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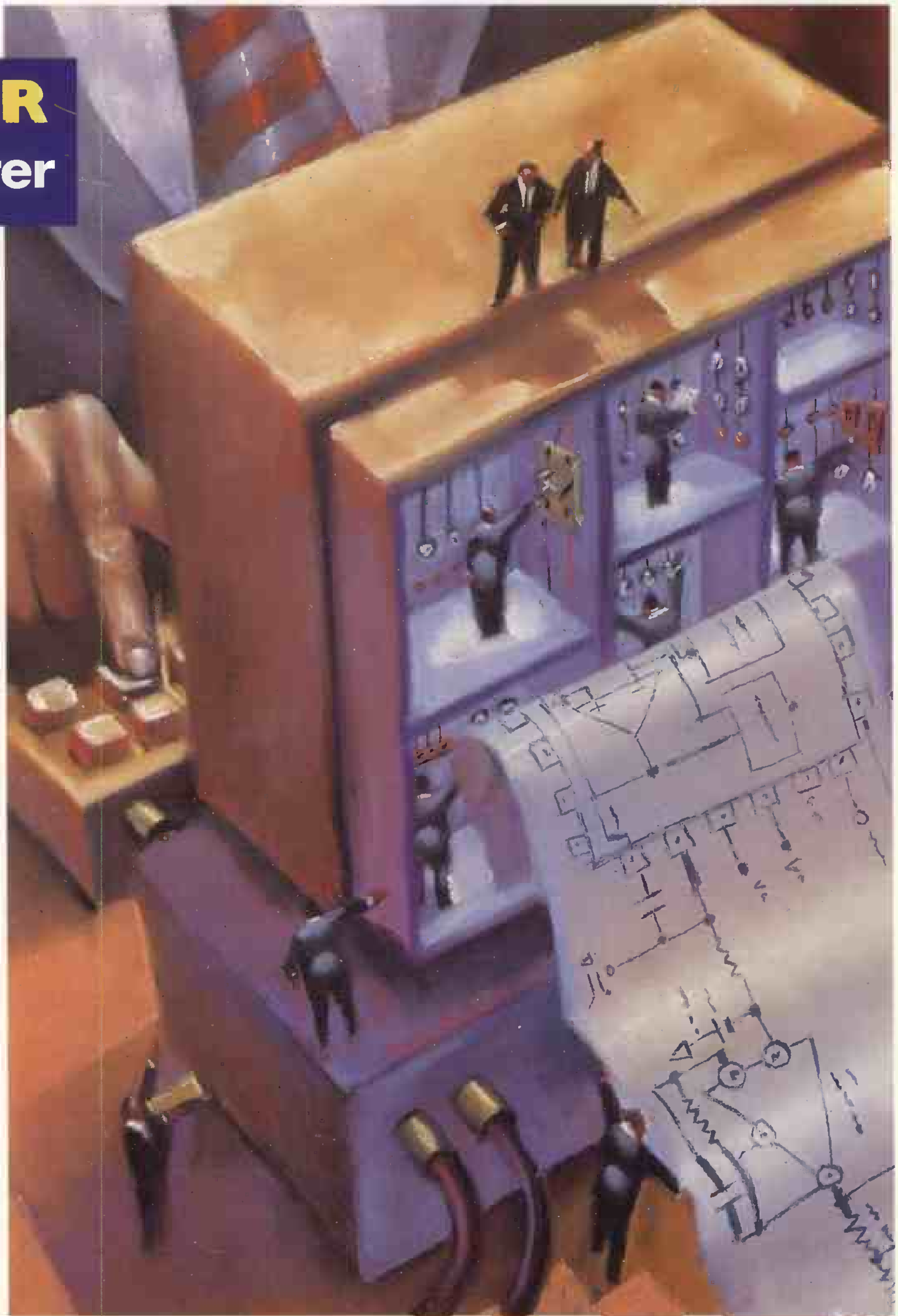

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FEATURES

Cover: Jamel Akib

£4000 WRITING EXCELLENCE AWARD
By writing the best rf design article of 1995, you could win a Hewlett-Packard HP8647A signal generator worth £4000. Details on page 889.

REGULARS

In next month's issue:
Underground communication – Mike Bedford looks at methods of making wireless links in caves, and presents a working transceiver design.
A new class of mosfet power amplifier – Bengt Olsson challenges the 30-year-old complementary output stage.
THE DECEMBER ISSUE IS ON SALE FROM November 24

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REED BUSINESS PUBLISHING

Abuse of the licence fee

Nobody doubts for a moment the ingenuity and technical excellence of the BBC's digital audio broadcast system – DAB. Demonstrations show it to offer an improvement on standard fm, particularly for mobile reception. Why then should we think very carefully before endorsing the new broadcasting system, now scheduled to start about a year from now?

It is simply this. The technology, while feasible, takes little account of broadcasting requirements. It has been tailored too closely to the needs of the BBC's national network with little account of local broadcasting.

The precise details of the technology are involved; this will be borne out by the price of receiving equipment if it ever becomes available. The essence of the DAB system is this. Each DAB transmitter broadcasts six separate programmes simultaneously using subcarrier interleave. The frequency spreading inherent in the DAB system reduces the individual data rate per unit carrier frequency to the point where multipath interference is no longer a problem. However, it requires that six programmes are transmitted simultaneously from a single site.

Where used for local broadcasting, it implies that six stations are locked together in an inflexible bundle. Six franchises would have to be offered to serve a local area since DAB only represents efficient use of frequency and financial resource when fully occupied.

While this arrangement clearly suits Radios 1 to 5 plus another, it leaves local radio out in the cold.

The EC, which sponsored BBC DAB research, feels compelled to push the system to take advantage over emerging US technology in setting world standards. American digital sound broadcast technology takes as its starting point the elimination of transmission shortcomings from individual stations. And most broadcasting systems around the world operate like the Americans.

One cannot argue with the sense of promoting an home grown broadcast standard but it has to be something in tune with a market requirement. We don't get too many letters complaining about the difficulty in retuning car radios to national networks as drivers cross the country. In any case, autotuning RDS radios were going to cure that, weren't they? We do however receive complaints about the lack of stereo coverage on the BBC television network.

Nicam has to be the most undersold broadcasting benefit ever offered to the public. The sound engineering on most television programmes is simply stunning. Indeed, it puts many straight radio productions to shame. To start putting money into something which offers no tangible benefit to the average listener, which requires high priced receiving apparatus, which will be unusable for most local radio, is an abuse of the BBC's licence fee.

The BBC's plan is to update tv transmitters to stereo only when the existing installed equipment comes to the end of its normal service life. This means that great swathes of the country will have to wait up to 20 years or more to get stereo sound on BBC tv channels.

We should not accept further developments from the BBC until it has fully implemented its television stereo sound service.

Hello, hello

From my new post as editor of *EW+WW*, I would like to reaffirm Frank's belief, expressed in the final words of his parting comment last month. I do share his ambition to develop *EW+WW* towards applied electronic design and I would like to thank him for the excellent work he has already carried out to this end.

One small change I will be making is to open up this comment page to a wider audience from time to time. I will be inviting leading figures in the industry to make their contribution, an example of which is the above guest editorial from Frank. But I will also be considering thought-provoking, stimulating ideas and statements from you – the readers.

Martin Eccles

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Live '94 – surround sound to Internet

Watch tv through a window. Apple's new Performa 630 TV Plus includes a TV tuner for its Multimedia environment.

In the forecourt at Earls Court, where they usually park a yacht during the Boat Show, or a tank during the Royal Tournament, Live 94 had a large articulated-lorry trailer, its black canvas side used as a billboard for the show. Unimaginative, perhaps, but as a symbol of the show within, it couldn't have been more appropriate.

The key concept at Live 94 was the Black Box – it delivers marvels of technology, but no need to ask how. This was perhaps to be expected, since it was more of a punters' show than one for techies, but it made for an experience that was often curiously contentless or elusive. After all, we've all seen a television picture. So what, if there are lots of them together? Or if they're a different shape? The images and the programmes on the screen remain much the same. It became hard to work out what you were supposed to be looking at, or for. Surround-sound demonstrations were particularly frustrating, unable as they were to compete with the ambient cacophony.

Nevertheless, there were some notable debuts. Home cinema – hitherto shorthand for surround-sound – took two decisive steps towards fully replicating the real thing. Channel 4 and Nokia organised the first live PALPlus widescreen transmission at the show.

Using processors newly brought back from the International Broadcasting Convention at Amsterdam and hastily connected-up, history was made on the Thursday with a 1963 CinemaScope film called Bye Bye Birdie. The Widescreen TV Forum, a DTI-led consortium of British broadcasters and manufacturers, went one better with a live demonstration, transmitted from Croydon, of MPEG2 digital compression, to 8 megabit/s, the equivalent of four channels into one. As for sheer size,

Continued over page

Car navigation systems are on course

For drivers who find it difficult to glance at a map and navigate while keeping both hands firmly on the steering wheel, there's good news around the corner.

Driving through cities may never be the same again as more and different types of dynamic route guidance systems are being mass-produced and implemented even in the most basic car models.

According to BIS Strategic Decisions, the international consultancy group, by the year 2002 automotive manufacturers will be fitting more than 6 million vehicles a year with navigation systems as they leave the factory. Route guidance is becoming a vital part of an integrated traffic management system. Well-coordinated traffic management improves fuel efficiency, reduces pollution and eases congestion on the roads – and lightens the workload on the driver.

During 1996 and 1997, more European and North American car manufacturers will start to offer stand-alone navigation systems as options on their models.

Such systems are Renault's Carminat and Ford's Route Guidance System which use, Radio Data Systems, RDS.

Renault's Carminat consists of an on-board computer that connects to a Radio Data System - Traffic Message Channel (RDS-TMC) broadcast on fm. The information is displayed on an lcd screen on the dashboard.

Ford's system will also navigate the driver through the shortest or fastest route to a given destination. The advice will be given visually or audibly.

By 1995, Mercedes-Benz will equip its S-class models with a GPS-based system called Auto-Pilot System, APS. The vehicle's position will be determined by consulting digitised maps of the city roads and road network on CD-rom.

But probably the most interesting type of route guidance is coming from CellPort Labs in the US, called C/P Connect. The system is an in-vehicle local area network which interfaces with around 23 different devices in the car including the cellular phone.

The biggest difference is the wireless connectivity of this system. A vehicle's problems and accidents can be automatically relayed to the appropriate party like the police, ambulances or the rescue services.

In Europe, similar systems are already taking off, one is the German Copilot system.

Copilot is a vehicle guidance system which literally leads a vehicle from traffic light to traffic light after the vehicle destination has been typed in by the driver. It can book parking spaces on request and give information on public transport links and timetables. Trials are scheduled – or have already begun – in many German cities, among which are Stuttgart, Frankfurt, Bremen and Dusseldorf.
Svetlana Josifovska, *Electronics Weekly.*

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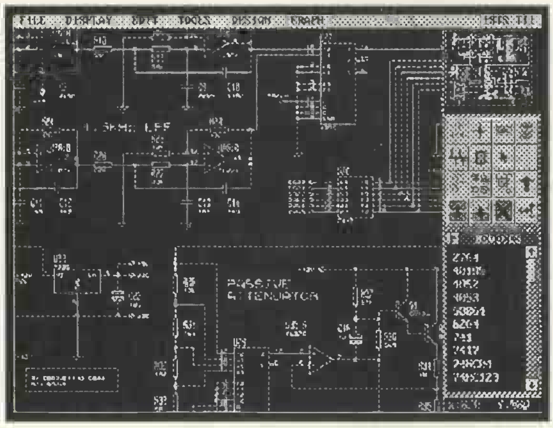

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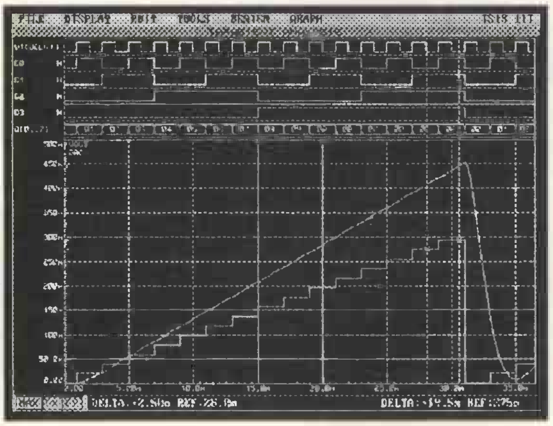

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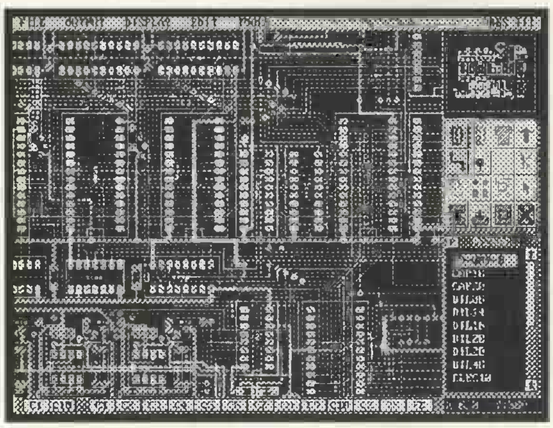

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CIRCLE NO. 105 ON REPLY CARD

Sharp was demonstrating two lcd projectors which produced impressively... er... sharp images on the big screen.

On the interactive front, attention has switched from games machines to more powerful and versatile CD-based systems. Panasonic was pitting its new 3DO against Philips' more established CDi.

But both are having to reckon with an aggressive contender from another sphere. Two of the largest stands on the main floor were those of Apple and Microsoft. Both majored on the same themes: education, entertainment and the home office – a triple whammy that searches out the weak spots of consoles marketed as tv add-ons. CD-rom in conjunction with a pc or Mac is beginning to develop a bandwagon momentum that could lift the computer from being the prerogative of the minority (now a very large minority) to being as natural an item of home technology as the telephone or the television.

Indeed it may well end up swallowing both. Apple chose the show to unveil its new Performa 630 TV Plus, a multimedia bag of tricks with built-in tuner that allows you to watch television, either as background desktop or in a small window, while working on something else. It may seem like a gimmick now, but it makes better sense as an interface for serious interactive tv shopping than the sofa, the widescreen and a remote control.

Compaq launched its new pc-based Presario. Also aimed at home users, its package includes fax, answering machine and access, via CompuServe, to Internet. The Internet was still an unofficial item on this year's Live agenda. Not for much longer, one suspects, with BT launching its own service in October.

Outside the home, the most significant pointer to the future was an inconspicuous display on the stand of in-car specialist Alpine. It consisted mainly of a small, bright lcd screen about inches (100mm) square, showing a road map, and formed part of a car navigation system, already in use in Japan. Much development work, by others as well as Alpine, is going into these at present. The aim is to combine location data from GPS with traffic information on RDS to guide motorists not only to their destinations, but around any hold-ups that may occur. Most of the elements for this are already available – the only one lacking for the UK being digital mapping, expected by 1996.

Live is by no means comparable with the Berlin Funkausstellung, but it does give one a sense of being at, or somewhere near, the leading edge of consumer electronics technology. These days, that is a pretty exciting and multi-faceted place to be.

Peter Willis

Cmos superseded by Leaps?

A new chip technology has been developed that significantly outperforms cmos and which is believed will eventually replace cmos for all logic circuits.

Called Leap – an acronym for lean integration with pass transistors – the technology is based on a type of transistor, developed by Hitachi. Unlike cmos circuits, where one transistor charges the output of a logic element and another discharges it, in Leap circuits a single 'pass transistor' both charges and discharges the output. Benchmark tests carried out by Hitachi show that Leap circuits beat cmos circuits on speed, silicon area and power consumption by factors of between 30 and 50 per cent.

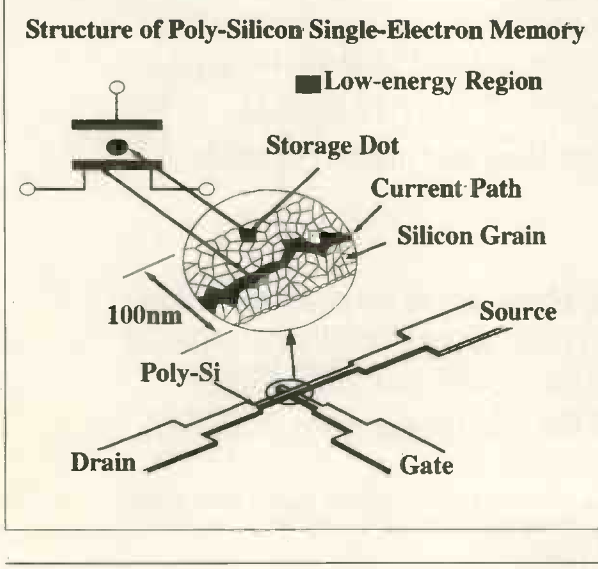
Hitachi is already planning to use Leap in its next-but-one low-power microprocessor, due in 1996 and called the SH4. The 32-bit device is expected to deliver 300Mips and consume just over one watt.

Dr Tsugio Makimoto, Hitachi's main board director responsible for semiconductors, believes that future performance will increase by a factor of two every three years, without increasing power consumption.

The building blocks of Leap circuits are not traditional logic gates such as nand and nor gates but more complex units with more inputs. They require different design methods and new libraries of standard functions.

Cmos technology for logic circuits has been around for the past 15 years. The concept of using a single mos transistor to pass on a charge state has also been around for some time. But it is only in the past year or so that researchers at Hitachi's labs have built pass transistor circuits and shown just how well they perform. Hitachi has applied for a patent on the technology.

Single electron switch using polysilicon



Hitachi is investigating the use of polysilicon to build single electron devices such as switches and memory elements based on quantum mechanical effects. Last year the company unveiled early results from small conducting islands separated by very thin insulators, which electrons can cross by quantum mechanical tunnelling. Recently, devices have been built using the jigsaw-like structure of polysilicon to form the islands. Hitachi has demonstrated memory operation using polysilicon, in which one of the polysilicon islands acts as the floating gate in a non-volatile memory element.

Solid state recording for camcorder

A camcorder is being developed which will use 256Mbit flash chips to store video sequences. Hitachi says the device will be the world's smallest and lightest camcorder and will fit into the palm of the hand.

Using video compression technology, Hitachi believes it will be able to store 30 minutes of digitised video in 400Mbyte of memory. The company says it hopes to launch a commercial chip-based camcorder within five years, at a price of around $1,000.

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New coating technique boosts floppies to 100Mbyte

Japanese photographic film maker Fuji Photo Film has developed a manufacturing technique which it claims will revolutionise floppy disks. By using a new metallic coating technique, Fuji claims it is possible to store between 100 and 200Mbyte of data – nearly 100 times more than the 2Mbytes for conventional disks.

The key feature is in the coating technique. "We coat the discs with very thin layer of material but we do it simultaneously with the undercoat," said Yasuhiro Abe, manager of the technical department of magnetic products at Fuji Film.

The thin, top layer is 0.1μm to 0.5μm thick, compared with the 2μm of conventional coatings. It consists of ultra-thin ferrite particles and is applied using a new process technology. The new double layer coating enables more information to be stored per unit area and data can also be read more quickly. Discs using the new coating can revolve at between 3000 and 5000rev/min and hold up to 2000 tracks per inch.

The new metal-particle (MP) process is a less expensive technique than the traditional metal evaporated (ME) coating currently used to prepare floppy disks. It has evolved from manufacturing photographic films for the last 30 years.

Two years ago Fuji used the technique on the 8mm and VHS camcorder market and produced camcorder tapes coated with MP 'super-double' layer.

Currently Fuji is co-working with a disk drive manufacturer to make the hardware needed for the high capacity disks. Fuji expects to find a manufacturer which will produce disk drives that read data ten times faster than current disk drives.

UK takes lead in optical IC production

A British start-up has cracked the problem of building integrated optical circuits and is about to set up the world's first production plant.

Bookham Technology has developed what founder Dr Andrew Rickman calls the cmos of optical circuits.

"We are already in prototype production," Rickman said, "and we are now investing in an assembly line. Investment in a clean room will probably come six months to a year after the assembly line has proved itself."

The chips are based on two fundamental breakthroughs: a way of building optical structures such as switches and amplifiers in silicon, enabling conventional IC production techniques to be used, and a way of coupling optical fibres to silicon waveguides on chips.

3D television nearer to reality

Researchers at De Montford University in Leicester believe they may have found a practical way to deliver three dimensional tv pictures.

Experimental work carried out by the team suggests that it could be much easier than previously thought to build a system which can capture, transmit and display full-parallax 3D tv pictures. It could mean that pictures in which the viewer can look round the image by moving his or her head will not now need special glasses.

According to a paper delivered by the team at an International Broadcasting Convention recently it should be possible to build a full colour 3D display using receivers with horizontal resolution of 2048 pixels and a vertical resolution of 1536 lines, requiring a transmission rate only about 1.5 times higher than that needed for 2D high definition tv (HDTV) pictures. To send this using current transmission methods, the data would need to be compressed by a modest factor of 16:1.

The work is based on a theoretical three-dimensional imaging system which uses a single integrated imaging device made up of arrays of large and small lenses to capture the picture, rather than multiple cameras.

The theory behind such devices has been known for some time, but it has only recently become possible to make planar arrays of micro lenses with a fine enough pitch. In the De Montford system, the finest micro-lens arrays have a lens spacing of 125μm.

The team used printed images to model the pictures which would be produced by such a system on a liquid crystal display. Using a viewing micro lens array with a lens pitch of 1.25mm the team found that the parallax information in the images was preserved even with a single pixel behind each lens in the array.

Karl Schneider
Electronics Weekly

Optical fibres have terahertz bandwidth potential

Optical fibres open up the possibility of networks with virtually unlimited bandwidth – in principle. The transmission range of optical fibres, covering wavelengths between 1.3 and 1.6μm, represents a bandwidth of some 50THz.

While 50THz would never be achievable down any significant length of fibre, today's networks still only make use of a tiny fraction of the bandwidth potential.

New all-optical devices being developed could make better use of fibre capacity. Most attention has focused on generating and decoding optical signals and it is now possible to generate optical pulse streams with repetition rates of a few Gbit/s and pulse widths of around a picosecond. This is achieved using mode-locked semiconductor lasers, which emit narrow band light.

In future bit rates of several hundred Gbit/s could be possible using post-connected passive optical multiplexers. A semiconductor laser generates interleaved pulses to multiply the bit rate before being fed into the fibre.

It has only recently become possible to build fast optical demultiplexers based on optical amplifiers using intra-band effects. An optical control signal pulls out the relevant pulses from the 'multiplexed input stream. Researchers at the Heinrich Hertz Institut in Germany believe such devices could handle signals to 100Gbit/s.

In principle, higher capacities could be achieved by combining tdm with optical frequency division multiplexing.

Several carriers of different wavelengths can give very high bit rates. Dr Alan Hill at BT's Martlesham laboratories says, "You may even be able to use thousands of carriers at distances of perhaps 1000km."

Switching signals in the optical domain is a thornier problem. Three types of switch are required for space, frequency and time switching.

Spatial switches are needed to correctly route a signal. Researchers from British Telecom's Martlesham labs, Cambridge University and University College London are developing 2×2 optical spatial switches that can switch the output from each of two input fibres into the correct output fibre, using liquid crystal shutters.

Optical frequency switches take an input signal on one carrier wave-length and translate it to a different wavelength. Such devices have been built in the lab using four-wave mixing, where the input signal is combined with control signals in a non-linear optical medium to generate the desired output wavelength through non-linear combination.

At the Heinrich Hertz Institut, intra-band effects in 'optically pumped amplifiers have been used to shift an optical carrier by a few THz while carrying a modulated 18Gbit/s signal. ■

Karl Schneider
Electronics Weekly

Best rf article '95

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

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

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

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

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

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CIRCLE NO. 106 ON REPLY CARD

RESEARCH NOTES

Jonathan Campbell

Deducing acceleration by induction

In addition to being miniscule, newly developed silicon inductive acceleration sensors are much easier to interface than their capacitive counterparts.

Aircraft navigation, to impact testing, to automotive safety systems – the range of uses for accelerometers is growing daily and there are already several technologies from which to choose. But researchers in the Department of Electronics, University of New South Wales, have just added a new one – a micro-accelerator fabricated from silicon wafers that uses electromagnetic field induction (*IEEE Electron Device Letters*, Vol 15, No 8, pp.272-273).

Measurements of acceleration ranging from 0-50*g* have already been demonstrated by the device, making it ideal for applications such as air-bag sensors in cars. The great benefit is that the associated electronics are much simpler than with conventional systems such as capacitive accelerometers.

In structure, the device developed by Ebrahim Abbaspour-Sani and colleagues, has much in common with an ordinary transformer, having primary and secondary windings of 12 turns each.

The lower section of its three part silicon construction contains the primary square coil, patterned from an evaporated aluminium layer using photolithography onto a thermally oxidised *p*-type silicon wafer . The 2.2x2.2mm seismic mass that senses the acceleration is suspended from the upper layer by two cantilever beams 600μm long. It is micromachined from *p*-type silicon and has the secondary planar coil sputtered onto it. Separating the two layers is a middle section providing the required spacing between the two coils and also assisting in the alignment process.

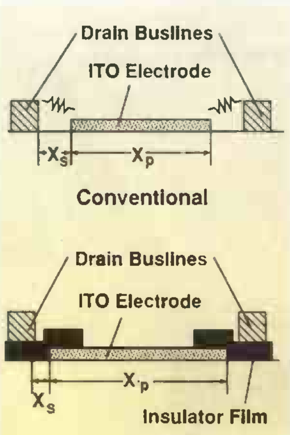
The three parts are assembled after fabrication, and signal conditioning circuitry, comprising a preamplifier, a precision rectifier and a dc amplifier, is added. In use, output voltage is related to the distance between the upper and lower coils, which will vary in response to acceleration.

Overall size is 4.2x4.2mm and the amplified voltage varies linearly between 0-9V, at a rate of 0.175V/g and a power consumption of less than 2.5mW. The range of detectable acceleration may be varied by selecting different designs in length, width and thickness of the cantilever beams. As the researchers point out, since no micro-machining of the lower part is required, the electronic circuitry for signal processing can be included on it using standard cmos technology.

The main drawback of the device is currently its large temperature drift. At room temperature, without temperature compensation circuitry, the amplified voltage drifts by ±0.2V/°C.

Buried technology widens market for lcds

Brighter, cheaper lcds: the buried indium-tin-oxide structure, ITO, allows the busline/electrode spacing X_s to be reduced so widening the electrode size X_p. Result is improved aperture ratio.



As information systems shrink, the demand continues to expand for thin-film transistor-addressed liquid-crystal displays (tft-lcds). Unfortunately, traditional devices are expensive to produce and, in portable equipment where power consumption is so important, brightness can suffer. But a new design of tft-lcd emerging from the laboratories of Hitachi in Japan *IEEE Transactions on Electron Devices*, Vol 41, No 7, pp.1120-1124) promises to tackle both those problems. A buried ito electrode (bi) structure helps reduce device cost by hugely improving the production yield. Power consumption for a particular brightness is cut by boosting the aperture ratio by almost a half.

Conventional technology has already put 10in diagonal 480 line panels in production, with 700 line displays under development. But production defects cause poor yields.

Intra-layer short circuits, resulting primarily from etching residues, are the main problem causing most pixel defects in lcd panels. Combined with aperture-ratio limitation, which has so far been a constricting factor in optimising the trade-off between power saving and brightness, the pixel defect problem continues to test the minds of display engineers.

The Hitachi design improves the aperture ratio while simultaneously reducing pixel defects, using the novel pixel structure with its buried electrode and an anodic oxidised Al-gate electrode.

In conventional designs, the pixel electrodes and the drain buslines are installed on the same plane and so are easily shorted by electrode etching residues. The Hitachi design buries the pixel electrodes under an insulator film, isolating them from the drain buslines. So, short circuits are almost completely eliminated and pixel defects are significantly reduced.

Normally it is also impossible to reduce the spacing between the pixel electrode and drain busline without risking short circuits. So the width of the pixel electrode is restricted and the aperture ratio limited. But in the new structure, since the pixel electrodes are isolated form the drain buslines, the spacing can be reduced to the resolution limit of the photolithographic process, and the pixel electrode can then be widened.

The Hitachi researchers used a 10in diagonal high-resolution lcd and 780 lines to evaluate their design. Their buried-ito-structure improvement, added to a reduced storage capacitance which also boosts the aperture ratio, showed a ratio increase from around 20% for the conventional device up to 29% for the tft lcd.

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CIRCLE NO. 107 ON REPLY CARD

Sorting through signals to pinpoint power line damage

The US Navy, the Star Wars missile programme, and the humble bat have all had a role in a monitoring system being developed at the University of Rochester to help electricity companies locate power lines damaged after natural upheavals.

The work follows catastrophes such as an ice storm that hit the Rochester area in March 1991, or the earthquakes that regularly rock Southern California. Utilities currently depend on customer complaints to find downed power lines and this slows repairs since company personnel must piece together scattered bits of information to track down faults. But University faculty students, working with engineers from Rochester Gas and Electric Corp., are close to perfecting an automated process that would allow engineers to find immediately which lines are down.

The idea is to outfit each power line with a transmitter that every so often transmits a distinct signal back to the utility signifying that the line is working. A missing signal would indicate that a line is down. The clever part has been to separate out all the signals coming from hundreds of power lines.

Several years ago the Navy turned to Edward Titlebaum, professor of electrical engineering, to solve the similar problem of how to ensure that signals from sonar-guided torpedoes remained distinct and could not become crossed.

Titlebaum, who admits to a fascination for bats and their own sonar systems, began studying ways to classify signals so that a particular one could always be identified, analogous to picking out a familiar voice in room crowded with people talking.

His solution was to develop a family of codes known as frequency hop codes, where a signal hops across several frequencies per second. Last year he showed how the codes could accommodate a virtually infinite number of users on networks of computers, cellular phones, radar and other applications where multiple users are drawing on the same resource.

Now Titlebaum's codes have been incorporated into a computer chip designed by students working with faculty members Alexander Albicki and Edwin Kinnen. In recent tests with RG&E, the team successfully inserted and then extracted distinct signals from the sea of electricity that surges along a utility's power grid.

The team has sent and received signals between engineers' homes and offices in the university's computer studies building just by plugging in their computer to wall outlets.

So far RG&E has not implemented the system because it is currently too costly. But University engineers are confident they can bring the cost down.

In the US, detecting power-line damage relies largely on reports from consumers but a new electronic detection system could be used to pinpoint problems automatically.

Camera catches the sleepy driver

In a few years from now, if the driver in the car in front of you suddenly jerks upright and you hear a distant buzzing and wonder where that distinct scent of peppermint is coming from, you could be seeing in action the next stage of an anti-sleep system being developed at Nissan. According to a paper from the 14th Enhanced Safety of Vehicles conference (*Automotive Engineer*, Aug/Sept), the company has been experimenting with a camera-based system that can trigger an alarm if a driver falls asleep.

In addition to an immediate sounding of a loud buzzer, Nissan has also been assessing the effects of filling the car with the smell of scented oils to keep the driver awake for long enough to find a service station.

A small 512×432 pixel ccd camera aimed at the driver's face is at the heart of the system, linked to a pc to allow measurement of the openness of the driver's eyes. What has made the use of a camera possible is that, instead of electrodes or other brain-monitoring equipment, the researchers have correlated eye openness directly with fatigue, There is an an alertness index that points directly to the number and amplitude of a2 brain waves.

In a related study, stimulation with the smell of peppermint was found to keep a driver refreshed for up to six minutes, and when used with a buzzer could keep stave off sleep for 15 minutes.

There's plainly some way to go before such a system finds its way into a car, but in the meantime, it won't do sales of Polos (the mint not the car) any harm.

Pancake-like disks have been sighted in the Orion Nebula: are they propylids?

Space discs shorten odds on Aliens

The probability of finding life elsewhere in the Universe may have been raised slightly by a recent discovery made by the Hubble Space Telescope. Astronomers using Hubble to scan the Orion Nebula found over half the stars (*J Br Astron Assoc*, 104, 4, 1994) showed pancake-like discs around them rather than shells.

The presence of discs indicates that the dust has too much spin to be drawn into the collapsing star, but instead will eventually agglomerate to make planets.

Dr Robert O'Dell of Rice University – who has christened these discs of planets-to-be 'proplyds' – and Zheng Wen at the University of Kentucky have measured the mass at the edge of one of the discs and found it to be several times that of the Earth. But at less than a million years old, The Orion Nebula discs have not had time to form into planets.

However, astronomers believe that finding so many proplyds amongst the young stars, indicates that planetary systems are even more numerous than we thought. The result is that the odds of finding a planet with life have just been reduced.

Colour-spectrum led technology for display panels

Leds that change colour in response to variations in applied voltage have been announced by Scientists at Lawrence Berkeley Laboratory, California. The devices, made from cadmium selenide nanocrystal and a semiconducting polymer, can change from red or yellow at low voltages to a distinct green at higher levels. The researchers believe such devices could become the basis for colour displays made up of multi-coloured pixels, using voltage control to dictate colour.

Key to the colour control is the Berkeley team's new approach to led design using a hybrid organic/inorganic nanocrystal composite structure built up by assembly of individual components. The colour change effect stems from the led's combination of two radically different materials with disparate dielectric constants and transport mechanisms, that support two different luminescent mechanisms (*Nature*, Vol 370, pp.354-356).

Layers of CdSe nanocrystals (<10,000 atoms) are built up on a substrate, up to a thickness of a few hundred Ångstroms, while the organic component is a layer of *p*-paraphenylene vinylene. This layer is chosen to ensure electrical stability and enhance carrier injection and confinement.

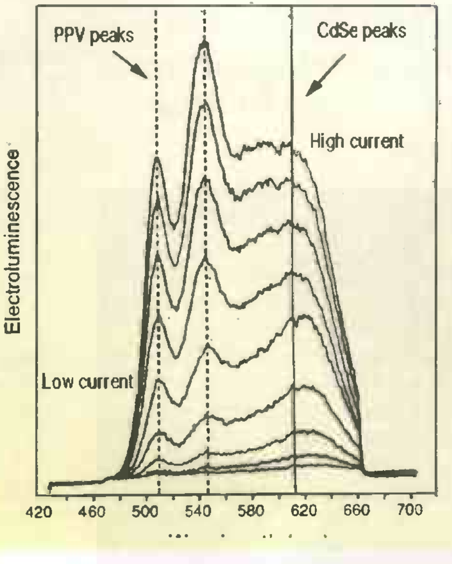
Light emission arises from the recombination of holes injected into the layer of the semi-conducting ppv with electrons injected into the multi-layer film of cadmium selenide. Starting-point luminescent colour of the CdSe layer can itself be altered from yellow to red by changing the nanocrystal diameter

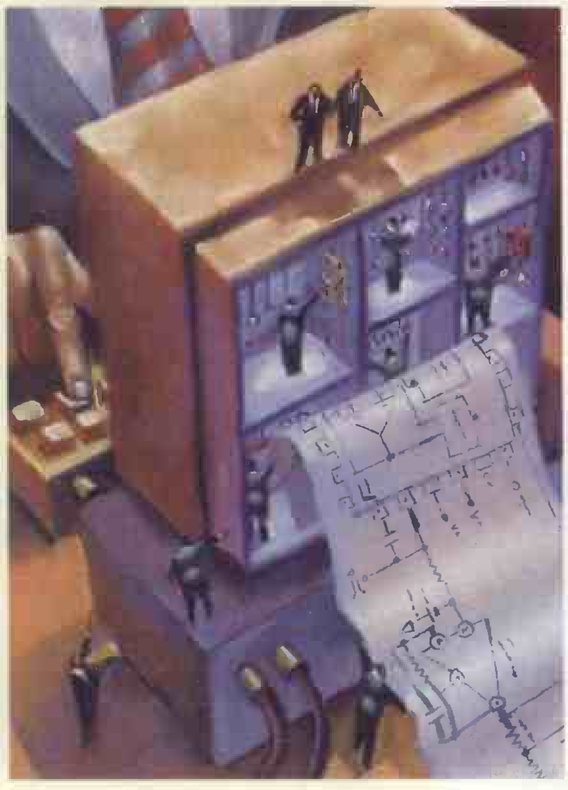
In samples with a very thin CdSe layer (one or two monolayers) significant ppv signal can be seen in addition to the CdSe luminescence. At lower voltages, the red or yellow CdSe emission predominates. But as the voltage increases, ppv luminescence becomes more intense. The researchers report that the phenomenon is reproducible over many scans, though if the bias is maintained for more than ten minutes, the ppv signal begins to fall off compared with the CdSe.

The Berkeley scientists are not completely sure of the explanation for the effect, but speculate that if the forms of the field dependencies of either charge migration or exciton (electron/hole pair) diffusion are different in two materials, then the recombination zone could move with voltage. Another possibility, says the team, is that the nanocrystals are behaving as typical bulk inorganic diodes, with falling emission intensities at high currents due to heating of the sample.

Electrical stability and efficiency problems still need to be overcome. But the devices do seem to offer the traditional advantages of bulk inorganic semiconductor diodes while combining the advantages of different colour emissions. By capitalising on the established advantages of organic polymers such as efficient hole transport and high breakdown voltages, the heterostructures should easily make larger areas.

Leds with voltage-dependent colour could form the basis of future flat-panel displays. At low voltages, CdSe predominates, in this case producing a yellow colour. At higher voltages, green luminescence of the ppv can be seen.





Circuits by design

Now that simulating and analysing circuitry by computer no longer needs the specialist skills it used to, engineers can spend more time on designing and less time on programming. Owen Bishop looks at the steps involved in taking a thermometer from concept to final design.

One of the best ways to learn about computer-aided circuit design and its benefits is to look at a specific example. The following charts the design of a small, hand-held thermometer incorporating a thermistor for sensing.

Over the range 0 to 100°C, the unit needs to be capable of reading temperature to the nearest degree. Whether or not a bridge is the best way of obtaining a readout is a matter of opinion, but a purely resistive circuit is most appropriate for an introduction to the software.

In the initial schematic, **Fig. 1**, an ntc thermistor reading 47kΩ at 25°C forms one arm of a conventional bridge. The thermistor is mounted on the case of the instrument, so its leads are short and there is no need for a compensating lead in the bridge.

Bridge excitement is carried out by a stable 2.5V dc source, the exact nature of which is irrelevant. The meter is a readily-available and inexpensive type with a coil resistance 650Ω, centre-zero indication, and a full scale deflection ±125μA.

Preset resistor R_{bal} is for balancing the bridge, so as to make the response symmetrical, and preset R_{scale} allows adjustment for full deflection at the ends of the range. To demonstrate how the circuit operates and how its design can be refined, I have chosen a circuit simulation package called *SpiceAge for Windows*.

Spiceage for Windows

Launched in 1992 in response to the rapidly growing popularity of the PC's GUI, *Spiceage for Windows I* was an extension of a GEM version of the package adpoted by the OU and in other educational establishments world wide.

Now in its fourth version *Spiceage for Windows* is capable of handling up to 60 nodes. The most recent version, just announced, has had noise and reflection coefficient analyses added to its list of functions.

Behind Spice

Spice is an acronym for simulation program with integrated circuit emphasis. The original Spice program was developed at the University of California in Berkeley in the late 1960s. It has been through several stages of development since then and a number of variants, including *SpiceAge* have been produced by various authors.

All versions of Spice require the circuit to be presented as a netlist – a list of the components, their characteristics and their interconnections. The circuit can then be subjected to a number of different forms of analysis with results expressed numerically or graphically.

Versions of Spice differ in the syntax of the netlist and in the range of analyses that can be performed. As might be expected, the recent versions such as *SpiceAge* simulate a wider range of components than the original Spice and allow more kinds of analysis to be undertaken, and with greater precision. Although the input to these programs is essentially a netlist, associated software is available for several versions of Spice – including SpiceAge – for generating the netlist from a schematic diagram.

Although Spice was developed specifically for simulating the action of integrated circuits, it is able to handle the full range of electronic components from resistors and capacitors to triacs and timers.

Netlist

When simulating, the first task is to compose the netlist. On schematic Fig. 1, number or name the nodes of the circuit. Names usually make the tables and graphs easier to interpret, but sometimes it is quicker to refer to numbered nodes.

In Fig. 1, two of the nodes have names – 'pos' for the positive rail and 'gnd' for the 0V rail) – and the rest have numbers. Key the netlist, **Table 1**, into the computer after clicking on File and then New. The netlist has an explanatory title, preceded by an asterisk so that it is not taken by the computer to be part of the netlist. Next come the statements of the netlist, which may be entered in any order. The basic format for a statement is:

Component	Component	Connection	Connection	Value
type	name	p1	p2	

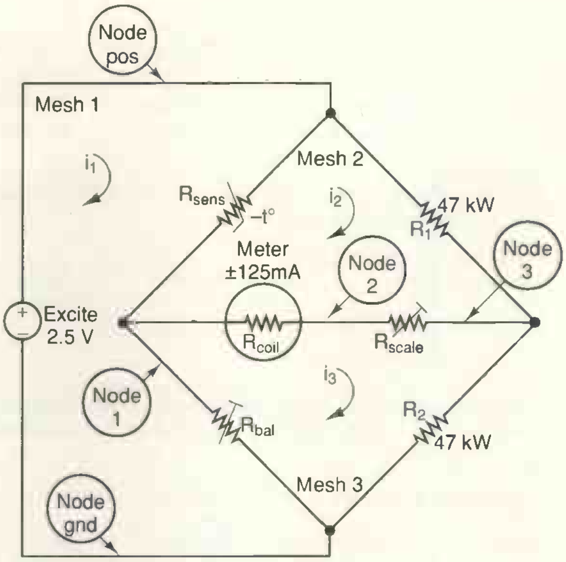


Fig. 1. Temperature measurement bridge used to illustrate the workings of Spice. Because the thermistor leads are kept short, there is no need for bridge compensation.

Component type is specified by one or more capital letters. In this netlist, '*V*' represents a voltage generator, and '*R*' a resistor. The component name is used to distinguish between components of the same type, such as 'R_{sens}' for the sensing thermistor, and 'R_{bal}' for the balancing preset. These names are helpful when interpreting graphs and tables, but meaningless to the analysing program. To the program, a resistor is just a resistor and, unless specified otherwise, all resistors are treated in the same way, no matter what their names.

Pins of a two-terminal device such as a resistor are identified by p_1 and p_2. We use these numbers to specify to which nodes the two terminals are connected. With non-polar devices, it does not matter which pin is p_1 and which is p_2. In the analyses, conventional current is taken to flow through the resistor from pin 1 to pin 2. A current flowing from pin 2 to pin 1 is represented by a negative value.

Components such as transistors have three connections in the netlist, and integrated circuits may have more. The final entry in the statement is specified by typing 'v=' followed by the numerical value. It is not necessary to specify units; the program knows that resistors are valued in ohms and voltage generators in volts. But the software recognises symbols for multiples and sub-multiples, such as 'k' for 'kilo', 'M' for 'mega', 'm' for 'milli' and 'u' for 'micro'.

The second statement of the netlist specifies a resistor – the thermistor – named R_{sens}. Its pin 1 connects to the 'pos' node, pin 2 connects to node 1, and its value is 42.890kΩ. This non-standard value is the resistance of the thermistor at 27°C, which is taken as the usual operating temperature for circuits simulated by *SpiceAge*. This value is calculated from the customary formula, as described in the panel.

Resistance of the thermistor at 0°C is 165174Ω and at 100°C it is 2975.65Ω – values needed later. Tolerance of inexpensive thermistors is only ±5%, so most of the figures in these values are not significant. But there is no harm in keeping them for the present and discarding them at the end of the analysis.

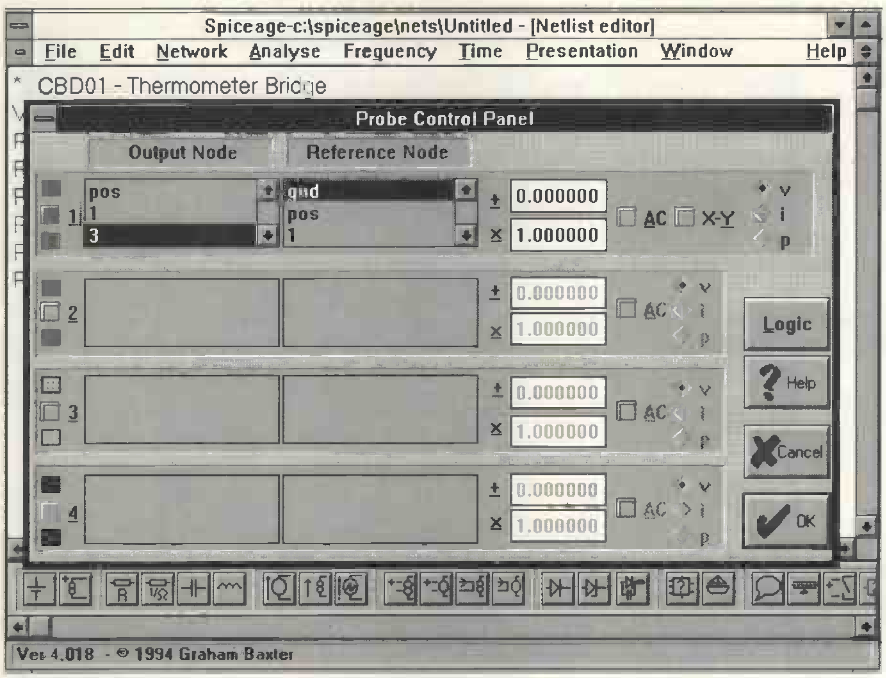


Fig. 2. Up to four probes can be attached by keying in the details in a dialogue box, called up by clicking on Time and then on Probes.

Table 1. In simulation, the first stage is to compile a netlist describing values of the various components in the circuit and how they interconnect.

CBD1 – Thermometer Bridge				
V Excite	**–out:gnd**	**+out:pos**	**Ex=None**	**Of=2.500000**
R Rsens	p1:pos	p2:1	v=42.8900k	
R Rbal	p1:1	p2:gnd	v=42.8900k	
R R1	p1:pos	p2:3	v=47.0000k	
R R2	p1:3	p2:gnd	v=47.0000k	
R Rcoil	p1:1	p2:2	v=650.0000	
R Rscale	p1:2	p2:3	v=100.0000	

Two of the resistors are presets; for these, type in any value between a very small one such as 0.001Ω and the full-scale value; the program does not accept zero resistance values. Here we choose 42.890kΩ for R_{bal} so as to begin with a balanced bridge, and 100Ω for R_{scale} – a convenient setting to start with.

Resistance of a thermistor

At any given temperature *T*, resistance *R* is given by:

$$R=R_{ref}e^{\beta(1/T-1/Tref)}$$

where R_{ref} is its resistance at T_{ref}, and temperatures are in kelvin.

The first statement of the netlist specifies the voltage generator, called 'Excite', with its negative output connected to ground and its positive output to 'pos'. Use '–out' and '+out' instead of 'p_1' and 'p_2' as pin designations for voltage and current sources. Since the voltage is a constant dc value, characterise it by stating that 'Ex=none', interpreted as 'no excitation', i.e. no waveform such as 'sine' or 'ramp'.

Constant dc voltage is provided by setting the offset to 2.5V.

Analysis

An analysis of dc quiescent voltages is obtained after 'attaching' probes to the circuit. Up to four probes can be attached by keying in the details in a dialogue box, called up by clicking on Time and then on Probes, **Fig. 2.** Attach Probe 1 to node 3, with its reference terminal connected to 'Gnd'.

Clicking on Analysis followed by DC quiescent, obtains **Table 2**. This shows the voltage at each node, with reference to 'Gnd'; the results are as expected in a balanced bridge. Node 'pos' is at 2.5V, nodes 1 and 3 are at half that voltage; node 2 is also at 1.25V because no current is flowing through the meter. An analysis of currents is rather more to the point, and is obtained by setting Probe 1 to measure current through R_{coil}. The table of results **Table 3** shows the currents; there is no current through R_{coil} and R_{scale}, because the bridge is balanced. All other currents are small, of the order of tens of microamps.

Much of the design of the bridge has already been done – this is widely used circuit. Use the simulation to find suitable full-scale values for R_{bal} and R_{scale}. The first step is to put the bridge out of balance by editing the netlist so that R_{sens} takes the value 2.976kΩ, its resistance at 0°C. Repeat the current analysis, which takes only a fraction of a second; it shows a current of –12.6μA through R_{coil}. Edit the netlist again, to make R_{sens} equal to 165.174kΩ, its resistance at 100°C; the current becomes +40.24μA.

At this stage it is apparent that the current for a full-scale deflection of 125μA is unobtainable, even if R_{scale} is reduced to zero. To increase currents in the circuit, increase the offset of the voltage generator to 10V by editing the netlist. This is far easier than soldering in a different bandgap voltage reference.

These alterations have the desired effect but the response of the meter is still not symmetrical; it does not swing equally in the negative and positive directions when R_{sens} is given its 0°C and 100°C values.

The next step is to adjust R_{bal} by keying in a series of different values and checking the magnitude and direction of the current through R_{coil}. This is the equivalent of adjusting the preset and observing the behaviour of the meter. You will soon home on a value of 13kΩ, giving ±117μA through R_{coil}.

Symmetry has been obtained but full-scale deflection is still not possible – even with R_{scale} reduced to 1Ω. Increasing the voltage to 12V, and subsequently increasing R_{scale} to 4.2kΩ gives a full-scale output, but this throws the response slightly out of symmetry. Increasing R_{bal} to 13.6kΩ produces a response of ±125μA almost exactly. This process of adjusting one preset, then the other, and gradually aligning the simulated circuit is analogous to similar operations performed on a real circuit with a screwdriver. The netlist now looks as in **Table 4**.

Using a simulator provides a valuable insight into the workings of the circuit, with practical implications. For example, when you try to make the scale response symmetrical, it you will find that altering R_{bal} by only 0.01kΩ makes an appreciable difference to the sym-

Table 2. Direct-current quiescent analysis of the thermometer shows the results you would expect from a balanced bridge.

Ref Node gnd [27.00C]

Node name	Volts	Node name	Volts	Node name	Volts
pos	2.5000000	1	1.2500000	3	1.2500000
2	1.2500000				

Table 3. Analysis of curents through the various elements of the thermometer bridge. Because the bridge is balanced, there is no current through R_{coil} or R_{scale}.

[27.00C]

Component name	amps	Component name	amps	Component name	amps
VExcite	55.74007u	R Rsens	29.14432u	R Rbal	29.14432u
R R1	26.59574u	R R2	26.59574u	R Rcoil	0.0000000
R Rscale	0.0000000				

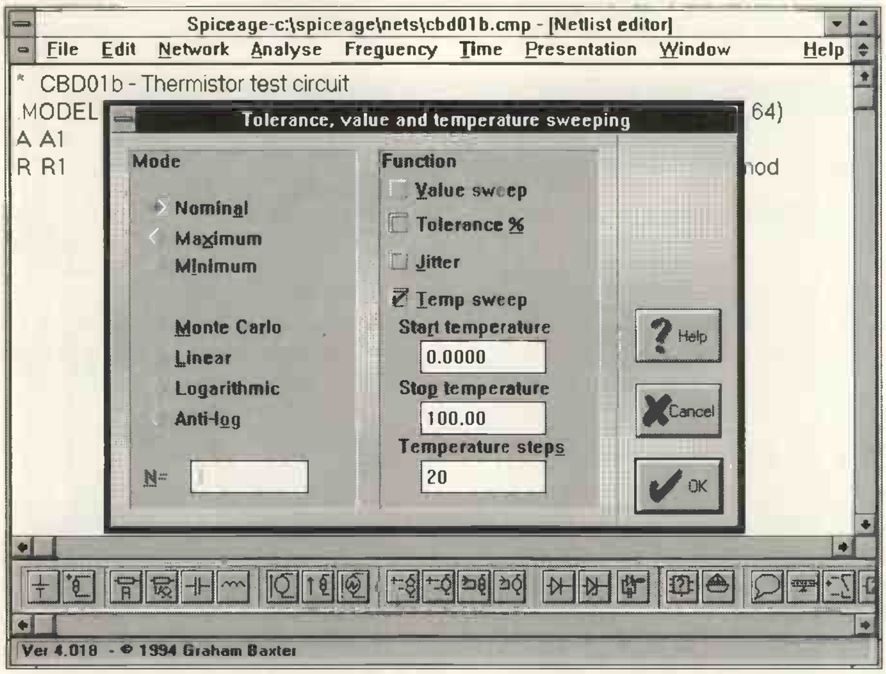


Fig. 3. Via the Analyse menu, it is possible to sweep the temperature of the circuit over a wide range and in incremental resolutions.

A complementary approach

A circuit simulator has many facilities for examining and analysing circuit behaviour. In essence, it simulates actual components joined together with real wires and powered by real voltage or current sources.

Building the circuit involves typing in the netlist and testing it involves attaching probes to it at key points. The simulator comprises the functions of soldering iron, component box, breadboard, power pack, signal generator, multimeter, oscilloscope, frequency meter, spectrum analyser and other instruments.

Overall, the approach is a practical one, a simulation of operations on the workbench. This appeals to many engineers who may, to begin with, regard cad with suspicion. Those of a more theoretical turn of mind may well prefer to use a more purely analytical approach. They may rather set down some equations and work on them mathematically.

But the problem is that, even for a simple circuit of only half a dozen components, the maths rapidly gets out of hand. This is where the computer helps, for computers are good at maths. They work very fast and they make no mistakes.

The difference between the two approaches is illustrated by tackling the thermistor circuit design problem, using the *Mathematica* package. This software is described by its author as 'A system for doing maths on a computer'. It is applicable to any field of study in which maths plays a significant part.

metry. This indicates that R_{bal} is best realised as a fixed 13kΩ resistor in series with a 1kΩ, 22-turn preset. Scale adjustment is not as critical; a 4.7kΩ preset is satisfactory for R_{scale}.

Before leaving the circuit, set Probe 1 to read power dissipation and repeat the analysis. In the thermistor, power dissipation is only 550μW, so the self-heating effect may be disregarded. Most power dissipation is in R_1, but is only 1.7mW so low wattage resistors may be used throughout.

Mesh analysis

In the complementary approach, consider Fig. 1 as a network with three meshes. Using Kirchhoff's Voltage Law, write the three mesh equations, following the standard procedure:

Mesh 1: $(R_b+165174)i_1-165174i_2-R_bi_3=12$
Mesh 2: $-165174i_1+(212824+R_s)i_2-(650+R_s)i_3=0$
Mesh 3: $-R_bi_1-(650+R_s)i_2+(47650+R_b+R_s)=0$

These are the equations at 0°C, when the resistance of the thermistor is 165.174kΩ. Resistance R_{bal} is abbreviated to to R_b, and R_{scale} to R_s. The set of equations is similar at 100°C, when the resistor of the thermistor is 2.976kΩ, but using i_4, i_5, and i_6 to represent the currents:

Mesh 1: $(R_b+2976)i_4-2976i_5-R_bi_6=12$
Mesh 2: $-2975i_3+(50626+R_s)i_5-(650+R_s)i_6=0$
Mesh 3: $-R_bi_4-(650+R_s)i_5+(47650+R_b+R_s)=0$

There are two more equations to write, representing the deflections of the meter at 0°C and 100°C. At 0°C, the meter reads –125μA:

$i_2-i_3=0.000125$
Similarly, at 100°C:
$i_6-i_5=0.00125$

There are now eight equations and eight unknowns – the six currents and the two resistances – so it is simply a matter of solving the eight simultaneous equations. Perhaps it is not so simple in practice, as an 8th-order determinant needs to be evaluated. This is where *Mathematica* comes to the rescue.

In **Table 5**, ln(2): = is followed by the input command, in *Mathematica* syntax. The command 'NSolve' instructs the computer to solve the following equations and produce numerical solutions. The first pair of curly brackets contains the eight equations.

Subscripts are placed in square brackets, for example r[b] for R_{bal}, i[3] for i_3. The asterisk represents 'multiply', though it is permissible to leave a space instead. The '==' is *Mathematica*'s symbol for equality. The second pair of curly brackets encloses a list of the variables to be evaluated.

A few seconds after Out(2):=, the result of the calculation is displayed. The first four statements list solutions for the currents and resistances. These are very close to the values obtained using *SpiceAge* – but not identical. This is because adjusting of the presets ceased once the result was within the required goal.

Return to *SpiceAge* and amend R_{bal} and R_{scale} to the values found with *Mathematica*; the currents displayed are the same as those in Table 3 to several significant figures.

The last four statements of Table 5 list another set of solutions. For this set, R_{bal} has to have a negative resistance of 13.5kΩ, which is not a practicable proposition. This emphasises that *Mathematica* is solving maths equations, not working with a simulated circuit. Always bear in mind the practical applicability – or otherwise – of the maths.

Non-linearity

The response of a thermistor is decidedly non-linear, especially over a range as wide as 0°C to 100°C. Return to SpiceAge to investigate this aspect of the circuit. The technique is to specify the way in which the resistance of Rsens changes with temperature. For ordinary resistors in the netlist, state the temperature coefficient. For example, the statement 'Te=0.00005' indicates a temperature coeffi-

Table 4. After adjustments to the software – the equivalent of tweaking with a screwdriver – the netlist looks like this.

*CBD1a – Thermometer Bridge, final version

V Excite	–out:gnd	+out:pos	Ex=None	Of=12.00000
R Rsens	p1:pos	p2:1	v=165.174k	
R Rbal	p1:1	p2:gnd	v=13.6000k	
R R1	p1:pos	p2:3	v=47.0000k	
R R2	p1:3	p2:gnd	v=47.0000k	
R Rcoil	p1:1	p2:2	v=650.0000	
R Rscale	p1:2	p2:3	v=4.20000k	

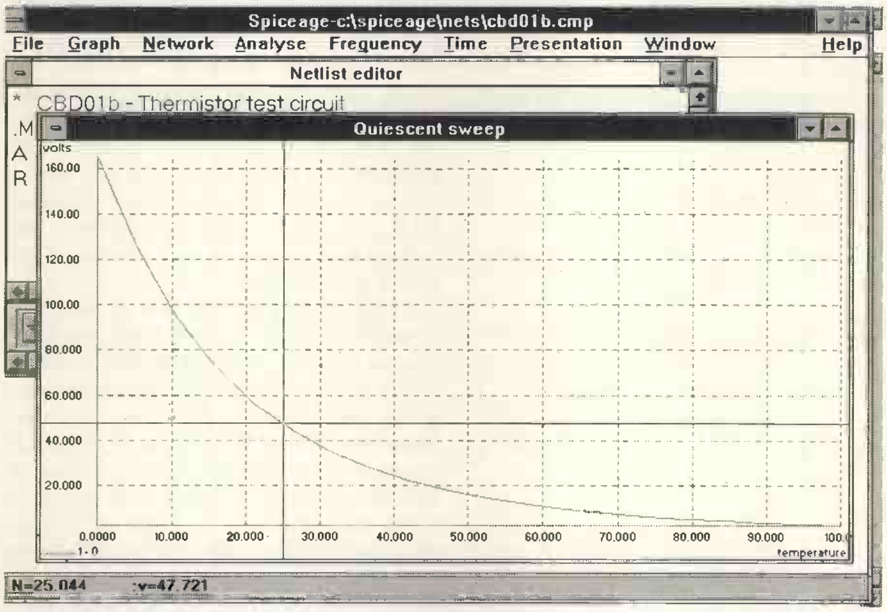


Fig. 4. Once the operating temperature sweep has been carried out, results can be displayed in graphical form.

Cdb01

```
In[2]:=
   NSolve[{(r[bl+165174)*i[1]-165174*i[2]-r[b]*i[3]==12,
   -165174*i[1]+(212824+r[s])*i[2]-(650+r[s])*i[3]==0,
   -r[b]*i[1]-(650+r[s])*i[2]+(47650+r[s]+r[b] )*i[3]==0,
   (r[b]+2975)*i[4]-2975*i[5]-r[b]*i[6]== 12,
   -2976*i[4]+(50626+r[s])*i[5]-(650+r[s])*i[6]==0,
   -r[b]*i[4]-(650+r[s])*i[5]+(47650+r[s]+r[b])*i[6]==0,
   i[2]-i[3]==0.000125, i[6]-i[5]==0.000125}, {r[b], r[s],i[:
   i[2],i[3],i[4],i[5],i[6]}]
Out[2]=
 {{i[1] -> 0.000247858, i[4] -> 0.000895195,
   i[2] -> 0.00019016, i[5] -> 0.0000651684,
   i[3] -> 0.0000651596, i[6] -> 0.000190168,
   r[s] -> 4092.04, r[b] -> 13518.2},
  {i[1] -> 0.000280428, i[4] -> -0.00091281,
   i[2] -> 0.00019016, i[5] -> 0.0000651492,
   i[3] -> 0.0000651596, i[6] -> 0.000190149,
   r[s] -> 47129.3, r[b] -> -13517.7}}
```

Table 5. Solving an eighth-order determinant is made easy by SpiceAge's assistant – Mathematica.

Table 6. In this test-circuit representation, a constant-current of 1mA connects directly across resistor R_1.

```
*CBD1b – Thermometer test circuit
MODEL   resmod   res      (TC1=0 TC2=0.00015 TCE=-4.64)
A A1    -out:0   +out:1   v=1.00000m
R R1    p1:p     p2:0     v=42.8900k        Mo=resmod
```

Table 7. Netlist with the separately evaluated thermistor model added.

```
*CBD1c – Thermometer Bridge with model thermistor
V Excite   –output:gnd +output:pos   Ex=None     Of=2.500000
R Rsens    p1:pos    p2:1      v=42.8900k        Mo=resmod
R Rbal     p1:1      p2:gnd    v=13.6000k
R R1       p1:pos    p2:3      v=47.0000k
R R2       p1:3      p2:gnd    v=47.0000k
R Rcoil    p1:1      p2:2      v=650.0000
R Rscale   p1:2      p2:3      v=4.20000k
.MODEL     resmod    res       (TC1=0 TC2=0.00015 TCE=-4.64)
```

cient of 50ppm/°C. Assume that tempco can be ignored for all resistors except Rsens in the present example. For Rsens, the relationship between temperature and resistance is not linear. Substituting the calculated resistance at 27°C into the formula (see box), making 27°C the reference temperature, and assuming that b has the data sheet value of 4090, the relationship is:

$$r=0.05146e^{4090/(t+273)}$$

SpiceAge allows for an exponential response of the form:

$$r=R[1+TC1(t-t_{27})+TC2(t-t_{27})]e^{TCE(t-t_{27})/100}$$

where R is the nominal resistance at 27°C, or other reference temperature, and TC1, TC2 and TCE are coefficients to be specified by the user.

Models

Special types of components are describable in Spice as models. Transistors and other devices with complicated patterns of behaviour are usually described in this way. In the first statement in **Table 6**, the description for the model of the thermistor is set out as a separate statement, beginning '.MODEL'. This has the model name 'resmod'. Suitable values are entered for TC1, TC2 and TCE. The test circuit of Table 6 shows a constant current source ('A') of 1mA connected directly across resistor R_1, which has the nominal resistance of R_{sens} at 27°C. This netlist models the behaviour of the thermistor according to the second equation for r.

Setting the program the task of modelling thermistors, which all have a peculiar temperature response, is somewhat unfair. Earlier versions of Spice do not even include an exponential coefficient. In particular, the exponential coefficient (TCE) in the second equation for r requires the temperature variable to be in the numerator of the index, whereas it is in the denominator of the index in the first equation. In spite of this, it is not difficult to find a set of coefficients that model the thermistor with reasonable precision.

There appear to be no easy mathematical routines for fitting the coefficients to the known response of the thermistor. The best approach is by trial and error, using the test circuit.

Using the first of the equations for r, calculate the resistance at, say, ten temperatures between 0°C and 100°C. The task is to fit the model to these values. Begin by making TC1 zero, give TC2 a very small positive value of about 0.0002, and give TCE a small negative value of about –4.

Next, set up a temperature sweep on the test circuit. From the Analyse menu, click on Tolerance & Temperature... This brings up a dialogue box, **Fig. 3**, in which you select Nominal mode, using the nominal value of the resistor, taken from the netlist), and key in the details of the temperature sweep. Select 0 as the starting temperature, 100 as the stop temperature, and 20 temperature steps.

Returning to Analyse, click on Quiescent Sweep. The graph displayed, which will be similar to **Fig. 4**, shows how the voltage across the model varies with temperature. With a current of 1mA, the corresponding resistance is 1kΩ/V; read the scale on the y-axis as kΩ.

Compare the graph obtained with the calculated values. This is easily done using the cross-hair cursor and reading the values from the panel at the bottom of the screen. Next edit the netlist to correct any discrepancies and repeat the analysis. Gradually the quiescent sweep gives a graph approximating very closely to the calculated values. Fig. 4 shows the curve finally obtained. The curve indicates that the resistance of the model departs less than 2% from the actual value over the range 0°C to 50°C, and is reasonably close to it above this range.

Testing linearity

Edit the netlist, Table 4, to replace R_{sens} with a thermistor model by adding the model description and by adding Mo=resmod to the specification of R_{sens}, **Table 7**. With probe 1 set to monitor current through Rcoil, running a temperature sweep as described above gives **Fig. 5**. As it happens, response is very close to linear and the meter needle lies at zero when the temperature is about 54°. At 0°, the current is almost exactly –125μA, as required. It slightly exceeds 125μA at 100°, which is due to the departure of the model from the ideal at the upper end of the scale. However, we now know enough to be able to build the thermometer with the confidence that it will work. Any slight discrepancies at the ends of the scale can be corrected by adjusting the real R_{bal} and R_{scale}. For low precision readings, with an accuracy of about 5%, we could use an evenly divided scale, but a slightly non-linear scale is to be preferred. ■

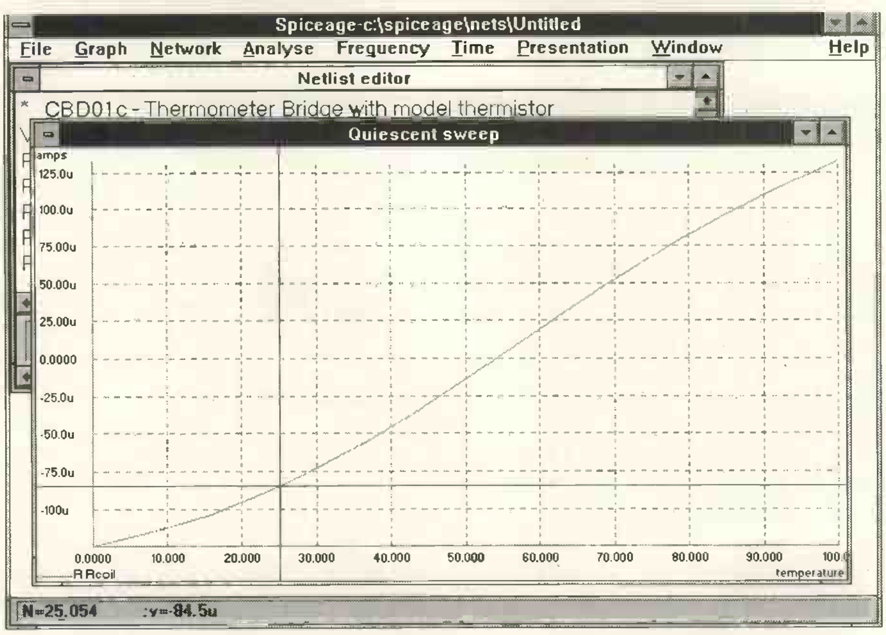


Fig. 5. Sweeping over the operating temperature range indicates that the thermometer is reasonably linear.

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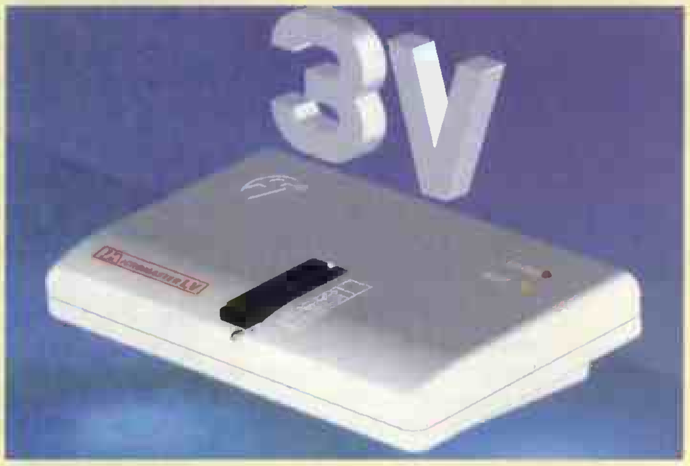

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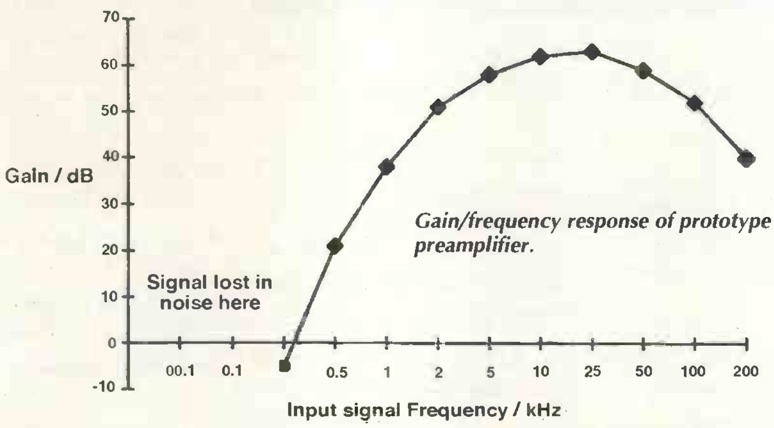
Micropower infra-red receiver

Unable to find a high performance infra-red receiver at the right price, Simon Bateson set about developing this circuit. Designed to provide long battery life, it consumes just 60µA at 4.5V.

Photodiodes are often required to detect relatively low-powered pulses in the presence of intense continuous sunlight or 100Hz illumination. This was described recently in *EW+WW* May 1994 on page 367. Loading the diode with a gyrator is better than using a resistor but there are still two problems. Firstly; although the gyrator exhibits a high ac impedance, there is no actual power gain from the transistor. Secondly, the considerable photocurrent induced by sunlight, of the order of several milliamps, must come from the power supply. Such a circuit is unsuitable for low power battery operation.

The photodiode load – a gyrating common-base circuit?

The circuit shows that the photodiode operates in photovoltaic mode into a 1kΩ resistive load. Transistor Q_1 saturates at a very low current and the collector sits around 0.3V. To signals with a fast rise time, such as those generated by remote control transmitters, Q_1 emitter appears as a low impedance and the photodiode current is diverted into it. Thus, Q_1 acts as a common-base amplifier and the collector current cuts off rapidly. Even at low collector currents the response is fast because of the lack of Miller and storage effects.



Gain/frequency response of prototype preamplifier.

Amplifier/comparator system

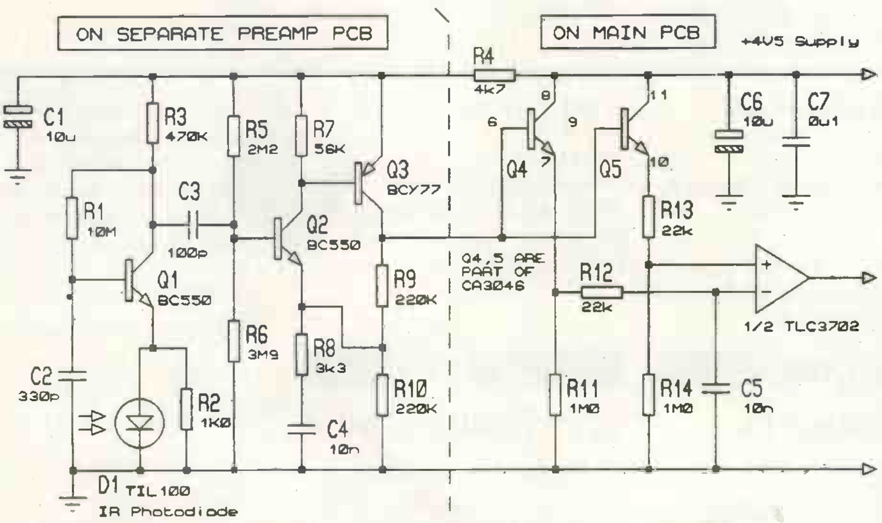
The amplifier is a series-feedback pair operated at a low current, adapted from *The Art of Electronics* by Horowitz & Hill. It has a good gain-bandwidth product considering the power consumption, but the usual narrow positive-going output spikes can 'hang up' and develop long tails when the circuit is heavily overloaded. This happens when the infra-red transmitter is brought in contact with the photodiode. To overcome this, Q_4 rectifies the amplifier output to provide a variable threshold for the LinCMOS micropower comparator. Q_5 acts as a buffer, matched and thermally linked to Q_4 to avoid degrading the comparator's offset specification.

Using typical keyring-type transmitters with *MC145028* type decoders, reception is reliable at all distances between direct contact and about eight metres. It can also be greatly improved by using a simple lens.

With a standard *TIL38* emitter pulsed at 1A (50µs on and 25ms off time) at a distance of 25 metres in daylight, the output at *Q3* collector reached 140mV peak against a background noise of 3.6mV rms (1MHz noise measuring bandwidth). A considerable data transmission range is possible but would depend on individual circumstances.

Gain/frequency response of the prototype preamplifier is shown here. For high speed data transmission, a better frequency response can be achieved at the expense of current consumption by scaling components appropriately. With this level of gain, layout is important and the preamp must be built on its own pcb. The supply decoupling resistor should be on the main board, otherwise power supply noise can couple into the high impedance output.

The entire circuit consumes less than 60µA at 4.5V, including 22µA for the spare comparator in the *TLC3702*. Reception is barely affected by sunlight, direct fluorescent or incandescent light. n



Low-power, infra-red preamplifier. For higher supply voltages, adjust the bias chain to produce 1.6V at Q_2 base.

Speed of the receiver is restricted if pcb layout is poor. Readers wanting a photocopy of the author's prototype layout should send an sae marked 'IR receiver' to EW+WW's editorial offices.

A new era in *magnetics*

Engineering magnetic films to only a few atoms deep can have profound effects on the magnetism. John Gregg and Sarah Thompson examine these effects, and how we can tailor them for practical use.

Magnetism is a familiar, everyday concept that provides us with magnetic bearings, computer disk technology and audio tapes, among other necessities of modern life. So what remains to be discovered or explored in this seemingly very mature and well-understood scientific field?

The answer lies in the technological ability to make very small magnetic systems that paradoxically show spectacularly different physical properties to those of everyday materials.

These small magnetic systems are new materials which nature never intended us to see. So how are they made? One possibility is to squeeze one spatial dimension by making very thin films of a magnetic material. The term 'very thin' needs some qualifying here: magnetic tapes and disks already use thin films, but these are thousands of atoms thick and behave very much like ordinary magnets. What we mean by thin in this instance are films just a few atoms deep, produced using state-of-the-art high vacuum technologies such as sputtering and molecular beam epitaxy, as outlined in the first and second panels.

These films are quasi two-dimensional in behaviour and exhibit new physical properties that reflect this reduced dimensionality.

More interesting still is the idea of multi layer sandwiches built up of many such layers – engineering the magnetism on a quantum scale. By careful choice of magnetic metals and geometry of the device, unusual magnetic spin configurations can be designed in which the spins associated with the atoms point in different directions in different magnetic layers, or even in the same layer. In some extreme cases the magnetic moments can even be made to adopt modulated spiral structures along the film's growth direction.

The way in which a sample magnetises is sensitive to the symmetry of its crystal structure. If you take a single crystal and apply a magnetic field along different directions, it will be easier to magnetise the sample for some orientations than for others. This preference for the magnetisation to lie in certain orientations is called the anisotropy, and is dependent on factors such as the strain on the crystal, or on macroscopic considerations such as the shape of the sample. In the particular case of thin films, the shape anisotropy tries to make the magnetism lie in the plane of the film to avoid the magnetic flux lines from

Magneto-optic recording

The greatest density of magnetic information can be stored when the magnetic moments lie perpendicular to the plane of the film. Thin film technology provided the ability to create such materials, but a new solution was required to read and write the information at such high density. The result is a combination of optical and magnetic effects using lasers.

Old information is first erased by shining a high intensity laser beam onto the magnetic bit that is to be erased. Locally, the bit is heated up above its Curie temperature (the temperature at which the material becomes non-magnetic). The film is then placed under a small coil which can be magnetised either up or down, and as the region cools and becomes magnetic again it orients itself in the direction of the field in the coil. The magnetic bit has been written.

A laser is also used to read the information, but this time it is lower powered and so does not heat and destroy the magnetism. The polarisation direction of linearly polarised light is rotated by reflection from magnetic material by the so-called Kerr effect. So the polarisation rotation of the reflected laser beam is analysed and used to determine the magnetic information that is written on the disk. The magnetic bit has now been read. Although this process allows information to be stored at very great density, the complexity of the read/write process means that it is more suited to data storage and retrieval than to fast access hard disks.

* Sarah Thompson is an EPSRC Research Fellow at the University of York, and John Gregg is a Royal Society Research Fellow at the University of Oxford.

Molecular beam epitaxy (mbe)

MBE is the state-of-the-art technique for making thin film crystals one atomic layer at a time. It was developed to its present high level of sophistication over the past two decades in response to demand from the semiconductor industry. Now its immense power and versatility is being exploited to make new magnetic materials by selectively layering different elements and encouraging the atoms to arrange themselves in a different way to that in which nature intended.

The technique works by evaporating the film material (in our case these are magnetic metals) very slowly (roughly two atomic layers per minute). This gives the atoms plenty of time once they have 'landed' on the growing surface to wander round and find their correct 'crystalline home' before they get buried by the next atomic layer.

Because the deposition rate is so slow, bombardment of the growing surface by impurity gases in the evaporation chamber can cause serious levels of impurity in the films. So to overcome this, the process must be conducted in an ultra-high vacuum (uhv) at background pressures so low that a freshly prepared and atomically clean surface would, if left to itself, stay clean for about twenty minutes or longer.

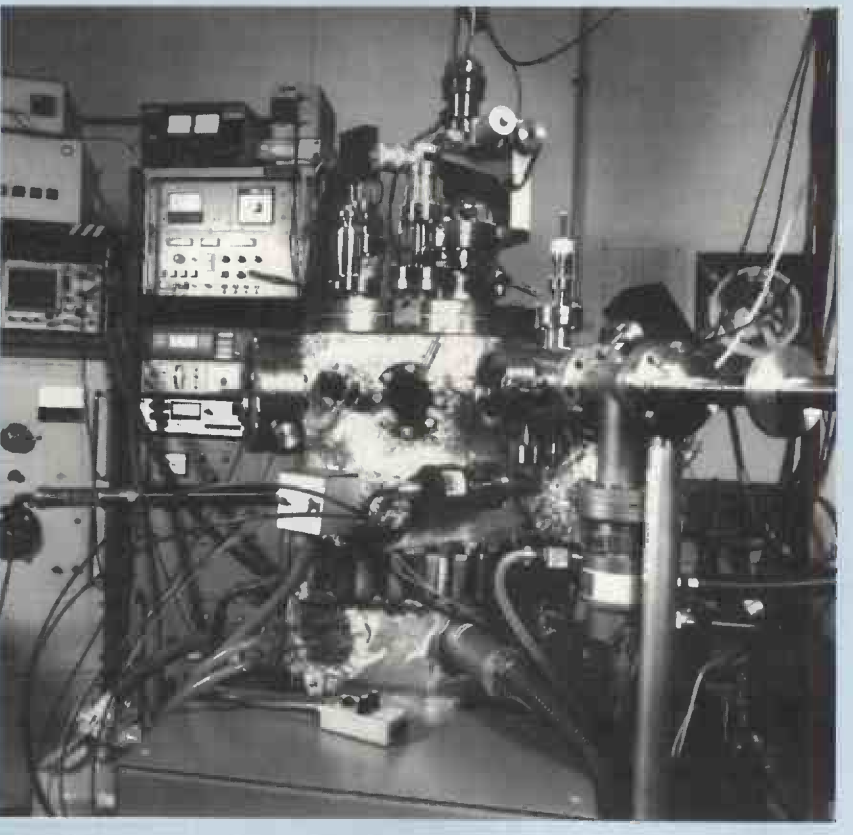
Low melting point metals are evaporated from Knudsen cells (a tall heated crucible) which have the advantage of excellent thermal stability and hence good control of the emerging atomic flux.

Knudsen cells have an upper working temperature limit determined by the crucible materials and are hence not suitable for evaporating all metals. Higher temperature metals are evaporated by using a focused electron beam incident on a slug of the metal. The beam is rastered to produce a molten puddle on the metal surface, the metal thus effectively acting as its own crucible so as to minimise contamination. Electron beam sources require more sophisticated control electronics to maintain stable film growth conditions.

The thickness of the growing film is monitored by presenting a quartz crystal to the flux of atoms. This quartz is as close as possible to the growing film position and is the frequency determining element in an oscillatory circuit. As atoms deposit on the quartz surface, the extra mass loading causes the crystal resonant frequency to drop in proportion to the film thickness.

In addition to monitoring the quantity of material deposited, we can also measure its crystal quality. This can be done as the film is being made by using a technique called reflection high energy electron diffraction (rheed) in which high energy electrons are bounced off the growing surface at a glancing angle and then examined using a fluorescent screen. Just as light diffracts from the surface of a compact disk to give rainbow patterns by virtue of the closely spaced fine lines on the disk, so the way in which the electrons diffract from the rows of atoms on the growing surface tells us in great detail about the atomic scale character of the film.

The picture shows a typical rheed pattern. The slight spottiness of the streaks gives us information about surface roughness, while the streak spacing tells us the spacing of the atoms.



The York molecular beam epitaxy plant (mbe) which is used to make magnetic trilayers and multilayers. Visible on the right is the sample transfer arm and loadlock used to pretreat and install the substrates without breaking vacuum in the main chamber. On the left is the fluorescent rheed screen and above is the sample manipulator.

The other parameter we need to monitor closely is the chemical composition. In particular we are especially interested to know if the interfaces are sharp or if the two materials in successive layers are chemically interdiffused since the properties of sharp and diffuse interfaces are quite different. Again, we are able to use electrons to probe this information.

Not all of the electron beam is reflected and diffracted, but some of the electrons penetrate two or three atomic layers into the surface. The disturbance they create ejects other electrons which are closely bound to atoms. We capture these and measure their energy. These so-called 'Auger' electrons have very distinct energies that are different for each element; by measuring the energies of all the ejected electrons we can determine which elements are present in the top few layers and their relative concentrations. Moreover, the electron energies characteristic of a particular element are slightly shifted depending on the state of chemical bonding of the atom, so we can even deduce the chemical environment of our atoms and hence get a very comprehensive picture of the chemical structure.

A reflection high energy electron diffraction pattern from a growing magnetic film. The pattern is obtained by bouncing a collimated electron beam off the film at grazing incidence and examining the resulting diffraction spread on a fluorescent screen. The separation and intensity of the vertical stripes provides information on the quality of epitaxy and the atomic separation. The spottiness of the stripes (intensity variations along the stripes) tells us about the surface roughness.

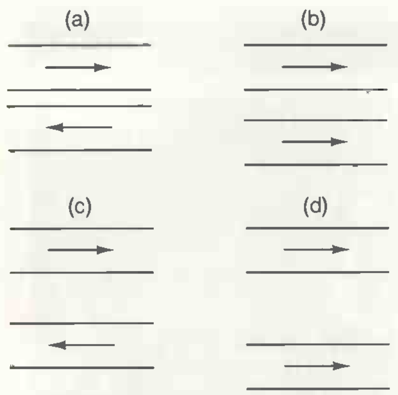


Fig. 1. Oscillatory exchange coupling: Magnetic sandwiches of cobalt, copper and cobalt in (a) to (d) are identical in construction, except for thickness of the copper layer which gradually increases from (a)-(d). The sign of the exchange coupling between the magnetic layers oscillates as a function of this thickness. It is positive for (b) and (d) which are thus ferromagnetically coupled, and it is negative for (a) and (c) which are antiferromagnetically coupled.

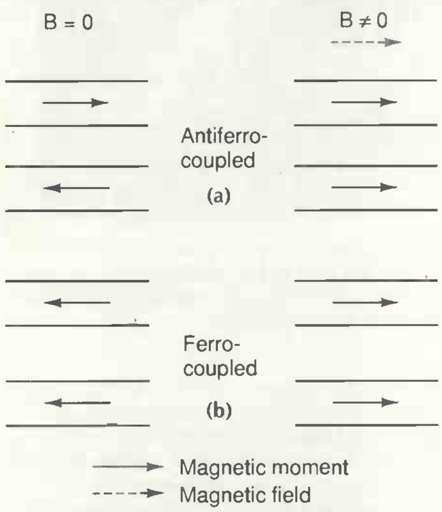


Fig. 2. Behaviour of ferromagnetically and antiferromagnetically coupled magnetic sandwiches with magnetic field applied. For both sandwiches, all magnetic moments have rotated to point along the field. However, in case (b), although both layers are turned by the field, relative orientation of the magnetic moments in the sandwich (and hence its other related physical properties such as electrical resistance) is unchanged. This shows why antiferromagnetic coupling (case a) is more useful for device applications.

spreading into the surrounding space, since this would cost magnetic energy.

Engineering the magnetism

The ability to alter chemical and structural properties on an atomic scale, to control what elements are inserted and where, and to control their crystal structure and physical shape, allows us to create entirely new materials with new magnetic properties. With some insight into the underlying physics, we can tailor these properties to our requirements.

A classic example of the potential of carefully engineered multi-layer magnetic materials is the development of the latest magneto-optic high density data storage discs (third panel). To get high information packing density, and to enable the 'read' process to work, magneto-optic storage relies on all the magnetic moments lying perpendicular to the surface – exactly the opposite to what is normally expected of a thin film since this implies the maximum amount of costly flux leakage out of the surface.

When very thin layers, no more than a few atoms thick, of certain materials such as cobalt and platinum are layered alternately, something very unusual occurs at the interface of the two materials. The interaction between these two metals introduces a new interface anisotropy which is bigger than the shape anisotropy and encourages the moments to lie perpendicular to the plane. Provided there is enough of the interface compared to the bulk of the film (i.e. very thin layers) then this new anisotropy wins and we have perpendicular magnetic moments as desired. A major triumph of the new magnetic engineering.

Layering magnetic and non magnetic materials in this way, we see interesting phenomena which result from the interactions between the magnetic layers across the non-magnetic spacers. In general we would expect the magnetic layers to act relatively independently of one another and just to respond individually to externally applied magnetic fields. In the absence of such a field they would be randomly oriented while in the presence of a strong field they would be aligned parallel to it.

However, an entirely new type of magnetic layered structure was discovered in the late 1980s in which the situation is quite different. These multilayers consist of alternate layers of a ferromagnet and a non-magnetic metal such

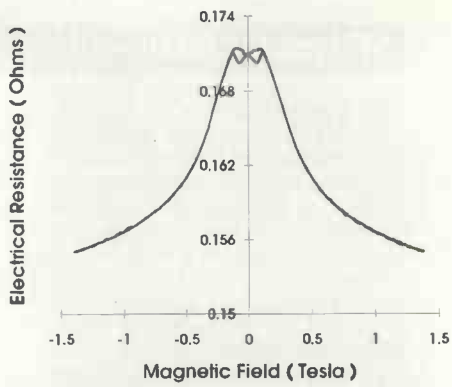


Fig. 3. Giant magnetoresistance. The thin film that produced these curves was prepared by co-sputtering cobalt and silver onto a glass substrate in the presence of rf bias. The film consists of tiny islands of magnetic cobalt in a sea of silver. Because the size of these islands is of the same order as the electron spin correlation length in metal, the electrical resistance is unusually sensitive to the application of external magnetic fields. For the example shown, resistance varies by about 10% in a field of 1 tesla. This factor is at least ten times higher than that observed in normal metals.

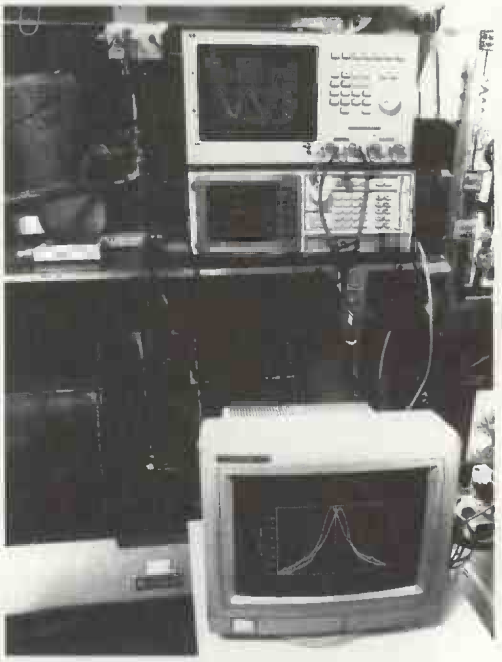


Fig. 4. The thermal conductivity of a giant magnetoresistive material is also tunable by applying a magnetic field. Here, the sample in the continuous flow liquid helium cryostat is being measured at temperatures between 4 and 400K and at fields up to 1.4T supplied by the iron cored nmr electromagnet. The measurement technique involves injecting an ac thermal wave and monitoring the thermal phase-shift across the sample with a germanium diode chip thermometer and a Stanford SR850 digital lock-in detector. The thermal resistance as a function of magnetic field is seen on the computer screen. Like the electrical resistance, the thermal resistance is high in zero field and low in high field, thus indicating that both properties of the material are determined by very similar carrier scattering processes.

The realisation of magnetic transistors

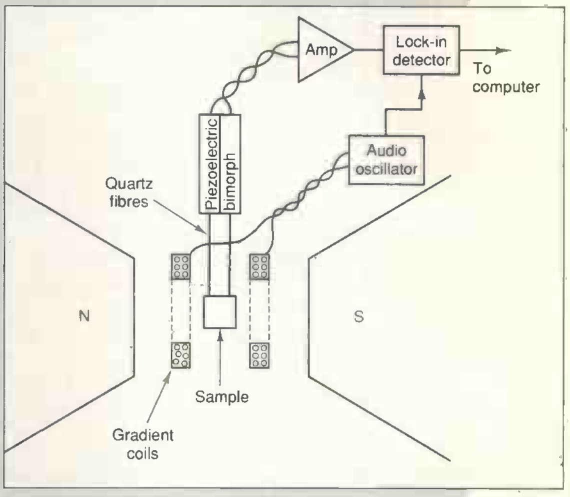
A ferromagnetic thin film may be regarded as an electron spin polariser which, when in contact with another metal and when a battery is connected, emits predominantly a current of minority electrons (ie electrons with their spins antiparallel to the magnetic moment direction in the film) in rather the same way as a piece of Polaroid passes only one polarisation of light. This then disturbs the balance of spin-up to spin-down electrons in the normal metal film. A second magnetic film connected to the other side of the normal film and through which the current also flows also acts as a spin polariser and can pass only spin up or spin down electrons depending on the orientation of its magnetisation. But since the concentrations of the two carrier types are different in the intermediate layer, the potential at which the second magnetic layer sits varies depending on which type of carrier is carrying the current, that is on whether the second magnetic layer is parallel or antiparallel to the first.

Measuring the magnetism

A magnetic film just one atom thick contains a very small number of atoms compared to most everyday magnetic objects, so how to measure its magnetism is not immediately obvious. Fortunately an ingenious instrument has been developed for the job called an alternating gradient force magnetometer, or afgm.

Its operation is simple but sensitivity is high: the sample is placed between pole-faces of an electromagnet. A magnetic field is applied to generate a magnetic moment in the sample. The interest is in seeing how this moment varies with the strength of field applied, with sample history and with time or temperature.

How is the moment measured? Superimposed on the steady field from the electromagnet is an oscillating field gradient which applies to the sample an oscillating force that is proportional to the magnetic moment. The sample is mounted on a quartz fibre attached to a piezoelectric 'bimorph'. When the sample wobbles in response to the alternating gradient, the bimorph generates an electrical signal at the same frequency which is then processed by phase sensitive detection. Size of this signal is directly proportional to the film's magnetic moment Considerable sensitivity enhancement can be obtained by making the oscillating gradient frequency coincide with a mechanical resonance of the sample assembly. A Princeton Measurement Corporation magnetometer for making such magnetic measurements exists in the University College of North Wales at Bangor under the management of Dr Kevin O'Grady.



Alternating gradient force magnetometer (Princeton Measurements Corporation). The sample sits between magnet polefaces on a quartz fibre assembly attached to a piezoelectrical bimorph which converts sample motion into an electrical signal. An ac current in the gradient coils causes the sample to oscillate in harmony with amplitude proportional to its magnetic moment. Phenomenal sensitivity of the instrument derives from lock-in detection of the electrical signal, combined with operating the instrument at a mechanical resonance of the sample assembly.

as copper. Contrary to expectation, very strong coupling was observed between adjacent magnetic layers.

Moreover the sign of this coupling was an oscillatory function of the thickness of the intervening copper layer (**Fig. 1**). For one thickness of copper (say three atoms thick), the layers want their magnetism antiparallel, while for a larger thickness (say five atoms) the magnetisations are parallel. For a further increase in thickness they are again antiparallel, and so on. So with no external field applied the magnetism of the layers is not at all random as for the case of an uncoupled magnetic stack. Instead the moments are well aligned (or antialigned).

The forces which cause this alignment are due to a quantum mechanical effect known as 'exchange interaction' which takes place between electrons in adjacent magnetic layers. The resultant coupling is thus known as 'oscillatory exchange coupling'. Its discovery opened up a wide field of research into layered systems in which the orientations of the respective magnetic layers could be controlled.

The most interesting exchange coupled magnetic sandwiches are those in which the coupling favours antiparallel alignment of the magnetic layer moments since this can then be upset by applying a large enough external magnetic field (**Fig. 2**). When the magnetism flips from antiparallel to parallel under the influence of the applied field, other properties

Giant magnetoresistance

In a multilayer the thicknesses of the individual layers are small enough to be on the same scale as the electron correlation length, so the current path will take in all the layers in a typical stack of magnetic layers, not just the top one.

The current carrying electrons have themselves a magnetic 'spin' and in a ferromagnet this interacts with the magnetism of the material. The electrons then split into two different populations, those with their spins pointing parallel to the magnetism (called the majority carriers because there are more of them) and those with spins antiparallel to the magnetic layer moment (the minority carriers). The minority carriers are heavily scattered, while the majority carriers are barely scattered at all.

If all the magnetic layers are aligned then the electrons consist of two classes: a privileged class which are majority carriers in all the layers and hence travel through the entire structure with little scattering and hence low electrical resistance; and a permanent minority class who are scattered wherever they go. The low resistance of the majority class shunts out the high resistance of the minority class, so the overall device resistance is low. If, on the other hand, adjacent magnetic layers are antiparallel or even if the layer orientations are random, none of the electrons is privileged and all spend equal amounts of time as majority and minority carriers. So the electrical resistance of the device is high, since it is dominated by the scattering which the carriers undergo in their 'minority capacity'.

Thus changing the magnetic configuration from antiparallel to parallel layers by applying a large field lowers the electrical resistance of the multilayer (**Figure 5**). In theory, changes of about a factor of two in resistance are obtainable and devices have been made which are not far off this performance. Hence giant magnetoresistance.

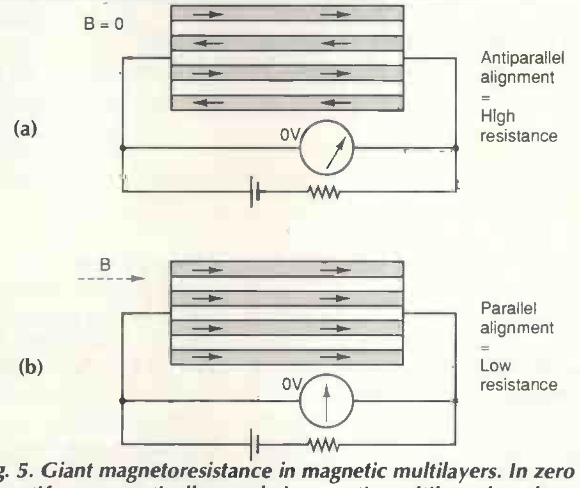


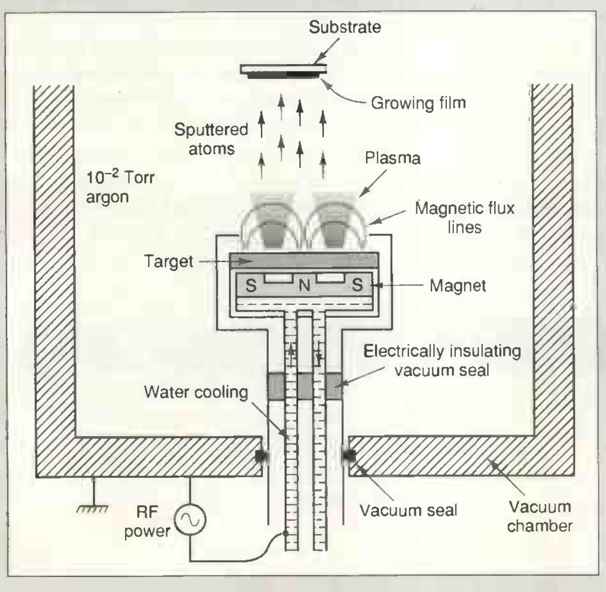
Fig. 5. Giant magnetoresistance in magnetic multilayers. In zero field, an antiferromagnetically coupled magnetic multilayer has alternate magnetic layers pointing in opposite directions (case a). All the electrons are scattered (in one layer-type or the other) irrespective of their spin polarisation, and the resistance is high. When a magnetic field is applied, the magnetic layers line up (case b). The electrons with spin antiparallel to the magnetisation are significantly scattered in all layers and are bad electrical conductors. The electrons with spin parallel to the magnetisation are hardly scattered at all and have a low electrical resistance: hence so does the whole device.

Making magnetic films by sputtering

Sputtering is a fast and effective way of making thin films where high crystalline quality is unimportant. Good single crystal thin films (such as may be grown epitaxially on single crystal substrates by mbe) are often necessary for unravelling and understanding the fundamental phenomena which determine the properties of new thin film materials, thus making it possible to 'engineer' the film structure to enhance particularly desirable features of the material's behaviour. However, for many practical purposes, such epitaxy is unnecessary and this is where sputtering comes into its own. Its speed and ability to work in cruder vacuums than mbe are not only attractive in a laboratory environment but also make it ideal for industrial manufacturing processes where high throughput is important. A typical industrial application of sputtering is in the aluminisation of compact disks.

The physics of sputtering is more complex than mbe: a disk of the metal to be sputtered (the 'target') is mounted onto a magnetron sputter-gun (so-called because its magnetic field profile closely resembles that of the rf generator of that name!). To operate the magnetron, the chamber is backfilled with a noble gas like argon to a pressure of about 10^{-5} atmospheres. Radiofrequency power is then applied to the magnetron which strikes an electrical discharge in the gas. This discharge is confined by a doughnut shaped magnetic field to the region near the target, and positive argon ions from the discharge bombard the target metal (hence its name!) and knock off metal atoms which fly across the chamber and coat onto the substrate. Because they are generated by an electrical discharge, these atoms are highly energetic compared to their thermally evaporated counterparts and so, despite the speed of deposition, they tend to give high quality microcrystalline films.

Sputtering systems need to operate at relatively good vacuum but with a high sputtering gas throughput. These conflicting vacuum requirements are best met by using a turbo-molecular pump, which is essentially a multi-stage turbine whose rotational speed is so fast that the turbine blades have a speed comparable with the speed of the gas molecules being pumped. The turbopump running the plant was provided by Leybold UK who specialise in vacuum equipment for such applications.



Cross-section through a magnetron sputter-gun. Radio frequency (or high dc voltage) is applied to the water-cooled target assembly. This strikes a discharge in the argon with which the chamber has been backfilled. The plasma is confined to the region close to the target which then gets bombarded by positive ions and releases metal atoms which fly across the chamber and coat onto the substrate to form a thin film.

of the sandwich change, and it is here that their potential for application lies.

In our thin film structures, the thicknesses of the individual layers are now comparable with the free path travelled between collisions by electrons carrying electrical current, so the behaviour of the electrical carriers is simultaneously influenced by the electronic band structure of the various metals used and also by the way in which the magnetism behaves. This magnetic behaviour has interesting knock-on effects on useful properties of the films such as optical refractive index, electrical resistance or thermal conductivity, and this is potentially exploitable to make novel magnetic devices.

As the drive towards automation and miniaturisation continues, there is an increasing demand for a wide range of sensors that are themselves in miniature. What better starting point then than a thin film? In addition, where such sensors are to be incorporated into an electronic component, they could be efficiently deposited in the same process. One can envisage many markets for such sensors, including the automobile or the aircraft industries.

The industry which is leading the way in the application of these films, though, is – once again – the magnetic recording industry. This is due to the dramatic discovery in the new exchange-coupled films of 'giant magnetoresistance'.

In most metals, the electrical resistivity increases by less than 1% when a strong magnetic field is applied. Even so, this magnetoresistance has been extensively employed to sense magnetic fields using such alloys as nickel-iron in which the effect is particularly pronounced. **Figure 3** shows how different is the behaviour of a giant-magnetoresistive material when a magnetic field is applied. The change in electrical resistivity is around ten times greater, at about 10%, and the resistance decreases with applied field. An entirely different mechanism is responsible for this effect, which is made possible by our ability to engineer the sample on a scale comparable with the electron correlation length which is a few nanometres or greater. The same novel quantum mechanical effects give rise to other strange properties. For example the thermal conductivity in a silver/cobalt film is tunable by applying a magnetic field, **Fig. 4**.

The recording industry has good reason to be interested in materials with a high magnetoresistance. The most common method of reading the magnetic information on a computer hard disk is with a small coil in which a voltage is induced as the disk spins past the head. Unfortunately all induction heads are by definition speed dependent, and hence information cannot be written at the same density right across the disk, so limiting its capacity. If, instead, a magnetoresistive element were used to read magnetic bits, then this would not be speed dependent at all. The greater the sensitivity of the element, the smaller the magnetic bits can be, and the greater the storage potential of the disk.

Since the discovery of giant magnetoresistance (gmr) there has been an enormous amount of international research activity to produce a gmr material that gives a large resistance change over a small field range, this being the requirement for making a sensitive read head. In June, IBM announced that it had succeeded, and that such a read head would be in production by the end of the century.

With its trial head, the company has been able to increase the storage capacity of a disk by 20 times, and to decipher information which was written at high density to disk some years ago, but which until now was impossible to read.

The fundamental mechanisms which underlie this so-called giant magnetoresistance are subtle but are explored in more detail in the fourth panel. The effect, though, in a multilayer material is that when all the layers are aligned, the overall device resistance is low, and when adjacent magnetic layers are antiparallel – or even if the layer orientations are random – the electrical resistance of the device is high.

The mechanism responsible for the gmr depends on the fact that regions of the material are pointing in different directions, and not on the existence of the exchange coupling. Consequently we can think up other ways in which to tailor a multilayer so that at some point in the magnetising process there is a switch from the state where the layers are antiparallel to the state where they are parallel,

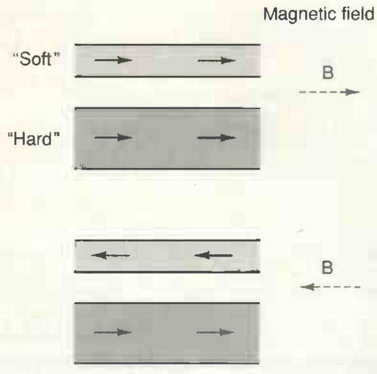


Fig. 6. Spin valves: A spin valve is very similar in structure to the device of Fig. 2 with the exception that the top and bottom magnetic layers are now made of different materials: the bottom layer is 'hard' and its magnetisation remains fixed even when a large magnetic field is applied. In contrast, the top layer is 'soft' and can be flipped in very small fields, thus giving a magnetic sensor that works in a lower magnetic field regime than those previously discussed. IBM has recently selected a system of this general type for developing a giant-magnetoresistive magnetic read head.

and hence a switch from the high resistance to the low resistance state.

A wide variety of structures called spin valves have been designed to do just this. In a spin valve, layers of magnetic material with very different magnetic properties are separated by thicker layers of non-magnetic metal so that they are sufficiently far apart to be uncoupled and hence to respond independently to externally applied magnetic fields. One of the layers is 'soft' – that is, its magnetism can be reversed by a small applied field. The other is 'hard', so its magnetic moment remains fixed unless a very large field is applied (**Fig. 6**). The electrical resistance of the device can then be changed by magnetically switching the soft layer parallel or antiparallel to the fixed hard layer.

The field at which the soft layer flips can be chosen to be very small, so such systems have the ability to produce a large resistance change in very small applied magnetic fields such as those generated by a magnetic bit on a hard disk. It is a material of this type that IBM has selected for its magnetoresistive head.

Giant magnetoresistive films do not need to be multilayers at all. We can achieve the same effects in alloys and this possibility lends even more scope and versatility to the range of magnetic devices which may be constructed. The example of Fig 3 was not itself a multilayer but a thin film alloy of silver containing little clusters of cobalt atoms which are about 3nm diameter on average. These small single domain ferromagnetic cobalt particles act like the layers in the multilayer film.

Materials like cobalt and silver are immiscible, so we have to force the two metals to mix in this way by co-evaporating or co-sputtering. However, there are many different ways that can be used to produce mixtures of immiscible materials, some of them, such as ball-milling and melt spinning, being very well suited to industrial production.

Magnetic Transistors

The mechanism behind giant magnetoresistance described under Giant magnetoresitstance, page 897 is tantalisingly similar to the semiconductor picture, where devices essentially operate by manipulating two different families of charge, the holes and electrons. This suggests that analogous magnetic transistors might be feasible.

By choosing two different magnetic materials, one hard and one soft as in the spin-valve device discussed above, the magnetic layers may be switched from parallel to antiparallel with an external field. Thus the device generates an electrical voltage that depends on the magnetisation direction of one of the constituent films (**Fig. 7**).

Moreover, the device 'remembers' its magnetic history, and hence can be used as a non-volatile magnetic switch memory device. The magnetic fields that cause the layer magnetisation reversals can be generated by pulses in small adjacent current-carrying wires, so a current amplifier configuration is possible, or indeed an assembly of several such devices can be used to make a processor.

The great promise of this technology lies in the fact that, unlike conventional semiconductor devices, fabrication of metal spin transistors becomes easier the smaller they are made. Indeed, device performance actually improves with miniaturisation. Packing density improvements of 100 are being forecast over semiconductors. This is enabled by the comparatively low power needed to run metal film devices and the ease with which the power may be dissipated, since the component materials are metallic.

Where else could the exploration of thin film technology lead? So far we have only discussed 'squeezing' nature in one dimension. This still leaves us with two dimensions to 'engineer', and we can do this with new developments in nanolithography. We can take a thin film or a multilayer, and with current techniques we can etch it into arrays of magnetic dots as small as 50nm in diameter.

With a careful choice of film material, we can make these dots single magnetic domains which behave like 'giant' magnetic atoms, with a single 'giant' spin. These 'giant' atoms can then be given special symmetries by appropriate choice of the dot shapes, and arrays of chosen symmetry of such atoms may be formed to create yet another class of new 'materials' whose interactions are established and whose properties are pre-programmable. This is a new and very unexplored field with great potential for finding and exploiting novel magnetic devices. ■

This author wished to acknowledge the support of the Magnetic Materials Initiative of the EPSRC, The Royal Society, The Worshipful Society of Fishmongers and the European Community.

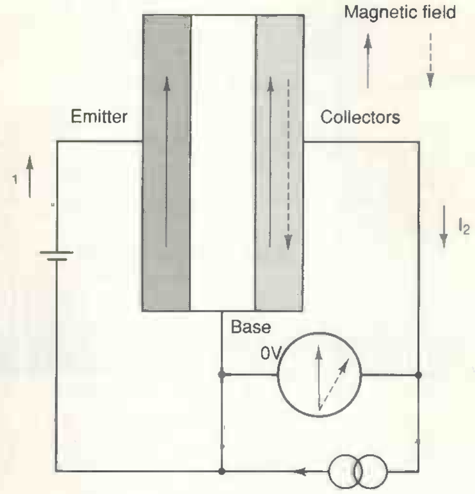


Fig. 7. Magnetic transistors: The geometry of the magnetic transistor is again like that of Fig. 2, except that this time there is an electrical connection to the non-magnetic interlayer to create a three terminal device. The battery creates a current I1 which flows from the first magnetic layer (the emitter) to the nonmagnetic layer (the base). This current is spin polarised because of the magnetism of the emitter and causes an excess of spin-down electrons in the base. If a current I2 is now drawn out of the second magnetic layer (the collector) and returned to the base, the collector-base potential varies depending on whether the collector and emitter magnetisations are parallel or antiparallel; this is because the direction of the collector magnetisation determines whether I2 will be carried across the base/collector interface by spin-up or spin-down electrons and different voltages are required to drive each electron type across the interface because of their different concentrations in the base layer.

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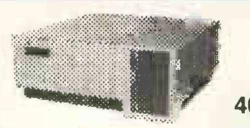

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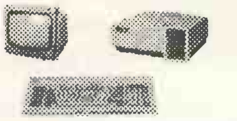

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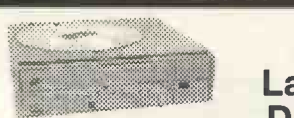

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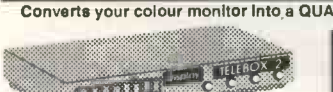

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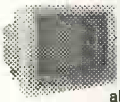

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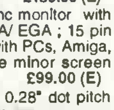

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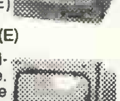

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NEW WAVE MICROWAVES

8: High-T_C superconductor circuits

Superconductor operating temperatures have risen phenomenally over the past decade, and in the race for useful devices, cooling systems are becoming ever cheaper and more efficient.
Mike Hosking describes current superconductor technology from the microwave engineer's viewpoint.

In 1911, H Kamerlingh Onnes, the first person to liquify helium, demonstrated superconductivity in mercury at the University of Leiden and started what is a continuing quest for new superconductor materials and a theoretical understanding of them. However, in the ensuing 75 years up to 1986, the critical temperature T_C at which superconducting behaviour starts, was only increased from that of 4.2K for mercury to about 23K for the widely-used niobium-tin alloy and marginally higher for niobium-germanium. Liquid helium, with its associated costs, was still required as the cryogen.

But in 1986 two scientists JG Bednorz and KA Müller of IBM discovered repeatable superconductivity, at about 33K, in a new type of compound consisting of barium, lanthanum, copper and oxygen. A year later they were both awarded the Nobel Prize for this work. The material discovered is a type of ceramic, called a perovskite and one technique of investigating new forms has been to systematically substitute other elements for the lanthanum and barium. Thus began a race, which is still continuing, between teams of researchers around the world, to find materials with higher values of T_C; the exciting question has been raised, but not answered, as to whether room temperature superconductivity will be possible?

Early in 1987, another milestone was reached by MK Wu and his research team at the University of Alabama, who produced a compound: yttrium-barium-copper oxide (YBCO) which had a critical temperature of 93K. This has turned out to be one of the most widely-used materials and is relatively easy to produce; furthermore, it exceeds the liquid nitrogen (LN2) barrier of 77K, opening the way for more efficient and convenient cooling at lower cost.

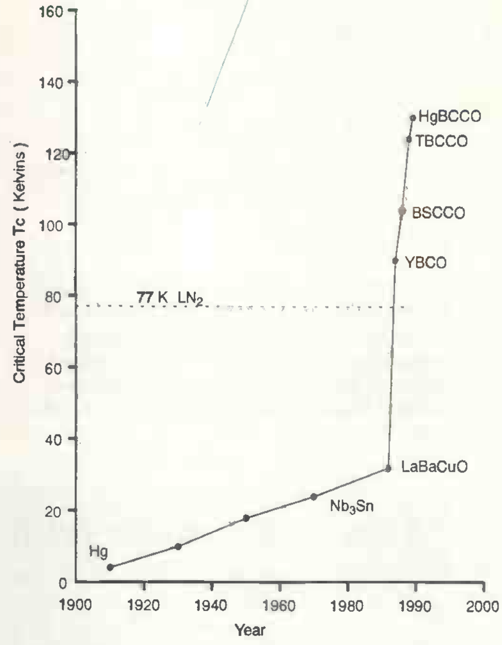

Fig. 1. Since the discovery of superconductivity in 1911, the progress made in increasing the critical temperature was relatively modest until the discovery of the new high-T_C materials in 1987.

Although many other compounds are under investigation, the most widely used, in addition to YBCO, is BSCCO (bismuth-strontium-calcium-copper oxide) with a T_C of 105K and another material, using thallium and barium, with a T_C of 125 to 128K (TBCCO).

Considerable work is also in progress using mercury as the replacement element (HgBCCO) giving a T_C of about 135K. Although higher critical temperatures, even approaching room temperature, have been observed in other materials, they have so far been short-lived events and have not been reproducible. **Fig. 1** shows the progress made in raising the value of T_C over the past 83 years and serves to demonstrate just how recent is the new high-T_C technology.

Kelvin versus centigrade

Absolute zero temperature, measured in kelvin, is approximately –273°C and the boiling point of liquid helium is 4.2K at an atmosphere; the corresponding temperature for liquid nitrogen is 77.4K, i.e. about –196°C. However, there is a great cost and energy-saving advantage in being able to use liquid nitrogen as the cryogen: not only is it about $^1/_{20}$th the price of liquid helium, but its latent heat of vaporisation is about 60 times higher, so that LN_2 overall is considerably more cost-effective.

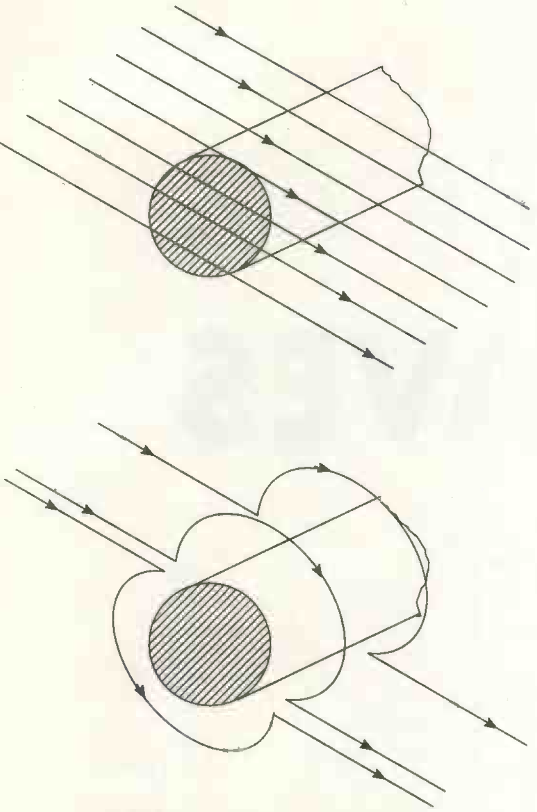

Fig. 2. *The true test of superconductivity is the expulsion of magnetic field from the interior of the conductor as well as a zero dc resistance. If the conductor was just a 'perfect' one, the magnetic flux would become trapped inside at the transition to the 'perfect' state.*

High-T_C superconductor characteristics

As with all superconductors, these new materials exhibit an abrupt change in resistivity when cooled to their critical temperature and below. At dc, the resistance of the conductor falls to zero: attempts to measure resistance in the earlier materials have revealed none down to levels of 10 to 23Ωcm and dc currents set up in superconducting rings have continued to flow for years without measurable change. However, just this phenomenon alone would also be present in a 'perfect' conductor and so a further characteristic, called the Meissner effect, must also be present in the test for a true superconductor.

As shown in **Fig. 2**, if an external magnetic field is applied to a conductor, it will penetrate throughout the material at all temperatures. However, a superconductor cooled below T_C will suddenly expel all of the internal magnetic flux, causing the field to flow around the surface. This is diamagnetism and, in a superconductor, is called the Meissner effect: most often demonstrated at science fairs by a levitating magnet.

In fact, the magnetic field is not completely expelled from the material, but decays exponentially into the surface and leads to an important superconductor parameter: the (London) penetration depth, (not to be confused with wavelength) which is analogous to the skin depth and is the distance at which the field has decayed to $1/e$ of its value at the surface. Unlike the skin depth, it is not a function of frequency; it is an intrinsic property of the material and varies with temperature below T_C. Typical values are from about 1500Å to 10μm depending upon type and quality of material (1 angstrom unit, 1Å=10^{-8}cm and 1 micron, 1μm=10^{-6}m). If $\lambda(T)$ is penetration depth at temperature T and $\lambda(0)$ the value at 0K, the variation with temperature is given by:

$$\lambda(T)=\frac{\lambda(0)}{\left\{1-\left(\frac{T}{T_C}\right)^4\right\}^{0.5}}$$

which thus increases rapidly with increasing temperature and approaches infinity as $T \to T_C$.

Surface resistance

So far, the above comments have applied to dc currents and fields under which conditions the superconductor does have zero resistivity. However, this is not the case with time-varying fields and, in an analogous way to normal conductors, the superconductor develops a surface resistance, R_S. At microwave and millimetre wave frequencies (depending on type and quality of material) this surface resistance determines the limit at which the superconductor has no loss advantage over that of a normal conductor and is one of the most important properties of the material to be specified.

Figure 3, for example, shows a plot of R_S against temperature for a melt processed thick film of YBCO on a YSZ substrate and demonstrates the dramatic reduction in resistivity at the critical temperature. However, at the measurement frequency of 13.4GHz, the resistivity does not decrease to zero but reaches a finite value depending on temperature; thin films would exhibit even lower resistance.

Forms of superconductor

One of the reasons why microwave applications are an area of great potential interest for these new high-T_C superconductors is that the highest quality materials and lowest losses are achieved for thin films, deposited on a suitable substrate. They are thus in a form compatible with other types of integrated or monolithic circuit. Such films are normally produced by laser ablation or by sputtering. In the former, a pulsed laser is focused onto a quantity of the material which evaporates and condenses onto a nearby substrate to form a thin film. Sputtering causes the target material to be bombarded by energised ions and transfers the material atom by atom to the substrate, as opposed to droplet or vapour form. An early restriction was the area of substrate which could be coated (1cm^2 or so) but this has increased to wafer sizes in excess of 3in diameter. In each case, a vital stage of the manufacturing process is a high temperature oxygen annealing cycle.

Not all of the conventional range of substrate materials is suitable for use with the high-T_C superconductors: either for reasons of 'poisoning' the films due to atom migration or because of an unacceptable electrical performance at microwave frequencies. At present, the main substrate materials used for microwave circuits are lanthanum aluminate ($LaAlO_3$) which has a dielectric constant, ε_r of about 24; magnesium oxide (MgO) with ε_r=10; yttria-stabilised zirconia (YSZ) with ε_r=30; and, more recently, crystals of neodymium gallate having ε_r=22. With reference to the earlier article on microstrip circuits: the above values of ε_r are generally much larger than those found in normal circuitry. So, as the track width required for a particular characteristic impedance of the transmission line decreases with increasing dielectric constant, finer lines and good definition are required from these superconducting circuits. A further consideration on choice and development of substrate materials is that of their loss tangents, which generally lie around 10^{-4} to 10^{-5}. Reduction of the surface resistance results in lower conductor losses and hence higher Q-factors, but a poor value of tanδ is often the limiting factor on overall performance due to the low dielectric Q-factor (1/tanδ).

Thick films of YBCO have also been produced by normal screen-printing techniques on YSZ substrates, followed by a firing and sintering stage at about 950°C and then a highly controlled oxygen annealing cycle. Such films have a higher surface resistance than thin films, restricting their present operating frequencies to the lower microwave bands. However, there is an advantage in that much larger area circuits can be readily manufactured and these are not confined to planar forms. Coaxial and cylindrical resonators, together with long wires have been successfully produced from thick films of YBCO: two examples being shown in **Fig. 6**. The cylindrical resonator operated in the TE_{011} mode and had a Q-factor of 715,000 at 77K; a similar copper resonator at the same temperature had a Q of 70,000. The coils were for a helical resonator which gave a Q-factor of 20,000 at 20K and 300MHz. It is also possible to produce the superconductors in bulk form, starting with a finely ground powder of the chemical constituents, moulding them to the required shape and then firing and annealing.

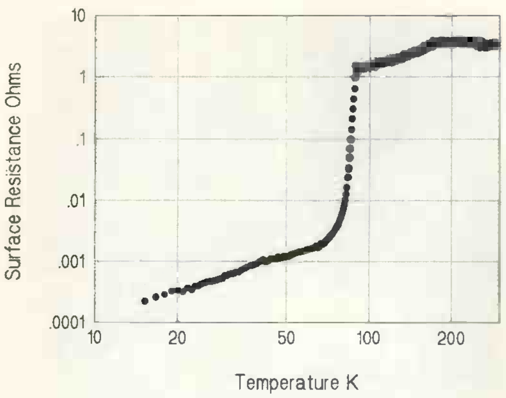

Fig. 3. To illustrate the abrupt transition to the superconducting state, the graph shows the variation of surface resistance with temperature for a melt-processed thick film of YBCO on a yttria-stabilised zirconia substrate. The onset of T_C is about 90K and the frequency of measurement is 13.4GHz. Courtesy of Bryan Tonkin, Physics Dept., University of Portsmouth.

Earlier in this series I pointed out that the microwave current in a conductor was confined to a surface layer of a thickness called the skin depth δ_S where:

$$\delta_S = \sqrt{\frac{2}{\omega\mu\sigma}}$$

and ω, μ, σ are the radial frequency, permeability and conductivity of the conductor respectively.

The surface resistance is given by $R_S=1/\sigma\delta_S\Omega$ per square and thus it can be seen that this property is proportional to the square root of frequency.

However, for a superconductor, the surface resistance is found to be proportional to the square of the frequency and so is increasing much more rapidly. This is shown in **Fig. 4**, where copper is compared with various forms of YBCO at 77K. Even though the superconductor loss may be several orders of magnitude lower than that of copper at low frequencies, a cross-over frequency is eventually reached where the two are equal. Such a consideration is important when considering specific applications and is an area of materials science wherein the quest for improvements in quality is being continuously pursued.

Paired electrons

The theory describing the superconducting state, referred to as the BCS (Bardeen, Cooper and Schrieffer) theory, was another Nobel prize-winning milestone and forms the basis of today's physical understanding. In this theory, as the superconductor is cooled to below T_C, electrons start to combine into what are referred to as Cooper pairs: each pair consisting of electrons with opposite spins and opposite and equal momenta, so that the net energy is zero.

Coupling force keeping the electrons together is provided by phonon interaction with the crystal lattice. However, above absolute zero, not all of the electrons are in pairs and this gives rise to two-fluid models of the material behaviour. At 0K, all of the electrons are in these superconducting pairs and above T_C all electrons are normal; between these two temperatures there exists a mixed state of normal conduction electrons and the Cooper pairs.

When dc current flows, the least energetic path is via the paired electrons, which effectively 'short out' the normal ones and offer no resistance. The electrons do, however, possess mass and the inertial effect of this gives rise to an inductive component. Under ac conditions, there will be an out of phase 'voltage' developed across this inductance leading to dissipative losses in the normal electrons.

Critical current and magnetic field

In addition to raising the temperature above T_C, the superconducting state can also be destroyed by increasing the current density in the material above a critical value J_C, or by increasing the external magnetic field above a critical value H_C; or by a combination of all parameters. The lower the material temperature below T_C the higher the values of current and magnetic field which the material can tolerate. Thus, there exists a three-dimensional operating region for a superconductor, as indicated in **Fig. 5**.

Apart from the power applications of superconductors (not considered here) current density and magnetic field can be important factors in microwave applications, particularly in the planar types of integrated circuit. In these, such as the microstrip form discussed in this series, current is flowing in a thin film of, say, $10^{-5}cm^2$ cross section and so high concentrations of current can occur for modest levels of microwave power. To a large extent, H_C and J_C depend on the material quality and, in thin films of BSCCO, currents well in excess of $10^6A/cm^2$ at magnetic field strengths in excess of 20T can be achieved.

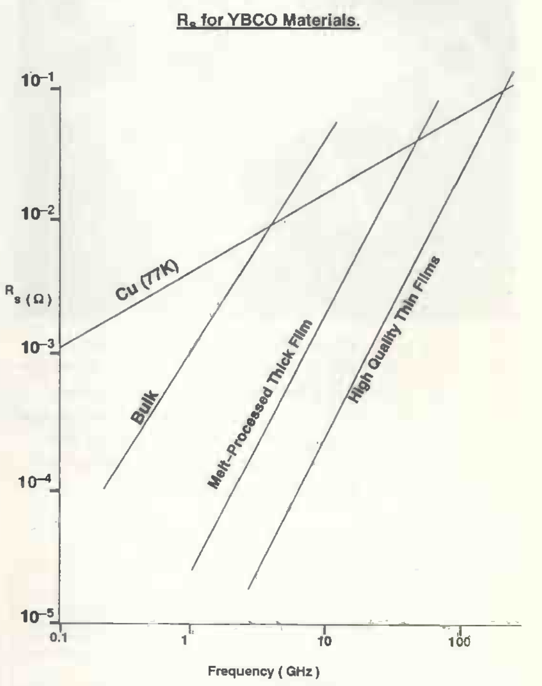

Fig. 4. The resistivity of a normal conductor increases as the square root of frequency, whereas that of a superconductor increases as the square. Thus there comes a cross-over point at which the superconductor no longer has an advantage in this respect. (N McN Alford and TW Button, Adv. Mat. 3 1991).

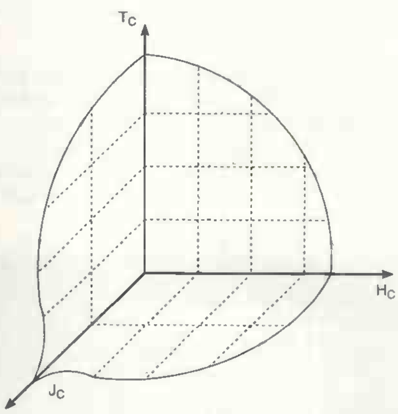

Fig. 5. The critical temperature for the onset of superconductivity is reduced by the presence of any external magnetic field or current flowing in the material. Each of these quantities has a critical value and the superconducting state can be destroyed by exceeding these values, or by a combination of current and field.

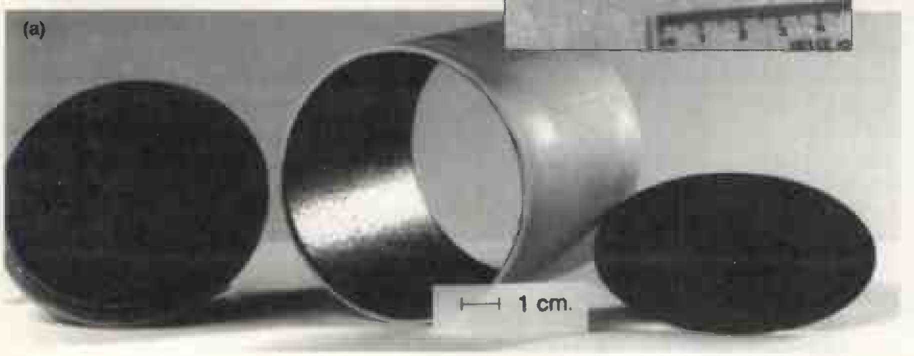

Fig. 6. (a) A cylindrical resonant of YBCO on YSZ having a Q-factor of 715,000 at 77K and (b) coils for a helical resonator operating at 300MHz. These components serve to illustrate the capability of thick films and bulk material to produce three-dimensional shapes. Courtesy of ICI Tioxide.

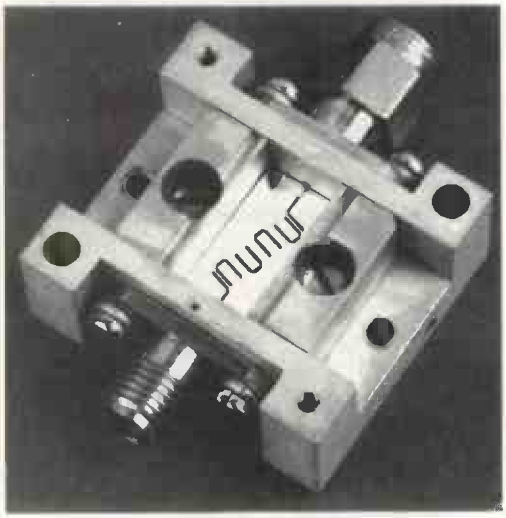

Fig. 7. Example of a thin film circuit for a four-pole filter, using gadolinium-barium-copper oxide on a lanthanum aluminate substrate. Insertion loss and selectivity far exceed that which would be obtainable with a normal microstrip circuit. Courtesy GEC-Hirst and GEC-Marconi Research Centres.

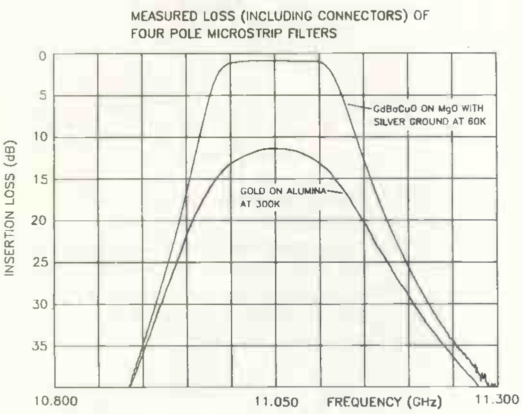

Fig. 8. Comparative performance of the four-pole filter of Fig. 7 with a normally conducting circuit. The measured insertion loss includes the connector losses and the fact that the superconducting circuit had a normal ground plane. Courtesy GEC-Hirst and GEC-Marconi Research Centres.

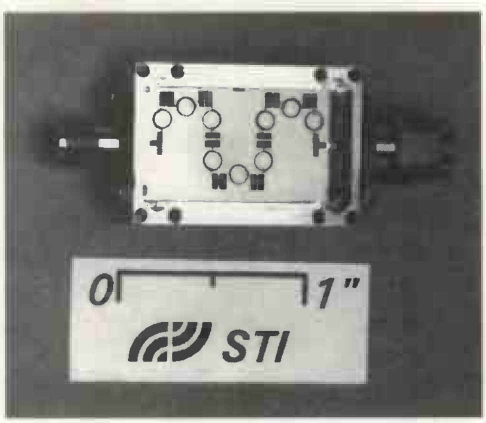

Fig. 9. The first high-T_C filter for a cellular base station replaced a large waveguide coupled resonator. This is a lumped element, nine-pole design in microstrip operating at 867MHz. Courtesy of Superconducting Technologies.

Microwave applications

Resonators: Due to the skin effect, the loss of a normal conductor increases with frequency and becomes significant at microwave and millimetre wave frequencies. This affects not just the transmission line attenuation, but also the achievable Q-factor of resonant structures.

If, by the use of superconductors, such losses can be reduced, then this technology offers improved performance to such devices as filters and resonators. Most microwave bandpass filters consist of coupled resonant structures, with the passband amplitude and phase response determined by the nature and weighting of the coupling and the stop band rejection by the number of resonators.

The use of superconducting resonators with lower insertion losses and high Q-factors would allow channelising filters to be designed with steeper skirts and thus afford systems performance advantages to satellite communications channels and to base stations for mobile communications.

An example of the design of a microstrip filter designed, developed and measured at GEC-Marconi Research Centre and fabricated at GEC-Hirst Research Centre using superconductor is shown in **Fig. 7**. The substrate is 0.25mm thick MgO, 7 by 14mm in size and the superconductor circuit is a four-pole, equiripple bandpass design with a 1% bandwidth centred at 11GHz. The superconductor material is gadolinium-barium-copper oxide (GdBaCuO), but the ground plane is normal silver.

Even though limited in performance by having a normal conductor ground plane and including the connector losses, the insertion loss of the filter at 60K was only 0.82dB; allowing for connector loss, the insertion loss was 0.48dB. An identical filter design using conventional gold conductors on alumina substrate had a room temperature insertion loss of 11.72dB and, even when cooled to 77K, still showed a loss of 6.5dB. **Figure 8** compares amplitude responses of the two filters.

Another promising filter application is that in cellular base stations for mobile communication systems where adjacent channel interference is a problem. As many of these stations operate below 1GHz, normal filters can be bulky and lossy; cavity waveguide filters are often used to obtain the low losses required to maximise the out-of-band cut-off and selectivity. High Q-factors obtainable with superconductors allow multi-pole bandpass filters to be designed with resulting improvements in skirt selectivity and insertion loss.

Claimed to be the world's first of such filters is the Superconducting Technologies example shown in **Fig. 9**. This is a ninth order, equiripple Chebychev design centred at 867MHz and replaced an existing waveguide filter more than three feet long. Each resonator section of the filter consists of a lumped element inductor (the ring shapes) and a capacitor: a lumped element being a component much smaller in size than the guide wavelength. At the input and output, tracks can just be seen a series interdigital filter. Such resonators had Q-factors of between 20,000 and 25,000 resulting in an insertion loss of below 0.3dB and an unmeasurable second harmonic rejection greater than 60dB.

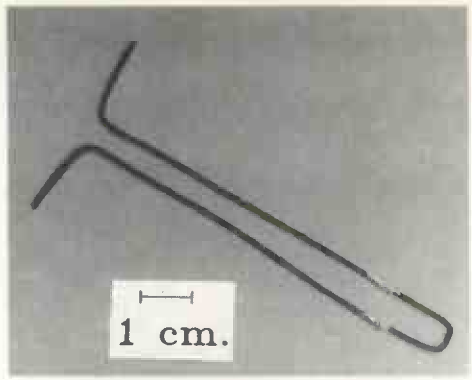

Fig. 10. The first microwave component in bulk YBCO was this short dipole, made by ICI and designed and tested by Birmingham University which demonstrated greatly reduced losses in the feed network compared with an eqivalent copper antenna.

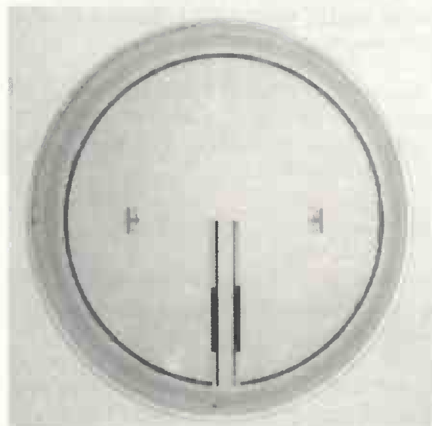

Fig. 11. The very sensitive pick-up coils used in medical MRI can benefit from high-T_C technology. This resonant antenna shown here is the first such thin film circuit. Courtesy of Superconducting Technologies.

Antennas: The half wavelength dipole is a widely-used form of antenna, either in its own right for low frequency communications or, at microwave frequencies as an element in phased arrays. Its directivity is 1.64 (or 2.15dB), with a lesser value of power gain depending upon the internal losses. Its radiation resistance is 73.1Ω.

Decreasing the length of the dipole does not greatly affect its directivity but gain decreases and the radiation resistance becomes small. For example, when the dipole length, l, is less than about $^1/_{20}$th of a wavelength λ such that a linear current distribution can be assumed, the radiation resistance is given by:

$$R_{rad} = 80\pi^2\left(\frac{l}{\lambda}\right)^2$$

So a dipole length of 0.05 would have a radiation resistance of about 0.4Ω; hence a high Q-factor. In addition, the reactance of the dipole becomes capacitive and, depending on the conductor diameter, can be very large. The overall result is that the small antenna becomes impractical to match to the normal transmission line impedances and any such matching network would be lossy and would require very critical toerances, thereby negating the initial size advantage. However, if the matching network was made superconducting, then the losses could be regained and a single short dipole has been shown (by a team from ICI and Birmingham University) to improve gain by 6dB over that of a similar copper one.

Shown in **Fig. 10**, this is the very first microwave device to be made from bulk YBCO wire and is an electrically short dipole in a parallel-wire transmission line, with signal connection contacts shown in silver. The length of short-circuited transmission line behind the contact points appears as an inductance to match the capacitive reactance of the dipole. Since this first demonstration, much further work has been done internationally to develope many other forms of printed antenna using films of superconductor.

When many dipoles, or other resonant shapes, form an antenna array, the signal distribution feed network can contribute very large losses. So that this is another area where the use of superconductors could prove of significant advantage, perhaps as the feed array to a satellite communications antenna dish.

Signal Processing: High speed analogue signal processing devices use single and tapped delay lines to perform such time domain functions such as up-chirp and down-chirp for matched filtering, correlation and Fourier transformation. Long delay lines with very low insertion loss are thus possible instead of the sometimes tens of decibels with conventional transmission lines.

Furthermore, by suitable choice of conductor and dielectric thickness of microstrip or coplanar lines, it is possible to slow the phase velocity over and above that already afforded by normal propagation in the dielectric, thereby achieving longer delays.

As penetration depth is independent of frequency in the microwave range, as opposed to skin depth which is strongly dependant, superconducting circuits can show dispersionless behaviour up to about 1THz, although geometry-related modal dispersion may occur. This makes them eminently suitable for high-speed computer interconnects and fast signal processors operating at picosecond rates. Conventional transmission lines as interconnects are too lossy and produce pulse distortion due to dispersion at well below these rates.

Medical: Cryogenic magnets are already in use, with conventional niobium superconductors requiring liquid helium cooling, for body scanning magnetic resonance imaging (MRI). But sensitive pick-up coils are required to detect the small resonances from the atoms of the body. Once again, the high-T_C materials have been successfully employed to produce these coils, typically operating at rf frequencies. Such a coil is shown in **Fig. 11** and is in the form of a TBCCO fine spiral on a 3in wafer of lanthanum aluminate, with central contacts, requiring cooling only with liquid nitrogen.

Conclusion

There are many more electronic applications for these high-T_C materials, let alone the high-power side: presented here are some of the main microwave ones. Many researchers and developers are working with Josephson junctions and superconducting quantum interference devices (squids).

There is also a rapidly-growing technology in active devices operating up to microwave frequencies called flux-flow transistors and the integration of Josephson devices with cmos circuits, but these would require a further article to do them justice. And do not forget cryostat technology which, with the added impetus given by the prospect of wider applications, is aimed at miniaturisation and economy.

We are at an exciting and very fast moving period in the discovery, development and exploitation of these new high-T_C superconductors and I am sure that the next few years will witness their consolidation into many commercial systems. ■

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CIRCLE NO. 118 ON REPLY CARD

DMM front-end *a new look*

To solve the bandwidth and drift problems associated with traditional dmm input conditioning stages, Keithley has completely redesigned the front end of its latest 6½-digit instrument.

Almost all digital multimeter input stages operate in the same way. They have an input attenuator followed by protection circuits and switches for routeing the signal depending on which range is selected.

Although tried and tested, this configuration has a number of inherent disadvantages. Involving a high resistance value, the input attenuator has a relatively high parasitic capacitance, which can limit high frequency performance of the meter. The routeing switches can couple stored charge at their output back to the input, possibly affecting the circuit under test. In addition, the traditional zener method of protecting against overload causes heating in the limiting resistor. As a result, overload recovery is slow.

In the new model *2000* dmm, all these problems have been minimised by adopting an entirely new approach to dmm input circuitry. This patented solid-state front end eliminates the common attenuator and involves power mosfet switches together with overload sensing op-amps. In addition, input resistance is specified at over 10GΩ for ranges to 10V and 10MΩ for ranges above.

A common shared protection circuit covers dc volts, ac volts and ohms. This solid-state protection provides low offset error and drift, fast response and recovery to overloads to 1kV, flat frequency response and low impedance.

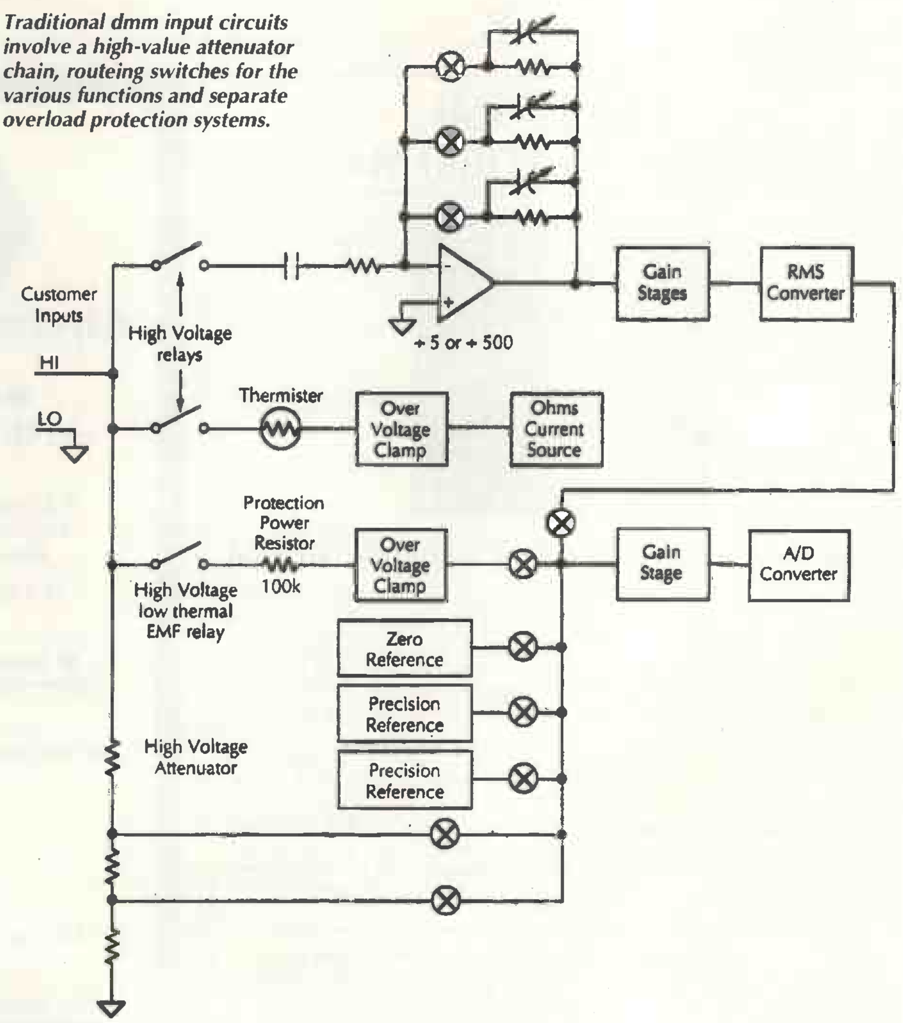
Traditional methods for protecting up to 1000 volts require three separate circuits, which increase cost and complexity. Within the *2000*, error-sensing op-amps turn off the power-mosfet switches rapidly to isolate the input. On removal of the over voltage, the switch is turned on equally rapidly. In fact overload recovery is within milliseconds.

High-performance dmms incorporate zero and gain correction of the front end to minimise errors. This slows down reading rates. With many instruments, the user has to remove the input leads so that the meter can self calibrate itself prior to use. The *2000*'s dc circuit has an inherently low offset drift front end. It requires no autozeroing or autocalibration to achieve its electrical performance.

The ac front end of the *2000* is unique for two reasons. The traditional ac front end takes the applied signal and divides it down by at least a factor of five. This simplifies the protection circuitry. However it has the disadvantage that low-level performance and frequency response are degraded. Frequency response degrades due to the complexity around the front end summing node, which is due to having to switch gains.

The new front end is configurable for unity gain or divide by 100. The unity-gain buffer is protected by the solid-state overload circuitry while the divide-by-100 circuit is protected in the traditional way. This results in better low level performance and a divide-by-100 circuit with a flatter frequency response.

Traditional dmm input circuits involve a high-value attenuator chain, routeing switches for the various functions and separate overload protection systems.

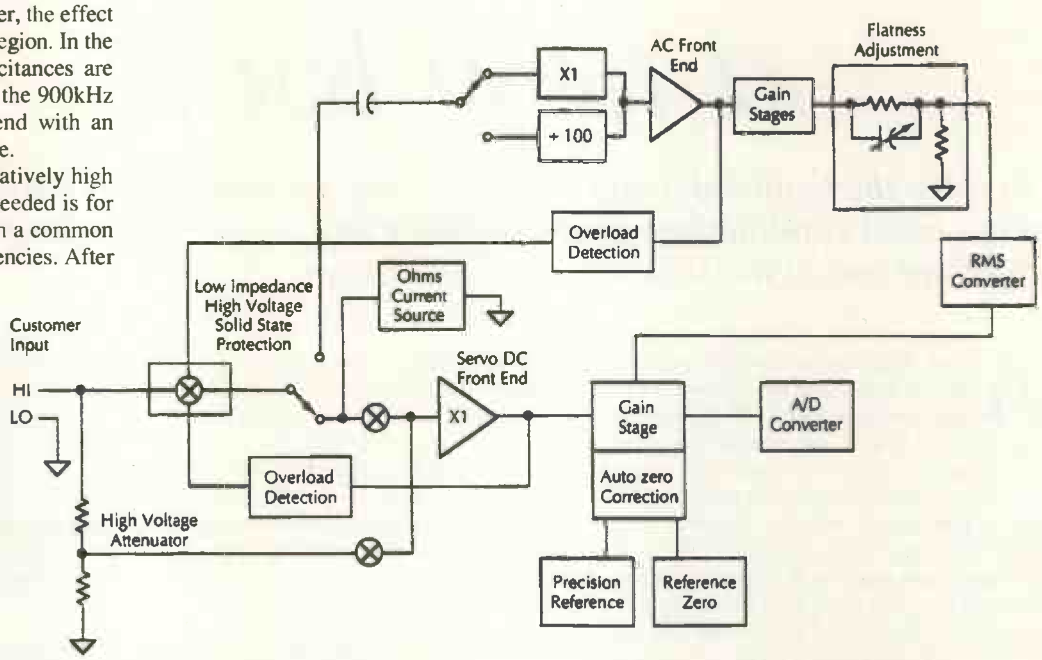


The second unique aspect of the front end is the divide-by-100 circuit. The classic problem with this type of circuit is that parasitic capacitances around the high-value input resistor cause frequency flatness errors. The classic solution is to add larger capacitors to dominate the parasitic capacitors. These capacitors are tweaked, resulting in a flatter response.

Due to the large capacitors combined with the 1MΩ input impedance however, the effect of the tweak is in the 1 to 10kHz region. In the new meter, these parasitic capacitances are reduced so the effect is raised to the 900kHz region. This results in a front end with an inherently flat frequency response.

Since the front end is flat to relatively high frequencies, the only trimming needed is for high frequency. This is done with a common rf circuit adapted for lower frequencies. After the front end conditioning and gain have been applied, the signal is attenuated by 2% with a resistive divider. Then, with a series of switches and capacitors, the high frequency gain is tweaked.

Traditionally, ac tweaks have been at lower frequencies and have had to be highly accurate. The 2000 tweak is a coarse 50kHz adjustment. Due to its location, every range can be compensated, resulting in a flatter overall instrument response. This again is an important improvement to the traditional method; one digitally-controlled circuit is used to tweak all ac ranges. The outcome is an ac bandwidth of around 900kHz as opposed to the common figure of 10kHz. ■



Claimed to be five times faster than its nearest competitor, Keithley's new 2000 dmm incorporates radically different input circuitry involving overload sensing op-amps and power mosfet switches.

CIRCLE NO. 121 ON REPLY CARD

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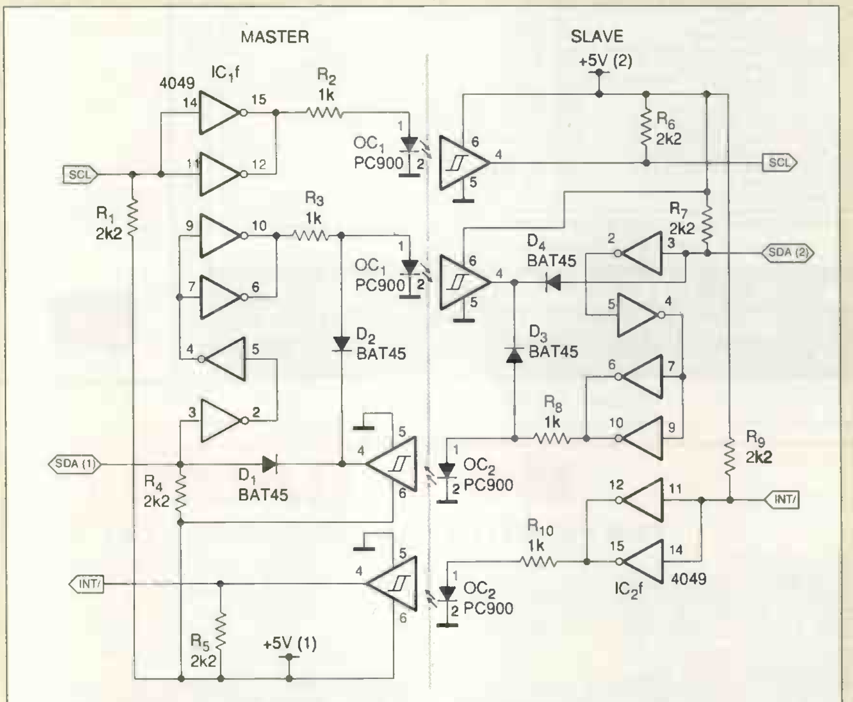
Bidirectional I²C bus isolator

On occasion, it is desirable to isolate the lines of an I²C bus to expand it or simply to separate earthing, particularly when high-resolution a-to-d converters are in use. Without precautions, simple buffers or opto-isolators in a ring do not work, since the circuit latches in one state or the other. This solution avoids that problem.

Delay in gates and opto-isolators is the core of the circuit. Suppose the $SDA_{(1)}$ input is high; pins 6 and 10 of IC_1 are low, isolator OC_1 pin 4 is off and its output, which is $SDA_{(2)}$ is high. When $SDA_{(1)}$ is low, the isolator conducts, as does D_4, and $SDA_{(2)}$ is also low. This level goes through the inverters of IC_2 and, were it not for D_3, would activate OC_2 and latch the circuit up. However, the signal through the IC_2 inverters is delayed, D_3 conducts and prevents OC_2 conducting. Diode D_1 is reverse-biased and prevents latch-up. Since the circuit is symmetrical, it is bidirectional. It works up to 100kHz, providing a swing of 0.8-5V due to the action of D_1 and D_4. Lines SCL and INT are the clock and interrupt lines and are unidirectional.

Falko K Kuhnke
Braunschweig, Germany



Opto-isolator for I²C bus avoids the problem of latch-up, using gate propagation delay.

The winner of the Thurlby-Thandar Instruments Spectrum Analyser shown opposite will be announced next month

Interfacing a signal riding high

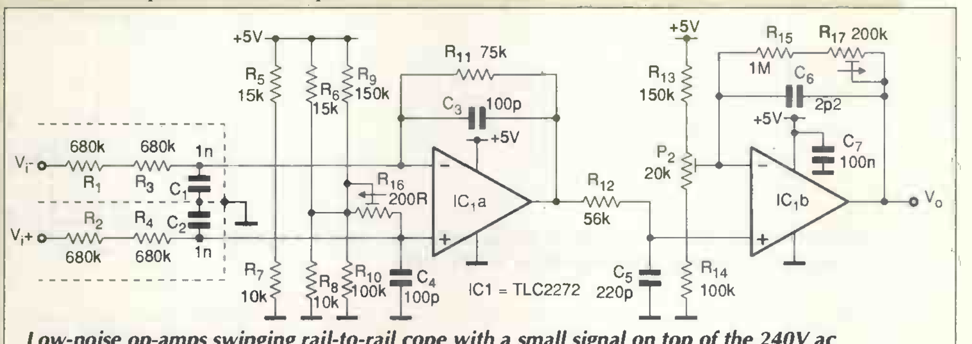
The use of a 'rail-to-rail' dual op-amp, the *TLC2272*, was the answer to the problem of interfacing a signal superimposed on a 240V ac mains live wire to a costly a-to-d converter, only a +5V supply being present.

Series resistors R_{1-4} isolate and protect the op-amp input, even when the 5V supply is off since, with R_{5-8}, there is 245:1 attenuation of the input to ensure that the signal stays within the common-mode range of the first stage. Resistors $R_{9,10}$ lower the impedance of the negative input to below that of the positive one, the difference being compensated by the trimmer R_{16}. This takes account of errors in the 1% tolerance resistors and allows adjustment for common-mode rejection. Capacitors C_{1-4} improve noise immunity, as do R_{12} and C_5. This first stage has a gain of 13.5 for differential signals.

In the second stage, gain is 18, trimmed by R_{17}, to give an overall gain for the whole amplifier of unity at a bandwidth of 0-8kHz. Trimmer R_{17} sets the output to 2V with no input.

First-stage input screening is needed, as shown, and R_{16} must be adjusted to reduce the 50Hz component to an acceptable 20mV at the output – a cm rejection ratio of about 90dB. Low-noise, high-impedance op-amps are necessary; the *LMC6482* is a possible alternative.

CJD Catto
Cambridge



Low-noise op-amps swinging rail-to-rail cope with a small signal on top of the 240V ac mains waveform, using only a 5V supply.

Serial word generator aids testing

When breadboarding a new circuit designed to work under microprocessor control, it can be difficult to generate the serial word needed for programming or control. This circuit uses a serial-to-parallel converter to generate up to a 32-bit word or more, which can be changed.

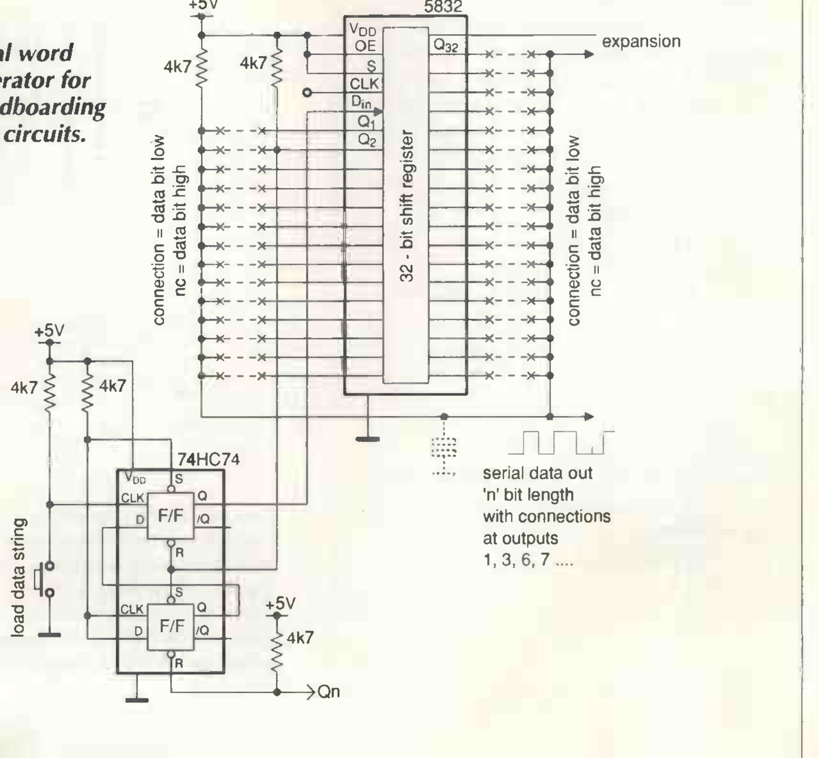
Setting the top flip-flop of the *74HC74* by the push-button causes it to insert a 1 into the shift register. The pulse is clocked through the register, each open-collector output to the data line being pulled low when its link is in place. When the pulse is at the Q_2 output, i.e. Q_2 goes low, resets the top flip-flop and sets the bottom one to prevent the loading of more data until the end of the load cycle — when the pulse has reached whichever output is made Q_n.

Since the circuit produces a random pattern at switch-on, at least 32 clock cycle should be allowed to clear the register before using the outputs or loading data. A low-value capacitor might be needed to prevent spikes between a series of 0s at the output. The *5832* bipolar outputs will sink over 100mA.

The circuit is a development of an idea by Noor Singh Khalsa.

Raymond Dewey
Allegro Microsystems Inc.

Serial word generator for breadboarding new circuits.



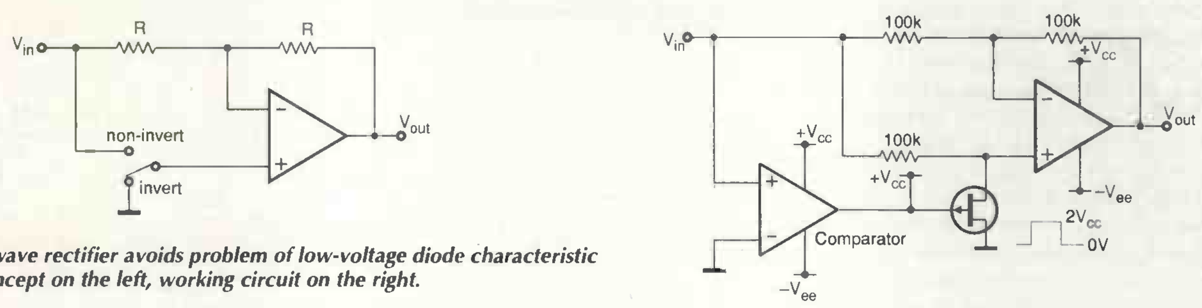
Full-wave rectifier needs no diodes

Ordinary silicon diodes do not conduct at less than 0.6V and will not, therefore, rectify lower voltages than that. The full-wave rectifier circuit shown here uses no diodes.

In the basic arrangement, left, the switch is in the lower positio and, the circuit is an inverter with a gain of –1. With the switch up, both inputs are driven and, since the gain of the non-inverter is $1+R/R$, the gain is now $+2-1=+1$. It remains to effect the switch changeover such that gain is –1 on negative half-cycles of an alternating input and +1 for the positive halves.

In the practical circuit on the right, a fet acts as the switch, taking its drive from the input via the op-amp comparator. Since it is a p-channel device, when its gate is at 0V, it is cut off and the circuit is an inverter, the reverse being the case when the gate drive is at or above cut-off. The value of R should be a hundred times $R_{DS(on)}$.

K N Sunil Kumar
Visakhapatnam
India



Full-wave rectifier avoids problem of low-voltage diode characteristic – concept on the left, working circuit on the right.

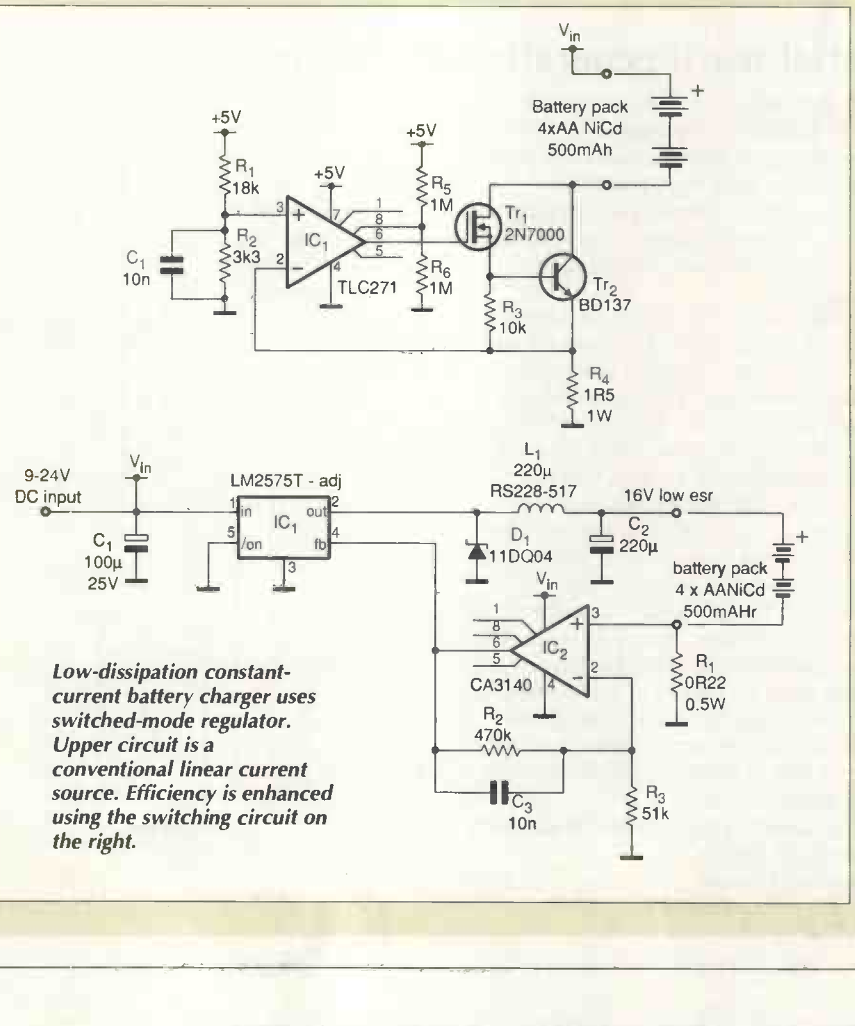
Switched-mode, constant-current charging

This switched-mode, constant-current battery charger eliminates the problem of power dissipation in the series element often associated with constant-current charging circuits of the linear variety.

The National Semiconductor *LM2575T-ADJ*, IC_1, is an adjustable 1A voltage regulator, configured here as a current regulator. Voltage generated across C_2 varies as the battery charges to maintain around 1.2V across R_1, which is fed back to the regulator. Gain in op-amp IC_2 allows a low value of R_1 to further reduce total power dissipation the circuit, C_3 being necessary to slow down the response of IC_1's internal control loop.

With these values, charging current is constant at 520mA, easily changeable to a different value by altering R_3. For much higher rates of charge, the 3A version of IC_1, the *LM2576-ADJ*, can be used with up-graded D_1 and L_1. No heat sink is needed for IC_1 with these values, dissipation being a constant 0.35W, with a conversion efficiency of over 85%.

Huw Jones
Gyrus Medical Ltd
Cardiff



Low-dissipation constant-current battery charger uses switched-mode regulator. Upper circuit is a conventional linear current source. Efficiency is enhanced using the switching circuit on the right.

Two ICs make biquad filter

A pair of current-conveyor ICs, *PA630* from Phototronics*, form a single-input filter giving low-pass, band-pass, all-pass and band-stop characteristics. Each chip contains the conveyor and a buffer.

Advantages of the circuit shown are: the use of only two ICs, against up to seven in earlier designs; very high input impedance; very low output impedance; and operating frequency up to several hundred kilohertz.

The table gives conditions and characteristics for each filter type.

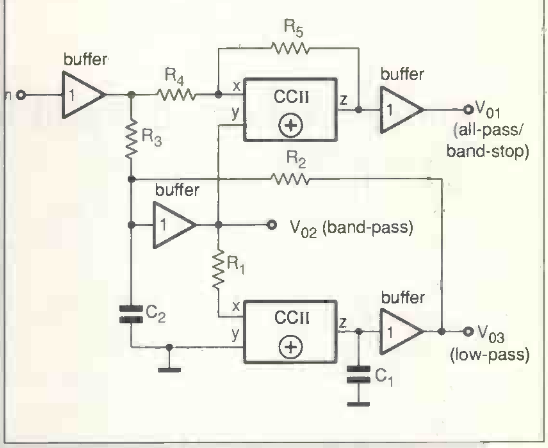
R Senani
Delhi Institute of Technology
India

**Phototronics, PO Box 977, Manotick, Ontario, K4M 1A8 Canada, Tel. 010 613 692 2247, fax 010 613 692 2605.*

Using only two chips, this biquad filter produces all four characteristics at frequencies up to several hundred kilohertz.

Pole frequency : $\omega_o = \sqrt{\dfrac{1}{R_1R_2C_1C_2}}$

Type of filter	Condition for realisation	Gain factor H_o	Other parameters : (pole Q, bandwidth)
Low-pass	—	$-\dfrac{R_2}{R_3}$	$Q_o = \dfrac{R_3}{R_2+R_3}\sqrt{\dfrac{R_2C_2}{R_1C_1}}$
All-pass	$\dfrac{2R_4}{R_5} = 1+\dfrac{2R_3}{R_2}$	$-\dfrac{R_5}{2R_4}$	
Band-pass	—	$\dfrac{R_2}{R_2+R_3}$	$BW = \dfrac{R_2+R_3}{R_2R_3C_1}$
Band-stop	$\dfrac{R_3}{R_2} = \dfrac{2R_4}{R_5}$	$-\dfrac{R_5}{2R_4}$	



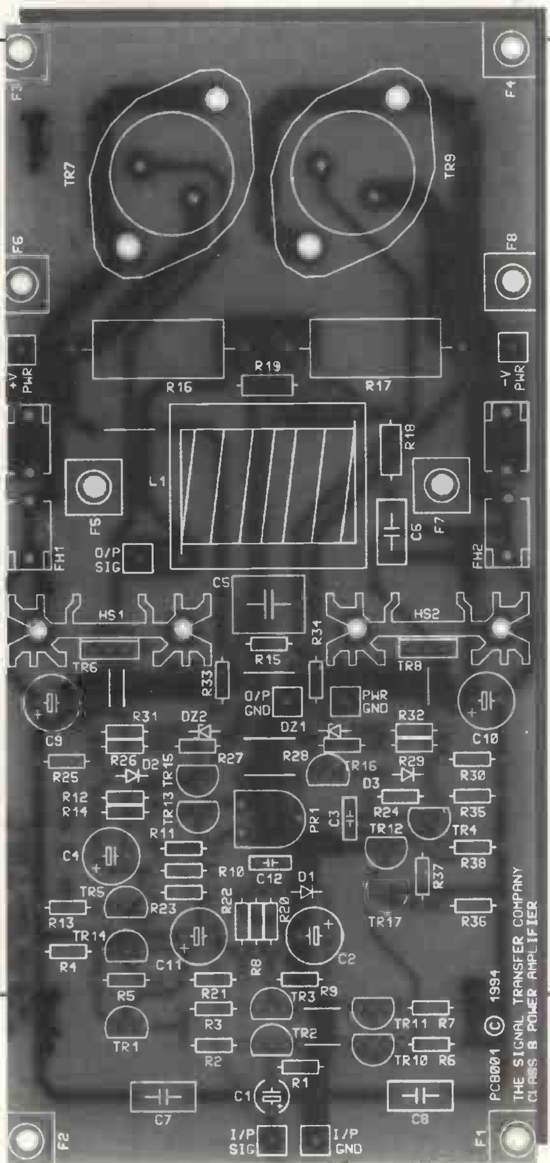
PCBs for Douglas Self's power amplifier series

Circuit boards for Douglas Self's high-performance power amplifier are now available via *EW+WW*.

Detailed on page 139 of the February issue, Douglas Self's state-of-the-art power amplifier is the culmination of ideas from one of the most detailed studies of power amplifier design ever published in a monthly magazine. Capable of delivering up to 100W into 8Ω, the amplifier features a distortion figure of 0.0015% at 50W and is designed around a new approach to feedback.

Designed by Douglas himself, the fibreglass boards have silk-screened component IDs and solder masking to minimise the possibility of shorts. Sold in pairs, the boards are supplied with additional detailed constructional notes.

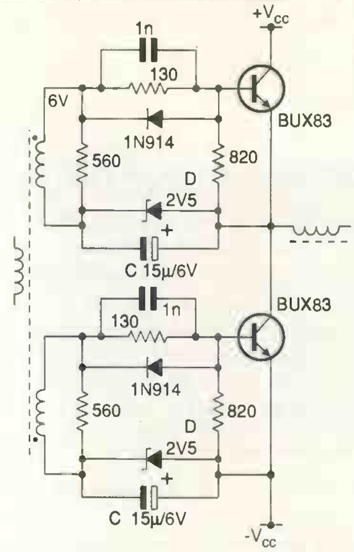
Each board pair costs £45, which includes VAT and postage, UK and overseas. Credit card orders can be placed 24 hours on 081-652 8956. Alternatively, send a postal order or cheque made payable to Reed Business Publishing to EW+WW, The Quadrant, Sutton, Surrey SM2 5AS.



Safe-switching bridge power stage

Long life for transistors in a bridge or half-bridge power stage depends on rapid switch-off during the dead time. In this circuit, switch-off is achieved by a stabiliser diode and a capacitor for each transistor and forces switch-off at any mark:space ratio. When, say, Tr_2 switches on, current from the transformer cause voltage drop across reference-diode *D* (not a zener), which charges *C* quickly, since its impedance is low. During the dead time, applied voltage is zero and the volts across *C*, less the *IN914* voltage, forces a reverse current in Tr_2, switching it off. When the negative cycle starts, what is left of the voltage across C reinforces it and ensures Tr_2 switch-off. Transistors having a reverse breakdown of 10V provide best results. A zener must not be used, since reverse-bias conduction is not permitted.

G Mirsky
Moscow
Russia



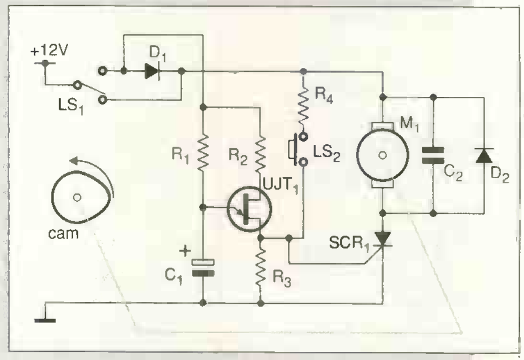
Adding a reference diode and a capacitor to each device in a bridge ensures fast switch-off during the dead time.

Battery-powered liquid feed timer

Electromechanical switching allows the flow of liquid (water, in this case) for five minutes when activated by a coin. The device is battery-powered, since there was no mains supply.

As seen in the diagram, the circuit is at rest, D_2 and SCR_1 passing only leakage current. When the coin closes LS_2, SCR_1 switches the motor on and the cam rotates 180°, opening the liquid valve. As the cam turns, it switches LS_1, momentarily interrupting the motor supply and turning SCR_1 off. Timing circuit R_1C_1 now receives voltage and UJT_1 eventually fires and restarts the motor, which drives the cam back to the start and turns off the water valve. The motor stops as LS_1 is again switched and the motor supply interrupted.

M J Nicholas
Bournemouth
Dorset



Timing circuit operates valve to supply liquid for a given time when operated by a coin.

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CIRCLE NO. 124 ON REPLY CARD

Cautionary tales for circuit designers

Some circuits look plausible and are disastrous, some work in spite of appearances and some could work with a following wind. Ian Hickman tells of his sufferings.

Often, a quick glance at a circuit diagram is enough to tell what it is supposed to do, and a little longer one will usually enable one to judge whether it will actually do it. Sometimes, though, there is a hidden catch, and the circuit won't work; at other times, it turns out that a seemingly unlikely circuit will work. Here is a selection of circuits for you to ponder, most but not all falling into the former category, which I have collected over the years.

Garbage in – garbage out

Firstly, a scheme for deriving an equal mark/space ratio square-wave from one with an unequal ratio. Everyone knows that if you divide a frequency by two you get an equal mark/space ratio. But **Fig.1 (a)** contains a howler that anyone can see through almost immediately. At least, almost anyone, as it was submitted by someone who presumably thought it would work (although he obviously hadn't tried it) to the readers' design ideas section of one of the controlled circulation magazines (now defunct). It was not the April issue!

Many readers wrote in to say it didn't work, one submitting two alternative circuits that do, also shown in Fig. 1. The circuit of Fig. **1(b)** operates over at least a 10:1 frequency range, given the appropriate component values, delivering a 50:50 ratio output. That of Fig. **1(c)** also operates over a range of 10:1 or more, with the further advantage that one edge of the output square-wave is coincident with that of the asymmetrical input waveform.

Re-inventing the wheel

Next, a circuit from the early sixties – when logic circuitry still used discrete components. We were working on missile test equipment, which was the first to make extensive use of digital measurement techniques. All measurement results, whether volts, frequency, period, d*V*/d*t* or whatever it was, were read out on a purpose designed-dvm – also part of the project.

A colleague charged with designing part of the digital circuitry had an arrangement of gates which I don't recall exactly but which was something like that shown in **Fig. 2(a)**, using various diode logic gates of the type illustrated in Fig. **2(b)**. It incorporated the bright idea of feeding back a gate output to an earlier gate, to which was applied a short 'take measurement' command pulse. This neatly ensured that the output gate was held open for the duration of the measurement, however long that took. Unfortunately, it didn't; even substituting a transistor And gate at gate B didn't help, and a discussion ensued among us all as to why not.

I pointed out that the 'gain' through a diode gate or even an emitter follower gate was just a little less than unity, so that when the signal was fed back to an earlier one after passing through a couple of gates, it was impossible for it to hold itself on. Our colleague went away to think about it and decided that the answer was to include an inverting transistor gate as shown in Fig. **2(c)**. These were only used where essential on cost grounds, but here it would serve to include the necessary gain in the loop and so was justified. Unfortunately, it had the incidental property of inverting the logic signal, so that didn't work either.

Finally he came up with the solution: the input logic was allowed to be an inverting transistor gate (requiring an inverted 'end of measurement' signal)

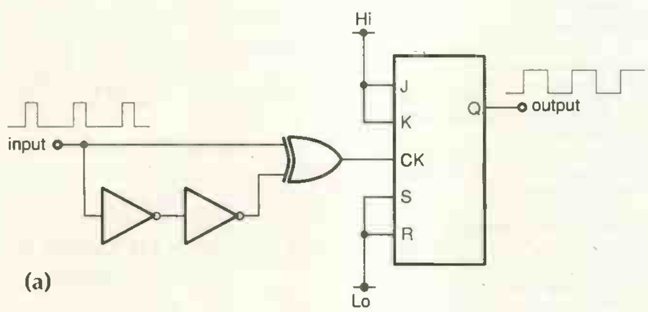
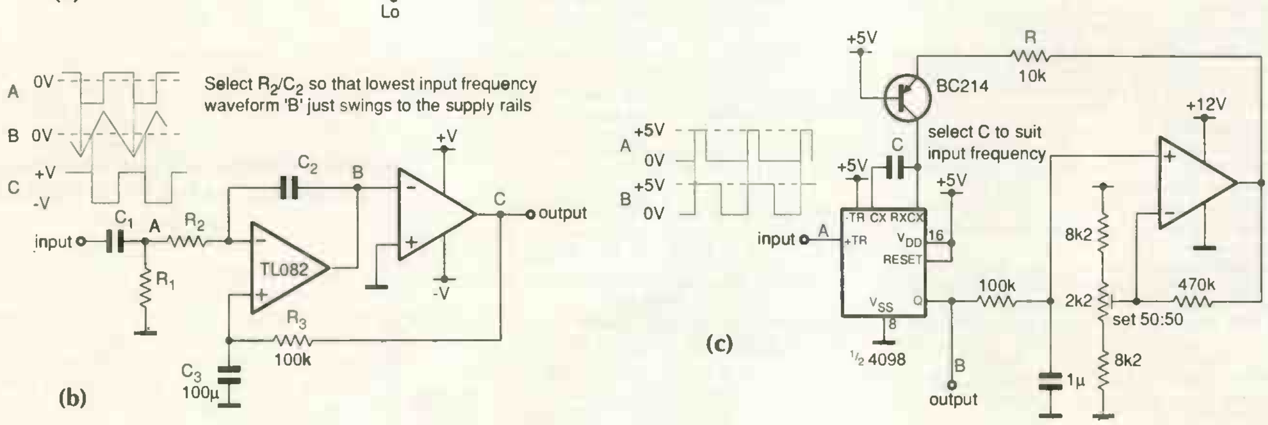


Fig.1. (a) Flawed circuit for deriving an equal mark/space ratio squarewave from on asymmetrical one. Alternative circuit at (b) works. as does that at (c), which also has an edge that is coincident with the input waveform.



while gate B would be similar. It worked a treat.
"Congratulations", said someone, "you've just re-invented the Eccles-Jordan flip-flop!"

A bootstrap too far

Now for another lame-duck circuit; one which was actually proposed (in an article about bootstrapping by someone who should have known better) in the august pages of this very magazine, quite a few years ago now when op-amps were less common and discrete transistor circuitry still the norm. The scheme for bootstrapping the base bias circuit in **Fig. 3(a)** is well known and very effective, particularly if the load on the emitter follower's output is light and it is provided with a constant-current long tail so that its gain is very close to unity. The input resistance still includes a shunt contribution from the transistor's collector/base resistance, but if the collector voltage were to follow the emitter voltage (and hence the input signal voltage), this component would also be bootstrapped out of sight. Fig. 3**(b)** shows the arrangement, which is elegant and also impossible.

It cannot work, for if the base current is negligible and the load on the circuit's output likewise, then the collector current must at

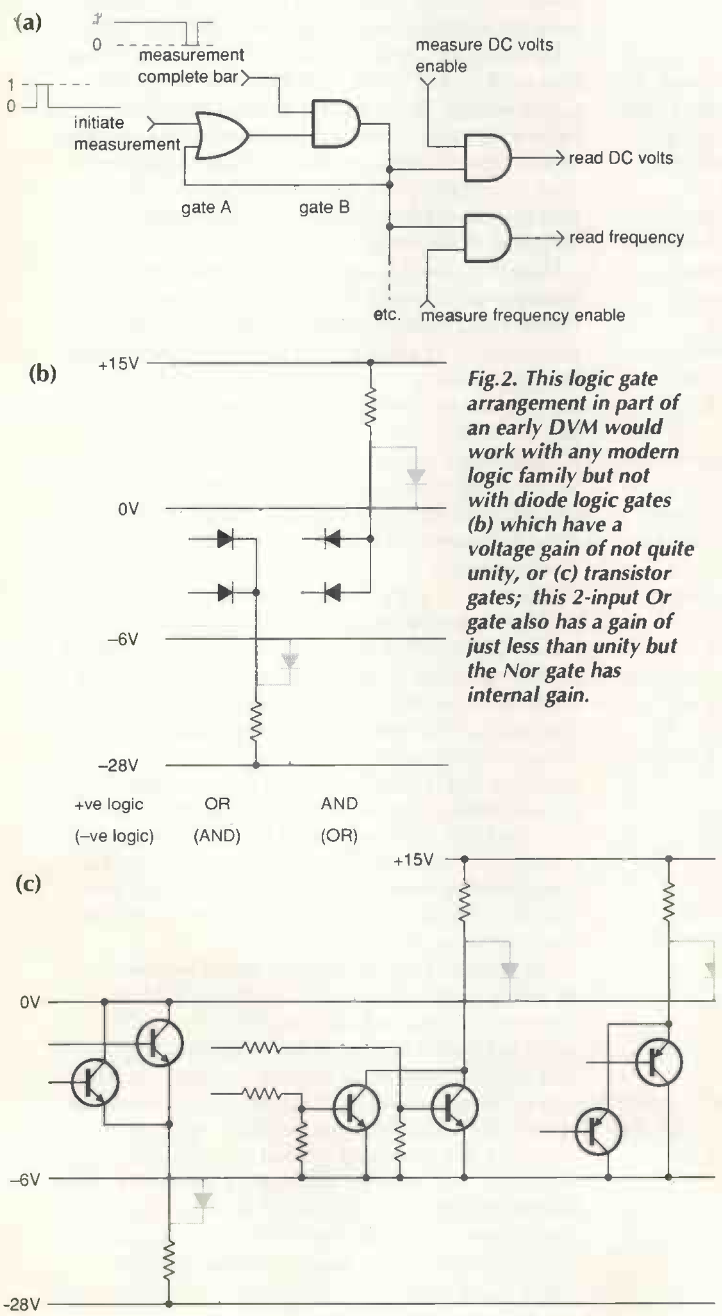


Fig.2. This logic gate arrangement in part of an early DVM would work with any modern logic family but not with diode logic gates (b) which have a voltage gain of not quite unity, or (c) transistor gates; this 2-input Or gate also has a gain of just less than unity but the Nor gate has internal gain.

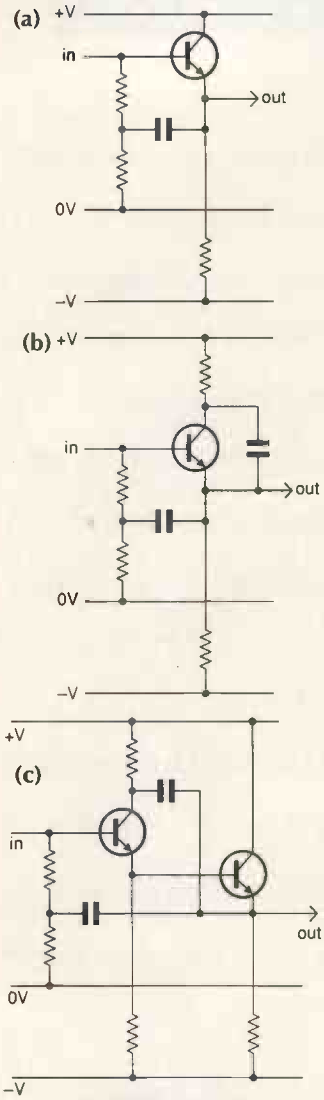


Fig.3. Bootstrapping an emitter follower's bias circuit, as in (a), to raise its input impedance works, but bootstrapping its own collector (b) for the same purpose doesn't, unless there is additional current gain in the loop, as at (c).

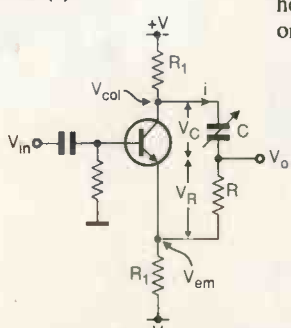
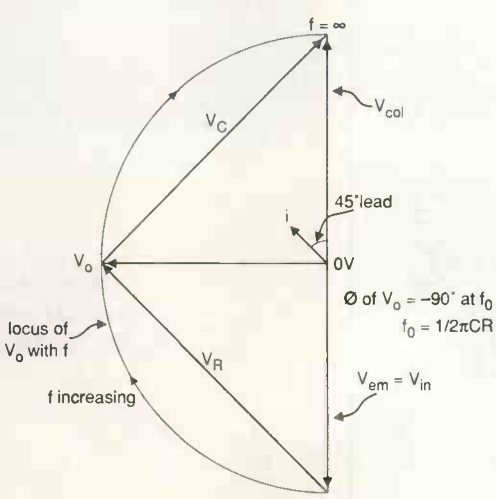


Fig.4. Simple all-pass filter stage with unity gain at all frequencies and a phaseshift varying from 0 to -180° as C varies from zero to infinity, or as the frequency varies from zero to infinity for a fixed value of C. Differing emitter and collector impedances are unimportant.



every instant equal the emitter current. So when the drop across the emitter resistor increases, so must the drop across the collector resistance; there is nowhere else for the extra emitter current to go. Ergo, the collector voltage must fall.

What if the emitter follower has a constant-current tail? In this case, the collector current cannot change and so neither can the voltage drop across the collector resistor, but the current through the transistor can change. The emitter can only follow a positive-going input by discharging the collector bootstrap capacitor. On a negative-going edge, the transistor will cut off and the constant tail-current will transfer to the capacitor, charging its bottom plate negative-wards until the transistor cuts on again while the collector voltage remains undisturbed.

Collector bootstrapping to get rid of the reduction of input impedance due to r_{cb} is however useful and effective; it just requires some extra current gain, to allow the input emitter follower's collector current to differ from its emitter current as in Fig. **3(c)**.

Impedances and currents

This point about the equality of an emitter-follower's emitter and collector currents is often overlooked. At one time I was required to design a calibrated variable phase-shift circuit. Having been impressed by an article called *The selectoject*, a circuit providing a tunable audio frequency band-pass or notch characteristic, as required, in the (then) *Wireless World*, I borrowed the basic idea and used the circuit shown in **Fig. 4**. Since the emitter and collector resistors are equal, the signal voltages at those electrodes must be equal in amplitude and in antiphase. When the reactance of C numerically equals R, the current in the branch RC leads the voltage across it by 45°. Thus the output voltage is variable over the range –180°, through –90° to 0° as C is varied from infinity down to zero. Its range was, of course, less than that, but the circuit worked very well, considering the performance of the transistors then available and given that the operating frequency was well above the audio range. But naturally its performance wasn't quite perfect. Said a colleague "I don't see how you can expect it to be, given the unbalanced source impedances driving the ends of the series CR. The bottom end is driven by the low output impedance of an emitter follower and the top end by the collector output impedance, which is high". So here is a circuit that works, even though at first glance one might think it wouldn't.

Same – but different

Some circuits work at times but not at others, that is to say one example works but another build of the same design does not. **Figure 5** looks like a 'spot-the-difference' puzzle, the only difference in fact being the addition in **(b)** of D_3. Figure **5(a)**, which appeared in the readers' design ideas section of one of the controlled circulation magazines, is a stabiliser circuit designed for use with a bank of NiCd cells. Although these have a fairly constant voltage during the discharge cycle, there is some voltage sag, especially if the operating temperature range is wide. This is undesirable if the battery pack is powering sensitive measuring equipment, so the stabiliser circuit shown was developed.

Error amplifier Tr_1 controls the compound pass transistor stage Tr_3/Tr_4, comparing the fraction of the output voltage across R_4 and part of R_5 with the reference voltage across zener diode D_1. Regulation is good, and so is stabilisation, since the reference voltage across D_1 (which should be a type with a sharp knee, suitable for use at low current) is derived from the stabilised output rather than the raw supply.

However, the circuit is bistable; if no output voltage, then no drive to Tr_3 and if no drive to Tr_3 then no output voltage. So R_2 ensures start-up when the batteries are first connected, or following the removal of an extended short circuit at the output, and provides short-circuit protection by limiting the drive to the pass transistor when the output current reaches a level sufficient to drop about 600mV across R_1.

Diode D_2 provides a path for recharging the cells, since it was envisaged that the circuit might be incorporated within the battery pack. This prevents the danger of damage or even fire if the battery were accidentally short-circuited, since with large cells there is a lot of stored energy and, on short-circuit, this can be released in a very short time. Housekeeping current on no-load is a miserly 55µA, but when the battery pack is not in use this can be reduced even further to a negligible 4µA or so (via R_2) if the battery pack be stored with the terminals short-circuited.

Capacitor C_1 is the clever part of the circuit. As explained, on extended short-circuit, the drain on the battery falls to a few microamps but, when the battery pack is connected to an instrument, it may find itself suddenly in parallel with a large decoupling electrolytic. This will cause the output terminal voltage initially to drop to zero. After this the capacitor will rapidly be charged up at the short-circuit current determined by R_1, if and only if Tr_1 is still supplying collector current for Tr_3. To fulfil this condition, C_1 maintains the voltage at Tr_1 base long enough for the terminal voltage to recover to a level (a volt or two) from which it would then build up to the rated output anyway.

The prototype circuits reliably turned on into a load including a 2000µF capacitor, so all seemed well. Some years later I had occasion to use this circuit again and built it up exactly as in Fig. 5(a), only to find that on connecting a capacitor greater than a few tens of microfarads at most, output voltage would not recover; the circuit remained sullenly switched off.

Solving this teaser took several cups of coffee before the light dawned. The stated purpose of C_1 is to hold up the voltage at the base of Tr_1 while the short-circuit current set by R_1 (not the short-circuit current via R_2) starts to charge up any external capacitance which might be connected. But, unfortunately, there is

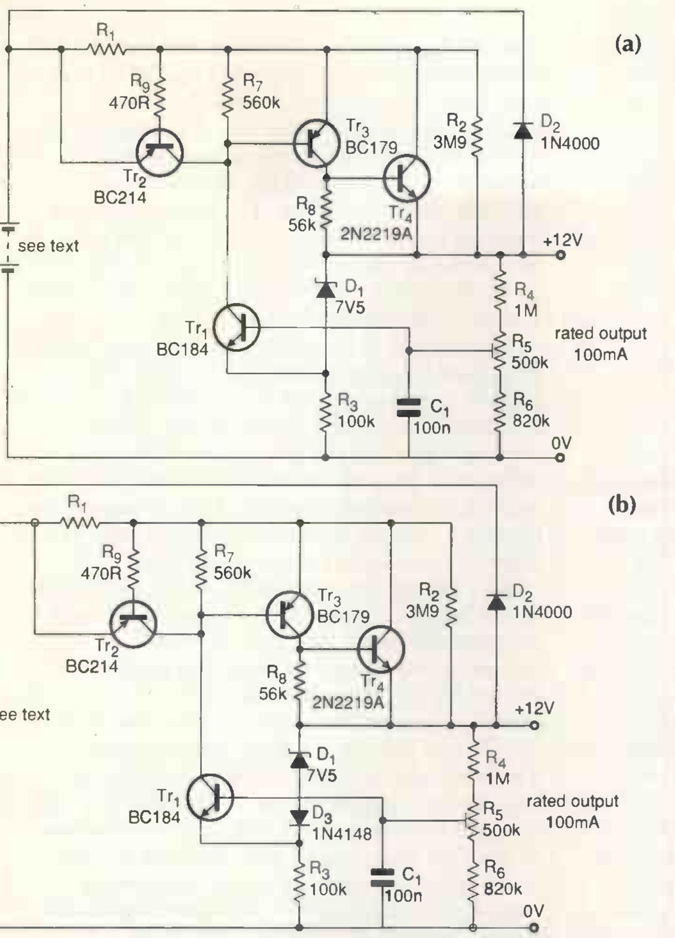


Fig. 5. Circuit of a current-limited power supply (a) whose output voltage may fail to recover when connected to a load including a large electrolytic. Adding D_3 cures the problem.

a discharge path for C_1 via the base emitter junction of Tr_1 in series with D_1 working as a normal diode in forward conduction), on by way of the momentary short across the output terminals due to the external electrolytic, back to the other terminal of C_1. Adding D_3, as in Fig. 5(b), cleared the fault entirely.

It is still not clear to me why the first prototypes worked, with as much external capacitance as one cared to throw at them, while the later ones would only stand a few tens of microfarads. Clearly, a case of minor differences between characteristics in devices which are nonetheless all individually within specification. Would the problem have shown up on a CAD simulation package such as Touchstone or *Mathcad* or *Spice*? I wonder. It depends on what limit values are built into the library models for the various parameters of the devices used, such as the extrinsic base resistance r_{bb}' of Tr_1, etc.

Differentiate and oscillate

Now for the real peach of a circuit shown in **Fig. 6(a)** which, if it really worked would be extremely useful. By way of introduction, remember that if you want to make a high-Q filter, be it low-pass, band-pass or high-pass, you need at least two poles. And if we are talking about an *RC* active filter and want it tunable, that usually means a two-gang potentiometer or variable capacitor. Thus the *CR* product of the frequency determining sections can be varied in step, providing, say, a 10:1 tuning range for a 10:1 variation of the variable elements.

This is not to say that you can't make a variable-frequency filter or sinewave oscillator using a single variable element; on the contrary you certainly can and Ref. 1 gives an example, while Ref. 2 describes no less than five such circuits. But the price you pay includes among other things, a reduced tuning range; an n:1 variation of the tuning element R gives only a $\sqrt{n}$:1 tuning range. The circuit of Fig. 6 seems to break through this limitation. And it should

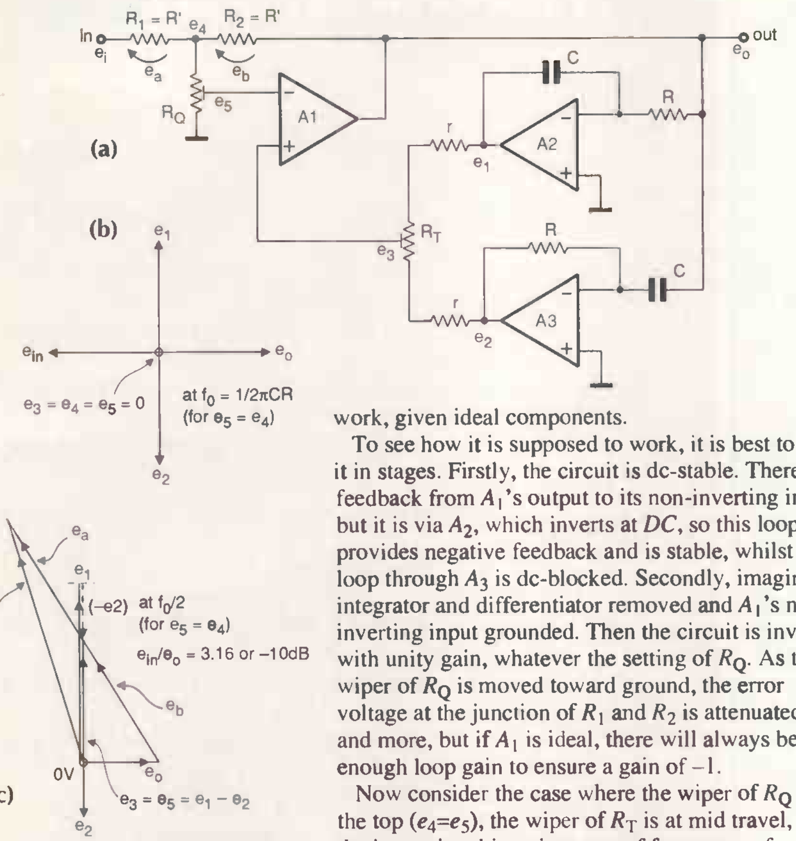


Fig.6. Variable-Q band-pass filter with constant centre frequency gain of unity (a) and its vector diagram (b) showing operation at the centre frequency, at minimum Q. Vector diagram at (c) shows operation at one octave below the centre frequency, at minimum Q. If you can stop it oscillating, it could be useful.

work, given ideal components.

To see how it is supposed to work, it is best to take it in stages. Firstly, the circuit is dc-stable. There is feedback from A_1's output to its non-inverting input, but it is via A_2, which inverts at *DC*, so this loop provides negative feedback and is stable, whilst the loop through A_3 is dc-blocked. Secondly, imagine the integrator and differentiator removed and A_1's non-inverting input grounded. Then the circuit is inverting with unity gain, whatever the setting of R_Q. As the wiper of R_Q is moved toward ground, the error voltage at the junction of R_1 and R_2 is attenuated more and more, but if A_1 is ideal, there will always be enough loop gain to ensure a gain of –1.

Now consider the case where the wiper of R_Q is at the top (e_4=e_5), the wiper of R_T is at mid travel, and the input signal is a sinewave of frequency of f_0=1/(2π*CR*). At this frequency, both the integrator and the differentiator have a gain of unity, the integrator output e_1 leading e_o by 90° (it is an inverting integrator) and the differentiator output e_2 lagging by 90°. Net voltage at the wiper of the tuning control RT, e_3, is zero and e_o=$-e_1$. If the wiper of R_T moves towards the integrator output, e_3 is zero at a somewhat higher frequency, or at a lower frequency if moved towards the differentiator output. In the limit, at the end of R_T's travel, for zero output at the wiper one output must be $(R_T+r)/r=N$ times the other, i.e. at $\sqrt{N}$ times f_0. So the tuning range is from $f_0\sqrt{N}$ to $f_0/\sqrt{N}$, or N:1. As r is made smaller relative to R_T, the tuning range becomes larger and larger.

Figure **6(b)** shows the situation at the band-pass centre frequency and Fig. 6(c) at a frequency one octave lower, for the case where e_5=e_4 (minimum Q). Clearly, as e_3 increases off tune, so e_o becomes smaller relative to e_i, so e_4 is no longer zero. Still, off-tune the output will not be far below unity so long as e_3 is small compared to e_5. The allowable detuning while still meeting this condition gets smaller as e_5 becomes a smaller proportion of e_4. Finally, as the wiper of R_Q approaches ground and e_5 tends to zero, any departure whatever from exact equality of e_1 and e_2 (i.e. any departure from the exactly on-tune condition) will result in a fall in e_o. Put another way, in these circumstance, A_1 will produce whatever output is necessary to keep the signal at its non-inverting input equal to that at its I input. If e_1 does not equal e_2, the only way it can arrange this is if e_o is near zero. The circuit provides a range of Q variable up to infinity, but with the on-tune response remaining at unity independently of the value of Q.

In principle all is fine; in practice the circuit is likely to oscillate – it certainly did when I tried it. The problem is the loop through A_1, A_3 and back to A_1. Integrators are splendid, docile circuits, since the demanded (closed-loop) gain falls with frequency at 6dB/octave, the same rate as the open-loop gain of an internally compensated op-amp. Thus the gain within the loop is constant until well beyond the unity-gain frequency and stability is therefore assured.

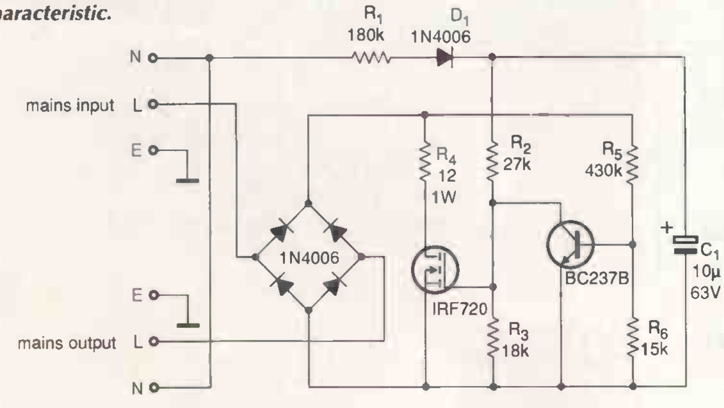
Differentiators are a very different kettle of fish: the demanded gain rises at 6dB/octave, while the open-loop gain falls at the same rate. Eventually, the demanded gain exceeds the open loop gain and all bets are off. A_1 output then effectively connects directly to A_3 input, with both op-amps contributing 90° of phase shift. At a high enough frequency, additional poles appear in the op-amps open-loop responses and oscillation results. Perhaps with a very high performance A_3 with a little capacitance across its feedback resistor and a little resistance in series with its input capacitor, one could turn it into an integrator at some frequency well above the band of interest, ensuring the stability of the circuit as a whole.

Fast fuse – if it works

Finally, another very useful circuit. I have had it on file for some time but have not made it up myself. However, at a recent gathering of engineers I fell into conversation with someone who had, and he claimed it didn't work. Did he substitute different components or values, or just get the wiring wrong? Or is there really a problem? I can't see any reason why it shouldn't work in principle (though a tolerance exercise on the component values might not come amiss) so in my book it remains a definite maybe.

The circuit, **Fig. 7**, is an electronic mains fuse, but faster than a fuse, a thermal trip or a magnetic breaker. In fact, it is not so much a fuse as a limiter since, if the load tries to draw more than the rated 100W, the circuit exhibits a re-entrant foldback characteristic; circuit operation should be clear from Fig. 7. Such a device is clearly a must for the lab. bench, so at the first opportunity I shall try it out. In the meantime, evaluating the viability of this circuit is (as it says in so many text books) 'left as an exercise for the reader'. (Hint: what voltage will the peak current through, say, a 125W resistive load drop across R_4 plus the power mosfet, and is this enough to turn on the n-p-n transistor?) ■

Fig.7. Fast-acting limiter allowing a maximum output of 100W, beyond which it exhibits a re-entrant foldback characteristic.



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2 Wiliams, P. *RC oscillators: single-element frequency control. Wireless World,* December 1980, pp 82-84.

LCR measuring transformed

Turning a transformer ratio-arm bridge back-to-front and discarding the transformers may seem perverse, to say the least, but Ian Hickman does just that to make an accurate component bridge.

One can design a general-purpose component bridge to cover a wide range of values of resistance, capacitance and inductance using a few close tolerance resistors as standards, some op-amps and a little ingenuity. Furthermore, basing the design on the principle of the transformer ratio arm bridge allows digital readout of the measured values. **Figure 1** shows the principle.

Oscillator

Using fixed-frequency operation at $\omega=10^4$ (1.5915kHz) simplifies matters. The oscillator is a state-variable filter, since it makes three outputs in quadrature available. As **Fig. 2** shows, it is simply a filter with zero damping, not a conventional oscillator, but there is no practical difference between an oscillator and a filter with infinite Q. An integrator produces a 90° phase lag (the integral of a cosine wave is a sinewave), but each of the two integrators in the loop apparently produces a 90° *lead*. This is because they operate in the inverting connection, and the relative phases are therefore as shown in Fig. 2. The bandpass output is labelled 0° because at the filter's resonant frequency $\omega=1/CR$, it would be in phase with an external input applied via a resistor at the inverting input of A_1. If R is 100kΩ and C is 1nF, then the nominal resonant frequency is 1.5915kHz.

Using a *TL084* quad op-amp, the circuit oscillates at the required frequency – in fact at 1.5914kHz, which is surely good enough. With the ±12.5V rails used, the amplitude is ±11V peak, amplitude stabilisation being provided by slight clipping of the peaks in each op-amp. Evidently, excess phase shift in the op-amps, which must be minimal at this frequency, together with layout strays, ensures that the overall loop phase-shift does not fall short of the 360° needed for oscillation, but the very small degree of clipping indicates that it barely exceeds what was necessary. Had it not performed, a few picofarads in parallel with R_4 would have persuaded it.

Figure 3(a) shows the Lissajous figure, produced by the bp and lp outputs applied to an oscilloscope in X/Y mode, the clipping appearing as the slight flat tyres at the bottom and left-hand side of the circle; both outputs measured

0.4% thd. Although all three amplifiers were clipping, the bp and lp outputs each show only the clipping occurring in that particular stage, because the harmonics making up the dent in the input waveform to each integrator are attenuated much more than the fundamental at the output. However, the hp output from A_1 does not benefit from this, so both its own clipping and that fed back from A_3 can be seen in Fig. **3(b)**, which shows the distortion meter residual output when measuring the hp output thd, which was 0.12%. Here, the time base speed has been adjusted to 157μs per division, corresponding to 90°. That the clipping in A_1 and A_3 occurs in quadrature is clearly evident.

Nulling the minor terms. While the arrangement of **Fig. 4** operates very like a transformer ratio-arm bridge, it has a number of drawbacks for use as a general-purpose component bridge. For example, it works well for pure resistors, capacitors and inductors, but there is no provision for nulling out the self-capacitance or self-inductance of a resistor, or the loss component of an inductor or capacitor. Also, although R_v can be calibrated directly in terms of conductance, giving a linear scale, a direct-reading resistance scale would be more useful; while a reciprocal scale could be used to read resistance directly, it would be very open at low values and very cramped at high values. On the other hand, the circuit of **Fig. 4** reads capacitance values directly. As the capacitance at Y_x is increased, R_v must be advanced pro-rata, not pro-reciprocal, to maintain balance. Thus, apart from some provision for nulling the "minor term", that is the quadrature or loss component, Fig. 4 is basically what is required for capacitance measurements.

For resistance and inductance, the variable facility, R_v and A_6, need moving to a position

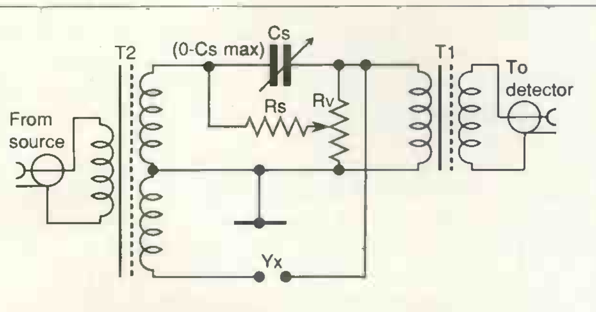


Fig.1. Reversing the connections to a conventional transformer ratio-arm bridge paves the way for a transformerless circuit arrangement.

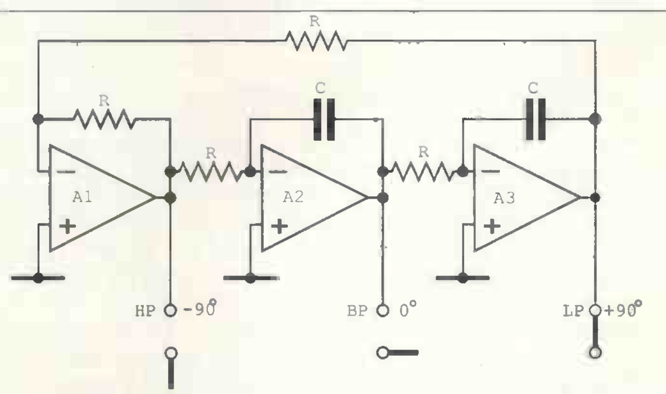
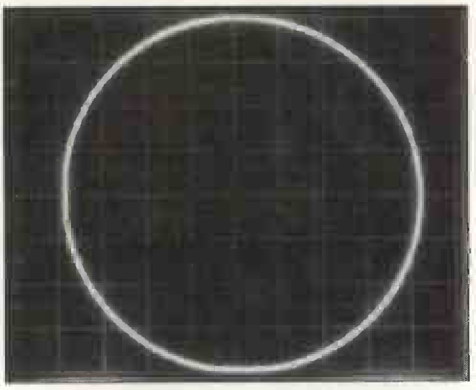
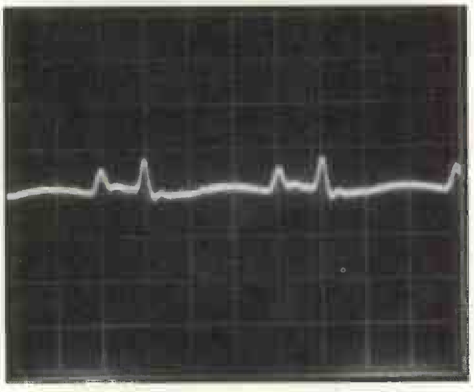


Fig.2. State-variable filter with zero damping and no input functions as an oscillator, but only just.



(a)



(b)

Fig.3a,b. Signals from oscillator 0° and +90°outputs, shown driving X and Y plates of an oscilloscope to display Lissajous figure, which indicates some clipping at left and bottom.

between S_1 and A_4, along with the ×10 step attenuator. Now, when R_v is set to zero, zero voltage is applied to Y_x, and so resistance and inductance read directly on a linear scale fitted to R_v.

At the expense of slightly more complicated switching, all of the foregoing can be arranged, as shown in **Figs 5, 6**. In Fig. 5, note that an 11Vpk-pk inverted (–180°) version of the 22Vpk-pk 0° output of A_{1b} is always provided by the inverting amplifier A_3, and that a quadrature component, leading or lagging as selected by S_3, can be added to it as required. In the resistance position of S_1, this waveform is applied to R_{15}, the 1MΩ resistance standard *STR*, which is connected to the virtual earth at the input of A_5. Meanwhile, a 0° 11Vpk-pk sinewave appears at the top of R_{13} and a proportion of this goes via A_4, the ×10 step attenuator S_2 and A_2 to the unknown resistor at the test terminals Y_x. With S_2 in position 1, resistors from 1MΩ down to zero can be measured with good resolution, since R_v is a ten-turn potentiometer fitted with a digital dial.

Nevertheless, the resolution would be too limited accurately to measure resistors in the low kilohms and ohms range, so S_2 selects other ranges down to 0-10Ω max in position 6. In the event that the 'resistor' under test has a significant reactive component, adjusting R_v alone does not produce a deep null to indicate complete balance. In this case, adding a cancelling quadrature component, by advancing R_{12} clockwise, i.e. in the direction indicated in Fig. 5, nulls the minor or quadrature term. Since, at maximum, this quadrature component can equal the in-phase component at the output of A_3, the instrument can measure 'resistors' with a phase angle up to 45° – and of course the output from A_3 will then be 3dB greater than 11Vpk-pk.

Inductance measurement is the same as that for resistors, with two differences. To allow for the 90° phase lag of the current relative to the applied voltage, when measuring inductors S_{1c} selects the output of A_{1c}, which is advanced by 90° relative to the A_{1b} output which was used for resistive unknowns. As before, S_3 and R_{12} allow for phase angles up to 45° from the ideal, i.e. for an inductor Q of down to unity – or even an inductance with a shunt negative resistance component! The other difference concerns the inductance standard *STL* (R_{14}); Fig. 5 shows a value of 10kΩ, which provides inductance ranges of 0-1H down to 0-10μH.

Capacitance measurements are made rather differently. Whereas for both *R* and *L*, the voltage applied to the unknown was adjustable both in steps (S_2) and continuously (R_v) with a fixed voltage applied to the standard, for *C* measurements the variable voltage is applied to the standard *STC* (R_{16}) while the voltage applied to the unknown capacitor is varied only in ×10 steps. To allow for the leading nature of the current through a capacitor, S_{1a} selects the lagging voltage from A_{1a} in place of an in-phase voltage. The result of this rearrangement is that again the digital read-out dial of R_v reads the value of the unknown *C* directly, as it did for *R* and *L*. Resistor R_{12} provides for balancing the capacitor's loss component, down to a tanδ of unity, or of course a capacitive susceptance including a negative conductance component.

Amplifier A_2 must be a special breed of op-amp, capable of driving capacitances up to 10μF. Many op-amps get very unhappy when faced with large capacitive loads – in this context, 'large' meaning a few hundreds or even a few tens of picofarads – and may oscillate unless special precautions are taken. Here however there are no problems, since A_2 is that remarkable op-amp the *TLE2027*. This was described in an earlier Design Brief [2], where it was shown driving 23Vpk-pk into 1μF at 318Hz. Here, it is required to drive up to 10μF at 1591Hz, but only at 110μVpk-pk in position 6 of S_1, or up to 11Vpk-pk in position 1 where the maximum capacitance load is only 100pF.

Indication

Any unbalance of the bridge results in current flowing in the virtual earth of amplifier A_5, which thus provides a signal to the detector stage, shown in Fig. 6. A 1MΩ log. potentiometer precedes a 40dB amplifier A_7, driving the loudspeaker, a 3Ω type with output transformer being used as it was to hand; a reasonably sensitive 64Ω speaker would do as well. The 2000μF capacitor C_6 provides additional smoothing for the 25V dc power supply, which is split into ±12.5V supplies for the op-amps by the *TLE2426* 'rail splitter'. Diode D_1 is a led 'On' indicator.

With the bridge measuring unknowns of

Transformer ratio-arm bridge without transformers

The transformer ratio arm bridge was described in an earlier Design Brief*, where its use in the conventional manner was described, i.e. with the detector connected to the centre-tapped balance transformer. Being a passive linear network, however, it can also be used 'back to front', with the source connected to the centre-tapped balance transformer instead, as seen in **Fig. 1**, although this simplified circuit will only measure resistors and capacitors (lossy or otherwise). It measures inductors either by arranging switching to connect C_s to the other end of the centre-tapped winding, or by connecting a fixed capacitor of value $C_{smax}/2$ in parallel with the unknown susceptance Y_x: the shunt inductive component of Y_x is then measured as an equivalent negative capacitance. Similar arrangements for R_s enable negative conductance components of Y_x to be measured.

At balance, there is no current through the lefthand winding of T_1, and so no voltage across it. Thus this winding represents a virtual earth and T_1 could nowadays be replaced by the virtual earth at the input of a suitably fast inverting op-amp; it would be nice to be able to eliminate T_2 also. It turns out that not only is this possible, but one can actually eliminate C_s as well.

Note that with the voltage applied to C_s in the phase shown in Fig. 1, balance can be achieved with a capacitive Y_x, while with a voltage in the opposite phase applied to C_s (from the other end of the centre-tapped winding of T_2, the two halves of which are perfectly coupled), inductive unknowns are catered for. If a voltage in quadrature were available, it would be possible to balance a resistive Y_x against C_s, or alternatively (much more useful) a reactive unknown could be balanced using a resistive standard. It was this thought that gave me the basic idea for the design of a universal laboratory *LCR* bridge, something I had been promising myself for some time.

Bridge operation

In basic principle, the circuit is simple, as illustrated in **Fig. 4**. A 0° phase signal is applied to the standard resistance R_s via R_v and A_6 and thence to the virtual earth at the input of A_5. Thus the effective value of R_s is adjustable over the range infinity down to $R_{smin}=R_s$, i.e. from zero conductance up to G_{smax}. In the resistance position of S_1, a 180° signal is applied to the unknown terminals Y_x, from the unity gain inverting amplifier A_4. When R_x is a resistance equal to the effective value of R_s (a conductance Y_x equal to $1/G_{s(effective)}$), the bridge is balanced, all of the current via R_s is just swallowed via Y_x by A_4, and no current flows in or out of the virtual earth of A_5. Therefore the detector registers no signal, indicating the point of balance. As R_s is effectively variable from infinity down to $R_{smin}=R_s$, unknown resistors down to the same value can be measured. If an attenuator with steps of ×1, ×10, ×100... is fitted between S_1 and A_4, resistors down to $R_{smin}/10$, $R_{smin}/100$ etc can be measured, extending the range of the bridge to much lower values of resistance, whilst at the same time, keeping down the output current demanded from A_4.

If the output of A_4 is made to lag by 90° on that shown, by selecting the C position of S_1, then the (leading) current through the capacitor will again be in antiphase with that via R_s, enabling capacitive susceptances to be measured. Similarly, in the L position of S_1, inductors can be measured.

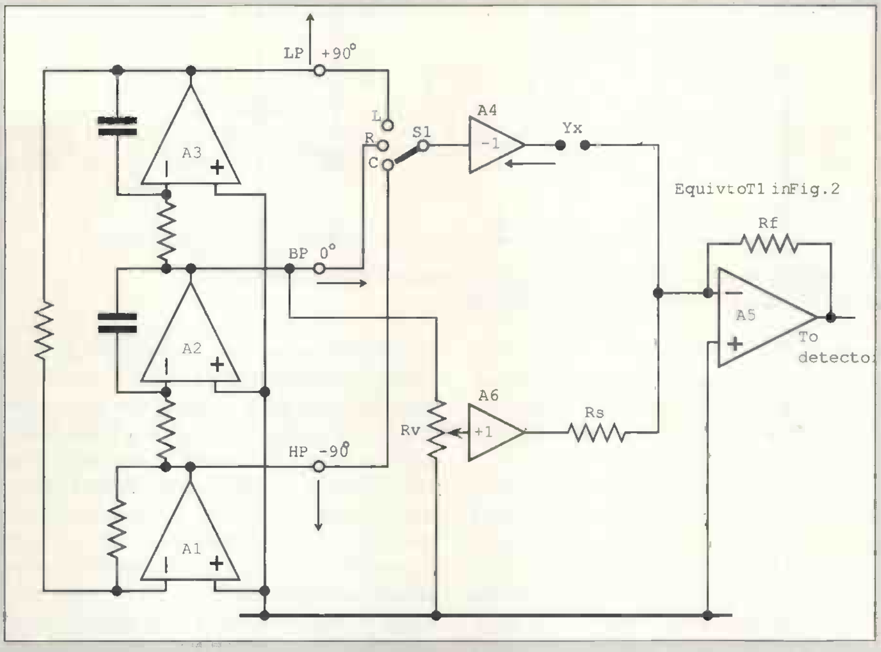


Fig.4. Basic, but flawed, circuit of new bridge. Good for pure rsistance or reactance, but offers no facility for nulling reactance of a resistor or losses in capacitors and inductors.

both high and low impedance, it was desirable to keep both electrostatic and magnetic hum fields out of the instrument's metal case, so I used one of the very inexpensive dc supplies built into a 13A plugtop case. As supplied, the circuit was as in **Fig. 7(a)**, but this was modified by removing the output voltage switch and just making room for an additional 470µF capacitor. After modification, the supply was as in **Fig. 7(b)**, its output permanently connected to the case of the bridge by a length of audio screened lead.

This pseudo-ratio-arm-bridge proved easy to use, providing a resolution of 0.1% of full scale on any range, thanks to the ten-turn digit dial on R_v. When measuring resistors, the setting of the quadrature control R_{12} is at or very near zero. The null obtained is deep and complete, with no sound audible in the loudspeaker other than a very slight trace of mains hum, due to the absence of stabilised supplies, which are, as it turns out, superfluous.

Measuring capacitors or inductors gives just as deep and complete a null as far as the fundamental is concerned, once the loss component has been nulled out with R_{12}, S_3 being in the C or L position as appropriate. However, as the reactance of a capacitor at harmonics of the drive waveform differs from that at the fundamental, whereas that of the capacitive

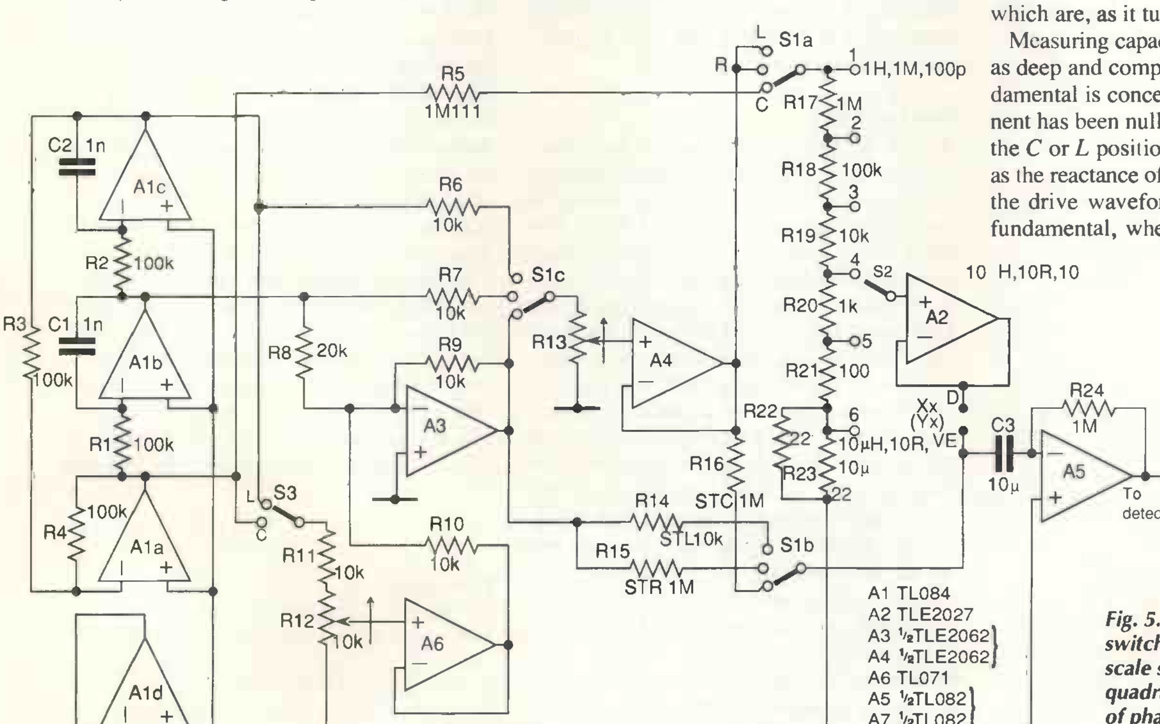


Fig. 5. Final bridge circuit, with switching to give a reasonable scale shape and to provide a quadrature term to take account of phase angles.

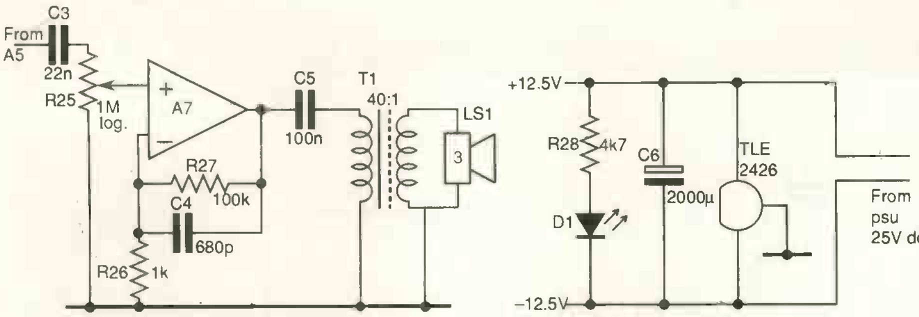


Fig.6. Bridge detector (a) and power supply (b), modified from a commercial plug-top dc supply to provide more smoothing.

standard *STC* (R_{16}) does not, some audio output at harmonic frequencies remains at balance. The ear easily distinguishes between the fundamental and the harmonic tone and thus a true balance is readily achieved.

When measuring an inductor, as its reactance rises with frequency, harmonics are not accentuated and, as with resistors, are inaudible at balance. Setting *STL* (R_{14}) at 10kΩ, rather than 1MΩ as for *STR* and *STC*, limits the maximum capacitance that can be measured to 1H, but provides ranges down to 10µH maximum, permitting in principle measurements down in the nanohenry range. To test low-inductance measurements, four turns were wound on a two-hole balun core type *FX2754*, which uses 3C85 material; the expected inductance, given the core's A_L of around 3500nH/turn, was 56µH.

On range 5 of S_2, the measured value was 52µH with the dissipation or quadrature control R_{12} set near to zero, indicating a high value of Q. Reducing the number of turns to one gave a measured value of 3.9µH with R_{12} set at about 30%, indicating a Q of 3, against an expected value of 3.5µH. Ideally, the measured value would have been one sixteenth of 52µH or 3.24µH, but the measured value includes the inductor's leads and the bridge's terminals and internal test circuit wiring.

Remember also that this bridge, like the transformer ratio-arm bridge from which it is derived, measures an unknown as a parallel combination of susceptance and conductance. When the Q is low, the inductor is effectively an inductance and a resistance in series; the impedance is higher than the reactance of the inductor alone. Measured in parallel terms, this appears as a rather higher value of inductance in parallel with an even higher value of resistance. If the quadrature control R_{12} were calibrated, a series/parallel conversion could be applied to the results, to obtain the actual value of inductance and its effective series resistance – and thence its Q.

This does, however, point up a limitation of inductance measurements carried out at such a low frequency as 1.59kHz. At this frequency, the Q of an air-cored inductor, or one with a slug but a return path in air, such as an rf choke, will be so low that balance will not be obtained in the *L* position of S_1, but only in the *R* position. On the other hand, small mains transformers may have a primary inductance of many tens of henrys. This being so, one might find it more convenient to use the 1MΩ *STR* also as the inductance standard *STL*, giving inductance ranges from 0-1mH up to 0-100H. An even better scheme would be two '*L*' positions on S_1, providing inductance ranges all the way from 0-10µH to 0-100H. Extending the idea even further, an alternative value of 10kΩ for *STC* would extend capacitance measurements up to 1000µF.

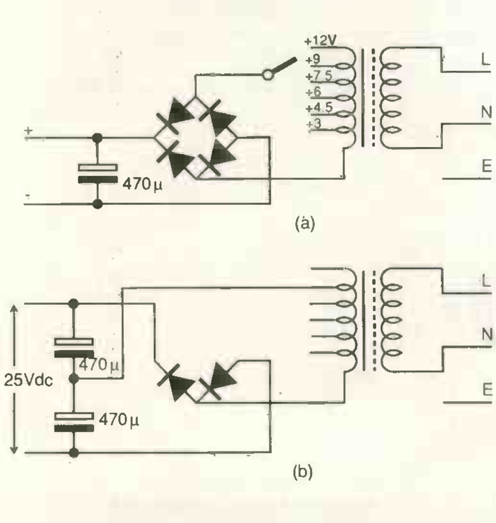


Fig. 7. Circuit of a commercial plug-top power supply, a), modified for voltage doubling, b).

Practicalities

A little attention to construction pays dividends in bridge performance. Recommended precautions are few, but necessary. Firstly, all the earth returns shown in Fig. 5, other than those associated with A_1 must go to a single star earth point, which can conveniently be the N (neutral) terminal; this was situated on the front panel, between and slightly below the D and VE (drive and virtual earth) terminals.

Secondly, resistors associated with S_1 and S_2 should be on the switches themselves and, untidy though it may look, connections from the switches and R_{12}, R_{13} routed directly in fresh air to the appropriate points on the circuit board – definitely no neat cableforms. Thirdly, the board layout should be such that A_2 output is as close as possible to the rear spill of the D terminal, say less than an inch of stout wire, and S_{1b} wiper should be returned direct to the rear of the VE terminal, from whence the lead to C_2. As far as accuracy is concerned, all resistors should be 1% or better, but more importantly R_6 and R_7 should both equal R_{13}, R_{11} should equal R_{12}, R_9 and R_{10} should each be half of R_8, and R_5 should equal the resistance from S_{1a} wiper to ground. The *SVF* oscillator frequency should of course be as close to $\omega = 10^4$ as possible.

Performance

I intended the bridge for use as a general-purpose lab. component bridge. Nevertheless, the instrument shares the same attribute as the transformer ratio-arm bridge – the ability to measure without error the series component of a pi network whose shunt arms are grounded. To verify this, a 56pF capacitor was measured, the value reading 56.5pF. After connecting 100pF capacitors from D and VE to N, repeating the measurement gave 56.4pF; so much for high impedance circuits.

Repeating the test on an 8.2Ω resistor gave 8.35Ω, and 8.3Ω after connecting 4.7Ω resistors between D and VE to ground. Sensitivity was noticeably reduced, but interpolating between the points at which the tone just reappeared each side of the null gave the reading.

This verifies that for both high and low impedances, the instrument can measure the series element of a pi network, even when the shunt arms present a lower impedance than that being measured. Thus for example, a component on a pcb can be accurately measured without disconnecting it, if the far ends of other components connected to it are grounded.

As a lab. instrument in occasional use, the bridge has proved very satisfactory. For more concentrated use, especially by unqualified personnel, an automated version would be preferable. Such a scheme could be readily implemented using voltage-controlled amplifiers, as follows. Detector stage R_{25} and A_7 would be replaced by two synchronous detectors, one driven from the 0° degree phase and one from the 90° phase. The dc outputs of the synchronous detectors would be filtered and amplified, and fed back to control two vcas, fed with the said 0° and 90° degree signals. Given that the vca output was linearly proportional to the control voltage, as is the case for a four-quadrant multiplier, the two control voltages represent the real and imaginary components of the unknown directly, and could be indicated on digital panel meters.

It is true that the results will be the components of the equivalent shunt representation of the unknown, but often this will not matter. For resistive components with a phase angle at the test frequency of less than 5.3°, capacitors with a tanδ of less than 0.1 and inductors with a Q greater than 10, the error in the value of the component as indicated will be less than 1%. ■

References

1. Hickman, I. The transformer ratio arm bridge. *Electronics World+Wireless World*, August 1994, pp 670-672.
2. Hickman, I. Differentiating op-amps. *Electronics World+Wireless World*, April 1994, pp 322-325.

poles&ZEROES

Looking at s-plane poles and zeroes can help circuit designers choose the best – and cheapest – solution, as Steve Winder explains.

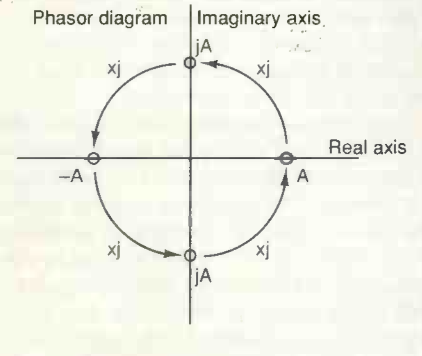


Fig. 1. Phasor diagram. Zero is at the centre. The horizontal axis is real while the vertical is imaginary.

Poles and zeroes are often used to describe certain analogue functions. These include filters, phase-locked loops and control systems. The concept of poles and zeroes has not always been very well explained and this article is an attempt at an introduction to the subject.

Even the most simple of circuits can be described in terms of its output response to a certain stimulus at the input. The output voltage divided by the input voltage is called the transfer function; it is a mathematical expression that describes how the input is transferred to the output.

The transfer function can be used to describe the time response; this is usually described in terms of the response to a short impulse at the input. Alternatively, the transfer function can be expressed in terms of its response to the frequency of the input signal, and is usually described in terms of a sine wave applied at the input. The time domain and the frequency domain responses are inter-related, and one can be transformed into the other using a Laplace transform.

Poles and zeroes do not exist. They are just a mathematical concept associated with the Laplace equations that describe a circuit's response in terms of the 's plane', which will be discussed later. Unfortunately for the beginner trying to understand the subject, engineers often think of poles and zeroes as having an entity of their own. For example, when a text says that 'a pole has been created by adding a capacitor at the input...' it is hard to imagine what this pole is. However, if the text read 'the frequency response has been changed by adding a capacitor at the input... and this can be modelled by creating a pole at position x in the s plane', it would require more words, but it may be clearer.

Suppose that the transfer function of a circuit is $1/(1+s)$. Do not worry about what s is for now. If there was a chance that s equals -1, the denominator $(1+s)$ would be equal to zero and the transfer function would be infinity. The s plane model of the transfer function would say that there is a pole at $s=-1$. More complex transfer functions may have an $(s+1)$ in the numerator (above the divide line) and s^2, or higher, in the denominator of the equation. In this case the transfer function would be equal to zero when s equals -1. It is hardly surprising, then, that the s plane model of the transfer function would say that there was a zero at $s=-1$. To find the poles and zeroes you have to factorise the equations in terms of $s+x$ or $s-x$ as appropriate. As with any equation, if the numerator and denominator (top and bottom, respectively) of the transfer function have factors that are equal, these can be cancelled. In other words a pole and a zero at the same location cancel each other.

Now, you may be thinking why use poles and zeroes at all. Poles and zeroes do have their uses when more complicated circuit responses are described. In the case of a filter, pole positions are often used to calculate component values. The design of transitional filters is based wholly on pole positions; by looking at the positions of poles in Bessel and Butterworth designs it was possible to place poles to produce a hybrid design that had features of both 'parents'. Poles are also very useful in describing the impulse response of a circuit. In control systems the position of a pole can indicate the degree of stability that the system has. Some engineers design circuits in terms of the poles and zeroes in the s plane, then select the components and circuit configuration that can be modelled by them.

The important step is to describe the s plane, the home of poles and zeroes. Before that, let's have a review of complex numbers and how they work.

Complex numbers

Complex numbers have real and imaginary parts. One way to think about complex numbers is by using a graph. Imagine a graph having two axes crossing at the centre, the point where the axis cross is known as the origin. The horizontal axis represents real numbers, positive to the right and negative to the left. So a point at +5 on this axis would be some way to the right of the origin. The vertical axis represents imaginary numbers which have a label j. This axis is positive above the origin and negative below it. The number j5 represents a point some distance above the origin.

Movement of points along the real and imaginary axis are 90° apart. A complex number with real and imaginary parts can be represented by, say, 3+j4. To plot the point represented by this number move along the real axis three units to the right of the origin, then move vertically by four units. The final point is five units from the origin, by Pythagoras, and about 53° from horizontal using the tangent rule (tanq=opposite/adjacent).

Let's now do some mathematics using the graph. If we subtract 3 from the point at +5 it moves left towards the origin and ends up at +2. This is a simple concept that is easy to grasp. Simple mathematics can also be used to add and subtract imaginary numbers, eg j5+j2 = j7; the point moves further from the origin, along the imaginary axis. A complex number is formed when an imaginary number is added to a real number; doing so moves the point representing the real number vertically above its position on the real axis.

A real number can be converted into an imaginary number by multiplication. Although j was described as a label, it is actually multiplied by the real number to make it imaginary (j is used by electrical engineers: mathematicians use i for imaginary). When a number is multiplied by j this is the same as introducing a 90° rotation about the origin. To understand this notation consider the phasor diagram given in **Fig. 1**, which is a graph with zero at the centre, a horizontal real axis and a vertical imaginary axis. Place a point on the positive real axis to represent a real number then, keeping the same radius from the centre, rotate 90° anti-clockwise until it is above the imaginary axis: this action is the same as multiplying the real number by j. If this action is repeated the real number is multiplied by j squared and the point ends up above the negative real axis, or –1 times the original real number. Since j squared was equal to –1, j must be the square root of –1. This is important; but is also a difficult concept to grasp, because one of the first things learnt at school is that you cannot take the square root of a negative number.

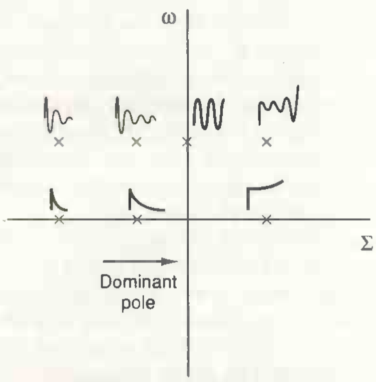


Fig. 2. S-plane complex frequency diagram with two axes. Real numbers are on the horizontal Σ axis and imaginary numbers on the vertical ω axis.

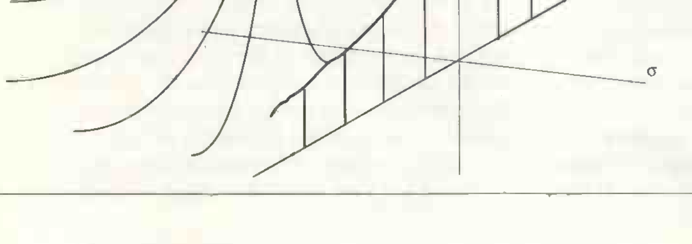
Practical use of complex numbers

Consider what happens when a sinusoidal voltage is applied across a passive component. In terms of the resultant current flow through the component, a real current is one that is in phase with the applied voltage. If the current through a circuit is entirely real, then the impedance must also be real, since $Z=V/I$, and the impedance is resistive. Power dissipation is given by $P=V\times I$.

Now consider an imaginary current. Impossible? No, this current does exist; only it is 90° out of phase with the applied voltage. This occurs when the component is an inductor or capacitor. In the case of the inductor, the applied voltage is 90° leading the current flow, and since $Z=V/I$ the impedance is $+j\omega L$, where ω is $2\pi F$; the frequency of the applied voltage. A capacitor is just the opposite. Applied voltage lags the current flow and the impedance is $j/\omega C$. If the current is imaginary, no power is dissipated in the component (hence the imaginary impedance) but energy is stored.

The storage of energy in reactive components can be seen in two ways. The first way is a dc effect. This is illustrated by the fact that a capacitor stores charge and is used to smooth power supply ripple. The magnetic energy stored in an inductor results in the sometimes dangerous back emf that is produced when the current path is broken. A safe way to demonstrate this is to connect a mains neon indicator across a high value inductor, say 1H, then briefly connect a carbon battery across the terminals; when the circuit breaks, the neon flashes. This is also the basis of car ignition systems.

Fig. 3. In this description of a first order filter, the s plane is three dimensional and the pole can be imagined as the pole of a tent.



Another illustration and application of the stored energy in inductors and capacitors, but one that is harder to visualise, is resonance. This is an ac effect. Resonance can occur when inductors and capacitors are connected in series or in parallel. High current flows or voltages are produced, depending whether the circuit is series or parallel connected. Stored energy is transferred from one component to the other and resonance occurs when the energy stored in the capacitor is equal to that in the inductor. In other words, resonance occurs when the magnitude of their reactances are equal; in a series connected circuit the $+j\omega L$ of the inductor cancels out the $-j1/\omega C$ of the capacitor and allows a large current to flow. Any circuit resistances (such as winding resistance) cause the oscillations to be damped, reducing the current flow. This illustrates an important feature of complex numbers; when they are added, the real parts and imaginary parts are added together separately. The imaginary parts cancel each other out if they are equal in magnitude but are of opposite sign.

A complex number can be converted into a number with magnitude and phase. Some calculators have this process, and its inverse, built into a function key operation. The magnitude of a complex number is the square root of (the real part squared added to the imaginary part squared); the phase angle is the arctan of (the imaginary part divided by the real part). To see where this may be useful consider some calculations involving an inductor. Inductors are never pure inductance (superconductors excepted) because the windings have some resistance, so the measured phase difference between the applied voltage and the current flow is never 90°.

To find the power dissipated in this inductor when a certain ac voltage is applied across it, we must find the magnitude of the impedance so that the current, and hence power dissipated, can be calculated. If the values of inductance and resistance are known, the impedance is complex and is given by: $Z=R+j\omega L$. The magnitude and phase angle of the inductor's impedance can be calculated: the magnitude of impedance is the square root of ($R^2+\omega L^2$); the phase angle is arctan ($\omega L/R$). The smaller R becomes, the more the inductor behaves like a pure inductance. The power dissipated in the winding resistance is given by $P=I^2R$, the current being found by dividing the voltage by the magnitude of the impedance.

If the values of inductance and resistance are not known, they can be calculated from the magnitude and phase angle of current through the inductor. The resistance is given by the impedance magnitude multiplied by the cosine of the phase angle. The inductance can be calculated by multiplying the impedance magnitude by the sine of the phase angle, then dividing by (2π multiplied by the frequency of the applied voltage).

The s plane

The s plane is a complex frequency diagram with two axis. Real numbers are on the horizontal axis which is labelled Σ. Imaginary numbers are on the vertical axis that is labelled jω. The signals represented by various points in the s plane are shown in **Fig. 2**. A signal represented by a point far to the left of the imaginary jω axis quickly decays. As the points approach the imaginary jω axis the signals they represent approach a steady state response. Any point to the right of the jω axis represents an exponentially increasing signal, hence instability. If the point lies on the real Σ axis there is no oscillation associated with the signal. Signals with some oscillation in their waveform are represented by points in the s plane above and below the real axis. As the frequency of the oscillation increases, so the points on the s plane are further from the real axis.

If any signal can be represented by the sum of a number of exponentially decaying sinusoids then any signal can also be represented by the sum of a number of points in the s plane. The s plane, then, is a graphical way of describing a complex signal.

The s plane arises from the use of the Laplace transform, which is similar to the Fourier transform. The purpose of both transforms is to take a time domain signal and convert it into the frequency domain. The Fourier transform works for repetitive signals, which is useful for calculating the harmonic content because it sums an infinite number of sinusoidal signals. The Laplace transform is slightly different in that the signals being summed are exponentially decaying sinusoids. This allows for signals that are not continuous and have zero amplitude before the interval being considered.

An exponentially decaying sinusoid can be expressed as e^{-st}. The s is a complex frequency given by Σ+jω. If this equation is expanded we get $e^{-(\Sigma+j\omega)t}=e^{-\Sigma t}e^{-j\omega t}$. The $e^{-\Sigma t}$ part is the exponential decay while the $e^{-j\omega t}$ part is the steady state oscillation. If Σ=0, the result is that the signal expression is $e^{-j\omega t}$ which is a continuous sinusoid and the Fourier and Laplace transforms are equivalent.

Describing components in the s plane

One reason why the s plane representation of signals is used, is the ease of describing components in the s domain. When a steady sinusoidal signal is applied across a capacitor or inductor the current that flows is 90° out of phase with the voltage and the reactances are said to be imaginary. A purely imaginary point in the s plane is when s=jω. An inductor's reactance, jωL, becomes sL and a capacitor's reactance, 1/jωC becomes $1/sC$.

Real poles

Consider a simple *RC* filter. A resistor is in series between the signal source and the load; a capacitor is connected across the load. There is a common connection from one side of the signal source to the load, and this is the reference for measurements. The resistor and capacitor form a potential divider. As the frequency of an applied signal increases, its amplitude across the load reduces. If the load has a high impedance, so that changes to the circuit currents are insignificant, the transfer function that describes the circuit is: $(1/sC)/(1/sC+R)$, which simplifies to $1/(s+1/CR)$. Now, CR is the time constant of an RC network, sometimes described by τ. The denominator becomes equal to zero when $s=-1/CR$ and this makes the transfer function equal to infinity; a pole is said to have been created. The pole is entirely real because there are no oscillatory components in the circuit.

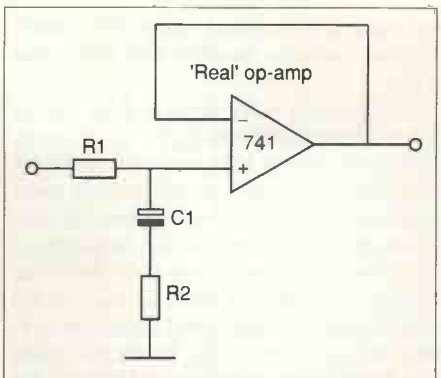


Fig. 4. Compensated first order filter having a lead-lag network at its input.

Now if an impulse is applied to this circuit we can intuitively see that the output will be stepped with an exponential decay. The decay rate depends on the time constant of the circuit, given by τ=CR, as the charge stored in the capacitor C drains away through the resistor R.

If a sinusoidal signal is applied to the *RC* filter circuit, the half power (–3dB) frequency occurs when the capacitor's reactance is equal in magnitude to the resistance. This is when ω, or $2\pi f$, is equal to $1/CR$. The frequency f is known as the cut-off frequency. It is, at the moment, hard to see the relationship between the pole position and the frequency response. The pole is on the negative real axis in the s plane while the frequency axis is vertical.

The key to visualising the effect of the pole position on the frequency response is to imagine a canvas tent. The pole behaves like the pole in a tent, it holds up the enveloping canvas. This is where the s plane becomes three dimensional, see **Fig. 3**. Consider the height of the canvas at points along the imaginary jω axis. It is highest at the zero frequency point, where it crosses the real Σ axis. It falls slowly in amplitude to start with and then by greater amounts until its rate of fall is proportional to the distance moved along the jω axis. The height of the canvas at the Σ axis is the dc attenuation which is zero; the attenuation at other points along the jω axis is relative to this. At high frequencies, attenuation increases by 6dB/octave, or in other words, doubling the frequency halves the output voltage.

Increasing the circuit's time constant makes it slower in response to impulses. It also has a lower half power frequency. You may have noticed that the distance of the half power frequency along the imaginary axis was the same as the distance of the pole from the origin, along the real axis. This was not coincidence, but by Pythagoras Theorem the distance from pole to half power frequency is √2 times the distance to zero frequency. The amplitude falls in proportion to distance, hence falls by 1/√2, or 3dB.

So now we have a way of converting a single real pole into a frequency response. A scale drawing of the s plane can be used, by measuring the distance from pole to origin and pole to frequency. The amplitude at any frequency is the input signal multiplied by the ratio of the pole to origin distance divided by the pole to frequency axis distance. Alternatively the ratio can be calculated.

Complex poles

An LC lowpass filter comprises a series inductor and a shunt capacitor, with resistive source and load. The combination of the two networks gives a potential divider. The transfer function, V_{out}/V_{in}, has a quadratic equation (of the form as^2+bs+c) in the denominator. This has two solutions for s which make the denominator equal to zero. Both solutions are complex, so the two poles do not lie on the real axis in the s plane but they are symmetrically placed about it. Each pole position is described by real and imaginary parts. The real part is the same for both poles and describes the amount of damping in the circuit. The magnitude of the imaginary part is the same for both poles, but one has a positive sign and the other is negative, and describes the frequency of oscillations produced if the circuit is excited by an impulse. Remember that it is the component values which determine the poles in the equation and the response of the circuit. The pole positions in the s plane are just another way of describing the circuit's transfer function.

Not surprising then, that the simple *LC* circuit described above is known as a two-pole filter. Filters using more reactive components are known as three-pole, four-pole, etc, as the number of components increases. The number of poles describes the filter 'order'. Odd order filters have one real pole and a number of complex poles depending on the filter order. Even order filters have only complex poles.

Active two-pole filters use feedback from the output to create a resonant circuit which is modelled by complex poles. Care has to be taken with active filters because their operation depends upon the response of an operational amplifier. Filter design is based on the amplifier having a perfect response; an output with constant amplitude and in-phase with the input. Peaking and ripple in the passband can occur if the op-amp response is poor.

To find the frequency response of a circuit that has complex poles, follow the same procedure as with the real pole. The response at frequency ω is given by the product of all the pole to the origin distances divided by the product of all the pole to jω axis distances. If the poles are equally spaced and lie on a semicircle, the product of distances to points along the imaginary axis remains almost constant, up to the point where the semicircle crosses the imaginary axis. A circuit represented by

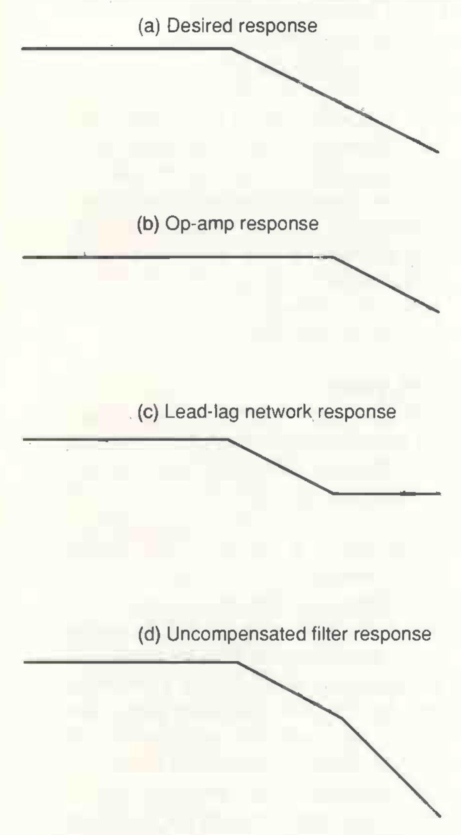


Fig. 5. Bode plots illustrating frequency compensation.

this arrangement of poles is known as a Butterworth filter and has a response that is maximally flat in its pass-band.

Zeroes

Simple *CR* circuits, which are described as having a real pole, also have a zero. The zero is never drawn on an s plane diagram because it is located at minus infinity on the real axis. Consider the transfer function $1/(s+1/CR)$. The pole is at $s=-1/CR$, because this makes the transfer function equal to infinity. However, as the value of *s* becomes larger, the transfer function becomes smaller and approaches zero.

Now consider what happens if the zero is moved closer to the pole on the negative real axis. The frequency response of the circuit that is represented by this arrangement of poles and zeroes will show a decay to start with, due to the pole; and then, as the frequency is increased further, the response will flatten out because the effect of the zero cancels the effect of the pole. Moving the zero closer to the pole can be achieved by inserting a resistor between the shunt capacitor and ground, this will be explained in detail later. Ultimately, by placing a zero at the same point in the s plane as an existing pole, the effect of the pole is removed.

Zeroes are also produced when the equation that describes a circuit response has a frequency dependent element in the numerator. Take the case of an elliptic lowpass filter, this has parallel *LC* elements in series with the source and load. When the resonance frequency of the *LC* circuit is reached it presents a high impedance so the signal is severely attenuated. There is said to be a zero at the resonant frequency; in the s plane this zero is on the positive and negative imaginary axis. Choosing the zeroes carefully lets harmonics of certain passband signals be removed.

Just as it was possible to calculate the circuit's frequency response from measurements of the pole positions in the s plane relative to the frequency axis, it is also possible to determine the effect of zeroes in this way. The frequency response is a constant, multiplied by the zero to frequency distance and divided by the pole to frequency distance. In the case of the real pole and real zero being close together, the response at high frequencies is flat; this is because the distances from the pole, and from the zero, to the frequency axis are approximately equal and the transfer equation equals a constant.

A pole-zero application

A previous article[1] described how lowpass filters were dependant upon the type of op-amp used. The article showed how the performance of a Sallen and Key low-pass filter was very dependent on the gain-bandwidth product of the op-amp used and on the order of the filter. It was also shown that the filter type has a bearing on the performance; Butterworth and Chebyshev (0.5dB ripple) types were used as examples. The Chebyshev filter has a steeper skirt response than the Butterworth type and is more demanding on the op-amp. The criterion used was for an accurate cut-off frequency and for no more than 2dB attenuation in the pass-band. There are times when a more demanding specification is required, e.g. the phase may be important.

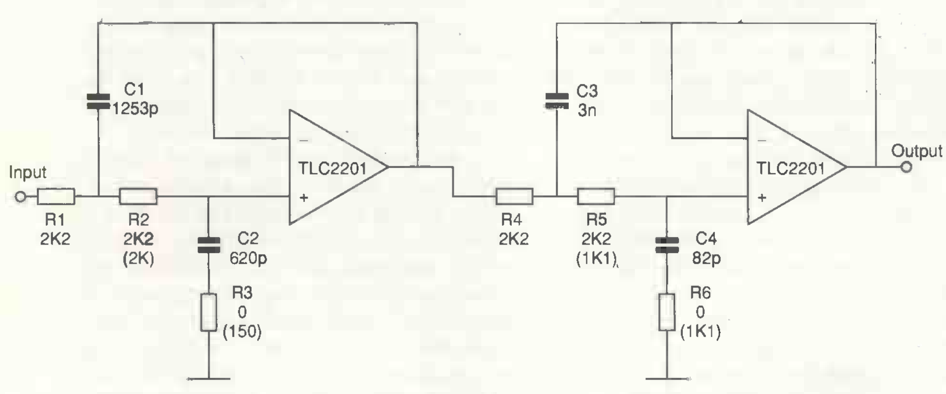
If two circuits are connected in series the overall frequency response is the product of the two individual responses. The frequency response of a first order low-pass active filter will therefore be a product of the *RC* network response and the op-amp's response. This is valid since the high input impedance of the op-amp will not affect the response of the *RC* network. If the op-amp has a perfect response within the filter pass-band then the final product will behave correctly. If the op-amp's response produces a change of amplitude or phase within the filter's pass-band, then the filter produced will have an incorrect cut-off frequency or, in the case of higher order filters that use feedback, will have ripple in the pass-band.

Using op-amps with a very high gain-bandwidth product may provide the solution to obtaining a near perfect response. There may be several disadvantages of this approach: the circuit's physical layout may be critical to avoid oscillation; the device may draw a higher than desired supply current; the device may be too noisy; or it may cost too much. High speed devices are often expensive and power hungry, and many have a minimum gain for stability.

It is much harder to design circuits when power sources are limited, such as when circuits have to run off batteries. If low-cost, low-power or other op-amp devices with insufficient gain-bandwidth product have to be used there are a few options that could be considered. One option is to design the filter with a cut-off frequency that is higher than required, hoping that the finished circuit will perform closer to the actual requirement. This is not an ideal solution and may be time consuming to develop. This solution may be made easier by the use of an analogue circuit simulator, such as *ECA2*[2] or *Analyser III*[3], allowing many designs to be tried before buying components. Another option, which will now be explained, is to modify the filter design by adding components that compensate for the op-amp's frequency response.

Operational amplifiers have an in-built feedback capacitor, for stability, and that causes the frequency response to be limited. A modern op-amp suitable for a filter is the *TLC2201* which has a gain-bandwidth product of 1.8MHz. This op-amp is a low-noise device and operates on single or dual 5V supplies. The effect of the in-built feedback capacitor

Fig. 6. 150kHz 0.5dB Chebyshev lowpass filter. Resistors $R_{3,6}$ are zero ohms in the uncompensated filter, otherwise, $R_{2,5}$ are reduced by the value of $R_{3,6}$, respectively.



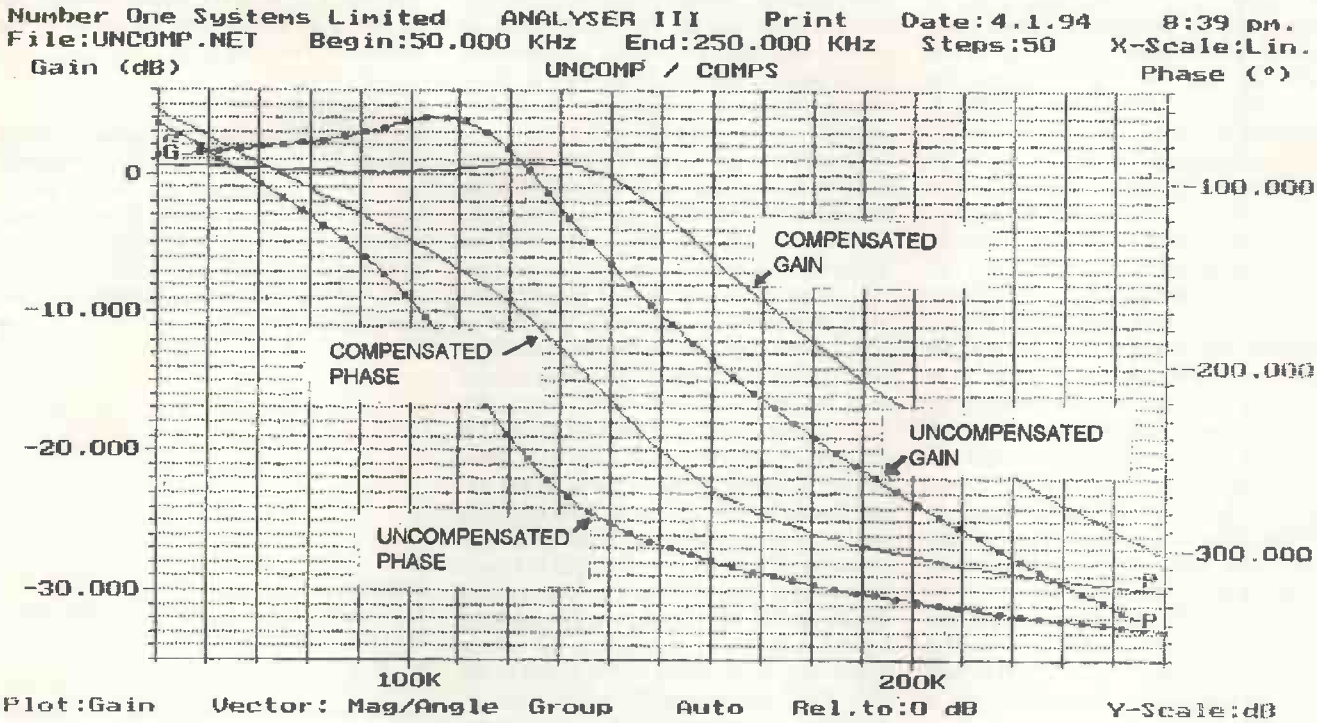


Fig. 7. Simulated responses of compensated and uncompensated circuits.

can be modelled by a pole in the s plane, placed on the negative real axis at $-2\pi(1.8^6)$, the same distance from the origin as the bandwidth. To remove the effect of this pole, it is necessary to place a zero at the same frequency. In other words, the filter roll-off must be stopped where the op-amp roll-off starts.

To overcome the effect of the op-amp's internal lowpass filter, it is necessary to modify the response of the *RC* network. What is required is a circuit having a falling response above the desired cut-off frequency, but a flat response beyond the amplifier's own cut-off point. This can be done by adding a resistor in series with the shunt capacitor at the amplifier's input, as shown in **Fig. 4**. This is known as a lead-lag network, because of its phase response, and is often used in control engineering or in phase locked loops. In terms of the pole-zero explanation, the extra resistor causes the zero to be moved from infinity and placed on top of the op-amp's pole; this effectively cancels the pole. The transfer function of the modified *RC* network is

$$(R_2+1/sC)/(R_1+R_2+1/sC).$$

By cross-multiplying, this can be re-written as

$$(s+1/R_2C)/\{s+1/(R_1+R_2)C\}.$$

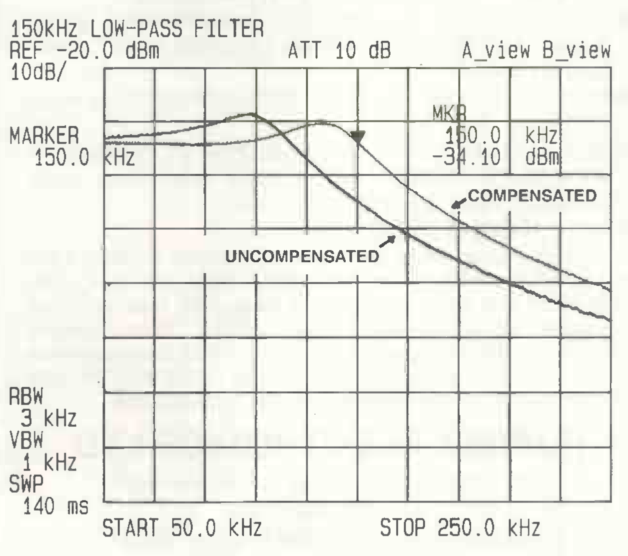
Examination of this transfer function shows that there is a real pole at $s=-1/(R_1+R_2)C$ and a real zero at $s=-1/R_2C$.

Instead of the input signal decreasing indefinitely, as the capacitors reactance falls, the signal's minimum level is limited by the potential divider action of the two resistors. In effect, with this modified circuit, the *RC* network provides part of the overall filter response and the op-amp provides the rest. This is illustrated in **Fig. 5**, using Bode plots (straight line approximation to a frequency response). This is rather a simplistic diagram because it does not show the phase response, which is also important. In the compensated network, shown in Fig. 4, resistor R_2 is given by the formula: $R=0.5\pi FC$. Frequency F is the unity gain bandwidth of the amplifier.

The value of R_1, found earlier, must be reduced by the value of R_2 because it has been shown that the pole position in the compensated circuit depends on the sum of R_1 and R_2. With no compensating resistor the pole position was just dependant on the value of R_1. Since the unity gain bandwidth has a high tolerance, 20% or more, R_1 and R_2 may need to be trimmed if the application is critical.

In a simple phase locked loop filter, comprising a *CR* network, the phase shift through the circuit increases to a maximum of 90° at high frequencies. This can cause oscillations if other circuits in the loop also introduce phase shifts, since a 180° phase shift in a negative feedback circuit is equivalent to positive feedback. The compensated filter (or lead-lag network) has a phase shift which increases to about 45° at the cut-off frequency, then decreases to 0° at high frequencies. By this action, the chance of loop instability is reduced.

Fig. 8. Frequency response of compensated and uncompensated filter circuits.



The compensation idea may be extended: as a more complex example, a fourth order Chebychev lowpass filter (with 0.5dB ripple) using the *TLC2201* op-amp was designed to have a 150kHz cut-off frequency, see **Fig. 6**. A fourth order filter has two pairs of complex poles, but the real part of these is set by the combination of R_2 and C_2 for one pair of poles, and R_5 and C_4 for the other pair. Compensation can be carried out in exactly the same way as before; by changing the values of R_2 and R_5, and by introducing resistors R_3 and R_6 in series with capacitors C_2 and C_4.

This filter was simulated using the Analyser III electronic circuit analysis program[3]. The uncompensated circuit showed peaking in the frequency response, just below the cut-off frequency. The values of the compensating resistors, R_3 and R_6, were then calculated and added into the netlist. Resistors R_2 and R_5 were reduced by an equivalent value, to maintain the correct filter cut-off frequency. In both cases, the nearest preferred values were used. The result of adding this compensation was that a correct filter response was obtained. A graph showing the simulated response of both compensated and uncompensated circuits is given in **Fig. 7**.

The graph in Fig. 7 also shows the phase response. Compensation has improved the phase response, which is linear with frequency up to about 100kHz. Previously the rate of phase change began to increase at frequencies above about 75kHz. A linear rate of phase change indicates a constant delay for all frequencies passing through the circuit. The effect of a non-linear rate of phase change is to cause the broadening of impulses passing through the circuit, which may be undesirable.

Having simulated the circuit and proved that the compensated filter would give a satisfactory response, the circuit was breadboarded using Veroboard. A frequency response of both compensated and uncompensated filter circuits is given in **Fig. 8**. The compensated filter has an accurate cut-off frequency, but shows a slight peaking in the pass band. This peaking did not show up in the circuit simulation and is due to the tolerance in the op-amp's gain-bandwidth product. This tolerance may be as high as 50% and its effect can be counteracted by fine adjustment of the compensating resistors. Another important point is that a practical filter must have its calculated input resistor value reduced by the source impedance, because this becomes part of the filter. The 2.2kΩ resistor, R_1, shown in the circuit diagram was actually a 600Ω impedance signal generator and a 1.6kΩ resistor, connected in series.

There are some good books that cover this subject in greater detail[4,5] for those of you who want to find out more.

Conclusions

The s plane description of a circuit is useful because the positions of poles and zeroes indicate both the impulse and the frequency response. The practical application is a useful one because it allows low-power or low-cost operational amplifiers to be used in filter circuits that would otherwise demand high performance devices. ■

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2. Those Engineers Ltd, *ECA2*.
3. Number One Systems Ltd, *Analyser III*.
4. R J Maddock, *Poles and Zeroes*, Holt, Rinehart & Winston.
5. P Lynn, *An Introduction to the Analysis and Processing of Signals*.

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Discrete active devices

High-power mosfets. Motorola's high-voltage TO-264 family of n-channel TMOS power mosfets dissipate 300W internally and control loads up to 5kW, 500V, 600V and 1000V devices now being available. The higher power reduces the need to parallel devices for high current and low R_{DS}(on) avoids the need for series gate resistors. On resistance for the 1kV version, *MTY10N50E*, is 1300mΩ at 10A, that of the 600V *MTY25N60E* 210mΩ at 25A and of the *MTY20N50E* 240mΩ at 20A. Motorola Inc. Tel., 0908 614614; fax, 0908 618650.

Digital signal processors

PAL/NTSC genlock. Raytheon's *RC6100* horizontal line genlock chip is meant for digital video signal processing and conferencing, supporting both PAL and NTSC formats. It provides a high-speed tracking sync. separator, a glitch filter and pixel clock generation and forms the timing reference for analogue acquisition and reconstruction. Input composite video can be in a choice of eight pixel-clock frequencies, since the chip supports CCIR601, $4F_{sc}$, square pixels and VGA in both formats. Ambar Components Ltd. Tel., 0844 261144; fax, 0844 261789.

Small, high-voltage mosfet. A TO-215AA mosfet by Ixys, the *IXTU01N80* has a blocking voltage of 800V, drain current of 100mA and on resistance of 80Ω. It is claimed to be the smallest available at 800V. IXYS Corporation. Tel., 0101 408 982 0700; fax, 0101 408 496 0670.

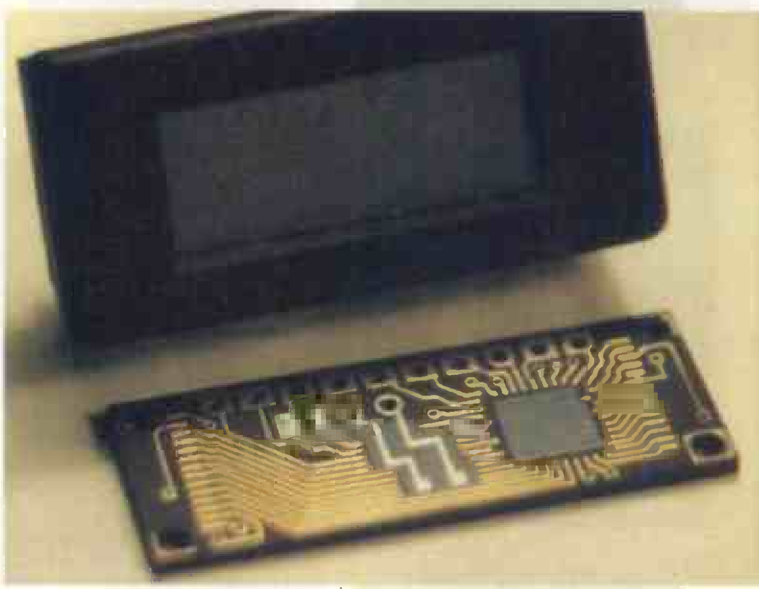
Medium-power mosfet. *ZVP4424* by Zetex is a TO92 p-channel power mosfet rated at 240V, 200mA and exhibiting an on resistance of 11Ω for a gate/source voltage of –3.5V. The range of gate/source voltage is ±40V and, since pulsed-current handling is up to 1A, the device is suitable for use as a telephone hook-switch or other interfaces where transients are to be expected. Its 1.4V threshold for a –1mA drain current makes the device a good interface between high voltage circuitry and standard logic. Turn-on and turn-off delays are 1ns and 26ns. Zetex plc. Tel., 061-627 5105; fax, 061-627 5467.

Low-noise jfet. Claiming it to be the lowest-noise dual jfet available, Linear announces the *LT1169*, which generates a noise voltage of 6nV/√Hz and a maximum input bias current of 10pA held over the –11V to 13.5V common-mode range. Input impedance is 10^{13}Ω; input capacitance 1.5pF. This device is unconditionally stable with capacitive loads of up to 1000pF; it has a 0.5mV input offset, a gain of 4 million, slew rate 2.4V/µs and gain/bandwidth product of 3.3MHz. Linear Technology (UK) Ltd. Tel., 0276 677676; fax, 0276 64851.

RF, n-p-n power. Six new power transistors by Motorola provide rf power in the 800-960MHz frequency range of 2-36W. *MRF857* is a 2.1W cw type with 3.3pF output capacitance in a stud package and *MRF862* the 36W device showing 75pF at the output. Between the two are *MRF857S, 858, 860 and 861* in various packages. Motorola Inc. Tel., 0908 614614; fax, 0908 618650.

Linear integrated circuits.

3.3V low dropout regulators. A second family of 3.3V low dropout regulators from National is now available. *LP2952-3.3* and *LP2953-3.3* deliver 250mA and show a 470mV dropout at full load, quiescent current at a load of 1mA being 130µA. Both devices have a 3.3V tap, eliminating external resistors. National Semiconductor GmbH. Tel., 01049 814110382; fax, 01049 814103515.



High-speed op-amps. True voltage-feedback op-amps from Linear, *LT1361/2* and *LT1364/5* are dual/quad types offering high speed while maintaining good dc accuracy. *LT1361/2* have 50MHz gain/bandwidth and slew at 800V/µs with a 1mV input offset and 1µA input bias, while *LT1364/5* are 70MHz types slewing at 1000V/µs with 1.5mV and 2µA. Differential gain for the 1364/5 into 150Ω is 0.06%; diff. phase is 0.04%. Linear Technology (UK) Ltd. Tel., 0276 677676; fax, 0276 64851.

SM voltage regulators. *ZMR250* and *ZMR500* fixed voltage regulators by Zetex provide 2.5V and 5V respectively, with thermal shutdown and current limiting, both types being in SOT23 SM packages. In standby, the devices take 25µA and 50µA, are unconditionally stable and need no external components. Both types supply 50mA and input stabilisation is 10mV worst case for voltages between 4.5V and 20V(*ZMR250*) or down to 7V for the *ZMR500*. Zetex plc. Tel., 061-627 5105; fax, 061-627 5467.

Low dropout regulators. Low-dropout voltage regulators from Semtech in the *EZ108X* range are pin-compatible with Linear's *LT1083/4/5/6* devices. Input/output differential is low; input is up to 6V and the output is either 3.3V fixed or 1.3-4V adjustable at currents between 1.5A and 7.5A. Regulation and stabilisation are 0.1% and 0.015%. Current limiting and thermal shutdown are provided. Semtech Ltd. Tel., 0592 773520; fax, 0592 774781.

Smallest DPM. Using a thick-film hybrid microcircuit manufactured by *CorinTech*, the *Lascar DPM 1* digital panel meter measures 30 by 14 by 13mm and is claimed to be the world's smallest. All the circuitry is on the thick-film hybrid, which is on ceramic and uses very fine tracks, all the resistors being printed onto the substrate and automatically laser trimmed as the circuit is powered-up, replacing trimming potentiometers. Instead of a surface-mounted IC the device is a bare die bonded to the substrate and connected to the rear side through printed-through holes in the ceramic. CorinTech Ltd. Tel., 0425 655655; fax, 0425 652756.

600V high-side driver. *IR2117* from IR is a high-side driver IC with a floating channel designed for bootstrap operation. It copes with offset voltages up to 600V, has a maximum offset supply voltage transient of 50V/ns and has an undervoltage lockout. On and off times are 125ns and 105ns and the device drives mosfets and igbts with a gate drive of 10-20V. It is also compatible with cmos outputs. International Rectifier. Tel., 0883 713215; fax, 0883 714234.

Datacoms dc-to-dc converter. Designed for the dc-input modem market, Gardners's new converter takes a 40-60V dc supply and provides three outputs: 5V,1A and ±12V,0.2A. All outputs stay within 5% of nominal in all conditions of line, load, temperature or cross-regulation and are overcurrent-protected. Ripple and noise are under 50mV and 100mV for 5V and 12V outputs. Gardners Ltd. Tel., 0202 482284; fax, 0202 470805.

Logic building blocks

16-bit ALU. Logic Devices's *L4C381* arithmetic and logic unit performs 16-bit addition, subtraction and logic operations including and, or and exclusive-or in under 15ns. Clock-to-result delay is 11ns, the overflow, zero detect, propagate, generate and carry flags coming in less than 15ns. Input and output registers provide pipe-lined operation and can be by-passed to allow flow-through working. Ambar Cascom Ltd. Tel., 0296 434141; fax, 0296 29670.

Microprocessors and controllers

OTP microcontroller. Toshiba has a one-time-programmable, 4-bit chip which can be programmed on a standard eprom programmer by using the appropriate pin adaptor. Features of the *TMP47P443VN* include an 8-channel, 8-bit a-to-d converter, three timers and a watchdog, an 8-bit serial port, zero-crossing detector, eight 20mA outputs and a pulse output. There are 4Kbyte of program memory and 256nibble of ram. Working at 5V and 6MHz, instruction cycle time is 1.3µs. Toshiba Electronics (UK) Ltd. Tel., 0276 694600; fax, 0276 691583.

More memory for H8/300. Hitachi's *H8/300* family of 8-bit microcontrollers now has two new members provided with more memory: the *H8/3256* and *H8/3247* are upgrades of, and compatible with the H8/325 series, having 2K-byte of ram and 48K-byte(3256) or 60K-byte(3257) of rom. Further, the devices now operate at voltages down to 2.7V at 5MHz. Hitachi Europe Ltd. Tel., 0628 585000; fax, 0628 585200.

Mixed-signal ICs

Video conferencing ICs. AT&T has a second generation of its *AVP* video codec chipset. The chipset is for desktop and group videoconferencing to H.320 standard compression and decompression. Each set in the series includes system controller, video decoder and video encoder, the new chipsets being software-compatible with the original AVP chips. New features include interpolation for better quality, higher data rates up to 1.92Mb/s and 30frames/s video encoding. AT&T Microelectronics. Tel., 0734 324299; fax, 0734 328148.

3V RS232 serial port. *MAX3241* by Maxim is an RS232 interface IC that allows true RS232 performance from a 3-5.5V supply taking only 1mA (1µA when shut down). It is a complete three-driver, five-receiver port using four small capacitors as the only external components and running at data rates of up to 120kb/s while maintaining 5V outputs. Maxim Integrated Products UK Ltd. Tel., 0734 845255; fax, 0734 843863.

CT1 transceiver. Fujitsu's *MB15A31* is a dual serial-input PLL frequency synthesiser for CT1 transceivers which incorporates digital blocks and analogue rf functions for both receive and transmit paths, all in the one chip. Both paths have 19-bit latches, programmable divider, phase detector with polarity change, 1GHz dual modulus prescaler, charge pump and vco, an analogue switch being used in the transmit block. A receive mixer has a gain of 21dB (–3dBm). Power consumption is 52.5mA at 3V and there is a power-down facility for the transmit PLL and vco. Fujitsu Microelectronics Ltd. Tel., 0628 76100; fax, 0628 781484.

Optical devices

Laser inspection. Optics by *Lasiris* generate a range of light patterns such as single or multiple lines, dot arrays and concentric circles from the output of helium-neon lasers or laser diodes, although they can be used with other types of laser. Viewed by a camera, the patterns allow inspection of parts, ranging in size from pcb components to car bodies, for alignment or edge detection. Gaussian distribution along the length of the stripes produced is eliminated. Laser Lines Ltd. Tel., 0295 267755; fax, 0295 269651.

Optical sensors. *Isocom Components* has a range of slotted interrupter and reflective sensors using infrared leds and photosensors in single or dual packages, the dual type being intended for direction sensing and, with an interface, speed measurement. A transistor detector is standard, but a Darlington can be supplied, as can logic-compatible output. Isocom Components Ltd. Tel., 0429 863609; fax, 0429 863581.

Optical sensors. Slotted optical switches in the *QVA* and *QVB* series by Quality Technologies use an infrared led facing an n-p-n phototransistor over a 3.18mm gap. Internal apertures of 1.27mm or 0.25mm confer the high resolution needed for use in encoders and position sensors. Various combinations of aperture size, lead spacing and output current are available and the devices come in board-mounting or panel-mounting form. Quality Technologies UK Ltd. Tel., 0296 394499; fax, 0296 392432.

Oscillators. Providing ecl square-wave output with transition times of less than 4ns, ACT's *HEV1500 AA* series of voltage-controlled crystal oscillators operates in the 10-180MHz frequency range. Stability options range from ±15ppm to ±100ppm and temperature ranges are 0-70°C to –55-105°C. Advanced Crystal Technology. Tel., 0635 528520; fax, 0635 528443.

Programmable logic arrays

Mixed-logic plds. Lattice Semiconductor *GAL16LV8ZD* and *GAL20LV8ZD* re-programmable logic devices interface with both 3.3V and 5V logic and consume only 45mA at 62.5MHz while operating from 3.3V. A power-down facility reduces current to 50µA. Erasure time is under 100ms. Micro Call Ltd. Tel., 0844 261939; fax, 0844 261678.

Power semiconductors

600V bridge driver. *IR2132* from IR is a three-phase, floating-channel bridge driver IC providing 10-20V gate drive for igbts or mosfets at offsets up to +600V. The device has three independent high-side and low-side referenced output channels for three-phase use. Inputs are compatible with 5V cmos or LSTTL and a ground-referred op-amp gives analogue feedback of bridge current by an external sense resistor, a current-trip facility also coming from this resistor. On and off times are 120ns and 94ns and dead time is 0.8µs. International Rectifier. Tel., 0883 713215; fax, 0883 714234.

2W dc-to-dc converter. Packaged in 24-pin dips, Semtech' *MP9000* series of 2W converters have a regulated output, 500V dc isolation and short and thermal protection. Acceptable input in 5V and 12V to give outputs of single 5V-24V and dual 12V and 15V at up to 200mA. Output ripple is 100mV pk-pk, the single unit having a pi filter at the output and the dual type at both input and output. Semtech Ltd. Tel., 0592 773520; fax, 0592 774781.

1.25A switcher. Cherry's *CS-3972* 1.25A, 60V current-mode switching regulator can be configured as a buck, boost, flyback, forward, isolated and non-isolated type, using a single-ended switch. Output switch, band-gap reference, voltage regulator, error amplifier, 40kHz oscillator, control and protection circuits are all integrated. The output transistor is quasi-saturated in the on condition, so that turn-off delay is lessened, as is power dissipation. Minimum input voltage is 3V at 6mA. Clere Electronics Ltd. Tel., 0635 298574; fax, 0635 297717.

PASSIVE

Trimmer capacitors. Murata's *TZCX3* miniature trimmers are made with an alumina stator instead of polymer to obtain better stability; capacitance setting drifts by less than ±1% after passing a solder bath. Values cover the 1.5-20pF range at temperature coefficients of 200ppm-500ppm, depending on value, Q lying between 300 and 500. Murata Electronics (UK) Ltd. Tel., 0252 811666; fax, 0252 811777.

Surface-mounted resistors. Low-value chip resistors by Dale Electronics are made from a solid metal strip instead of the film normally used, conferring high stability, low inductance and good power handling in these *WSL Series* resistors, whose values lie in the 0.01Ω to 1Ω range at 1W or 0.5W. Tolerance is ±1% and temperature coefficient ±75ppm/°C. Vishay Components (UK) Ltd. Tel., 0915 144155; fax, 0915 678262.

Shielded coils. Miniature surface-mounted chip coils in a new range from Murata measure 2 by 1.25mm and 0.9mm to 1.25mm high, depending on value. They are externally shielded by ferrite to eliminate emi. The *LQG21N* range comes in values from 0.1µH to 4.7µH, the 4.7µH type having a Q of 60 at 10MHz. Self-resonance occurs at 340MHz for the 0.1µH coil and 47MHz for the 4.7µH one. The smaller values possess 0.17Ω dc resistance and are rated at 250mA. Surtech Interconnection Ltd. Tel., 0256 51221; fax, 0256 471180.

Linear potentiometer. Model *LT/MLT* miniature linear potentiometer by Control Transducers is only 9.5mm in diameter and is designed for use in hostile conditions. It is made in extruded aluminium with guide rails for element and wiper assembly and has two wipers to assist in reducing the effect of vibration. MystR is the element material, which confers low noise and an expected life of 10^9 operations. Stroke lengths available are 13-255mm in 10 ranges at an independent linearity better than ±0.1%. Resistance is 1kΩ nominal. Control Transducers. Tel., 0234 217704; fax, 0234 217083.

Flatpack relay. A PCMCIA-compatible solid-state relay from CP Clare, the *TS* series is half the height of a standard relay at 2.3mm and will fit in a 'credit-card' modem, which can therefore house the Data Access Arrangement circuit with hookswitch, pulse dial relay and ring detect. CP Clare International nv. Tel., 01032 12/39 04 00; fax, 01032 12/23 57 54.

Chip capacitors. Kyocera has the *0402* surface-mounted chip capacitor in the *MLC* range, now available in NPO, SL, X7R and Y5V dielectrics. In NPO, the range is 0.5-120pF at 50V; in SL 82-220pF at 50V; in X7R 220pF-10nF at 16-50V; and in Y5V 2200pF-47nF at 16-50V. Insulation resistance is greater than 10GΩ or 500MΩ minimum. AVX Ltd. Tel., 0252 336868; fax, 0252 346643.

Connectors and cabling

Filtered BNCs. Two ceramic chip capacitors in *Genalog*'s BNC connectors provide filtering to reduce emi/rfi leakage. Contacts are gold-plated phosphor bronze and the inner insulator is polypropylene. The glass-fibre outer has a UL94V-0 rating. There are vertical and right-angled versions in standard die-cast and lower-cost plastic types available. Genalog Ltd. Tel., 0580 753754; fax, 0580 752979.

Displays

Colour lcd. A single edge light on NEC's new 9.4in colour lcd module is responsible for a power reduction to 4.8W from the 8W of the earlier unit. *NL6448AC30-10* is an active-matrix, thin-film-transistor colour lcd resolving 640 by 480 pixels and displaying 4096 colours at a contrast ratio of 110:1 and with a response time of 40ms. The single light still provides a luminance of, typically, 90cd/m^2. Circuitry for driving the light and the crystals is integrated. Horizontal viewing angle is 45° and in the vertical direction 30°. Weight is 680g. and thickness 12.5mm. NEC Electronics (UK) Ltd. Tel., 0908 691133; fax, 0908 670290.

"Largest" colour lcd. Sharp claims a world size record for its 21in thin-film transistor colour lcd, which is just over an inch thick. Dot configuration is 640 by 480 and pixels are in rgb vertical stripes. Sharp Electronics GmbH. Tel., 01049 40 23760; fax, 01049 40 23762510.

Monochrome LCD. The Orion *OEM-6448* is a monochrome 9.5in liquid-crystal display to the VGA graphics standard, only 7mm thick and weighing 315g. Its contrast ratio is 18:1 and there are up to 64 grey-scale levels. Hardware available includes a range of PC cards to provide fluorescent back lighting and variable contrast. Options are brightness control and contrast temperature compensation. EAO-Highland Electronics Ltd. Tel., 0444 236000; fax, 0444 236641.

Filters

Chip filters. *Murata* announces two-pole and three-pole chip band-pass filters for mobile, portable and cordless telephones. They are surface-mounted and designed to work in the 500MHz-3GHz range to customers' specification. The three-pole type have an insertion loss of 4dB maximum and, as an example, an 836.5MHz centre-frequency filter with a bandwidth of ±12.5MHz will give at least 30dB attenuation at ±77.5MHz. The two-pole version at the same frequency has a 2.8dB insertion loss and 20dB at the same offset. Murata Electronics (UK) Ltd. Tel., 0252 811666; fax, 0252 811777.

Sealed switch. The *Grayhill series 39* family of miniature pushbutton switches are now available in a sealed version, which has an O-ring seal to protect the switch during flux cleaning, since the switch is meant for board mounting. Configuration is spdt break-before-make, but the switch is usable as dpst. Contacts are rated at 0.125A at 220V ac, contact resistance 25mΩ and insulation resistance 1GΩ. EAO-Highland Electronics Ltd. Tel., 0444 236000; fax, 0444 236641.

Hardware

Custom boxes. If one's interest as a design engineer stops dead when the circuit performs as intended, reality can be restored by *Bafbox*, who can supply a range of enclosures which can be tailored for size and drilled with the required holes in a week, for short runs. Larger numbers will be handled later, including modifications from the smaller pre-production batch. Bafbox Ltd. Tel., 0280 705777; fax, 0280 706320.

Instrumentation

Sound-level meter. B&K's *Type 2260* sound-level meter is intended for use in environmental monitoring and product testing. It is a hand-held instrument, but also incorporates the features of a PC. This instrument is a digital filter analyser giving 1/1octave analysis with centre frequencies from 31.5Hz to 8kHz, software providing level distribution and cumulative distribution on broadband channels and in individual octave bands, together with a 15s graphical level v. time profile. It is expandable by means of standard interfaces and programmable by PCMCIA program cards. Two phase-matched channels have 80dB dynamic range, an unweighted output being provided for taping, and results are down-loaded via the RS232 interface. Bruel & Kjaer (UK) Ltd. Tel., 081 954 2366; fax, 081 954 9504.

Cable-tv signal-level meter. A 48-861MHz signal-level meter by Alban Electronics, the *Promax MC-560* is designed to cope with present and future cable systems, automatically measuring ratios of video-to-audio, and carrier-to-noise and carrier-to-Nicam, results being presented on a backlit lcd. Its microprocessor holds the CCIR frequency plan, but the user is allowed to make his own with the help of a PC, or Alban will do it. Alban Electronics Ltd. Tel., 0727 832266; fax, 0727 810546.

Distortion meter. Leader's new automatic distortion meter, the *LDM178*, uses a high-pass filter with three spot frequencies at 315Hz, 1kHz and a user-set frequency to eliminate wow and flutter, an automatic level control assisting distortion measurement in tape equipment. Two panel meters show output level alongside distortion in dB. Thurlby Thandar Instruments Ltd. Tel., 0480 412451; fax, 0480 450409.

Radio clock. *ADC-60* provides standard time in ascii or bcd to any computer serial port to the accuracy of the 60kHz MSF transmissions. Its internal clock locks to MSF or to the German DCF at 77.5kHz when MSF is down for maintenance, the internal source free-running if no transmission is received. Two software packages are supplied: one a TSR to run under Dos and the other to run minimised under Windows. Amdat. Tel., 0272 699352; fax, 0272 236088.

Communicative counter. A 1.3GHz counter-timer by Thurlby-Thandar, the *TF830-ARC* is provided with an RS232 serial interface for control by and reading to an external controller. Measurements may be made on frequency, period, frequency ratio, pulse width and event counting. A technique of reciprocal counting is used which results in a high resolution at all frequencies and eliminates the normal ±1 count error. Resolution is at least seven digits and frequencies down to 0.001mHz can be measured. Thurlby Thandar Instruments Ltd. Tel., 0480 412451; fax, 0480 450409.

Crystals

Low-profile crystals. ACT's *4EX* series of leaded crystals measure under 4mm in height and have an HC-49/U footprint. They are in a metal can and cover the frequency range 3.2-50MHz in preferred values or in specified frequencies to order. Stability is ±100ppm and ageing ±5ppm/year. Temperature range is –40°C to 125°C. Advanced Crystal Technology. Tel., 0635 528520; fax, 0635 528443.

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

Smart DPMs. 'Intelligent' digital panel meters in the *DPM* range from ITT can be configured for various measurement and data acquisition requirements, basic instruments each having 34 ranges of voltage, alternating and direct current and temperature measurement. Extra functions include data-logging memory, thermocouple linearisation, alarm monitoring and interfaces for printers, message displays, recorders, Pcs and PLCs. Resolution is 40,000 with a scaleable display from –19999 to +999999. ITT Instruments. Tel., 0256 311877; fax, 0256 23659.

Navigation systems

Compact GPS receiver. Rockwell's new *MicroTracker GPS Receiver* is a five-channel unit with differential GPS for accuracy in positioning down to less than 5m. The differential working eliminates the effect of selective availability which, being translated, means a reduced accuracy of up to 100m, errors having been introduced by the US during the Gulf War. MicroTracker also has 30s time to first fix and works accurately at altitudes from –1000ft to 45000ft. Telecom Design Communications Ltd. Tel., 0256 332800; fax 0256 332810.

Power supplies

Small converters. Maxim claims its *MAX860/861* to be the world's smallest charge-pump dc-to-dc converters. They use no inductors and the only externals are two capacitors that either invert or double the input. Switching frequency is selectable in six ranges between 6kHz and 250kHz. Inputs of 1.5-5.5V are inverted or doubled to 2.5-5.5V into an output impedance of 12Ω. Logic-controlled shutdown is provided. Package is 8-pin SO. Maxim Integrated Products UK Ltd. Tel., 0734 845255; fax, 0734 843863.

High-voltage supply. Applied Kilovolts has a floating, variable high-voltage power supply, the *HP3F*, intended to power imaging electron detectors, supplying 0-3.5kV at up to 100μA. Ripple and noise are less than 20mVpk-pk and injected ripple on the accelerating 2.5kV supply less than 10mV pk-pk. Power required is 24V from earth, with controls and monitor signals also earthed. Applied Kilovolts Ltd. Tel., 0273 439440; fax, 0273 439449.

Radio communications products

Quadrature modulator. *RF2412* by Anglia is a low-cost universal modulation chip to generate AM, FM, PM or compound carriers at vhf/uhf. Modulating IF at 50-150MHz, it offers an rf output at 200-1000MHz. The chip includes all components including diff. amplifiers for baseband input, a 90° hybrid phase splitter, limiting local oscillator amplifiers, two balanced mixers, a combining diff. amplifier and an rf amplifier driving into 50Ω. A single 5V supply is needed. Anglia Microwaves Ltd. Tel., 0277 630000; fax, 0277 631111.

Transducers and sensors

Temperature sensors. *Elmwood's* range of temperature sensors are flexible, wire-wound or etched-foil types either in flexible dielectric layers or on film dielectrics, designed to conform to uneven surfaces and to measure the temperature of a surface rather than a point; maximum thickness is 0.04in. They are provided with pressure-sensitive adhesive backing and cover the range –200°C to 235°C. Radiatron Components Ltd. Tel., 01784 439393; fax, 01784 477333.

Peak rf power sensor. Rohde & Schwarz introduce the *NRV-Z31* peak power sensor , which will find application in the measurement of transmit power of cellular radio equipment, sync. pulse power for television, emc test signal peak power and peak power of line-frequency modulated signals such as microwave ovens and diathermy equipment. The unit is effectively a probe for the R&S range of power meters and covers the frequency range of 30MHz-6GHz. Three models exist: model 02 handles rf bursts down to 2μs wide; 03 does the same for prfs from 100Hz, giving up to seven readouts per second; and 04 is meant for use with GSM, PCN and DECT radio. Rohde & Schwarz UK Ltd. Tel., 0252 811377; fax, 0252 811447.

Accelerometer. TI has a new acceleration sensor for the ±0.4 to ±10g range at up to 50Hz, designed for use in measuring vehicle acceleration in anti-lock braking, traction control and suspension systems. It uses a metal beam in a capacitive circuit and IC circuitry, a technique claimed by Texas to be more durable than micromachined silicon. Texas Instruments. Tel., 0234 270111; fax, 0234 223459.

Vision systems

Colour ccd camera. Sony's *XC-777P* is the company's smallest and lightest colour camera, being based on the 440kpixel 1/3in imager, all associated circuits being contained in the one housing, which measures 22 by 22 by 89mm, weighing 75g. Resolution is 460TVL and minimum sensitivity 4.5lux. There is an electronic shutter speed as well as a flickerless mode to overcome beating under fluorescent lights. Power consumption is 2.3W. Sony Computer Peripherals & Components. Tel., 0932 816000; fax, 0932 817001.

PCMCIA video camera. *PC Card Camera* by VVL is claimed to be the first integrated PCMCIA video camera. It is palm-sized and may be integrated into notebook and pen-based PCs, being controlled through the Windows VVL Snap package interfacing to the PCMCIA port through Card and Socket Services. It will display motion video in real time on the PC screen at five frames per second, the resulting images being captured and saved as .TIF files, if required. VLSI Vision Ltd. Tel., 031-539 7111; fax, 031-539 7140.

LabView/HiQ bundle. *LabView* graphical instrumentation software and *HiQ* numerical analysis and data visualisation software, both for the Mac, are now combined by National into a package called *LabSuite* at a considerable saving in cost. *LabSuite* now provides laboratory automation, automatic testing, process monitoring and control, physiological monitoring, personal instrumentation, numerical analysis, data visualisation and reporting. National Instruments UK. Tel., 0635 523545; fax, 0635 523154.

External data acquisition. As a more convenient alternative to plug-in, internal boards, Strawberry Tree's DATAShuttle is a small external unit taking inputs directly from sensors, its output going to a PC parallel port. Eight analogue inputs and eight digital i/o lines interface to the PC, there being no need to open up the PC or adjust calibration; sensors connect directly to the DATAShuttle. Quicklog PC is supplied with the system and the unit works with WorkBench for Windows. There is a high-speed parallel pass-through interface to provide a port for other devices, avoiding the need to disconnect them to use the port for data acquisition, and up to 15 DATAShuttles can be linked in series to give 120 analogue inputs and 120 digital i/os. Inputs are automatically calibrated for thermocouples and rtd compensation and gain, resolution and acquisition rates are software selectable on all channels. Adept Scientific Micro Systems Ltd. Tel., 0462 480055; fax, 0462 480213.

Compact cameras. Henderson's new range of board cameras includes pin-hole versions and types designed to take a range of interchangeable lenses from 3.6mm to 16mm. The cameras are on a single printed board measuring 42mm square and are sensitive down to 0.5lux. There are several enclosures, including an ABS case, a spherical globe and smoke-detector types. A 12V power supply can be fitted up to 15m from the camera. Henderson Security Electronics Ltd. Tel., 0684 274874; fax, 0684 294845.

COMPUTER

Computer board-level products

100MHz backplane cpu card. The *Blue Inferno* single-board computer by HM Systems uses IBM's 66MHz and 100MHz processors. The board takes up to 62M-byte of ram, has a fast IDE local bus and video, full i/o facilities, 1Mb of dram and a video cache. There are two serial ports, a 16-byte fifo buffer and an enhanced parallel port. H M Systems plc. Tel., 081-209 0911; fax., 081-209 0912.

Two-Pentium motherboard. SPC has the *ASUSTeK PC1/E-P5NP4*, a PC motherboard using two Dual Pentium P54C processors on one board, which is intended for multiprocessing on desktop workstations and low-level servers. It is compatible with PCIbus and EISAbus and its zif sockets take 90MHz or 100MHz P54Cs. The motherboard uses Intel's Neptune 82430N PCIset chipset, up to

512Kbyte of 3.3V write-back memory and a 1M-bit flash eprom containing P54C PCI bios. SPD Ltd. Tel., 0420 563588; fax, 0420 562206.

Computer systems

People-proof computer. The Dynapro *ErgoTouch* computer appears to be proof against most of the disasters a hostile world has to offer. It is completely sealed and operated by a touch screen, the whole being impervious to the public at large and most chemicals. Non-technical, novice or casual users can use it to get information or control things. It provides vga graphics and all the features including "plenty" of ram and a large hard disk needed to run as a stand-alone computer or in a network. Standard PC tools can be used for development of applications. Enclosures are custom-designed in any colour and the touch surround can be provided with any arrangement of buttons or function keys. An optical port allows software input or output. The computer can be mounted almost anywhere, since it is light in weight. Advanced Modular Computers Ltd. Tel., 0753 580660; fax, 0753 580653.

Development and evaluation

8-bit development kit. Toshiba has introduced a development kit to provide a low-cost method of developing applications for 8-bit microcontrollers. *TLCS-90* contains the *TDB90* development board based on the TMP90C141N controller, terminal software, a monitor and a full-feature assembler, the kit being connected to a PC by a printer cable. The *TDB90* board has 1Kb ram, 6-channel, 8-bit resolution a-to-d conversion, up to 54 i/o pins, stepper-motor control, timers and counters, a serial i/o channel and a prototyping area. Purchasers can buy 10 *TMP91P640N-10* one-time programmable devices at a special rate. Toshiba Electronics (UK) Ltd. Tel., 0276 694600; fax, 0276 691583.

Computer peripherals

GPIB instrument control. For *H-P 9000* series *700* workstations with EISA slots, National announces the *GPIB-HP700-EISA* interface kit which features 5.5Mbyte/s data transfer rate for both read and write using HS488 and 1.3Mbyte/s with three-wire GPIB. The kit includes the EISA-GPIB board and NI-488.2M software for HP-UX v.9, which has over 50 GPIB-related routines and functions. National Instruments UK. Tel., 0635 523545; fax, 0635 523154.

Port expander. Technology Concepts uses IDT's *R3051* risc processor in its *SUPERport* snap-together modular system which supports up to 256 RS232 or RS422/485 serial ports on a PC, for use in multi-user systems running UnixWare, SCO Unix or Multiuser DOS. The ISA bus controller card with the R3051 drives the whole stack of ports simultaneously at up to 115.2Kb/s. IDT's 3051 family includes four devices with on-chip cache sizes of 2.5, 6, 10 and 20K-byte and frequencies of 16-40MHz. Integrated Device Technology. Tel., 0372 363734; fax, 0372 378851. Technology Concepts Ltd. Tel., 0633 872611; fax, 0633 879329.

Programming hardware

Universal programmer. Ice Technology's *Speedmaster LV* low-cost universal programmer operates at both 3.3V and 5V. Without the use of adaptors, the instrument will program devices with up to 40 pins, adaptors being available for unusual packages. Optional 16-bit and 8-bit 3.3/5V emulator cards plug into the unit to provide a rom/ram emulator to test code in the system before programming. Ice Technology Ltd. Tel., 0226 767404; fax, 0226 370434.

Software

Analogue filter design. Information on ripple and attenuation levels, pass-band and stop-band limits and termination impedances given to *Filtech*, a new filter synthesis package from Number One Systems, are the only inputs an engineer must provide to obtain a completed design. Filtech analyses synthesised filter circuits independently and displays a plot of the calculated frequency response superimposed on specified limits. It will handle both active and passive filters of up to sixth-order at frequencies of less than 1Hz to more than 1GHz. It is interactive, in that its recommendations can be overridden, so that filter type or order can be altered and the effects shown on screen. A further feature is that Filtech will, if required, force all or some components to fit one of the preferred-value sequences and show the effect on response. Number One Systems Ltd. Tel., 0480 461778; fax, 0480 494042.

Visualising numerical data. *PV-WAVE Advantage* by Visual Numerics combines graphics and numerical analysis to provide easier assimilation of large amounts of data, enabling patterns and trends to be observed and analysed. The package runs under Unix and a new version under Windows, while the *PV-WAVE Point & Click* is a Unix-only version. Workstation Source Ltd. Tel., 0734 759292; fax, 0734 757522.

LETTERS

Letters to *"Electronics World + Wireless World"* Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

Proof of the pudding...

Peter Barnes is quite right to apply all his critical faculties to features in the media which put forward new scientific ideas, and sometimes challenge orthodoxy.

As the purveyor of a "rubbishy theory" suggesting how power system electrodynamic fields and natural ionising radiation might cause disease, I would remind him that yesterday's heresy often becomes today's dogma, and that proof takes time.

An excellent example of how scientific truth will out, was the way Marconi had to demonstrate ('The man who started ripples in the ether', *EW+WW*, September) that radio waves could reach over the horizon before his armchair critics – who included many eminent physicists of the day – were silenced.

In my own case, I hope Mr Barnes noticed the report in *Research Notes* (*EW+WW*, September) which also linked production of gamma radiation with intense electric fields in thunderstorms – fields which also exist near high voltage lines!

All I can say at this stage is that others have confirmed that there is an ionising radiation anomaly near power lines, and that reports of the death of the idea at the hands of a well known body may be exaggerated. Serious shortcomings have recently become apparent in the radiation measuring techniques used to test the idea.

Anthony Hopwood
Worcester

...is in the heating

In his comments (*Letters*, September) R L Tufft uses the words "sham" and "hoax" in reference to the article "Electricity without Magnetism" (*EW+WW*, July, 1992, pp. 540-542). He said the hypothesis seemed to be of "monumental importance" but asks "where is the power?".

As author and co-inventor, I had been intent on saying no more on the subject until I had built my own prototype version advancing on the three prototypes built by J S Strachan. But I am now forced to comment.

First, may I say that the title of the *EW+WW* article was an editorial change that reflected more the opinion of my colleague Strachan who suspected that the special properties of the polymer dielectric used in the capacitor stack were responsible for the heat-to-electricity conversion. My belief is that the nickel film forming the capacitor plate electrodes is the seat of an action enhanced by the magnetic polarisation in the nickel. Magnetism seems to be essential as a magnetic field deflects electron heat flow to confront an opposing electric potential. I see the polymer with its electric polarisation as serving only as if it gives the device the character of an electrolytic capacitor, so enhancing the current in the oscillations in the transverse-to-heat-flow direction.

So, two years on, where do we stand? Well, the *EW+WW* article attracted research interest and within a short period Strachan had a new working prototype, the performance of which I captured on a video recorder which has not failed to arouse interest by those in academia and corporate research who have seen it. As a result the primary research effort at this time is being undertaken by a group at MIT in the USA.

Strachan has been unable to pursue the project owing to priority research on a DTI-funded Smart I and Smart II award concerning his laser technology developments. Recently I, in my retirement, have begun my own experiments. But I too have now won a Smart I award to research another project only marginally related to the thermoelectric converter. So much of the forward progress now depends upon the initiative of others.

As to the "hoax" aspect, was mhd a hoax back in the 1960s? MHD was the technology of seeding ions into the flow of hot gas in passage through a magnetic field to develop an electrical output in the mutually orthogonal direction. Our thermoelectric invention is simply a solid-state version of that technology. Heat carried by electrons in passage through thin nickel film is deflected laterally by the domain magnetism in the nickel and by activating transverse current flow which always takes the path of forward emf (least resistance – in fact negative resistance) so the electrons cool as heat converts into electrical output.

Note that in the August 1994 issue of *Physics World* an article entitled 'Metals blow hot and cold' reported that electrons activated by laser-generated sound pulses in metals (nickel being mentioned specifically) exhibit unexpectedly very high temperatures indicating a non-equilibrium state in the electron-phonon interaction. Our thermoelectric invention was a spin-off discovery from research by Strachan aimed at setting up

Linear thinking

The articles on audio amplifier design by Doug Self have been fascinating, and easy to read. Particularly interesting for me was the article concerning output stage linearity ('Common-emitter power amplifiers: A different perception?', July 1994, pp. 548-552).

If an amplifier output stage consists of complementary mosfets in a common source arrangement, and the gate is driven by a current source, then do we not have perfect dc linearity in the output? The high output impedance from connecting the load to the mosfet drains will cause no problems with output impedance, as discussed in the article.

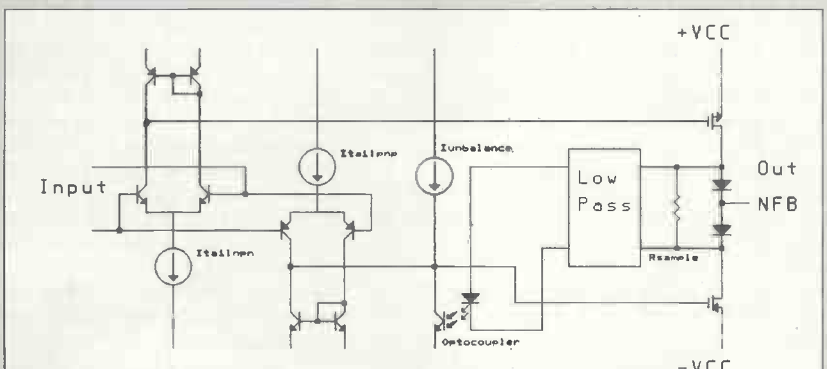
Also, the output drive capability of the current source directly limits the slew rate of the mosfet stage. According to some rather simplified *Spice* simulations, the speed of the stage can be higher than that of a voltage-driven mosfet because the current drive is less likely to cause oscillations.

One method of achieving current drive is to use a differential pair with a current mirror, the tail current of the differential pair being the maximum possible gate current, also determining the slew rate. A complementary differential pair, also with a current mirror, can be used to drive the other mosfet.

Biasing the stage seems to be tricky. A solution is to sample the output current with a resistor, and unbalance the gate currents with a current bias. The output current signal is clipped and severely low-pass filtered, so that audio signals do not affect quiescent current. This signal is then applied to an optocoupler that restores dc gate balance and proper quiescent current, while maximising gate drive impedance.

I am no expert in audio amplifier design. My main interests are in the medical side of electronics and amplifier design is one of my several hobbies. I welcome any comments on this method of linearising power output stages.

Doug Eleveld
Groningen
The Netherlands



acoustic oscillations in a pvdf stack of nickel-aluminium coated laminations.
Harold Aspden
Southampton

Remember the ZX!

The Government asks what we should do to interest young people in technology. Sir Clive Sinclair knew, didn't he! Remember the *ZX Spectrum*?

Playing games was never enough in those days. Any *ZX Spectrum* owner could create their own protection from the death rays of Zargon if they were prepared to learn *Z80* machine language – and many were.

Z80 machine-code programming manuals were commonly seen alongside pop-music counters in those days. Articles written by teenagers for teenagers appeared in magazines like *Your Spectrum*, displaying a knowledge of machine code programming that certainly exceeded my own. The competition between school boffins to crack the latest game equalled anything to be found on the rugby field. A couple of Christmas and birthday presents could provide a full development system small enough to carry in a cycle saddlebag but powerful enough to act as host to develop *Z80*-based microsystems. I doubt if many adults could have afforded a BBC-B system to do this?

Granted, a *286* can now be bought for about one third of a week's wages here in the UK. But its not the same. 'The Complete Commented IBM rom disassembly' is not a title you are likely to see. Such information *is* available, but you have to search it out and know what you are looking for. It certainly won't be easily accessible for teenage programmers, who no longer display their hacking skills in PC magazines. I can imagine the looks of horror if they did! These are business machines: Spectrums were for fun.

So what have we got now? Sega and Nintendo and a coming generation whose idea of computer literacy is to know the latest buzz-words and run custom software under Windows – about as close as you can get to playing 'Nintendo' without actually doing it.

I say: 'Come back uncle Clive. All is forgiven – even that dreadful car'.
Stephen Dyke
Bedfordshire

Lodge-ing a protest

Marconi's first transatlantic experiment of December 1901, backed up by the recordings on-board SS Philadelphia, clearly rates as the most significant event in the history of radio. Furthermore, no-one would wish to deny Marconi's place of pre-eminence in bringing radio into a practical and commercial reality.

But there has always been a dichotomy between the public perception of these events and the more informed judgement of scientists and engineers. So it is a pity that John Powell Riley, in his article on Marconi (*EW+WW*,'The man who started ripples in the ether', pp. 778-782) seems to be perpetuating the popular misconceptions and is gratuitously critical of British scientists and engineers. In particular, notwithstanding the experiments of Hertz, it was the 'Maxwellians' – FitzGerald, Heaviside and Lodge – who understood electromagnetic radiation better than anyone else in the world at that time. Marconi was initially secretive as his early experiments contained nothing novel.

Riley's otherwise interesting article is also quite misleading in that Oliver Lodge was granted the first patent for tuning (11,575 May 10, 1897). The later 7777 Marconi patent should not have been granted, and was also subsequent to that of K F Braun (1862, January 26, 1899). For this reason, it was not recognised elsewhere in the world.

In 1911 that fact was also accepted in this country when Lodge's patent was extended for seven years so that the Marconi Company was effectively forced to purchase it for a large but undisclosed sum of money. Much later, in 1943, the US Supreme Court decided that Lodge's 1897 patent was the only valid tuning patent.

The year 1894 was, of course, not only the start of Marconi's experiments but also the year in which Oliver Lodge gave the first public demonstration of signalling by radio at the British Association on 14 August. This demonstration involved both the single-point and iron-borings coherers *he* invented on the basis of Branly's scientific experiments. Readers looking for more information should see our recently published book (*Oliver Lodge and the Invention of Radio*, eds P Rowlands and J P Wilson, PD Publications, Liverpool, 1994).
J P Wilson
Keele University
P Rowlands
Liverpool University

Intellectual property is right

I have been surprised by the strong feelings aroused by my article 'Patently unclear' (*EW+WW*, May 1994, pp. 433-436). I imagined that reactions would range from indifference, to enlightened avidity, but to be taken to task by callers and respondents for daring to imply that there is such a thing as "intellectual property", and that it is right and proper for inventive creativity to be protected, is a sign that some people don't live in the real world.

Anyone who has sweated out an innovative technical solution, or by dint of intellectual prowess and ingenuity created a new approach, has every right to protect the effort involved. Inventive steps may be inspirational. But often as not they emanate from sheer hard work.

Why should this effort be rewarded by immediate plagiarism, which is what happens in practice when no mechanism exists to protect or exploit the idea. Creative technology is seldom a passive, neutral process. Anyone who doubts this is clearly naive.

Try pacing around your living room at 3.00am with the umpteenth cup of coffee, your head reeling with a jumble of seemingly un-correlatable data, knowing that if you don't crack the creative technical impasse occupying your mind every waking and sleeping hour, three years of hard work is down the drain. The elation in creating a pathway forward really drives home what intellectual property is all about.
Barrie Blake-Coleman
Salisbury

In and outs of informality

I am sorry my article 'The ins and outs of oscillator action' (*EW+WW*, July,1994, pp. 586-589) proved uncongenial to Mr Dawe (*Letters*, August).

Dealing with the points he raised, let me start with the one of least consequence, that the language in which it was couched did not "strictly belong to the science of electronics". It was not meant to. I would be the first to agree that the expressions to which he takes exception would be out of place in the *Proceedings* of one of the learned Institutions or even in the less formal ambience of *Electronics Letters*. But a magazine with such a wide circulation and range of readers as *EW+WW* aims to interest, inform, stimulate and even – dare I say it – entertain its readers? Had the Editor thought the language used too inappropriate he was quite capable of amending it.

Perhaps Mr Dawe comes from a generation which does not remember the symmetrical transistor. This was developed as a cross point switch and worked equally well (perhaps 'equally badly' would be a better expression) in the normal and inverted modes, since the emitter and collector are identical structures.

The modern rf npn epitaxial planar transistor is designed for optimum operation in the normal mode and performance with the collector and emitter interchanged is "much worse". How much worse I cannot say as it is not specified for such use and I have not measured it. I am surprised, though, that Mr Dawe wonders what $1/f$ noise is.

Although "what the article explains is not at all clear" to Mr Dawe, the basic message was that when the anode voltage of a valve swings below the cathode voltage, anode current does *not* flow, whereas when the collector voltage of a transistor swings below that of the emitter, collector current *does* flow. Thus the amplitude stabilisation action of the two circuits is fundamentally different and the removal of forward conduction of the base-collector junction removes one possible source of close-in noise.

Another transistor circuit, which avoids bottoming of the collector (as typically found in a single-transistor Hartley circuit) by a different arrangement, is the current switching or class D oscillator, a single transistor version of which has been described[1].

The pseudo-valve circuit could certainly be considered for use at higher frequencies, and this is in no way contradicted by the choice of 2.5MHz for the operating frequency of my Fig. 4. Such a low operating frequency was deliberately chosen because of the attendant arrangements for monitoring the circuit action. For example recording the emitter current waveform required intrusions into the circuit which would not normally be required.

Figure 5 shows spikes of emitter current just a few nanoseconds wide, stretching the 250MHz capability of my oscilloscope to the limits. Yes, certainly there are "practical problems in making measurements" which must be addressed seriously. Regarding oscillator close-in noise: from being 'flat' (varying as $f([0])$ with frequency offset f from carrier) in the 'far out' region, the spectral noise density rises as $f([-n])$ as the carrier frequency is approached, n increasing from 1 (flicker phase) through to 4 (random walk fm) as the offset decreases – though not all of these regions will appear separately identifiable in a typical oscillator[2, 3]
Ian Hickman
Hampshire

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1. 'Oscillator tails off lamely?', Ian Hickman, *EW+WW*, Feb, 1992.
2. 'Design principles and test methods for low phase noise rf and microwave sources' Dieter Scherer, Hewlett-Packard RF and Microwave Symposium.
3. 'Phase noise in signal sources' . P Robins, IEE Telecommunications Series, No 9, Peter Peregrinus, 1984.

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Measuring true power via rms conversion

In power measurement, problems arise – and much confusion persists when the load is not a pure resistance. Inductive and/or capacitive circuit components introduce a phase shift between the voltage across, and the current through, a given load.

This phase shift becomes more pronounced as the reactance of these components rises, with increasing frequency. They then become a greater portion of the total impedance of the load. In a purely inductive circuit, the voltage leads the current by 90°. Conversely, the voltage lags current in a purely capacitive circuit.

There are three primary ways of defining and measuring the sine-wave power dissipated in a given load impedance – apparent power, P_a, average power, P, and reactive power, P_r.

Apparent power, measured in volt-amperes, is simply the product of the rms value of the voltage across a given load times the rms value of the current through the load. That is:

$$P_a = V_{rms} \times I,$$

where V is in volts and I is in amperes. The volt-ampere rating is often used in specifying electrical equipment since volt-amperes may be used to directly compute the current requirements of individual pieces of equipment.

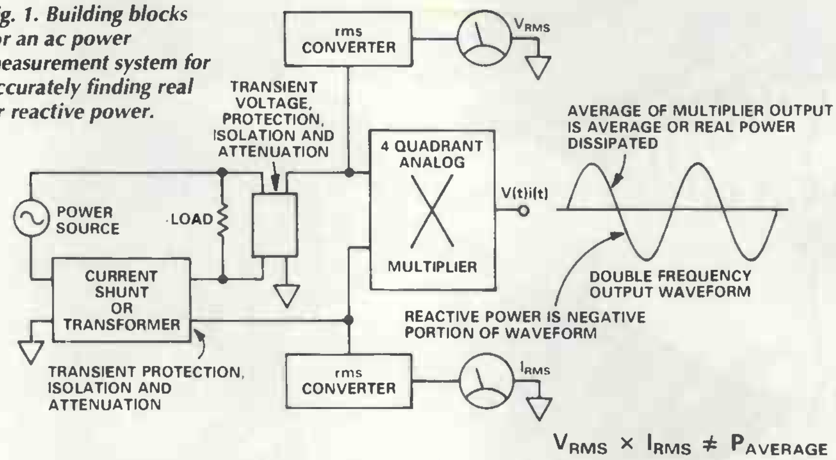


Fig. 1. Building blocks for an ac power measurement system for accurately finding real or reactive power.

Average or real power, measured in watts, is equivalent to the apparent power multiplied by the cosine of the phase angle separating the voltage and current waveforms. That is:

$$P = P_a \cos\theta = V_{rms} I_{rms} \cos\theta,$$

where V is in volts, I is in amperes and θ is in degrees.

Most commonly used, average power specifies the overall power consumption of a particular circuit. This is regardless of the dissipation of its individual components – some of which may be reactive. The cosine of the phase angle, θ, is also referred to as the power factor and is the ratio of a circuit's average power to its apparent power. A highly reactive load exhibits a low power factor with a correspondingly low power consumption.

Because of the importance of defining power consumption within individual reactive components in a circuit, a third power specification, reactive power was created. Reactive power, in VAR. (volt-amp reactive), is used to directly measure the peak power consumption of individual inductive components in a circuit, even though their average power consumption (ideally) is zero.

Reactive power is very important to electrical power companies since they must still supply this energy during a portion of every cycle, even though (on the average) no energy is actually dissipated.

Reactive power is given by:

$$P_r = P_a \sin\theta = V_{rms} I_{rms} \sin\theta,$$

where V is in volts, I is in amperes and θ is in degrees.

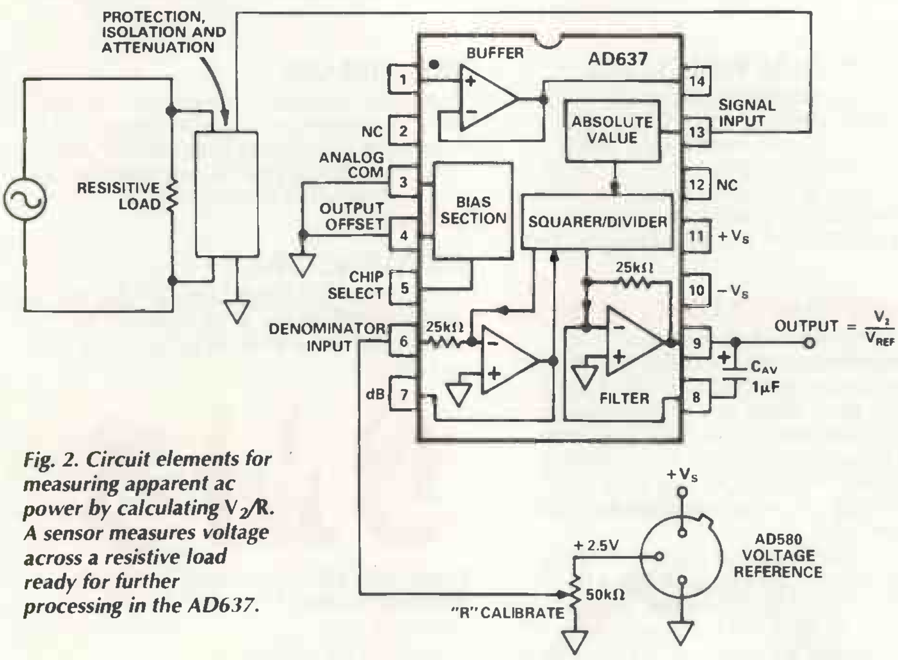


Fig. 2. Circuit elements for measuring apparent ac power by calculating V_2/R. A sensor measures voltage across a resistive load ready for further processing in the AD637.

Practical power measurement

The fact that averaging is carried out in performing rms computation means that whatever phase information existed in the original signal will be lost after rms computation. This fact precludes the use of rms converters for measuring power into non-resistive loads. Measurement of complex power is normally carried out using analogue multipliers, since they will preserve the voltage/current phase information.

Figure 1 shows the building blocks for a practical power measurement system which can accurately measure both real and reactive power. As shown, rms converters are used for real-time monitoring of the rms value of the voltage and current waveforms being processed. With their dc outputs, the converters can directly drive either analogue panel meters or dvm chips.

Output of the analogue multiplier is *VIcosq*. At this point, unfiltered multiplier output equals instantaneous power dissipation through the load. As shown, if the output is low-pass filtered, it will then equal the average or real power dissipated.

Likewise, if only the negative half cycle of the output waveform is detected and filtered, this output will respond to the reactive power dissipated in the load.

Figure 2 shows a practical circuit for measuring apparent power by calculation of V_2/R. A voltage sensor measures the voltage across a resistive load. The *AD637* rms converter then squares this voltage ready for scaling by the denominator input voltage at pin 6. Denominator voltage must be set to give the required output voltage scaling for each particular load resistance.

Since *VR* varies with the value of *R*, the circuit must be recalibrated each time the value of load resistance is changed. One volt per milliwatt or one volt per watt would be practical scale factors for this circuit. Because a squaring operation is being performed by the *AD637*, the scaling voltage must be carefully chosen to provide sufficient headroom to allow the rms converter to process the maximum full-scale input level without clipping. Thus, there will be a tradeoff between maximum input level and low-level sensitivity.

This information is extracted from AD's book *RMS-to-DC Conversion Application Guide*.

Analog Devices, *Station Avenue, Walton-on Thames, Surrey KT12 1PF. Tel. 0932 247401, fax 232222.*

PWM output from semiconductor pressure sensors

For remote sensing and noisy environment applications, frequency modulated or pulse width modulated output is more desirable than an analogue voltage. Both fm and pwm outputs inherently have better noise immunity in these types of applications.

Generally, fm outputs are more widely accepted than pwm outputs, because pwm outputs are restricted to a fixed frequency. According to Motorola application note *AN1518* however, obtaining a stable fm output is difficult to achieve without expensive, complex circuitry.

With either an fm or pwm output, a microcontroller can be used to detect edge transitions to translate the time-domain signal into a digital representation of the analogue voltage signal. In conventional voltage-to-frequency conversions, a voltage-controlled oscillator may be used in conjunction with a microcontroller. This use of two time bases, one analogue and one digital, can create additional inaccuracies.

For either fm or pwm outputs, the microcontroller is only concerned with detecting edge transitions. If a programmable frequency-stable pwm output could be obtained with simple, inexpensive circuitry, a pwm output would be a cost-effective solution for noisy environment/remote sensing applications while incorporating the advantages of frequency output.

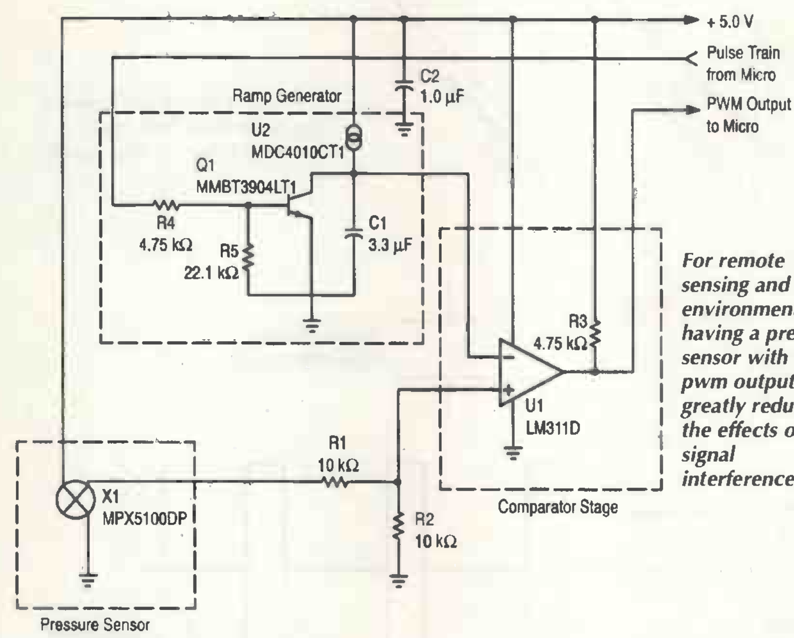
In the pulse width modulated output pressure sensor design shown, simple, inexpensive circuitry creates an output waveform with a duty cycle that is linear to the applied pressure. Combining this circuitry with a single digital time base to create and measure the pwm signal, results in a stable, accurate output.

Two additional advantages of this design are that an a-to-d converter is not required, and since the pwm output calibration is controlled entirely by software, circuit-to-circuit variations due to component tolerances can be nulled.

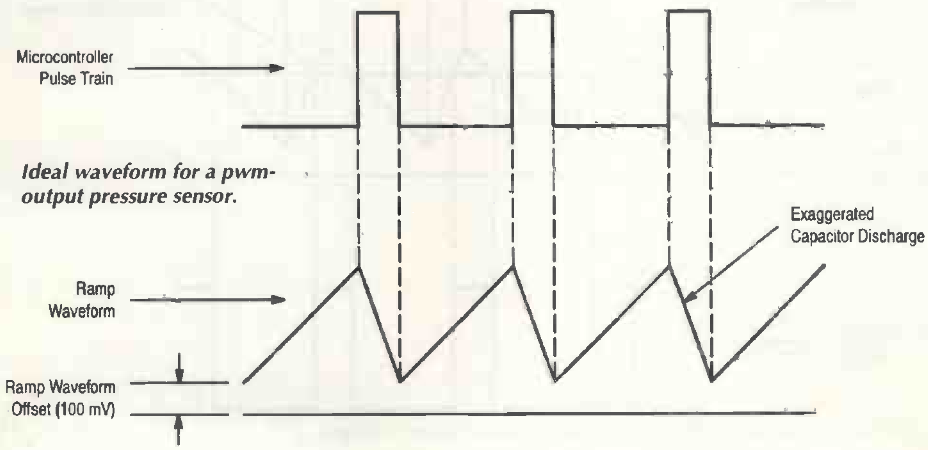
Circuit elements

Main elements of the pwm sensor system are the pressure sensor, a ramp generator, a comparator, and a microcontroller. Ramping is produced by a transistor switch, constant current source, and capacitor.

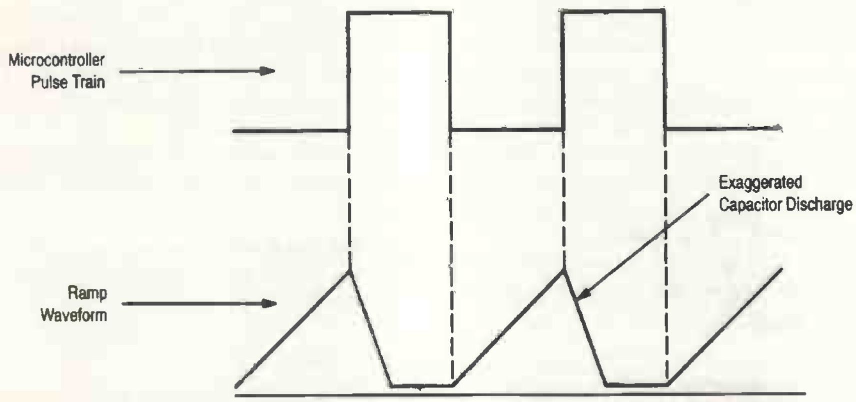
Pressure transducers in the *MPX5000* series are temperature compensated and calibrated. They are available in full-scale pressure ranges of 50kPa (7.3lb/in^2) and 100kPa (14.7lb/in^2). With the recommended 5.0V supply, the transducer produces an



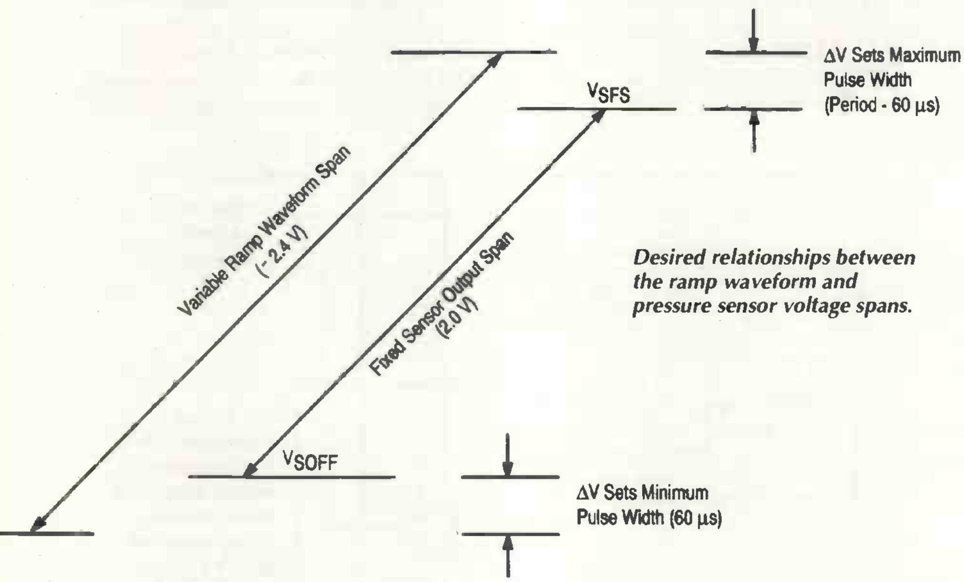
For remote sensing and noisy environments, having a pressure sensor with a pwm output can greatly reduce the effects of signal interference.



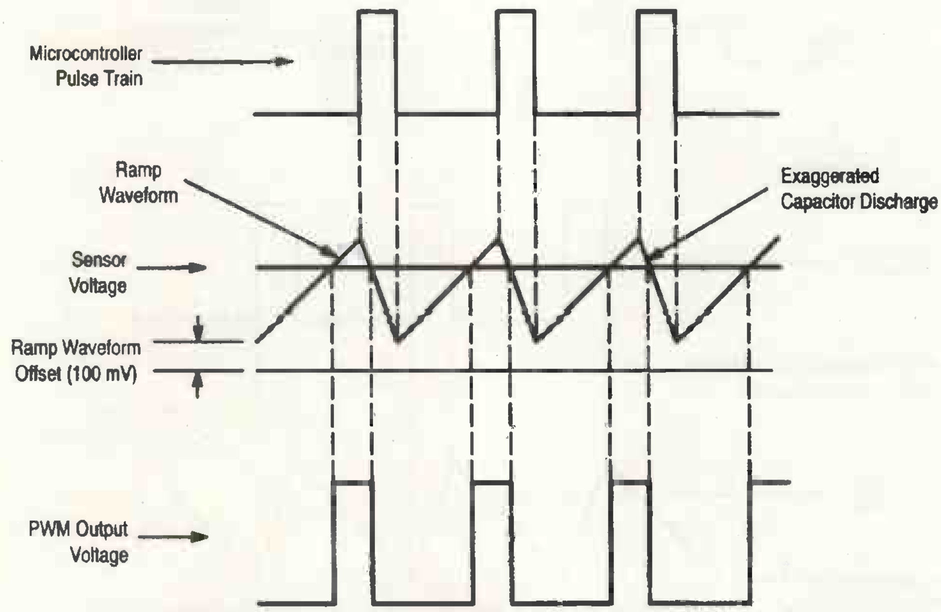
Ideal waveform for a pwm-output pressure sensor.



In this non-ideal pwm output, flats in the ramp valleys are caused by allowing the ramp capacitor to discharge completely. Best waveforms are produced when one ramp cycle begins immediately after its predecessor ends.



Desired relationships between the ramp waveform and pressure sensor voltage spans.



Relationships between pwm output and pressure sensor voltages.

output of 0.5V at zero pressure to 4.5V at full scale pressure.

Note that output of the pressure sensor is attenuated by $R_{1,2}$. This yields a span of 2.0V ranging from 0.25V to 2.25V at the non-inverting terminal of the comparator.

A pulse train output from the microcontroller drives the ramp generator transistor base. This pulse can be accurately controlled in frequency as well as pulse duration via software.

The ramp generator uses a constant current source to charge the capacitor. It is imperative to remember that this current source generates a stable current only when it has approximately 2.5V or more across it. With less voltage across the current source, insufficient voltage causes the current to fluctuate more than desired; thus, a design constraint for the ramp generator will dictate that the capacitor can be charged to only about 2.5V, when using a 5.0V supply.

Constant current charges the capacitor linearly by:

$$\Delta V = \Delta t / C$$

where t is the capacitor's charging time and C is the capacitance.

As shown in the ideal ramp waveform diagram, when the pulse train sent by the microcontroller is low, the transistor is off and the current source charges the capacitor linearly. When the pulse sent by the microcontroller is high, the transistor turns on into saturation, discharging the capacitor.

Duration of the high part of the pulse train determines how long the capacitor discharges, and thus to what voltage it discharges. This is how the dc offset of the ramp waveform may be accurately controlled. Since the transistor saturates at approximately 60mV, very little offset is needed to keep the capacitor from discharging completely.

Pulse-width modulated output is most linear when the ramp waveform period consists mostly of the rising voltage edge. If the capacitor were allowed to completely discharge, a flat line at approximately 60mV would separate the ramps, as indicated in the diagram showing the non-ideal ramp.

These 'flats' may result in non-linearities of the resulting pwm output (after comparing it to the sensor voltage). Thus, the best ramp waveform is produced when one ramp cycle begins immediately after another, and a slight dc offset stops the capacitor from discharging completely.

Flexibility of frequency control of the ramp waveform via the pulse train sent from the microcontroller allows a programmable-frequency pwm output. Using the previous equation, the frequency – or inverse of period – can be calculated with a given capacitor so that the capacitor charges to a maximum ΔV of about 2.5V. Remember that the current source needs around 2.5V across it to output a stable current.

Importance of software control becomes evident here since the selected capacitor may have a tolerance of +20%. By adjusting the frequency and positive width of the pulse train, the desired ramp requirements are readily obtainable and the effects of component variances can be nulled.

For this design, the ramp spans approximately 2.4V from 0.1V to 2.5V. At this voltage span, the current source is stable and results in a linear ramp. To summarise, increasing frequency and/or pulse width reduces the span of the ramp and dc offset.

In the comparator stage, the *LM311* is designed specifically for use as a comparator and thus has short delay times, high slew rate, and an open-collector output. A pull-up resistor at the output is all that is needed to obtain a rail-to-rail output.

As the circuit shows, the pressure sensor output voltage is input to the non-inverting terminal of the op amp and the ramp is input to the inverting terminal. When the pressure sensor voltage is higher than a given ramp voltage, the output is high; likewise, when the pressure sensor voltage is lower than a given ramp voltage, the output is low. Relationships are shown in the diagram.

Since pressure sensor voltage is attenuated and does not reach the ramp's minimum and maximum voltages, there will be a finite minimum and maximum pulse width for the pwm output. These widths are design constraints dictated by the comparator slew rate.

Minimum positive and negative pulse widths are kept at 20μs to avoid nonlinearities at the high and low pressures where the positive duty cycle of the pwm output is at its extremes. Depending on the speed of the microcontroller used in the system, the minimum required pulse width may be larger.

Microcontroller details

For this application, the microcontroller requires input capture and output compare timer channels. The output capture pin is programmed to output the pulse train that drives the ramp generator, and the input capture pin detects edge transitions to measure the pwm output pulse width.

Since software controls the entire system, a calibration routine may be implemented that allows an adjustment of the frequency and pulse width of the pulse train until the desired ramp waveform is obtained.

Depending on the speed of the microcontroller, additional constraints on the minimum and maximum pwm output pulse widths may apply. For this design, the software latency incurred to create the pulse train at the output compare pin is approximately 40μs.

Consequently, the microcontroller cannot create a pulse train with a positive pulse width of less than 40μs. Also, the software that measures the pwm output pulse width at the input capture pin requires approximately 20μs to execute.

Referring to the relationship diagram, the software interrupt that manipulates the pulse train always occurs near an edge detection on the input capture pin (additional software interrupt). Therefore, the minimum pwm output pulse width that can be accurately detected is approximately 20μs+40μs, or 60μs. This constrains the minimum and maximum pulse widths more than the slew rate of the comparator mentioned earlier.

An additional consideration is resolution of the pwm output. It is directly related to the maximum frequency of the pulse train. In this design, 512μs are required to obtain at least 8-bit resolution. This is determined by the fact that a 4MHz crystal yields a 2MHz clock speed in the microcontroller. In turn, this translates to 0.5μs per clock tick.

There are four clock cycles per timer count. This results in 2μs per timer count. Thus, to obtain 8-bit resolution, the difference between the zero pressure and full scale pressure pwm output pulse widths must be at least 2μs×256. But since an additional 60μs is needed at both pressure extremes of the output waveform, the total period must be at least 632μs. This translates to a maximum frequency for the pulse train of approximately 1.6kHz.

With this frequency, voltage span of the ramp generator, and value of current charging the capacitor, the minimum capacitorvalue may be calculated with the earlier equation.

Calibration

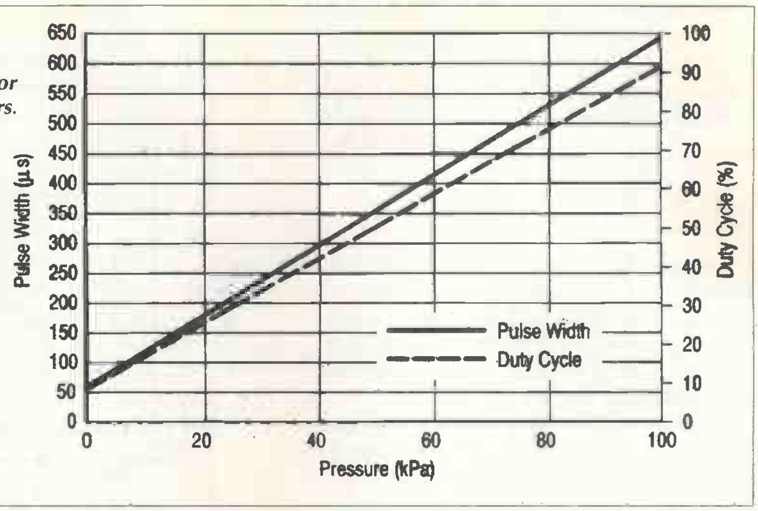
Start with a pulse train that has a pulse width and frequency that creates a ramp with about 100mV dc offset and a span smaller than required. In this example the initial pulse width is 84μs and the initial frequency is 1.85kHz.

Decrease frequency of the pulse train until the ramp span increases to approximately 2.4V. The ramp span of 2.4V ensures that the maximum pulse width at full scale pressure will be at least 60μs less than the total period. Note that by decreasing the frequency of the pulse train, a dc offset will begin to appear. This may result in the ramp looking nonlinear at the top. If the ramp begins to become nonlinear, increase the pulse width to decrease the dc offset.

Repeat the steps in the previous paragraph until the ramp spans 2.4V and has a dc offset of approximately 100mV. The dc offset value is not critical, but the bottom of the ramp should have a 'crisp' point at which the capacitor stops discharging and begins charging. Simply make sure that the minimum pulse width at zero pressure is at least 60μs.

***Motorola**, 8 Tanners Drive, Blakelands, Milton Keynes MK14 5BP. Tel. 0908 614614, fax 618 650.*

PWM output pressure sensor linearity errors.



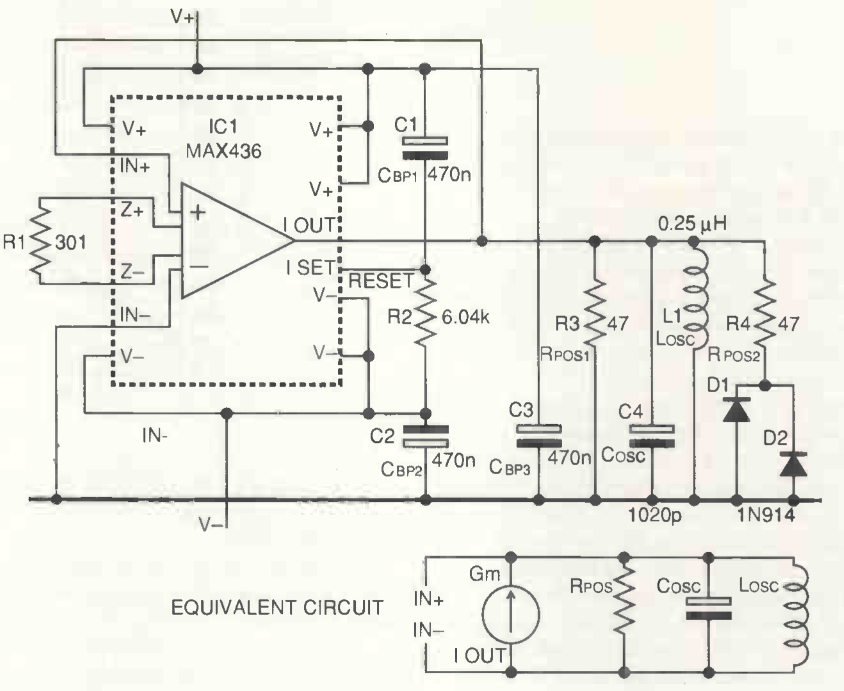
LC oscillator features 1% distortion

At the heart of many oscillators is a parallel-resonant *LC* tank circuit whose impedance is infinite at the resonant frequency of $1/(2\pi\sqrt{LC})$Hz. Infinite impedance implies an absence of parallel damping resistance, so once it starts, an ideal tank circuit should continue oscillating indefinitely.

The actual tank circuit has parasitic resistances that dissipate energy, causing the oscillations to die out. You can counteract this effect by adding a 'negative' resistance, which cancels the net parallel parasitic resistance. Negative resistance is easily synthesized with a wideband transconductance amplifier.

Connect the transconductance amplifier's positive input to its output and its negative input to ground as shown. Now, a positive voltage applied to the output causes current to flow out of the amplifier, in proportion to the applied voltage. The circuit acts like a resistor whose current flows in the opposite direction; hence the negative value.

Negative resistance – easily simulated using a transconductance op-amp – produces an LC oscillator with 1% thd.



Source impedance of the IC's current-source output – at 2.5kΩ minimum – is compatible with the 50 to 300Ω load resistance in applications for which the IC is intended. Load resistance, in this circuit R_3, also resembles that in a typical application.

Load resistance R_3 should be much smaller than the tank-circuit parasitics, yet larger in absolute value than the transconductance amplifier's negative resistance. Resistor R_1 sets the negative resistance in terms of the amplifier's transconductance: $g_m=8/R_1$, where the factor of eight is inherent in the IC.

Negative resistance is therefore $(R_1)/8$, which must be less than R_3. Choosing 47Ω for R_3 yields $R_1<8R_3=376\Omega$. A reasonable value for R_1, therefore, is 301Ω. As intended, the parallel combination of negative resistance ($-(R_1)/8=-37.6\Omega$) and positive R_3(47Ω) yields a negative resistance of –189Ω that shifts the oscillator's complex-conjugate pole pair to the right half plane.

By itself, the combination of tank circuit and regenerative element – negative resistance – simply drives the output amplitude to saturation. To achieve steady oscillation the circuit needs an amplitude limiter. Resistor R_4 serves that purpose; it appears, in parallel with R_3, only when the amplitude is sufficient to turn on one of the diodes D_1 or D_2.

Maxim, 21C Horseshoe Park, Pangbourne, Reading RG8 7JW. Tel. 0734 845 255, Fax 0734 843 863.

Evaluation for PC colour H.261 video i/o chips

An evaluation and prototyping system produced for GEC Plessey's *H.261* video compression and decompression chipset is described in *Application Note AN146*.

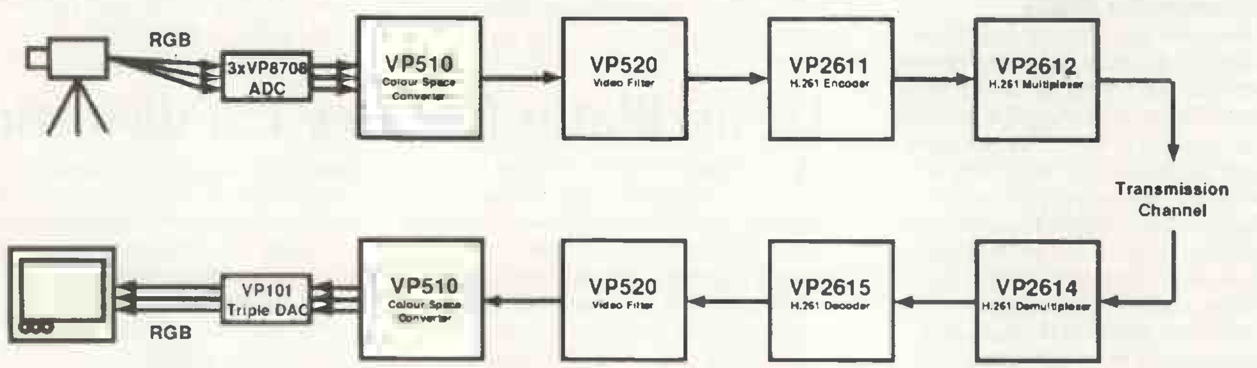
The system is a software configurable IBM PCIXT/AT compatible expansion card supporting coding/decoding of CIF and QCIF images at data rates up to 2Mbits/s and frame rates up to 30Hz. It incorporates three 8-bit video a-to-d converters, provide 24-bit colour accuracy, and triple 8-bit video d-to-a converters for rgb display. All ram requirements are fully localised.

Video, in rgb format, is input to the board from a source which can be gen-locked to the composite sync signal provided by the *VP520* video filter in the decode path. Red green and blue data is sampled at the system clock frequency of 27MHz using an individual a-to-d converter for each channel. This data is colour space converted, filtered and coded to H.261 specification and passed to the transmission channel.

Normally, the transmission channel is a simple link to the decoder section of the board. However this can be intercepted and output to another evaluation board, a network/isdn terminal adaptor or a different H.261 decoder if desired. It is also possible to input H.261 data from another system and decode and display using the *VPB261*.

There is also the option of alternative video i/o formats via two headers on the pcb. Further information in the note covers software, PC interfacing and control signals. Each of the components shown is described in a separate paragraph.

***GEC Plessey Semiconductors**, Cheney Manor, Swindon, Wiltshire SN2 2QW. Tel. 0793 518000, fax 518411.*



System blocks of a PC expansion card for evaluating Plessey's H.261 video chipset via the PC. In addition to taking in rgb, compressing it, and feeding it into a PC, the card also performs the reverse function to drive a monitor or video recorder.

Field Electric Ltd.

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THE WEAPON OF PEACE

As Nigel Cook reports, four out of five ideas of Winston Churchill on the future of war, published in 1925, have become reality. And the fifth? Read on...

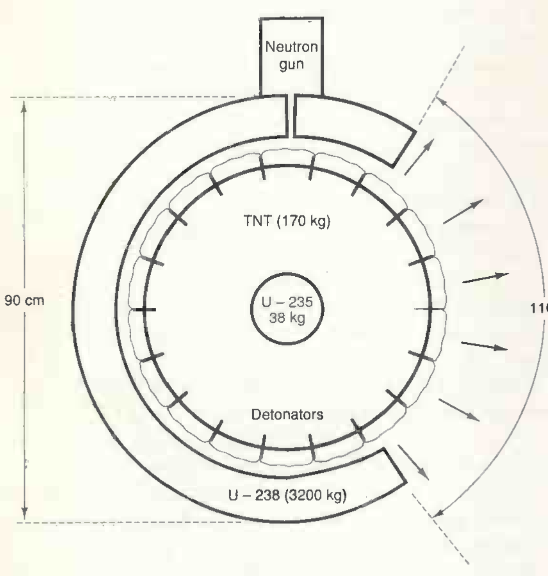


Fig. 1. Vertical cross section through the directed EMP weapon. Prompt gamma rays are radiated horizontally from a circular aperture to maximise the net (asymmetrical) Compton current parallel to the ground. The bomb weighs 3.4 metric tons and is 90cm across. (The bomb dropped on Hiroshima was 3m long and has a mass of 4.1 metric tons)

In an article published in 1925, Winston Churchill posed five ideas on the future of war. The first four have all become reality. The first is nuclear weapons; "Might not a bomb no bigger than an orange be found to possess a secret power – nay, to concentrate the force of a thousand tons of cordite and blast a township at a stroke?" Next is guided missiles and rockets; "Could not explosives even of the existing type be guided automatically in flying machines by wireless or other rays, without a human pilot, in ceaseless procession upon a hostile city?" Thirdly, poison gas and chemical warfare; "only the first chapter has been written in this terrible book." And finally, biological warfare; "Blight to destroy crops, Anthrax to slay horses and cattle, plague to poison not armies but whole districts".

The US Department of Defense has just undertaken the development of the fifth and final suggestion. Churchill's article entitled *Shall We All Commit Suicide?*, offered as the last word on the technology of war a suggestion which previously seemed pseudo-scientific fantasy. But, just like his other ideas, science has finally caught up. Churchill's fifth idea; "It might have been hoped that the electromagnetic waves would in certain scales [frequencies] be found capable of detonating explosives of all kinds from a great distance."

A need has recently arisen for a new weapon which could stop nuclear reactor plutonium production in threatening countries seeking nuclear weapons. According to Pentagon sources, it could also be used to effectively halt conventional warfare without killing or injuring anyone (by destroying the electronic components of weapons), or, indeed, actually detonate all the explosives as Churchill imaginatively suggested 69 years ago. Nuclear reactors cannot be attacked with conventional explosive weapons without the risk of releasing radioactivity which could injure civilians.

Harold Smith, assistant to US defence secretary Les Asin, summarised the requirements in December 1993: "We need a weapon today that will bring a reactor to a standstill, that would not contaminate the surrounding atmosphere." Ashton Carter, assistant US secretary of defence in charge of counter-proliferation, added during the same interview: "We're talking about a new mission." To accomplish this, they have authorised the development of a new bomb which releases an electromagnetic pulse powerful enough to destroy all electronic equipment targeted, without producing early fallout.

The EMP weapon is not an essentially secret invention and can therefore be discussed here in some detail. Like the neutron bomb, the weapon itself is a very fundamental concept to nuclear design, and the special features pertain only to the yield, height of burst, and an outer radiation shield. To optimise the EMP, the fraction of the bomb's total yield which appears in prompt gamma rays must be maximised. Prompt gamma rays are the only source of gamma radiation emitted at a high enough rate (or power) to create a charge separation in the atmosphere sufficient to produce a damaging EMP. About 3.5% of the energy of nuclear fission is released in this form. The shorter the interval of time over which the fission reaction occurs, the greater the rate of prompt gamma emission, the larger the electric field, and the greater the frequency of EMP. Research recently declassified shows that the tamper of a low yield fission bomb absorbs over 85% of the prompt gamma rays.

To meet these objectives the EMP weapon deploys a pure fission implosion bomb with no heavy uranium tamper. This is conventionally used to reflect neutrons back into the fission reaction and to protract the explosion process by inertia, thereby increasing the per-

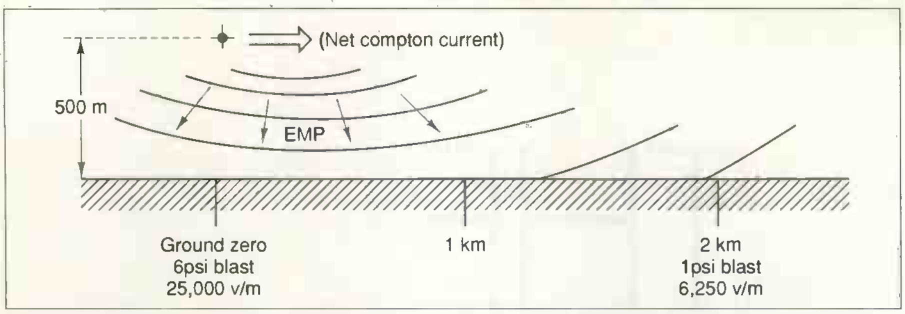


Fig. 2. Directed EMP weapon is used operationally like the neutron bomb. However, the EMP weapon minimises the dangerous effects to people and maximises the high-power microwave (100MHz) electromagnetic pulse emission. This couples immense currents in cables, aerials and power lines which rise faster than lightning and burn out electromagnetic components.

centage of mass fissioned and with it the total yield. The absence of a tamper therefore lets the core expand and the reaction be quenched after only a kiloton of TNT equivalent energy has been generated.

The resulting one kiloton burst at 500m altitude is identical to the neutron bomb; in that the blast, fallout and thermal effects are virtually negligible on the ground. The difference is that the neutron bomb is a complex thermonuclear devise to maximise the neutron radiation yield; the EMP weapon is basically a simple pure fission device maximising the relative yield of prompt gamma rays.

For the purpose of countering nuclear reactors, it is anticipated that EMP bombs will use highly enriched uranium-235 rather than plutonium, since it is too expensive for making a great stockpile of bombs and is many thousands of times safer than plutonium. In fact, the concentration of U-235 needed to cause a radiation injury when ingested is more important for chemical poisoning to the kidneys.

With the source of gamma radiation known, the problem is then using it to produce a directed EMP. The principle is simple: the prompt gamma rays are scattered on average every 300m in the air surrounding the bomb. The scattering mechanism is the Compton effect, whereby an electron in the air is ejected from its atom, leaving the latter ionised.

Since the electron has a far smaller inertia than the ion, it travels outward faster. This separation of charges creates an electric field which, according to Maxwell's equations, is greatest perpendicular to the direction of the so-called 'Compton current'. An ordinary nuclear explosion in the air produces an EMP due to the variation of the air density with altitude, because a vertical asymmetry in the otherwise spherical Compton current is produced. An asymmetry in the radial current is essential for EMP or any radio emission: a radio aerial can be almost any shape except a sphere.

The problem with the natural vertical EMP (which has been understood since the 1960s) is that, if the weapon is air burst (to avoid fallout), the minimum fields occur directly below the detonation, and the maximum fields occur at a long distance in a circular region around ground zero. This effect again occurs in radio transmission: the minimum reception occurs directly above or below a vertical aerial, while the maximum fields are radiated horizontally.

The EMP weapon creates an artificial horizontal asymmetry in the Compton current by absorbing the gamma radiation travelling upwards and downwards from the bomb in a natural outer shield. The prompt gamma rays are all emitted within 10ns (for a one kiloton bomb), which is well before the bomb has destroyed the shield by heat and hydrodynamics. The idea of introducing such artificial asymmetry into nuclear weapons was first put into practice in the successful Ming Blade underground nuclear test at the Nevada Test Site in 1974. This was done to confirm the theoretical model used for surface burst nuclear weapon EMP, so that cold war missile silo equipment could be protected. Of greater interest today are the data from Dining Car, a 1975 nuclear test at Nevada where military hardware was for the first time deliberately subjected to an EMP from a real nuclear explosion. Since the end of the cold war, the Defense Nuclear Agency in America has classified the results of such tests, and even its secret manual entitled *Capabilities of Nuclear Weapons*.

The design of the EMP weapon is shown in **Figs. 1** and **2**. It is a simple and yet highly controllable invention. The heavy radiation shield, while maximising the radiation flash environment high in the air, absorbs most of the downward directed radiations and thus avoids producing casualties on the ground. The exact variation of EMP around ground zero is precisely determined by the solid angle through which radiation is allowed to escape from the bomb, and by height of burst. As the emission angle is increased, a greater amount of prompt gamma radiation escapes. However the symmetry of the radiation field also increases with the emission angle, which means that a smaller fraction of the gamma radiation is then radiated as EMP. On balance, the optimum angle is 110°, for which about one sixth of the prompt gamma radiation is emitted into the air.

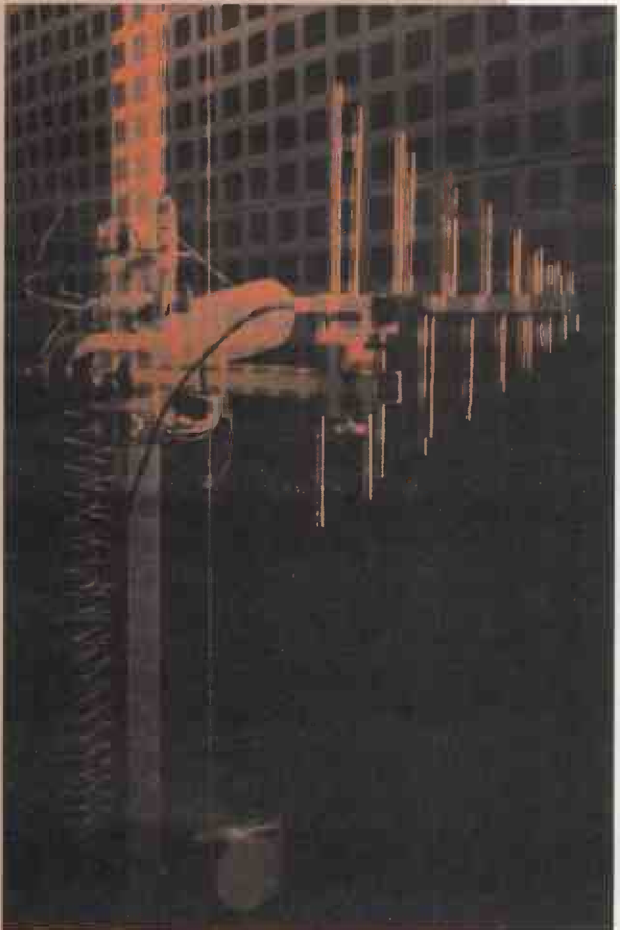
Burst at an altitude of 500m to avoid early radioactive fallout and to achieve a merged and uniform EMP on the surface below, this bomb would blanket a square kilometre with a peak EMP of 25kV/m. At greater distances, e.g., outside the radius of radiation absorption high in the air, the field decays inversely with distance. Therefore, the EMP falls to 6.25V/m at 2km ground radius, and to 2kV/m at 6km.

Experience in 1962 on Hawaii, 1,300km from the 1.4 megaton Starfish Prime nuclear test (detonated over Johnston Island), showed that an EMP of just a few kV/m can cause marked effects even on old electronic systems. For example, 300 street lights were fused in 30 series connected loops, dozens of burglar alarms were set off, and circuit breakers initiated power cuts in different circuits. Except for fuses, electronic equipment was not permanently affected since it takes about 1 to 2J to burn out a valve. However, microelectronics are a crucial component of nuclear reactors and modern weaponry, and they are thousands to millions of times more vulnerable than valve technology. For example, an MC17 silicon chip (data input gate) is burned out, according to the previously secret *Capabilities of Nuclear Weapons*, by an EMP of just 0.08mJ. Furthermore, promising to fulfil Churchill's prediction exactly, we find that various kinds of explosive detonators are fired off by an EMP of between 0.02 and 0.6mJ.

These effects would readily occur out to a distance of between 2 and 6km from ground zero. For comparison, serious skin burns, caused for a one kiloton bomb by a thermal exposure of 5cal/cm^2, occur only to a ground radius of 500m; and the blast wave effect even at ground zero, where the peak overpressure would be 40kPa or 6lb/in^2, would not be sufficient to structurally damage concrete buildings (for instance a nuclear reactor), owing to the very short duration of the blast from a one kiloton bomb. The Pentagon will therefore soon have at its disposal the first true weapon of peace. ■

References

1. Winston Churchill's 1925 article is reprinted in his book *Thoughts and Adventures* (Butterworth, 1932).
2. Assistants to US Secretary of Defense are quoted on the EMP bomb project in various American newspaper reports, e.g., Steve Komarow, *Technology: Pentagon's Driving Force, USA Today*, International Edition, Tuesday 28 December 1993.
3. Secret Pentagon manual recently released under the US Freedom of Information Act: Philip J. Dolan (ed.), *Capabilities of Nuclear Weapons*, Defense Nuclear Agency of the US Department of Defense, 1972. Revised 1978 and 1981.
4. The declassified figure of "over 85%" in the text of the article is based on Chapters 1 and 7 of *Capabilities of Nuclear Weapons*. These state that although thermonuclear weapons release only 0.1% of their total energy as prompt gamma rays, small fission bombs of one kiloton release 0.5%. Since the book states that 3.5% of the total energy of the explosion is in prompt gamma rays, the tamper obviously absorbs the other 3%, which is 85.7% of the total.
5. References to the nuclear test EMP are from Chapter XI of Samuel Glasstone and Philip J Dolan, *The Effects of Nuclear Weapons*, US Departments of Energy and Defense, 3rd ed., 1977.



Scattering knowledge for low-power amps

Put simply, design of low power rf amplifiers means selecting a bias point then making use of scattering and noise parameters. Here, Norm Dye and Helge Granberg analyse both noise and scatter, then deal with bias considerations and power gain. **From the book RF Transistors: principles and practical applications.**

Scattering parameters tell "everything" there is to know about small signal amplifier design with one exception – noise. Impedance matching, gain, input and output vswr, and stability can all be expressed by mathematical equations involving S-parameters. They are basically a means for characterising n-port networks using the concept of travelling waves.

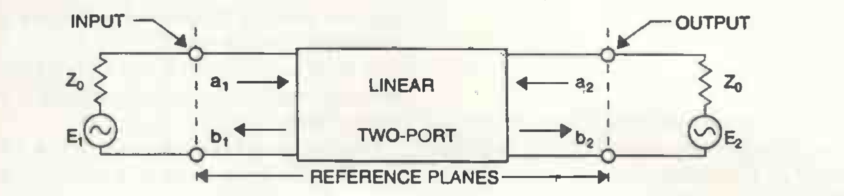
A travelling wave created by a generator (source) and launched on a transmission line toward a load is referred to as an incident wave. Any mis-matches encountered by the incident wave will result in a reflected wave which travels back down the transmission line toward the generator.

For a two-port network such as a transistor, if the network is embedded in a 50Ω measuring system, the S-parameters become simply the coefficients of the incident and reflected voltage waves (**Fig. 1**).

S_{11} and S_{22} in a 50Ω system are the input and output voltage reflection coefficients, related to input and output vswr by the formula

$$\text{vswr} = \frac{1+|\Gamma|}{1-|\Gamma|}$$

where $|\Gamma|$ is the magnitude of the voltage reflection coefficient. The quantity $|S_{21}{}^2|$ is the power gain of the transistor at the specified bias conditions and frequency – and, of



$$\text{vswr} = \frac{1+|\Gamma|}{1-|\Gamma|}$$

Fig 1 Two-port S-parameter definitions course, with 50Ω source and load terminations.

Noise parameters

Three basic noise parameters completely describe the noise characteristics of a low power transistor. These are the minimum possible noise figure obtainable from the transistor NF_{min}, the equivalent noise resistance of the transistor – called R_n, and the optimum source reflection coefficient Γ_{opt}. Sometimes four basic noise parameters are referred to, because the quantity Γ_{opt} is a complex number and is often referenced by stating its magnitude and angle. Also the quantity R_n is sometimes normalised to a specific characteristic line impedance by dividing the quantity by Z_0. In this case, the normalised noise resistance is always specified using the lower case letter r: $r_n = R_n/Z_0$

A given value of noise figure, NF, can be determined:

$$NF = NF_{min} + 4r_n \frac{|\Gamma_s - \Gamma_{opt}|^2}{\left(1-|\Gamma_s|^2\right)\left(1+|\Gamma_{opt}|^2\right)}$$

Once the three noise parameters are known, this shows that the noise figure of a transistor amplifier for a specific bias condition and frequency is entirely dependent on the source impedance seen by the transistor – Γ_s. If the value of NF is specified, the locus of points representing possible values of Γ_s are circles on the Smith chart. The radius of a noise circle will increase with increasing values of NF, with the circle having zero radius being located at the point of Γ_{opt}. The centres of all the NF circles will lie along the Γ_{opt} vector which originates at the centre of the Smith chart and terminates at the location of Γ_{opt}.

Finally, the centres of the noise figure circles are located at the points determined by:

$$C_{Fi} = \frac{\Gamma_{opt}}{1+Ni}$$

where Ni is a noise figure parameter defined by:

$$Ni = \frac{NF_i - NF_{min}}{4r_n} \bullet |1+\Gamma_{opt}|^2$$

and NF_i is the value of the desired noise figure circle. Likewise, the radii of the circles are given by:

$$R_{Fi} = \frac{1}{1+Ni}\left[N_i^2 + N_i\left(1-|\Gamma_{opt}|\right)^2\right]^{0.5}$$

Optimum source reflection coefficient (Γ_{opt}), noise resistance (r_n), and minimum noise figure (NF_{min}) remain the same as described.

Plotting noise figure circles is a tedious operation best left to computers with programs that work in conjunction with Smith chart displays.

Most rf low power transistor manufacturers have automated equipment and computer programs for generating noise (and gain) circles and will provide users with the required information as part of the job of selling their transistors. But if the manufacturer does not supply circles for particular conditions, the alternatives for systematic design of a low noise amplifier are few.

Biasing considerations

Choosing the bias point is less difficult than designing a suitable bias network. First, the manufacturer supplies a curve showing f_τ versus collector current for a bipolar transistor. For good gain characteristics, the transistor should be biased at a collector current that results in maximum or near-maximum f_τ while for best noise characteristics a low current is generally most desirable.

Finally, the maximum signal level expected at the input of the transistor should be considered. The bias point must be at sufficiently high current (and voltage) to prevent the input signal from swinging the collector current out of the "linear" region of operation. The transistor should have been chosen to have sufficient operating current to prevent the input signal from driving the transistor into the saturated region of operation – also an operating condition that would prevent class A (or linear) operation.

If the amplifier needs to work over a range of temperature, a bias network must be designed that maintains the dc bias point as operating temperature changes.

Two basic internal transistor characteristics have a significant effect on the dc bias point – ΔV_{BE} and $\Delta\beta$..

The base-emitter voltage of a bipolar transistor decreases with increasing temperature at the rate of about 2.5mV/°C. Emitter voltage V_E tends to minimise the effect because as base current increases (as V_{BE} decreases), collector current increases causing V_E to increase too.

But as V_E increases, collector current tends to decrease, according to $\Delta I_C = \Delta V_{BE}\, I_C/V_E$, where ΔI_C is the change in I_C, ΔV_{BE} is the change in base-to-emitter voltage, and V_E is the quiescent emitter voltage.

Similarly, the transistor's dc current gain typically rises with increasing temperature at the rate of about 0.5%/°C. Further bias circuit complications arise from the fact that most semiconductor manufacturers give β the least control of any major dc specification. It is not uncommon to have a bipolar transistor with a range of β that exceeds 5 or 6 to 1. That is to say, the ratio of guaranteed maximum β to minimum β is 5 or 6 to 1.

A more normal range is 4:1. Only by special selection can the manufacturer achieve a guaranteed range of 2:1.

Change in collector current for a corresponding change in β can be approximated by:

$$\Delta I_c = I_{c1}\left(\frac{\Delta\beta}{\beta_1\beta_2}\right)\left(1+\frac{R_B}{R_E}\right)$$

where I_{C1} is the collector current at $\beta = \beta_1$, β_1 is the lowest value of β, β_2 is the highest value of β, $\Delta\beta$ is $\beta_2 - \beta_1$, R_B is the parallel combination of the resistors R_1 and R_2 in a bias network, and R_E is the emitter resistor.

The equation shows the advantages of minimum spread in dc current gain: the smaller the value of $\Delta\beta$ (both with temperature, and from transistor to transistor), the lower will be the resulting change in collector current.

Once a transistor is specified, the only control left for the designer is resistance ratio R_B/R_E – unless a more complicated bias network, such as a constant current source, is chosen. Obviously, the smaller this ratio, the

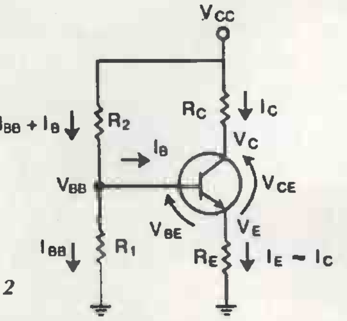


Fig 2

Choose the operating point for the transistor:
I_C=10mA, V_C=10V, V_{CC}=20V, β=50

Assume a value for V_E that considers base stability:
V_E=2.5V

Assume $I_E \approx I_C$ for high-beta transistors.

Knowing I_E and V_E, calculate R_E

$$R_E = \frac{V_E}{I_E} = \frac{2.5}{10\times10^{-3}} = 250\Omega$$

Knowing V_{CC}, V_C and I_C, calculate R_C:

$$R_C = \frac{V_{CC}-V_C}{I_C} = \frac{20-10}{10\times10^{-3}} = 1000\Omega$$

Knowing I_C and β, calculate I_E:

$$I_B = \frac{I_C}{\beta} = 0.2\text{mA}$$

Knowing V_E, V_{BE}, calculate V_{BB}:

$$V_{BB}=V_E+V_{BE} = 2.5+0.7 = 3.2\text{V}$$

Assume a value for I_{BB}, the larger the better, see text:
I_{BB}=1.5mA

Knowing I_{BB} and V_{BB}, calculate R_1:

$$R_1 = \frac{V_{BB}}{I_{BB}} = \frac{3.2}{1.5\times10^{-3}}\Omega = 2133\Omega$$

Knowing V_{CC}, V_{BB}, I_{BB} and I_B, calculate R_2:

$$R_2 = \frac{V_{CC}-V_{BB}}{I_{BB}+I_B} = \frac{20-3.2}{1.7\times10^{-3}} = 9882\Omega$$

less the collector current will vary. However, the lower the value of R_B/R_E, the lower is the current gain of the amplifier. A practical rule of thumb is to keep the ratio less than, but close to, 10.

Figure 2 shows a typical bias circuit together with a simple guide to the necessary calculations (see Chris Bowick's *RF Circuit Design*, Indianapolis, Howard Sams & Co, 1982)

Power gain

Transducer power gain, G_t, is defined as the power delivered to the load, divided by the power available from the source, and is given by:

$$G_t = \frac{|S_{21}|^2\left(1-|\Gamma_S|^2\right)\left(1-|\Gamma_L|\right)^2}{\left|(1-S_{11}\Gamma_S)(1-S_{22}\Gamma_L)-S_{12}S_{21}\Gamma_S\Gamma_L\right|^2}$$

which can be manipulated to to give

$$G_t = \frac{1-|\Gamma_S|^2}{|(1-\Gamma_{IN}\Gamma_S)|^2}\bullet|S_{21}|^2\bullet\frac{1-|\Gamma_L|^2}{|(1-S_{22}\Gamma_L)|^2}$$

where,

$$\Gamma_{IN} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L}$$

or,

$$G_t = \frac{1-|\Gamma_S|^2}{|(1-S_{11}\Gamma_S)|^2}\bullet|S_{21}|^2\bullet\frac{1-|\Gamma_L|^2}{|(1-\Gamma_{OUT}\Gamma_L)|^2}$$

where,

$$\Gamma_{OUT} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_S}$$

Comparing the two equations for G_t, the first relates it to input term $\{1-|\Gamma_s|^2\}/|1-\Gamma_{IN}\Gamma_S|^2$, device term $|S_{21}|$ and output term $\{1-|\Gamma_L|^2\}/|1-S_{22}\Gamma_L|^2$, where the input term is dependent on output quantities.

The second equation shows a similar expression except that the output term depends on input quantities. If source reflection coefficient Γ_s is made equal to the conjugate of the transistor input reflection coefficient Γ_{IN} – ie the transistor input is conjugately matched – we can obtain operating power gain G_p. The importance of G_p is that it is independent of the source impedance because Γ_s was forced to equal S_{11}^*. Operating power gain is:

$$G_P = \frac{1-|\Gamma_S|^2}{1-|\Gamma_{IN}|^2}\bullet|S_{21}|^2\bullet\frac{1-|\Gamma_L|^2}{|(1-S_{22}\Gamma_L)|^2}$$

The above equations for G_t can be solved for known values of load and source reflection coefficients. But the complication is that the load reflection coefficient depends on the source reflection coefficient, and vice versa. So the equation can be solved, but only through an iterative process. At the root of our problem is the term S_{12} – the cause of the interaction of input and output. In some cases, S_{12} is small enough to be considered as zero: such a network is called a unilateral network. But it can not always be neglected.

To obtain an exact solution ($S_{12} \neq 0$), we could use the above equation for G_p and develop a process for determining the load reflection coefficient. But for now, we will assume the network is unilateral, and will work with the simpler equations while developing a technique for determining source and load impedances to obtain desired amplifier performance. Later we can return to the situation when S_{12} is not 0, which is generally the fact in real life.

Impedance or admittance coordinates

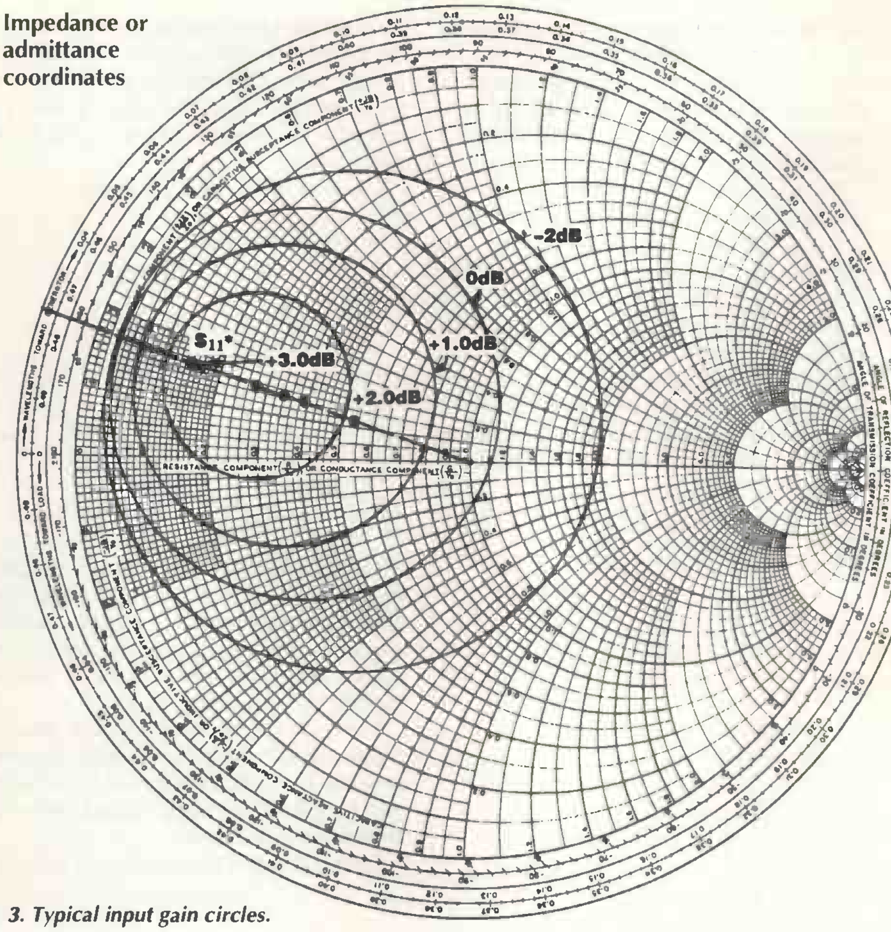


Fig. 3. Typical input gain circles.

One way to verify if a network can be considered unilateral is to calculate a term called the "unilateral figure of merit". This quantity, U, is defined by:

$$U = \frac{|S_{11}||S_{21}||S_{12}||S_{22}|}{\left(1-|S_{11}|^2\right)\left(1-|S_{22}|^2\right)}$$

Defining G_{tu} as the transistor power gain with $S_{12}=0$ and G_t as the actual transistor power gain, the maximum error introduced by using G_{tu} instead of G_t is given by:

$$\frac{1}{(1+U)^2} < \frac{G_t}{G_{tu}} < \frac{1}{(1-U)^2}$$

To illustrate the use of this equation, take the *MRF571* at 1GHz and a bias condition of 6V and 50mA. From the data sheet $|S_{11}| = 0.60$, $|S_{12}| = 0.09$, $|S_{21}| = 4.4$ and $|S_{22}| = 0.11$. So:

$$U = \frac{(0.60)(0.09)(4.4)(0.11)}{\left[1-(0.60)^2\right]\left[1-(0.11)^2\right]} = \frac{0.0261}{(0.64)(0.988)} = 0.0413$$

with the minimum and maximum errors being calcuated as –0.35dB and + 0.37dB.

Frequently, the errors are less than 0.25dB and, as such, are sufficiently small to justify using G_{tu}.

Returning to the expression for G_t and assuming $S_{12} = 0$, the equation becomes:

$$G_{tu} = \frac{1-|\Gamma_S|^2}{|(1-S_{11}\Gamma_S)|^2}\bullet|S_{21}|^2\bullet\frac{1-|\Gamma_L|^2}{|(1-S_{22}\Gamma_L)|^2}$$

and can be broken into three sources of gain; $G_0 = |S_{21}|^2$ which is the contribution of the transistor itself, $G_S = (1-|\Gamma_S|^2)/|1-S_{11}\Gamma_S|^2$ which is the "gain" achieved by the input circuit; and $G_L = (1-|\Gamma_L|^2)/|1-S_{22}\Gamma_L|^2$ which is the "gain" achieved by the output circuit.

If the circuit design is narrow-band and maximum gain is wanted, all that is required is to set $\Gamma_s = S_{11}^*$ and $\Gamma_L = S_{22}^*$.

For broad-band, with a certain amount of gain across a band of frequencies, then circuits must compensate for the variations in gain with frequency of the device itself. Two ways can achieve this: feedback, or selective mismatching.

In selective mismatching, the input and output "gains" are varied (by matching) to compensate for the gain variations with frequency of the transistor, represented by $|S_{21}|^2$.

As Γ_s is varied to other values – thereby causing gain G_s created by the input matching network to vary between 0 and $G_{s,max}$, for a given value of G_s the locus of points representing values of Γ_s is a circle. And in a manner similar to noise circles, the centre of the circle having zero radius is located at the point S_{11}^*. Radius of the gain circles will increase with rising values of G_s. Again the centres of all the gain circles will lie along the S_{11}^* vector which originates at the centre of the Smith chart and terminates at the location of S_{11}^*.

The situation is identical for the output matching network. Another set of "gain" circles (G_L) can be drawn whose centres lie along the S_{22}^* vector which originates at the centre of the Smith chart and terminates at the location of S_{22}^* Typical gain circles are shown in **Fig. 3**.

In a manner similar to noise circles, the gain circles for either the input network or the output network can be drawn on a Smith chart using the following formulas (for the input network):

$$d_s = \frac{g_s|S_{11}|}{1-|S_{11}|^2(1-g_s)}$$

$$R_s = \frac{(1-g_s)^{\frac{1}{2}}\left(1-|S_{11}|^2\right)}{1-|S_{11}|^2(1-g_s)}$$

where

$$g_s = \frac{G_s}{G_{s,\max}}$$

and

$$G_s = \frac{1-|\Gamma_s|^2}{|1-\Gamma_s S_{11}|^2}$$

G_s is the gain represented by the circle, d_s is the distance from the centre of the Smith chart to the centre of the constant gain circle along the vector S_{11}^*, R_s is the radius of the circle and g_s is the normalised gain value for the gain circle G_s.

Likewise, for the output network:

$$d_L = \frac{g_L|S_{22}|}{1-|S_{22}|^2(1-g_L)}$$

$$R_L = \frac{(1-g_L)^{\frac{1}{2}}\left(1-|S_{22}|^2\right)}{1-|S_{22}|^2(1-g_L)}$$

and

$$g_L = \frac{G_L}{G_{L,\max}}$$

$$G_L = \frac{1-|\Gamma_L|^2}{|1-\Gamma_L S_{22}|^2}$$

where G_L is the gain represented by the circle, d_L is the distance from the centre of the Smith chart to the centre of the constant gain circle along the vector S_{22}^*, R_L is the radius of the circle and g_L is the normalised gain value for the gain circle G_L.

These circles represent different values of G_s, the gain created by the input matching network, or G_L the gain created by the output matching network. Negative gain circles can be drawn for both cases.

Input and output gain circles can be used in two kinds of amplifier designs: designing an amplifier with a specified amount of gain; and designing a broad-band amplifier having a specified gain over a band of frequencies. In either case, gain or loss from the input and/or output matching networks can be allocated in whatever manner desired provided the gains (or losses) are actually realisable. The maximum available gain can be determined at any frequency by conjugate matching. Obviously, more gain than this value at a specified frequency can not be achieved.

Half the loss (or gain) is often assigned to both input and output circuits, although this is not essential. ■

Next article – practical examples.

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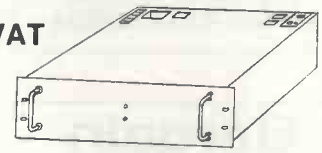

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