters instead of voltage. It requires you to input three conversion parameters, A, B and C, and the unit of the parameter. It then converts a voltage into others using the following equation. Value V is the voltage measured by the a-to-d converter. Obviously, A, B and C have to be specified from a calibration.

If the automatic data logging mode is used, the user has to key in the time interval, total number of scans and the name of the data file. Auto Start and Auto Stop initiate and terminate the automatic data logging. After an automatic data logging is completed, you can view the time histories of the a-to-d conversion and digital input data by clicking the Plot data.

Technical support
The i/o system hardware described last month together with the Visual Basic source code, DLLs and executable files is available from the author in various forms. Please direct your enquiry to Dr Pei An, 11 Sandpiper Driver, Stockport, Manchester SK3 8UL, UK Tel/Fax: +44-(0)161-477-9583. Pei's e-mail number is PAN@FS1.ENG.MAN.AC.UK.

List 1. Producing DLLs using Turbo Pascal for Windows is relatively easy, but they need to follow a specific structure.

```
LIBRARIES name
FUNCTION function_name1 (a1,a2...:INTEGER):INTEGER;EXPORT
BEGIN
          (instructions of the function)
END;
FUNCTION function_name2 (b1,b2.....:INTEGER):INTEGER;EXPORT
BEGIN
          (instructions of the function)
END;
EXPORTS
   function_name1        index 1,
   function_name2        index 2;
BEGIN
END.
```

ferent. Program instructions execute only when a particular event calls that section of code into action. Events in Visual Basic applications can have several sources as shown in **Table 1**.

Orientated around objects

In object-oriented programming systems, or oops, an object combines programming code and data into a single unit. Once defined, an object takes a life of its own. You simply pass information to the object without knowing how the object was created or how it works.

In Visual Basic, there are two primary objects, namely forms and controls. A form defines a window on-screen which is what users see when running the application. A control is an object you place onto a form.

Each control performs a specific function. Visual Basic provides several predefined controls. Each control is represented by an icon in the Tool Box. Each object has an associated set of event procedures. You chose appropriate event procedures and write codes for that procedure.

Writing dlls using Turbo Pascal

Using Turbo Pascal for Windows , writing dlls is as easy as writing Turbo Pascal programs. There is a special structure for generating dlls, shown in **List 1**. The italic and upper case words are reserved words of the Turbo Pascal for Windows. Others are words which are specified by users.

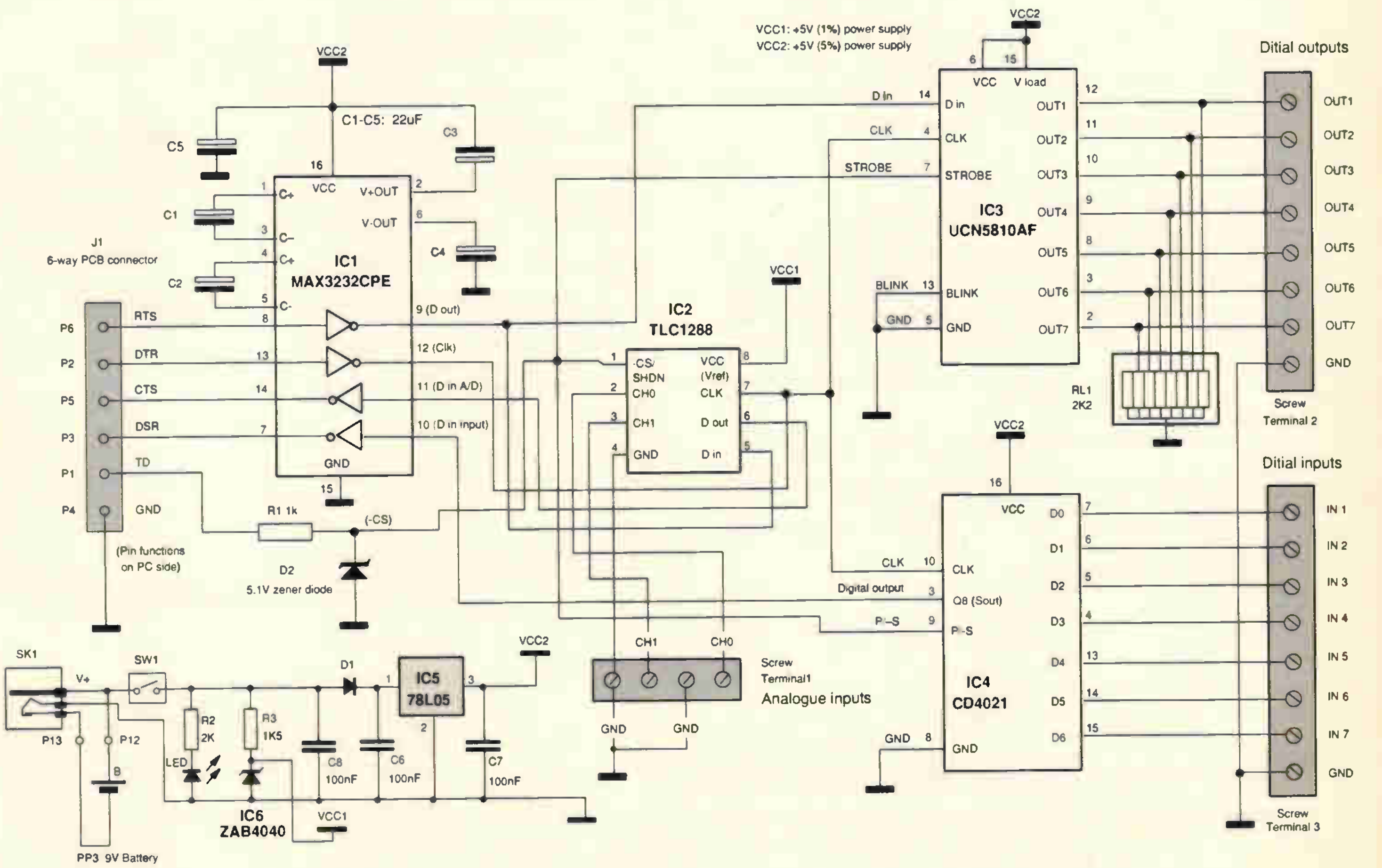
In this structure, the first line the declaration 'LIBRARY' determines that the program is a dll program. It is followed by a name which is specified by users of this library. This example library contains two functions:

```
FUNCTION function_name1
(a1,a2.....:INTEGER):INTEGER;EXPORT
```

and

```
FUNCTION function_name2
(b1,b2.....:INTEGER):INTEGER;EXPORT
```

You can see that the two functions are defined as integers and declared to be 'EXPORT'. Next, an 'EXPORTS' instruction is issued,



Circuit diagram of the RS232 data logger/controller (repeated from last month's issue).

which exports function_name1 index 1 and function_name2 index 2.

The program should be made executable. This is done by choosing the 'Make' function contained in the 'Compile' manual of the TPW editor screen. Note that the dll cannot be run within the Turbo Pascal environment.

List 2 is the dll for the serial data acquisition and control system. It contains four basic functions which control all the operations of the RS232 data logger and controller.

This library can be used by any window programming language, including Visual Basic and Visual C. Here I will show how a Visual Basic program calls the dll functions.

Assume that the above dll is stored in C:directory and the file name is 'RSLOGGER.DLL'. If that is the case, the declarations shown in **List 3** are required in the Visual Basic program.

After this declaration, the four functions, RS232(), Configure_RS232(), AD_converter () and Inputdata() can be called anywhere within the Visual Basic program. In this way, the Visual Basic program and dll functions are connected together. Some example instructions are given in **List 4**.

The functions,

```
RS232( x:integer ):integer (x can be
    0,1,2,3 or 4)
```

are concerned with addresses of the COM ports installed on your computer. Routine RS232(0) returns the number of installed RS232 ports on your pc. The port address of COM1 is returned by RS232(1), the address of COM2 by RS232(1), etc.

```
Configure_RS232(RS232_address:integer)
    :integer
```

List 3. Declarations required for the main program, assuming the file name 'RSLOGGER.DLL' and C:directory as the file's destination.

```
Declare Function RS232 Lib
    "C:\rslogger.dll" (ByVal x As
    Integer) As Integer
Declare Function Configure_RS232 Lib
    "C:\rslogger.dll" (ByVal
    RS232_address As Integer) As
    Integer
Declare Function AD_converter Lib
    "C:\rslogger.dll" (ByVal
    RS232_address As Integer, ByVal
    mode As Integer, ByVal com_byte
    As Integer, ByVal Others As
    Integer) As Integer
Declare Function inputdata Lib
    "C:\rslogger.dll" (ByVal
    RS232_address As Integer) As
    Integer
```

List 2. Dynamic link library for the data capture subsystem allows data to and from the COM port to be accessed by Windows applications.

```
Library rslogger;
{Window DLL for the RS232 data logger/controller written
in Turbopascal for windows. Hardware and software designed
by Dr. Pei An. All rights preserved, 10/96
RS232 connection details: Note: unlisted pins are not used
in the circuit
DTR (bit 0 of 04 offset register of UART, modem control):
  Clock signal, inverted in the circuit by RS3232
RTS (bit 1 of 04 offset register of UART, modem control):
  Data out signal, inverted in the circuit by RS3232
TD  (00 offset register of UART, data register): -CS
  signal, not inverted (normally low)
CTS (bit 4 of 06 offset register of UART, modem status):
  serial A/D data input, inverted in the pc
DSR (bit 5 of 06 offset register of UART, modem status):
  serial digital data input inverted in the pc
GND (ground)  }
uses
                    Windos, wincrt;
Function RS232(x:integer):integer; export;
{Universal auto detection of COM base address
   $0000:$0400 holds the printer base address for COM1
   $0000:$0402 holds the printer base address for COM2
   $0000:$0404 holds the printer base address for COM3
   $0000:$0406 holds the printer base address for COM4
   $0000:$0411 number of parallel interfaces in binary}
var
                    number_of_COM, COM1, COM2, COM3, COM4
                    :integer;
begin
      number_of_COM:=mem[$40:$11]; {read number of
                    parallel ports}
      number_of_COM:=(number_of_COM and (8+4+2)) shr 1;
      COM1:=0; COM2:=0; COM3:=0; COM4:=0;
      COM1:=memw[$40:$00];        {Memory read procedure}
      COM2:=memw[$40:$02];
      COM3:=memw[$40:$04];
      COM4:=memw[$40:$06];
      case x of
                    0:   RS232:=number_of_COM;
        1:          RS232:=COM1;
        2:          RS232:=COM2;
        3:          RS232:=COM3;
        4:          RS232:=COM4
       end;
end;
 Function Configure_RS232(RS232_address:integer):integer;
                    export;
 {Configure RS232 serial data format, Baud rate: 115200,
                    Data length: 5, Stop bit: 1, no
                    Parity check}
 {To achieve 115200 bit rate, a frequency divisor must be
                    loaded into the UART}
 var
                    ij:integer;
 begin
                    port[RS232_address]:=0;
         for ij:=1 to 10000 do ij:=ij;
                    port[RS232_address+3]:=128;{Loading
                    serial data format, first bit of the
                    register is 1}
         port[RS232_address+0]:=1;  {LSB of the divisor is
                    1}
         port[RS232_address+1]:=0;  {MSB of the divisor is
                    0}
         port[RS232_address+3]:=0;  {Load divisor}
         Port[RS232_address+1]:=2;{2=Generate interrupt
                    when TD buffer is empty}
         for ij:=1 to 1000 do ij:=ij;
 end;
Function AD_converter(RS232_address, mode, outputdata,
                    Others:integer):integer; export;
{RS232_address, Base address of the selected RS232 port}
{mode: select analogue multiplexer mode}
{outputdata: digital output word (DB0 to DB6 bit, 7 bit in
                    total)}
{others: for further expansion}
var
   ii, Single_differential, Odd_sign, dummy_byte:byte;
   IO_data: array[1..12] of byte;
   data:array[1..12] of integer;
   Digital_data:array[1..12] of byte;
   binary_weight, dummy:integer;
Procedure delay;
{A short delay}
var
                    ij:integer;
begin
                    for ij:=1 to 6 do ij:=ij;
end;
Procedure AD_control(datax:byte);
{procedures for controlling A/D converter, serial-in latch
                    and parallel-serial shift register}
var
                    ij:integer;
begin
                    port[RS232_address+4]:=1+2*datax;
                    {CLK=0, Dout=datax, start bit=1}
                    port[RS232_address+4]:=0+2*datax;
                    {CLK=1, Dout=datax, start bit is
                    clocked into the A/D converter}
          delay;
          port[RS232_address+4]:=1+2*datax; {CLK=0,
                    Dout=datax}
          delay;
```

configures the RS232 port specified by RS232_address to a mode required by the RS232 data logger/controller. In,

```
AD_converter(RS232_address, Mode,
    Output_byte, Other :integer)
    :integer,
```

data is read from the a-to-d converter and latches Output_byte to the seven outputs. The function returns the a-to-d conversion result in integer. RS232_address should be supplied. Mode 1, 2, 3 or 4 selects the input mode of the a-to-d converter while Output_byte is an integer ranging from 0 to 127. The value 'Other' is reserved for future expansions.

```
Inputdata(RS232_address:integer)
    :integer
```

reads the seven inputs into the computer. RS232_address should be supplied.

Applications of the logging system

The present RS232 data logger/controller can be used with virtually any type of computer as long as it has a standard RS232 port. It combines a-to-d converter, digital input and digital output in a single package, therefore it provides a versatile solution for digital control applications.

Low power consumption makes it possible for battery powered applications. Together with a lap or palm top computer, it can be used as a mobile data logging and control centre. Many functions can be added to the system. including a multichannel analogue multiplexer to provide more analogue input channels and sensors and signal condition circuits to measure temperature, pressure, light intensity, magnetic field, weight, flow rate of fluids, biosignals, etc. ■

List 4.

RS232_address = RS232(x)
dummy = Configure_RS232(RS232_address)
Voltage = AD_converter(RS232_address, 1, com_byte, 0) * 5.02 / 4096
Data_input = inputdata(RS232_address)

```
end;
Procedure Configure_mode;
{Assign values for Odd_sign, Single_differential}
{Mode 1, Single mode, Channel 0
 Mode 2, Single mode, Channel 1
 Mode 3, Differential mode, Channel 0 positive, Channel 1
                 negative
 mode 4, Differential mode, Channel 1 positive, Channel 0
                 negative}
begin
                 case mode of
                 1:  begin Odd_sign:=1;
                 Single_differential:=0; end;
                 2:  begin Odd_sign:=0;
                 Single_differential:=0; end;
                 3:  begin Odd_sign:=1;
                 Single_differential:=1; end;
                 4:  begin Odd_sign:=0;
                 Single_differential:=1; end;
        else  begin Odd_sign:=1; Single_differential:=0;
                 end;
        end;
end;
Procedure configure_output;
{Assign IO_data[ii] according to OUTPUTDATA, ii=1 to 12,
                 OUTPUTDATA should be 0-127}
var
                 ij:integer;
begin
                 for ij:=1 to 4 do IO_data[ij]:=0;
        IO_data[5]:=1-Outputdata and 64 shr 6;
        IO_data[6]:=1-Outputdata and 32 shr 5;
        IO_data[7]:=1-Outputdata and 16 shr 4;
        IO_data[8]:=1-Outputdata and 8 shr 3;
        IO_data[9]:=1-Outputdata and 4 shr 2;
        IO_data[10]:=1-Outputdata and 2 shr 1;
        IO_data[11]:=1-Outputdata and 1;
end;
begin
   configure_mode;
   configure_output;
   Binary_weight:=4096;
   port[RS232_address]:=0;   {TD sends data. -CS goes from
                 low to high for a short period of
                 time. then goes back to low}
   repeat delay until port[RS232_address+2] and 1 =0;
   for ii:=1 to 120 do delay;
   AD_control(0);                {Start}
   Digital_data[7]:=(1-(Port[RS232_address+6] and 32)  shr
                 5);
   AD_control(Single_differential);     {}
   Digital_data[6]:=(1-(Port[RS232_address+6] and 32)  shr
                 5);
   AD_control(Odd_sign);
   Digital_data[5]:=(1-(Port[RS232_address+6] and 32)  shr
                 5);
   AD_control(0);
   Digital_data[4]:=(1-(Port[RS232_address+6] and 32)  shr
                 5);
   for ii:=1 to 12 do
   begin
        Binary_weight:=binary_weight div 2;
        Port[RS232_address+4]:=0+2*IO_data[ii]; {CLK=1,
                 Dout=Datax[ii]}
        delay;
        Port[RS232_address+4]:=1+2*IO_data[ii]; {CLK=0,
                 Dout=Datax[ii]}
        delay;
        dummy_byte:=(Port[RS232_address+6]);
        data[ii]:=(1- (dummy_byte and 16)  shr 4) *
                 Binary_weight;
        if ii<4 then  Digital_data[4-ii]:=(1-(dummy_byte
                 and 32)  shr 5);
   end;
   dummy:=0;
   for ii:=1 to 12 do dummy:=dummy+data[ii];
   AD_converter:=dummy;
   Dummy:=0;
   Binary_weight:=1;
   for ii:=1 to 7 do
                 begin

                 dummy:=dummy+digital_data[ii]*binary_
                 weight;
                     Binary_weight:=binary_weight*2;
                 end;
        Port[RS232_address+7]:=dummy; {input digital data
                 is stored in the scratch-pad register
                                offset 07 of
                 the UART}
end;
Function Inputdata(RS232_address:integer): integer;
                 export;
{Read digital input data (7-bit) from the scratch-pad
                 register, offset 07 of the UART}
var
                 ij:integer;
begin
                 for ij:=1 to 10 do ij:=ij;
                 Inputdata:=port[RS232_address+7];
end;
Exports
        RS232            index 1,
        Configure_RS232            index 2,
        AD_converter               index 3,
        Inputdata        index 4;
begin
end.
```

Charge pumps get new life

Charge-pump voltage converters are an old idea brought up to date by integration, the need for smaller power supplies and the attraction of inductorless circuitry explains Philip Darrington.

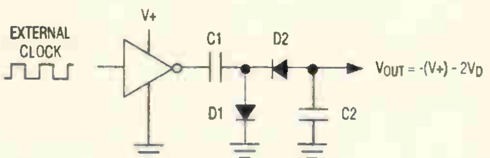
Charge pumps, diode pumps, switched-capacitor voltage multipliers, call them what you will; they all reduce to a couple of capacitors and two switches, in the simplest case at least (Sir John Cockroft and ETS Walton, who originated the circuit in 1929, used multiple diodes and capacitors to obtain 710kV for the first successful proton accelerator in 1932, accuracy and efficiency not being of prime importance). In early circuits of this type, currents were low and noise obtrusive; their use as sources of power was therefore unattractive, except that the circuits used no inductive components – always an inducement.

In recent years we have seen something of a revival in the use of charge pumps to supply small amounts of power in portable, battery-powered equipment, the avoidance of inductors again being a powerful argument. There are new integrated circuits that overcome the drawbacks of inaccuracy, low output current and noise so effectively that their use in rf work is now attractive and current consumption approaches that of supplies using inductors. Maxim produces such ics and the *Maxim Engineering Journal*, vol.24, contains an article on the subject, of which this is a resumé.

Figure 1(a) is about as basic a charge pump as you can get. On positive-going pulses, C_1 charges by way of D_1; as the input goes negative, D_1 cuts off and the charge on C_1 passes to C_2, building up as succeeding pulses arrive. Charge on C_2 is held by D_2 and is the output voltage, which is smoothed to some

Choice of pump capacitor

Data sheets for charge pumps usually mention only one or two capacitor values, but it is worth pointing out that they will work well with a wider range of capacitors, in particular at low output currents. In general, specify the smallest value to give the required level of output voltage, current and ripple. Ceramic types can be used at lower output currents; some manufacturers now make ceramics with 10µF capacitance at low cost.



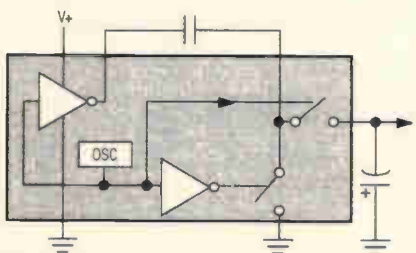


Fig. 1. Charge-pump circuits in original form (a) and in modern guise (b), where the diodes are replaced by cmos switches.

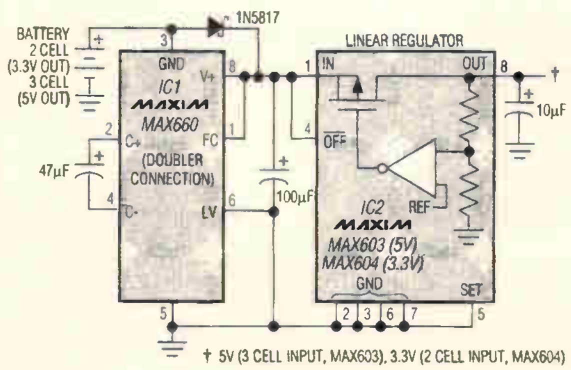


Fig. 2. Basic charge-pump circuits are unregulated; one way of regulating the output is to use a separate regulator. This combination provides 200mA at 3.3V from a two-cell battery or 150mA from three cells.

extent by the presence of C_2. In the ic circuit of Fig. **1(b)**, the diodes are replaced by cmos switches, driven by the internal clock, but the principle of operation remains the same. Output currents of ic pumps are increasing as input current decreases. As an example, the circuit of **Fig. 2** puts out 100mA at 3.3V from a two-cell battery of the AA type or one lithium cell, holding the 3.3V output for inputs down to 2.2V; with inputs over 2.4V, it will supply 200mA for a short time. Efficiency is nearly 80% with a low input voltage; just over 50% with a higher input.

Regulation

There is no inherent regulation in the basic charge pump. One way round this is the Fig.2 circuit, in which a separate regulator comes after the pump, but it is also possible to regulate the charge pump ic itself either by combining the two ics in Fig.2 into one or by modulating the pump operation by varying switch resistance. Linear regulation is less noisy and so is preferred for jobs such as bias generation in GaAsfet rf amplifiers; on the other hand, modulation costs less and gives more output current, other things being equal, since there is no need for a series-pass transistor. **Figure 3** is an example of the modulation method and **Fig. 4** its efficiency plotted against input voltage. The step occurs at the point where the circuit automatically changes from being a doubler to behaving as a tripler under the control of the amplifier and feedback; highest efficiency in each zone is given by the lowest input voltage, losses being smallest under that condition.

Although there is no series pass transistor, losses are the same as in a linear regulator, since there is an energy loss each time the pump capacitor voltage changes in a cycle when they are placed in parallel rather than in the series configuration for charging.

Current requirements

Current draw in this type of voltage converter is small, which is useful when they are to be used in small, hand-held equipment that spends much of its time asleep. For this reason, the light-load operating current is often more important than full-load efficiency as far as battery life is concerned; it makes little sense for the power supply to use as much current as the load

For a charge pump, supply current is generally proportional to switching frequency, so that current needs are smallest at low frequencies, but this means more ripple, less output current and larger pump capacitors, at least in older designs. Pin-settable switching frequency is sometimes seen. Newer types take advantage of the application by supplying output current when the load needs it, the circuit of Fig. 3 being one such, where the SHDN (shutdown) pin turns the oscillator on or off to give a no-load supply current of 75μA. Full output current is 50mA with 0.22μF pump capacitors.

A technique that allows a very small quiescent current and high output current when required is 'on-demand' switching, in which the control amplifier, for example in the *MAX619* of Fig. **3**, monitors the output voltage and switches the oscillator on only when the output falls below 5V.

Application

Charge-pump voltage converters find a nearly ideal application in the provision of a programming voltage for flash memory chips; in volumes of the size of credit cards the absence of electrolytic capacitors is virtually essential. The circuit in **Fig. 5** shows an ic for this purpose that supplies a 12V V_{pp} to program 2byte words of flash memory, while the circuit already seen in Fig. 2 will give 5V for 5V flash. As can be seen, the charge-pump converter allows voltage conversion in situations in which linear regulation has, until recently, been the only solution.

All, however, is not unalloyed rapture; there is also the flip side, whose name is noise. When a capacitor is connected to another at a different voltage, as happens at switch-over in the pump circuit, current flow is only limited by capacitor esr and switch resistance of around 5Ω, so filtering is needed to reduce it.

Figure 6 shows a circuit using the *MAX850* in which the effect is small and is used to provide negative bias for GaAsfet rf amplifiers. It gives

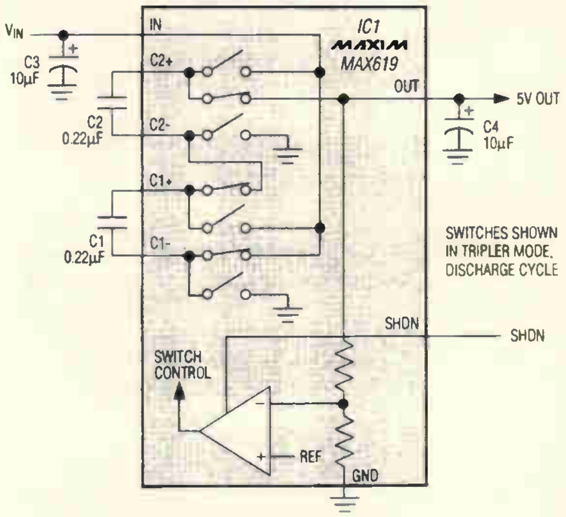


Fig. 3. Regulation comes on-chip in this MAX619 in the form of switch-resistance control feedback. The circuit doubles or triples input voltage for increased efficiency.

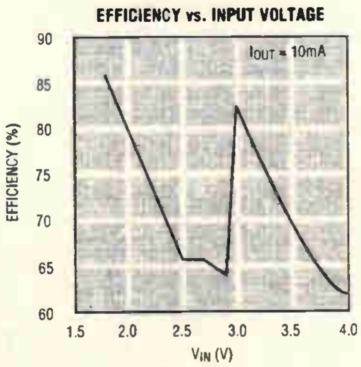


Fig. 4. Efficiency of the Fig.3 circuit varies, depending on whether the circuit is doubling or tripling, the changeover being automatic.

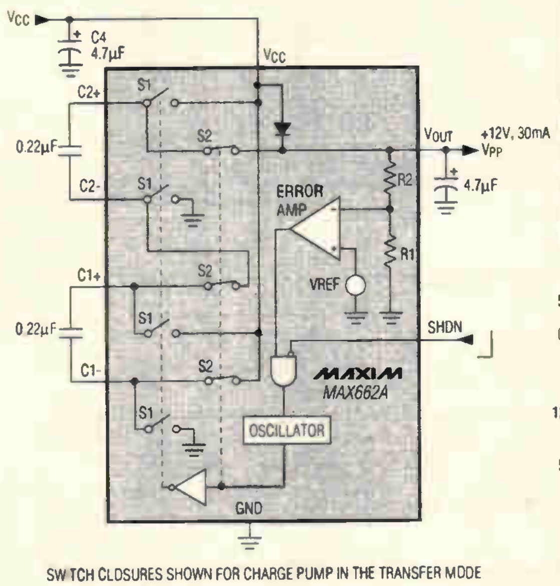


Fig. 5. 12V programming supply for flash memory cards, regulated for loads of 30mA. Ceramic capacitors replace electrolytics to give a physically small circuit.

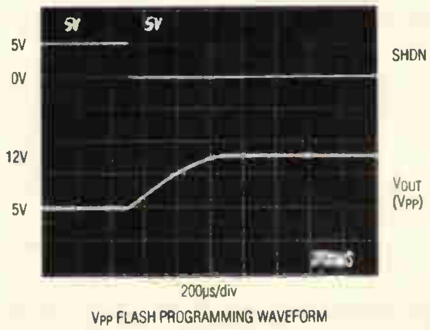
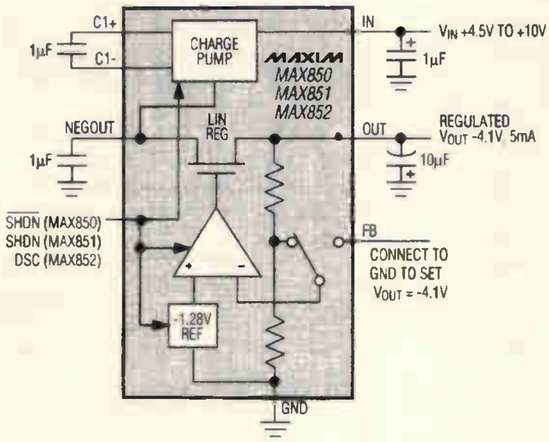
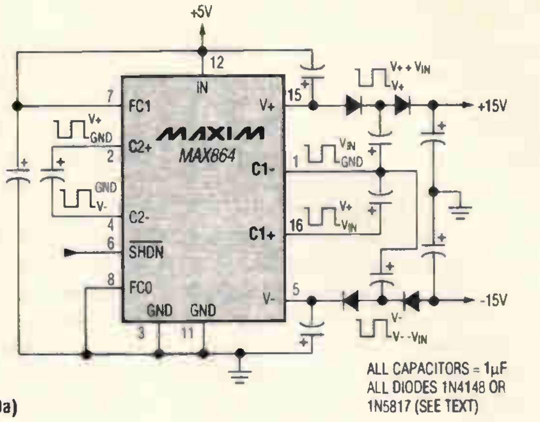


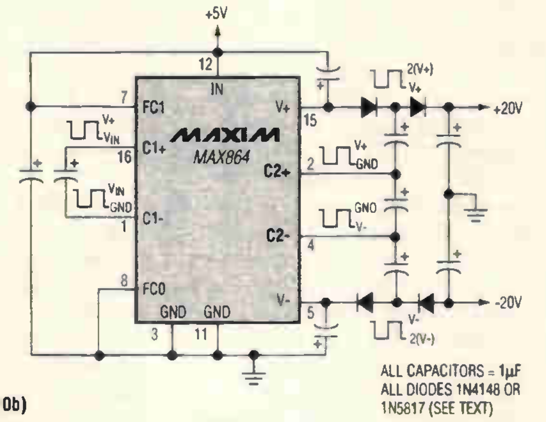
Fig. 6. Negative-output converter, working at high frequency to allow the use of small capacitors. The linear regulator reduces output ripple and noise to 2mV pk-pk.





(10a)

Fig. 7. Tripling (a) or quadrupling (b) the input voltage by external diode/capacitor networks still results in a space-saving circuit.



(10b)

–4.5V out and runs from +5V in. It consists of an inverting charge pump followed by a negative-output linear regulator. It works at 100kHz to allow the use of small external capacitors and the regulator section reduces noise and ripple to 2mV pk-pk.

Enhancements

Although discrete diodes are no match for the cmos switches used in the ics mentioned as regards voltage drop, you can make a very small 5V to ±20V converter by using a boost stage of discrete diodes for use in ccd power supplies, lcd bias and varactor tuners. A *MAX864* used on its own will generate ±10V less losses from 5V input or ±6.6V from 3.3V in; additional diode-capacitor stages approximately double the outputs to $\pm 4V_{in}$ or increase them by 1.5 to about $\pm 3V_{in}$, depending on the connection of the external network, as shown in **Fig. 7**. The 1µF filter capacitors give under 100mV of ripple and supply 20mA; slightly larger capacitors will lower the ripple considerably. At an operating frequency of 100kHz, and with 1µF capacitors, no-load supply current is 7mA. Pin-programming a lower frequency will reduce that to 600µA, but then you will need to use 10µF capacitors.

Figure 8 shows the generation of a negative output greater than the positive input voltage, a feat not normally possible, with the use of additional diodes; the circuit shown gives -8V or more from inputs of 2.5-5.5V.

Low-power computer peripherals can obtain power from the computer's serial port. Mice and other devices use the modem signals DTR and RTS, but the circuit in **Fig. 9** takes power from the TX line of a three-wire port. It gives 8mA output, enough for a cmos microcontroller and a few bits of logic. TX idles at a negative voltage, so that the ic's input polarity is reversed, the negative input between output and ground enabling it to pump backwards. The 4.7V zener provides regulation. ■

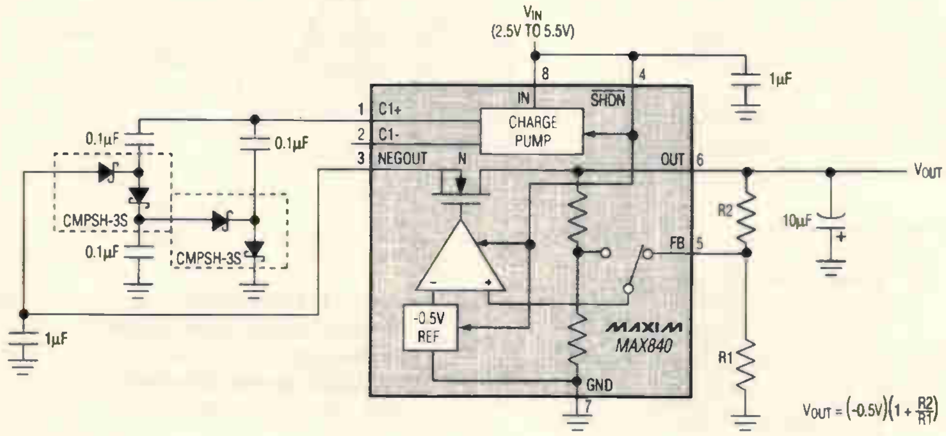


Fig.8. External circuit produces a negative output at a voltage greater than the positive input.

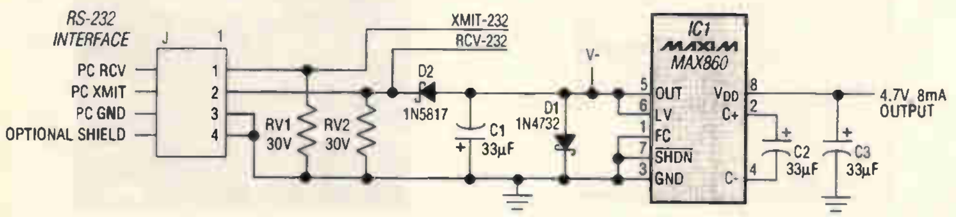


Fig.9. Taking power from the TX line of a computer's serial port, this MAX950 with external diodes will power a microcontroller. The zener shunt assists with regulation.

Technical support
Maxim, Unit 3, Theale Road Technical Park, Station Road, Theale, Berkshire RG7 4XX, tel. 01734 303388, fax 305577.

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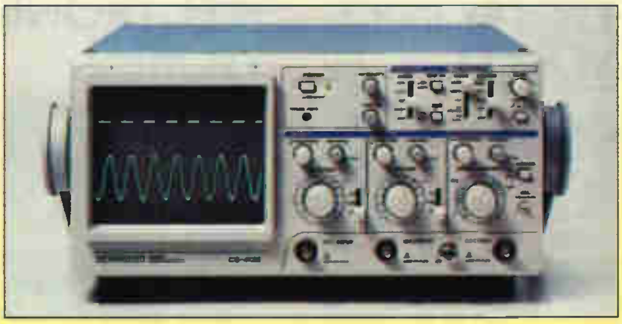

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5mV to 5V/div	DC-20 MHz –3 dB or 10 Hz to 20 MHz –3 dB on AC range
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Crosstalk	–40 dB
Operating modes	CHI: CH1 single-trace CH2: CH2 single-trace ALT: alternate CH1 and CH2 display CHOP: chopping display of CH1 and CH2 ADD: Combined waveform of CH1 and CH2
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Sensitivity	same as vertical axis (CH2)
Response	DC: DC to 500kHz –3dB AC: 10Hz to 500kHz –3 dB X-Y phase matching within 3° at 50kHz
Operating modes	CHI: Y axis, CH2: X axis
Sweep	
Modes	NORM: trigger sweep AUTO: auto free-running with no signal
Sweep time	0.5 μs to 0.5s/div ±3% (0.2μs/div uncal.), 1-2-5 steps, 20 ranges w. fine adjustment
Sweep magnify	x10 ±5% (20ns/div uncal.)
Triggering	
Trigger sources	VERT MODE: input signal selected in VERTICAL mode CHI: CH1 input signal CH2: CH2 input signal LINE: commercial power supply EXT: EXT. TRIG input signal
External triggering	
Input impedance	1MΩ and 22 pF approx.
Coupling modes	AUTO, NORM, FIX: AC coupling TV-FRAME: TV-LINE:
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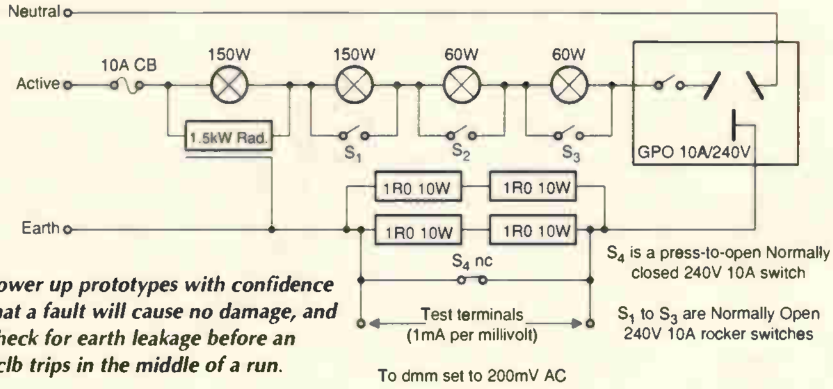
CIRCUIT IDEAS

Fire up prototypes without the fire



Power up prototypes with confidence that a fault will cause no damage, and check for earth leakage before an eclb trips in the middle of a run.

The moment of expensive truth when a prototype is first connected to the mains is fairly often accompanied by flashes, loud reports and the disgusting smell of baked resistor. This need not be; this circuit carries out the switch-on gently and gracefully, indicates trouble before it happens and has done for twenty years with no fuse blowing, circuit-breaker tripping or other regrettable events.

Protection and indication are both performed by (a) a 1.5kW radiator on the ceiling and (b) a 150W bulb in parallel with it, also on the ceiling where it can be seen, the two being wired in series from the live mains input to the socket to be used via three lamps of different powers with shorting switches.

Various combinations of the shorting switches let you limit the possible load current to less than the radiator element. Provided that the load is under 1500W, a fault is shown by the 150W bulb across the radiator lighting up brightly and staying that way. Switched-mode power supplies cause the bulb to light on the initial current surge, but it then fades out unless a capacitor or diode is faulty, in which case it stays lit.

Earth leakage testing is the reason for the 1Ω resistors. At the college at which I teach, there is a room with 15 computers, several programmable logic controllers also being in use. All are connected via earth leakage circuit breakers which appear to trip in a somewhat random manner except that it is usually in the middle of a plc programming session.

Voltage across the resistor network is simply measured on a true-rms digital meter set to 200mV or, in our case, 20mV full scale for better resolution. Other voltage measuring methods are possible, but this works well.

Do not be tempted to increase the value of the resistor network for improved resolution; a 10Ω resistor would overheat in the presence of a short.

H Peter Harle
Mt Druitt College of TAFE
Mount Druitt
NSW Australia

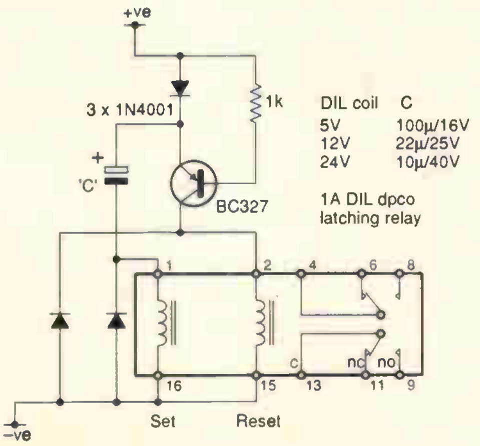
2μA relay switching

Needing to switch a battery-powered, remote rf amplifier into a receiving antenna feeder without my physical attendance in all weathers, I designed this simple latching relay circuit. It hardly affects battery life and switches automatically when power is applied to the amplifier or removed.

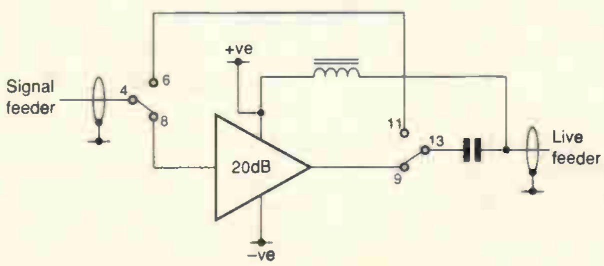
Applying power to the amplifier via the rf feeder charges the capacitor through the 'set' coil. The transistor is reverse-biased and signals from the antenna go through the amplifier. When power is removed, current flows through the 1kΩ resistor, the transistor discharges the capacitor *C* through the reset coil and the rf amplifier is by-passed.

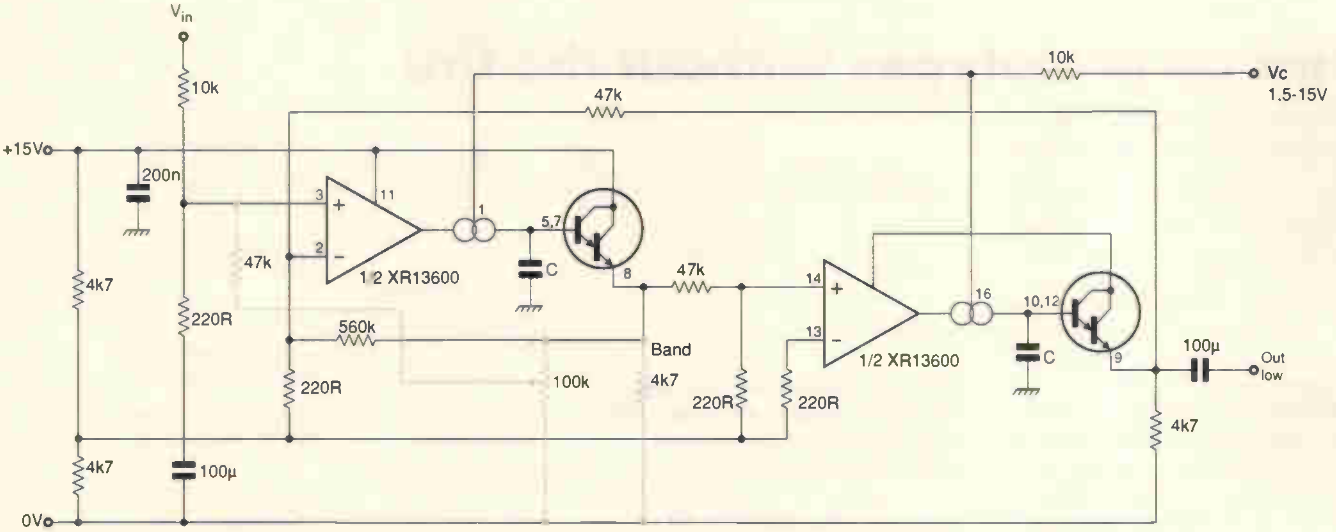
The circuit only needs 2μA to overcome capacitor leakage and there is no need for constant relay energising or diode switching currents. On a 6V supply and using the *RS 369-595* dil latching relay, turn-on current is 25mA for a few milliseconds; initial capacitor charge is used again at turn-off.

Graham Maynard
Newtownabbey
Northern Ireland



Latching relay circuit for remote antenna amplifier switching uses only 2μA and gives a clean switchover. With a power relay and the C values shown, it would be useful for power control use.





State-variable active filter using transconductance amplifiers is voltage controlled and, with the additions shown dotted, becomes a voltage-controlled sine-wave oscillator.

Simpler voltage-controlled filter/oscillator

Using transconductance op-amps renders the design of voltage-controlled filters and oscillators comparatively simple.

Originally designed as a state-variable active filter with its frequency range determined by the value of *C* and the control voltage, the circuit shown functions as an sine oscillator with the additions shown dotted.

For a control voltage range of 1.5-15V, and *C* at 22nF, the frequency range is 6Hz-400Hz; with *C* at 330pF, the range is 400Hz-22kHz.

Distortion when working as an oscillator is reduced by the addition of the potentiometer, the output frequency then depending on the pot. setting and the value of *C*.

Kamil Kraus
Rokycany
Czech Republic

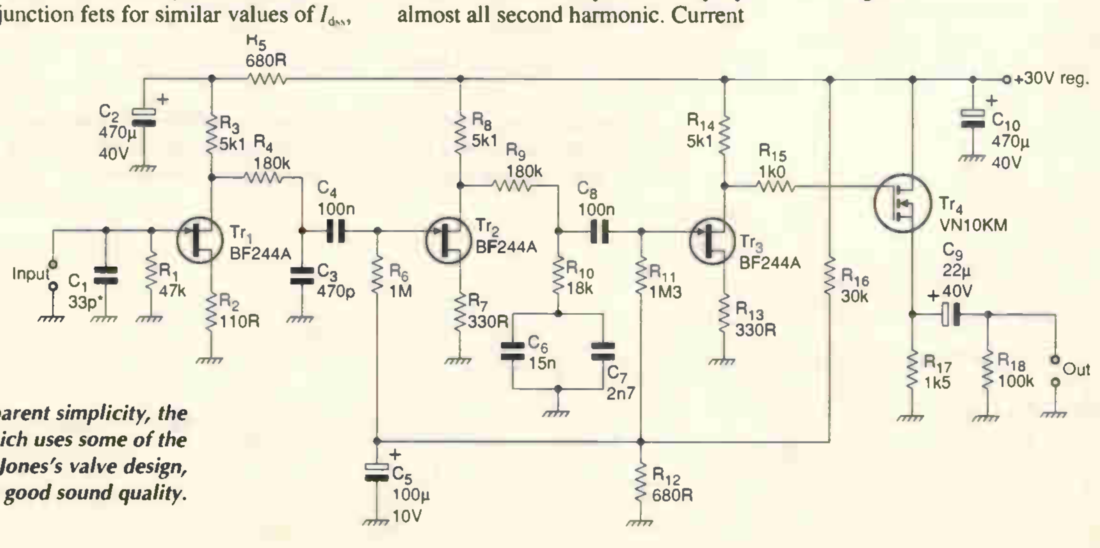
Fet preamplifier has no overall feedback

Preserving some of the important features of Morgan Jones's excellent valve preamplifier design, such as passive RIAA correction and zero overall feedback, this all-fet circuit produced good sound quality in informal listening tests.

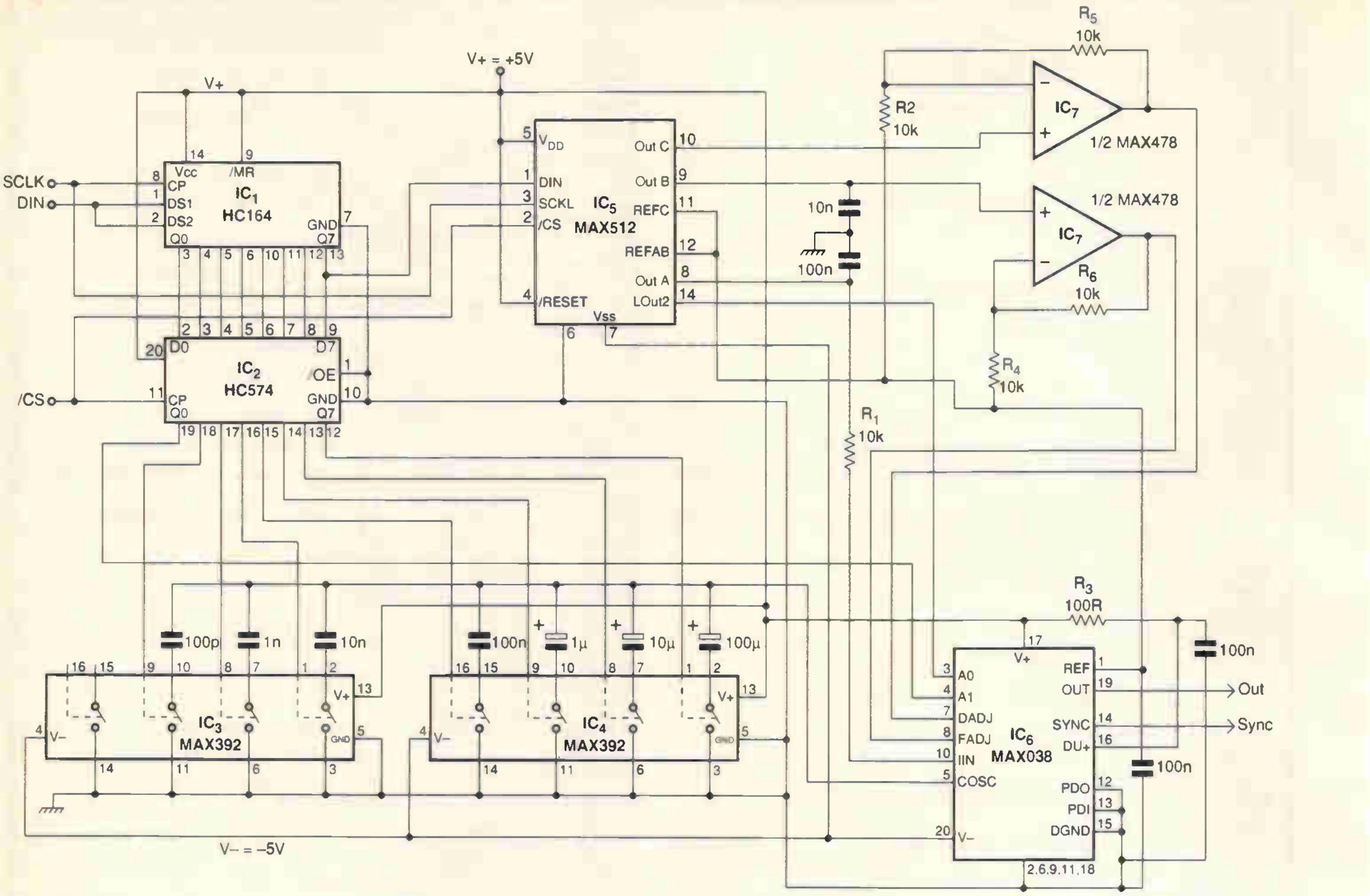
I found it necessary to select the junction fets for similar values of I_{dss}, namely 4.8mA±0.2mA. It seems that, if the transistors are from the same batch from the same manufacturer manufacturer, their other characteristics should match well.

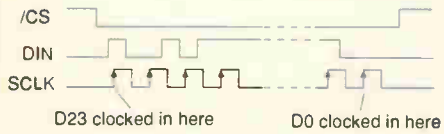
Voltage gain of the circuit shown is 38dB at 1kHz, distortion being less than 0.2% for an output of 1.2Vpk-pk, almost all second harmonic. Current consumption is 20mA per channel and the circuit derives its power from an *LM317* regulator. Adjust C_1 for correct termination of the pickup/tonearm.

Laszlo Gaspar
Nottingham Trent University
Nottingham



In spite of its apparent simplicity, the phono preamplifier, which uses some of the ideas of Morgan Jones's valve design, produces good sound quality.





Three-wire serial interface controls frequency, duty cycle and waveform of sine, square and triangle generator. Bypass all supplies with 1μF electrolytic in parallel with 1nF ceramic.

£100 WINNER

2.5MHz generator with serial control

An SPI or QSPI three-wire serial interface controls waveform, duty cycle and frequency of a *MAX038* waveform generator through a *MAX512* triple digital-to-analogue converter.

In **Fig. 1**, Dac A of IC_5 and R_1 form the coarse frequency control by providing the current to charge whichever capacitor is connected to C_{OSC} on pin 5 of IC_6. Duty cycle and fine frequency control come from Dac B and Dac C outputs of IC_5 to DADJ and FADJ on IC_6. Quad analogue switches $IC_{3,4}$, controlled by octal latch IC_2, connect one of seven capacitors for each frequency range to C_{OSC} on IC_6. Frequency and duty cycle are determined by the formulae:

$$f_{coarse}=\frac{V_{ref}A}{C\times 256\times 10000},$$

$$f_{fine}=\frac{-V_{ref}B-128}{128\times 0.0343},$$

$$Duty(\%)=\frac{V_{ref}C-128}{128\times 0.0575},$$

where V_{ref} is in volts, C_{OSC} in farads, frequency in hertz and A, B and C are the dac codes.

Dac outputs can swing between $\pm V_{ref}$ or ±2.5V, but should be limited to ±2.4V from dac B and ±2.3V from Dac C. The processor in control of affairs determines frequency by counting SCLK pulses or measuring their period.

As regards the input, SCLK low-to-high transitions clock in serial bits, each 24-bit word being latched into IC_2 by a low-to-high on /CS, as seen in **Fig. 2**. One */CS* transition latches the first 16 into IC_5 and the last eight into IC_2, the IC_5 bits consisting of seven for control and eight for data. Seven of the IC_2 bits select the range capacitor, the eight and the latched auxiliary control bit from IC_5 LOUT together form a two-bit word to select sine, square or triangular waveforms.

Since there is vestigial resistance to ground through the analogue switches of about 20Ω, the triangular waveform is slightly distorted. If this is important, it can be avoided by using reed relays instead or, if only one range is needed, hard wiring the capacitor.

Terry Millward
Reading

Output frequency ranges, in hertz, for the various C_{osc} capacitor options.

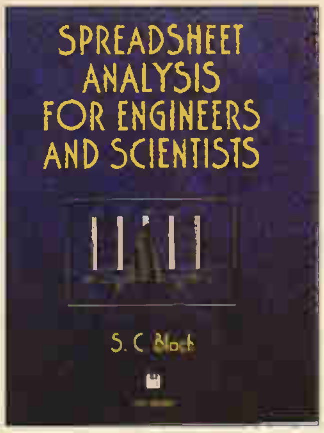
C_{OSC}	Frequency Minimum (Dac A code=1)	Maximum (Dac A code=255)
100μF	9.77E–3	2.49
10μF	9.77E–2	2.49E+1
1μF	9.77E–1	2.49E+2
100nF	9.77	2.49E+3
10nF	9.77E+1	2.49E+4
1nF	9.77E+2	2.49E+5
100pF	9.77E+3	2.49E+6

Input word bit decriptions, * is active high.

D_0 clocked in last and, together with LOUT from IC_5 (i/p bit D_{22}), selects waveform:

	square	triangle	sine
D0	0	0	1
D22	0	1	x

D_1 connect 100pF to C_{OSC}
$D_{2,3,4}$ 1nF, 10nF, 100nF resp.
$D_{5,6,7}$ 1μF, 10μF, 100μF resp
$D_{8\text{-}15}$ dac data lsb-msb
$D_{16,17,18}$ load dacs A, B, C resp.*
$D_{19,20,21}$ shut down dacs A, B, C resp.*
D_{22} select o/p waveform
D_{23} not used



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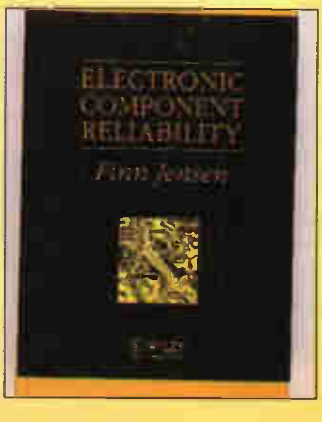

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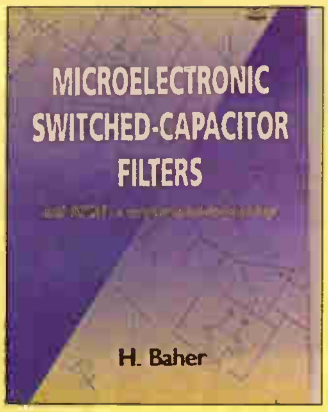

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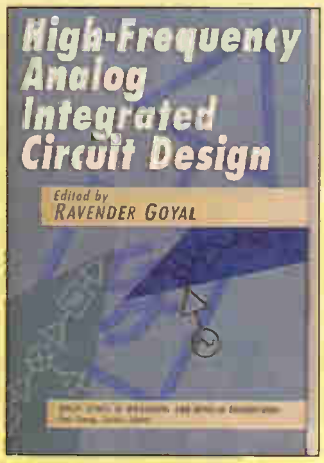

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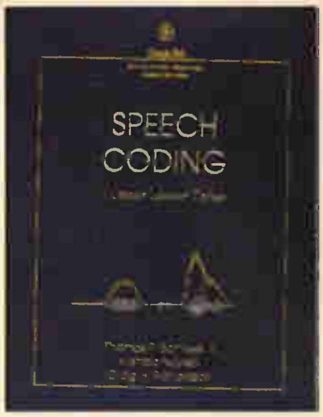

perform a wide variety of speech coding and speech processing experiments. The text presents and explains a set of basic speech coders analytically and in terms of the specific parameters controlling each coder. The manual leads the student through the experimental process of understanding how speech 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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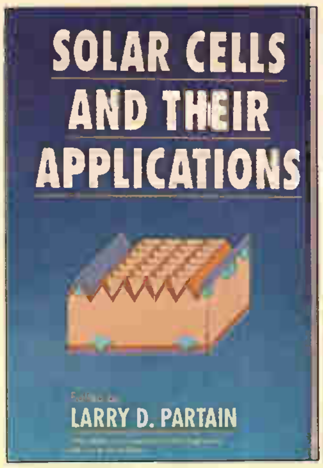

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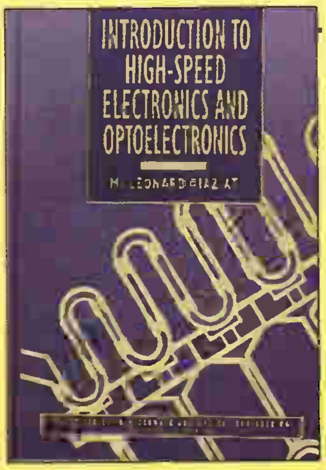

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Discoverer of the electron, J J Thomson, was Cavendish Professor of Experimental Physics, Cambridge University.

Celebrating 100 years of electronics, Tom Ivall looks back to J J Thomson's 1897 discovery of the first elementary particle.

Centenary of *the electron*

"The assumption of a state of matter more finely subdivided than the atom is a somewhat startling one..."

Perhaps the Victorian public audience at London's Royal Institution, hearing this remark by Professor J J Thomson of Cambridge University, was indeed somewhat startled by the idea. The occasion was an evening meeting on Friday 30 April 1897, and the remark came near the end of the professor's lecture on 'Cathode Rays' – a hot scientific topic at that time.

Then right at the end of the presentation, in a brief, almost casual fashion, Thomson mentioned a set of results from his experimental work on cathode rays at the Cavendish Laboratory. These suggested he had established the existence of a particle which was subatomic and had a definite mass and electric charge. He called it a corpuscle.

Some authorities maintain that the formal announcement of the discovery was actually the printed report of a British Association meeting held in September 1897, at which Thomson gave a further account of his work. But in any case this public disclosure did not have much impact on the electrical science and technology of the day. In 1897 engineers were rushing ahead with electrical machines, the telegraph and telephone, radio communication and many other useful devices, quite unaware that it all depended on the great mobility of this very corpuscle. Electric current was seen as a kind of fluid.

The birth of particle physics

Now, of course, we understand the particle as the central fact of electronic engineering. Its behaviour in solids, gases and vacua has been intensively studied and exploited. What we also accept, almost dismissively without thinking much about it, is the tremendous advance that this discovery made in our understanding of the structure of matter.

Up till then the atom, as pictured by Dalton, was the basic unit of matter, indivisible and indestructible. With this new discovery the atom was 'split' – or at least a bit was chipped off it. Thus the electron was the first elementary particle to be identified. We had the beginning of the modern science of particle physics.

The most immediate effect was in chemistry – a new concept of how atoms were held together in molecules. The electronic theory of valency provided a fresh explanation of the mechanism of chemical bonding. A planetary model of the atom emerged, with outer electrons which formed bonds by transfer or sharing with other atoms.

Later this picture of orbiting particles was modified by quantum mechanics. Here the electrons were seen as having the dual properties of particles and waves. This wave mechanical concept of the electron was in a sense the completion of the discovery. Indeed when Dr Arnold Lynch of University College London recently gave an IEE lecture on Thomson's 1897 investigations he entitled it 'Half the electron'. Yet physicists are now considering whether the electron might not be fundamental, as originally thought, but made up of even smaller particles.

Exploring the electron

So how did Thomson make his discovery? He was investigating cathode rays, which were already thought to be streams of negatively charged particles. He established that they were indeed real particles by showing that they had mass and electric charge. He also found that this mass was only a tiny fraction of that of the lightest atoms then known, so the particles could not themselves be atoms.

Furthermore, he discovered that the velocity of the particles was very much lower than the speed of light, so the cathode rays could not be some form of electromagnetic waves.

By measuring the deflection of cathode rays in known magnetic and electric fields, Thomson was able to calculate the ratio of the charge to the mass of the assumed particles, e/m. This ratio came out as extremely large – about 7×10^7 coulombs/gram. In 1897 the smallest particles possible were thought to be hydrogen atoms. So, either the charge was unconvincingly high, assuming the charged particles to be equivalent to hydrogen atoms, or the mass was very low – much less than that of hydrogen atoms. Experiments and deductions showed the latter alternative to be the case.

Later on, Thomson inserted an available estimate for e in the ratio in order to calculate the actual value of the mass, m. Here he made use of earlier work of physicists in attempting to discover a fundamental unit of electricity.

Related findings

Throughout the 18th and 19th centuries there had been much scientific interest in the discharge of electricity through the residual gases in evacuated glass vessels and tubes. Glow effects were observed from as early as 1709 onwards.

In 1838, Faraday noted that the appearance of the discharges changed with reducing air pressure. Plücker observed a difference between the light from the ordinary discharge and a separate greenish glow that appeared on the glass. Both he and, later, Crookes found that this fluorescent area on the glass could be deflected by a magnet. The direction of this bending, of course, was an indication of the sign and direction of the flow of current carriers.

Goldstein described what was apparently originating from the negative electrode as the 'cathode stream', but really strong evidence of the existence of rays came around 1870, when Hittorf in Germany and Crookes in Britain placed an obstacle inside the tube. This cast a corresponding shadow on the area of green fluorescence on the glass.

But at this stage nobody really knew whether these cathode rays were waves in an aether or particles. Throughout the 1880s there was a great controversy about it. In the main, German scientists favoured waves and their British contemporaries particles. Thomson was fully aware of this controversy.

It was left to the French physicist Perrin to settle the matter. He showed that when cathode rays were directed onto a metal cylinder it gradually acquired a large negative charge. He concluded that the rays were streams of negatively charged particles, and everyone agreed.

If they were in fact negatively charged particles, what kind of particles? Were they molecules, atoms or something smaller? Answers to these questions would establish beyond doubt the reality and nature of these mysterious entities.

Proof of the divisible atom

This is where Thomson's experiments turned out to be a huge step forward. It was his work alone that demonstrated that the particles could not be Dalton's indivisible atoms but had to be something smaller – in short, subatomic.

Using various adaptations of the Crookes tube, he did a series of measurements on cathode rays that he hoped would throw more light on their nature. He measured the positions where the beams made patches of fluorescence on the tube glass. He measured the magnetic and electric fields which deflected them. He measured electric charges and heat energies.

From the numerical results he was able, as already mentioned, to calculate the e/m ratio and the velocity of the assumed particles.

Probably the most significant of Thomson's measurements – crucial to the whole discovery – was the deflection of the cathode rays by an electric field. He was able to achieve this because, technologically, he had an advantage over earlier scientists in having better vacuum pumps available, which allowed him to reduce the gas pressure inside the tube to a much lower value – about 1.3 N/m^2, which is around 10^{-2}mmHg).

This meant that, for the first time, a static electric field could be maintained inside the tube for deflection without being discharged through too much residual gas made conductive by the cathode rays. Previously, other scientists had thought that their inability to deflect the cathode rays electrostatically showed that the rays were really waves in an aether.

Reading Thomson's original lectures and papers (see reference) is a somewhat difficult task. He did not present the different stages of his work in chronological order and, of course, he used the concepts, terminology and measurement units that were current in the late 19th century. The following outline, therefore, although sticking to the basic principles and experimental facts, is a version written more in line with the language of modern electronics technology.

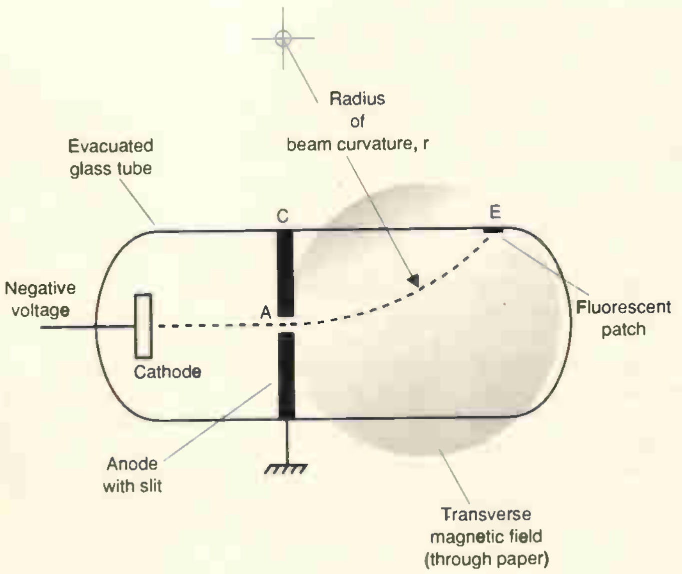

Fig. 1. Measurement principle used in establishing the charge-to-mass ratio of the electron. The real cathode-ray beam and fluorescent patch were much more diffuse than suggested here.

Thomson's experimental hardware

In his experiment to measure the magnetic deflection of cathode rays, Thomson used a form of Crookes tube arranged as in **Fig. 1**.

Coming from a cold cathode at a negative potential, the cathode rays were accelerated by an anode, at earth potential, which contained a slit. Passing through the slit, a thin beam of rays was deflected by a transverse magnetic field. Visualise the flux lines as passing through the paper. The deflected beam struck the inside surface of the glass and produced a fluorescent patch on it at a certain distance from the anode, CE.

A very uniform transverse magnetic field was produced from a pair of Helmholtz coils, as in the Helmholtz galvanometer. These coils are equal, coaxial, circular coils parallel to each other, spaced apart by a coil radius.

When a charged particle enters a transverse magnetic field it experiences a force at right-angles to its direction of motion. This force depends on the magnetic flux density B and on the charge e and velocity v of the particle. As a result the particle travels in a circular path and the force is centripetal. If the field is strong enough this becomes a closed circle and the constant centripetal force is mv^2/r, where m is the mass of the particle and r is the radius of the circle so formed.

In this state the centripetal and magnetic deflecting forces are equal, so one can write

$$mv^2/r=Bev$$

from which the ratio of charge to mass comes out as

$$e/m=v/Br$$

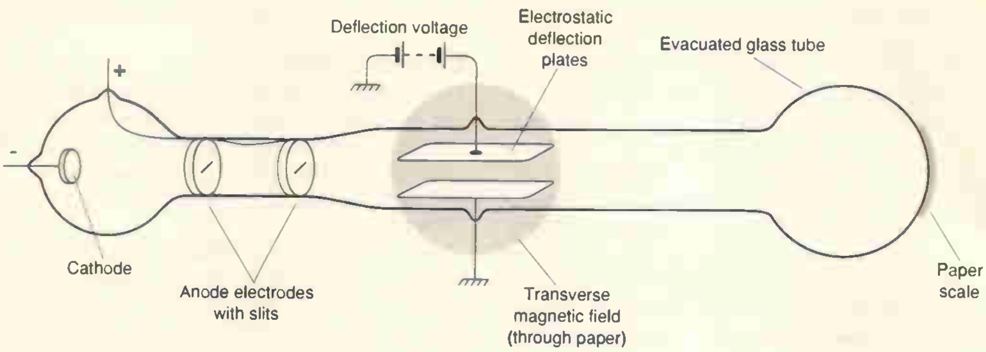

Fig. 2. By balancing the opposed magnetic and electrostatic deflections of the cathode-ray beam in this tube, Thomson was able to calculate the velocity of the electrons.

Thomson had the value of B from measurement. He found r, first by measuring the distance CE in Fig. 1, then, knowing the distance CA from the construction of the tube, by calculating the radius of the beam's curvature: $2r=(CE^2/CA)+CA$.

Finding the beam's velocity

But in the above formula for e/m the velocity of the particle, v, remains to be found. To arrive at v Thomson devised a form of Crookes tube as shown in **Fig. 2** – almost a precursor of the modern oscilloscope tube.

The cathode rays originated from a cold cathode. They were accelerated by an anode system formed by two circular metal electrodes which contained slits to channel the rays into a thin stream.

After passing through the magnetic and electrostatic deflection system, which allowed the fields to be varied, the beam ended on the inside of the glass bulb seen on the right. Here it produced a fluorescent patch, the position of which could be measured by a vertical paper scale pasted on the outside of the bulb – a simple calibrated screen for measuring deflection.

Again the uniform transverse magnetic field was generated by a pair of Helmholtz coils. The aluminium electrostatic deflection plates were about 5cm by 2cm and approximately 15mm apart.

The method of finding v with this tube was to balance the opposing forces exerted on the particles by the crossed magnetic and electric fields, so that their deflection angles were exactly nulled and the beam remained undeflected.

Thomson applied a direct voltage to the two deflection plates. The resulting electric flux lines were perpendicular to both the magnetic flux lines and the beam direction. Voltage polarity was chosen so as to oppose the magnetic deflection of the beam. He then adjusted the current in the Helmholtz coils until the observed vertical position of the fluorescent patch showed that the beam was not deflected.

In this state the electric field of field strength E exerts a force on the particles of Ee and this equals the force exerted on them by the magnetic field, which is Bev, as described above. So, if $Ee=Bev$ it follows that the velocity $v=E/B$.

The electric field strength E is V/d, where V is the applied voltage and d the distance between the deflection plates. Thus, Thomson was able to obtain a value for v and insert it into the equation for e/m given above. He found, in fact, that v was only about a tenth of the speed of light – an indication that the cathode rays were not electromagnetic waves.

Finding values for charge and mass

But, of course, e/m is only a ratio. Thomson arrived at actual numbers for e and m by assuming for the charge e a value which the Irish physicist Johnstone Stoney had proposed in 1874 as a fundamental unit, an 'atom of electricity'. It was Stoney who had already, in 1891, suggested the name electron for this fundamental unit, from the well known Greek origin. Thomson had persisted in calling his particle a corpuscle.

Electrolysis studies by Faraday, Arrhenius, Clausius and others had suggested that a definite amount of electricity was associated with a definite quantity of matter. Faraday gave the name ions to the dissociated products of electrolysis. Since these ions were moved under the influence of electricity it seemed reasonable to suppose that they were themselves charged. Thomson himself, using a cloud chamber, investigated the charges carried on water droplets.

Stoney had proposed that the charges were carried in integral multiples of a certain fundamental unit of charge, and this minimum charge, which he called the electron, he calculated to be about 10^{-19} coulomb. But he probably didn't think of his electrons as possessing mass.

When Stoney's proposed value for e was put into Thomson's calculated ratio for e/m it yielded a value for the mass of the particle. This turned out to be only about $^1/_{2000}$ of the mass of the hydrogen atom, a fraction later refined to $^1/_{1837}$.

Thomson followed up the initial experiments by trying different residual gases in the tube, different materials for the cathode and other electrodes and various accelerating potentials between anode and cathode. He found that the ratio e/m was independent of all these variations. It was obviously a universal relationship.

Supporting evidence

Other scientists validated his discovery in various experiments which gave better measurements of e/m. Lorentz obtained a more precise value for the electron mass from calculations based on the Zeeman effect. Millikan, in what became known as the oil-droplet experiment, greatly improved on earlier attempts by Thomson to measure charge directly.

The more precise value he obtained for e of 1.6×10^{-19}C, when inserted in e/m, resulted in a rest mass figure of 9.11×10^{-31}kg. This value has recently been refined to 9.1093897×10^{-31}kg. A recent precise value for e is 1.60217733×10^{-19}*C*.

Further work by Becquerel, Lenard, Rutherford and other scientists, including Thomson himself, showed that the negatively charged particles observed in the photoelectric effect, beta rays, thermionic emission and other phenomena were in fact the corpuscles Thomson had discovered and what we now call electrons.

In 1906 J J Thomson, later Sir Joseph, was awarded the Nobel Prize for physics. ■

Reference

J J Thomson, Cathode Rays, *Philosophical Magazine*, 5th series, Vol. 44, No. 26, October 1897, pp. 293-316.

Centenary Exhibition

The Science Museum, London, is running a special exhibition commemorating the discovery of the electron. Called Life, the Universe and the Electron, it continues till 5 April 1998 and includes historical apparatus, a film, interactive displays, push-button demonstrations and modern electronic devices.

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Dual-port sram. IDT announces the first synchronous dual-port sram family, which allow pipelined operation to give better performance in data transfer. They are also available in flow-through versions for fast random-access to individual memory locations and rapid transition from write to read operation. Synchronous operation means that processors do not need to hold control signals for long periods; during a synchronous write function, the processor needs only to apply address and data to the chip for 5ns as opposed to the 35ns of asynchronous devices. The dual-port structure allows the entire contents to be read from both ports simultaneously with no delay from multiplexing. Devices now available are 512Kb *IDT709279/70927/709089/70908*, 256Kb *IDT709269/709079* in X16 and X8 form. Integrated Device Technology. Tel., 01372 363734; fax, 01372 378851.

Microprocessors and controllers

New PICs. *PIC16C76/77* 6.25mips eprom-based microcontrollers by Microchip have 8K by 14 words of program memory, 376byte of data ram and a low-power 5-channel (76) and 8-channel (77) 8-bit a-to-d converter with sample-and-hold, $\pm$1 lsb accuracy and 16ms acquisition time. Cycle time is 200ns and there are 35 single-cycle instructions. Timing includes a real-time clock, one 8-bit and two 16-bit counter/timers. I/O functions occupy 33 of the pins and include an 8-bit asynchronous slave port, a synchronous serial port for SPI and I^2C, a 6.25million-bit/s usart with baud-rate generator, two pwm outputs to 8-bit resolution and a 16-bit capture and compare facility. Arizona Microchip Technology Ltd. Tel., 01628 851077; fax, 01628 850259.

Low-power controller. Toshiba's *TMP87PH48U* works on supplies down to 2.7V and has, optionally, a 32kHz secondary oscillator for low-power use. Power consumption is 1.75mA at 3V and 4.2MHz, 20μA with the 32kHz oscillator, 10μA in sleep mode and 0.5μA in stop mode. Features include 16kByte of otp memory, 512byte of ram, uart and I^2C serial ports, 16-channel, 10-bit and four-channel, 12-bit a-to-d converters, two 16-bit, two 8-bit, interval and watchdog timers. Instruction cycle time is 0.5μs minimum and there are 412 instructions including multiply, divide and bit manipulation. Development software has ANSI C and C-like compilers and there is a real-time emulator. Toshiba Electronics UK Ltd. Tel., 01276 694600; fax, 01276 694800.

Motors and drivers

Stepper/driver/gearbox. *VEXTA 2-Phase CSK.SH* series stepping motor-controller-gearbox combination by Oriental Motor provides finer step angles than is given by independently operated motors and are therefore meant for use in applications where higher step resolution in needed. An additional feature is that the vibration found in low-speed operation of independent motors is reduced by the gearing. Six models are available in single or double shaft form and six gear ratios. Oriental Motor UK Ltd. Tel., 01252 519809; fax, 01252 547086.

Stepper driver/controller. Latest in Ericsson's family of stepper motor driver/controllers is the *PBL 3776*, for high-current motors. Being a two-channel chopper device, it will control a two-phase stepper with cmos or bipolar, external output transistors. Preferred current handling range is 3-10A/phase. Power needed is 5V for logic drive and 10-45V for output; channels are closely matched for positional accuracy. Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095.

Driver/a-to-d converter. Full-bridge pulse-width-modulated motor driver ic from Allegro, the *A3955*, incorporates a 3-bit, 8-level, non-linear digital-to-analogue converter and is meant for eight-step digital microstepping of two-phase bipolar stepper motors. It will put out up to $\pm$1.5A continuously at up to 50V, internal fixed off-time pwm current control regulating load current to the required level by means of an external *RC* network, a sensing resistor setting current limit. A mixture of a slow recirculating current decay and fast regenerative mode allows low ripple and heating while keeping good regulation. There is full protection and no requirement for power-up sequencing. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.

Microwave components

Isolators and circulators. Ocean Microwave Corp. includes in its range of passive microwave components a number of ferrite isolators and circulators for 900MHz and 1800MHz cellular PCN networks. There are flanged and thin pill-type drop-in units for strip-line circuitry and low-power surface-mounted types for hand-held equipment, also double-junction flanged and smd versions providing high isolation and four-port circulator features. Smd types handle 1W in 2.4-2.5GHz models with 20-25dB isolation and there is a 40dB isolation type. Maximum power rating in the range is 160W. Anglia Microwaves Ltd. Tel., 01277 630000; fax, 01277 631111.

Optical devices

Optocoupler. Sharp has a new, low-current optocoupler for telecoms. Switching current is 50μA, a level needed for connection detection in ISDN systems. Current transfer ratio is 60% and dark current 3μA, suitable to connect to a cmos microcontroller in a terminal or telephone. Hero Electronics Ltd. Tel., 01525 405015; fax, 01525 402383.

Transformers. Stontronics offers open-form toroidal and chassis-mounted transformers giving low voltages in the 6VA to 2.5kVA range with independent, series, parallel or centre-tapped outputs. Both types have two independent windings for 120V and 240V working and two output windings. Use of a double-section bobbin allows compliance with the relevant standards of insulation. Stontronics Ltd. Tel., 01734 311199; fax, 01734 311145.

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PASSIVE

Passive components

Miniature metal-film resistors. Dubilier has the *DMR 16T* range of metal-film resistors rated at 0.4W in values of 10Ω to 1MΩ to a tolerance of ±1%. They measure 1.9mm by 3.7mm and withstand soldering heat and solvents, temperature cycling, humidity and have high terminal strength. Maximum working voltage is 200V. Dubilier Ltd. Tel., 01371 875758; fax, 01371 875075.

Hand-held multimeters. Kenwood's first venture into these instruments is the new four-member family of *DL* multimeters, *DL-91/92/94/97*. Each has digital and bar-graph lcd indication and can be used in manual or automatic modes of operation; three of them have the hold facility, which retains the reading when the probe is removed. The 0.5% *DL-91* handles voltage and current dc/ac, resistance, diode and conductance check. *DL-92* is similar, but copes with frequency and maths functions to 0.3%, the *DL-94* offering the same functions but to 0.1%. *DL-97* does all that plus temperature and capacitance and has a dual display to show two functions at the same time, all to within 0.06%. It also has an RS232C interface. Kenwood UK Ltd. Tel., 01923 218794; fax, 01923 212905.

Chokes. *LHL10/13/16* are radial-leaded inductors by Taiyo Yuden designed for use in power supplies and are in values from 3.3μH to 150μH, rated at 0.028-4.5A, depending on the type. Tolerances can be ±5%, ±10% and ±20% and minimum *Q* between 30 and 140. The chokes are in plastic encapsulation and are available taped or in bulk packs with 5mm cropped leads. Taiyo Yuden UK. Tel., 01494 464642; fax, 01494 474743.

Thermistors and varistors. Thomson-CSF's ISO-approved thermistors and varistors are now available in the UK from Kestronics. Ntc thermistors with resistances at 25°C of 1Ω to 1MΩ work at temperatures between –55°C and 150°C and are in the form of chips, leadless and leaded discs, leaded chips and with metal casings, a special package allowing operation to 200°C. Ptc types work up to 130°C at maximum operating voltages from 30V to 265V with nominal 25°C resistances of 1.1Ω to 220Ω. The varistors have ratings of 14Vrms/22Vdc to 625Vrms/1000Vdc. Kestronics Ltd. Tel., 01727 812222; fax, 01727 811920.

Ultra-low-*Z* electrolytics. With what Semicom claims to have the lowest impedance on the market, the *MV-DX* series of electrolytics also have a high *C/V* ratio and come in surface-mounting or through-hole versions. As examples, a 10V, 100μF type has an impedance of 0.27Ω at 100kHz and 20°C with ripple current of 230mA at 100kHz and 105°C, while a 35V, 2700μF component exhibits 0.015Ω and 3200mA for the same conditions. Capacitors in the series are suitable for high-frequency work such as switched-mode power supplies and filters. Semicom UK Ltd. Tel., 01279 422224; fax, 01279 433339.

Audio products

Audio interface. I2S, IEC958 and proprietary Japanese audio data formats are brought together by the Philips *TDA1373H* ic, which operates in the digital domain, no analogue stages being used. Total harmonic distortion for a 20kHz bandwidth are –113dB for 20-bit data and –95dB for 16 bits. The ic is also a sample-rate converter for 44.1kHz cd and recordable cd data, 48kHz for professional audio and digital cassettes and 32kHz for digital stereo, the rate being set by external crystals or by the internal pll that tracks the sample rate of the input, mainly for decoding IEC958 data for cd recording. Arrow-Jermyn. Tel., 01234 270027; fax, 01234 214674/791501.

Cameras

Imaging chip. Vision's *5402* range of monochrome, medium-resolution cmos sensors that are said to provide significantly enhanced quality and performance over ccd types. They give composite video output with a minimum of external circuitry and the range includes packaged sensors, camera modules with or without lenses and miniature cameras in plastic or metal housings. VLSI Vision Ltd. Tel., 0131-539 7111; fax, 0131-539 7141.

Digital camera chipset. Sony's *SS-1* three-chip set performs processing and driving functions for a lightweight digital camera for security, multimedia or manufacturing. It is suitable for use with high or low resolution ccds and offers a "window" feature which allows a pc's screen area to be defined, background brightness or peripheral lighting thereby leaving the selected image unaffected. The set comprises the *CXD2163* processor, the *CXA2006* camera head amplifier and the *CXD2480* driver, which has all timing and control functions for PAL and NTSC standards. Pronto Electronic Systems Ltd. Tel., 0181-554 5700; fax, 0181 554 6222.

Connectors and cabling

Filtered BNCs. GTK has a range of filtered BNC sockets with a number of body styles including low-profile types standing 5.73mm above the board, a number of top or side entry models being available for through-hole or surface mounting. Capacitance is 9400pF and the sockets come in 50Ω and 75Ω versions. GTK (UK) Ltd. Tel., 01344 304123; fax, 01344 301414.

Low-profile connectors. Connectors from Stocko in the *Series 80000* range are compact units with board-to-top height of 8mm and a through-board height of 6mm, contacts being on a 2mm pitch in 2-16 ways. Crimp contacts of the indirect connectors are of pre-tin-plated phosphor-bronze and those of the soldered type of pre-tin-plated brass; bodies are in natural PC66 nylon or coloured UL94 V-O inflammability specification. Stocko (UK) Ltd. Tel., 01707 650882; fax, 01707 642735.

Coax. connector packs. Vitelec *Connectapaks* are designed to ease the stocking and use of coaxial connectors, each pack containing 100 crimped BNC plugs in a plastic box, which has separate compartments for the parts; to reduce wastage, there are 5% more pins and ferrules than needed. Electrospeed stocks the packs for a range of sizes including RG58, 59, 179 and 174, for 0-4GHz working at 50Ω, 0-1GHz at 75Ω. Electrospeed. Tel., 01703 644555; fax, 01703 610282.

Crystals

Telecoms crystal. The *SA-315H* crystal from Seiko Epson, intended for use in telecommunications, is a 1.6mm height, straight leaded type with a frequency stability of ±3ppm in the 0-50°C range of temperatures. Frequency range is 12-27MHz to a tolerance of ±10ppm at 25°C; over the –40°C to 85°C, stability is ±15ppm. Ageing is 1ppm and drive level 100μW to 2mW maximum. In a hermetically sealed can, it measures 8 by 3.1mm and withstands flow soldering. Advanced Crystal Technology. Tel., 01635 528520; fax, 01635 528443.

Hardware

Case frames. Vero's emc-screened *KM7-II* 3U case frames are interchangeable with standard versions, but provide more than 40dB attenuation at up to 200MHz. The frames hold 100mm high Eurocards in two depths and in 42, 60 and 84HP width, with ventilated or unventilated top covers. A conductive elastomer seal fitted into the side plates and selective masking of the top covers provides ground continuity, rfi fingers being fitted to electrically connect front panels and body. Variations in colour, width, ventilation and other features can be offered. Vero Electronics Ltd. Tel., 01703 266300; fax, 01703 265126.

Modular bgas. *3M Textool* ball-grid array sockets come in four different platform types. The s-ram platform is for the 14 by 22mm rectangular bga package, while Platforms I, II and III are for packages up to 23mm, 35mm and 50mm respectively. Modular design allows rapid and inexpensive modification to meet a range of size variations. BFI IBEXSA Electronics Ltd. Tel., 01622 882467; fax, 01622 882469.

Test and measurement

Tetra test set. Marconi Instruments announces the *2968 TETRA Radio Test Set*, which is a single instrument to test the future Terrestrial Trunked Radio (Tetra) digital trunking pmr standard. It is said to be the only instrument of its type to contain Tetra signal generator, modulation analyser and spectrum analyser in the one enclosure. It conforms to the European Telecommunications Standards Institute specifications and tests the rf, audio and dc parameters of mobiles, base stations and direct-mode terminals, also showing a graphical display of tdma burst profile against the Tetra limits. Marconi Instruments Ltd. Tel., 01438 742200; fax, 01438 727601.

Interference DF. *DDF190* is a new member of Rohde & Schwarz's digital direction finder family, this one having

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an upper frequency limit of 3GHz. It is intended for use in locating sources of interference and unauthorised transmissions. Two antennas cover the whole frequency range from 20MHz to 3GHz and the instrument is remotely controlled, being compatible with commercial receivers with unregulated if output of 10.7 or 21.4MHz. Operating as a correlative interferometer, the instrument copes with all modulation types and, if it is used in a vehicle, an electronic compass indicates north relative to the vehicle; vector df is also available. Rohde & Schwarz UK Ltd. Tel., 01252 811377; fax, 01252 811447.

Instrumentation interfaces. First products in National Instruments' DAQ series of interfaces are announced, combining stand-alone instrument quality with the facilities provided by pc-based instruments. *DAQScope*, *DAQMeter* and *DAQArb* are PCI, ISA and PCMCIA interfaces, compatible with Windows 3.1, '95 and NT, to provide oscilloscope, digital multimeter and arbitrary waveform generation by means of *VirtualBench-Scope/DMM/ARB* virtual instruments, working with *LabView* and *LabWindows/CVI*. Instrument drivers are available for use with C, C++ and Visual Basic. National Instruments UK. Tel., 01635 523545; fax, 01635 523154.

Ferrites for rfi suppression. *Zemrex SF* suppression ferrites are now made in a greater number of sizes and characteristics to meet the demand for closer tolerances in rfi suppression. Cables to be protected feed through the ferrite apertures with no need for earthing and there are split-shell models for retrofitting. The *SF-F* series fits round ribbon cables of 18-41 ways with optional nylon snap-on cases to anchor the ferrite to the cable. A number of variants fit cables of different sizes and shapes and the *SF-C* type fits on the pins of D connectors. Warth International Ltd. Tel., 01342 315044; fax, 01342 312969.

Dso upgraded. Gould's *Classic 6000* digital storage oscilloscope has a member of new features as standard: a full-colour display; 50K memory; fast Fourier transform capability; and pulse-width triggering. £4,600 will get you a 200MHz bandwidth instrument with four low-noise inputs, live waveform processing, user-defined channel scaling and a compression technique, *TruTrace*, that gives a variable-intensity display to emulate that seen on an analogue instrument. Channel markers on the colour display are matched in colour to the relevant traces to ensure you read the right one. Gould Instrument Systems Ltd. Tel., 0181-500 1000; fax, 0181-501 0116.

Low-cost 20MHz oscilloscope. Feedback has a new dual-channel, 20MHz oscilloscope, the *OS5020P*, believed to provide the best value for money in the UK. There is an XY display mode available and the two inputs can be alternate or chopped, one of them being invertible. Sensitivity is 1mV/div with 10x magnification. Two probe kits are supplied as standard in the price of £289. Feedback Test and Measurement. Tel., 01892 653322; fax, 01892 663719.

Signal-conditioning oscilloscopes. New from Nicolet, the *Integra 10/20* oscilloscopes, which offer signal conditioning and analysis within the instruments. Both instruments are four-channel types combining true 12-bit resolution at a 1Msample/s acquisition rate with flexible triggering and accurate differential or single-ended amplifiers. In addition, the Integra 20 has optional colour display and thermal plotter, a floppy-disk drive, signal-conditioning amplifier options and a dsp board for real-time data analysis. Waveform display is flexible and includes roll mode, X/Y, compression, min/max and interpolation, while a split-screen mode allows simultaneous display of a whole record with a magnified section. There are rear inputs for transducers of various types. Nicolet Technologies Ltd. Tel., 01908 679903; fax, 01908 677331.

Waveform generator/editor. Thurlby Thandar's pc-based, 20MHz *GS2020 GeneratorScope*, developed with the Open University, provides three functions: a dual-channel oscilloscope; a digital waveform generator; and a four-output power supply. Control software runs under Windows 3.1 or 95, each section having its own window in the normal Windows manner so that it can be used in other applications such as word processors or spreadsheets. *WaveCad* software, supplied with the instrument, is used to create arbitrary waveforms using drawing tools or equations or both. Waveforms captured by the oscilloscope section can be modified and combined with others using cut, paste, copy and a number of other editing facilities. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Clamp meter. Metrix claims its *MX120* clamp-on meter is the only one to measure current in a single or multicore cable without separating the conductors. You simply clamp the meter round the sheath of flat, round, two-core, three-core or twin-and-earth cable up to 25mm diameter to measure currents up to 40A or up to 200A in single cores. Cable type selection is by rotary switch and readout by lcd. Metrix Electronics plc. Tel., 01384 402731; fax, 01384 402732.

Literature

Telemetry. To provide an insight into the subject of radio telemetry, Wood & Douglas has produced a manual on the subject, *The Ideal Partner*, covering typical applications, terminology, antennas, radio links, operating modes, interference and overcoming problems in all these areas. It is free from W&D or you can download from the company's home page at http://www.woodanddouglas.co.uk. Wood and Douglas Ltd. Tel., 0118. 9811444; fax, 0118 9811567.

Maplin. Maplin's 1997 catalogue of electronic components is now available, listing and describing more than 18 000 products from international sources. It comes free to anyone holding a Maplin account. Maplin Electronics. Tel., 01702 554155; fax, 01702 554001.

Power supplies. XP's tenth catalogue is available; by now, probably, on interactive cd-rom too. The paper one has more than 280 pages to cover a very wide range of power products including dc-to-dc converters and switching power supplies and an eight-page supplement giving details of new and updated products. Also included are details of the 'configuration facility', in which XP engineers labour to produce specified power supplies in a couple of days. Both paper and CD are free. XP plc. Tel., 01734 845515; fax, 01734 843423.

CD-Rom relay catalogue. *Designers' Databank* on a free cd-rom (minimum needed Windows 3.1, 486, 8Mb, double-speed drive) lists and describes over 2000 of Siemens relays. A search engine enables a match to requirements to be easily found among the Schrack, OEG and Potter & Brumfield brands. It puts exportable data and drawings on screen and provides a help facility; it also generates an order form. Siemens plc. Tel., 01344 396685; fax, 01344 396665.

Networking cabinets. A range of enclosures for intended for use in networking applications is described by Vero in a new catalogue, which contains data on *IMRAK 400* wall-mounted cabinets; *IMRAK 600* types; *IMRAK 1400* for lan, wan and telecoms; and the *VERAK IP* and *EMC* enclosures for industrial networks and general electronics use. There is a comprehensive range of accessories, including cable management modules for copper and

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fibre cables. Vero Electronics Ltd. Tel., 01703 266300; fax, 01703 265126.

Electrolytics. *EVOX Rifa* electrolytic capacitors are the subject of a new catalogue from Campbell Collins. It has 76 pages and contains design information including operation and application notes and a glossary of terms and definitions. There is a rapid selection chart. Campbell Collins Ltd. Tel., 01438 369466; fax, 01438 316465.

Materials

Thermal interface pads. Chomerics offers *V-Therm* thermally conducting, electrically isolated pads, made from boron-nitride-filled silicone elastomer, to form the interface between components and heatsinks/heat spreaders. The material conforms to surface irregularities to fill air gaps and is also available with an adhesive backing, thicknesses available being 0.51mm, 1.02mm and 1.52mm, others being made to order. Thermal conductance is more than 5W/m-K and dielectric strength 6000Vac/mm, with volume resistivity $10^{14}\Omega$-cm. Parker Hannifin plc, Chomerics Division. Tel., 01628 486030; fax, 01628 476303.

Protection devices

Sub-miniature fuses. Fuses measuring 3.18mm in diameter and 7.11mm long in the *MQ* and *MS* ranges by Belrose are moulded, fast-acting, board-mounted types rated at 125V ac or dc and 125mA-15A (MQ), 370mA-7A (MS). Blow time of the MQ range is 5s maximum at 200% load, while that of the MS types is 30s at 200%. All ratings are recognised under the Components Program of UL and CSA certified. Europa Components & Equipment plc. Tel., 0181-953 2379; fax, 0181-207 6646.

"Green" cleaners. *Cirozane*-based precision cleaners by Chemtronics are said to be the first high-performance replacements for cfcs, offering all their features except that of destroying the ozone layer. They can be used without the need to make allowances on plastics and metals and on equipment under power that must be kept running. Three types are available: a cleaner/degreaser; a flux remover; and a contact cleaner. Rocol Ltd. Tel., 0113 2322600; fax, 0113 2322740.

Power supplies

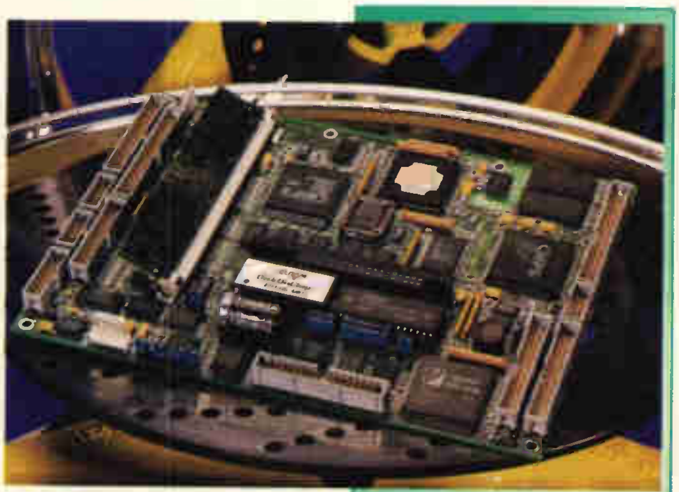
Linear psus. With low noise in mind, Computer Products has produced two families of linear power supplies, the *PM300CE* and *PM500CE* series, which meet EN55022/11 and FCC Part 15 Level B standards for radiated and conducted noise. They are available in single, dual and triple output form and all take standard 230Vac input, power ratings being 2.5-15W for the 300 series and 1-10.5W in the 500 versions. Outputs are 5V, ±12V and ±15V in various combinations and the units are all in non-conductive plastic cases with threaded inserts for chassis or panel mounting. Connections are by screw terminals. The 500 series provides lower noise: line regulation of ±0.02%, load regulation of ±0.04% and ripple/noise of 1mV. Computer Products, Power Conversion Ltd. Tel., 00353 24 93130; fax, 00353 24 03257.

Dc-to-dc converters. *WP06R* converters by Power Convertibles provide input-voltage ranges of 9-18V, 18-36V and 34-75V and a standard operating temperature range of –25°C to 71°C with no derating. Models available have 5-6W outputs of ±5V, ±12V and ±15V in single, dual and triple versions. Units are metal-cased for reliability and emi shielding and the 32 by 20 by 10mm size is obtained by a 200kHz switching frequency and flyback operation. Power Convertibles Ireland Ltd. Tel., 00353 61 474133; fax, 00353 61 474141.

Switches and relays

Rocker switches. Cherry Hirose has a new family of compact rocker switches, the *Type RS1*, which is in dpst and spst forms and can handle inrush currents up to 160A at 250V ac. Normal current rating is 16A and mechanical life is 20 000 cycles at operating temperatures between –25°C and 85°C. The range includes remote switching types. Cherry Electrical Products Ltd. Tel., 01582 763100; fax, 01582 768883.

Solid-state relays. The Douglas Randall and Crydom Europe range of scr-output solid-state relays (all now Crydom) now includes lower current types with load currents of 0.1A to 1.5Arms, control voltages being 3.5-10Vdc at 10-50mA. Operation is by voltage (*SDV* range) or current (*SDI* types), and there is the option of random or zero-crossing turn-on. Crydom. Tel., 01202 897964; fax, 01202 891918.



Transducers and sensors

Current sensors. Zetex has a range of current sensors that employ the magnetoresistive effect of thin-film Permalloy, the *ZMC05/10/20*, which detect and measure alternating and direct currents of 5A, 10A and 20A respectively. These devices are in the form of an ic, the magnetic field being generated by an internal conductor, a thin-film Wheatstone bridge on silicon forming the sensor, which is provided with an internal magnet to counteract the effect of external magnetic disturbances. The isolated conductor resistance is 0.7mΩ and the 10A and 20A types handle an overload of 300A for 10ms. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5467.

Pressure sensors. If earth moving is your forte, Control Transducers have the very pressure sensor. Its *Model SA* is designed for general industrial use and, apparently, is very popular with makers of earth-moving equipment. Pressure range is up to 490bar or 7100lb/in^2 and the transducer is fitted with terminations and mounting to suit the application. Another of the company's devices is meant to handle hydrochloric acid and other unfriendly fluids that attack stainless steel, from which the other types are made; it has a Teflon coating, which leaves accuracy unaffected. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

Embedded pc. Increased levels of integration allow the *TPC Series* of processor cards by Datasound Laboratories to satisfy the needs of users who require low module counts in applications such as point-of-sale terminals and vehicles. The cards have graphics, keypad, a-to-d converters and digital input/output and, although this has reduced the need for external expansion, there is a range of i/o modules accessible through a 16-bit ISA AT-compatible PC/104bus and PCMCIA v.2.1 controller supporting all types of card. Processor options are 486DX66, 486DX100 and 586 133, supported by a CS4041 chipset. The card drives both crt and flat-panel displays and there is 1Mbyte of video memory. Memory is up to 32Mbyte of fast-access dram and 2Mbyte of eprom or flash eprom. Datasound Laboratories Ltd. Tel., 01462 675530; fax, 01462 482461.

COMPUTER

Four computers in a box. *System 2000* by BVM is a self-contained IP65 sealed, rfi-proof, stainless steel unit containing up to four RP2000 16MHz 68302 single-board computers and a mains or 24-28V dc power supply. Each RP200 can be fitted with two IndustryPack standard i/o mezzanine boards to provide the choice of digital, opto-isolated, d-to-a and a-to-d i/o and is meant to be mounted directly on plant or machinery to give distributed intelligence for process control and data capture. Software support for the RP2000 sbc includes FieldLink, BVM's 2Mb/s autorouting FieldBus networking software. BVM Ltd. Tel., 01489 780144; fax, 01489 783589.

Dual Pentium single-board pc. Newly announced by Tri-Map, the *Peak-6020* is claimed to be the first industrial dual Pentium-Pro single-board computer and has been developed for use with multi-processor operating systems such as Windows NT, Unix, Solaris and OS/2, its 32-bit performance suiting it to data, signal and image processing. It is fitted with four 72-pin simm sockets to take from 8Mb to 256Mb of FPM, EDO or BEDO dram and full error checking and correction. The Intel 82440FX PCI chipset is on board to support PCI, Bus-PCI v2.1. There are parallel, serial and Intel usb ports with keyboard connectors and a programmable watchdog timer for fail-safe operation. Tri-Map International Ltd. Tel., 01705 424800; fax, 01705 424801.

Long-life embedded pc. With guaranteed hardware support for the next nine years, the Contec *Box-PC* is designed as an industrial embedded pc for working in high levels of vibration, for networking and with ease of operation and expansion in mind. The Box-PC includes opto-isolated digital i/o, two PCMCIA slots, RS-232 and RS232/485 ports, a bay to take a 2.5in or 1.8in IDE hard disk drive or flash disk, interfaces for a keyboard, PS/2 mouse and printer, expansion facilities for more i/o and VGA/lcd interfaces. Gothic Crellon Ltd. Tel., 01734 7776161; fax, 01734 776095.

Development and evaluation

PIC emulator. RF Solutions has added to its family of Microchip PIC emulators a low-cost daughterboard module for development of embedded applications needing lcd output. It is usable with the company's *ICEPIC* and *ICEPIC2* emulators to enable real-time emulation of applications using the 16C923/4 PIC processors, both of which have lcds. The combination of *ICEPIC* and daughterboard provides source-level debugging in assembler and C at up to 10MHz. Plugging directly into the *ICEPIC* main pod, the daughterboard needs no extra hardware and the combination runs under Windows, the connection to a pc being by RS-232. RF Solutions Ltd. Tel., 01273 488880; fax, 01273 480661.

Computer peripherals

Touchpad keyboard. Cherry has a new keyboard with a touchpad to replace a mouse. *Model 11900* is intended for use when a mouse might be a bit awkward, as in rack-mounted computers, and when space is restricted, since the keyboard is only 16.5in wide and 7.5in deep. It uses a programmable, 105-key Windows '95 layout with the touchpad on the right below the numeric keys. Cherry Electrical Products Ltd. Tel., 01582 763100; fax, 01582 768883.

Software

Systems design software. i-Logix announces version 1.1 of its *Statemate MAGNUM* tool for automatic systems-level design, software design and implementation. It integrates with other tools in the embedded systems design process, including Doors, RTM, Matrix and Saber and runs executable virtual prototypes, supporting Windows '95 and NT 4.0. Engineers graphically specify, model and analyse behaviour, functions and operator interface, the software automatically generating source code in C, Ada, VHDL and Verilog from a graphic model. i-Logix UK Ltd. Tel., 01249 446448; fax, 01249 447373. ■

LETTERS

Letters to "Electronics World" Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

EMC enforcement issues

I noted with interest the news item 'EMC law proving difficult to enforce' on page 444 of the June issue. The problems reported have been apparent to some of us for nearly twenty years – from the time when the development of the 89/336 EMC Directive was first mooted. Until last year. I seriously considered that the curse of Cassandra had been applied to my – and others' – efforts to get the true situation recognised.

It was only last year that people generally began to listen. Note that I do not by any means deny that emc is important; it is the half-baked implementation that is causing all the trouble.

Many emc measurements – both emission and immunity – are extremely difficult and costly to perform. This is true even now, when the measurement processes have been placed under computer control. While this has greatly decreased the time required to perform the multiplicity – up to several hundred – of measurements required for one 'test', the capital cost of the automatic equipment is beyond the resources of all but the largest manufacturers and those test houses which can ensure full-time use of the equipment.

The problems have been exacerbated by the accreditation authorities, who require sources of error and uncertainty in measurement to be controlled and allowed for. EMC measurements are, however, notably imprecise and unrepeatable, because in many cases the results depend critically on the exact physical disposition of the equipment – both that under test and the test fixtures themselves – and cables. If this problem were openly faced, very few measurements would be acceptable for accreditation.

Under these circumstances, the statement attributed to Mr Holland indicates a degree of confidence that is shared by few people who actually know anything about emc measurements.

It is also true that, where finished goods are made up from sub-assemblies that are themselves subject to conformity with the Directive, and the finished goods are found to be in violation of the relevant requirements, then all the sub-assemblies have to be investigated, although this may eventually prove inconclusive: in emc the whole may bear very little resemblance to the sum of its parts.

This complication is a direct consequence of the 'interpretations' placed by Directorate-General III of the European Commission on the vastly over-simplified language of the Directive itself. Presumably the consequences of this interpretation, in requiring a long chain of investigation, were either not realised by DG III or were glossed over. We can form an estimate of the skills of DG III in this highly technical field by noting that there have been several issues of 'guidance notes' since 1991, which have contained nothing less than hilarious errors, or which would be hilarious if the subject and context were not so serious.

The latest issue, promised for February, is still awaited. A draft was posted on Usenet and this showed that earlier standards of accuracy and lucidity had been maintained.

The situation is gradually being improved, as more and more people become convinced of the extent and nature of the problems, but it will be ten years or so before reasonable practicality will be achieved.

Meanwhile there will be unjustified prosecutions, false convictions, false acquittals and much money transferred from the electronics industry to the legal profession. Let us hope that the none of the doomsday scenarios used to 'sell' the EMC Directive to the politicians, of mad robots, runaway driverless trains etc, actually occur.

John Woodgate
Rayleigh
Essex

EMC ethics

In the recent editorial I wrote for *Electronics World* about the new EMC legislation, I predicted that companies would use the laws to shop their competitors.

I see from *New Electronics* that this has in fact happened (page 9 of the issue dated Tuesday 25 February). In this report, a company with its own emc testing facilities 'shopped' one without such facilities. There was no altruistic motive of protecting the ether from pollution. In my view it was done solely in order to defeat competition.

We expect the laws on, for example, car tax or tv licence evasion to be enforced by the officers of the relevant authorities. Most would find it distasteful if the State relied solely on the enemies and neighbours of these law-breakers to inform on them, in order to uphold the law – even if this did justifiably happen from time to time.

In the case of the CE marking laws, it is freely admitted that in this country they are 'complaint-led' – a euphemism for relying on informers. It is clear that the laws are being implemented in such a way as to give such complainants an economic incentive to become informers. This is a thoroughly contemptible state of affairs.

We have already become a nation of unpaid tax-collectors on behalf of the State, thanks to VAT, but do we want to become a nation of snitches as well?

Rod Cooper
Sutton Coldfield

Q & A

Outputs in phase quadrature?

In the June issue, Alan Scrimgeour asked:

Q I am looking for a circuit to phase shift by 90° the components of a signal with frequencies in the range 10Hz to about 350Hz. Although simple integration or differentiation can achieve this. they do so at the expense of a frequency dependent change in the signal amplitude which I cannot use.

A Commencing in *EW* April 1996 and ending in the following issue, I wrote a two-part article describing exactly the filter that Alan Scrimgeour is looking for. The article was entitled 'Designing an SSB Outphaser', parts 1 and 2 (pp 306-310, and 392-4 respectively).

David Gibson
Somewhere
Anywhere

Better pcb photography please

Apologies to readers for the poor reproduction of two of the photographs in the article 'Prototyping Circuit Boards' in the previous issue. This was due to a production malfunction on two of the photographs of PnP circuit board transfer film – not in the quality of the material presented.

In the left-hand column on page

Obituary – Howard Hutchings

Howard J Hutchings BA (Hons), MPhil died on Friday 23 May 1997 aged 51.

Howard started his teaching career at Hull College in 1974 where he taught across a range of electronics subjects until 1985, he then joined what is now the University of Lincolnshire & Humberside as a senior lecturer in the School of Engineering and Information Technology. There he accumulated considerable expertise in real-time signal processing.

A prolific publisher of articles, Howard selected a particular series to compile his book 'Interfacing with C' in which his practical, no-nonsense approach to engineering is evident. In this book he reserves a special acknowledgement for his wife, Margaret, "a constant source of encouragement."

Howard Hutchings was highly respected by his colleagues at the University and was acknowledged not only as a teacher of the highest calibre but also as the school expert in digital signal processing.

Howard leaves a wife, Margaret, and three children.

He will be sadly missed.

Peter J Hale and Dr Martin Dickinson

562, both of the bottom two photographs were very badly reproduced. The photographs should have indicated that the results from PnP are comparable to other methods of prototyping. Besides being a much more rapid process, PnP works out about 30% cheaper than the ultraviolet method – a fact not mentioned in the article.
Rod Cooper
Sutton Coldfield

More scope needed

Around 1970, Tektronix were making large, expensive and high-performance oscilloscopes, involving special tubes and other components. They were invaluable on the GEC-Marconi-Elliott computer production floor. One reason was that a 1µs spike showed up on a 10s scan because of the great brilliance available.

I believe few, if any, of the workshop oscilloscopes now sold can achieve this, yet such spikes could be shown on a computerised raster oscilloscope using computer enhancement to steady and widen the image of the spike.
Bernard Jones
London

Resistance to criticism

In the June issue, David Markie's letter began:
I was reading through the article 'Resistors in C' in the April edition, and was getting on fine until I came across "...it is advisable to use at least a 486DX... because of 57600 calculations... a lesser pc will take several minutes". Continuing the extrapolation, an old eight bit 2 MHz processor running interpreted Basic would have taken around an hour - it didn't. So I thought something's wrong here and I started to look at the code listing.

The author replies:
I am glad that Mr Markie got on fine until he came across "It is advisable..." because by that time he had almost finished the article. I am sorry that he did not re-read it prior to putting pen to paper, for most of his letter states exactly what was in the article.

He missed the reason for displaying the higher ratio pairs, and the simplicity of the published listing. I am well aware of the short-comings of nested loops, as stated in the article. However, I am sorry that Mr Markie has made the assumption that this method has also been used in the more complex version of the special offer, and therefore advised readers to resist this offer.

The listing for this version is both long and complex, and not for publication. But the original program, written quickly purely to get a calculator up and running, I feel has merit in it's size and simplicity.

Not being aware of another program that calculates parallel resistors, and having had the use of this one for some years, my intention was to make it available to others, bring to light any similar programs that may exist, provoke comments (of the constructive kind), and to gauge the need and interest in such a programme prior to completing a windows version.

As a final comment, it is easy for a person to belittle another's ideas but it is something else to have the idea and do something about it. What was it they said about Marconi?
John Lavender
Yate
Avon

Q & A

Power amp instability

Q In essence, Charles Coutas' question was, why do some audio power amplifiers go unstable when you connect a piece of coaxial cable to their input?

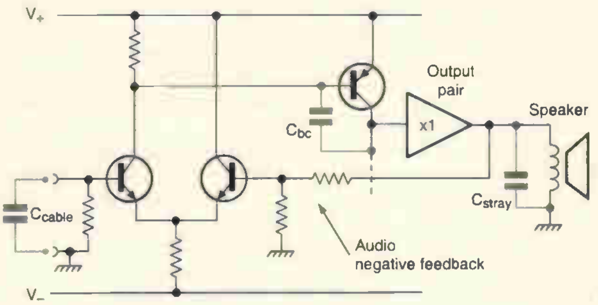
A Nowadays, audio power amplifier circuits are based on the same long-tailed pair differential input circuit, and have the same input to output zero dc offset performance, as the standard operational amplifier circuit. So it should be possible to possible to explain any instability using the same dry mathematical techniques.

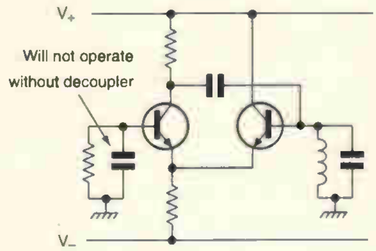
It is possible however to redraw this circuit to show that it contains a 'parasitic' oscillator circuit – a kind of emitter coupled oscillator popular with rf designers because untapped coils can be used – of which only one element is normally missing, the decoupling capacitor at the base of the first transistor.

Since this happens to be the transistor at the input of the audio amplifier circuit, it becomes more obvious why stray capacitance here, such as screened cable self capacitance tends to cause hf instability.

One way of reducing this problem is to reduce negative feedback. A lower-gain, high-linearity video amplifier transistor used as the driver may well help. It may also help to use an op-amp to drive the output pair. In a commercial op-amp, all the problem elements have been carefully balanced out for you.

However since input impedance with high levels of negative feedback is very high at audio frequencies, and low at the instability frequency, the simple expedient of series input resistor should not be dismissed as invalid.
A Ziemacki
Rotherham
South Yorkshire





On the left is an rf oscillator configuration favoured by designers because it does not rely on tapped coils. To the right of it is a simplified power amplifier layout, in which you can see similarities with the oscillator. Capacitance C_{cable} is self explanatory, C_{bc} is the transistor's base-collector capacitance, of about 100pF and C_{stray} comes from the speaker leads, crossover capacitors, etc.

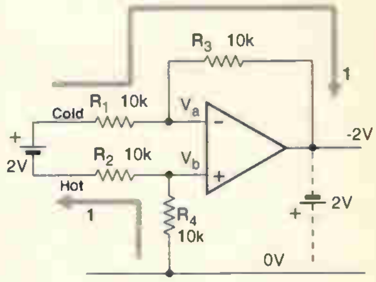
Difference of opinion

I am afraid I must disagree with Douglas Self's analysis of the standard single-op-amp differential amplifier. In particular, his results suggesting that a floating source sees equal impedances of 10kΩ each input. I would like to present a simple analysis to disprove this.

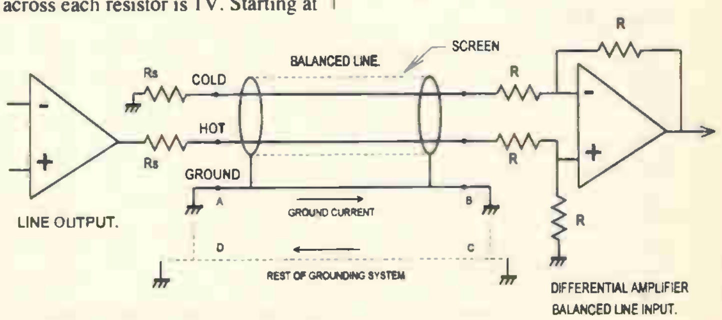
Consider the circuit as given in his Fig. 3. A good floating source, for analysis, can be a 2V battery across the hot and cold terminals. This can represent an ac signal at a particular instance in time. The differential gain of the op-amp is unity.

Both inputs of the op-amp are forced to be at the same potential by feedback; thus V_a=V_b. Since the positive side of the battery is connected to the inverting side of the op-amp, the op-amp output will be at –2V. The output of the op-amp can be considered to be a 'virtual battery' between the output pin and the 0V rail. Both op-amp inputs are high impedance, thus no current flows into or out from them. The current flow can be analysed now.

The two 'batteries' are in series, thus generating 4V in total. The four resistors R_1 to R_4 are also connected in series, so the voltage dropped across each resistor is 1V. Starting at R_4, the voltage at the non-inverting input, V_b, must be –1V since one end of the resistor is connecting to 0V and 1V is dropped across it. The hot



Doug Self's original Fig. 3, below, with the circuit redrawn to help understanding, below.



input is therefore at –2V, due to the 1V dropped across R_2. The input battery voltage is 2V, so the cold input is at 0V.

Resistor R_1 drop 1V, so the inverting input voltage, V_a, is at –1V. This complies with the requirement for having V_a=V_b; they are both at –1V . Finally 1V is dropped across resistor R_3, giving an output of –2V.

So far my analysis has agreed with Douglas Self's. Since the cold input is held at 0V and the hot input is at –2V, the voltage on one line has no signal on it while the other has the full signal voltage.

Since the current flow is 4V/40kΩ, or 0.1mA for a 2V signal applied at the input, the input impedance is 20kΩ. This also agrees with Douglas Self's analysis.

However, Douglas's contention that both hot and cold inputs have an impedance of 10kΩ is wrong (reference his Table 1). The cold input has an impedance of 0Ω and can be considered a virtual earth. This must be true since the cold input is always held at 0V, even though current is flowing through the node. The hot input has an impedance of 20kΩ since there are two 10kΩ resistors in series to the 0V rail and the non-inverting input impedance of the op-amp is very high – ideally infinite.

An input line connected to the standard one-op-amp differential amplifier will be unbalanced, as Douglas suggested. Since the current flow is balanced, to a floating source, there will be no magnetic coupled interference to other cables. The important point is that the unbalanced electric signals on this line will generate fields that may couple into other lines, i.e. cause interference to them.

At high frequencies there will be a signal path from the floating source to ground through stray capacitance. The floating source model is therefore only good at audio; it is not a good model at hf or above. For the *TL072* op-amp used in the example this is not a problem, but high-speed op-amps can be used up to vhf.

Steve Winder
Ipswich
Suffolk

Disk preamp design thoughts

I was pleased to see some discussion regarding preamplifier design. I noted in particular the comments made, almost as a throw away line by Douglas Self about the performance of the *LT1028*. He said that the distortion products approach 0.01% without also noting that this figure is quoted for operation at a closed-loop gain of 1000, with an output voltage swing of ±10V.

I imagine that this would be an extremely undesirable set of operating conditions for the first stage of an RIAA preamplifier. I have perused the data sheets for the *AD797* and have attached distortion plots for this rather good op-amp, which show distortion products sticking their noses above the fence at 3V rms output.

When either of these op-amps are operated under the conditions likely to be encountered in the first stage of an RIAA preamp, distortion is down in the noise.

Some time ago I built an RIAA preamplifier using *LT1028*s for the first stage and *NE5534AN* in the second stage. This was to replace a rather venerable preamplifier and an external moving-coil head amplifier, which used dual paralleled *2SB737* and *2SD786* transistors in a symmetrical arrangement at the inputs with a few other bits and pieces. These devices were chosen for their very low base resistances.

The choice of the *LT1028* was simply due to some being to hand at the time. The *AD797* was not then available. I built this to suit my own requirements. As I have a number of phono cartridges and a fairly large collection of LPs, I decided to make the preamp switchable for both moving magnet and moving coil inputs.

And I agree with Doug Self that noise performance can be bettered.

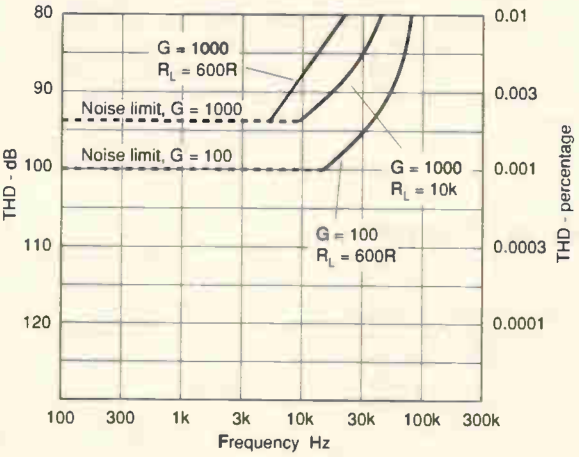
However the difference in cost at the end of the day is not all that much when the price of a bunch of 2SB737s and other odds and ends is taken into account and is certainly nothing like seven times as much overall,in the whole project, at least not in this country.

I have found that both the *LT1028* and *AD797* as well as most other really high performance op-amps are prone to burst into indignant protest unless considerable care is taken with circuit board layout, grounding and bypassing of supply rails, etc.

In the end, I found it easiest to use a double-sided board with a ground plane on the component side, thus allowing low impedance bypassing and easy separation of signal and bypass grounds.

The requirements for the ultimate in low-noise performance for a moving-magnet *versus* a moving-coil input are quite different. Moving-coil alternatives typically present a series coil resistance of 1.5 to 3Ω or so and a low and well damped inductive reactance of about 20-50μH, although I do possess a maverick, a Denon *Karat 23R*, which has a 28Ω coil resistance and an inductance of about 100μH as well as an output about 15dB above that of my Ortofon *MC20*.

The moving magnet cartridge – a Decca *London Gold* – is also rather different from most as it has a series coil resistance of about 1.8kΩ, rather than the usual 500 to 600Ω. Its inductive component is difficult to measure as the Q is very low, however the total reactance at 1000Hz is about 1.95kΩ. As with the Denon, this cartridge produces a relatively high output and it sounds good to me.



Curves for the AD797, showing distortion performance.

With a moving-coil preamplifier, the input noise current density is not particularly significant, the E_n of the amplifier and the Johnson noise generated by the cartridge and the resistors in the feedback path are dominant. Accordingly an amplifier with the lowest possible input noise voltage and the ability to drive a very low resistive load at the output is beneficial. Also, as the required output voltage swing of the first stage is not large, distortion is not likely to be a serious problem with a suitable amplifier.

On the other hand with moving-magnet cartridges input noise current density becomes significant. But the rise caused by increasing cartridge reactance with frequency is substantially suppressed by the RIAA de-emphasis response) and the input end of the amplifier tends to be surpassed by the Johnson noise generated by the cartridge.

In any event, good signal-to-noise ratios are relatively easier to achieve as the signal level is generally 30dB or more above that for the moving-coil case and in reality a properly designed moving-magnet preamp will have a signal to noise ratio which comfortably exceeds the noise levels introduced by such things as the medium, microphony and turntable noise.

A good fet amplifier such as the *AD745* with an input noise voltage of

Q & A

Edible batteries

Q In the June issue. David Heaton asked if anyone remembered a description of bio-galvanic batteries.

A Long ago, the Dutch electronics magazine Elektuur published a description of this bacteria power source. It was republished many years later as part of an insert, under a heading along the lines of 'xx years ago in Elektuur'.

I kept the relevant part of that insert on file, but alas there is no indication of publication dates.

Here you have a translation of the description:

A mixture of brown powdered chaff of rice, together with single-celled fungi like yeast will provide a few milliamps of current at some tenths of a volt.

The mixture is put into a small plastic box. Add a little water, and collect the energy using a copper strip as positive terminal. The negative should be aluminum.

The process will keep itself going for several years. Twelve cells in series are sufficient to feed a small radio receiver, a little lamp or a tiny motor...

The yeast bacteria are absolutely harmless, and since they are the most primitive form of life, this could hardly be called cruelty to animals.

At RCA, where the cell was developed, the search for cells with a higher power output continues.

Steven Bolt
Vogelenzang,
The Netherlands

2.9nV/√Hz could provide a marginal improvement, as could several paralleled fets with large arrays on the chip such as the 2SK147 or its complement 2SJ72 in conjunction with an op-amp. This is because the current noise density term then becomes insignificant and an E_n of 1.5nV/√Hz, or better, should be achievable.

Nothing can be done about Johnson noise apart from chilling all the resistive bits with liquid nitrogen. At ambient temperatures we are stuck with about 0.128nV/*R*nV/√Hz.

● In response to Nanno Herder's comments regarding 'feedforward error correction' in the April issue, page 320, I agree that at the end of the day it is a negative feedback scheme as gain must be introduced in the loop. This is especially the case for a source follower output stage.

However, the solution adopted by Robert Cordell has merit. Firstly the error correction amplifiers only have to deal with small voltage levels. They are implemented, very economically, with a couple of very fast small signal transistors and a few resistors. These amplifiers are not in the direct signal path but introduce a difference output into a resistive summing node at the input of the driver, somewhat like servo amplifiers.

The pole introduced by the feedback circuit does not introduce significant additional lag at the output stage as when properly compensated for stability, the gain falls away at very high frequencies. An output slew rate in excess of 300V/µs is achieved, with closed loop distortion better than 0.001% up to about 30kHz. In addition, square-wave response is good to an absurd 500kHz or so.

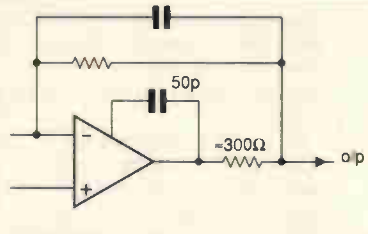
Measured improvement in transfer linearity the output stage a tone with the introduction of this error correction is better than an order of magnitude with open-loop output stage distortion of less than 0.1% at 20kHz.

The speed of the class A voltage amplifier is also exemplary and no special tricks are employed apart from the use of appropriately biased cascode stages throughout and a fairly nifty common mode rejection mechanism. Surely the proof of the pudding is in the eating. This design was first published in 1984 and I have seen few others since that come even close to this level of performance.

William de Bruyn
Victoria
Australia

AD797 instability

I noticed in a reply to a letter about his preamplifier design, Douglas Self said that he had not used the excellent *AD797* operational amplifier because of difficulties with hf stabilisation. My own experience is similar, in that I find the device prone to oscillations extending to vhf if the 50pF distortion neutralisation capacitor is used.



Putting a resistor in series with the output of the AD797, within the overall feedback loop, can solve the device's instability problems when used in this configuration.

I think that I know a satisfactory cure. That is to put a resistor in series with the output, but within the overall feedback loop, as shown in the schematic . The appropriate value of this resistor depends on the rest of the circuit, but 300Ω is a reasonable starting value.

I normally experiment a bit and use a value about 10% larger than the one at which oscillation sometimes sets in. I say 'sometimes' because this is a non-linear and bistable effect. If the resistance is to small, the circuit can work, but may be kicked into oscillation by, say, static discharge on the input, or heavy machinery arcing in the neighbourhood. This oscillation is hard to stop, once started.

David Thomas
Heidelberg
Germany

Its simple really

I have been reading the various 'debating' letters of Doug Self and Allen Wright over the past few months, and while I feel a certain percentage of the arguments put forward. are inconsequential, I am prepared to buy into two of the arguments; those of electrolytic capacitors and of the famous *5532* op-amp.

While it is true to say that common electrolytics are not the ideal audio device, it is also true to say that if carefully implemented, they are an acceptable and cost effective means of coupling audio stages, and for blocking DC. The unfortunate reality is that they do have a finite amount of leakage, and this can be a problem in high impedance circuits, causing dc offset and possibly distortion. The other difficulty is the relatively high impedance they can exhibit at high frequencies. These effects combined can cause non-linearity in some cases, and the designer must keep this in mind, but failure to believe that these devices can be successfully implemented in high quality audio designs is purely unreasonable.

The new style of low leakage and low impedance electrolytics – as used in switchmode applications – are entirely suitable for audio use.

I love this term "unpleasant sonic signature". I could use that to describe a number of compositions I have heard (though I would be using it out of context). I feel it is unfair, however, to describe the *NE5532* in this manner. It is true that many commercial designs use this chip extensively, and it remains a popular choice, but poor implementation has caused distress to those who treat op-amps like the universal fix-all.

I understand the *5532* is internally compensated for voltage gains over 3. For gains under this, external compensation is required (important for tone stages). The output is configured to drive loads down to 600Ω, but optimum distortion figures will not be obtained that way. The output stage is not symmetrical; few op-amps are. really. The addition of compensation capacitors drops the slew rate considerably. Given all that, my experience is that the 5532 is still a device with good audio performance. There are now superior devices. The *OP-275* springs to mind. but these are just not as readily available.

The solution to the problem of the effects of speaker cables would appear obvious to me. Keep them short.

Bruce Colledge
Ferry Hills
Australia

Loudspeakers exposed

In 'Loudspeakers exposed' in the June 1997 issue, John Watkinson makes a number of illfounded statements concerning almost every aspect of the audio reproducing chain, particularly speaker testing and design.

It seems that Mr Watkinson thinks that listening tests cannot be objective. Actually, listening tests can and should be objective, at least in the sense of being repeatable.

A possibility is random sequence A/B/X testing, which can achieve remarkable sensitivity and repeatability if properly conducted. Such a test can even be undertaken by a single individual, if properly equipped and honest. It can handle tests for insertion detectability as well as differential equipment comparisons.

The author seems to think that speakers must be omnidirectional and placed well into the room, in order to achieve "better reverberation," whatever this means. Now, while one must certainly take the characteristics of the room into account when designing and placing speakers, this is not enough; one must also decide what listening experience to try and simulate.

As an example, one possibility is to consider the listening room to be like a private box at the recording venue. One must then, among other things, try to minimise at least the long-delay reflections from behind the speaker, since that wall isn't supposed to be there. This then suggests speakers should definitely be directional and/or placed closer to one wall for this use, and bookshelf speakers might actually be a good choice.

It is certainly not a "hallmark of a good speaker" that you can listen to it indefinitely. Colourations deliberately designed into modern mass market equipment actually reduce listener fatigue by emphasising certain portions of the audio spectrum and suppressing others.

On the other hand, a very good speaker, without audible colourations or distortion, should induce roughly the same amount of fatigue as the original material would in the live listening situation. A horrible instrument perfectly reproduced is still a horrible instrument.

Yes, indeed, many loudspeaker designs use resonances to achieve extended low frequency response, among them the closed box and reflex enclosures. The closed box doesn't avoid resonance but it is less sensitive to manufacturing process errors, since it is the least complex of the two systems.

A bass-reflex design can be made to have very good transient response, if correctly matched to the room's influence. More often than not, this will probably mean you should tune the port resonance rather low, and maybe overdamp the box slightly. The distance from the speaker to the port is usually short enough, compared to the wavelengths in the frequency range where the port contributes, that the delay in itself will introduce no insurmountable philosophical problems.

A speaker enclosure cannot be considered to be just a pressure containment vessel. Its function is much more complex than that of, say, a welding gas container. A speaker box must contain fast pressure variations, and the pressure may vary over the box volume, not just over time, and using too simple a shape may introduce difficulties with damping structural resonances. The pressures contained are not very great, either. The largely rectangular box type is fairly well mastered these days, and wall vibrations are not really problematic with modern materials and properly applied bracing.

The scarcity of square aerosol cans is probably not related to any acoustic properties of their shape. While spherical loudspeakers may be well enough designed to offer good performance, the comparison to static pressure containers is of little relevance.

Finally, in a picture of what the author considers a good example of a bad speaker, and I generally agree, spikes are noted in passing as "damaging to furniture." However, the worst aspect of spikes is the colouration they introduce. First, spikes are not quite rigid, and will resonate with the cabinet weight at a frequency somewhere between a few hundred and a few thousand hertz, often causing audible colourations.

Also, when pressing down, the speaker vibrations work against the stiffness of the spike and the floor, but when pushing up, they only work against the cabinet weight. For sufficiently strong vibrations, this amounts to rectification, and will introduce distortion. Of course, one can bolt the speaker to the floor, but this may make redecoration difficult, and cannot be recommended.

Hans J Albertsson
Stockholm
Sweden

But what about the marketing angle?

I've just finished reading an editorial by Mr. Watkinson disguised as an article about loudspeakers.

I'm afraid it is full of his own personal pet dislikes and contains some truths and some half truths about loudspeakers. But most significantly it contains the typical pro audio moan about hi-fi purchasers, i.e. "I'm the expert and you are ignoring me."

I have worked in professional audio for the past ten years and have run into this smug attitude so many times. It does the professional no good.

Here are some facts about high-end audio purchasers which Mr Watkinson needs to know.

1) The number of domestic loudspeaker purchases in the UK above £5000 next year will be about a hundred. No need to tool up for cylindrical containment vessels then.

2) The number of tri-amped or quad-amped systems sold next year will be about zero. Audiophiles do not like committing their electronics to their loudspeakers. The argument has run for thirty years and includes motional feedback, negative output impedance amplifiers, and digital input speakers. All fail. Why? People don't buy them.
3) Electrostatic tweeters, liked by the John, I have no particular problem with. Quad failed as a company due to lack of advertising and was willing to sell out to a holding company.

The point is that very good passively crossed-over loudspeakers can be made in square wooden boxes with bass reflex ports. It just takes skill. If you make a very good one, advertise it relentlessly.

If you are very patient, you will sell a few. If you make a Hawksford current-mode loudspeaker with motional feedback, three way Lipschitz-Vanderkroy linear phase crossover for six external current mode poweramps, and throw in-a few dsp chips for left over error correction, you won't sell *any*.

Interesting isn't it? Manufacturers make what people want to buy. Anyway, I have had my rant but I would like to see the loudspeaker design submitted by Mr. Watkinson, as I am sure it will be better than all the stuff we are using at the moment.

Dave Mate
Warwickshire

Spherical speakers and cheap amplifiers

May I thank John Watkinson for an interesting, informative and amusing contribution in June's issue. I do agree with a number, of the issues raised.

On the subject of pressure vessels,we have been in engaged in speaker design for a number of years and have developed a speaker system housed in a sphere as he suggests. From the outset, as Mr. Watkinson pointed out, a rectangular box was not.the ideal enclosure to withstand the pressure waves developed from the bass or mid range driver.

One other reason to chose a curved interior is to reduce,the possibilities of standing waves building up between flat surfaces which can cause unwanted resonances.

Apart from the mechanical deficiencies of a rectangular box, wood has its limitations. We have found it impractical to isolate the internal sound energy from the sides of a wooden box, and the radiation which emanates to the outside – even in MDF is very distorted.

Though low in volume, the effect is quite prominent. I can only liken it to intermodulation distortion in amplifiers where a small amount can be very offensive.

To limit contributions from the enclosure-we have employed a-material specifically developed for speaker enclosures called Polybimin which is cast into a 15 litre sphere and forms the bass midrange reflex cabinet.

The cabinet is not-the only important element, we use a driver unit manufactured by a company which has been developing speakers for over 50 years and employs the latest in materials technology. Our driver has a carbon-fibre and Kevlar cone, though this unit is expensive, we have found it to be well worth the extra cost.

The complete units enjoy the advantages Mr Watkinson suggests and out perform conventional speakers by a considerable margin. Since they have gone into production I have wondered whether it is practical to provide a significantly better speaker by introducing active electronics.

I have to disagree with Mr Watkinson when he suggests that "power amplifiers are so cheap today." Many electronics engineers.– myself included – are currently working to design amplifiers which can drive loudspeaker without adding their own subtle nuances.

Class B transistor amplifiers have improved dramatically since the sixties. But Class A .still has the favour of the public, because of the lack of intermodulation distortion synonymous with transistor class B. This point. is reinforced by the recent renaissance being enjoyed by manufacturers of valve .amplifiers. However, the price penalties for these more exotic technologies are significant.

So while it may appear an amplifier is a simple and cheap item in detail it exhibits.considerable complexities, on a par with any other part of the audio chain.

J R Charlesworth
Ikon Audio Ltd

Loudspeaker reverberations

John Watkinson's article on loudspeakers, *EW* June, contains some interesting ideas and some useful challenges to accepted practice. However, he falls into an error common amongst those who approach monitoring from a domestic viewpoint.

He states: "This renders the use of heavily damped rooms for monitoring suspect This approach has no psychoacoustic basis and has simply evolved as a practical way of using loudspeakers having poor directivity."

Any acoustic textbook will show that a reverberant enclosure will produce standing waves with their associated pressure *maxima* and *minima*. The exact characteristics of these depend on the dimensions of the enclosure, the wavelength of the sound, the position of the sound source and the position of the listener or measuring microphone.

Below 200Hz the difference in sound-pressure level between nodes and antinodes can be as much as 20dB – which makes nonsense of any speaker curves.

The solution for a single listener is to be in the near field of the speaker, i.e. where the direct sound field is greater than the reverberant.

In a control room, as found, say, in a recording studio, there will often be more than one listener,. For example, there may be a recording engineer, the producer and perhaps one or more artists. There is not enough room at the console for all these so some will be sitting usually at the back of the room. It is important that all listeners hear as nearly as possible the same sound balance. Since pressure *maxima* occur at a pressure boundary, there will be a large increase of low frequencies at the back wall if there is no low-frequency absorption.

High-frequency absorption is necessary because delayed reflections interfere with the direct sound and lead to 'smearing' of transients, upon which stereo image and general clarity depend.

These points are valid, of course for any loudspeaker, even an ideal one.

Control-room design is complex. The continuing number of papers at AES conventions is proof. It certainly involves more parameters than simply overcoming one problem of practical loudspeakers.

S W Davies
Aylesbury
Bucks

Loudspeakers explored

I read with interest John Watkinson's article 'Loudspeakers exposed'. The author's discussion of traditional loudspeaker design is very superficial. I appreciate the difficulty involved in presenting the subjective aspects of sound reproduction in print; the endless correspondence in your journal over the years bears witness to this. The author refers to "countless learned papers," but there is not a single reference for those who wish to read further.

The proposition that active loudspeakers are the way forward is an interesting suggestion, but is it possible to produce a satisfactory design? I suggest that *Electronics World* could run a design competition for a system, perhaps sponsored by a manufacturer of drive units.

I am forced to the conclusion that the results would have to be judged subjectively by a listening panel, but the rules should insist on a reasonable standard of technical justification for the design, to prevent entries that have been arrived at by trial and error.

The winning design should of course be published so that readers can build and compare it with their existing speakers.

Jim Adams
Sandy
Bedfordshire ■

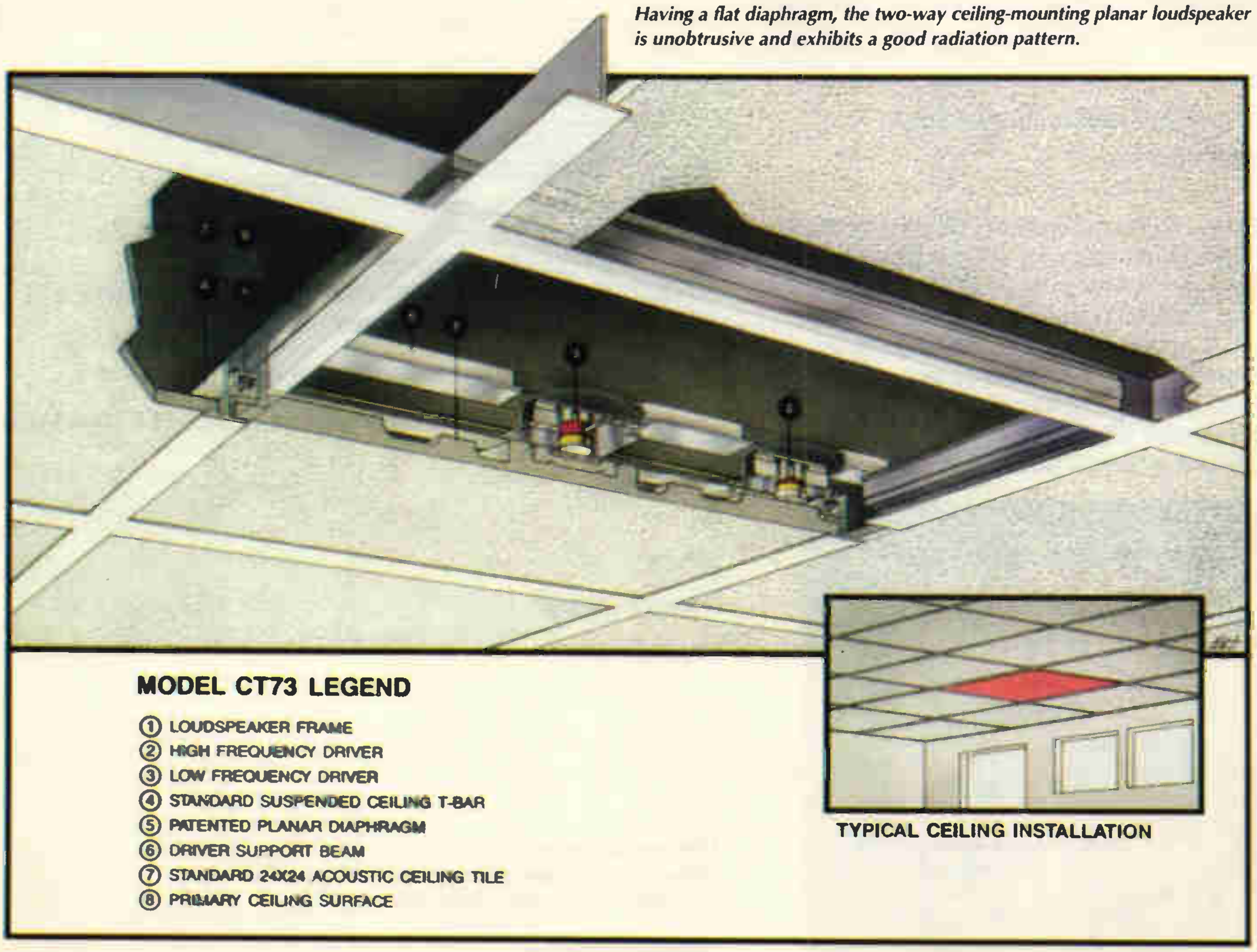
Flat but not flat

Rapidly gaining popularity, flat-panel loudspeakers are visually less intrusive than cone or dome alternatives, and they offer good omnidirectional performance. Gregory DeTogne explains.

Throughout the history of audio, loudspeakers using conical, domed, or compression driver designs have dominated virtually every segment of the marketplace. Still unknown to many, however, is the fact that there is an alternative to these well-known technologies which has been around for decades, and is currently making a play for a larger market share. Less obtrusive in style and configuration, this alternative product category departs from tradition in that it is based around a flat-panel, or planar design.

Within a dome-and-cone world, flat loudspeakers may sound like a notion which defies inviolable laws of physics. But it's not. The first truly practical planar loudspeaker was manufactured 67 years ago in the US by the Automatic Musical Instrument Company of

Having a flat diaphragm, the two-way ceiling-mounting planar loudspeaker is unobtrusive and exhibits a good radiation pattern.



Grand Rapids, Michigan. An electrostatic device used by the company in a coin-operated precursor to the modern juke box, this planar patriarch performed poorly by today's standards. It worked nonetheless, establishing a path for others to follow.

But others were slow in coming, if they came at all. As time passed, the further refinement of planar loudspeaker designs was stymied by cost. The materials required in the manufacturing process were simply too expensive to encourage a proper profit incentive.

More appropriate materials

All that changed in the 1960s when the materials required to produce planar loudspeakers on a large scale became widely available at more reasonable prices. Built using a number of plastics and charged films, flat electrostatic devices appeared in number, among which the most famous was the Quad *ESL 63*.

But even with the availability of cheaper materials at a manufacturing level, planar loudspeakers remained the red-headed stepchild of the audio world. Still widely perceived – accurately or not – as esoteric and expensive, their demand on electronics and somewhat unstable nature at the time often made ownership an ordeal.

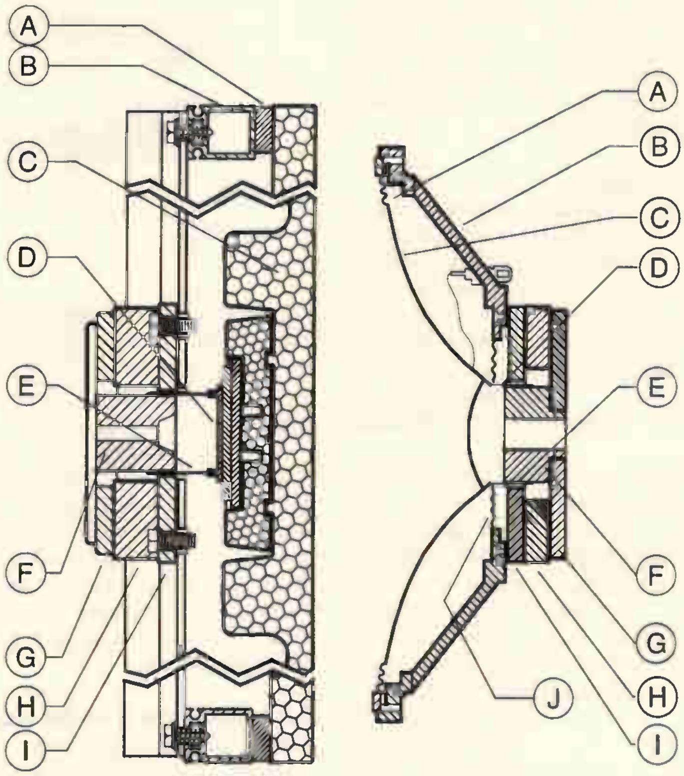
Meanwhile, back in the mainstream audio world of the mid-to- late sixties, a wellspring of advancements were made in traditional loudspeaker design such as air suspension and the development of long-throw transducers. At the same time, the evolutionary path taken by planar loudspeakers diverged into a three-way split.

One group, including manufacturers such as KLH, Martin Logan, and Infinity opted to maintain a conventional high-voltage approach. A second faction, of which Magnapan was representative, decided to seek out ways in which energy could be created using magnetic materials. A third, and considerably smaller band of individuals sought to construct planar systems from other types of materials to produce a reliable, cost-effective product capable of delivering true hi-fi performance.

Among this latter group was a physics professor named Jose Bertagni. Working in his native Argentina, Bertagni experimented with materials including thin sheets of wood and plastic to produce diaphragms to accompany his omnidirectional planar loudspeaker designs.

Diaphragm material

Ultimately, Bertagni discovered an expanded polystyrene-based formulation composed of individual beads which could be heat-formed into complex shapes while maintaining uniformity in structure and an even density. By 1970, he was granted his first US patent, and subsequently was presented with 22 others in countries around the globe.



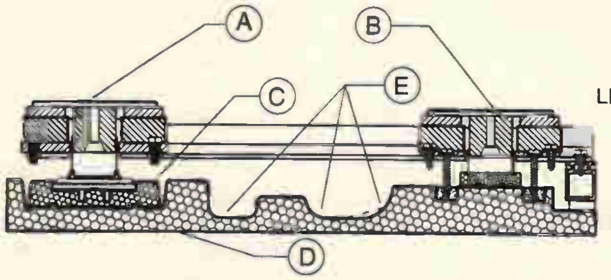
Conventional conical transducer, right, compared with Sound-advance's planar alternative.

LEGEND
- A SUSPENSION(SAS) / SURROUND (CONE)
- B FRAME
- C DIAPHRAGM
- D HAMMER (SAS) / COLLAR (CONE)
- E VOICE COIL
- F POLE PIECE
- G REAR PLATE
- H MAGNET
- I FRONT PLATE
- J SPIDER

Since then, 16 more patents have been granted worldwide for other Bertagni designs. Bertagni began manufacturing his flat-panel loudspeaker designs in Buenos Aires, but by 1975, he moved his operations to the US in Southern California to escape the increasingly hostile Argentinean political climate.

Today, the company is based in the Southern California town of Santa Ana, where an engineering team led by Bertagni's sons, Alex and Eduardo, continues to devise new planar designs (the elder Bertagni died in 1992). Having operated under the names Bertagni Electroacoustic Systems (BES) and Bertagni Electronic Sound Transducers (BEST) International for a number of years, the company is currently known as Sound Advance Systems.

While manufacturing under the BES name, the Bertagni family sold their products main-



LEGEND
- A LOW FREQUENCY DRIVER
- B HIGH FREQUENCY DRIVER
- C COUPLING DISC (DENSITY 'A')
- D DIAPHRAGM (DENSITY 'B')
- E INTERMODULATION TRAPS

Dual-driver planar loudspeaker in cross-section highlights the grooves and channels used as intermodulation traps.

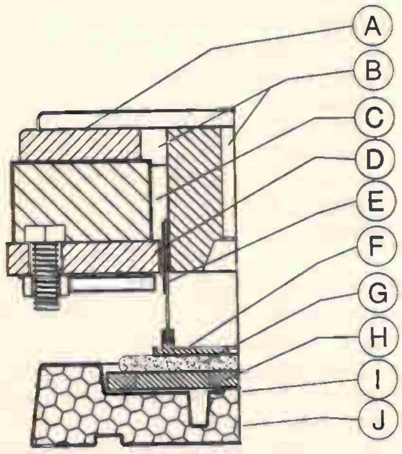
ly to high-end consumer audiophiles. Their success with BES Geostatic loudspeakers in this market funded research to develop other applications of their planar technology, and by 1980, the company had expanded its marketing efforts to include commercial, pro sound, and residential applications.

Benefits of planar

The advantages of planar loudspeakers over conventional designs are essentially twofold. First, their flat configuration allows sound to be dispersed from the entire surface of the diaphragm as opposed to just the centre. This translates into omnidirectional performance characteristics which spread sound out evenly across the entire listening environment, not just directly underneath the loudspeaker source.

Secondly, flat-panel transducers can be easily concealed. Within Sound Advance's current product line-up, models are offered which can be installed within a wall or ceiling in a fashion rendering them completely invisible. They can even be painted over or wallpapered. Models designed expressly for distributed overhead systems mimic the size and exact appearance of tiles used in drop ceilings.

In a most basic sense, here's how the current generation of Sound Advance planar loudspeakers work: instead of propagating sound by passing it through a cone or dome-shaped structure, an electrical signal is sent through a copper-wire coil, which produces a magnetic field causing vibrations. This coil is attached to the back of a specially-fabricated diaphragm made of polystyrene.



Planar motor and driver with the diaphragm facing down.

Vibrating back and forth, the diaphragm bends like an archer's bow to produce a wide variety of pitches which vary according to where it's struck by the coil. As previously mentioned, the sound itself then emanates from the entire surface of the diaphragm. When paint or wallpaper is applied over these loudspeakers, it merely becomes part of the vibrating surface.

The difference is in the diaphragm

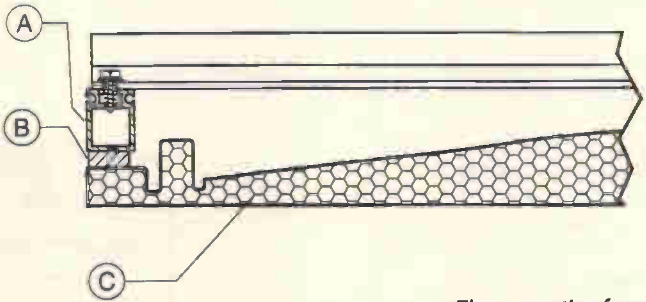
Within this operating scheme, it is important to remember that the diaphragm is what sets this transducer technology apart from all other types of radiating sources.

During the manufacturing process, special additives are added to the raw polystyrene bead material used by Sound Advance which enables it to meet ASTM E84 flame-spread and smoke-density ratings. Then the material is formed under heat and pressure via a wet molding process which also provides the shaping needed to create grooves and channels called intermodulation traps which reflect return energy from the diaphragm edges away from the coil.

An environmentally-controlled curing period follows, during which the diaphragm is stretched and bonded to its supporting frame with proper amounts of tension. Dampening weights are added at this point in time as well, at a number of points in each system to tune the motional response of the entire motor structure.

Upon completion of this process, the motor structure houses a variable-thickness diaphragm which rigidly complies to lower frequencies while allowing a carefully-controlled, progressive decoupling of the diaphragm's surface from the driver to reproduce higher frequencies.

Given its very nature, the entire diaphragm assembly radiates energy at lower frequencies while a controlled and tailored reduction in effective diaphragm surface area occurs as the frequency increases. This produces exceptionally wide angular coverage at higher frequencies as a result.



The supporting frame of the planar loudspeaker seen in cross section shows the equivalent of the conical loudspeaker's surround.

Conversely, low frequencies move the diaphragm backwards and forwards from a central position in a motion dampened completely at the edges. Mid and high frequencies travel from within the composite material toward the outer edges, where they are extinguished inside the intermodulation traps.

Collectively, all of this activity energises a large number of the individual polystyrene beads comprising the diaphragm's base material. When each bead comes into contact with an adjoining bead, it radiates just like a miniature point source.

Over the years and through the various name changes, the company which is now Sound Advance Systems has seen its planar loudspeakers installed in the White House, King Fahd's palace in Saudi Arabia, the Greek Theatre in Los Angeles, the Mirage Hotel in Las Vegas, the Hard Rock Cafe in New York City, the Hotel de Coronado in San Diego, the Ritz-Carlton in Dana Point, California and the Hollywood Bowl.

Other players

Other manufacturers currently involved in the flat-panel loudspeaker race in the UK include New Transducers Limited (NTL) and NCT. Designers of the well-hyped NXT line of components, NTL promises to bring their first flat-panel products to market in the very near future.

As for the future of flat-panel loudspeakers in general, based upon all evidence, the technology has matured to the point where it is here to stay. In addition to the applications described thus far, those imbued with the entrepreneurial spirit are already aglow with an abundance of ideas.

Projection television screens could easily double as a loudspeaker, for instance. Or how about having the headliner of your automobile serve as a loudspeaker system too? Then there's the notebook pc equipped with sound panels that slide out of the sides next to the display screen. And don't forget a talking microwave oven, combination solar-powered toaster/portable stereo, and... well, let's just say the possibilities are endless for now, and leave it at that. ■

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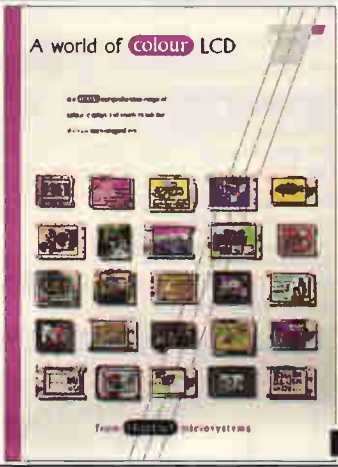

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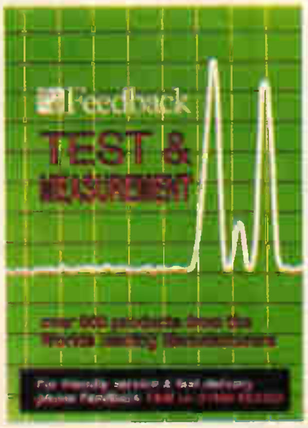

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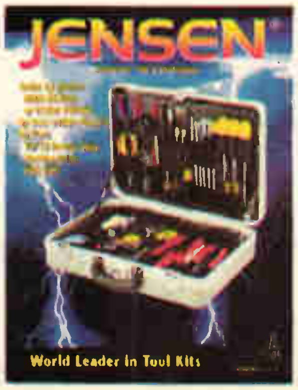

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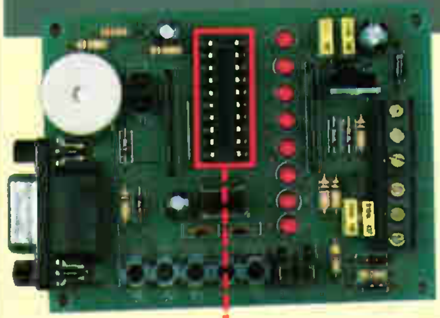
Analogue comparator	-	-	-	-	-	YES	YES
Data pointers	1	1	1	2	2	1	1
Package Pins (DIL)	40	40	40	40	40	20	20

C51 Microcontroller Starter System



- ✔ Optimising C Compiler
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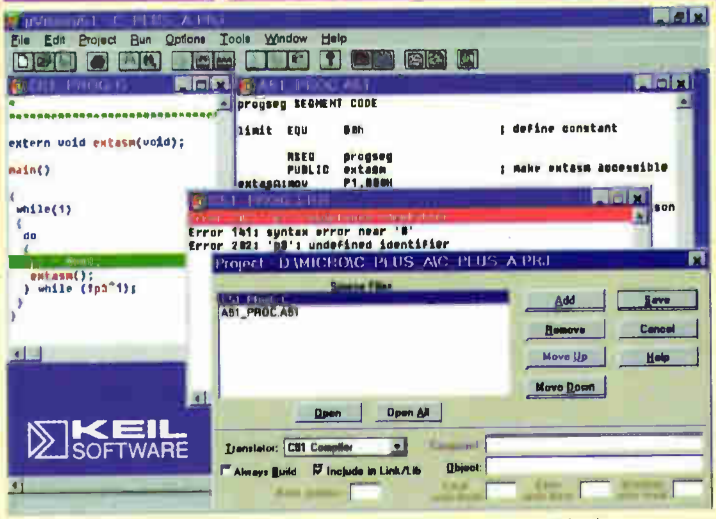
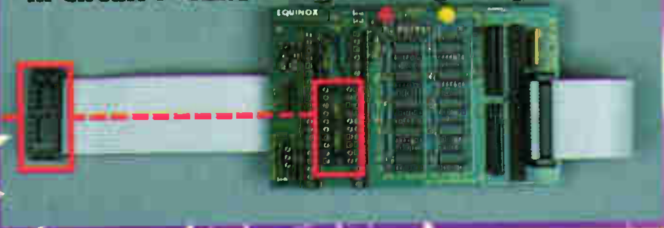


▲ KEIL Integrated Development Environment - C compiler + Assembler output restricted to 2K total program code.

In-Circuit Parallel Programming Adaptor



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