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A new class of mosfet power amplifier – Bengt Olsson challenges the 30-year-old complementary output stage.

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The European quality alternative
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 returning 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 swaths 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 E+W+W 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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Watch tv through a window. Apple's new Performa 630 TV Plus includes a TV tuner for its Multimedia environment.

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 AutoPilot 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.

Live ‘94 – surround sound to Internet

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 no pictures together? Or if they’re a different shape? The images and 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 8 megabit/s, the equivalent of four channels into one. As for sheer size,

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UPDATE

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 have switched from games machines to more powerful and versatile CD-based systems. Panasonic was launching its own service in October.

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Outside the home, the most significant pointer to the future was the 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.

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Peter Willis

Cmos supersedes 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.

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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November 1994 ELECTRONICS WORLD+WIRELESS WORLD
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 a 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 rotate 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 a high-speed 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 re-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.

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.

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 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 labs says, “You may even be able to use thousands of carriers at distances of perhaps 100km.”

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 2x2 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 wavelength 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
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 planar aerials... The list will hopefully be endless.

All articles accepted for publication will be paid for – in the region of several hundred pounds for a typical design feature.

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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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).

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 constraining 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 busline. 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 from 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 10 in 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.

In addition, 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. Seperating 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

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 (b) 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.
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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

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 512x432 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 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.
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 that using both materials could make 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 Angstroms, while the organic component is a layer of p-paraphenylen 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. 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.
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.

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 adopted by the OU and in other educational establishments worldwide. 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.

Circuits by 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 excitation 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.

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 numerical 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 type Component name Connection Value

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_sensor’ for the sensing thermistor, and ‘R_bal’ for the balancing preset. These names are helpful when interpreting graphs and tables, but meaningless to the analyzing 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 p1 and p2. 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 p1 and which is p2. 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 ‘=’ 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_sensor. 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 used 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.

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.

<table>
<thead>
<tr>
<th>Component type</th>
<th>Component name</th>
<th>Connection to Node</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>V Excite</td>
<td>-out: pos</td>
<td>+out: pos</td>
<td>None</td>
</tr>
<tr>
<td>R Rsens</td>
<td>p1: pos</td>
<td>p2: 1</td>
<td>42.890kΩ</td>
</tr>
<tr>
<td>R Rbal</td>
<td>p1: 1</td>
<td>p2: gnd</td>
<td>42.890kΩ</td>
</tr>
<tr>
<td>R R1</td>
<td>p1: pos</td>
<td>p2: 3</td>
<td>47.000kΩ</td>
</tr>
<tr>
<td>R R2</td>
<td>p1: 3</td>
<td>p2: gnd</td>
<td>47.000kΩ</td>
</tr>
<tr>
<td>R Re coil</td>
<td>p1: 1</td>
<td>p2: 2</td>
<td>650.000Ω</td>
</tr>
<tr>
<td>R Rscale</td>
<td>p1: 2</td>
<td>p2: 3</td>
<td>100.000Ω</td>
</tr>
<tr>
<td>Reference Node</td>
<td>-out: pos</td>
<td>+out: pos</td>
<td>None</td>
</tr>
<tr>
<td>Output Node</td>
<td>pos</td>
<td>ref</td>
<td>0.00000</td>
</tr>
<tr>
<td>Reference Node</td>
<td>-out: ref</td>
<td>+out: pos</td>
<td>1.00000</td>
</tr>
</tbody>
</table>

Resistance of a thermistor
At any given temperature T, resistance R is given by:

\[ R = R_{ref}(\frac{1}{1 + \frac{T - T_{ref}}{T_{ref}}} \] where \( R_{ref} \) is its resistance at \( T_{ref} \), and temperatures are in kelvin.

Two of the resistors are preset; 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.
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 Rcoil. The table of results Table 3 shows the currents; there is no current through Rsens and Rscale, because the bridge is balanced. All other currents are small, of the order of tens of microamps.

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 pragmatic 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.

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

<table>
<thead>
<tr>
<th>Ref. Node</th>
<th>gnd</th>
<th>Name</th>
<th>Volts</th>
<th>Node</th>
<th>Volts</th>
</tr>
</thead>
<tbody>
<tr>
<td>pos</td>
<td>2.5000000</td>
<td>1</td>
<td>1.2500000</td>
<td>3</td>
<td>1.2500000</td>
</tr>
<tr>
<td>2</td>
<td>1.2500000</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**Table 3. Analysis of currents through the various elements of the thermometer bridge.**

<table>
<thead>
<tr>
<th>Component</th>
<th>Rsens</th>
<th>Rscale</th>
</tr>
</thead>
<tbody>
<tr>
<td>VExcite</td>
<td>55.74407u</td>
<td></td>
</tr>
<tr>
<td>R R1</td>
<td>26.59574u</td>
<td></td>
</tr>
<tr>
<td>R Rcoil</td>
<td>0.0000000</td>
<td></td>
</tr>
</tbody>
</table>

**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.

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**ELECTRONICS WORLD + WIRELESS WORLD November 1994**
metry. This indicates that $R_{wa}$ 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 550pW, 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: $(R_2+2976)i_4-2976i_5-R_6i_6=12$

Mesh 2: $-2975n+(50626+R_s)i_3-(650+R_2)i_6=0$

Mesh 3: $-R_2i_4-(650+R_2)i_5+(47650+R_s+r_{b})i_6=0$

These are the equations at 0°C, when the resistance of the thermistor is 165.174kΩ. Resistance $R_1$ is abbreviated to $R_{bal}$, and $R_{scale}$ to $R_s$. The set of equations is similar at 100°C, when the resistance of the thermistor is 2.976kΩ, but using $i_4$, $i_5$, and $i_6$ to represent the currents:

Mesh 1: $(R_b+2976)i_4-2976i_5-R_6i_6=12$

Mesh 2: $-2975n+(50626+R_s)i_3-(650+R_2)i_6=0$

Mesh 3: $-R_2i_4-(650+R_2)i_5+(47650+R_s+r_{b})i_6=0$

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:

$i_4-i_5=0.000125$

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, In(2):= is followed by the input command, in *Mathematica* syntax. The command ‘N Solve’ 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 $R_{bal}$ and $i[3]$ for $i_3$. The asterisk represents ‘multiply’, though it is permissible to leave a space instead. The ‘$\Rightarrow$’ 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 $R_{sens}$ changes with temperature. For ordinary resistors in the netlist, state the temperature coefficient. For example, the statement ‘Te=0.00005’ indicates a temperature coefficient.

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

<table>
<thead>
<tr>
<th>R</th>
<th>Rsens</th>
<th>Rbal</th>
<th>Rs</th>
<th>Rscale</th>
</tr>
</thead>
<tbody>
<tr>
<td>p1:pos</td>
<td>p2:pos</td>
<td>p1:1</td>
<td>p2:2</td>
<td>p2:gnd</td>
</tr>
<tr>
<td>v=650.000</td>
<td>v=47.000k</td>
<td>v=13.600k</td>
<td>v=47.000k</td>
<td></td>
</tr>
<tr>
<td>Rsens</td>
<td>Rbal</td>
<td>R2</td>
<td>Rcoil</td>
<td>Rscale</td>
</tr>
<tr>
<td>p1:pos</td>
<td>p2:pos</td>
<td>p1:2</td>
<td>p1:2</td>
<td>p1:2</td>
</tr>
<tr>
<td>v=650.000</td>
<td>v=47.000k</td>
<td>v=850.0000</td>
<td>v=4.200000</td>
<td></td>
</tr>
</tbody>
</table>

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

```
In[2]:= N\[Solve\](\{r[b]+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[1]
,i[2],i[3],i[4],i[5],i[6]\})
```

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


Table 6. In this test-circuit representation, a constant-current of 1mA connects directly across resistor R1.

<table>
<thead>
<tr>
<th>MODEL</th>
<th>resmod</th>
<th>res</th>
<th>V</th>
<th>p1:pos</th>
<th>p1:1</th>
<th>p1:3</th>
<th>p2:pos</th>
<th>p2:2</th>
<th>p2:2</th>
<th>p2:2</th>
</tr>
</thead>
<tbody>
<tr>
<td>A1</td>
<td>out:0</td>
<td>+ out:1</td>
<td>v=1.000000m</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>R1</td>
<td>p1:p</td>
<td>p2:0</td>
<td>v=42.8900k</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

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

<table>
<thead>
<tr>
<th>MODEL</th>
<th>resmod</th>
<th>res</th>
<th>V</th>
<th>p1:pos</th>
<th>p1:1</th>
<th>p1:3</th>
<th>p2:pos</th>
<th>p2:2</th>
<th>p2:2</th>
<th>p2:2</th>
</tr>
</thead>
<tbody>
<tr>
<td>*CBD1b - Thermometer test circuit</td>
<td>resmod</td>
<td>res</td>
<td>V</td>
<td>p1:pos</td>
<td>p1:1</td>
<td>p1:3</td>
<td>p2:pos</td>
<td>p2:2</td>
<td>p2:2</td>
<td>p2:2</td>
</tr>
<tr>
<td>A1</td>
<td>out:0</td>
<td>+ out:1</td>
<td>v=1.000000m</td>
<td></td>
<td></td>
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<td></td>
</tr>
<tr>
<td>R1</td>
<td>p1:p</td>
<td>p2:0</td>
<td>v=42.8900k</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

\[ r = 0.05146e^{4090/(t+273)} \]

SpiceAge allows for an exponential response of the form:

\[ r = R(1 + TC1(t-27) + TC2(t-27)^2)e^{TCE(t-27)/100} \]

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 (1mA) connected directly across resistor R1, which has the nominal resistance of Rsens at 27°C. This netlist models the 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 easy 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 Rsens with a thermistor model by adding the model description and by adding Mo=resmod to the specification of Rsens in 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°C. At 0°C, the current is almost exactly -125μA, as required. It slightly exceeds 125μA at 100°C, 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 confidence that it will work. Any slight discrepancies at the ends of the scale can be corrected by adjusting the real Rbal and Rscaie. 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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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.

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 $Q_3$ 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.

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

Low-power, infra-red preamplifier. For higher supply voltages, adjust the bias chain to produce 1.6V at $Q_3$ base.

---

Gain/frequency response of prototype preamplifier.
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**CIRCLE NO. 115 ON REPLY CARD**
A new era in magnetics

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 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 geometry 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 behaving as sputtering and molecular beam epitaxy, as outlined in the first and second panels.

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.
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 as a stainless steel.

**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.

**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.

**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.
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 agfm.

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.

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.
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 aluminiumisation 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 $10^{-2}$ 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 gas, 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 causes 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 sensitive to the electronic state of the sample, in a similar manner to that in which giant magnetoresistive devices are used to sense magnetic fields. 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 field required, and the greater the storage density possible.

The recording industry has good reason to be interested in materials with a high magnetoresistive response. 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 fields required. This is particularly true for giant magnetoresistive devices, which have a background of over 2000 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.

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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 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.

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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 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 magnetoresistance, 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 case 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.

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FLOPPY DISK DRIVES 3.5", 5.25"

- 5.25" CDC 94205-51 40mb HH MFM I/F RFE tested
- 5.25" SEAGATE ST -238R 30 mb RLL I/F Refurb
- 3.5" FUJI FK-309-26 20mb MFM I/F RFE

For more details, refer to the full listing provided in the document.
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.

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).

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. 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 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 1/50th the price of liquid helium, but its latent heat of vaporisation is about 60 times higher, so that LN2 overall is considerably more cost-effective.
High-Tc 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 235cm 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:

$$\lambda (T) = \frac{\lambda (0)}{(1 - T/T_C)^{1/2}}$$

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 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 (1cm² 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 (LaAlO₃) which has a dielectric constant, $\varepsilon_r$ of about 24; magnesium oxide (MgO) with $\varepsilon_r=10$; yttria-stabilised zirconia (YSZ) with $\varepsilon_r=30$; and, more recently, crystals of neodymium gallate having $\varepsilon_r=22$. With reference to the earlier article on microstrip circuits: the above values of $\varepsilon_r$ 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 $\tan \delta$ is often the limiting factor on overall performance due to the low dielectric Q-factor ($1/\tan \delta$).

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 TE₄₁₁ mode and had a Q-factor of 715,000 at 27K; 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.
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 $\delta_s$ where:

$$\delta_s = \frac{2}{\omega \mu_0 \sigma}$$

and $\omega$, $\mu$, $\sigma$ are the radial frequency, permeability and conductivity of the conductor respectively.

The surface resistance is given by $R_s = \frac{\rho_s}{\delta_s^2}$ 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, $10^{-6}$ 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 $10^6$ A/cm$^2$ at magnetic field strengths in excess of 20T can be achieved.

**Critical temperature for the onset of superconductivity**

$$T_C = \frac{2}{\omega \mu_0 \sigma}$$

where $T_C$ and $H_C$ are the critical temperature and magnetic field respectively. 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, $10^{-6}$ 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 $10^6$ A/cm$^2$ at magnetic field strengths in excess of 20T can be achieved.
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 achieveable 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 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 communications 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.

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 1/8th of a wavelength λ such that a linear current distribution can be assumed, the radiation resistance is given by:

\[ R_{\text{rad}} = 80\pi \left( \frac{1}{\lambda} \right)^2 \]

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 tolerances, thereby negating the initial size advantage. However, if the microwave device was made superconductive 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 electically 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 to match the capacitive reactance of the dipole. Since this first demonstration, much further work has been done internationally to develop 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 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-chip and down-chip 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 to perform such time domain functions (squids).

Furthermore, by suitable choice of conductor and dielectric thickness of microstrip or coplanar line, 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 dependent, superconducting circuits can show dispersionless behaviour up to about 10Hz, 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-Tc 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 TBBCCO 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-Tc 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-Tc superconductors and I am sure that the next few years will witness their consolidation into many commercial systems.
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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 routing 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 routing 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 100kΩ 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 autocali-
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.
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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 IC₁ are low, isolator OC₁ pin 4 is off and its output, which is SDA(2) is high. When SDA(1) is low, the isolator conducts, as does D₄, and SDA(2) is also low. This level goes through the inverters of IC₂ and, were it not for D₃, would activate OC₂ and latch the circuit up. However, the signal through the IC₂ inverters is delayed, D₃ conducts and prevents OC₂ conducting. Diode D₁ 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 D₁ and D₄. Lines SCL and INT are the clock and interrupt lines and are unidirectional.

Falko K Kuhnke
Braunschweig, Germany

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**YOU COULD BE USING A 1GHz SPECTRUM ANALYSER ADAPTOR!**

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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 $R_{14}$ 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.

CID 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.
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 position 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

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, IC1, is an adjustable 1A voltage regulator, configured here as a current regulator. Voltage generated across C2 varies as the battery charges to maintain around 1.2V across R1, which is fed back to the regulator. Gain in op-amp IC2 allows a low value of R1 to further reduce total power dissipation the circuit, C3 being necessary to slow down the response of IC1's internal control loop.

With these values, charging current is constant at 520mA, easily changeable to a different value by altering R3. For much higher rates of charge, the 3A version of IC1, the LM2576-ADJ, can be used with up-graded D1 and L1. No heat sink is needed for IC1 with these values, dissipation being a constant 0.35W, with a conversion efficiency of over 85%.

Huw Jones
Gyys Medical Ltd
Cardiff
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

<table>
<thead>
<tr>
<th>Type of filter</th>
<th>Condition for realisation</th>
<th>Gain factor</th>
<th>Other parameters: (pole Q, bandwidth)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Low-pass</td>
<td></td>
<td>$\frac{R_2}{R_3}$</td>
<td>$Q_0 = \frac{R_1}{R_2 + R_3} \sqrt{\frac{R_4 C_2}{R_1 C_1}}$</td>
</tr>
<tr>
<td>All-pass</td>
<td>$\frac{2 R_4}{R_5} = \frac{1}{4} \frac{2 R_3}{R_2}$</td>
<td>$\frac{R_2}{2 R_4}$</td>
<td></td>
</tr>
<tr>
<td>Band-pass</td>
<td></td>
<td>$\frac{R_2}{R_2 + R_3}$</td>
<td>BW $= \frac{R_2 + R_3}{R_2 R_3 C_1}$</td>
</tr>
<tr>
<td>Band-stop</td>
<td>$\frac{R_1}{R_2} = \frac{2 R_3}{R_3}$</td>
<td>$\frac{R_2}{2 R_4}$</td>
<td></td>
</tr>
</tbody>
</table>

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

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.

*Phototronics, PO Box 977, Manotick, Ontario, K4M 1A8 Canada. Tel. 010 613 692 2247, fax 010 613 692 2605.
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, Tr2 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 1N914 voltage, forces a reverse current in Tr2, switching it off. When the negative cycle starts, what is left of the voltage across C reinforces it and ensures Tr2 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

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, D2 and SCR1 passing only leakage current. When the coin closes LS2, SCR1 switches the motor on and the cam rotates 180°, opening the liquid valve. As the cam turns, it switches LS1, momentarily interrupting the motor supply and turning SCR1 off. Timing circuit R1C1 now receives voltage and UJT1 eventually fires and restarts the motor, which drives the cam back to the start and turns off the water valve. The motor stops as LS1 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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Maxcom’s MX170B is a compact 3.5-digit instrument featuring six measurement functions plus over-range and high-voltage indicators. In addition to reading dc and ac voltage, dc current and resistance, the MX170B also has a diode-test function and three battery-test-under-load ranges. Weighing only 150g and measuring just 70 by 116 by 24mm, this is a truly pocket-sized instrument.

The MX170B has a liquid-crystal readout and is accurate to ±0.5%, ±2 digits on its 2V range.
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MINIOYCLPS PIR 52x64x40mm runs on PP3 battery complete. £49.99 ref MAG15.


DOS PACK Microsoft version 5 Original software but no manuals or comprehensive manual! 5.25 FDD only.

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, dV/dr 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)
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

![Logic gate diagram](image)

Figure 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.
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 negatively 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 $r_{eb}$ is however useful and effective; it just requires some extra 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 output 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.

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 $-180°$ 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.

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 $R_4$ 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 voltage 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 $R_3$) if the battery pack is stored with the terminals short-circuited.

Capacitor $C_1$ is the clever part of the circuit. As explained, on an 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 D3 cures the problem.

a discharge path for C1 via the base emitter junction of Tr1 in series with D1 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 C1. Adding D3, 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 an $n:1$ tuning range. The circuit of Fig. 6 seems to break through this limitation. And it should
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 $Q$. As the wiper of $Q$ 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 $Q$ is at the top ($e_3 = e_2$), the wiper of $R_1$ is at mid-travel, and the input signal is a sine wave of frequency of $f_0 = 1/(2\pi CR)$. At this frequency, both the integrator and the differentiator have a gain of unity, the integrator output $e_1$ leading $e_0$ by $90^\circ$ (it is an inverting integrator) and the differentiator output $e_2$ lagging by $90^\circ$. Net voltage at the wiper of the tuning control $RT$, $e_3$, is zero and $e_0 = -e_1$. If the wiper of $R_1$ moves towards the integrator output, $e_3$ is zero at a somewhat higher frequency, or at a lower frequency if moved toward the differentiator output. In the limit, at the end of $R_1$'s travel, for zero output at the wiper, one output must be $(R_1+R)/N$ times the other, i.e. at $N$ times $f_0$. So the tuning range is from $f_0/N$ to $f_0/N$, or $N:1$. As $r$ is made smaller relative to $R_1$, 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 $e_0 = e_2$ (minimum Q). Clearly, as $e_0$ increases off tune, so $e_0$ becomes smaller relative to $e_2$, so $e_3$ is no longer zero. Still, off-tune the output will not be far below unity so long as $e_3$ is small compared to $e_2$. The allowable detuning while still meeting this condition gets smaller as $e_3$ becomes a smaller proportion of $e_2$. Finally, as the wiper of $Q$ approaches ground and $e_3$ 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 $e_0$. Put another way, in these circumstances, $A_1$ will produce whatever output is necessary to keep the signal at its non-inverting input equal to that at its input. If $e_3$ does not equal $e_2$, the only way it can arrange this is if $e_3$ 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 so high that it can work at a much higher 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 its 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 claimed that 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 triac 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?)

References

INSTRUMENTATION

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 \( R = 100k\Omega \) and \( C = 1\mu F \), 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 \( A_2 \) 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 \( A_2 \) 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 \( R \) 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 \( Y \) is increased, \( R \) 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, \( R \), 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.](image)

![Fig. 2. State-variable filter with zero damping and no input functions as an oscillator, but only just.](image)
INSTRUMENTATION

between $S_1$ and $A_4$, along with the $\times10$ step attenuator. Now, when $R_v$ is set to zero, zero voltage is applied to $Y_x$, and so resistance and attenuator. Now, when $R_v$ is set to zero, zero

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.

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 11V-pk-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 $A_{16}$, which is advanced by 90° relative to the $A_{16}$ output which was used for resistive unknowns. As before, $S_1$ 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 ($R_{18}$); Fig. 5 shows a value of 101dI, which provides inductance ranges of 0-1H down to 0-10pH.

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 ($R_{19}$) while the voltage applied to the unknown capacitor is varied only in $\times10$ steps. To allow for the leading nature of the current through a capacitor, $S_{1a}$ selects the lagging voltage from $A_{16}$ 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 1nF of unity, or of course a capacitive susceptance including a negative conductance component.

Amplifier $A_3$ 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 $A_3$ is that remarkable op-amp the TLE2027. This was described in an earlier Design Brief 2, where it was shown driving 23Vpk-pk into 1pF at 318Hz. Here, it is required to drive up to 10μF at 1591Hz, but only at 110Vpk-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 $A_3$, which thus provides a signal to the detector stage, shown in Fig. 6. A 1MΩ 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 $C_{0}$ provides additional smoothing for the 25V dc power supply, which is split into ±12.5V supplies for the op-amps by the TLE426 '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 $C_{0}$ to the other end of the centre-tapped winding, or by connecting a fixed capacitor of value $C_{0}$ in parallel with the unknown susceptance $Y_z$; the shunt inductive component of $Y_z$ is then measured as an equivalent negative capacitance. Simple arrangements for $R_v$ enable negative conductance components of $Y_z$ 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 $T_{1}$ also. It turns out that not only is this possible, but one can actually eliminate $C_{0}$ as well.

Note that with the voltage applied to $C_{0}$ in the phase shown in Fig. 1, balance can be achieved with a capacitive $Y_z$, while with a voltage in the opposite phase applied to $C_{0}$ (from the other end of the centre-tapped winding of $T_{1}$, 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 $Y_z$ against $C_{0}$ 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. 932
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 \( R_2 \), 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 \( R_{s\text{min}} = R_s \), i.e. from zero conductance up to \( G_{s\text{max}} \). 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 \( 1/G_{s\text{effectiv}} \)), 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 \( R_{s\text{min}} = R_s \), unknown resistors down to the same value can be measured. If an attenuator with steps of \( x_1, x_{10}, x_{100} \ldots \) is fitted between \( S_1 \) and \( A_4 \), resistors down to \( R_{s\text{min}}/10, R_{s\text{min}}/100 \) 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.

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 \( R_e \). 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 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
INSTRUMENTATION

The maximum capacitance that can be measured at balance is limited by 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 (R1) 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 nanoheny range. To test low-inductance measurements, four turns were wound on a two-hole balun core type F32754, which uses 3C85 material; the expected inductance, given the core’s A, of around 3500H/turn, was 56H.

On range 5 of S2, the measured value was 52μH with the dissipation or quadrature control R12 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 R12 set at 350°, 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 point, which can conveniently be 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 R12 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 hence 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 S1, 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 0-1kH up to 0-100H. An even better scheme would be two ‘L’ positions on S1, 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 100pF.

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 A1, 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 S1 and S2 should be on the switches themselves and, unduly though it may look, connections from the switches and R12, R13 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 A2 output is as close as possible to the rear spill of the D terminal, say less than an inch of stout wire, and S1b wiper should be returned direct to the rear of the VE terminal, from whence the lead to C2. As far as accuracy is concerned, all resistors should be 1% or better, but more importantly R6 and R7 should both equal R12, R13 and R11 should equal R12, R13 and R10 should each be half of R6, and R9 should equal the resistance from S1a wiper to ground. The SVF oscillator frequency should of course be as close to 10kHz 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 8.8Ω 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 reap- peared 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 R3A and A1 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

**JPG Electronics**

- Large range of CMOS TLI. 74HC 74F Linear Transistors kits.
- Rechargeable batteries, capacitors, tools etc.
- Standard grade capacitors with screw terminals.
- Nickel Metal Hydride AA cells high capacity with 500mAH compression trimmer.
- High power chargers as above but charges AAA, D size and C size with solder tags.
- Computer grade capacitors with screw terminals.
- Digital and Analogue Simulation.
- Microcontroller (£9.95 10+, £7.95 100+).
- £0.85 package.
- 38000uf 20v memory. 1000mA £3.85 2A £4.40.
- £50.00 per 100.
- Nickel Metal Hydride AA cells high capacity with 500mAH compression trimmer.
- High quality photo resist copper clad epoxy glass.
- High intensity red, green or yellow 5mm diam £30p each.
- High power chargers as above but charges AAA, D size and C size with solder tags.
- 6012 inches £6.20 each.

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

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 interrelated, 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...', 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 s 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 \(s\times\) 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 rep-
Fig. 3. In this description of a first order filter, the $s$ plane is three-dimensional space, where the real and imaginary axes are at 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 ($\tan\theta = \text{opposite/adjacent}$).

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, e.g. $3+j2 + 5+j2 = 7+j4$; 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, so doing 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° anti-clockwise 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 $-i$ times the original real number. Since $j$ squared was equal to $-1$, $j$ must be the square root of $-1$. This is an important concept to grasp, because one of the first things learnt at school is that you cannot take the square root of a negative number.

Practical use of complex numbers
Consider what happens when a sinusoidal voltage is applied to a 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 $P=VI$.

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 $V=IZ$, the impedance is $j\omega L$, where $\omega$ is 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.

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 $-j/\omega C$ 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 magnitude of the current.

A complex number can be converted into a number with magnitude and phase angle (a polar representation), but 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 magnitude squared added to the imaginary part squared; the phase angle is the arc-tan 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 voltage is applied, 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 $(R^2+\omega^2L^2)$; the phase angle is $\tan^{-1}(\omega L/R)$. The smaller $R$ becomes, the more the inductor behaves like a pure inductance. The power dissipated in the winding resistance is given by $P=VR$. 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 $2\pi$ multiplied by the frequency of the applied voltage.

The s plane
The s plane is a complex frequency diagram with two axes. Real numbers are on the horizontal $\Sigma$ axis and imaginary numbers on the vertical $\omega$ axis.

Fig. 2. s-plane complex frequency diagram with two axes. Real numbers are on the horizontal $\Sigma$ axis and imaginary numbers on the vertical $\omega$ axis.
The frequency of an applied signal increases, its presence for measurements. The resistor and ammeter across the load reduces. If the load is a common connection from one side of the series between the signal source and the load; consider a simple RC filter. A resistor is in components in the s domain. When a steady sinusoidal signal is used, its ease of describing components. 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 $\Sigma + j\omega$. If this equation is expanded, we get $e^{(1+j\omega)t} = e^{e^{j\omega t}}e^{-\omega t}$. The $e^{j\omega t}$ is the exponential decay while the $e^{\omega t}$ part is the steady state oscillation. If $\Sigma = 0$, 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 $\Sigma = 0$. An inductor’s reactance, $j\omega L$, becomes $sL$ and a capacitor’s reactance, $1/j\omega C$, becomes $1/sC$.

Real poles
Consider a simple RC filter. A resistor is in series with 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 $(1+1/CR)$, which simplifies to $1/(s+1/CR)$. Now, $CR$ is the time constant of an RC network, sometimes described by $\tau$. 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 on the negative real axis because there are no oscillatory components in the circuit.

If an input 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 $\tau = CR$, as the charge stored in the capacitor $C$ drains away through the resistor $R$.

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 $\omega_0$ or $2\pi f_0$ is equal to $1/CR$. The frequency $f_0$ 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 plane becomes three dimensional, see Fig. 3. Consider the height of the canvas at points along the imaginary $j\omega$ axis. It is highest at the zero frequency point, where it crosses the real $\Sigma$ 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 $\omega$ axis. The height of the canvas at the $\Sigma$ axis is the dc attenuation which is zero; the attenuation at other points along the $\omega$ 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 coincident, but by Pythagoras Theorem the distance from the pole to half power frequency is $\sqrt{2}$ 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 frequency 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 is known as an $L-C$ network. The transfer function, $V_o/V_i$, 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 oscillation 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 $\omega$ is given by the product of all the poles to the origin distances divided by the product of all the pole to $\omega$ axis distances. If the poles are equally spaced and lie on a semi-circle, the product of distances to points along the imaginary axis remains almost constant, up to the point where the semi-circle crosses the imaginary axis. A circuit represented by
Fig. 5. Bode plots illustrating frequency compensation. 

(a) Desired response
(b) Op-amp response
(c) Lead-lag network response
(d) Uncompensated filter response

Fig. 6. 150kHz 0.5dB Chebyshev lowpass filter. Resistors R3,6 are zero ohms in the uncompensated filter, otherwise, R2,5 are reduced by the value of R3,6 respectively.

ANALOGUE DESIGN

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 delay 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 remains a constant.

A pole-zero application
A previous article described how lowpass filters were dependent 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 phase shift or gain error 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 or Analyser III, 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
can be modelled by a pole in the s plane, placed on the negative real axis at \(-2\pi(1.8)\), 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

\[
\frac{1}{(R_2+1/sC)(R_1+R_2+1/sC)}
\]

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

\[
\frac{s+1/R_2C}{s+1/(R_1+R_2)C}
\]

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.
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 R2 and C2 for one pair of poles, and R3 and C4 for the other pair. Compensation can be carried out in exactly the same way as before; by changing the values of R2 and R6, and by introducing resistors R1 and R5 in series with capacitors C3 and C4.

This filter was simulated using the Analyser III electronic circuit analysis program. The uncompensated filter showed peaking in the frequency response, just below the cut-off frequency. The values of the compensating resistors, R1 and R6, were then calculated and added into the netlist. Resistors R2 and R5 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 frequency response of both compensated and uncompensated filter 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 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 shown 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.

References
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2. Those Engineers Ltd, ECA2.
3. Number One Systems Ltd, Analyser III.
5. P. Lynn, An Introduction to the Analysis and Processing of Signals.
A-to-d and d-to-a converters

Fast, low-power a-to-d. TDA8760 is a 10-bit analogue-to-digital converter, claimed by Philips to be the first commercial a-to-d to sample at 50Msample/s while dissipating only 850mW from 5V. It will digitise analogue signals with component up to 20MHz and, when sampling a 4.43MHz full-scale input at 40Msample/s, signal-to-noise ratio is better than 56dB with a THD better than <−65dB. Since input capacitance is only 4.5pF, no input buffer is needed. The tri-compatible outputs are tri-state, and can be programmed to give two-tot-complement coding. Philips Semiconductors (Eindhoven). Tel., 01034 40 722091; fax, 01034 40 724825.

Discrete active devices

High-power mosfets. Motorola’s high-voltage TO-246 family of n-channel TMOS power mosfets dissipate 3000W internally and control loads up to 8kW, 500V, 600V and 1000V devices now being available. The higher power reduces the need to parallel devices for high current and low Rdson avoids the need for series gate resistors. On resistance for the 1kV version, M2Y10N06E, is 1300mΩ at 10A, that of the 600V M2Y25N050E 210mΩ at 25A and of the M2Y20M08E 240mΩ at 20A. Motorola Inc. Tel., 0908 614614; fax, 0908 616850.

Digital signal processors

PAL/NTSC genlock. Raytheon’s RS7000 horizontal line driver chip is meant for digital video signal processing and conforming, 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 CCIIR651C, -54C, square pixels and VGA in both formats. Ambiar Components Ltd. Tel., 0844 261144; fax, 0844 261789.

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ACTIVE

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 65Ω. 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 40V, 250mA and exhibiting an on resistance of 11Ω for a gate/source voltage of −3.5V. The range of gate/source voltage is 88V 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 1012Ω, 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 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 MRF860 the 36W device showing a 1.5pF. This device is unconditionally stable and need no external components. Both types slewing at 1000V/ps with 1.5mV common-mode range. Input are tri-state, and can be programmed to give two-tot-complement coding.

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 47mV 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 81411036; fax, 01049 81413515.

High-speed op-amps. True voltage-feedback op-amps from Linear, LT1361/2 and LT1364/5 are dual/quadruple types offering high speed while maintaining good dc accuracy. LT1361/2 have 50MHz gain/bandwidth and slew at 800μV with a 1mV input offset and 1μA input bias, while LT1364/5 are 700kHz types slewing at 1000μV with 1.5mA 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 3.3V 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 LT1030/2/4/5/6 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%.

Smalllest DPM. Using a thick-film hybrid microcircuit manufactured by CorinTech, the Lascar DPM1 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 655635; fax, 0425 652756.

Current limiting and thermal shutdown are provided. Semtech Ltd. Tel., 0592 773520; fax, 0592 774731.

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/µs 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 modern microprocessor, this converter's new control generates a 40-650V dc supply and provides three outputs: 5V, 12V and 24V. All output voltages are nominal in all conditions of line, load, temperature or cross-regulation and are overcurrent-protected. Ripple and noise are less than 50mV and 100mV for 5V and 12V outputs. Gardners Ltd. Tel., 0202 482384; fax, 0202 470685.

Logic building blocks
16-bit ALU. Logic Device's L4C831 arithmetic and logic unit performs 16-bit addition, subtraction and logic operations including AND, OR and XOR. Bit addition, subtraction and logic arithmetic and logic unit performs 16-bit operation on the data available at one time. All outputs stay within 5% of the appropriate pin adaptor. Features include for programmable partition and transistor detector is included. The pattern allows inspection of the workpiece. Optics by Lasiris Information. Control Transducers Ltd., 0202 482384; fax, 0202 471680.

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 car components to car bodies, for alignment or edge detection. Gaussian distribution along the length of the strips produced is eliminated. Laser Lines Ltd., Tel., 0202 482384; fax, 0202 471680.

Optical sensors. Isocom Components has a range of slotted interrupter and reflective sensors used for infrared data transmission in single or dual packages, the dual being intended for direction sensing. Isocom Components can be configured as a bump, kick, flyback, forward, isolated and non-isolated type, using a single-ended switch. Switch output, band-limited reference, voltage regulator, error amplifier, 4kHz 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. Clero Electronics Ltd. Tel., 035 295747; fax, 035 297717.

This is a text-based document containing information about electronic products and technologies. It appears to be a section from a magazine or a newsletter, discussing various products and their specifications. The text is formatted in a typical magazine layout with paragraphs, headings, and sometimes bullet points. There are also some tables and figures, although they are not transcribed in this text. The document seems to be aimed at an audience interested in electronics and technology, possibly engineers, technicians, or hobbyists in the field. The content covers a variety of topics, including datacoms, logic units, optical devices, and other electronic components. The text is written in a professional tone and uses technical terminology. It includes contact information for companies and further resources for those interested in learning more about the products discussed.

The document is structured in a way that is typical for technical publications, with clear sections and subsections. Each product or technology is described with specific details, such as technical specifications, performance metrics, and contact information for companies that produce or distribute these products. This layout is designed to provide a comprehensive overview of the products and technologies, allowing readers to make informed decisions about purchasing or using them.

In conclusion, this text is an informative resource for those involved in the field of electronics, providing detailed information about various products and their applications. The use of technical language and the inclusion of contact details make it a valuable reference for professionals and enthusiasts alike.
Chip capacitors. Kyocera has the 0402 surface-mounted chip capacitor in the ML range, now available in NPO, SL, X7R and Y5V dielectrics. In NPO, the range is 8.5-120pF 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 500MQ minimum. AVX Ltd. Tel., 0252 336866; fax, 0252 346643.

Connectors and cabling

Filtered BNCs. Two ceramic chip capacitors in Genalog's BNC connectors provide filtering to reduce em/rfi leakage. Contacts are gold-plated phosphor bronze and the inner insulator is polypropylene. The glass-fibre outer has a UL94V-1 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 lcid. A single edge light on NEC's new 9.4in monochrome lcid 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 lcid 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/m2. 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.

Monochrome LCD. The Orion OEM-6445 is a monochrome 5.5in 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-Heinagol Electronics Ltd. Tel., 01049 40 23760; fax, 01049 40 23762510.

NEW PRODUCTS CLASSIFIED

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Cable-tv signal-level meter. A 48-861MHz signal-level meter by Alban Electronics, the Promax MC-500 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 lcid. 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 DF101, uses a high-pass filter with three spot frequencies at 315kHz, 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.

Communicative counter. A 1.3GHz counter-timer by Thurlby-Thandar, the TF830-ARC is provided with an RS232 serial interface for control and to send the output to a printer or computer, measuring ratios of video-to-audio, and carrier-to-noise and carrier-to-Nicam, results being presented on a backlit lcid. 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.

Sealed switch. The Grayhill series 30 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 125A at 220V ac, contact resistance 250mΩ and insulation resistance 1GΩ. EAO-Highland Electronics Ltd. Tel., 0444 236300; fax, 0444 236641.

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/ octave 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 PC/MIA 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. Burrel & Kjaer (UK) Ltd. Tel., 081 954 2366; fax, 081 954 9554.

Crystal Technology. Tel., 0635 528520; fax, 0635 528443.

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-48/U footprint. They are in a metal can and cover the frequency range 3.2-500MHz in preferred values or in specified frequencies to order. Stability is ±100ppm/year. Temperature range is -40°C to 125°C. Advanced Crystal Technology. Tel., 0635 258250; fax, 0635 258443.

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 ±77.5kHz. 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.

Hardware

Custom boxes. If one’s interest as a design engineer stops dead when the circuit performs as intended, reality is allowed to make his own with 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.

Dos and the other to run minimised versions of a PC. This instrument is a Communicative counter. A 1.3GHz counter-timer by Thurlby-Thandar, the TF830-ARC is provided with an RS232 serial interface for control and to send the output to a printer or computer, measuring ratios of video-to-audio, and carrier-to-noise and carrier-to-Nicam, results being presented on a backlit lcid. 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

LD178, uses a high-pass filter with three spot frequencies at 315kHz, 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 TSP to run under Dos and the other to run minimised under Windows. Amdal Tel., 0272 699352; fax, 0272 236088.
Transducers and sensors

**Temperature sensors.** Elmocontrol’s range of temperature sensors are flexible, wire-wrap 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 0.044m. They are provided with pressure-sensitive adhesive backing and cover the range -20°C to 235°C. Radiatron Components Ltd. Tel., 01784 335033; fax, 01794 477333.

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

**Accelerometer.** Ti has a new acceleration sensor for the ±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 ccid camera. Sony’s XC-777P is the company’s smallest and lightest colour camera, being based on the 440pixe1/3in imager, all associated circuits being contained in the one housing, which measures 22 by 22 by 89mm, weighing 75g. Resolution is 480TVL and minimum sensitivity 4.9lux. 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., 0132 816000; fax, 0132 871001.

PCMCIA video camera. PC Card Camera by VVL is claimed to be the firstly 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. VVL Vision Ltd. Tel., 031- 539 7111; fax, 031-539 7140.

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 0.8lux. 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 board-level products

100MHz backplane cpu card. The Blue Inferno single-board computer by HM Systems uses IBM’s 60MHz and 100MHz processors. The board takes up to 62MB 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. HM Systems plc. Tel., 081-209 0911; fax., 081-209 0912.

Two-Pentium motherboard. SPC has the ASUS10K PC16-EPSNP4, 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 EISA, can be fitted up to 150Mbytes of memory 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 Dynapar Ergo Touch 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. TLI-86 contains the TDB80 development board based on the TMP90C141N controller, terminal software, a monitor and a full-feature assember, the kit being connected to a PC by a printer cable. The TDB80 board has 1K0 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 TMP91P40N-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-HP/700-EISA interface kit which features 5.5Mb/s data transfer rate for both read and write using HS488 and 1.3Mb/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 523514.

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 RD051 drives the whole stack of ports simultaneously at up to 115.2Kbs. IDT's 3051 family includes four devices with on-chip cache sizes of 2.5, 5, 10 and 20K-byte and frequencies of 16-40MHz. Integrated Device Technology, Tel., 0372 363734; fax, 0372 378651. 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 3.3V 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 370343.

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 POWERware. The package runs on 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.

For more details, contact POWERware, 14 Ley Lane, Marple Bridge, Stockport, SK6 5DD, UK.

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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 rays to natural ionising radiation. The articles on audio amplifier design by Doug Self have been fascinating, and easy to read. Particularly interesting for me was the article 'Linear thinking' by Doug Eleveld, Groningen, which discusses the use of op-amps in audio circuits. Self has been intent on saying no more on the subject until I had built my own prototype version advancing on the three prototypes built by JS 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-electrical 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 thermo-electric 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 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 thermo-electric 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, 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

November 1994 ELECTRONICS WORLD + WIRELESS WORLD 949
acoustic oscillations in a pvdf stack of nickel-aluminum 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 Z80 if they were prepared to learn Z80 machine language - and Zargon if they were prepared to do some assembly. Any ZX Spectrum owner could create their own keyboard and a display - Z80-based microsystems. I doubt if many ZX Spectrum owners knew, didn’t he! Remember the ZX!

Southampton

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 concern, which is what I was accused of. I never said it didn’t “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 - be taken to task by callers and correspondents for being an expression of opinion, 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 intermediate stations he invented on the basis of Brantly’s scientific experiments. Readers looking for more information should see our recently published book (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 aversion, to be taken to task by callers and correspondents for being an expression of opinion, 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 intermediate stations he invented on the basis of Brantly’s scientific experiments. Readers looking for more information should see our recently published book (Lodge and the Invention of Radio, eds P Rowlands and J P Wilson, PD Publications, Liverpool, 1994).

Barrie Blake-Coleman
Salisbury

References
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CIRCLE NO. 136 ON REPLY CARD
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:

\[
P_a = V_{\text{rms}} I_{\text{rms}}
\]

where \( V \) is in volts and \( I \) is in amperes. The volt-amperes 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.

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:

\[
P = P_a \cos \theta = V_{\text{rms}} I_{\text{rms}} \cos \theta,
\]

where \( V \) is in volts, \( I \) is in amperes and \( \theta \) 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, \( \theta \), 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:

\[
P_r = P_a \sin \theta = V_{\text{rms}} I_{\text{rms}} \sin \theta,
\]

where \( V \) is in volts, \( I \) is in amperes and \( \theta \) is in degrees.

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 VImax. 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 V2/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.

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

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 AN15/18 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 environments/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.
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 $R_2$. 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:

$$\Delta V = \frac{AV}{C}$$

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 \(\Delta V\) 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 \(LM331\) 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 20\(\mu\)s 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 40\(\mu\)s. Consequently, the microcontroller cannot create a pulse train with a positive pulse width of less than 40\(\mu\)s. Also, the software that measures the pwm output pulse width at the input capture pin requires approximately 20\(\mu\)s 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 20\(\mu\)s + 40\(\mu\)s, or 60\(\mu\)s. 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, 512\(\mu\)s 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 0.5\(\mu\)s per clock tick.

There are four clock cycles per timer count. This results in 2\(\mu\)s 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 2\(\times\)256. But since an additional 60\(\mu\)s is needed at both pressure extremes of the output waveform, the total period must be at least 632\(\mu\)s. 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 capacitor value 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 84\(\mu\)s 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 60\(\mu\)s 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 60\(\mu\)s.

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

**LC oscillator features 1% distortion**

A t the heart of many oscillators is a parallel-resonant \(LC\) tank circuit whose impedance is infinite at the resonant frequency of \(1/(2\pi FC)\)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.
APPLICATIONS

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 2.5Ω 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: \( gm = \frac{8}{R_i} \), where the factor of eight is inherent in the IC.

Negative resistance is therefore \( \frac{R_1}{8} \), which must be less than \( R_3 \). Choosing \( R_3 = 4752 \) yields \( R_i < \frac{8R_3}{2} = 37.652 \). A reasonable value for \( R_1 \), therefore, is 30Ω.

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 PC/XT/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/sdn 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.
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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...

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-
The EMF 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 EMF 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 burglary 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 0.08J. Furthermore, promising to fulfill Churchill's prediction exactly, we find that various kinds of explosive detonators are fired off by an EMP of between 0.02 and 0.6J.

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/cm², 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/in², 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
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.
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

$$vswr = \frac{1 + |\gamma|}{1 - |\gamma|}$$

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
RF TRANSISTORS

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 $T_{opt}$. Sometimes four basic noise parameters are referred to, because the quantity $T_{opt}$ 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:

$$NF = NF_{min} + 4\frac{r_n}{(1 - |\Gamma_n|)^2}$$

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 — $\Gamma_n$. If the value of $NF$ is specified, the locus of points representing possible values of $\Gamma_n$ 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 $T_{opt}$. The centres of all the noise circles will lie along the $T_{opt}$ vector which originates at the point of the Smith chart and terminates at the location of $T_{opt}$.

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

$$C_n = \frac{T_{opt}}{1 + Ni}$$

where $Ni$ is a noise figure parameter defined by:

$$Ni = \frac{NF - NF_{min}}{4r_n} + |\Gamma_{op}|$$

and $NF_1$ is the desired noise figure circle. Likewise, the radii of the circles are given by:

$$R_n = \frac{1}{1 + Ni} (N^2 + Ni(1 - |\Gamma_{op}|)^2)^{0.5}$$

Optimum source reflection coefficient ($T_{opt}$), 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 of 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 $f_t$ 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 $f_t$ 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 — $\Delta V_{BE}$ 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 $\Delta I_C = \Delta V_{BE} I_C^2 / V_E$, where $\Delta I_C$ is the change in $I_C$, $\Delta V_{BE}$ 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 $\beta$ the least control of any major dc specification. It is not uncommon to have a bipolar transistor with a range of $\beta$ that exceeds 5 or 6 to 1. That is to say, the ratio of guaranteed maximum $\beta$ to minimum $\beta$ 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 $\beta$ can be approximated by:

$$\Delta I_C = I_C \frac{\Delta \beta}{\beta} \left( 1 + \frac{R_E}{R_E} \right)$$

where $I_C$ is the collector current at $\beta = \beta_1$, $\beta_1$ is the lowest value of $\beta$, $\beta_2$ is the highest value of $\beta$, $\Delta \beta$ is $\beta_2 - \beta_1$, $R_E$ 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, a clear 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
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:

$$G_t = \frac{|S_{21}|^2 \left[1 - |S_{12}|^2 \right]}{\left[1 - S_{11}\right]}$$

which can be manipulated to give

$$G_t = \frac{1 - |S_{21}|^2}{\left[1 - S_{11}\right]}$$

where,

$$\Gamma_m = S_{11} + \frac{S_{21} S_{12} \Gamma_L}{1 - S_{22}}$$

or,

$$G_t = \frac{1 - |S_{21}|^2}{\left[1 - S_{11}\right]}$$

where,

$$\Gamma_{OUT} = S_{22} + \frac{S_{21} S_{12} \Gamma_L}{1 - S_{11}}$$

Comparing the two equations for $G_t$, the first relates it to input term $\left[1 - |S_{21}|^2 \right]$, $\Gamma_m$ term $S_{21}$ and output term $\left[1 - |S_{12}|^2 \right]$, $\left[1 - S_{22}\right]$ $\Gamma_L$, 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 $\Gamma_L$ is made equal to the conjugate of the transistor input reflection coefficient $\Gamma_m$ – 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 $\Gamma_m$ was forced to equal $S_{11}^*$. Operating power gain is:

$$G_p = \frac{1 - |S_{21}|^2}{\left[1 - S_{11}\right]}$$

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 impedance because $\Gamma_m$ was forced to equal $S_{11}^*$. Operating power gain is:

$$G_p = \frac{1 - |S_{21}|^2}{\left[1 - S_{11}\right]}$$

where $\Gamma_m = S_{11} + \frac{S_{21} S_{12}}{1 - S_{22}} \Gamma_L$

or,

$$G_p = \frac{1 - |S_{21}|^2}{\left[1 - S_{11}\right]}$$

with the minimum and maximum errors being calculated as $-0.35$dB and $+0.37$dB.

Frequently, the errors are less than $0.25$dB and, as such, are sufficiently small to justify using $G_m$.

Returning to the expression for $G_t$ and assuming $S_{12} = 0$, the equation becomes:

$$G_m = \frac{1 - |S_{21}|^2}{\left[1 - S_{22} \Gamma_L\right]}$$

and can be broken into three sources of gain:

$$G_t = |S_{21}|^2$$

For broad-band, with a certain amount of maximum gain is wanted, all that is required is to set $\Gamma_m = S_{11}$ and $\Gamma_L = S_{22}$

To obtain an exact solution ($S_{12} = 0$), we could use the above equation for $G_m$ 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.

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:

$$U = \frac{|S_{11}| |S_{21}| |S_{12}| |S_{22}|}{1 - |S_{11}|^2 |S_{22}|^2}$$

Defining $G_m$ as the transistor power gain with $S_{12} = 0$ and $G_t$ as the actual transistor power gain, the maximum error introduced by using $G_m$ instead of $G_t$ is given by:

$$\frac{1}{1 + U} \leq \frac{G_t}{G_m} \leq \frac{1}{1 - U}$$

To illustrate the use of this equation, take the MRF571 at 1GHz and a bias condition of 6V and 50mA. From the data sheet $|S_{11}| = 0.60$, $|S_{12}| = 0.09$, $|S_{22}| = 4.4 - 1S_{22}$ and $S_{22} = 0.11$. So:

$$U = \frac{(0.60)(0.09)(4.4)(0.11)}{(1 - (0.60)^2)(1 - (0.09)^2)} = 0.0261$$

or

$$U = \frac{(0.64)(0.988)}{0.0413}$$

Figure 3. Typical input gain circles.
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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):

$$d_i = \frac{S_i|S_{11}|}{1 - |S_{11}|^2 (1 - g_s)}$$

$$R_i = \frac{(1 - g_s)^2 (1 - |S_{11}|^2)}{1 - |S_{11}|^2}$$

where $g_s = \frac{G_i}{G_{i,\text{max}}}$

and

$$G_i = \frac{1 - |R_i|}{1 - |R_i| |S_{11}|}$$

$G_i$ is the gain represented by the circle, $d_i$ is the distance from the centre of the Smith chart to the centre of the constant gain circle along the vector $S_{11}^*$, $R_i$ is the radius of the circle and $g_s$ is the normalised gain value for the gain circle $G_i$.

Likewise, for the output network:

$$d_L = \frac{g_L|S_{22}|}{1 - |S_{22}|^2 (1 - g_{L})}$$

$$R_L = \frac{(1 - g_{L})^2 (1 - |S_{22}|^2)}{1 - |S_{22}|^2}$$

$$G_L = \frac{G_L}{G_{L,\text{max}}}$$

$$G_L = \frac{1 - |R_L|}{1 - |R_L| |S_{22}|}$$

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 $G_i$, 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.

RF Transistors: Principles and practical applications is available by postal application to room 1333 EW+WW, Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS. Cheques made payable to Reed Books Services. Credit card orders accepted by phone (081 652 3614). 288pp Hardback 07506 9059 3 Cost £19.95 + Postage £2.50

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