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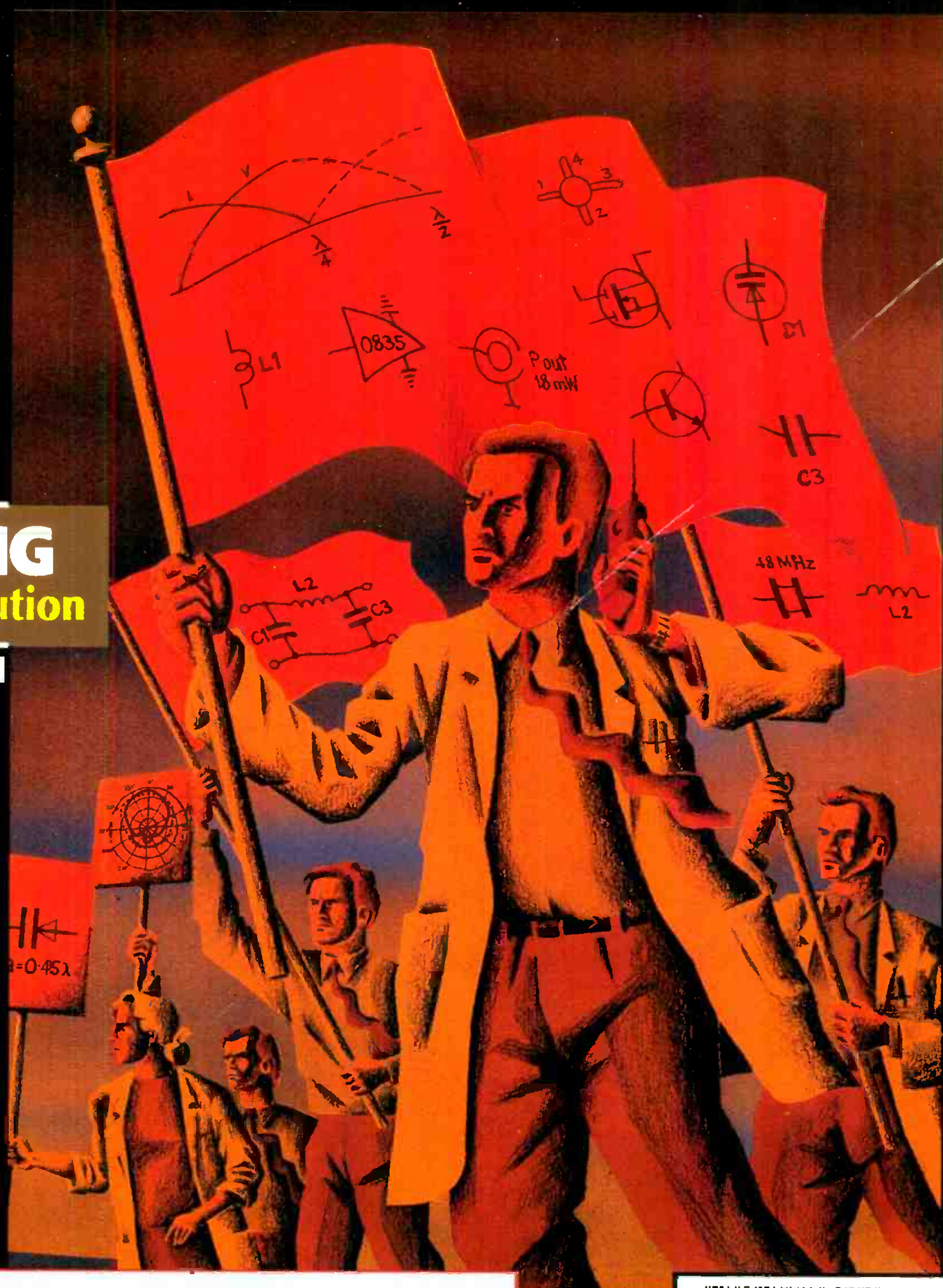

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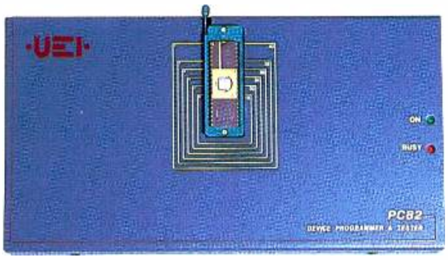

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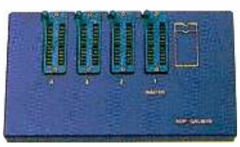

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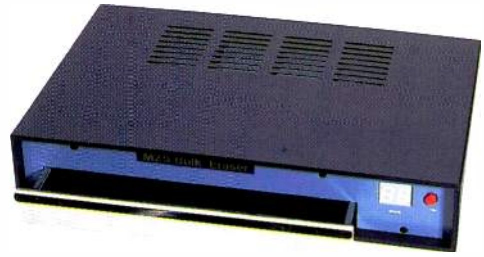

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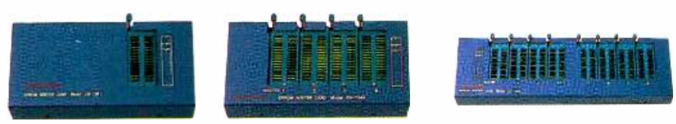

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REGULARS

In next month's issue: Field programmable gate arrays will become as important to design engineers as TTL logic did 20 years before. FPGA provides a flexible and reprogrammable alternative to TTL. Learn about it here.

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A degree in hair styling and other cultural studies

Did you watch the science documentary series *Life on Earth*? Of course you did. Make the most of it because no science based documentaries are likely to made on this scale again.

Not my warning but that of Sir David Attenborough, delivering the presidential address at this year's meeting of the British Association for the Advancement of Science. Sir David warned that ITV early evening schedules are to be regeared for mass audience with a single bottom line of viewing figures: "No programme will be shown in the evening before 10.30 pm unless it can command a minimum guaranteed audience of at least 8 million.

"That means that from next year onwards, no serious science programmes will be shown on independent television at times that most people are able to see them" said Sir David.

While I would be the first to agree that television (and electronics magazines) should, in general, be entertaining rather than overtly educational, the demotion of science from peak viewing times will do nothing to help the cultural role that science and technology should fulfil in an advanced society.

And it is not just ITV which denies science a place in the living room. BBC current affairs output is highly regarded. But does it deserve this regard? If the Pound loses half a Pfennig against the Deutschemark, programme producers will wheel in an industrial editor to comment on the effects of a possible interest rate rise on 'industry', a financial editor to comment on stock market reaction, a political editor to comment on the (usually) predictable mutterings from government and opposition benches, a foreign affairs editor to comment on the EC perspective and, finally, a home affairs editor interviewing the chairman of a building society.

Compare this with media coverage of IBM's announcement that it had developed a scanning, tunnelling electron microscope, at its Zurich research facility. Consider the manner in which the BBC reported news of an instrument that gives us direct 3-D sight of atoms in a lattice, a development so important that it will permanently change our view of materials for solid-state physics and biochemistry.

I don't think that there was any.

We now live in a country built from, and surrounded by, science and with little culture of its understanding. Indeed, there is almost an anti-culture. How many times have you heard the Great and the Good publicly boasting of their incompetence with household technology?

But there are worse things than being governed by terminally stupid people. Ask sixth form pupils about their career ambitions. Top of the list will be law, medicine, media, public relations and accountancy. Middle ranking interests include business and financial services. Chemistry, physics and maths are seldom the preferred choice.

There is a real possibility that physics will disappear from the UK entirely. University science places, particularly those in physics, are available with bargain basement A-level grades; funding rules are now strictly tied to capitation which means that university science departments are having to maintain their facilities on a shrinking budget while arts departments prosper.

The importance of science to society must be recognised. It seems unlikely that we shall be able to maintain our living standards based on an education system which produces only lawyers and hairdressers.

Frank Ogden.

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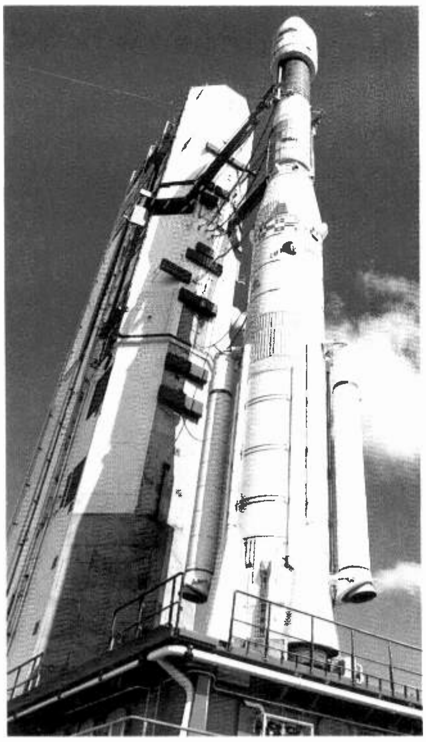
REGULARS

UPDATE



The hopes of Surrey University riding high: Ariane 52 carried two 50kg microsats built by SSTL and intended for mobile radio comms experiments. The £1 million programme took just 12 months to complete.

Engineers shift their footing

A fundamental restructuring of the engineering profession has been proposed. It represents the latest thinking of the Steering Group set up at the beginning of this year under the chairmanship of Sir John Fairclough, Chairman of the Engineering Council. Its thoughts were circulated this week to the engineering institutions, The Engineering Council and the Royal Academy of Engineering in a consultation paper.

The document considers the formation, role and organisation of a new single body to act as a focal point for the engineering profession, including Chartered Engineers, Incorporated Engineers and Engineering Technicians. The steering group has considered the relationship between a new single body and the institutions, selection of members for the new body, the registration of engineers and the relative merits of a statutory or non-statutory system.

The new single body will subsume The Engineering Council either by reforming it or replacing it, so forming a new relationship with the engineering institutions.

An alternative to one single body would be formation of a collegiate structure in which groups of institutions form intermediate bodies between the individual institutions and the central organisation. While this would support effective decision-making by focusing responsibility for sectoral issues directly onto the appropriate organisations it would introduce an extra layer of bureaucracy into the current two-tier system of Council and institutions.

Responses to the consultation document are now being sought and must be submitted by November 10, 1992. They will help the Steering Group prepare an interim report due to be submitted to the Council of Presidents of the Institutions early next year.

Julia King.

Shades of Dan Dare and the Mekon.

A talking head that appears out of a desk is just one of the futuristic ideas in telecommunications being researched at BT Laboratories, Martlesham near Ipswich, for business life in the 21st century.

The head is part of the Future Desk which may replace the familiar constructions of wood and metal with screens and surfaces capable of achieving 3D videoconferencing through holographic projections of a human being.

The Future Desk is just one of the concepts being pioneered by a group of technical and scientific experts in the systems research division who are working as a think tank for combinations of telecommunications technology that can be applied to the future.

Transistors to replace magnetrons?

Philips Semiconductors has developed a microwave power transistor capable of producing a 750W output pulse for radar transmitters. Designed for class-C operation, the MX1011BY700Y runs from a 50V collector rail and is characterised for a duty cycle of one per cent. It should reduce the output transistor count in airborne and vehicle radar systems operating between 1030 and 1090MHz. The associated gain is 6.3dB.

IBM to end PC manufacture?

US computer giant IBM is expected to announce plans to turn its $7bn PC and workstation systems manufacturing business into a separate subsidiary. Industry analysts say this will allow the computer maker to cut costs in its PC operation dramatically

US opens up for space telephones

US plans for satellite-based mobile telephone systems cleared a regulatory hurdle last month but there is still a lot of work to be done to demonstrate the technical feasibility of the low earth orbit satellite systems.

The US regulatory authority, the FCC, has granted licences to three companies for trials using the 1610-1626.5MHz and 2483.5-2500MHz bands agreed at WARC this year.

But one of the licence holders Motorola-backed Iridium, which plans to launch its first five satellites in 1996, has modified its system to reduce the total number of satellites needed for global coverage from 77 to 66. The other licence holders are Constellation and Ellinsat.

●Mobile telephone makers could face the prospect of conflicting standards for the next generation of pocket telephones if the US Government is forced to abandon plans for personal communications systems (PCS) in the 1.9GHz band.

A powerful microwave user group is currently lobbying Congress to argue the unsuitability of the 1850-1990MHz band for the new mass market PCS services. This band has been the front runner for the US system, in part due to its similarity to the DCS1800 system being proposed in Europe. Despite plans by the authorities to accommodate existing users of the band, pressure is growing to look to higher frequencies for the PCS services which will be introduced by 1995.

Richard Wilson, *Electronics Weekly*

Single chip GPS?

Single chip microwave receivers are about to slash the cost of high accuracy satellite based global positioning system technology.

Motorola has plans for a single chip GPS receiver next year.

According to a spokesman in the company's RF group the design of the downconverter chip will need at least two more iterations before it can be combined with an LNA into a single device.

The chip will take unidirectional or bidirectional signals from the GPS antenna and deliver a demodulated signal at baseband which can be processed to extract the positioning information. The accuracy of the system is defined by the signal generated by the GPS satellites. This is better than 100m for civil applications.

GEC Plessey Semiconductors is also expected to announce a GPS chipset two or three ICs before end of the year. Engineering samples of a two-chip GPS design are already being evaluated by UK companies. The two chips are a 1.575GHz low noise amplifier (LNA) and a dedicated GPS downconverter.

Current GPS receivers use discrete components and custom hybrids. They are about the size of a personal stereo and cost around $500. Single chip designs could drive prices below $100 which open up mass market applications in car navigation and electronic security systems. More than one European cellular telephone maker is investigating the incorporation of a GPS receiver in its handsets.

BiCMOS boost

IBM is developing a bicmos process which, it hopes, will double the performance of its current bicmos lines. The bipolar transistors in IBM's current lines have a cut-off frequency of less than 20GHz whereas the new process will allow frequencies up to 60GHz. Furthermore, the cmos section will be built on a 0.25µm process, instead of the current 0.5µm geometry.

An IBM spokesman said that this should double the switching speed of the cmos. He claimed that the new technology would be superior to gallium arsenide in all aspects including speed and density. First production will appear by the middle of the decade.

Optical radar provides night picture

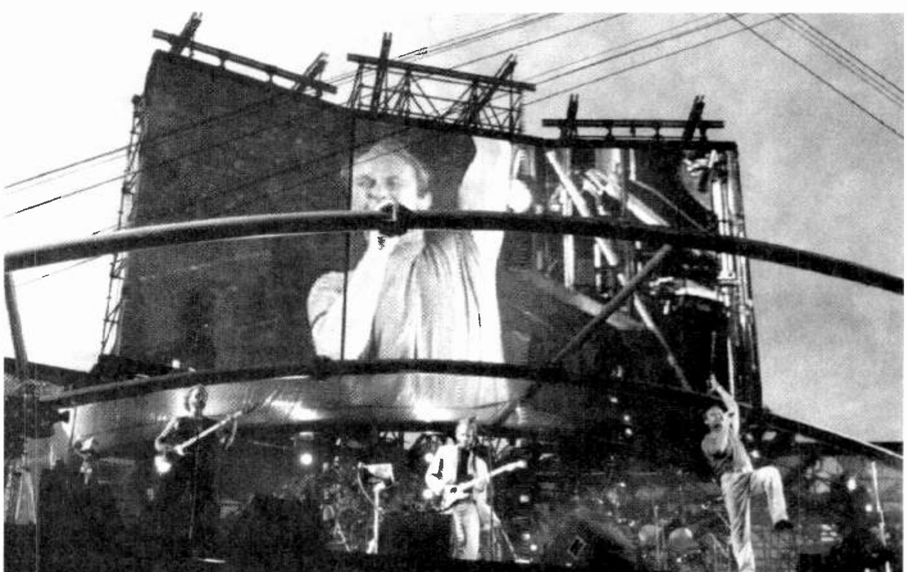
A laser radar that can spot power lines in the way of low-flying military helicopters is being developed as part of an Anglo-French research project.

British radar maker GEC-Ferranti Defence Systems and French firm Dassault Electronique are believed to be collaborating on the project, reportedly called Clara, to demonstrate a helicopter based optical-wavelength scanning radar.

Optical lasers have several advantages over infra red and millimetric systems. The most important is that laser light is easier to steer using conventional mirror optics which are also highly compact. A rapidly scanning pencil-beam laser radar system can provide television quality night vision systems for helicopter pilots; computerised image processing highlights hazards.

Laser radars use little power and allow fine beam, precision measurement at short ranges. Atmospheric absorption – basically clouds and fog – makes them of little use for long range air-to-air detection. But radar makers can exploit this to provide stealthy protection and prevent the enemy detecting the laser beam from afar.

The project will build on technology previously test flown by GEC Avionics.



A question of image: Rock Band Genesis on its latest US tour using three Sony Jumbotron screens, each measuring 5.6m high by 3m wide and weighing six tonnes, to ensure fans can both see and hear.

Dianagate raises questions of technology and law

The widely publicised telephone conversation alleged to have taken place between Princess Diana and an admirer has alerted the general public to the risks of saying anything private, personal or secret on any phone which relies on a radio link. It may also prompt the police and government to use existing laws to protect privacy.

Cellular phones are inherently insecure, as are most cordless phones that are used around the home and garden. For less than £200, anyone can buy scanners which allow them to listen to and record conversations. Eavesdropping on radio phone calls has now become a hobby.

None of the companies or government bodies involved – British Telecom (owner of Cellnet), Vodafone, Oftel, Scotland Yard, the Department of Trade and Industry, and Home Office – has taken responsibility for the problem. They argue that users of radio phones should always be aware of the inherent lack of security.

Both Cellnet and Vodafone now acknowledge that it is possible to eavesdrop on calls. "It is not illegal to buy or sell the equipment," says Cellnet, "and the law on use is unclear."

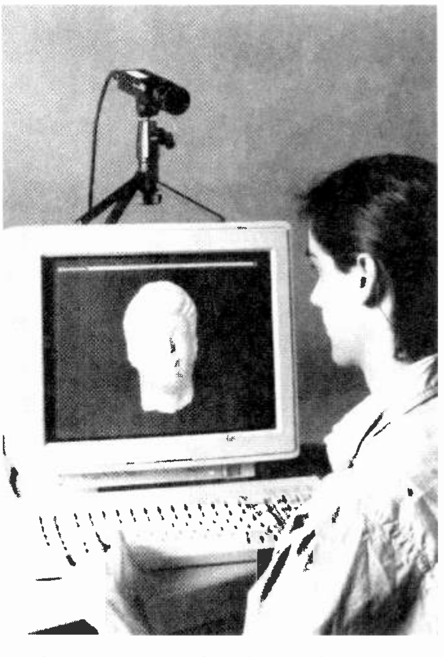
Under the Telecommunications Act, the Wireless Telegraphy Act and the Interception of Communications Act, electronic eavesdroppers can be jailed for up to two years. But the acts have seldom been used to curb such eavesdropping. The papers which have published transcripts of the "Diana" tapes, and made them available by telephone to inquisitive readers, have now pushed the law to new limits.

Britain's two cellular phone networks currently work with analogue frequency modulated speech signals at frequencies of around 900MHz. The inherently more secure digital GSM standard systems are not yet available, and neither are they likely to be for a couple of years.

The cellular network companies have played down the risk of eavesdropping with scanners, arguing that the two halves of a cellphone conversation, to and from the mobile phone, are carried on different radio frequencies. Also, these frequencies change as the caller moves between cells served by different transmitters.

In 1989 there was a spate of newspaper articles on eavesdropping, revealing what equipment is needed and how much it costs. An eavesdropper with a single-frequency receiver can hear both halves of a conversation because the cellphone call passes through the public telephone network, which provides a sidetone so that both callers, on a conventional or a cellular phone, can hear in their earpiece what they say into their mouthpiece.

The radio cells in an urban area may be only a few kilometres wide, so the frequency of a moving cellphone will switch every few minutes. Cells in rural areas are much larger because they need to cater for fewer simultaneous calls. So when a scanner locks onto a conversation it will hold it for much longer. **Barry Fox**, *New Scientist*.



Students at MIT's Media Lab are using a machine vision system to develop interactive interfaces between humans and computers. For example, the machine vision systems are used to track the position of a person's head in order to implement virtual holography – in which the computer can create a 3D view on the monitor based on the position of the person looking at it. The vision system will also be used to develop low bit-rate teleconferencing, and to replace the computer mouse by tracking a user's hand positions.

Intel processor gets speed boost

Intel is now delivering a 66 MHz 486 DX2 processor chip. Systems using it with a 33 MHz system clock can execute their applications, on average up to 70 percent faster than systems using a 33 MHz 486 DX CPU.

The 486 DX2 family operate at an internal frequency that is double the rest of the system. The company has already shipped some 300,000 50MHz DX2 chips.

The technology operates by doubling the external system clock signal it receives through the system bus and providing the internal 2x clock signal to on-chip sub units. As a result, blocks such as the CPU and cache can run at a 2x internal clock rate to execute instructions.

The CPU operates at maximum efficiency when incorporated with fast design elements such as second-level cache, burst memory controllers and write buffers. For example, second-level cache helps eliminate wait states by providing the CPU's on-chip cache with a high-speed pool of data. If an application or operating system causes the microprocessor to access off-chip data, the second-level cache can supply the data immediately rather than forcing the chip to wait for slower system memory to provide the data.

REGULARS

RESEARCH NOTES

Taking a cold look at gravity measurement

Physicists at Stanford University in California claim to have measured the gravitational constant with unprecedented accuracy using "fountains" of atoms chilled to within a few millionths of a degree absolute. The procedure is complex, but the principle is simple: if you cool an atom, then by definition you reduce its random motion. That in turn allows more time to observe its critical parameters and make accurate measurements of its behaviour. Since almost all fundamental physical quantities, such as frequency and time, are based on atomic measurements, a cool atom is an accurate atom.

Most of us, when we want to cool something down, immediately think in terms of extracting energy using a freezer or some other sort of heat pump. But when physicists want to cool something to within a fraction of a degree above absolute zero, different techniques are called for. They mostly now make use of intersecting laser beams to create what is irreverently called "optical molasses" or "optical goo" – a state where

Continued over page

Op amp that tests itself

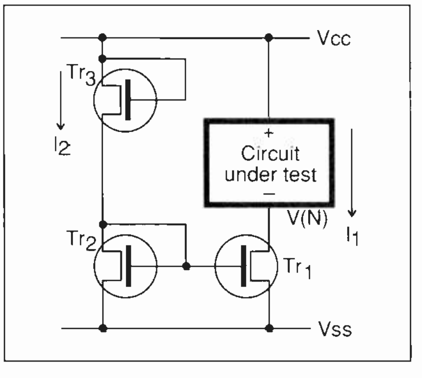
A novel method, just revealed, of testing cmos op-amps could eventually lead to completely self-testing chips.

The testing of all-digital circuits, although complex, is now a well established art. But a lot of research activity is now turning to the testing of mixed analogue and digital systems in which a simple go/no-go approach is inadequate or misleading.

In a paper recently published, M Roca and A Rubio *[Electronics Letters, Vol. 28 No. 15]* describe a testing technique based on the fact that the majority of open circuit or bridge faults in these chips result in some change of current consumption, I_{DD}. This is true even if the circuit continues to maintain some semblance of normal functionality.

Using computer simulation on a relatively simple cmos op-amp, Roca and Rubio modelled the effects of every likely open or short-circuit fault that would occur during manufacture. They used notional 10Ω resistors to model a fault involving bridged conductors and treated open-circuits in terms of transistors with a break in the drain-source circuit.

From this simulation it was found that bridged circuits nearly always produce a significant increase in total current consumption, while open-circuit elements nearly always lead to a decrease. Interestingly enough, Roca and Rubio found that a number of bridges, while increasing the power consumption of the chip, did not affect its functionality when working as a linear voltage follower. Such defects would not therefore be spotted with standard functional tests though they might well lead to subsequent breakdown through overheating.



This self-testing op amp uses a current comparator to provide a sensitive check on the consumption of the circuit under test. If the chip consumes more than its rated current, the voltage at V(N) rises sharply, indicating a fault.

No claim is made that I_{DD} testing will spot every fault; it would not, for example, spot an open-circuit input or output! But it is much simpler to implement than full function testing. The Spanish researchers suggest that it would not cost much and add very little to the overall consumption to build I_{DD} sensors directly into chips at the design stage, thus creating a sort of self-test capability.

The circuit they present uses a current comparator to provide a sensitive check on the consumption of the chip proper. Under normal operating conditions, transistor T_1 is near saturation, so the voltage drop between the circuit –ve and $V(N)$ is negligible. Should the chip consume more than its rated current, the voltage at $V(N)$ rises sharply, thus indicating a fault. A reduction would also be indicated, though with less sensitivity.

Roca and Rubio say that their standard cmos op-amp incorporating this self-test facility would reliably reject itself with any one of ten open circuits or 22 bridges.

Taking this strategy a stage further, why not build in an analyser that measures the precise I_{DD}, compares it with a look-up table of likely faults and then triggers a speech synthesizer? Just imagine switching on some piece of equipment with a faulty mixed-signal LSI chip only to hear it announce: "We apologise for the late arrival of your pulses. This is due to operating difficulties at the gate of *Tr99*".

OSCILLOSCOPES

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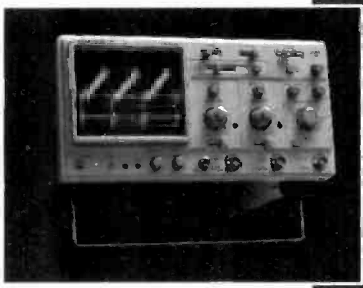

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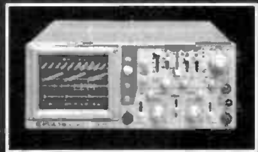

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photons can absorb energy from atoms and trap them in what amounts to suspended animation.

Optical goo takes advantage of the fact that atoms absorb certain frequencies of light much better than other frequencies. A laser beam tuned to a frequency either side of those critical frequencies will not disturb the majority of atoms. But if an atom happens to be moving towards the laser source the Doppler shift raises the laser frequency slightly, making the atom more likely to absorb a photon of energy, which then slows down the atom. By directing six lasers into a small space, the Stanford group can slow down the random movement of atoms in any direction, thus cooling all the atoms present.

Stephen Chu, one of the team, says that if you drop an atom into the optical goo, within a few milliseconds it comes into thermal equilibrium and for all practical purposes stands still. When he first did that in 1985, the effective temperature of the atoms was a few hundred milliKelvin. Later experiments showed that it is possible to cool atoms to within a few microKelvin, that is as near to absolute zero as makes no difference.

The Stanford group have not only trapped sodium atoms at almost zero K, they have also manipulated them in space by adjusting the frequency of some of the laser beams. In one experiment Mark Kasevich lifted the atoms upward and was able to observe them in free fall, a sort of atomic fountain that lasted a whole second. Atoms at room temperature normally move at supersonic speeds, so to be able to watch them drifting around at about 2m/s is what Kasevich describes as a "new regime in atomic physics".

An atom chilled to near absolute zero accelerates under the influence of gravity and so the frequency of its energy levels gradually changes. By measuring this change for a whole second, Kasevich has been able to measure the force of gravity to within three parts in 10^8. He and Chu expect to improve that precision to around one part in 10^{10} in the near future.

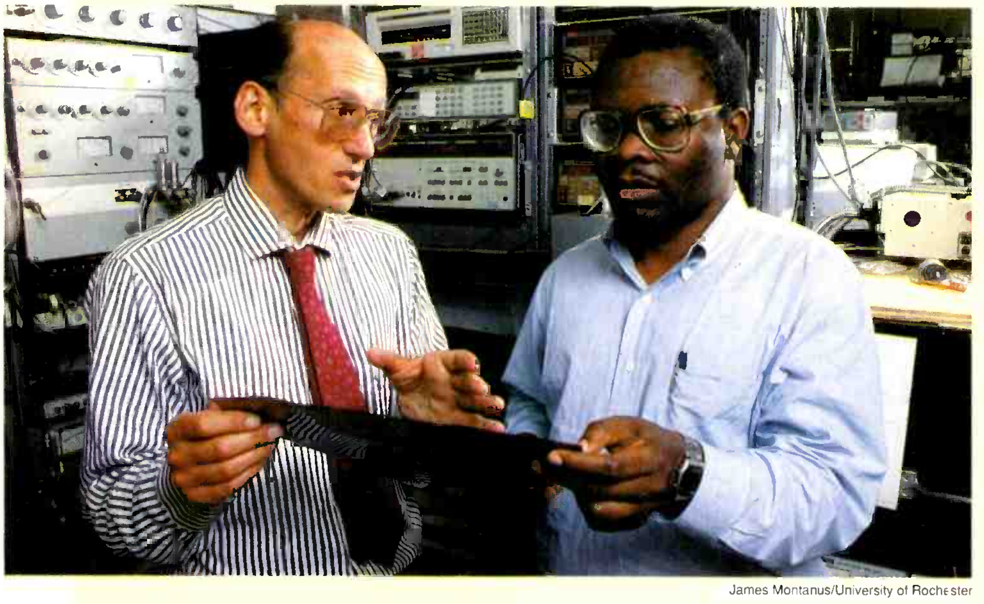
A portable gravity meter with the latter accuracy could measure changes in distance from the centre of the Earth to within a third of a millimetre; it could also detect underground anomalies such as water or mineral deposits with unprecedented ease.

Stable photo-conductive polymers

Materials developed at the University of Rochester, NY, and by a team at the Xerox Corporation could have implications for imaging technology. The new materials have formed the basis of the first patent application filed from the US National Science Foundation's Center for Photoinduced Charge Transfer at the university.

Scientist Martin Abkowitz at Xerox has evaluated thin films of polymer, only a fraction of the thickness of a human hair, fabricated by Samson Jenekhe, associate professor of chemical engineering at the University of Rochester. The materials were found to be sensitive to the red light emitted by the solid-state lasers widely used in printers and in some electronic copiers. Any material that responds efficiently to light from such lasers is of particular interest for these imaging applications.

Experiments have shown that the polymers are robust and remain intact and stable up to hundreds of degrees, much higher than most photosensitive materials used in copiers and printers. Previously fabricated high-temperature materials have been very difficult to process into useful devices or thick films, but the latest ones are much easier to work with while still retaining their robustness.



James Montanus/University of Rochester

Xerox (Martin Abkowitz, left) and University of Rochester (Samson Jenekhe, right) laying the ground-work for new imaging technology.

The new polymers are example of photoconductors, materials that act as electrical insulators when kept in darkness, but which become conductors when they are exposed to light. Photocopiers, laser printers, solar cells and even the human eye are based on such photoresponsive materials. In the field of imaging, photoconductors are used to create latent images in the form of electrical charge patterns.

Such a latent image is made by applying an electrical charge across the surface of a photoconductive material and then selectively discharging it with light focused by a lens or from a laser. The latent image is made visible when particles of toner – dry ink – are applied to the surface of the photoconductive material and then adhere to the charge pattern. This now-developed image is then transferred to paper and fused into the paper to make a permanent copy.

The polymers the University of Rochester and Xerox scientists are working on are double-stranded with molecular links between the strands. They are known as ladder polymers and include polybenzimidazolebenzophenanthrolines and polyquinolines.

Soldering techniques all in a flux

Scientists at Sandia National Laboratories in Albuquerque, NM are developing environmentally-friendly soldering techniques for producing electronic products without using ozone-destroying chemicals, principally CFC (chlorofluorocarbon) solvents. One of the main uses of CFCs at present is the cleaning of fluxes from printed circuit boards.

Today's fluxes, usually rosin-based, are necessary to promote wetting of the base metal by the solder alloy. For solder to be drawn into the gap of the soldered joint and to make an effective bond, the metal surface and the solder must be free from oxide or other obstructing films. But fluxes have to be removed after they have done their job; otherwise there is a risk of corrosion or circuit losses or possible fouling of moving parts.

Figures from 1989, the most recently available, indicate that globally more than 100,000t of the chlorofluorocarbon CFC-113 were used to clean fluxes from soldered printed circuit boards. That figure was approximately half the world's annual consumption of CFC-113.

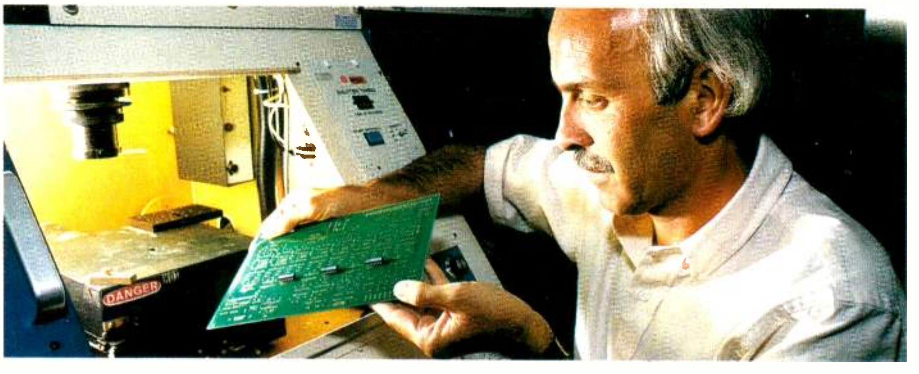
Researchers at Sandia's Center for Solder Science and Technology are developing CFC-free, fluxless soldering methods that do not require the use of any CFC solvents. These methods involve the development of a number of different techniques including the use of controlled atmospheres, thermo-mechanical surface activation and protective coatings.

Controlled atmosphere soldering makes use of various "clean" or chemically-reducing atmospheres to maintain solderable surfaces. As well as the use of a vacuum these atmospheres include inert or reducing gases, reactive plasmas or dilute acid vapour/inert gas mixtures. In controlled atmosphere soldering, researchers have demonstrated that hydrogen plasma cleaning of heavily oxidized copper can produce oxide-free, solderable surfaces at temperatures below 250°C.

Thermo-mechanical surface activation soldering uses different forms of energy to break up the surface oxide and to facilitate wetting of the underlying metal. The use of lasers, solid-state diffusion and ultrasonics are typical ways in which this can be achieved. These thermo-mechanical processes can be done either in air or in a controlled atmosphere.

Using a laser, for example, Sandia researchers have developed a soldering process for making closure joints on radar modules. This fluxless soldering involves the melting of a pre-placed solder preform with a directed laser beam under a protective gas cover. Rapid heating and cooling of the soldered joint produces a very fine microstructure with improved mechanical properties.

The Center for Solder Science is also working with the University of California at Berkeley and the State University of New York at Stony Brook to develop protective coatings that enhance fluxless solder wettability. This involves studying the microstructure and fluxless wettability of nickel-gold platings and also the bonding behaviour of organic inhibitors on metallic surfaces and their effect on subsequent solder wetting.



A Sandia scientist inspects a printed a circuit board produced through fluxless laser soldering. A laser soldering station is in the background.

Staring winkers open windows?

Scientists at Leicester University and De Montford University (formerly Leicester Poly) have developed a way of enabling severely disabled people to operate Windows programmes without the need for a mouse or a keyboard. Ergonomist Howell Istance has made use of existing technologies that make a cursor move wherever the operator's eyes are pointing.

Together with optometrist Peter Howarth, Istance has developed the art far beyond that of existing systems. These are mostly limited to distinguishing into which of four screen quadrants the eyes are looking. Such systems, as well as being relatively crude, also require specialised software. This has excluded a lot of disabled users from using much available computer software.

The Leicester equipment uses a Micromeasurements system *7000* binocular eye tracker, employing a pair of infra-red cameras that sample eye positions 60 times a second. As the eyes move up, down, right or left, the system records each eye position and sends the appropriate commands to the computer. The eye tracker can thus emulate a mouse and the cursor or pointer moves around the screen as the operator's eyes move.

Emulating the movement of a mouse is one thing; the more difficult task has been to provide an eye-driven equivalent of the operate button. Peter Howarth says that he is currently working on two different approaches, one based on staring at the required icon for an unnaturally long time and the other based on winking at it.

Normally when selecting items on screen, the eyes flick about continuously, and rarely settle anywhere for long. Howarth says that it is easy to simulate a command function by arranging the equipment to recognise an unnatural stare (it also sounds a wonderful way of prompting operators who daydream).

The second approach is even more interesting. Since the eye tracker is a binocular system, it can be programmed to recognize differences between the movement of one eye and the other. Howarth says that, normally, we do not move our eyes independently. The system can therefore ignore a blinking (or sleeping) operator, but instantly recognises some unnatural eye movement such as a wink. (The mind boggles at what complications this could lead to in the average office.)

The accuracy of the system at present is such that a person can play a game called "Windows Solitaire" with little difficulty. For screen targets of 20 pixels high by 50 pixels wide, the system performance is comparable to that of a mouse. For larger targets it can be superior. The only snag, says Howarth, is that you have to keep your head reasonably still. That may be enough of a restriction to prevent the system coming into widespread use among able-bodied operators, but for those who are currently unable to operate any sort of machine, it could prove a godsend. The researchers are keen to hear from anyone interested in discussing potential applications. ■

THE RF DESIGN REVOLUTION

Ten years ago, technologists were predicting an information explosion – not only an increase in information, but, equally, the means to communicate it. They envisaged that user freedom would require radio frequency technology operating in untried parts of the spectrum. But they needed a new generation of semiconductors to drive an RF revolution. A new design series by Tim Stanley.

Modfets, GaAsfets, tegfets, beam lead diodes, planar doped diodes, mmics and asics are keywords in the revolution's lexicon. Some of these device types are effective at hundreds of GHz; millimetre-wave devices return noise figures measured as fractions of a dB. Multiple functions on a single chip have, in many applications, obviated much circuit design effort. For example, a complete, digital paging receiver will soon be offered using only two ICs.

This series will consider devices from the application viewpoint. It will include low-noise discrete radio devices; radio systems ICs and building blocks including digital signal processing; RF power devices; passive devices; asics, and other devices for special applications, such as animal tagging and some non-communication RF applications in medicine.

We start at the front end, with discrete devices for low-noise receiver applications.

Modulation-doped field-effect transistors – modfets

The modfet is the fastest three-terminal semiconductor device in the world. Also known as the high electron mobility transistor (hemt),

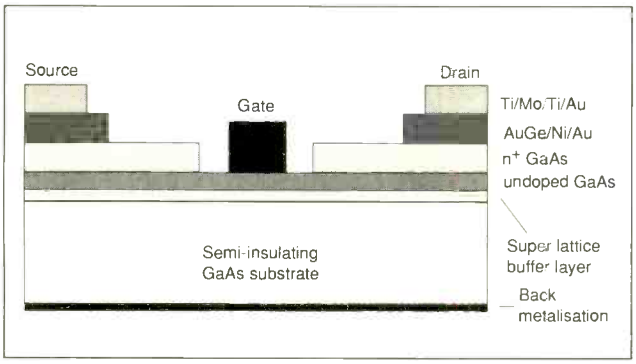


Fig. 1. GaAs mesfet structure.

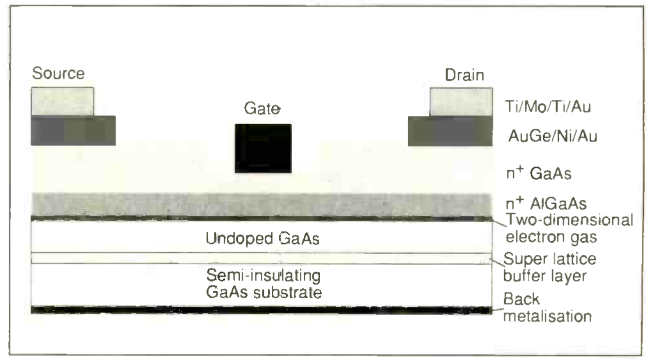
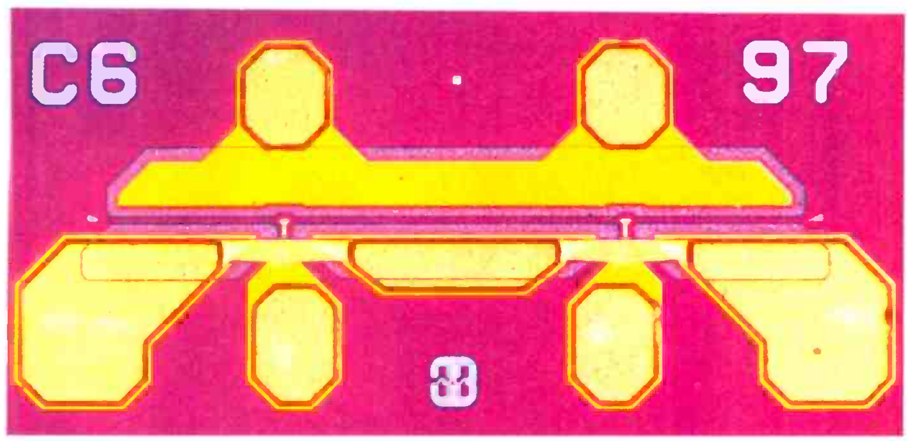


Fig 2 Modfet structure, showing the critically-narrow electron gas layer. This development of the GaAsfet yields a transistor operating at over 200GHz.



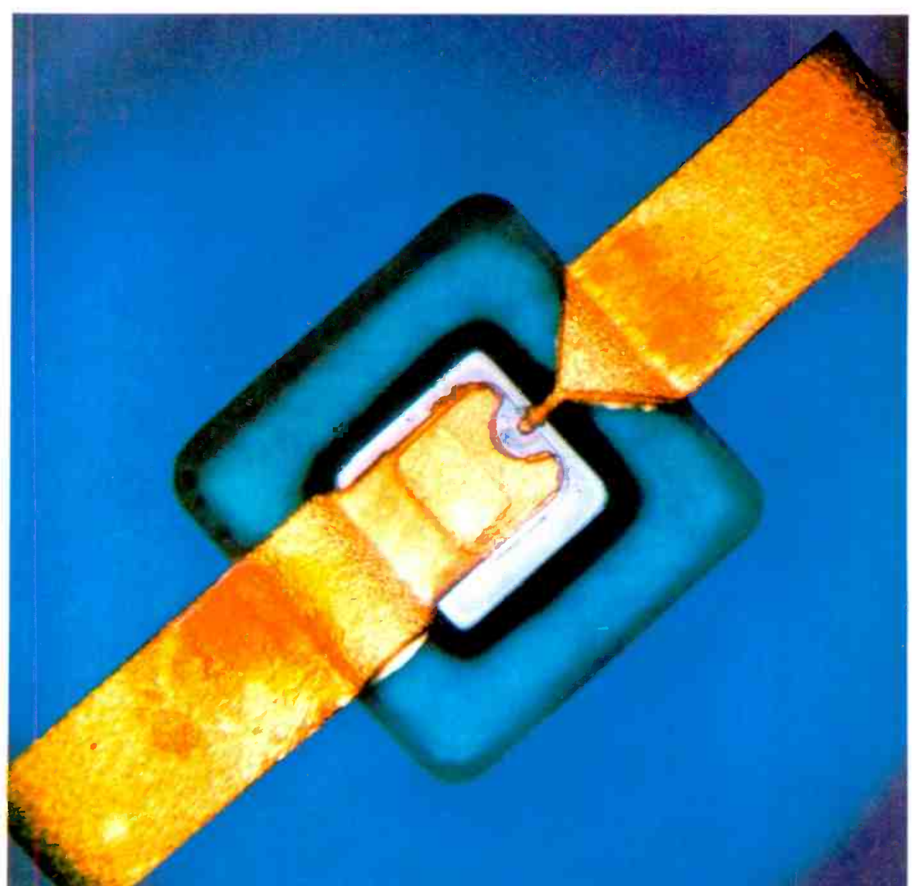
This Harris microwave gallium arsenide fet is the first in a series of standard fets optimised for Ku to Ka performance. At 18GHz, the n-channel HMF03400 offers 8.5dB typical maximum available gain (mag), 11dBm typical output power at mag tuning, 15dBm typical output power at 1dB gain compression and a typical 1dB compressed gain of 5.0dB. At 26GHz, a typical mag of 5.0dB is obtained

selectively-doped heterostructure transistor (SDHT) or the two-dimensional gas field-effect transistor (tegfet) the modfet currently leads technology for lowest-noise microwave applications. Present-day performance delivers noise of 0.6dB at 60GHz; transition frequencies exceed 250GHz with 400GHz devices under development. But how is it done?

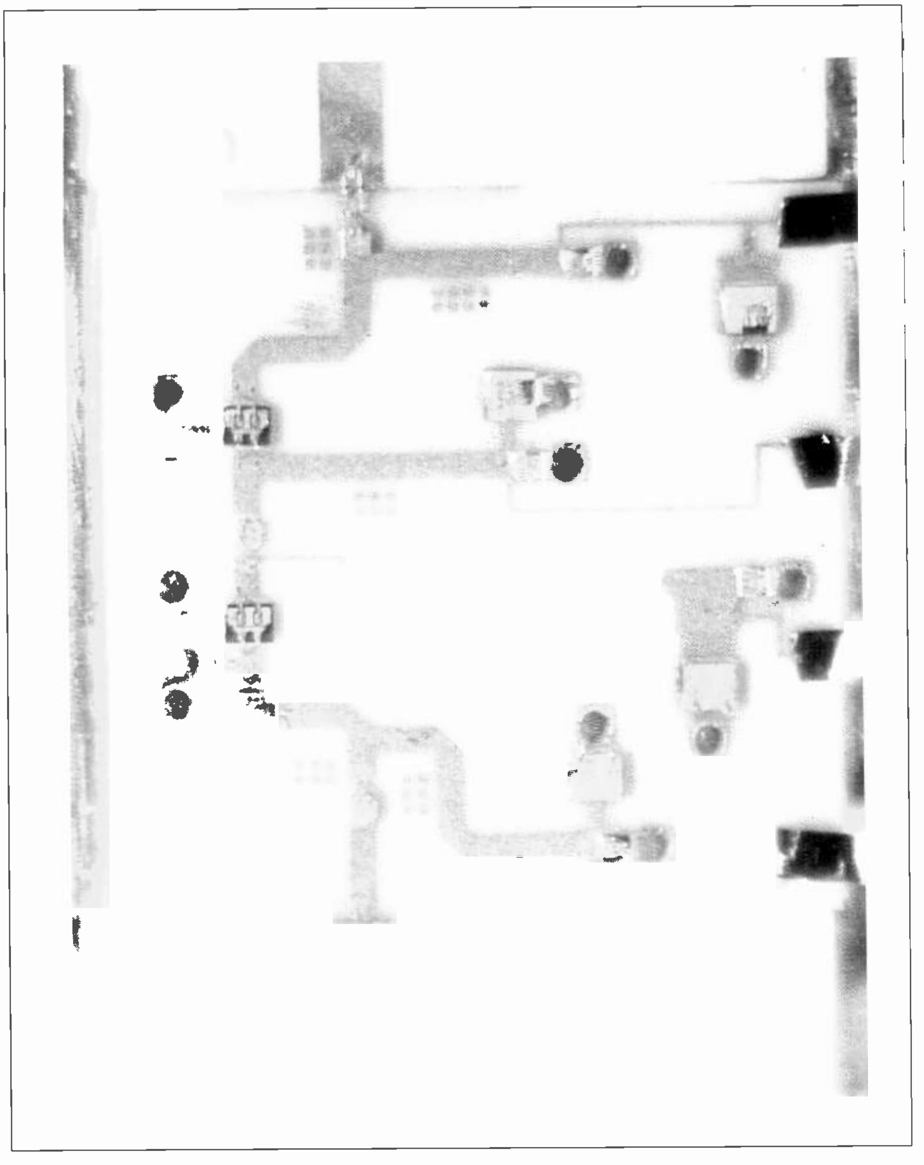
Gallium Arsenide fets have been with us for many years, a variant being the metal-on-semiconductor fet (mesfet) shown in **Fig 1**. This has been refined by reducing the gate length, using photo-lithography, to a practical limit of about 0.35μm. A noise figure of about 1.2dB at 12GHz is about the limit with this device. Electron beam lithography can further reduce gate geometry, but requires serially exposing each gate rather than a complete wafer, and therefore is not suitable for cost-effective mass production.

The modfet is a mesfet but with a very thin layer of electron gas between the AlGaAs and GaAs layers. This is achieved by constraining i.e. "modulating" the electrons in the gas by applying an external voltage across the heterojunction interface to constitute a "two-dimensional" electron gas, **Fig 2**. Hence the term modfet. This technique is more appropriate for high volume, low cost production.

Figure 3 shows noise performances for two variants of modfet compared with the mesfet demonstrating the former's superiority, particularly at millimetre-wave frequencies.



Beam epitaxy allows single crystals to be produced one layer at a time. This results in a barrier that is a single, high quality crystal rather than a relatively crude metal-semiconductor interface, eliminating many undesirable properties of Schottky diodes. This beam lead diode from GEC Plessey Semiconductors. requires just 280μW of LO power for maximum conversion efficiency.



Modfets as used in a front end radar system amplifier for use on the proposed Polar Platform satellite POEM 1. Produced by Matra Marconi Space, it achieves a noise figure of 0.7 dB with 21 dB gain at 5GHz. The modfet is described by them as having demonstrated "excellent reliability for space-borne, low-orbit applications."

Applications

DBS manufacturers have used the resultant improvement in receiver sensitivity to shrink dish size. The additional cost is about $2 to $5 for a packaged, 0.25μm gate length device. A low-noise down-converter by Matsushita, comprising a modfet followed by three GaAs I.C.s, achieves a system noise figure of less than 1.3 dB with an associated gain of at least 60 dB from 11.7 to 12.2 GHz.

Mitsubishi's recommended line-up of its discrete devices for the same application is shown in **Fig. 4**, giving overall converter block noise figures from 0.8 to 1dB depending on the choice of the first-stage device.

The modfet is also a unique device for operation at low temperatures, where high electron mobility is achieved. During the Neptune fly-past by Voyager II, workers at the National Radio Astronomy Observatory (NRAO) replaced the first stage of a three-stage 8.4GHz GaAs mesfet low-noise amplifier, operating at 15°K, with an AlGaAs/GaAs modfet, thus reducing the noise temperature of the LNA by a factor of two from 22°K to 11°K. This improvement allowed NRAO to accomplish the mission without the conventional helium-cooled maser technology.

Low-noise mixers – exit the Schottky?

The Schottky diode has a chequered history in two respects – the behaviour of practical devices has often been too variable for sustained agreement to be reached in understanding between theory and experimentation, and on the technological side, the sensitivity of the current voltage characteristics to the detailed fabrication conditions has been a persistent problem. A technique known as molecular beam epitaxy (MBE) can now produce single crystals an atomic layer at a time. This means that the active barrier is a single, high-quality crystal, rather than a relatively crude metal-semiconductor interface, with the result that many of the undesirable properties of Schottky diodes can be eliminated. Immediate advantages to the front end designer are lower noise and much lower oscillator drive level requirements.

An MBE structure of particular interest here is the planar-doped barrier diode (PDB), comprising a doping sequence of n-i-p-i-n layers. The p- layer is highly doped, but typically only about 30 layers thick. It attracts electrons to its acceptor sites, setting up a barrier to the motion of other electrons. An important char-

An engineer from Continental Microwave tests LNBs used in broadcast satellites. The popularity of the application has resulted in a mass market for GaAsfets, which has been instrumental in radically reducing production costs.

acteristic of a mixer diode is the tightness of the knee of the I-V characteristic around forward turn-on, which determines the mixing (and detecting) ability of the device.

Marconi Electronic Devices claims to be a world leader in this field, and a comparison of one of its PDB diodes with conventional Schottky is shown in **Fig. 5**. Of particular note is the low level of oscillator drive requirement – as low as 280μW for the minimum noise figure of 6dB as compared with about 1.3mW for a GaAs Schottky device. This reduction in LO drive has obvious advantages, not only in simplified oscillator stages but in reducing LO radiation. It appears that these PDB devices maintain the advantages over the more conventional types to about 100GHz.

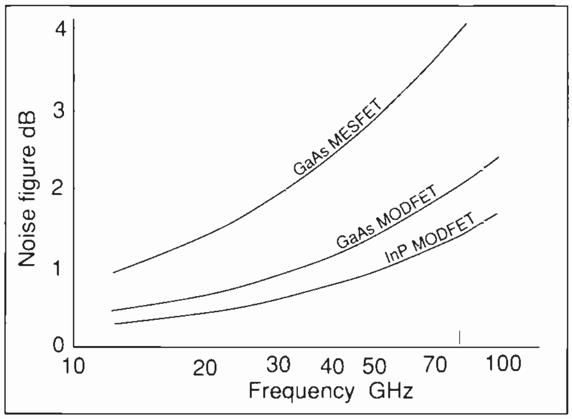


Fig 3 Exceptionally-low noise figures are attained by the modfet, here compared with its predecessor.

Lowering the noise floor

Noise figure is usually the most critical parameter of a microwave receiver front end. It may be described as the ratio, in dB, of the degradation of signal/noise caused by the device:

$$\text{Noise figure (dB)} = \frac{\text{signal at input}}{\text{noise at input}} - \frac{\text{signal at output}}{\text{noise at output}}$$

All units measured in dB

Looking at Fig. 3, we see that for, say, 20GHz, the noise figure of a GaAs mesfet is about 1.5dB, and for a modfet is about 0.5 dB. Hence, we might expect that the modfet device would give an advantage over the mesfet of about 1dB in noise figure, and hence of 1dB in signal/noise ratio at the receiver output (assuming the devices are similar in other respects).

However, this assumes that the input noise reference is that generated by a resistor at room temperature, i.e. 21°C which is 294°K. In practice, the actual source in a working system comes from the aerial and whatever noise it is picking up. The source noise from the sky is much less than 294°K at 20GHz for an aerial pointing skywards but not at the sun. It probably averages about 3°K. However, the aerial system itself is a noise generator, owing to the thermal movement of electrons in the aerial and feeder, etc. Some noise may also be picked up by side-lobes of the aerial (which is typically a dish at these wavelengths) from relatively warm objects such as the earth and buildings. It is difficult to predict what the equivalent noise temperature of an aerial system might be, but 100°K is a reasonable assumption.

We may recalculate the noise performance of the devices with reference to 100°K which gives new "noise figures" of 3.3dB and 1.3dB for the mesfet and modfet respectively. Hence, the advantage in this system would be 2dB (i.e. 58% as a power ratio) not 1dB (26%) as might have been assumed by simple comparison of conventional noise figures. The striving of the device designer for fractions of a dB in noise performance is more significant than initially meets the eye.

Even so, an advantage of 2dB may seem trifling. Perhaps so, if the circuit or system designer is concerned with a "one-off" radio link or small network. Consider, though, the cost savings in a large communications network such as a cellular telephone system, where a 2dB advantage in receiver sensitivity means that the transmitter output power rating can be 60% less (assuming no extraneous off-air interference) giving capital and running-costs savings. Multiply these by the hundreds of transmitters required for national coverage, and the cost effectiveness of these modern receiver devices can be appreciated.

Consider, also, a radar system, where the high, and expensive, transmitter power requirement can be reduced by improved receiver sensitivity. Again, the cost advantage is obvious, particularly where supply power and aerial size may significantly be limited, such as in a satellite, or an airborne radar system. As Fig. 3 shows, the advantage of the modfet increases significantly with frequency.

Next month ICs at the front end.

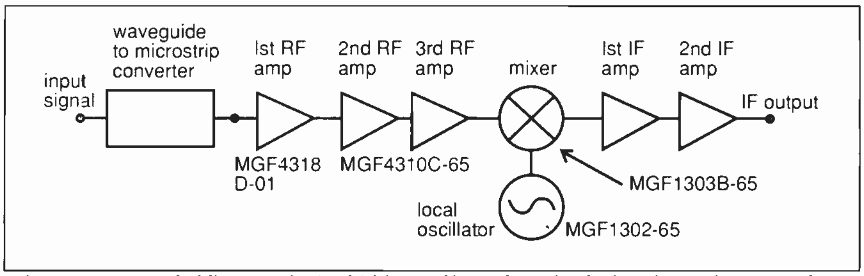


Fig. 4 Recommended line-up of Mitsubishi's modfet and mesfet devices for a Direct Broadcast Satellite receiver head. The super-low-noise first stage is achieved with devices of sufficiently low-cost to be ideal for consumer electronics.

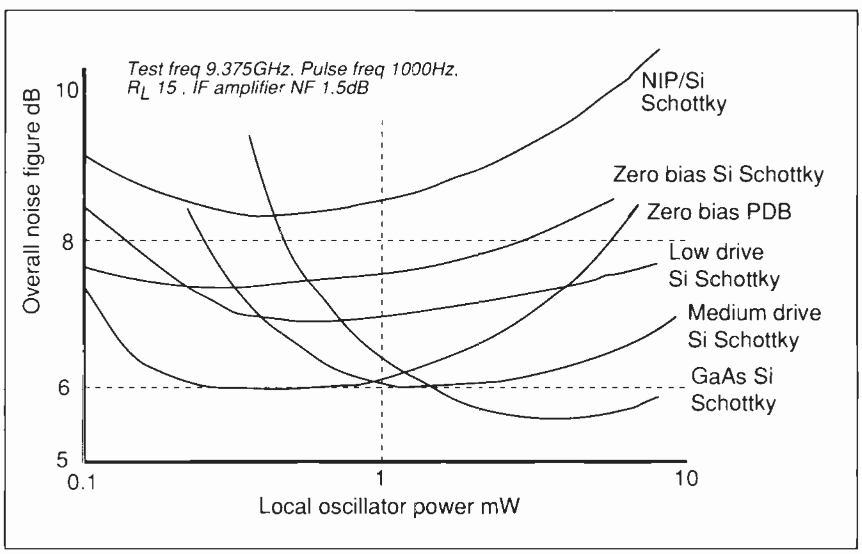


Fig .5. The planar barrier diode gives low-noise performance for receiver mixers with oscillator drive as low as 0.2 mW.

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Marconi/Saunders Signal Sources type – 6058B – 6070A – 6055B – 6059A – 6057B – 6056 – £250–£350. 400Mc/s to 18GHz.
Marconi TF1245 Circuit magnification meter + 1246 & 1247 Oscillators – £100-£300.
Marconi microwave 6600A sweep osc., mainframe with 6650 Pl – 18-26.5GHz or 6651 Pl – 26.5-40GHz – £1000 or Pl only £600.
Marconi distortion meter type TF2331 – £150, TF2331A – £200.
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Tektronix Plug-ins 7A13 – 7A14 – 7A18 – 7A24 – 7A26 – 7A11 – 7M11 – 7S11 – 7D10 – 7S12 – S1 – S2 – S6 – S52 – PG506 – SC504 – SG502 – SG503 – SG504 – DC503 – DC508 – DD501 – WR501 – DM501A – FG501A – TG501 – PG502 – DC505A – FG504 – P.O.R.
Ailtech Stoddart receiver type 17/27A – .01–32Mc/s – £2500.
Ailtech Stoddart receiver type 37/57 – 30-1000Mc/s – £2500.
Ailtech Stoddart receiver type NM65T – 1 to 10GHz – £1500.
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Tektronix Mainframes – 7603 – 7623A – 7633 – 7704A – 7844 – 7904 – TM501 – TM503 – TM506 – 7904 – 7834 – 7104.
Knott Polyskanner WM1001 + WM5001 + WM3002 + WM4001 – £500.
Ailtech 136 Precision test RX + 13505 head 2 – 4GHz – £350.
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Tektronix M2 gated delay calibration fixture. 067-0712-00.
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B&K 4815 calibrator head.
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Telequipment CT71 curve tracer – £200.
H.P. 461A amplifier – 1kc–150Mc/s – old colour – £100.
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Tektronix oscilloscopes type 2215A – 60Mc/s – c/w book & probe – £400.
Tektronix monitor type 604 – £100.
Marconi TF2330 or TF2330A wave analysers – £100–£150.
HP5006A Signature Analyser £250 + book.
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Tektronix plug-ins – AM503 – PG501 – PG508 – PS503A.
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CIRCLE NO. 116 ON REPLY CARD

Improved preamps put dat back on the road

The new dat recording machines combine portability with high quality recording. But microphones used may be degrading the result. Adrian Pickering and Max Hadley investigate.

Digital audio personal-stereo-type machines can now offer high recording quality while on the move, with performances hitherto only achievable with professional quarter-inch recorders such as a Uher or Nagra. The dat recording format is stereo, 16-bit linear sampling at 48kHz, of comparable quality to CD. But, since almost all recordings made with a portable dat machine are likely to be from a microphone, the performance of the microphone preamplifiers becomes critical. Unfortunately, manufacturers have not paid as much attention to this as perhaps they should – the result is often a dat machine with recording quality degraded by microphone preamplifiers more suited to a good cassette recorder!

The solution is to substitute a better-quality microphone preamplifier. This is relatively straightforward if there are no constraints on power consumption or size. Clearly, the whole purpose of using a portable dat machine is its convenience, so any ancillary equipment must not cause compromises here. The requirement is for a compact dat-quality microphone preamplifier with a low power consumption.

Preamp requirements

The low-noise preamplifiers traditionally employed in microphone input transformers have many advantages, and provide a balanced input and excellent common-mode rejection. But they can be very susceptible to stray magnetic fields which are often encountered in "out-and-about" recording locations.

Compact transformers with a good AF response and immunity to external fields are expensive and so a transformerless design is to be preferred. A balanced input would be desirable, though not absolutely essential since the microphone lead is typically only 1-2m long.

Microphone type does affect preamplifier design. Professional moving-coil microphones have low output levels. Commonly-available electrets give a higher output, but this is accompanied by their internal preamplifier noise.

Studio-quality capacitor microphones have high output levels and low noise, but are expensive and usually phantom-powered.

For portable use, powered microphones of any type are a nuisance as they need constant feeding with fresh batteries to assure an uninterrupted "take". So the moving-coil microphone still has much to commend it and the basic performance specification for the preamplifier thus emerges:

Low noise. The preamplifier should have the best noise and distortion performance attainable with the available power supply (see below). Resulting recordings should be at least as good as those made on portable quarter-inch recorders (typically, 60dB signal to noise ratio).

Gain. A typical professional 200Ω moving-coil microphone in face-to-face interviews generates a signal level of about -65dBu. 0dBu (0.775Vrms, 1mW in 600Ω) is the standard line level for professional equipment. Taking into account the extra gain available from the dat line inputs, the preamplifier gain needs to be about 45dB. Any higher gain would reduce headroom and generate excess noise.

Output. The dat with its unbalanced line inputs is used right next to the preamplifier so

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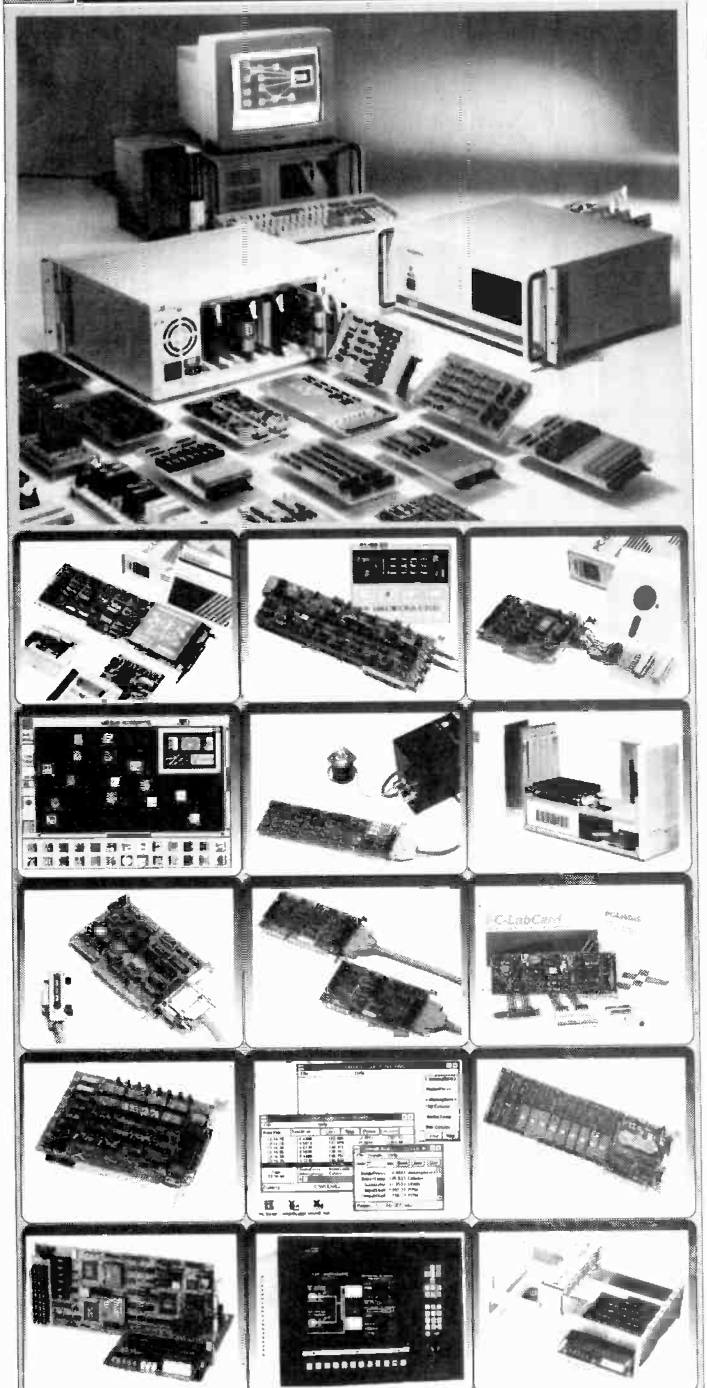

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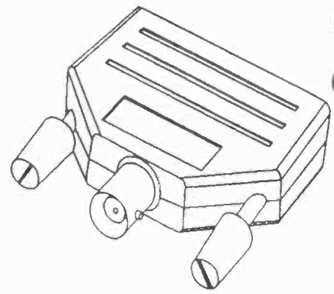

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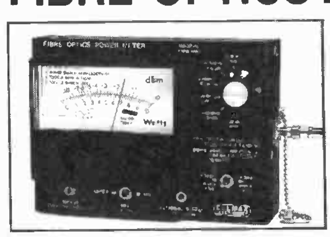

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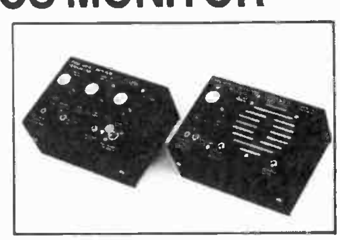

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that the output from the preamplifier can be unbalanced too.

Power supply. Difficult access to the dat power supply makes batteries necessary. There should be a single battery of a type commonly available from local shops and it should be replaced as rarely as possible. Equipment performance should gently degrade as the battery reaches the end of its life.

Users – radio journalists or electronic engineers – must have confidence in the equipment: conducting an interview can be stressful enough without having to worry about the equipment. Any doubts over reliability will lead to the machine rapidly being rejected in favour of equipment that worked before. Equipment must be robust and idiot-proof – and useable. However good the electronics, if equipment is difficult to use it will surely be discarded.

Rotary switches. Rotary switch types – least likely to be misoperated by something rubbing against the equipment – should show a clear indication of the position of the switch. If switches of any other type are used they should be guarded or stiff to operate.

Minimise the number of connectors. As the enclosure must be compact, small connectors and fragile, thin, cables are undesirable. Instead, robust leads should be wired direct from the enclosure and be long enough so that, during normal use, they plug straight into the microphone(s) via a locking XLR.

Positive on/off indication. A constantly-illuminated light should indicate that equipment is on and functional, and also helps to remind the user to switch the equipment off after use. The light should serve as a battery health indicator.

Mono mode. Most radio journalism is still carried out in mono, though dat provides stereo. The otherwise-redundant right channel could be used to record the mono signal at a 6dB lower level in case clipping occurs on the left; but the option switch will rarely be used and should be hidden. A stereo preamplifier must work in stereo or mono, saving battery power when mono mode is used.

Reverse battery protection. A battery replaced in a hurry may be inserted the wrong way round with equipment switched on. But the circuit must not be damaged or degraded.

Input and output open- and short-circuit behaviour. Leaving a microphone input open-circuit or shorting the input should not generate clicks or noise. The output should survive being shorted indefinitely.

Spatial consistency and simplicity. Common-sense rules should govern layout – eg the left microphone lead should be on the left side. Equipment status should be obvious from a quick glance at one face of the enclosure.

Minimum interacting controls. The wide dynamic range of the dat and preamplifier results in the recording level being uncritical. As the dat already has a level control, the preamplifier does not need one. In practice the gain is usually adjusted by moving the microphone.

Design and op-amp selection

A professional moving-coil microphone has a typical impedance of 200Ω. Since a microphone transformer is to be avoided, an amplifier with a very low equivalent noise input voltage is required. For common mode interference rejection, a good input configuration is a well-matched differential pair of transistors[1]. Low input voltage noise requires the input transistors to be bipolar types with a high current gain, h_{FE}, run with a comparatively high collector current and with a low base spreading resistance[2].

A suitable IC low-noise operational amplifier is a convenient solution. Among the best is the Linear Technology[3] *LT1028* which has an equivalent input voltage noise density, e_n, of 0.9nV/√Hz at 1kHz and an input current noise density, i_n, of 1pA/√Hz. Input transistors operate with 0.9mA collector current and the total supply current is typically 7mA.

The *LT1028* will operate at ±4.5V, which means it is possible to power it from a 9V battery. The specialised Analogue Devices *SSM2017* microphone preamplifier performs no better but has significantly higher power consumption (10mA) and requires split supplies of >±6V – difficult to use with a single 9V battery.

In the circuit (**Fig. 1**), the *LT1028* is used as a classic differential amplifier with the input resistors omitted and their function performed by the microphone source resistance, R_s. Gain is inversely proportional to R_s. Any resistances in the input generate their own thermal noise and should be avoided; only R_s is clearly unavoidable. The topology requires that the microphone be able to work into an (active) short-circuit.

It is a popular myth that all microphones must drive into a 1kΩ load. Although most manufacturers characterise their microphones with such a load, their actual behaviour changes little – and possibly even for the better – if they drive a lower impedance. So using an *LT1028* in the circuit, the largest single noise source is the resistance of the microphone itself (1.8nV/√Hz for 200Ω).

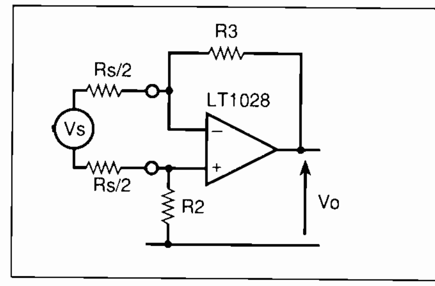


Fig. 1. Modified differential amplifier. The LT1028 is used as a classic differential amplifier with the input resistors omitted and their function performed by the microphone source resistance, R_s.

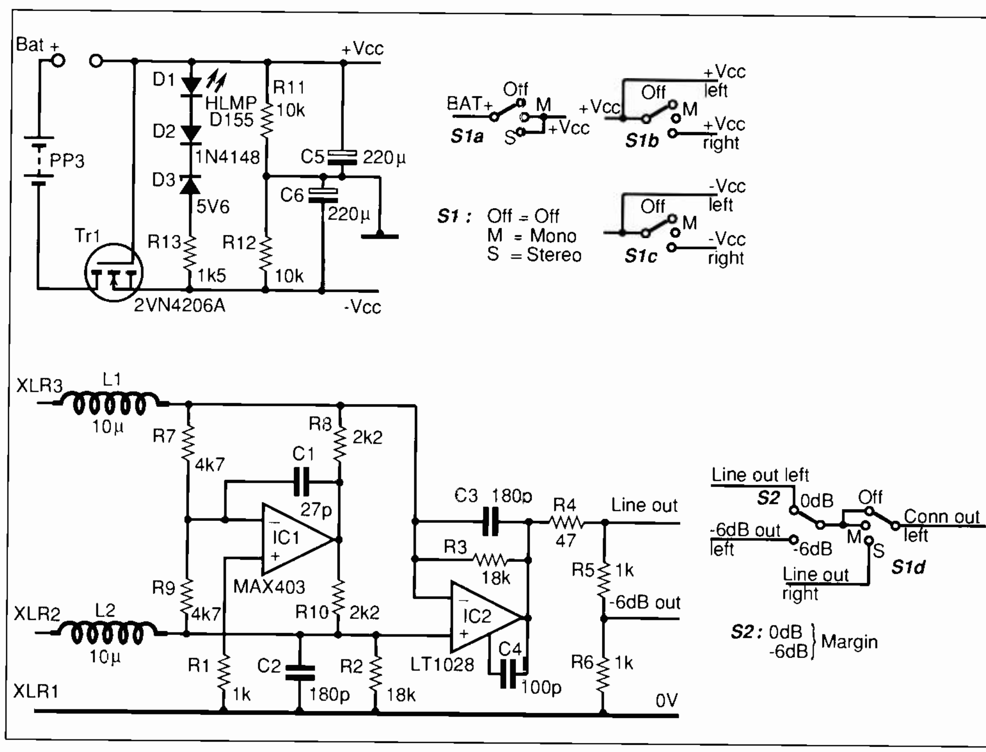


Fig. 2. Low noise, low power microphone preamplifier. Power supply (top left) to left and right channels. Switch SW1 state controls the battery and power distribution to each channel. Power distribution (top right). Rotary switch SW1 has three states: off, battery disconnected; on, mono mode, left channel only powered; on, stereo mode, both channels powered. Preamplifier single channel (bottom left). Left and right channels are identical. Power supplies and decoupling (100nF) are not shown. R_8 and R_{10} should be matched to better than 0.1%. Mono mode signal routing (bottom right). In Mono mode the right channel output is derived from the left channel. Option switch SW2 selects: right equals left, or right is –6dB left. The latter gives further peak overload protection during recording

Modifying the configuration

On its own, the configuration is not wholly satisfactory, as both inputs are driven, via the feedback resistors, at half output voltage. There is a considerable risk that the *LT1028* input could be driven outside its common mode input voltage range when powered from the 9V battery. Also, the operation of some single-point stereo microphones may be affected if both channels share a common earth. Hence the circuitry around *IC1* (**Fig. 2**) is added to force the inputs to lie symmetrically about ground by making their sum zero.

A low power operational amplifier, the Maxim *Max403*, is used to minimise the overall power consumption. Up to the frequency at which the *Max403* runs out of gain (10kHz), IC_2 sees almost no CM signal so its CMRR is unimportant and its distortion is minimised[4]. Note that although R_2 plays no part in controlling differential mode gain up to this frequency since IC_1 provides the microphone current (see box), the common mode rejection ratio (CMRR) at higher frequencies is still dependent on R_2 and C_2 matching R_3 and C_3.

The *Max403* output drives each input of the *LT1028* via a pair of closely matched resistors, R_8 and R_{10}. Its output drive current is split precisely between them. When the *Max403* counters a CM signal, any mismatch between R_8 and R_{10} generates a differential mode signal which will degrade the total CMRR. A match of better than 0.1% assures a CMRR of 54dB at 1kHz.

It is easy to find matched pairs better than this by running a digital multimeter down a short bandolier strip of 1% 50ppm/°C resistors. These are the only two components in the circuit with critical matching requirements. Actual value is not important though larger values will reduce the CM swing that the *Max403* can suppress.

The CM cancelling circuit will also affect the preamplifier noise level. Noise of the *Max403* is injected via R_8 and R_{10} and, being CM, is rejected by the *LT1028*. Also the resistors R_7-R_{10} contribute thermal noise. Because of the low impedance at the output of the *Max403*, R_8 and R_{10} each add uncorrelated noise as if they were in parallel across the input. R_7 and R_9 appear in series across the input. All these resistances are shunted by R_s which, at 200Ω, dominates. Usefully, they limit the effective R_s when the microphone is unplugged, preventing the gain falling below the *LT1028* stability limit (6dB). Output noise remains small because of the reduction in gain due to the larger effective R_s.

L_1 and L_2 provide RF rejection and, since the components also have a small resistance, they limit the gain if the input is shorted. The noise is noticeable in this case as the gain is very high due to a very low effective R_s.

The power supply circuit generates a signal earth from the PP3 battery supply. It also provides a pilot indicator using a high-efficiency Hewlett Packard led. The device is very bright with a power consumption of only 1.5mA when the battery is fresh, dropping to <1mA during battery discharge, dimming completely when the supply is about 7V. N-channel power mosfet, T_1, provides the reverse polarity protection as the voltage drop of conventional protection methods could not be afforded. When power is first applied to the "backwards" device, the inherent substrate-drain diode conducts and a gate-source voltage is generated across the load. T_1 thus bootstraps itself into its resistive operating region with a typical channel resistance of <2Ω.

The stereo preamplifier is housed in a plastic enclosure measuring 25mm x 70mm x 120mm with its interior screened using conductive nickel paint. On the front fascia are the rotary off/mono/stereo switch, the pilot led and the hard-wired left and right microphone leads. The left lead is long enough (1.5m) to be sufficient for mono interviews with the microphone directly connected. The mono margin option switch is accessible via the battery compartment. The output is connected via a short, hard-wired 3.5mm jack-terminated

Interview technique

Following a stint as a "Media Fellow" in 1990, an academic working in the mass media to promote science, I have been a freelance radio journalist for local FM radio, reporting on science and technology.

Shortly after I started, it became obvious that every radio journalist needs his or her *own* tape recorder. The quarter-inch tape machine is still the standard in radio programme production, its format allowing manual editing to the highest broadcast standards with the simplest of tools: chinagraph pencil and razor blade. But the low recording density that permits this is also a fundamental weakness. Portable quarter-inch tape recorders have to be physically large and are tape and power hungry. Nevertheless the StellaVox, Uher and Nagra portables are in wide use, the resultant material being immediately editable.

All other options involve dubbing to quarter-inch for subsequent editing. If this overhead is acceptable then the possibilities extend to high-quality, professional compact cassette or proprietary miniature formats, such as mini-Nagra, and dat.

Dat has become an accepted studio format. To a limited extent, professional portable dats are in use in radio journalism. But the newer consumer portable-stereo-type models, promise the same quality for significantly less money.

The Aiwa *HD-S1* and Sony *TCD-D3* have been tried and both were found to have noisy microphone preamplifiers. I currently use the Sony with a Beyer *M58* moving-coil, omni-directional microphone making mono recordings via the preamplifier described here. The result is dubbed to quarter-inch on a Studer *A807* for editing and subsequent broadcast.

Adrian Pickering

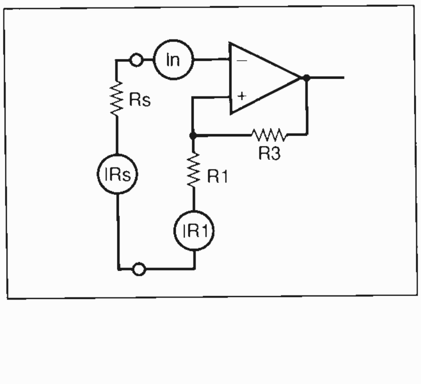
Operational amplifiers and noise

In the figure is an operational amplifier in a high gain configuration together with its equivalent noise voltage source. High gain means that $R_3 >> R_1$. The equivalent noise current sources of each input have been omitted as their contribution is negligible with bipolar operational amplifiers and low source impedances.

Calculations show the addition of the amplifier and resistor input noise densities to find the total equivalent input noise voltage density (e_{neff}) for a number of circuit choices. As the noise sources are uncorrelated, the noise powers add, therefore noise voltages are given by an RMS sum:

$$e_{\text{neff}} = \sqrt{e_n^2 + 4kT(R_1 + R_2)} \quad \text{V}/\sqrt{\text{Hz}}$$

Case 1. R_s = 200Ω, R_1 = 470Ω and a *RC4560* with e_n = 7nV/ Hz:



$$e_{neff} = \sqrt{49 + 11.32}$$

$$= 7.77 \text{ nV}/\sqrt{\text{Hz}}$$

The noise from the operational amplifier is dominant; try another type.

Case 2. R_s = 200Ω, R_1 = 470Ω and an *NE5534* with e_n = 4nV/ Hz:

$$e_{neff} = \sqrt{16 + 11.32}$$

$$= 5.23 \text{ nV}/\sqrt{\text{Hz}}$$

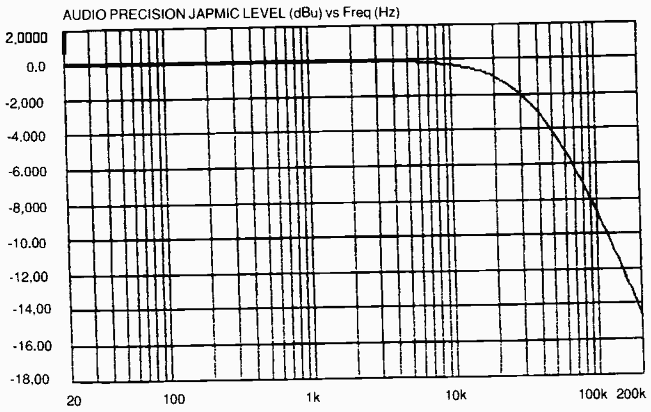
The contribution from the operational amplifier is comparable to the noise from R_s and R_1 and this is about the best that can be done without changing component values significantly.

Case 3. The inverting configuration can reduce the value of R_1 or even omit it, making a lower noise amplifier worthwhile. With R_s = 200Ω, no R_1 and an *LT1028* with e_n = 0.9nV/ Hz:

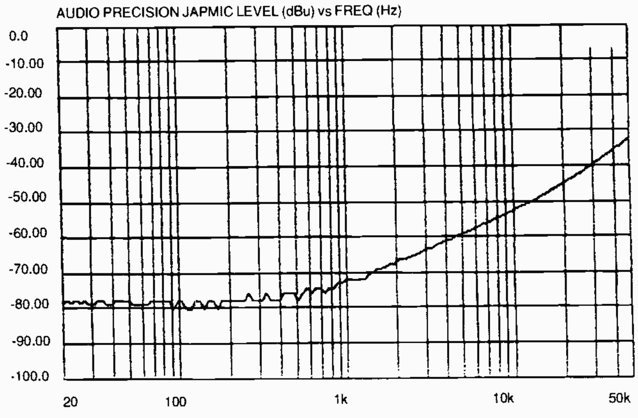
$$e_{neff} = \sqrt{0.81 + 3.38}$$

$$= 2.05 \text{ nV}/\sqrt{\text{Hz}}$$

The noise from R_s is dominant but irreducible. Choosing an operational amplifier with a low e_n means higher equivalent current noise, i_n. Fortunately, this is still insignificant in this configuration. This is the basis of the microphone preamplifier circuit presented here.



a)



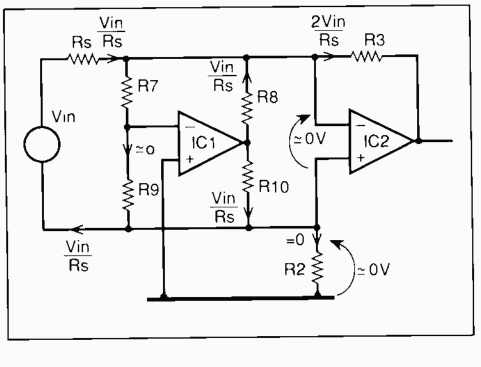
b)

Fig. 3. Performance measurements (a)Differential mode gain vs frequency. (R_s = 150Ω, input level -47.4dBu, analyser bandwidth 10Hz 500kHz.) Plot ref: 18 JUN 92 19:51:41 (b) Common mode gain vs frequency. (R_s = 150Ω, common mode input (-47.4+40)dBu, analyser bandwidth 80kHz.) Plot ref: 18 JUN 92 20:00:40

Differential mode gain

A signal source being "shorted" means there is no input voltage to define a voltage gain. But for the purposes of comparison with other designs the voltage gain here is defined as the output signal divided by the unloaded input voltage.

The feedback around IC_2, the *LT1028*, keeps both inputs at the same voltage. Hence the current flowing through the microphone is V_{in}/R_s and R_7 and R_9 have no volts across them. The feedback around IC_1 ensures that there are no volts across R_2 either. The only remaining route for the microphone current is from the output of IC_1 via R_{10}. IC_1's output voltage will be $V_{in}R_{10}/R_s$. The same voltage appears across R_8 so the same current will flow. Total current into the inverting input of IC_2 is thus $2V_{in}/R_s$ and this will flow through R_3, generating the output voltage $2V_{in}R_3/R_s$. The gain is therefore $2R_3/R_s$, which is the same as the unmodified differential circuit of Figure 1.



lead. The unit is used fastened to the base of the dat using a couple of self-adhesive Velcro strips.

Performance

The preamplifier prototypes work well (subjectively) and are in active service. They have also been used successfully with stereo electret microphones, despite the somewhat unorthodox load they present. **Figure 3** shows some objective performance measurements made using an Audio Precision System One analyser.

Figure 3(a) shows the gain with R_s=150Ω; 47.3dB at audio frequencies and 1dB down at 20kHz. The amplifier is DC coupled and there is no low frequency roll-off.

To measure the CMRR the input level was increased by 40dB and applied CM. Fig. 3(b) shows the output with this input. At low frequencies the output is below the apparent −77dBu noise floor, and so the CMRR is better than 117dB. Even at 20kHz the CMRR is 85dB.

THD measurements show that, with a 1kHz 0dBu output, the only measurable distortion is −112dB (0.00025%) of second harmonic. Once again, this is beneath the broadband noise floor.

Noise measurements with R_s = 200Ω, using a 20kHz brickwall filter and precision millivoltmeter show that the equivalent input noise (EIN) in this bandwidth is −128.2dBu (2.1nV/√Hz equivalent input noise voltage density). The figure compares well with the calculated value using the typical e_n and i_n figures for the *LT1028*. As the thermal noise of a 200Ω resistor at 25°C in a 20kHz bandwidth is −129.6dBu, the noise figure is 1.4dB at this input impedance.

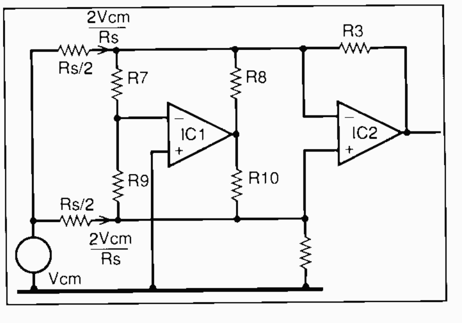
Power consumption is about 8mA per channel and an alkaline PP3 gives more than 40h use in mono mode. Output clip level is over +8dBu with a fresh battery, dropping to +5dBu just before battery failure (7V). At this point the dynamic range still exceeds 88dB, comparing well with the 90dB of the dat. For performance per milliwatt this throws down a strong challenge to any other design. ■

Common mode gain

Input current into the positive input to IC_1 is $2V_{cm}/R_s$. Once again, the only source for this current is via R_{10} and the same current will flow through R_8. Now, instead of adding, this current cancels the input current into the inverting input. No current flows through R_3 and so the output, and hence the gain, is zero. This is only true if $R_8 = R_{10}$. If $R_8 = R - \delta R$ and $R_{10} = R + \delta R$, then the common mode gain is:

$$A_{cm} = 2\frac{R_3}{R_5}\left(\frac{R+\delta R}{R-\delta R} - 1\right)$$

For R = 2k2, δR = 2.2Ω (0.1%) this gives A_{cm} = −10dB, giving about 54dB CMRR. Using a $3^1/_2$ digit DVM it is feasible to match R_8 and R_{10} to within ±0.5Ω giving A_{cm} = 23dB (better than 67dB CMRR).



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Acknowledgement

Thanks to Richard Dudley for making the Audio Precision analyser measurements.

CIRCLE NO. 122 ON REPLY CARD

CIRCLE NO. 123 ON REPLY CARD

PSU regulation boosts audio performance

Ben Duncan explains why audio power regulation is the sensible way to a better listening experience.

Power rails in most electronics are assumed to be stiff enough to hang any amount of circuitry off them without worrying – provided current capability is observed. In reality, stiffness is finite, with finite impedance causing spurious signal-related voltages to be developed along supply buses, commensurate with any change in signal current and/or voltage.

Most signal processing topologies exhibit power supply rejection (PSR), indicating that a change in rail voltage is only partially reflected at the output port. But the PSR ratio (PSRR) of any individual stage is finite too. The upshot is that communal supply rails always induce some interactions between stages. Power rail feed-through makes no distinction between logical sequence, so later stages can interfere with earlier ones, and vice-versa.

In general, with the predisposition to couple wholly unrelated stages and sections, the feed-through may be regarded as complex, even chaotic.

Devolving small signals

How big is this feed-through? Sensible, economically justifiable PCB conductor cross-sections and lengths realise baseline source impedances (Z_s) in the realms of low mΩ. This, combined with typical current fluctuations of tens of milliamps in moderately complex small signal circuits, produces power rail voltage fluctuations in the order of 1-300μV, or–120 to –70dBV.

The PSRR of any half-way-decent signal processing block will be at least –10dB at 20kHz and below, ranging up to –100dB with good circuits at 100Hz. So error voltages entering the signal path will be in the region of –80 to a hypothetical -220dBV.

For the bulk of electronics, pollution on this scale is of no concern. But if greater than –120dBR (ie –146 to –160dBV; (see terminology explanation box), it can be a problem in music reproduction – arguably the most critical class of processing. Here an error-to-noise ratio (ENR) better than –120dBR is needed.

Real requirements of regulation

To be positively inaudible, the output stage's supply should be suppressed so its contribution at the output results in SPLs below the hearing threshold. Assuming a maximum capability of 125dB SPL, then supply artefacts must be at a signal level that creates <–125dB SPL, or –125dBR relative to the system's electrical output.

If the PSR in the audio band can fall to 25dB worst case, but no lower, we need <–100dBR on the rails. To achieve this, with a modest maximum peak current of 20A, a PSU impedance below 25μΩ is required. If attained by capacitative decoupling alone, a hypothetical 250 pure Farads would be required to

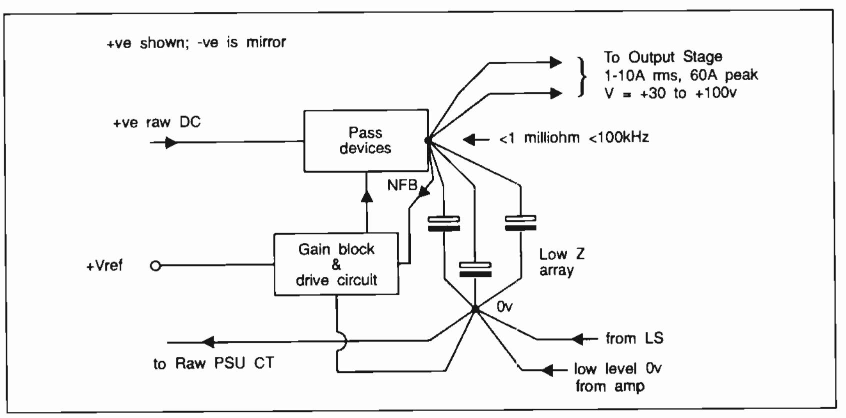

Fig. 1. Audio power regulation scheme. Vr_{ef} is external and independent. The active device's inductive Z_0 is compensated for, and heavily damped up to RF, with an ultra-low impedance reservoir array, tightly noded. Audio output current is drawn directly from across the array.

TERMINOLOGY

High sonic resolution: A subjective description indicating relative absence of intermodulation, adverse harmonic and other error signals occurring simultaneous with music program.

dBR: dB referred to a given reference voltage, in this case a power amplifier's maximum rms output, typically ranging 20V to 100V, or +26 to +40dBV.

dBV: dB referred to 1 volt rms, any impedance.

PSRR: Power Supply Rejection Ratio, normally expressed in decibels.

Time-population space: The sum of the operating hours of all the population of a given artefact.

maintain this specification down to 20Hz. Clearly, some kind of active regulation is needed.

Beyond a very low and preferably invariant flat Z_0 vs frequency, and low noise, regulator requirements specific to audio have not been widely documented. Stability needs to be verified over a wide range of supply currents and waveforms. Loudspeaker peak current-demand (in worst case conditions up to 60A with some studio monitors[10]) must be sourceable for musically significant periods, up to hundreds of milliseconds. With music and speech, the long term RMS draw from a class A-B amplifier will be around 1-10% of peak capability. But the regulator will need to be able to source 10-25% of peak current demand on a near continuous basis to pass steady-state sinewave tests performed (solely) by makers, repairers and reviewers. A high-power audio regulator demands the best possible load regulation, though line regulation is no more important than usual. Output voltage stability is not an issue, so long as any change is below audio frequencies and stochastic, ie true drift.

Short-circuit protection is also not so essential as in a bare lab or industrial supply. With abusive output loading, the amplifier's output devices will likely expire before the regulator's. To some extent, it was the misguided use of foldback current limiting to hide inadequate current/power capability that sealed the fate of audio power regulators in the early 70s. With modern, high SOA devices and supplies below ±50V, thermal protection on the pass device's heat-sink can be

History

The error voltage problem has been considered obliquely by many in reference to capacitative supply decoupling[1,2]. But the role of active regulators in increasing supply stiffness is sparsely documented. In 1980, Sulzer[3] published a "super regulator" for small signal audio (eg preamplifiers), aimed at markedly reducing supply impedance through liberal use of NFB. While developing a commendably low Z_s at the output node of well below 1mΩ at 1kHz, Z_s is non-flat throughout the audio band, and increases to around 12mΩ at 100kHz. The obvious drawback is that beyond powering a single, abruptly adjacent stage, the regulator's low Z_s will be swamped by the power rail's resistance and inductance, unless supply lines to all additional stages are extremely short and stout.

Since 1983, I have pioneered the principle of powering every stage in my reference standard audio designs[4] from individual regulators. The regulator acts to increase PSR while inter-stage isolation means Z_s is less critical than it might otherwise be. Subsequently, many critical listeners have reported that this scheme makes a lasting improvement to sonic quality. Objective evidence is available in the form of spectral measurements. It sounds extravagant, but this localised application is precisely what the originators of the monolithic regulator foresaw[5]. The question then arises, why not regulate a power output stage?

Two decades of unregulated power

The first transistor amplifiers ran from single supplies. They had mostly local feedback and poor PSR. For today's more critical and better-equipped listeners, regulated supplies would be needed, if only to reduce 100Hz (+ harmonics) ripple feed-through to acceptable bounds. By the late 60s, use of the long-tailed input pair and high NFB provided enough 100Hz rejection.

But a few quality amplifiers, (eg Quad's *303* and later models by Radford and Naim) employed a stabilised supply, ostensibly because it offers a tightly defined power output, irrespective of loading or mains voltage. This is valuable to avoid unexpected clipping if the system is sufficiently underpowered to need to be routinely used within 3dB of maximum output voltage swing. Another advantage is improved isolation between stereo channels if they must share a single supply[6].

In the early 70s, Robert Carver[7] argued that "saggy" unregulated supplies gave better music performance. They could satisfy the high transient excursions, without the cost penalties of a transformer and heat-sink rated to sustain full power continuously. The argument was relevant to steady-state testing, lab and industrial use, but not the majority of music. At the same time, amplifier power began to spiral above 100W, and split rail designs became the norm, requiring either two suitably high current regulators, or else a single regulator combined with rather dubious capacitative divider schemes (**Fig. 1**). Designers were meanwhile becoming increasingly aware, not just of speakers' tendency to dip to a fraction of their nominal impedance

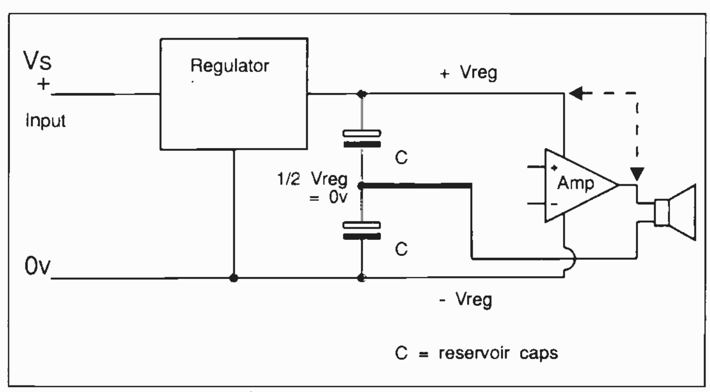

Fig. 1. Capacitative divider schemes overcame the need to have two regulators when split rails became the norm circa 1972. But, the capacitors need to be rated at rail-to-rail voltage, in the case of a DC fault (shown dashed). They must also be capable of passing above 60 peak amperes of output current. Single, large reservoir electrolytics do not meet modern audio quality requirements.

at spot frequencies, but also the ability of more adventurous passive cross-overs to phase shift the drive units' back-EMFs, increasing dynamic current demand by 200% or more.

Altogether, these factors indicated that an amplifier's ability to source high peak currents – around ten times but up to twenty times the maximum steady state – is important if sound quality is not to be compromised in some passages, when driving difficult speaker models. It followed that a regulator would need current capabilities and transient response commensurate with the amplifier itself. The conclusion is a doubling in complexity of amplifier circuitry and cost. Even in an up-market design, it appeared an inelegant route. Benefits were seen at the time solely in terms of steady state specifications at spot frequencies – and these could have been obtained simply by increasing NFB.

The regulator's disadvantage, other than budgetary, was to reduce the amplifier's short-term power or IHF rating, the "big number" beloved of unscrupulous makers. From the mid 70s, unregulated supplies based on increasingly large reservoir capacitors became the norm, and regulated supplies for output stages became a dim memory for most.

The easy way to build a good-sound power amplifier is in class A. Commercial examples above 50W are ecological disasters, both through their electricity consumption, and through employing hugely overrated PSUs. Overrating is partly to deal with almost any loudspeaker's worst case demand, and partly to yield higher sonic resolution. But class A's sonic benefits have little to do with the guaranteed absence of cross-over artefacts, and much more to do with the lack of thermal and power rail intermodulation[8].

Greg Ball's recent article *Distorting Power Supplies*[9] brings the picture up to date and explains with rigour why this should be so, with a cogent survey of the mechanisms at work in the unregulated supplies of today's class A-B audio power amplifiers.

The outcome is that energy saving class A-B can approach or meet class A standards, provided either the supply's or the amplifiers' PSR is vastly improved. To tackle the problem at its heart, we must deal with the former, alias PSU regulation.

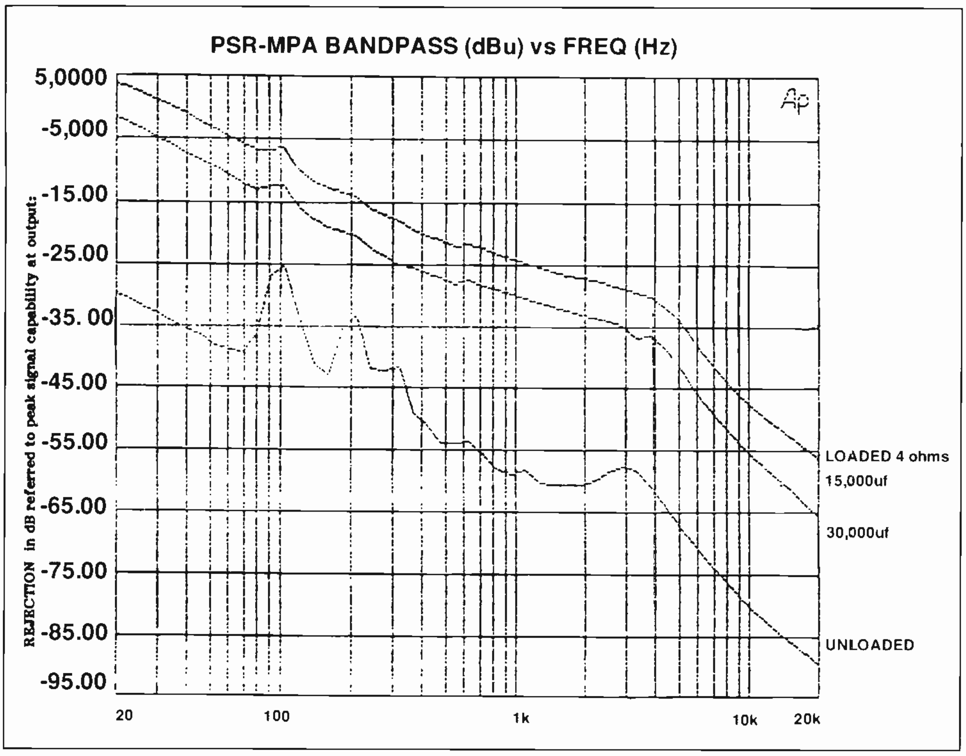

Fig. 2 Audio precision plot of unregulated PSU residue vs frequency, measured in 1/3rd octave bands. The PSU is powering an amplifier driving 4Ω at 8.75A peak (upper pair), and unloaded (lower). Upper pair shows punitive improvement after doubling reservoir from 15,000μF low impedance array to one of 30,000uF. At high frequencies, the residue of a single 15,000μF capacitor would be 10 to 20dB higher. Supply impedance can be inferred from the load impedance and residue's magnitude in dB. For the 4Ω plots, –25dB at 300Hz represents some 200mΩ.

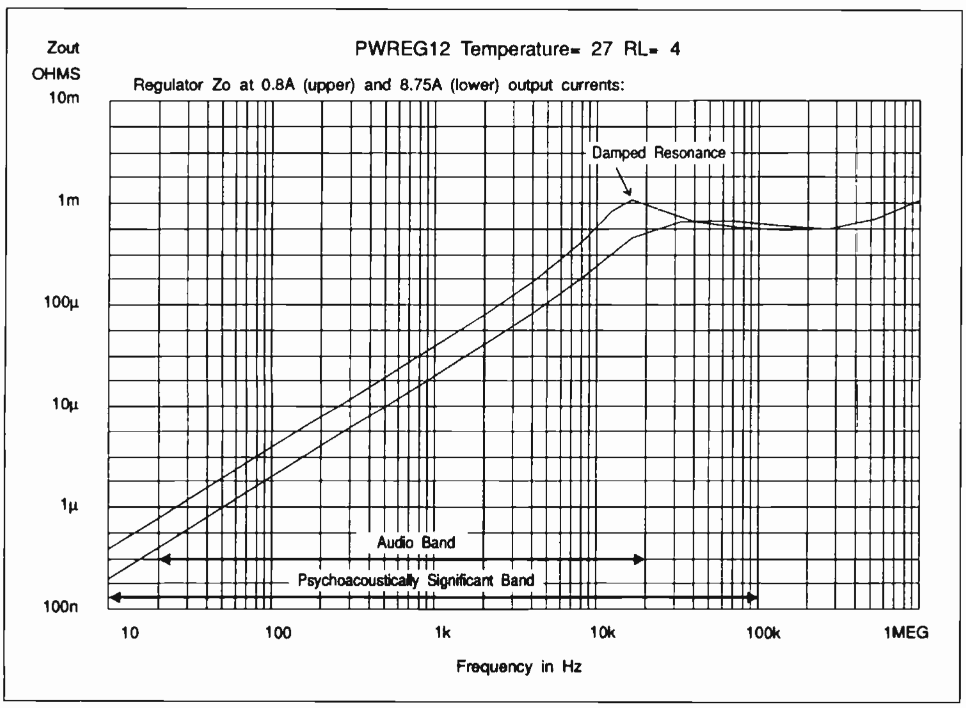

Fig. 3. A MicroCap III model predicts the output impedance of the regulator, using level II op-amp and BJT and lumped capacitor models for speed. Note the disparity in the resonant peak, which MC3 predicts at 17kHz. Further modelling reveals that the peak occurring at about 4kHz in Fig. 4 will shift up to this frequency when the prototype layout is tidied up for production.

enough to protect both the regulator and mains transformer from prolonged over-dissipation.

Design features and results

A design employing series pass devices, controlled by a drive circuit, which uses NFB to sample the output voltage (**Fig. 1**) differs from previous designs in subtle but crucial ways. Feedback ratio in this circuit is exceedingly high, on the order of 10^7 at 10Hz – and still 1000 at 100kHz – and a moderately large but extremely low impedance output decoupling array is part of the regulator and intimately coupled to it. A large capacitance is needed to source transient currents up to 60A or more (short of paying for added pass devices, larger heat-sinks and a mains transformer) – all for a requirement, albeit vital, that persists for just a few monets in any percentage of music played flat out.. Also the Z_0 of any regulator employing NFB looks inductive, as Z_0 rises with increasing frequency, commensurate with reducing open-loop gain. Using an output capacitor to compensate, flattening the Z_0 curve risks resonance. But a very low impedance array intimately connected through heavy bus bars with micro-ohms of series resistance damps the resonance, yielding an almost flat Z_0 across several decades of frequency. The method counteracts the trend for high order harmonics to be emphasised, as in conventional NFB regulators.

Critically, output current is drawn directly across the array, and all grounds are noded to it. Ultra-low inductance needed for RF stability is achieved by paralleling a large number of modern low-inductance units. These are connected in the prototype by multiple braids with resistances under 2μΩ/mm; by their proximity (within 100mm) to the pass devices, and by connecting the NFB sensing wire nodally. Array parasitics should be diminished as far as is practical by analysing the micro-impedance of interconnection geometry. Then conventional compensation capacitors in the order of a few pF can be connected around and within the control loop to tidy up any residue in the 10s and 100s of MHz. The compensation will degrade output impedance and rejection, but at frequencies where it will not bother audio.

No attempt is made to derive the reference voltage ($+V_{ref}$) from the input or output of the supply, as in conventional regulators. Instead, it is taken from the driver circuit's separate supply which is already provided by a conventional, small signal regulator. The key benefit is that regulator output is governed by a low-noise voltage that is inherently isolated from the main supply's large and highly complex ripple and signal-dependent voltage perturbations. Using only three modestly priced bipolar power transistors as pass devices gives a potential 45A capability. So the regulator is easily able to sustain any prolonged (but essentially short-term) audio demand that exhausts the output reservoir: current availability is then limited by the raw reservoir and mains transformer.

Turning to results, **Fig. 2** and **4** show the rejection and Z_s of the unregulated and regulated conditions when used to power a class A-B amplifier, driven at maximum with and without a resistive load. Z_s is estimated by dividing the residue voltage by the output current. The simulation model's predictions (**Fig. 3**) make an interesting comparison. Overall,

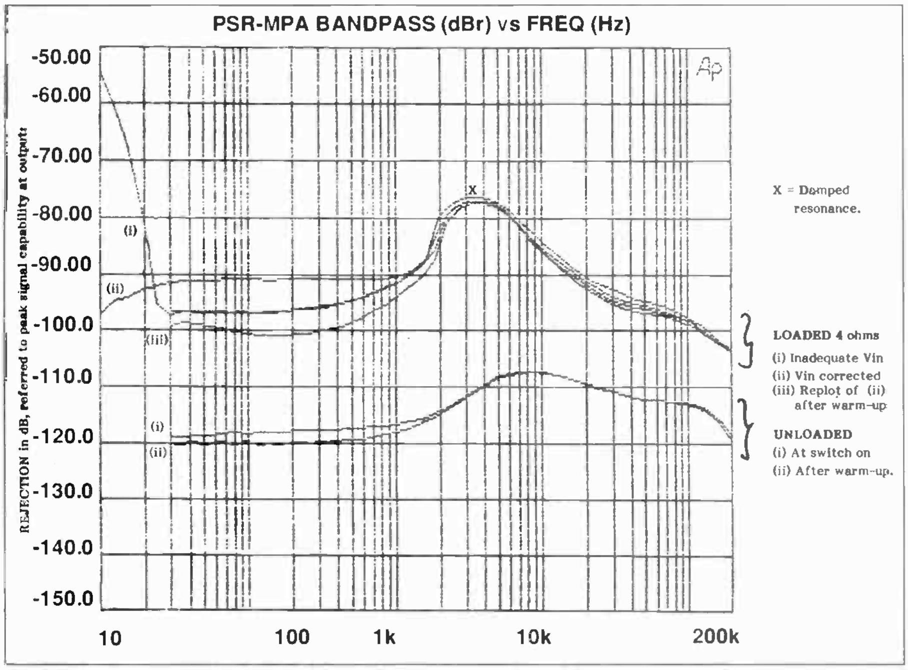

Fig. 4. PSRR of the completed regulator, driving 8.75A peak into 4Ω (upper plots) and open circuit (lower). Damped resonance centred at 4kHz shows effect of imperfect noding and layout in prototype. Rise below 30Hz shows dropout due to inadequate V_{in}, fixed by increasing transformer voltage. All plots include stochastic noise, explaining in part how they differ from simulations in Fig. 3.

the improvement in listening, static THD and driven SNR numbers, and the implied spectra is real and vast (**Tables 1** and **2**). Most of all, the regulator is not expensive – especially taking into account the savings made by being able to reduce the primary supply's component ratings by up to ten-fold. ■

Table 1. PSR in dB with full drive into 4Ω at 10Hz, 1kHz and 20kHz.

Condition	10Hz	1kHz	20kHz
Unregulated supply of 15,000µF	nil	–24	–57
+ regulated supply with 15,000µF output array	–92	–90	–92
PSR improvement	**+92dB**	**+66dB**	**+35dB**

Table 2. %THD+N reduction.

A given amplifier driven at 0.5dB < clip into 4Ω resistive

Without regulator:	0.002% < 100Hz rising to 0.04% at 20kHz
With regulator:	<0.006%, 20Hz - 20kHz (flat)

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ALL-POWERFUL – or all too much?

P-Cad does everything you will ever need from PCB cad. But John Anderson wonders if you can have too much of a good thing.

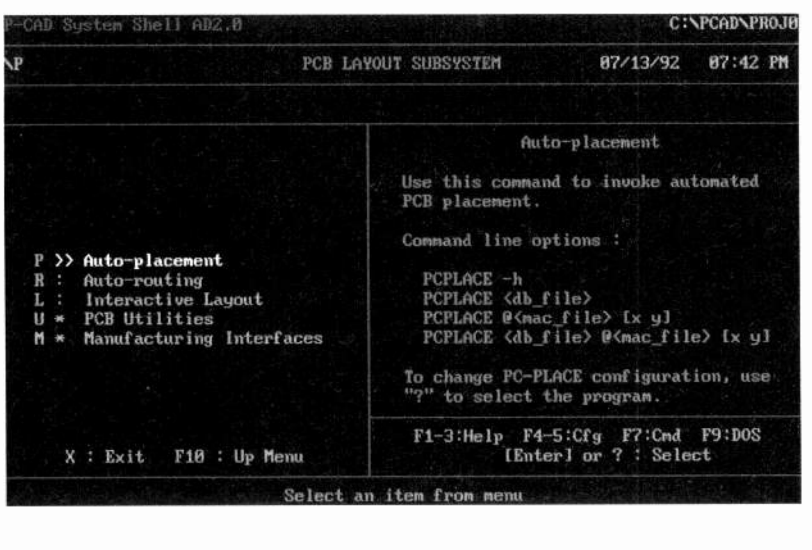
Few PCB design packages can claim to have *all* aspects of the process covered. But compared with the rest, *P-Cad* has everything, taking the design from a hierarchical schematic capture to the control programs for manufacture pick and place. Complexity is one of the penalties that has to be paid for this level of functionality. So a little formal software training could prove a worthwhile investment to get the most out of the package.

P-Cad itself is a mature PCB layout product, with a somewhat chequered commercial history. P-Cad Inc was acquired by Cadcam specialists Cadam in 1989, and Cadam was in turn acquired by IBM. So the software is actually the *de facto* IBM PCB cad program, with over 19,000 systems installed worldwide.

The user interface is based around a series of two-pane text screens, with options in the left hand pane selected using the cursor keys.

As each option is selected, the right hand pane changes to show either the structure of the program and database at that level, or details of the operational syntax. So the somewhat archaic text interface is actually quite a good route map for using the system.

An alternative is that each program in the system can be executed from the dos command line.



Auto-placement menu. The system works in the graphics editor environment, and with component footprints loaded, the user can move and fix the position of specific (or all) components.

Schematic entry

The schematic entry graphics screen is well organised with all controls normally mouse operated. There is a menu of commands on the right of the screen, with some commands being two layer, bringing up a different selection menu. Drawing, redrawing and utility of the schematic entry is generally very good.

But one area which can cause confusion is in informing *P-Cad* about the directory structure of the data and library files – even though *P-Cad* installation put the files there in the first place.

The reason is that entering schematic capture from the command line results in a small menu from which default set-ups, including directories, can be set. Inside the schematic editor there is no way to access information about the current directory or move to another, and to edit a file in a directory that is not configured requires typing in the whole path – tedious. One way to solve this problem is through the dos shell facility, the only way to view a file directory.

Placement

Placement is achieved by its own program, run either from the *P-Cad* shell or from the dos command line. The system works in the graphics editor environment, and with component footprints loaded, the user can move and fix the position of specific (or all) components. A grid is defined for the devices, the intersection of the grid lines being referred to as lattice points. Placement barriers can be defined to stop components being placed in particular areas.

During placement, horizontal vertical histograms can be displayed indicating the degree of congestion of tracks with that particular placement. Placement is aided through a complete series of commands, such as component alignment, gate swap to reduce track length and a suite of component move and rotate commands.

Track editing

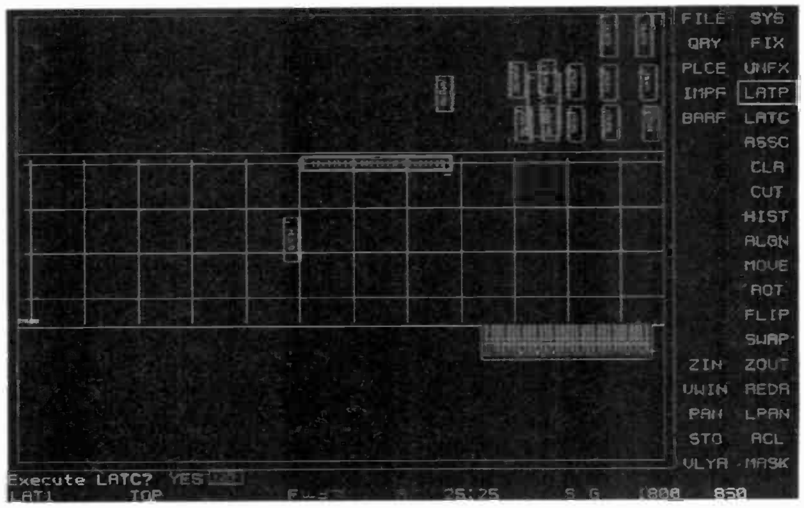
The schematic editor for PCB layout has the same user interface as the schematic and placement editors and operates with similar layered menus.

Full forward and back annotating transcribes schematic data to the PCB and PCB data back to the schematic so that consistency is ensured between circuit diagram and PCB.

Autoroute

P-Cad autorouter provides the usual memory and maze routing strategies, and includes a rip-up router suitable for more difficult routing tasks. Related facilities include via minimisation, and blind and buried vias (in multi-layer PCBs). Everything takes place on all layers on a 0.001in grid.

To neaten the final result and to help ensure that design rules are not broken, the router can work with a separate grid for the vias.



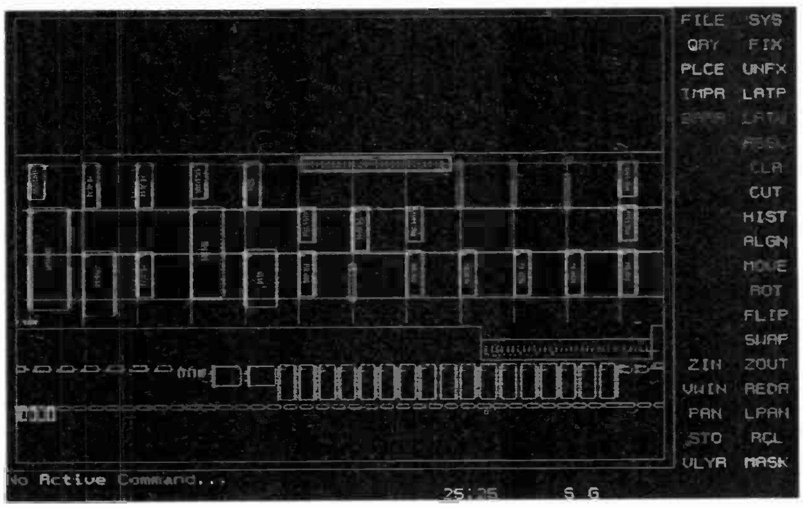
Components loaded and lattice defined in P-Cad.

Libraries

Comprehensive libraries cover most common components. For going beyond these, other libraries span the dominant technologies (TTL, cmos, ECL, linear etc) and major suppliers (National, Intel, Motorola etc). The package also reveals its insular US roots here, as libraries cover only the products of US corporations – an annoying weakness with more and more of the leading-edge electronic components coming from Japan.

But a nice touch is the library of standard PCB outlines that includes PC, Pcat, Apple, Multibus and several others.

Schematic library items can be added to the library, using the same graphics editor as for normal schematics.



Part-placed individual outlines can be interactively selected.

Installation and documentation

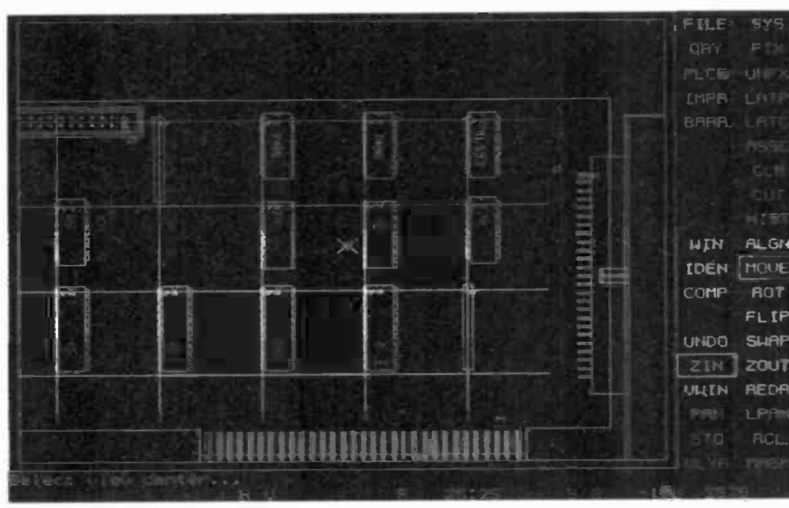
Installing P-Cad is not a good introduction to the package. The software uses a novel serial dongle system comprising a 250 x 80 x 40 box containing small plug-in keys for each section of the program. The box itself – which buzzes annoyingly – plugs into a separate mains outlet.

But the problem is that the dongle uses a 25-way D-shell for the serial interface without adapters for the 9-way serial connectors (introduced with the IBM AT). So two adapters are required.

Installation itself is fast and slick, the distribution discs being accepted in any order, and in about 15 minutes all was completed - all 20Mbytes of it! The promotional material suggests a 30Mbyte disc is required. But with this amount of software I would suggest 100Mbytes to avoid having dos struggle with fragmented clusters. Installation includes a fiddle with AUTOEXEC and CONFIG files, but the software tells you exactly what it is doing.

Documentation makes War and Peace look like a short story. No less than eleven 1 8 x 9in paperback books, with probably over 5000 pages, stack up to make a thickness of nearly 8in – more than a little daunting. But the software turns out to be surprisingly easy to use, and the great stack of books is actually very useful for learning and reference (and filling up your book shelves impressively).

The manuals are professionally published and well organised so that even with so much data, references can be found without too much difficulty.

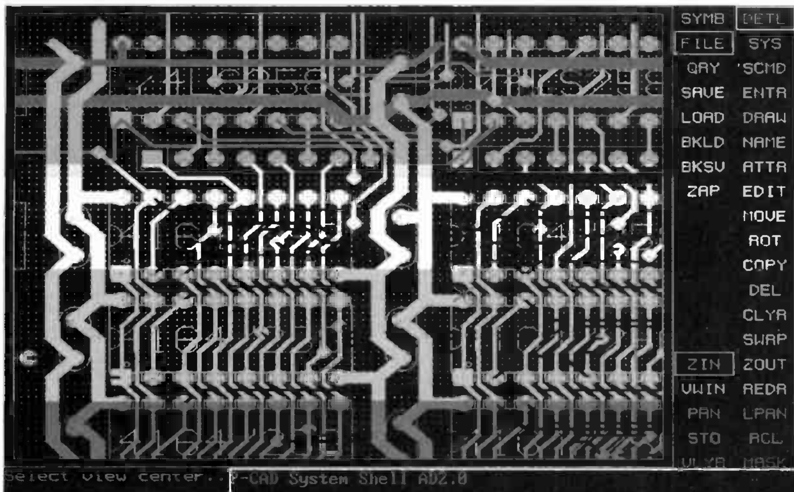


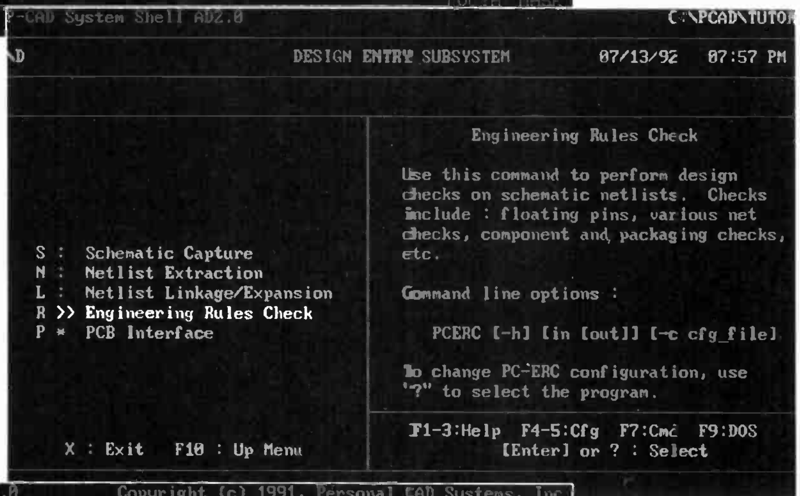
Final result of auto-placement with P-Cad

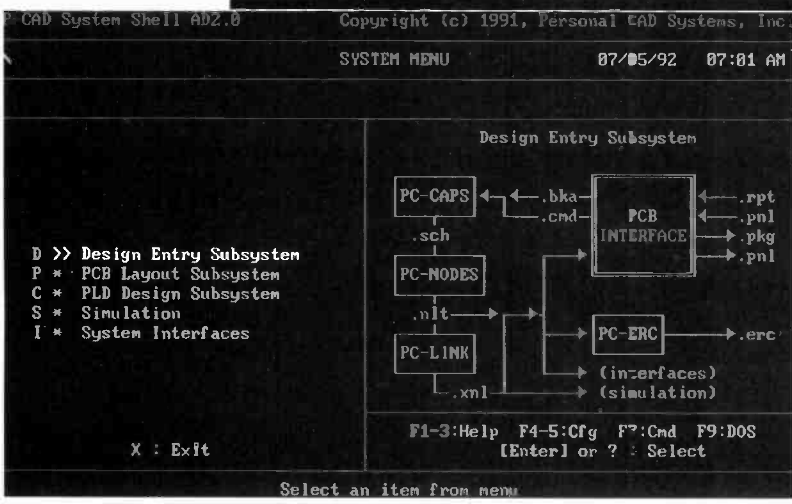
SUPPLIER DETAILS

P-Cad Associate designer AD-4 £2000 is written by Cadam Inc, Burbeck, California. It is available in the UK from KGB Micros Ltd. 162 Bestobell Road, Slough, Berks SL1 4SZ. Tel: (0753) 696069

Other products in the range:
AD-1 entry level similar to the *AD-4* but excluding auto-place and route £1500.
Master Designer MD-1 similar to *AD-4*, but with more features and larger databases £6500.
Professional PCB for Unix on Sun Spark or IBM 6500 £9980.







P-Cad's system menu – the shell giving access to most aspects of the package.

Schematic capture has a neat graphical editor.

Routing editor is very similar to the schematic editor. Note the complex autorouting.

SYSTEM REQUIREMENTS
XT or AT
640K
Additional 1m memory recommended
Dos 3.3
30Mbyte hard disc
25-way RS 232 port for dongle
Colour display and adapter (EGA, VGA, 8514 and superVGA)
Mouse or digitiser tablet

Little bits extra

Several export facilities are available – indeed one of the manuals is dedicated to export. Amongst the formats are DXF and IGES as well as the usual plotter, photoplotter and NC drill. But *P-Cad* always seems to offer that much more, and included in the exports are panelisation (grouping PCBs into panelled sets), and control output for pick and place and auto insertion manufacturing machines.

At first it looked as though the package was going to be especially interesting, with menu features detailing data capture and compilation for programmable logic devices (PLDs). But on closer inspection, this part of the menu system was inoperative, and the package needed to be upgraded for those functions.

There is also a simulation menu option – again this proved to be inoperative on my review software – and the sales literature indicates there is a logic simulator and a file export to *Spice* analogue simulation.

Importance of training

As expected, the product from an IBM company is good – though the price compared with the competition is certainly a premium.

In terms of facilities *P-Cad* seems to have the lot. Not surprisingly the wide range of basic commands, option settings and configurations can make this one of the most daunting software products to learn and use: the 11 reference manuals are evidence of the enormous volume of material.

No one can be expected to memorise this amount of data and put it together in a structured way to make the system work first time. As a result, to obtain full value from the product software training has to be a serious consideration.

KGB, the UK distributors, offer training for schematic capture (two days), interactive layout (two days) and auto-placement and routing (three days); each day costs £350 and should be costed into the price of the final product.

So *P-Cad* is undeniably a powerful product, taking the design from a hierarchical schematic capture to the control programs for manufacture pick and place. But potential users should think very carefully about whether they are up to using a package having this level of complexity and no on-line help. ■

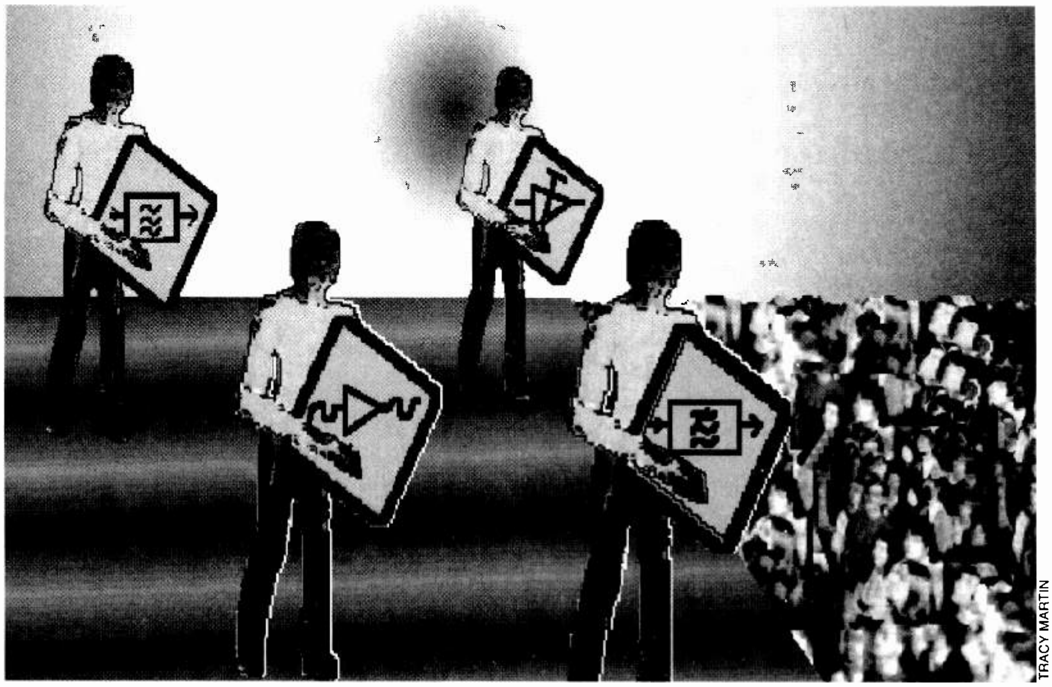
SPECIFICATIONS

Schematic and PCB editors:
100 layers
Cut and paste editor
On-line connectivity checking
User defined macros
6000 standard symbol and footprint library
1300 component, 2500 net and 32000 pin capability
Up to 300 schematic sheets in one design

Schematic capture:
Back annotation
Real time component drag with orthogonal rubber-banding

PCB layout:
0.001in or 10µm resolution
PCBs to 60 x 60 in
SMD support includes both sides of PCB

Autorouter:
Four routing algorithms
Up to 32 routing layers
Route by window via minimisation



Behavioural modelling wrapped up in a black-box

Bashir Al-Hashimi explains how behavioural models of analogue circuits are developed and uses an AM modulator/demodulator system to illustrate some of the models available.

One of the advantages of analogue behavioural modelling (ABM) is the ability to model analogue circuit functions using equations, tables and transfer functions. Designers can use the technique to simulate complex systems as a combination of "black-boxes" each of which performs a specific function.

In this way performance can be predicted before circuits are be built, a detailed component design of a system being described to the simulator using a basic set of components such as resistors, transistors and various voltage and current sources. Component connections are then expressed in terms of nodes – a type of simulation often called the *structure* or *primitive level* simulation.

In a simple example, take as a system under simulation a 2nd-order lowpass filter. To optimise performance, the effect of changing the filter type (Butterworth, Bessel,...) on system response needs to be examined.

Until recently, this meant a number of different type filters would have to be designed and simulated – a time consuming task. A more effective way to tackle the problem is to simulate the filter as a "black-box", where the input/output relationship can be expressed in terms of an equation. As a result filter component design is unnecessary at this stage, since the filter function has been modelled using a mathematical expression. At a later stage the model can be replaced by actual circuitry.

In this "behavioural level" simulation, a functional block can be described by its behaviour without worrying about its physical structure. The description can be an equation or data table, and a user has the ability to check and optimise a system without needing to perform circuit design. Complex systems can also be simulated quickly and more efficiently than at the primitive simulation level. Several simulators offer analogue behavioural modelling ABM)[1,] but here we will look in detail at *Pspice*[2]. (For a good introduction to *Pspice* see "Adding Spice To Technology"[3].)

Pspice ABM

ABM comes as an extra option to the basic *Pspice* simulator, allowing the user to describe analogue component or circuit operations using equations and tables. Descriptions are implemented using four functions: Laplace, frequency, table and value.

Laplace is usually used to describe the frequency response of circuits in terms of a transfer function in the Laplace transform variable. Frequency response of a circuit can also be written as a table using the frequency function, where each entry in the table consists of a frequency and the magnitude and phase of the response at that frequency. The table function allows circuit operation to be described by a look-up table, with the table consisting of input-output data pairs.

Value handles voltage or current sources whose output is any arithmetic function of voltages and currents elsewhere in the circuit.

Pspice performs ABM using two controlled sources: VCVS (voltage controlled voltage source) and VCCS (voltage controlled current source), identified by *Pspice* as a component starting with the letter E or G respectively. The E component will be used when an output voltage is required, the G component when output current is needed

Circuit modelling

In this section, a number of commonly used analogue circuits will be modelled as "black-boxes". These include filters, amplifiers and rectifiers. As indicated earlier, a 2nd-order lowpass filter provides a simple example to illustrate the development of a behavioural model.

The voltage transfer function of this filter is:

$$(V_{out}/V_{in})= A_1/(s^2+sA_2+A_1) \quad (1)$$

where s is the Laplace transform variable, A_1 and A_2 are expressions describing the filter characteristics and are given by:

$$A_1= 4\pi^2\omega^2c;$$
$$A_2= (2\pi\omega_o F_c)/Q \quad (2)$$

ω_o is the normalised frequency (rad/s), Q is the quality factor and F_c is the denormalised cut-off frequency of the filter (see box, "Laplace transform and s domain"). The filter transfer function has been written in terms of s, therefore, the Laplace function will be used to model the filter. Also, since the output of the filter is a voltage, the E component will be used.

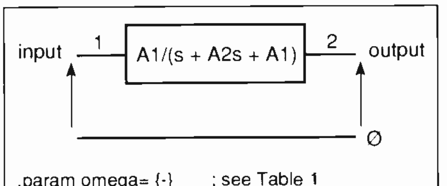
So Eq(1) is described in *Pspice* as:

E20Laplace{V(1)}={A1/(s*s+A2*s+A1)} (3)

where the input to the filter is a voltage at node 1, and the output is a voltage at node 2. Nodes have been chosen arbitrarily to show how the "black-box" filter model would be connected within a system.

The 2nd-order lowpass behavioural filter *Pspice* model is (black-box notation, shown in **Fig. 1**):

The simple nature of the transfer function allows changes in the model to be made easily. For example, changing the filter type simply involves entering the appropriate values of the parameters ω_0 and Q selected from Table 1 into Eq. (3). *Pspice* allows parameters to be declared and used in expressions within an input file, giving a great deal of flexibility when developing and modifying circuit models. In the model, note how the filter parameters are defined using the .PARAM command. This illustration demonstrates how only the transfer function in s, allows a model to be developed for a lowpass filter. Generalised models for filters are possible such as lowpass, highpass or bandpass[4], further simplifying the operation. The technique can be extended to other circuit functions where only the Laplace transfer function needs to be specified. Frequency response of our lowpass filter could have also been modelled using the "frequency" function (see box, "Examples of ABM circuit models", which also considers the modelling of amplifiers and rectifiers).




```
.param omega= {-}          ; see Table 1
.param Q= {-}              ; see Table 1
.param Fc= {-}             ; -3dB frequency point
.param pi= {3.142          ; constant
.param pi_square= {pi*pi              ; constant
.param A1= {4*pi-square*omega*omega*Fc*Fc
                                      ; see Eq (2)
.param A2= {2*pi*Fc*omega/Q}          ; see Eq (2)
E 2 0 Laplace {V(1)}= {A1/ (s*s+A2*s+A1)}
                                      ; see Eq (1)
```

Fig. 1. Black-box filter notation for a behavioural model and its equivalent Pspice code.

Laplace transform and s domain

The relationship between the input and output of linear systems can be expressed by a series of differential equations. Obtaining the system response usually means solving the differential equations – which can be tedious. A mathematical method, the Laplace transform provides an easier way of solving these equations.

For example, the Laplace transfer function of the RLC circuit shown in the figure is derived as follows. Input and output voltages of this circuit are related by the differential equation,

$$V_{in}= RI+ L\, dI/dt + V_{out}$$

using Kirchoff's voltage law. Using the Laplace transform method, d/dt= s, where s is the Laplace transform variable. The equation can then be expressed as

$$V_{in}= I(R + sL) + V_{out} \quad \text{(i)}$$

Current through the capacitor is given by

$$I= C dV_{out}/dt= sCV_{out} \quad \text{(ii)}$$

Substituting Eq (ii) into Eq (i) and simplifying yields

$$(V_{out}/V_{in})= (1/LC)/(s^2+ s(R/L)+ 1/LC) \quad \text{(iii)}$$

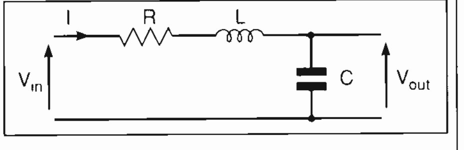
describing the Laplace voltage transfer function of the 2nd-order lowpass filter given in the figure. It is of the form of Eq (1) in the text.

Usually, ABM requires that ideal 2nd-orders systems are described in terms of ω_o and Q. To relate these parameters to the circuit components, Eq (iii) will be compared with the general lowpass Laplace transfer function,

$$T(s)=\omega_o^2/(s^2 + s(\omega_o/Q) +\omega_o^2)$$

giving

$\omega_o^2= 1/(LC)$ and
$Q= \omega_o(L/R)$



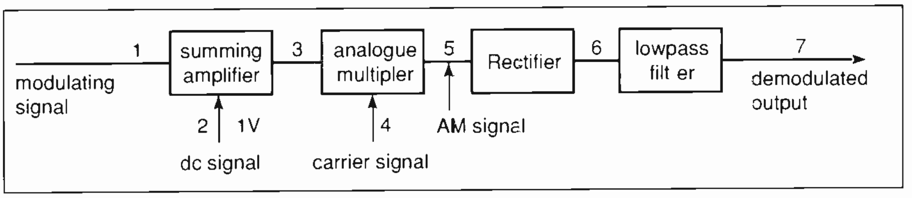


Fig. 2. Behavioural model example using an AM transmission system. The summing amplifier and analogue multiplier represent an amplitude modulated full carrier transmitter while the rectifier and lowpass blocks represent the receiving system.

Table 1. ω_o, Q of the various filter responses.

	ω_o	Q
Butterworth	1	0.707
Chebyshev (0.5dB)	1.231	0.864
1dB	1.05	0.957
2dB	0.907	1.129
3dB	0.841	1.307
Bessel	1.73	0.577

Behavioural level

Some of the behavioural circuit models developed earlier (and shown in "Examples of ABM Circuit Models") can be used to simulate an AM modulator and demodulator. These circuits have been selected because *Pspice* does not provide AM modulation directly – they also strike a reasonable balance of complexity to illustrate the usefulness of ABM while still allowing a comparison with the primitive simulation level.

The AM system (**Fig. 2**) consists of a summing amplifier and analogue multiplier – describing the modulator – and a rectifier and lowpass filter which represents the demodulator.

An AM modulated signal is described by:

$$V_{am} = A\sin\omega_c t\,(1 + m\cos\omega_m t) \qquad (4)$$

where ω_m and ω_c are the angular frequency of the modulating and carrier signals respectively, A is the amplitude of the carrier and m is the modulation index. Examination of the equation suggests that an AM signal can be modelled in two stages using ABM.

In stage one, the modulating signal ($m\cos\omega_m t$) is added to a fixed DC voltage of 1V to obtain ($1 + m\cos\omega_m t$). In the 2nd stage, the result of this addition will be multiplied by the carrier ($A\sin\omega_c t$). Both of these operations are easily achieved using the value function of the ABM.

The *Pspice* input file of the AM system is:

```
*This file simulates an AM system    ; comment
Vmod 1 0 sin (0 0.5v 1K 0 0 90)      ; modulating
                                       signal
Vdc 2 0 dc 1v   ; dc inputVcar 4 0 sin (0 lv l00K)
                                     ; carrier signal
Esum 3 0 Value= (V(1)+V(2)}          ; summing ampf
Emult 5 0 Value= (V(3)*V(4)}         ; AM signal
Erect 6 0 Table (V(5)}=
(-1.5,0) (0,0) (1.5,1.5)
                                     ; rectified signal
*2nd-order filter parameters
*.param omega= {1}     ; Butterworth filter
.param Q= (0.707)      ; Butterworth filter
.param Fc= {1K}        ; F3dB point
.param pi= {3.142}     ; Constant
.param pi square= ~pi*pi)            ; Constant
.param A1- {4*pi_square*omega*omega*Fc*Fc)
                                     ; see Eq (2)
.param A2= {2*pi*Fc*omega/Q}         ; see Eq (2)
Elow 7 0 Laplace {V(6)}= {(A1)/(s*s+ A2*s+A1)}
                                     ; filter model
*
* These resistors all needed to satisfy Pspice require-
*ment for a minimum of 2 nodes to every component.
R1 1 0 1G
R2 2 0 1G
R3 3 0 1G
R4 4 0 1G
RS 5 0 1G
R6 6 0 1G
R7 7 0 1G
* TRANSIENT ANALYSIS
.tran 0.2m 4m 0m 0.2m; time analysis range
.options itl5=10000    ; number of analysis points
.probe
.end
```

The modulating signal, DC voltage and carrier signal have been defined using various independent voltage sources as shown at the start of the above *Pspice* input file.

Simulating the demodulator involves modelling the operations of the rectifier and the lowpass filter. Both of these circuits have already been modelled.

In the case of the lowpass filter, the Laplace transfer function model will be chosen in preference to the frequency response table model (see box, "Examples of ABM circuit models"), because it offers greater flexibility. The filter has been assumed to be of the Butterworth type with –3dB frequency point at 1kHz. In the *Pspice* simulation of the AM signal and its frequency spectrum (**Figs. 4** and 5), *Pspice* correctly models the carrier at 100kHz and the two sidebands at 99kHz and 101kHz respectively. **Figure 6** shows the demodulated signal at the filter output which corresponds to the input signal.

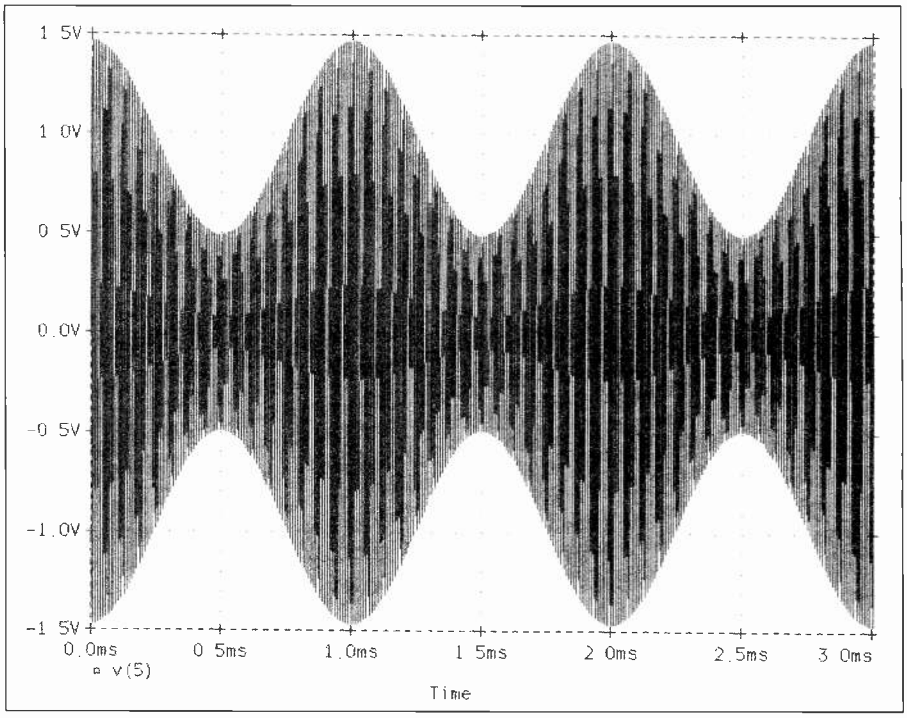


Fig. 4. Simulated AM signal provided by the first two behavioural blocks of Fig. 2.

ABM circuit models

This box includes the modelling of filters, amplifiers and rectifiers, which form the basis of an AM modulator/demodulator system shown in Fig. 2. Here the 2nd-order lowpass filter will be modelled using an alternative function to the "Laplace" function illustrated in the main text. Frequency response of the circuit is written as a table using the "Frequency " function. Each entry in the table consists of a frequency, magnitude (dB), and phase (°).

For example, a normalised 2nd-order Butterworth filter is modelled as:

E 2 0 Frequency {v(1)}=(0.1,0,-16)(0.5,0,-44)(0.75,-0.5,-68)

(1,-3,-90) (1.25,-5,-108)(2,-12,-137)

(3,-19,-151)(4,-24,-160)

To scale this filter to a specific cut-off frequency, all the frequencies in this table must be multiplied by the required cut-off frequency.

Also, if a different filter type is required, all the magnitude values need to be changed according to the chosen filter type. So the functions "Laplace" and "Frequency" can both be used to model frequency response, but clearly as can been seen from the filter example, the "Laplace" function offers more flexibility with respect to circuit modifications through at the expense of developing a transfer function.

In many applications, circuit operations must be modelled by arithmetic equations, for example a summing amplifier. Using the VALUE function, the amplifier is simply modelled as:

E 3 0 Value= {V(1)+ V(2)}

describing a summing amplifier with an output voltage at node three. The value of this voltage is equal to the summation of the voltages at nodes one and two. Some analogue circuit operations are best described by a look-up table. In this case, the function TABLE offers the best option for modelling, because it allows unrestricted pairings of parameters (unlike FREQUENCY which is always frequency , magnitude, and phase). An example is the half-wave rectifier modelled as:

E 2 0 Table {v(1)}= (-1,0) (0,0) (1,1)

The table has three entries, describing the full operation of the rectifier. The first value of each bracket is the input; the second value is the corresponding output. So in this example, the amplitude of the input signal at node one varies from –1V to +1V. Output at node two is 0V for inputs ≤0V, 1V for inputs ≥1V and input=output in between.

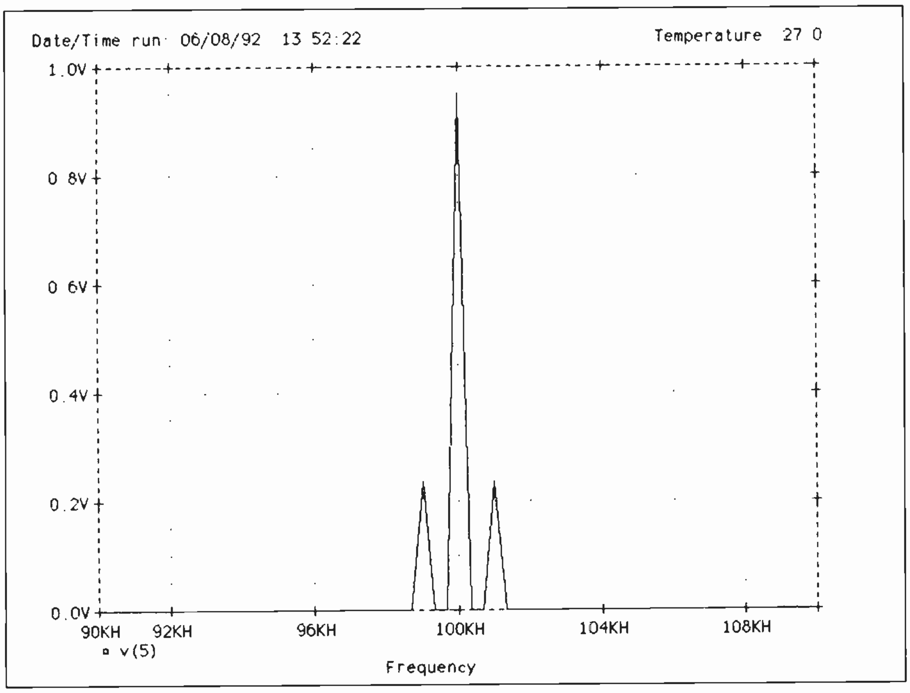


Fig. 5. Simulated AM frequency spectrum. Pspice correctly models the carrier at 100kHz and the two sidebands at 99kHz and 101kHz respectively

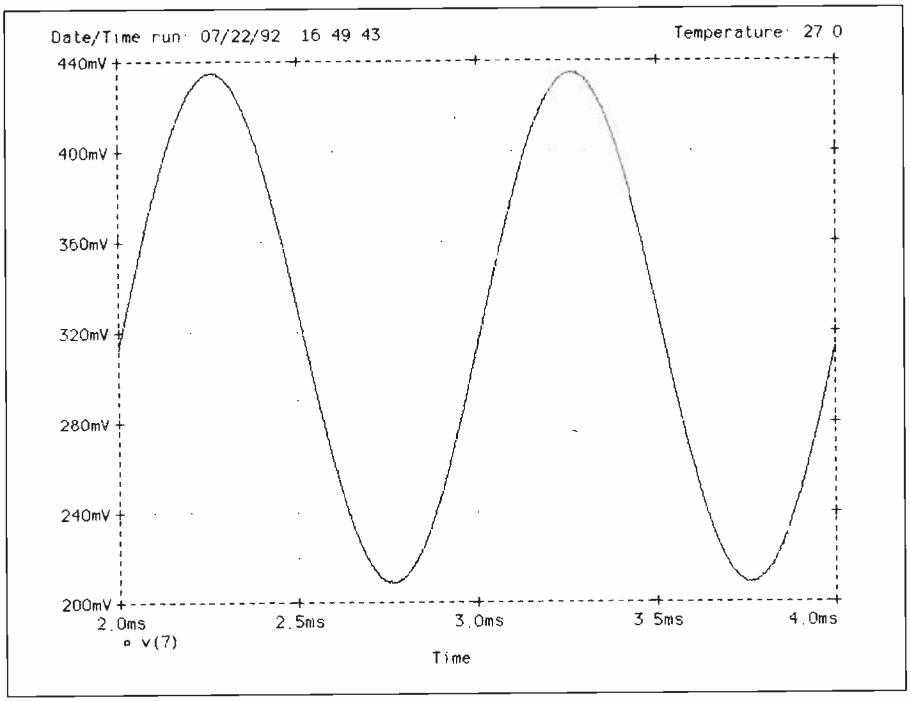


Fig. 6. Simulated demodulated signal, corresponding to the input signal.

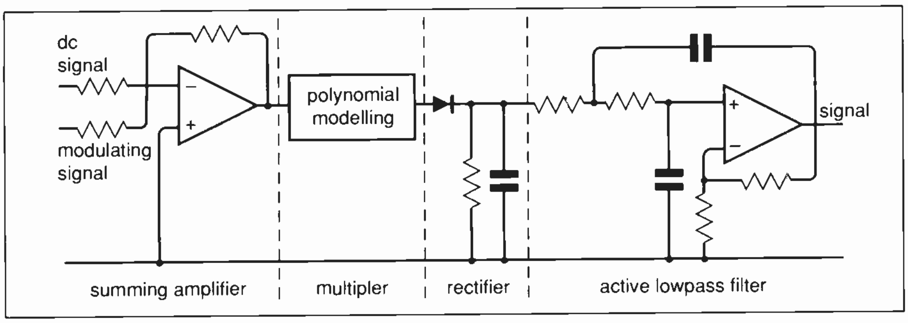


Fig. 7. Primitive level circuit of the AM system shown in Fig. 2.

Table 2. Comparison of the two simulation approaches.

	Behavioural	Primitive
Development	0.5day	1 day
Simulation	175s	220s

Simulation times were measured on PC 386SX.

Behavioural vs primitive

As a comparison, the same AM system was simulated using primitive level components, **Fig. 7**. The amplifier was simulated using a VCVS, in other words modelled using an equivalent circuit. There is no equivalent circuit available for an analogue multiplier. An ideal analogue multiplier could be modelled by a discrete transistor circuit[5] or by using polynomial sources. Polynomial sources are an accepted technique for use in the basic *Pspice* without ABM and offer a more convenient method for simulation of ideal circuits than actual circuit design (see *Pspice* manual for more information).

The rectifier was simulated using a combination of a diode, a resistor and a capacitor, and the filter was modelled using a VCVS, four resistors and two capacitors. The filter was assumed to be implemented using an active Sallen-Key circuit.

Both behavioural and primitive approaches have been shown to yield similar and satisfactory results. But there is a saving in development and simulation time associated with the behavioural approach (**Table 2**). Clearly optimisation of system performance is most effectively achieved by the use of ABM.

Finally, though this article has mainly dealt with modelling ideal circuits, ABM can be used to model practical components and circuits. Examples of these are given in the *Pspice* manual and MicroSim newsletter magazines: for instance, April 1991 issue dealt with modelling of a lossy transmission line. ■

Acknowledgment

Thanks to Alan Holden for his help in preparation of this article.

References

1. Swager, A,"Behavioural Models Expedite Simulation", *EDN*, Nov 21, 1991. 2. *Pspice*, Microsim Corporation, 20 Fairbanks, Irvine, California 92718.
3. Stewart, P, "Adding Spice To Technology", *EW + EE*, Jan 1992.
4. Al-Hashimi, B, "Pspice Behavioural Model of a Universal Filter", submitted for publication in EDN.
5. Gilbert, B., "A high Performance Monolithic Multiplier Using Active Feedback", IEEE Journal of Solid-states, Vol, SC-9, No.6, pp364-370, Dec 1974.

Circuits, Systems & Standards

First published in the US magazine EDN and edited here by Ian Hickman

Understand capacitor soakage to optimise analogue systems

Dark secrets of capacitors

This item from Bob Pease is essential reading for anyone designing low level analogue circuits where low frequency/DC performance is important. The big surprise for me was how poorly silver mica capacitors – considered by many to be one of the most stable and reliable types – performed in the soakage tests.

IH

Veteran circuit designers often had a shocking introduction to dielectric absorption when supposedly-discharged high-voltage oil-filled paper capacitors reached out and bit them. Indeed, the old oil-filled paper capacitors were notorious for what was once called soakage – a capacitor's propensity to regain some charge after removal of a momentary short. Today, few of these capacitors are still in use, but soakage is still a problem. How do you deal with it?

Nowadays, effects of dielectric absorption are likely to be noticed in more subtle ways: perhaps in the performance of an integrator that can not be reset to zero; or a sample/hold that refuses to work correctly. But whether effects are (literally) felt or merely observed in a circuit's behavior, dielectric absorption is a characteristic that every capacitor possesses. It is inherent in the dielectric material itself – though poor manufacturing or inferior foil electrodes can contribute to the problem.

Apt description

Soakage seems an apt term for dielectric absorption – considering what the capacitor seems to be doing. In a typical example a capacitor charges to 10V for a long time T and then discharges through a small-value resistor for a short time . If the short circuit is removed and the capacitor terminals monitored with a high-impedance voltmeter, the capacitor will be seen to charge back to 0.1%, 1% or as much as 10% of the original voltage. For example, a 1μF Mylar capacitor charged to 10V for 60s (T_{CHARGE}) and discharged for 6s ($t_{DISCHARGE}$) charges to 20 or 30mV after 1 min (T_{HOLD}). **Figure 1** shows a simple evaluation circuit for measuring the characteristic.

A capacitor exhibiting dielectric absorption acts as if during its long precharge time the dielectric material has soaked up some charge that remains in the dielectric during the brief discharge period. Charge then bleeds back out of the dielectric during the relaxation period and causes a voltage to appear at the capacitor terminals.

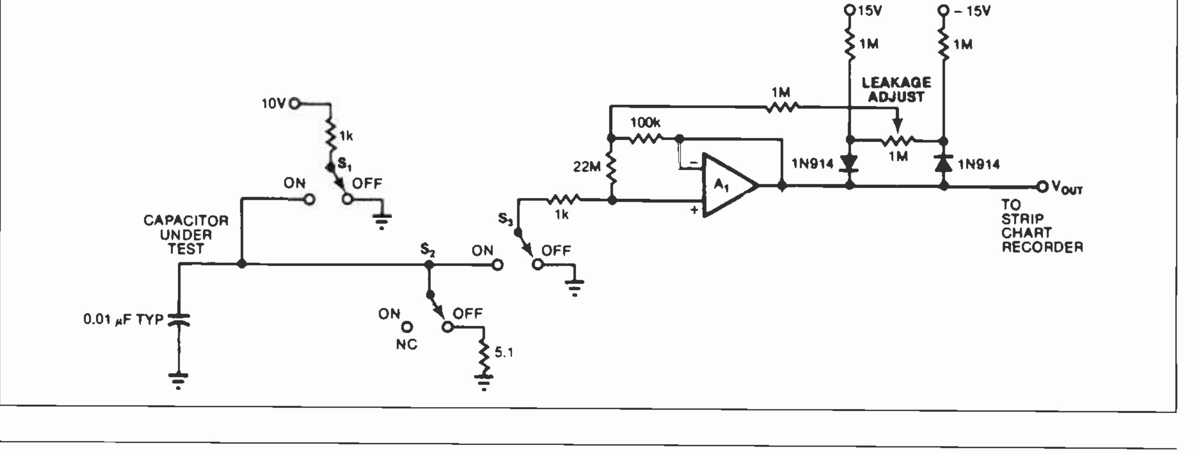


Fig. 1. A simple test fixture allows evaluation of dielectric absorption at low speeds. To use, start with all switches off and throw S1 and S2 on for 1min; then throw S1 and S2 off and wait 6s, throwing S3 on during the wait period. Next turn S2 on and watch V_{OUT} for 1min. To compensate for leakage leave all switches off for 1min and then thow S2 and S3 on. Monitor V_{OUT} for 1min and subtract this value from the V_{OUT} value obtained earlier.

Figure 2 depicts a simple model of this capacitor. When 10V is applied for 1min, the 0.006μF capacitor charges almost completely. But during a 6s discharge period it only partially discharges. Then, over the next minute, the charge flows back out of the 0.006μF and charges the 1μF capacitor to a couple of dozen millivolts.

The example indicates that a longer discharging time reduces soakage error but that discharging for only a small fraction of that time results in a larger error. Illustrating the point, **Fig. 3** shows the results of conducting the basic Fig. 1 test sequence for 1s, 6s and 12s discharge times. Note that the capacitor tries to remember its old voltage, but the longer it is held at its new voltage, the more effectively it forgets – in the Fig. 3 case, soakage errors equal 31mV at $t_{DISCHARGE}$=1s, 20mV at $t_{DISCHARGE}$=6S and 14mV at $t_{DISCHARGE}$=12s.

Do these low-speed tests have any bearing on a capacitor's suitability in fast millisecond or microsecond sample/hold applications? If the Fig. 1 experiment is repeated for T_{CHARGE}=T_{HOLD}=1000μs and $t_{DISCHARGE}$=10μs, very similar capacitor-voltage waveforms are seen but with about ten times smaller amplitudes. In fact, for a constant T:t ratio, the resulting soakage error decreases only slightly in tests ranging in length from minutes to microseconds.

Figure 4 circuit approximates the capacitor characteristic, which can be observed on actual capacitors by using test setup shown in **Fig. 5** Here, a sample/hold IC exercises the capacitor under test at various speeds and duty cycles, and a limiter amplifier facilitates close study of the small residual waveforms, without overdriving the oscilloscope when the capacitor is charged to full voltage.

Such experiments illustrate that if a certain amount of charge is put into a less-than-ideal capacitor, a different amount of charge will result, depending on the wait time. Thus, using low-soakage capacitors proves important in applications such as those involving high-resolution dual-slope integrating A-to-Ds. Sure enough, many top-of-the-line digital voltmeters do use polypropylene (a low-soakage dielectric) devices for their main integrating capacitors.

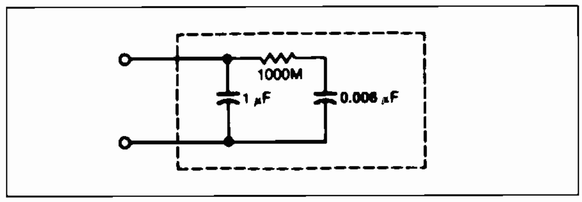


Fig. 2. To model the soakage characteristics of a 1μF Mylar capacitor, consider a circuit that incorporates a 0.006μF capacitor to represent the dielectric's charge-storage

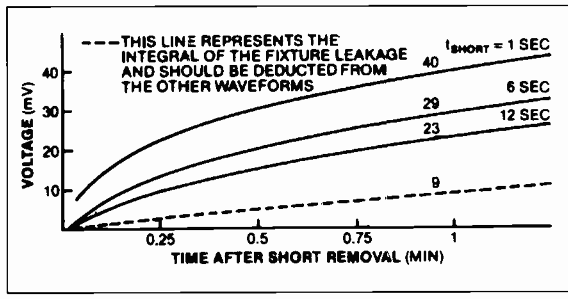


Fig. 3. Obtained using Fig. 1's test circuit, these dielectric absorption-measurement results for a 1μF capacitor show that longer $t_{DISCHARGE}$ times reduce soaking-caused errors.

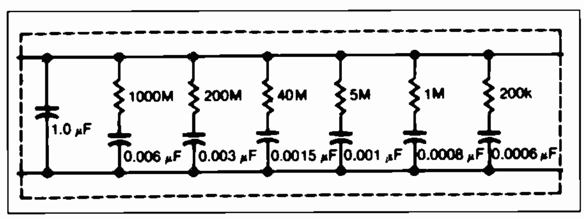
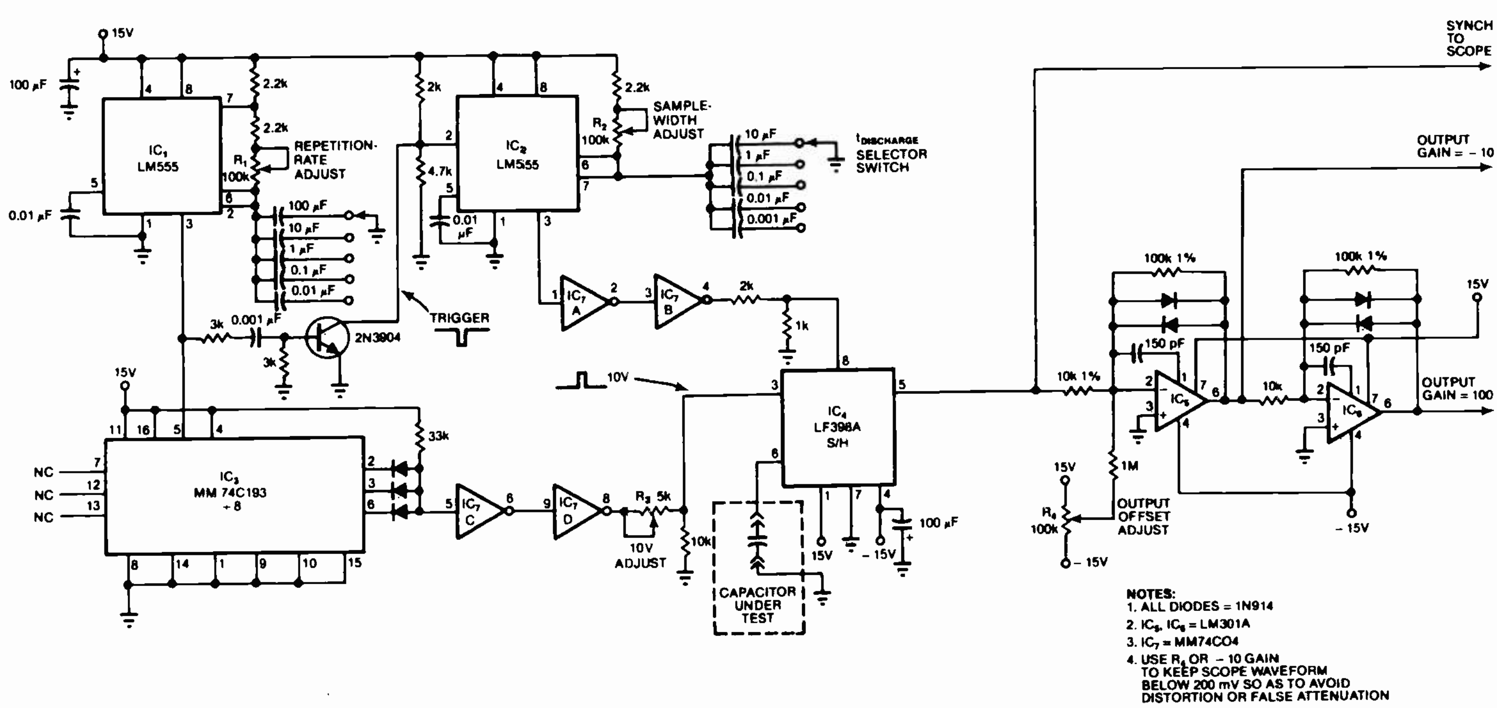


Fig. 4. More precise than Fig. 2's equivalent circuit, a capacitor model emplying several time constants proves valid for a wide range of charge and discharge times. This model approximates a Mylar capacitor.

Fig. 5. Capable of automatic sequencing of the dielectric-absorption tests, a circuit employing sample-hold and limiting stages allows measurements to be made for a wide range of T_{CHARGE}, T_{HOLD} and $T_{DISCHARGE}$ values. Figure 7 shows results obtained using this circuit.



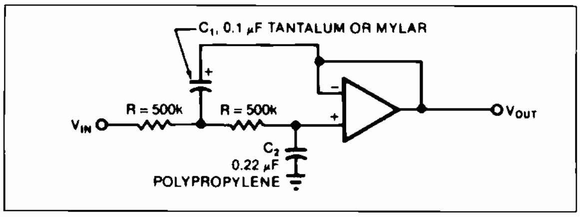


Fig. 6. Soakage can present problems when designing a fast-settling amplifier or filter. C1 can be a Mylar or tantalum unit, but making C2 a polypropylene device improves performance.

But dielectric-absorption characteristics are most obviously detrimental in applications involving sample/holds. Manufacturers guarantee how fast these devices can charge a capacitor in SAMPLE mode and how much their circuits' leakages causes capacitor-voltage droop during the HOLD mode. But they do not give any warning about how much the capacitor voltage changes because of soakage – a factor especially important in a data-acquisition system, where some channels might handle small voltages while others operate near full scale. Even with a good dielectric, a sample/hold can hurt accuracy, especially if the sample time is a small fraction of T_{HOLD}.

For example, although a good polypropylene device can have only 1mV hysteresis per 10V step if T/t=100ms/10ms, this figure increases to mV if the T/t ratio equals 100ms/0.6ms. Because most sample/hold data sheets do not warn of such factors, capacitors should be evaluated in a circuit(such as in Fig. 5) using time scaling suited to the application.

Other applications in which soakage can degrade performance are those involving fast-settling AC active filters or AC-coupled amplifiers. In the circuit shown in **Fig. 6**, C_1 can be a Mylar or tantalum unit because it always has 0V DC on it. But making C_2 polypropylene instead of Mylar noticeably improves settling. For example, settling to within ±0.2mV for a 10V step improves from 10 to 1.6s with the elimination of Mylar's dielectric absorption. Similarly, voltage-to-frequency converters benefit from low-soakage timing capacitors, which improve V/F linearity.

Dielectrics good at all speeds

Fortunately, good capacitors such as those employing polystyrene, polypropylene, NP0 ceramic and Teflon dielectrics perform well at all speeds. **Figure 7** shows the characteristics of capacitors using these dielectrics and others such as silver mica and Mylar. In general, polystyrene, polypropylene or NP0 ceramic capacitors furnish good performance, although polystyrene can not be used at temperatures greater than 80°C. Although NP0 ceramic capacitors are exxpensive and hard to find in values much larger than 0.01μF, they do achieve a low

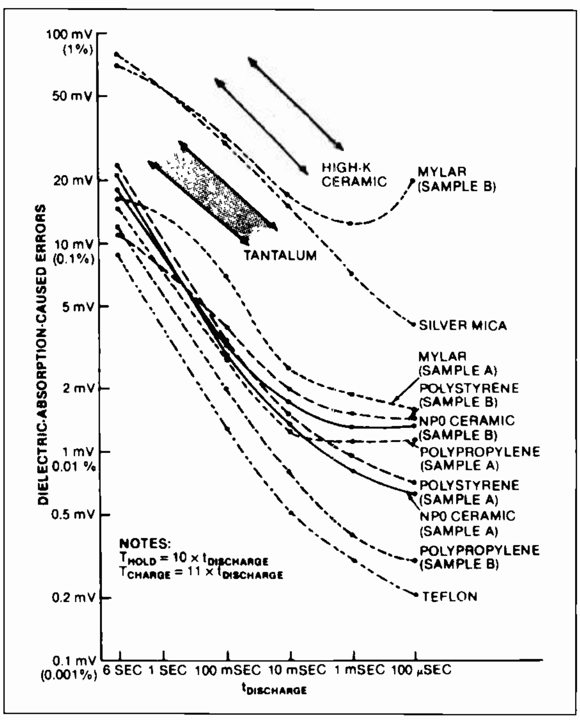


Fig. 7. Soakage-measurement results for a variety of capacitors illustrate the effects of $t_{DISCHARGE}$ values on dielectric-absorption-caused errors. Note that the curves for two different samples of NP0 ceramic capcitors intersect.

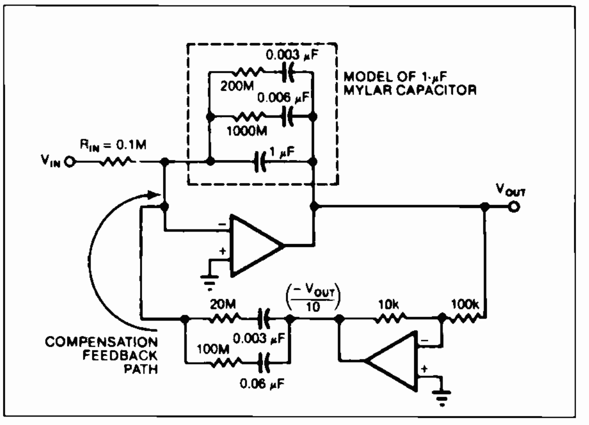


Fig. 8. An integrator can be compensated for dielectric absorption by feeding its inverted output back to the input through one or more experimentally chosen RC networks; they cancel the equivalent network inherent in the capacitor's dielectric material.

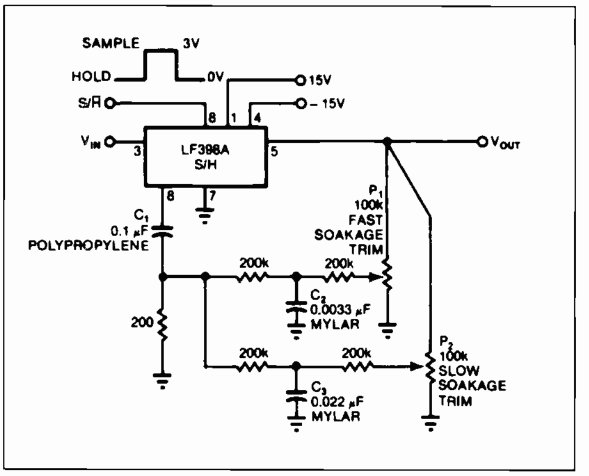


Fig. 9. Adding compensation circuitry to a sample/hold yields better-than-Teflon performance with a polypropylene capacitor. Using Teflon capacitors in such circuits can yield a 15-17 bit dynamic range.

temperature coefficient (a spec not usually significant for a S/H but one that might prove advantageous for precision integrators or voltage-to-frequency converters). Teflon is rather expensive but definitely the best material to use when high performance is important. Furthermore, only Teflon and NP0 ceramic capacitors suit use at 125°C.

If dielectric-absorption values in Fig 7 are studied, wide differences in performance for a given dielectric material can be seen.

For example, polypropylene *sample A* is about as good as *B* at t=6s, but *B* is four times better at high speeds.

Similarly, NP0-ceramic *sample A* is slightly worse than NP0-ceramic *sample B* at low speeds, but *A* is definitely better at high speeds.

Some Mylar capacitors (*sample A*) get better as speed increases from 1000 to 100μs, but others (*sample B*) get

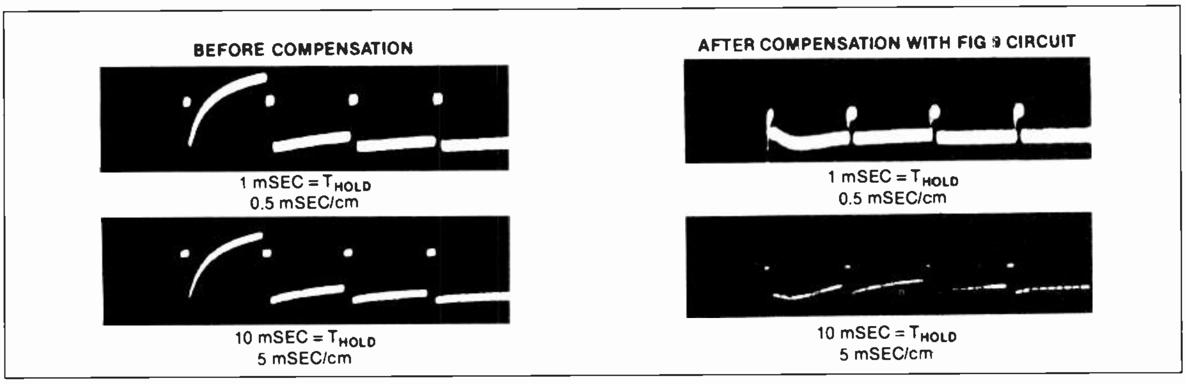


Fig. 10. Adding Fig. 9's compensation network to a sample hold circuit yields a ten-fold performance improvement for sample times of 50-2000µs; additional RC networks and trimming pots can extend the time range. The short pulses represent normal S/H jumps and occur during the sample time. The exponentially rising waveform during the hold time results from soakage. Note that soakage effects are still visible during the second hold period.

worse. So if consistently good performance from capacitors is required, they should be evaluated and specified for the speed at which they will be used in a particular application.

Keep in mind that because most sample/holds are used at much faster speeds than those corresponding to the 1min or 5min ratings usually given in data sheets, a published specification for dielectric absorption has limited value.

In addition, other dielectrics furnish various levels of performance. The following questions and answers reveal the differences.

● Any long word that starts with poly seems to have good dielectric properties. So how about polycarbonate or polysulfone?
No – they are about as bad as Mylar.

● Does an air or vacuum capacitor have low soakage?
It might, but many standard capacitors of this type are old designs with ceramic spacers, and they might give poor results because of the ceramic's hysteresis.

● If a ceramic capacitor is not an NP0 device, is it any good?
Most of the conventional high-K ceramics are just terrible—20 to 1000 times worse than NP0-and even worse than tantalum.

● Is silicon dioxide suitable for small capacitances?
Although Fig. 5 test setup, used in preparing Fig. 7 chart, only measures moderate capacitances (500 to 200,000 pF), silicon dioxide appears suitable for the small capacitors needed for fast S/Hs or deglitchers.

Cancellation improves accuracy

A practical method of getting good performance with less-than-perfect capacitors is to use a soakage cancellation circuit such as one of the form shown in **Fig. 8**, in which a capacitor of the type modelled in Fig. 4 serves as an integrator. (Only the first two soakage elements are shown.) The integrator's output is inverted with a scale factor of –0.1, and this voltage is then fed through one or more experimentally chosen *RC* networks to cancel the equivalent network inherent in the capacitor's dielectric material.

Figure 9 shows a practical sample/hold circuit with an easily trimmed compensator. The network provides about a ten-fold improvement for sample times in the 50 to 2000µs range (**Fig. 10**). Although the compensation is subject to limitations at very fast or slow speeds, the number of *RC* sections and trimming pots employed can be extended.

Simple circuits similar to Fig. 9 or Fig. 8 have been used in production to let inexpensive polypropylene capacitors provide better-than-Teflon performance. In turn, using these compensator circuits with a good Teflon capacitor furnishes a dynamic range of 16 to 17 bits.

Robert A Pease, *National Semiconductor Corp.*

BARGAINS – 10 New Ones This Month

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12/24V DC SOLENOID constructed so that it will push or pull, plunger is a combined rod and piston. With 24V is terrifically powerful but is still very good at 12V and, of course, with any intermediate voltage with increasing or decreasing power. It has all the normal uses of a solenoid and an extra one, if wired in series with a make and break, this could be a scribing tool for marking plastics and soft metals. We welcome other ideas and will give a £25 credit voucher for any used. Price **£1**, Order Ref. 877.
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Mic lines

Tim McCormick's article "Putting Mic Amplifiers on the Line", (*EW + WW*, May 1992), targets lack of operating point stability in discrete input transistor pair at high gain settings in convectional mic preamps (his Fig. 3), as a starting point for development of his incredibly complex – and hence impractical – final design. But existence of excellent low noise PNP transistors such as the Rohm *2SB737* (quoted $r_{b'b}$ of 2Ω and typ E_n of 0.4 nV/√Hz), permits allowable emitter degeneration of typically 20Ω per transistor to balance currents, linearise the stage and set maximum gain, without resorting to unbiased capacitors. The following IC differential amplifier can then be capacitor coupled, as is indeed his second stage, to provide remaining required gain: done, with a minimum of low cost components and few matched resistors, to achieve adequate CMR.

But the whole argument fades into insignificance with advent of the Bowers topology *SSM2017*, an eight pin minidip 750 pV/√Hz device. This requires only one external gain setting resistor, for dB to 60dB at low cost, permitting reduced assembly time, PCB real estate and well defined performance. A manufacturer would be hard pressed to ignore such a device in favour of a complex "roll your own" approach, unless some definite sonic virtues could be demonstrated.

Greg Ball
Coolangatta
Australia

Engineers of the world unite!

Congratulations on a splendid analysis of current industrial problems (*EW + WW*, Comment, August 1992). Unfortunately, such views hardly ever reach those in charge of events or general public.

UK electronic related industries, have had little rational planning of development (as evidenced by the Brema book "The Setmakers", a sorry tale indeed).

UK engineers are themselves much to blame, for quietly enjoying their work so much they allow, without effective comment, really important decisions to be taken by those in charge who sometimes have no up to date understanding of technology – or even no understanding at all.

Now the engineering community needs to inform the general public, government and especially their own employers of useful developments which *could* be provided if appropriate decisions were to be taken. If those who make or approve investments are unaware of skills and ideas being wasted, they can hardly be blamed for ignoring them.

Financial matters are discussed at interminable length in several media but engineering related matters almost never.

R H Pearson
Lincs

Visually challenging

I am writing on behalf of the Joint Working Group of the Institute of Food Science & Technology, Royal National Institute for the Blind, Guide Dogs for the Blind Association and British Computer Association of the Blind.

We aim to seek a viable method enabling blind and partially-sighted people to read information on food labels. Our major objective has been print on food packages, but now we are seeking a system which would benefit visually-impaired people, far beyond reading food labels, assisting shopping independence and providing them with greater access to information of all kinds.

Such a system needs to be operated by an individual, involving scanning text on any food package and converting it to a synthesised speech output. Ideally it should be light, portable and compact and needs to recognise as many character fonts and sizes as possible while being as inexpensive as possible (many potential users are elderly and on low incomes).

There are at least 1,000,000 visually-impaired people in the UK alone, and an even larger number throughout the English-speaking world. There are also those who need software appropriate to other languages. There must be someone among your readers who would see a threefold opportunity; a substantial new market waiting for a product, an exciting technical challenge and a chance to help blind people. What a wonderful combination!

J R Blanchfield
5 Cambridge Court
210 Shepherd's Bush Road
London
W6 7NL

Misguided CFA

Textbook writers on electromagnetic radiation, who carry out proper retarded-function mathematics, seem to concentrate on the distant magnetic effect of current flowing in transmitting aerial wire[1], or on effects of current in wire plus distant electric effect of charges collecting at wire ends[2,3]. Moullin[2], for example, considers a vertical wire, supposed to carry a uniform current, joining two conducting spheres. He shows, at distances sufficiently great, that induction fields become negligible; the radiation magnetic field vector due to current, and electric field vector due to charges on spheres, are in phase and in the fixed ratio $(\mu/\varepsilon)^{0.5}$ – the generally accepted result.

There seems to be a whiff of sleight-of-hand in these analyses. In low-frequency electromagnetic problems, magnetic effects of *all* parts of the circuit carrying current are considered. Yet in dealing with electromagnetic radiation, the contribution to radiation of current in the aerial wire is taken into account. But, mysteriously, the contribution to radiation of the return, displacement current flowing back outside the aerial wire is not considered. Why not? Displacement current certainly gets considered, but as an *effect* (of the inductive electric field), not as a cause. But it seems to have as much right to be considered as contributing to radiation as does current in the wire.

Could there be a lacuna here, in

RC modification

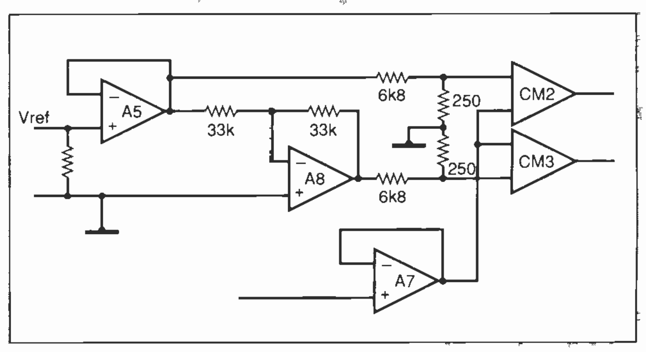
I would like to suggest a simple circuit modification to the RC oscillator in Dan Stiurca's article "Staying in control in an all-pass filter RC oscillator", (*EW + WW*, July 1992).

It produces constant control pulse widths regardless of V_{REF} and solves low amplitude harmonic distortion problems.

Offset nulling of $A_{5,7,8}$ is just one way of eliminating phase-shifting effects of op-amp/comparator offsets.

A R Pleasance
Bradford

Circuit modification where $A_{5,7,8}$ must have offset nulling capabilities



standard treatment of electromagnetic radiation, accounting for continued rumblings, in these columns, on a crossed-field aerial?

Here, inventors[4] do the opposite of other writers by considering the magnetic effect of displacement current between their *D*-plates, but ignoring magnetic effects of current in the short piece of conductor exciting the upper *D*-plate. Whatever the knee jerk response to the CFA it is difficult to criticise inventor accounts of it as incomplete, when standard texts show similar failings.

K Donaldson
Sevenoaks

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Cruel science – bad science

I find it increasingly incredible that the scientific world still uses, let alone places any value on, experiments conducted on live animals in an attempt to extrapolate the results to humans. Messrs Silvio De Flora and Francesca D'Agostini (*EW + WW*, Research Notes, July 1992) themselves admit difficulty doing exactly that.

Effects of UV radiation on human skin are already well known. Spectra produced by quartz-halogen lamps is not only well established, but is easily and painlessly measured.

As your report pointed out, Colin Driscoll of the Radiological Protection Board made public three years ago the risks of UV over-exposure from these lamps, so what is the point of such research? Are lives of poor creatures expended merely to enable De Flora and D'Agostini to write a career-enhancing paper? It is precisely such blatent disregard for non-human life which makes animal-based research a controversial issue.

I am involved with audio electronic field research, and would not dream of measuring distortion produced in an op-amp circuit in an attempt to determine distortion exhibited by a power amplifier, even though the two might be related. It would simply be bad science.

Bad science surrounds and permeates animal based research, degrading and defiling science in general. It should be shunned by those remotely connected with science or who care about creatures who share our precious planet. Surely it is time the scientific community stood up against a relic of barbaric times which has nothing to do with science.

P Williams
Hertfordshire

PSR PS

A M Wilkes, (*EW + WW*, Letter, August 1992), may wish to know amplifiers using his proposals concerning power supply and PSR – including a 20kHz pulse width regulating power supply in an internal diecast box – were produced some twelve years ago by Masaru Nagami of Sony. A pair of these 200W monoblocks, *TAN900* Esprit, have done daily duty in my home without fail. At a mere 10.5kg each, they have shrugged off challenges from quite a few goliaths with "vastly over-built power supplies" of a "naive, resource-wasteful band-aid" variety to quote Greg Ball (*EW + WW*, May 1992). Nothing sampled so far has persuaded me to change and a number of revered four man lift, room heater, models have proved, long term, to be noticeably less involving.

Recently I read that a design by John Franks, Chord Electronics' *SPM 1200*, has taken up the SMPS cause at 80kHz. Praise be, it has achieved at least one rave review – tempered by sheer incredulity at something so small and light which could deliver the goods. I have not tested it myself but am quite prepared to accept its authenticity.

Why are audibly superior and technologically advanced designs never commonplace? Several competent designers tell me they would love to use SMPS but their marketing people forbid it. In many countries weight and size are major consumer criteria. Hence 4mm steel chassis, 1cm thick panels, OTT toroids and generally wasteful specifying of major components. If he can think of some way to tilt the see-saw back from macho to music there might be some room for Mr Wilkes and others to advance the cause.

Geoffrey Horn
Oxford

America's Marconi

Echoing the title of KA Ioffe's article, "Popov: Russia's Marconi" (*EW + WW*, July 1992), Telsa could be considered as America's Marconi. Indeed, there were a few innovations in early radio Telsa had not pioneered.

Telsa tolerated Marconi using devices which he patented until 1915 when he applied for an injunction against Marconi. But the injunction was denied because very few people understood the significance of Telsa's work. In 1943, a review board decided Marconi's patents were invalid because they had been anticipated by Telsa.

A remarkable feature of the Branly coherer is that it remains in its high resistance state while energised by steady current from the relay battery, typically a volt or so. But it "coheres" in response to a single pulse with a potential in order of mV and a duration in order of a ms.

Early spark transmitters radiated energy virtually as a single electromagnetic pulse, as shown in the figure, and while well suited to coherer type receivers, waves were too few to be tuneable. At close range, pulses simply "shocked" receiver tuners into oscillation, making it impossible to eliminate interference.

Telsa, like Popov, employed a coherer to study electromagnetic pulses generated by lightning. But, Telsa employed a Branly type coherer modified to rotate axially by means of a clockwork motor, thus constantly restoring it to a high resistance state and used in conjunction with a paper-tape register.

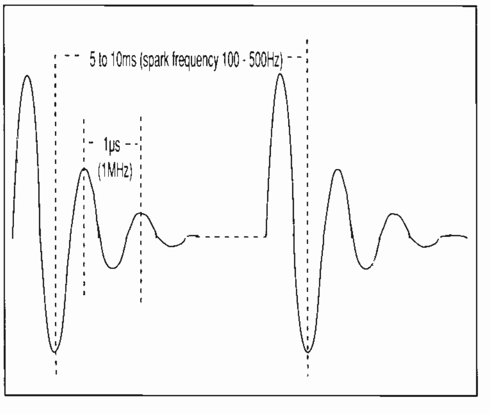
Telsa's "Colorado Springs Notes", 4 July 1899, state thirty minutes after the storm had passed over, discharges were too feeble to operate the register. But, after another 30min, the register began to operate again for some time until once again signals became too weak.

Continuing periodically for some hours, the signals finally ceased when it was estimated that the storm was more than 200miles away. Telsa believed this phenomenon was caused by "earth standing waves".

Unfortunately, I have been unable to repeat Telsa's observations, but, with a replica of his rotating coherer, in conjunction with an earth probe antenna, I have detected pulses. Presumably these are the same mysterious pulses which shocked my original LC very long wave tuner into oscillation (*EW + WW*, Letters, February 1991).

George Pickworth
Kettering

Wavetrains produced by replica of a typical 1901 spark transmitter.



Valve advice...

How can I read valve markings (eg. *6H6, 35Z5*) that have been rubbed off?

I have already tried breathing on the glass, hoping condensation formed will show up the markings – it works sometimes. It does not work with metal-bodied valves.

I do not know whether putting valves under a UV lamp will show up marking remains and do not want to try in case I damage the valve.

Stephen Shaw
South Africa

...interface information

Does anyone know of any software/interface available for use with an IBM compatible PC which can decode audible morse code or digital transmission and can tune into on my SW radio on dozens of different frequencies?

Michael Lec
Gt Yarmouth

If

Many scientists since Dingle (*EW + WW*, Letters, August 1992) have shown Special and General Theories of Relativity are self contradictory, inconsistent, riddled with anomalies and, as formulated, cannot be correct. But Relativity continues to be taught in universities as an undisputed fact. Some maintain there is an academic "mafiosi" which suppresses, prevents, and discourages publication of criticisms of invalidations of Relativity, see *Il Grande Grido - Ethical Probe on Einstein's followers in the USA*, RM Santilli 1984.

Of course if the Special Theory is incorrect, then consistent zero-velocity results of all famous Michelson-Morley type experiments are real, not apparent: the earth really is located at the centre of the universe. Geocentrist scientists, including the Tychonian Society, bring much recently discovered evidence from astronomy and quantum physics to support their cosmology. In addition, if Relativity is wrong, then weight has to be lent to claims of top anti-relativists such as Stefan Marinov, Howard Hayden, Peter Beckmann and Carl Zapffe, who state the electromagnetic doctrines of Maxwell and Hertz, as well as Newton's Third Law and the First Law of Thermodynamics are all suspect. This could put all of physics on to a new track, and lead to development of some new and revolutionary electrical devices, see *Deutsche Physik - International Glasnost Journal on Fundamental Physics*, Volumes 1-3 1992.

A Goldberg
London

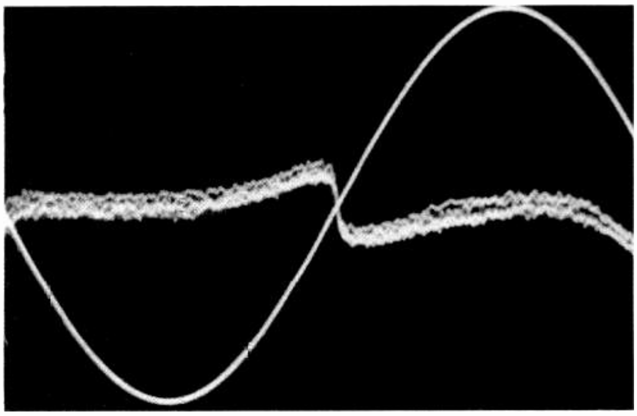
Distorted Logic

We are often told it takes a very good engineer to design a bad amplifier these days. Now with Doug Self's very apposite riposte (*EW + WW*, Letters, August 1992), I have had my confidence shaken.

I was turning over some old papers and came across a photo taken for a commissioned article on Audio Amplifier design, published in *Wireless World* as long ago as June 1969. It is a screen shot of a 'scope, showing clear hiatus in transfer characteristics at 1 watt ,"crossover distortion", also appeared in full colour on your front cover. It was, of course, taken from output of a DF meter, after sampling voltage across a dummy resistive load connected across speaker output terminals of a much favoured hifi amplifier. If my faded notes can be relied upon, it was a little less than 8mV p-p but because of meter inertia, would not read on a waveform analyser. Yet, it was clearly audible and uncomfortably so. And what, I hear you say, did it sound like? I can use a comparison I could not have used then - rather like quantization distortion!

Ironically I recall commenting we had a new, younger generation of audio engineers who appeared to be unaware of this type of distortion. Oh, dear. *Plus ça change.....*

Reg Williamson
University of Keele



Sudden infant death

Effects of electromagnetic pollution on various aspects of the human condition have been reviewed in Roger Coghill's article "Killing fields-biophysical evidence", (*EW + WW*, Febuary 1990). Increase in incidence of sudden infant death, reported by Coghill, among babies living near high high voltage power lines and electric railways can be explained in terms of effects caused by electrical discharges on levels of condensation nuclei and oxides of nitrogen in the atmosphere. Condensation nuclei are small particles of matter, which may or may not be charged, that are present in the air. They are of relevance in cloud formation, fog, smog and other atmospheric processes.

If the baby is sleeping in an environment with a significant level of condensation nuclei, then the exhaled air on cooling may not remain supersaturated but may condense onto the nuclei.

Sudden infant death can now be explained in terms of pooling of excess exhaled air, high in carbon dioxide, at the infant's face.

J A Corbyn
Fremantle
Western Australia

The alternative killing fields

I read with interest Alasdair Philips' article "Power Politics: Playing with Lives" (*EW + WW*, April 1992). I have read similar articles before and I remain unconvinced because, in any study concerning people's health, there is always one factor which researchers fail to consider; the physical position of the subject upon the earth.

The earth is not neutral in its effect upon us as dowsers have repeatedly shown. Serious illnesses frequently occur where noxious earth energies exist.

Michael J Cooney
London

OOOOOOps!

I don't suppose I will be first to point this out, but in my article "Squeezing into the Picture" (*EW + WW*, August 1992), superscripts seem to have gone astray under the sub-heading "Improving on 1000 years", 1,023 should read 10^{23}! Improving on 100,000,000,000,000,000,000,000 years!!

Andy Wright
Buckinghamshire

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Analog Electronics
Ian Hickman

Good all-round electronics designers are hard to find, according to the recruitment specialists. There are either bad all-rounders or good specialist (for example, microwave, power supply, microprocessors specialists). Many young designers have been lured away from the fundamentals of electronic design to more 'glamorous' digital work, yet there are many simple pieces of electronic equipment for which a purely analogue realisation is still cheaper, more reliable and more appropriate than a microprocessor-based solution. Analogue staff are in desperately short supply, and in many fields - telecommunications for example - analogue skills are very much in demand. Ian Hickman's latest book includes many examples from his large collection of circuits (built up over thirty years in commercial, professional and defence electronics), selected for their usefulness in a wide range of applications. Hardback 300pages.
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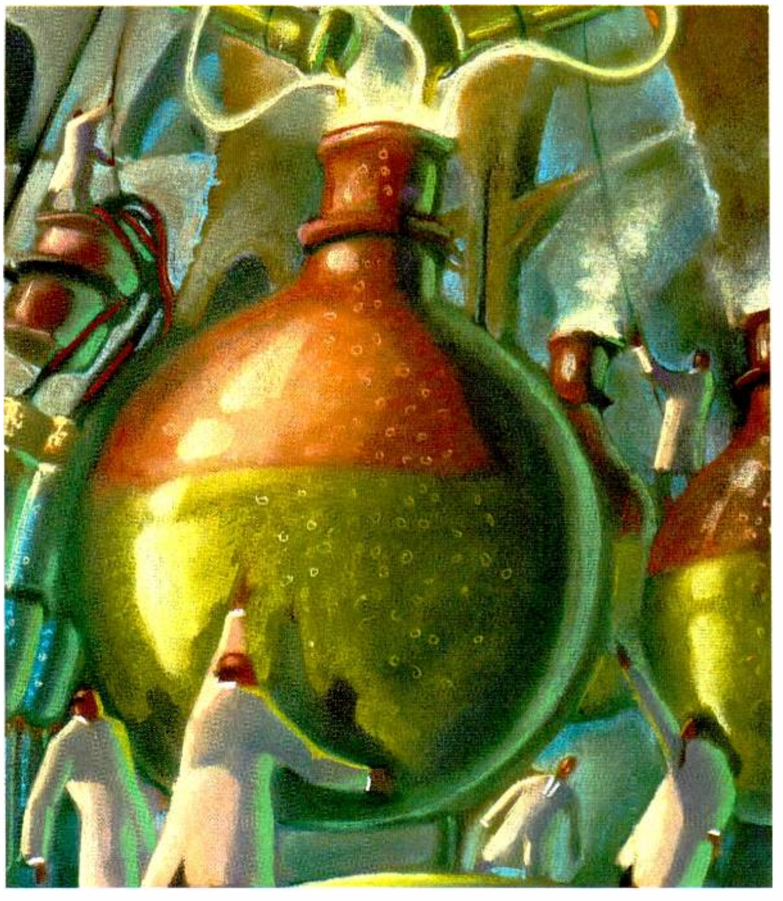


Genuine solutions to spurious arguments

In the third part of his series of articles on direct digital synthesis Ian Hickman turns his attention to spurious signals and looks at how to cope with them.

Several methods can be used to reduce the spurious signals which can appear in the output spectrum of a DDS. Some spurious signals are due to amplitude modulation. Where the output of a DDS is not actually used directly for transmitting but is offset to another frequency in a mixer, choose the DDS signal as the high level LO drive to the mixer and the other translating signal as the lower level 'signal' input to the mixer – assuming the lower level signal is itself spur-free.

This is spur avoidance rather than reduction, but the result is the same. Another spur avoidance scheme is the hybrid synthesiser (**Fig. 1**). Here, a DDS operating at around 10MHz (say 10.7MHz) is used as the reference oscillator of a PLL synthesizer. Using a reference divider ratio *m* in the range 4 to 6, the system operates with a comparison frequency of around 2 to 3MHz. A variable reference frequency permits a simple single loop design, while the high comparison frequency enables the loop gain to be maintained up to a higher frequency than usual, keeping the VCO noise sidebands in check and permitting frequency changes that are very rapid for a PLL type synthesiser. Frequency settling between PLL steps is achieved by adjusting the reference frequency.

With this scheme, the range over which the reference frequency must be varied to provide continuous interpolation between the steps of the main loop varies inversely with the main loop divisor. Power available in modern microcontrollers means the additional control overhead presents no problem. It is true that the arrangement means that 10Hz steps will in general not be exactly 10Hz. But the 32-bit frequency resolution typical of a DDS will provide more than adequate resolution to ensure that any frequency can be set to well within the frequency accuracy of a typical equipment, even at an output frequency of 1GHz.

The range of reference frequency adjustment required is only a few tens of kHz, enabling an inexpensive but highly selective 10.7MHz crystal filter designed for the VHF PMR bands to be used. Tests on one of the DDS chips already mentioned showed that with a 40MHz clock, all spurious outputs were greater than 80dB down everywhere within the bandwidth of the crystal filter (though not elsewhere), over the required range of reference frequency adjustment. The result is a synthesiser system with frequency agility and low close-to-carrier noise levels approaching that of a DDS, combined with the freedom from spurious outputs typical of a PLL system.

Reducing rather than avoiding

Various approaches are also possible for reducing rather than merely avoiding spurious output. In those applications where the DDS uses an off-chip D-to-A, spurious signals can be minimised by careful choice of D-to-A unit. For example, the *Q2334* provides a 12-bit output to the D-to-A. But as a result of extensive testing Qualcomm currently recommend the 10-bit Sony *CX20202A* D-to-A in applications using outputs up to the DDS's maximum output frequency.

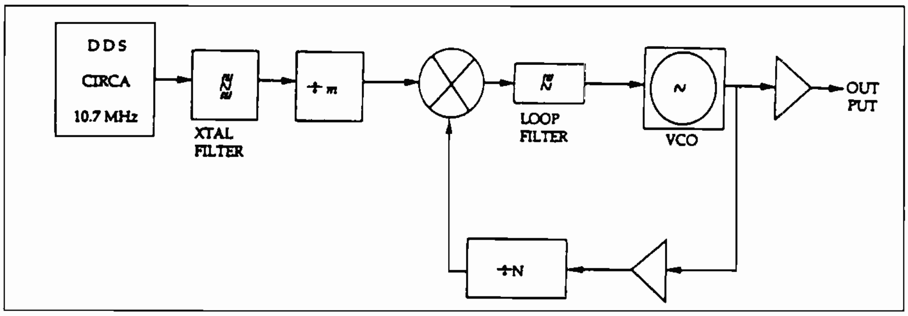
The recommendation highlights that in a DDS application, published D-to-A parameters such as glitch energy, integral linearity and settling time tell only part of the story. (At lower frequencies, more D-to-A bits is naturally a better choice – the TRW *TDC1012* 12-bit D-to-A is recommended for use with the *Q2334* at output frequencies lower than 20MHz. Another high performance D-to-A which should be considered for DDS applications is the Tektronix 12-bit *TKDA30*, which features a 100MHz update rate and 65dB spurious free dynamic range.)

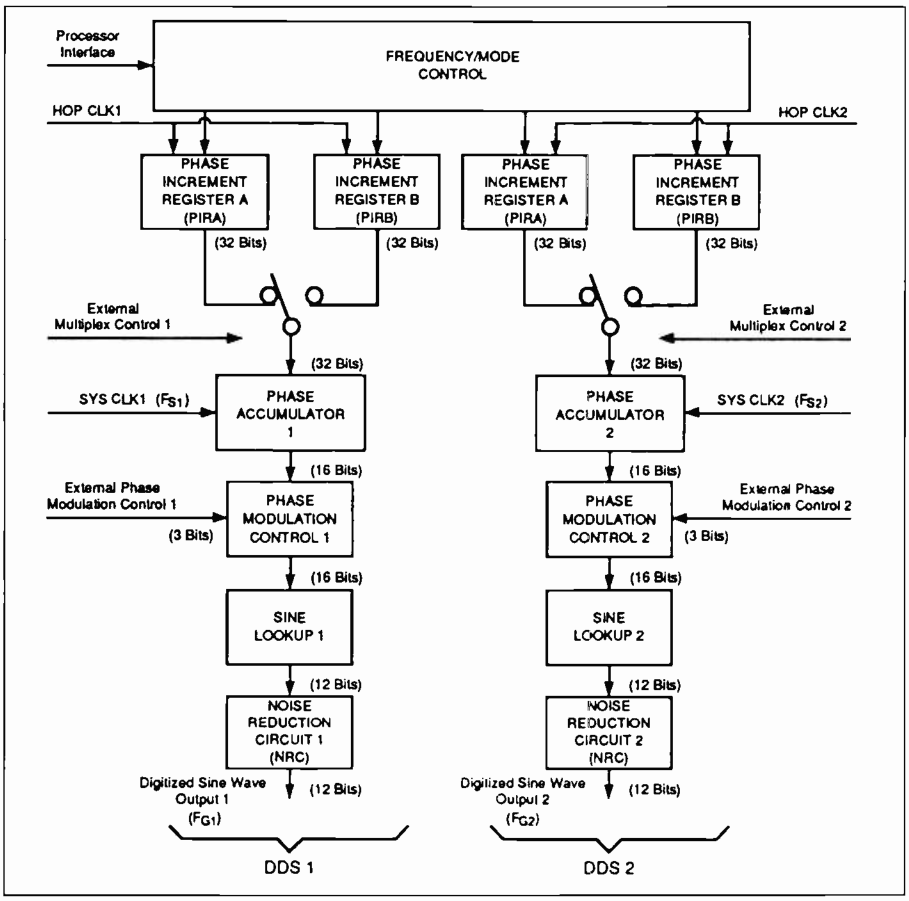
Several workers have described schemes for modifying the mechanism which results in the generation of discrete line spurs. Thus M Bozic[3] reported an arrangement (for use in a hybrid DDS/PLL synthesiser) for applying AM to the output of the DDS. This was done by modifying the look-up values within a conventional 8-bit low frequency TTL DDS. By adjusting the phase of the AM, the amplitude of the unwanted sidebands closest to carrier can be reduced at the expense of increasing the amplitude of other spurs further out where they are outside the PLL loop bandwidth.

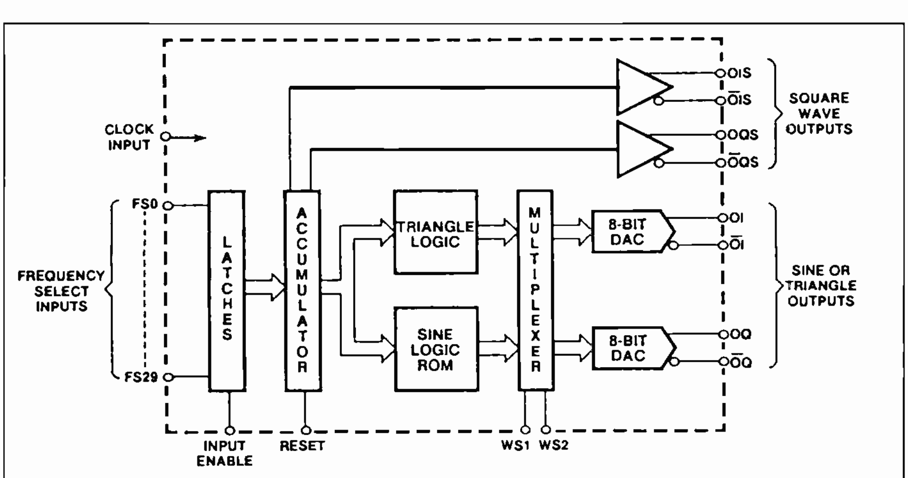
In the same forum, M P Wilson and T C Tozer proposed a potentially more versatile scheme[4]. Here, a pseudo random number is added to the rom output before it is passed to the D-to-A, breaking up the repetitive nature of the error process when there is a fairly simple ratio relating output and clock frequencies. Errors introduced into the D-to-A output are cancelled by the output of a second D-to-A which produces the complement of the pseudo random values.

NRC (noise reduction circuitry) is incorporated in the DDS of **Fig. 2** and may be enabled or not as required. It has the effect of reducing the amplitude of discrete line

Fig. 1. In this hybrid synthesiser, a DDS operating over a narrow frequency range (where all spurious signals are greater than, say, 80dB down) is used as the reference oscillator. If the DDS operates at around 10.7MHz, then an inexpensive but highly selective crystal filter such as is used in PMR (private mobile radio) can be used. (Institution of Electrical Engineers[1])







spurs at the expense of raising the general noise floor – as was mentioned earlier, the total spur energy tends to be independent of whether it is concentrated in one or a few lines, or in a continuous sea of low level spurs (**Fig. 4**).

While NRC reduces AM spurs, it cannot suppress them completely and its effect may not always be apparent: eg if operating at a frequency where there is no large spur. One use for the NRC is to permit the use of a cheaper 8-bit D-to-A. The resultant performance then approaches that of a 10-bit D-to-A without NRC.

Handling PM spurs

The spur reduction techniques mentioned so far are aimed at reducing AM spurs, although as mentioned previously (Direct Digital Synthesis, *EW + WW*, August pp. 630-634), PM spurs are potentially more troublesome, as they cannot be suppressed by limiting. Both AM and PM sidebands appear in pairs equally spaced about the carrier and it is by no means uncommon to find both AM and PM at the same modulating frequency.

Low level AM and PM are orthogonal, with the result that on one side of the carrier the AM and PM sidebands are in phase whereas

Fig. 2. The Q2334 contains two totally independent DDSs of the external D-to-A variety, and is capable of accepting a different clock frequency for each, or of generating different frequencies (with common resolution) using a common clock, or of generating two outputs at the same frequency in quadrature phase or any other relative phase. The two DDSs are controlled by a common microprocessor interface and each supports both phase modulation and phase coherent frequency hopping. Versions of the chip operating at clock frequencies up 50MHz are available, and by combining the two outputs it is possible to operate at up to 80% of the clock frequency[2]. (Qualcomm Inc. and Chronos Technology Ltd.)

Fig. 3a) The SP2002/A accepts clock frequencies up to 1600MHz (1400MHz for the military temperature range /B version) and provides RF outputs up to $F_{clock}/4$, with a choice of sine, triangle and square outputs, each available as an in-phase and quadrature pair with its inverses. This ECL chip runs on –4.5V and includes very fast on-board D-to-As to provide the I and Q sine or triangle outputs. 3b) To achieve its very high speed operation in silicon, the device stores only a part of a sinewave in rom. It clocks up and then down again over half a sinewave and then the MSB inverts the data for the other half cycle. The upper trace shows the device clocked at 10MHz and outputting the corresponding maximum frequency of 2.5MHz, at one of the 256 possible phases (FS29 set to 1, all others zero). In the lower trace, bits 28, 27 and 26 are also set, demanding an illegal output above $F_{clock}/4$;this results in the MSB inversion being out of phase with the half sinewave count. (GEC Plessey Semiconductors Ltd)

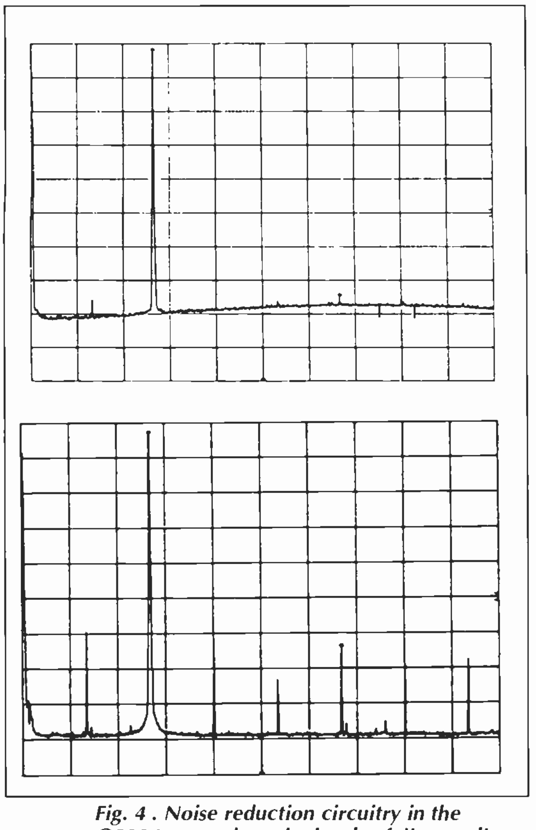


Fig. 4. Noise reduction circuitry in the Q2334 can reduce the levels of discrete line spurs at the expense of raising the general noise floor. (Reproduced by courtesy of Qualcomm Inc and Chronos Technology Ltd)

Types of DDS

Some manufacturers produce DDS chips which provide a sinewave output direct – though it is of course a stepwise approximation rather than a smooth waveform. Such DDSs only need the addition of a frequency setting word input, a clock source and a lowpass filter at the output to provide a complete digital-frequency-command to RF-output synthesiser system. A good example is the GEC Plessey Semiconductors *SP2002* (**Fig. 3**).

Other manufacturers' DDS chips provide the step values representing an output sinewave in *P* bit binary notation, for application to a separate off-chip D-to-A converter. In these instances, *P* is usually in the range 8 to 16 bits. A good example of this approach is the Qualcomm Q2334, Fig. 2, which outputs two sinewaves in 12bit binary form. There are advantages and disadvantages to using an internal D-to-A, but it is largely a matter of horses for courses.

For a DDS generating outputs at up to a few tens of MHz, an external D-to-A provides the user with greater flexibility (like choosing a lower resolution or slower but cheaper D-to-A where appropriate): for a DDS accepting clock frequencies up into the microwave region, an external D-to-A is not feasible, due to the delay and skew which would be produced in the longer data lines associated with an off-chip D-to-A converter.

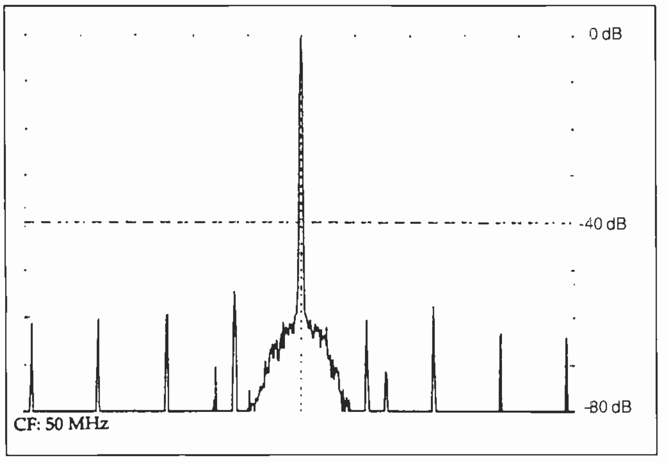
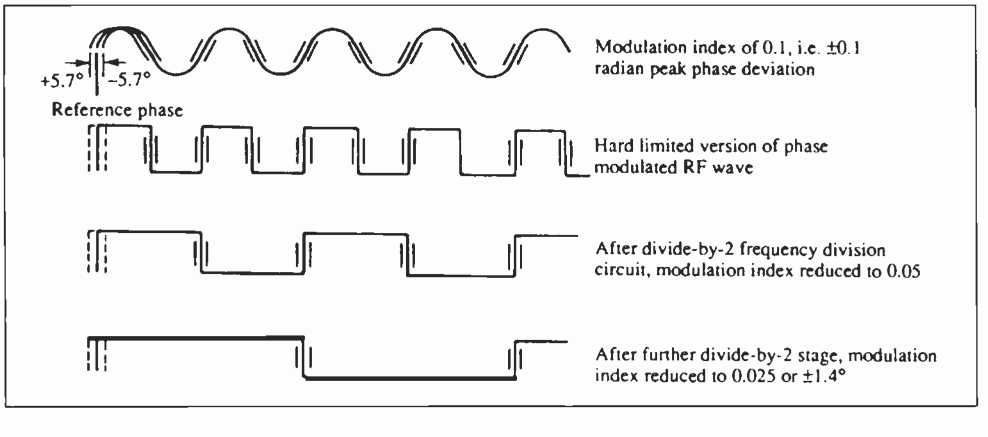


Fig. 5. Output of an SP2002 clocked at 400MHz with bits 28 and 13 set, giving a demanded frequency of 50.001526MHz (display centre frequency 50MHz, span 100kHz). (Institution of Electrical Engineers[1])

Fig. 6. Illustrating the reduction of phase deviation when a phase modulated signal passes through a frequency divider chain, showing for example how division by four reduces the modulation index by a factor of 4, corresponding to a 12dB reduction in PM sideband level. (Analog Electronics, Ian Hickman, Heinemann-Newnes).



on the other they are in anti-phase. This results in a pair of sidebands of unequal amplitude; for example the closest-in pair in **Fig. 5** differs by 6dB.

After converting from log to linear (dB to volts), simple arithmetic shows that in this case, one sideband pair has three times the amplitude of the other. It is not obvious whether the sidebands are mainly due to PM, with smaller AM sidebands responsible for the difference in amplitude, or the other way round. Hard limiting would suppress the AM sidebands, leaving the two PM sidebands equal and at an intermediate level between those shown, or both 10dB lower than the intermediate level, respectively. If the AM and PM sideband amplitudes are equal, they will cancel out completely on one side of the wanted output, leaving an isolated spur on the other side.

An isolated spur is indeed often encountered, but it is more likely to be due to an image or alias. It results in both AM and PM, thus hard limiting will convert the single spur into a pair of PM sidebands, each 6dB lower than the original spur.

AM can be suppressed by limiting, but the only way to reduce PM is by frequency division. Dividing the output frequency of a DDS by a factor of two will reduce the PM sideband amplitudes by 6dB, by halving the modulation index – this assumes that the modulation index is small in the first place (**Fig. 6**).

Of course, the technique buys PM spur reduction at the price of a reduced frequency coverage (but finer resolution) from the DDS, but with GaAs DDS chips such as those from Sciteq (see picture) providing output frequencies up to 300MHz or more, this may not represent a limitation in many applications.

Where it is, the range covered can be restored by up-conversion in a DDS plus mix/filter arrangement. Both upper and lower sideband outputs can be used, selected by appropriate filters. But limitations of practical filters mean that there will be a band just above and below the up-conversion carrier that cannot be used.

The difficulty can be avoided, and the output frequency coverage extended to about three times the available range of the divided DDS output, by using one or other of two up-conversion carriers (**Fig. 7**). Here, (assuming LO_2 is higher in frequency than LO_1) the LSB of the former can fill in the gap between the sidebands of the latter and vice versa, providing continuous frequency coverage of 3*N*, where *N* is the basic range of the DDS (or the range after division, if this is employed for PM spur reduction).

In another application describing PM spur reduction by division[5], the division process involves squaring up the input sinewave for input to the digital dividers (Fig. 6), thereby suppressing any AM spurs as well as reducing the PM ones. "Division" can be carried out in the DDS itself, simply by limiting the maximum commanded frequency to one half, one quarter or whatever of the rated maximum output frequency.

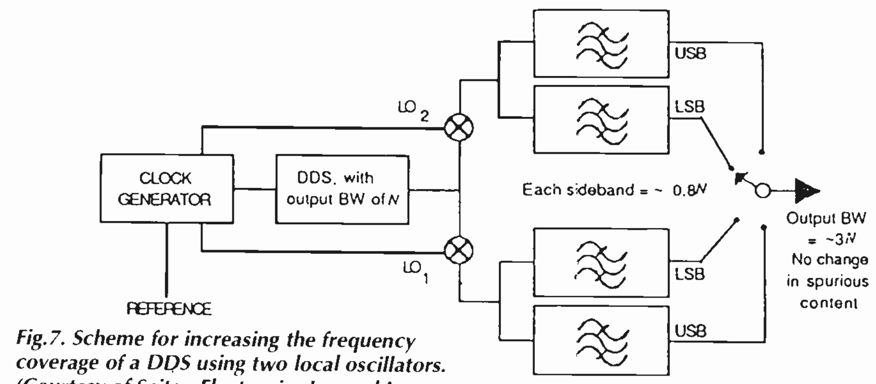
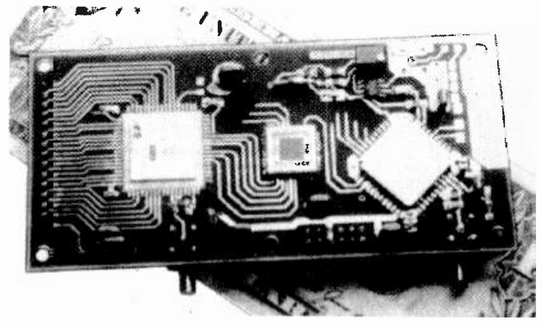


Fig.7. Scheme for increasing the frequency coverage of a DDS using two local oscillators. (Courtesy of Sciteq Electronics Inc and Lyons Instruments Ltd.)



The ADS332-101 is a complete GaAs DDS module on a PCB. It provides outputs up to 300MHz or more with sub-1Hz resolution from a 640MHz clock, with spurious outputs at –40dB or less, harmonics –35dB or less. Output power is –15dBm ±2dB. (Sciteq Electronics Inc and Lyons Instruments Ltd).

At least one manufacturer states that this is more effective in minimising PM spurs than using the whole range of the DDS followed by division, especially if the post-DDS reconstruction lowpass filter's cutoff frequency is reduced appropriately. But the approach does not give the finer frequency resolution provided by dividing the DDS output frequency.

Another scheme for spur reduction (or possibly it might be better described as spur avoidance) might be worth further investigation. The method uses two different clock frequencies, and relies on the fact that when the relation between the output frequency and the clock is a fairly complicated ratio, the spurious energy is dispersed among many lower level spurs rather than concentrated in just one or a few large ones.

Figure 5 shows the output of an *SP2002* DDS clocked at 400MHz with FSW bits 28 and 13 set, giving a demanded frequency of 50.001526MHz; several spurious signals are visible. If bit 14 or 15 or 16 ... is set instead of bit 13, the signals all move further and further away from the wanted output frequency, until with bits 28 and 21 set, the closest spur is about 800kHz away (**Fig. 8**, centre frequency 50.384MHz, span 4MHz).

If bit 13 is now set again (together with both bits 28 and 21) close-in spurs reappear, but only at a lower level, **Fig. 9** (span 50kHz). This is in line with the observation that the more complicated the output to clock frequency ratio, the more lines the spur energy is distributed over, and the lower their general level.

Thus if the 400MHz clock were used when generating frequencies in the region of 53.125, 59.375, 65.625 ... 96.875MHz and an offset clock of 425MHz when generating frequencies close to 50, 56.26, 62.5 ... 100MHz, a reduced level of spurious outputs (especially close-in spurs) might be obtained. A closer clock offset than 25MHz might be better and the reduction of close-in spurs would be particularly beneficial in a hybrid DDS/PLL application.

DDS sources can perform virtually all of the usual forms of modulation, thus implementing almost a complete transmitter exciter on a single chip. In addition to AM, FM and PM, FSK is simply achieved, with the added advantage over the FEK form that it is phase coherent and constant envelope. Thus FFSK (fast FSK, also known as MSK – minimum shift keying) is also easily produced. The more complex forms of modulation such as 16-PSK or 64-APK, etc are also relatively straightforward in a DDS system. ■

Next month: Details on modulation methods; other manufacturers of DDS chips and their products; more considerations relating to DDS operation and applications.

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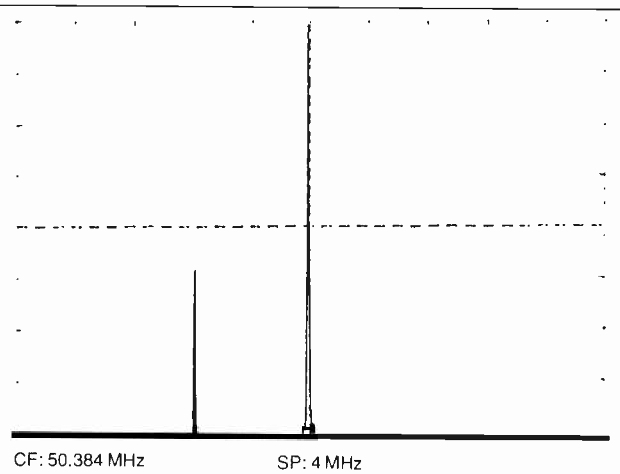


Fig. 8. As Figure 5 but bits 28 and 21 set, centre frequency 50.384MHz, span 4MHz. Nearest spur now about 800kHz away from the wanted output. (Institution of Electrical Engineers[1])

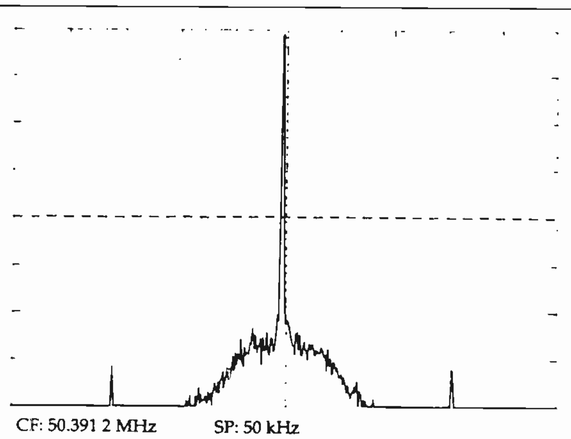


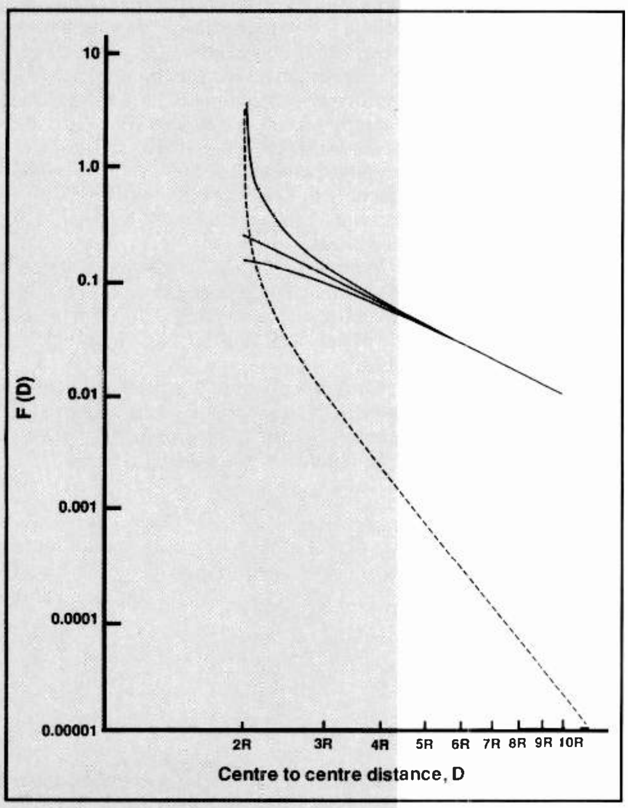
Fig. 9. As Fig. 5 but bits 28, 21 and 13 set, centre frequency 50.391MHz, span 50kHz. Close-in spurs now reappear, but only at a lower level. (Institution of Electrical Engineers[1])

Why Cavendish kept "Coulomb's" law a secret

Henry Cavendish experimentally proved "Coulomb's" law – before Coulomb. Leonid Kryzhanovsky suggests reasons why Cavendish might have preferred to keep his work to himself.

Most readers probably know that Henry Cavendish (1731-1810) was the first to prove experimentally what we call "Coulomb's law" in the early 1770s. This was long before the French scientist himself announced his work, but Cavendish never published an account of his own researches[2]. The story has intrigued historians of science since 1879 when Maxwell published Cavendish's electrical researches, some of which had never been disclosed before.

Many hypotheses have been framed as to why Cavendish would not publish his proof of the law of electrostatic force – mostly bearing on Cavendish's personality. For example, contemporaries speak of his haughtiness and neglect of others' opinions. But this cannot really be supported.



Forces of attraction between a charged conducting sphere and an uncharged conducting sphere of the same diameter (dashed line) obey the law 1/D (the monopole-dipole interaction). Spheres of varying charge state – solid line – did not seem to follow the simple inverse relationships of attraction. Was Cavendish unable to resolve the effect of induced charge?

True, he led a secluded life by devoting himself completely to science. But he did publish about twenty articles. He also invited colleagues to his house to demonstrate his experiments, amusing his guests with an "electric fish" made of wood and leather and placed in a bath with hidden Leyden jars to produce an electric shock.

He let people use his library – and he could cooperate with others as witnessed by his heading a committee on lightning protection.

Professor J L Heilbron suggests that Cavendish did not publish his results simply because of the incomplete state of his research, the want of an occasion and a growing interest in chemistry[3].

The first of these arguments seems to me especially convincing and I believe there might have been purely scientific reasons which kept Cavendish from making public his proof of "Coulomb's law".

Coulomb or Cavendish

Charles Augustin de Coulomb (1736-1806) formulated the "fundamental law of electricity" as follows: "The repulsive force of two small [conducting] globes charged with the same kind of electricity varies as the inverse square of the distance between the centres of the globes"[4].

He stated the same law for unlike changes (the attraction case), experiments with a torsion balance providing support for the law in both cases. Thus, electrostatic force is given by

$$F \propto 1/d^2 \qquad (1)$$

where d is the centre-to-centre distance. Coulomb's memoirs appeared in 1788.

On the basis of a single null experiment with an uncharged conducting sphere

"Cavendish cared more for investigation than for publication" – Maxwell[1].

Force of interaction between two spheres

D	F $q_1 = -q_2$	$F = 1/D^2$ $\lvert q_1\rvert = \lvert q_2\rvert$	F $q_1 = q_2$	F $q_1 \neq 0$, $q_2 = 0$
2.0	∞	0.25	0.154	3.357908
2.05	1.22538	0.237954	0.15012	0.3043072
2.1	0.759739	0.226757	0.146583	0.1625742
2.2	0.470269	0.206612	0.139794	0.0842546
2.3	0.352514	0.189036	0.133313	0.0552683
2.4	0.285277	0.1736	0.127089	0.0397156
2.5	0.240647	0.16	0.121091	0.0299590
3	0.134819	0.111111	0.094437	0.0100980
4	0.067097	0.0625	0.058457	0.0021602
5	0.041404	0.04	0.036680	0.0006811
6	0.028324	0.027778	0.027250	0.0002684
7	0.020656	0.020408	0.020165	0.0001228
8	0.015751	0.015625	0.015501	0.0000625
9	0.012415	0.012346	0.012277	0.0000345
10	0.010041	0.01	0.00996	0.0000203

embraced by two charged conducting hemispheres of a somewhat greater diameter, Cavendish arrived about 1773 at the law

$$F \propto q_1 q_2 / d^2 \qquad (2)$$

where $q_{1,2}$ are the values of charges of infinitesimal geometric size. It is Eq (2) rather than Eq (1) which is known as Coulomb's law.

Assume $q_1 \neq 0$ and $q_2 = 0$. Then it formally follows from Eq (2) that $F = 0$, ie there is no interaction between balls. Cavendish naturally knew this was not so, an uncharged ball being always attracted by a charged one. It is this evidence that might have kept him from publishing Eq (2).

Difficulties which Cavendish might have found with $q_1 \neq 0$ and $q_2 = 0$ are easily countered by making recourse to electrostatic induction (Faraday's term) – a phenomenon well known in the 18th century for long rods[5]. But to transfer the phenomenon onto small balls involved overcoming a psychological barrier.

Reasoning presented in the foregoing paragraph has been recently suggested by Boris Khasapov in a private communication. When I was a student, I also thought of the particular case where $q_1 \neq 0$ and $q_2 = 0$ and resolved mentally the paradox. But I knew nothing of Cavendish's story and I did not think of how to tackle the case quantitatively (an aspect I shall turn to later).

Electrostatic induction

Results of Robison[6] and Coulomb[7] are distorted since researchers have not accounted for electrostatic induction. But, as far as Eq (1) is concerned, there is not the paradox yielded by Eq (2) for $q_1 \neq 0$ and $q_2 = 0$, only a discrepancy between Eq (1) and experimental results.

Coulomb published the law named after him without the numerator although he had made experiments to verify the relation:

$$F \propto q_1 q_2 \qquad (3)$$

These experiments – simple in principle and clever – involved touching a charged conducting ball with an uncharged conducting ball of the same size, and assuming the initial charge will be divided in half between balls. Accordingly, force F should be reduced by one half upon each touch, which he observed[8].

But Coulomb could not help thinking of the case when $q_1 \neq 0$ and $q_2 = 0$, which is why he did not combine Eqs (1) and (3) established by him into Eq (2) to obtain a real "Coulomb law". He could afford to publish Eq (1) since it followed from a separate series of experiments (without charge division). At the same time, Cavendish could not afford to publish Eq (1) alone since his experiment yielded at once Eq (2) with its paradox which he probably could not resolve.

In connection with the paradox, it is interesting to evaluate the force of interaction between like conducting balls for $q_1 \neq 0$ and $q_2 = 0$. A comparison can be made by plotting the force of interaction F (in relative units) as a function of centre-to-centre distance D expressed as a multiple of radius R of a sphere (see figure). A straight line represents an idealised Coulomb law where charges $q_1 = \pm q_2$ of spheres are concentrated at their centres (dependence $F = 1/D^2$ assumed). The upper solid line represents an attraction for $q_1 = -q_2$, and the lower solid line represents the repulsion for $q_1 = q_2$. The dashed line relates to the case of particular interest: $q_1 \neq 0$ and $q_2 = 0$. The latter three dependencies include electrostatic induction. Data used for construction of the curves are shown in the Table.

The dashed line shows that forces of attraction between a charged conducting sphere and an uncharged conducting sphere of the same diameter obey the law $1/D$ (the monopole-dipole interaction) quite precisely from $D = 10R$ to $D = 3R$, at which point it begins to increase rapidly with decreasing D. The force remains finite even for $D = 2R$.

Both the graph and table are due to Professor JA Soules of Cleveland State University who obtained them by a numerical method[7]. He has treated $q_1 \neq 0$ and $q_2 = 0$ upon my request, for which I am especially grateful.

Correct conclusion

In conclusion, I would like to remind readers of Benjamin Franklin's (1706-1790) experiment and a correct conclusion incorrectly drawn from it by Joseph Priestly (1733-1804). In 1755 Franklin discovered that a cork ball on a silk thread dipped into a charged metal can was not attracted by its walls and, when brought into contact with them, acquired no charge. From a vague analogy with Newton's work on the force of gravity within a hollow sphere, Priestley suggested, in 1767, Eq (1) for the electrostatic force[9].

Professor Soules writes in the cited paper that a theoretical approach is more satisfactory in establishing Eq (1) than any experiment – but, from a modern standpoint only. Indeed, Franklin's experiment confirms Maxwell's equation: div $D = \rho$ – a generalised form of Coulomb's law (take a closed surface passing inside walls of can). But in the 18th century, neither Maxwell's equations nor Gauss's theorem had been formulated.

So Cavendish could have rightfully claimed to be the first discoverer of what we call Coulomb's law.

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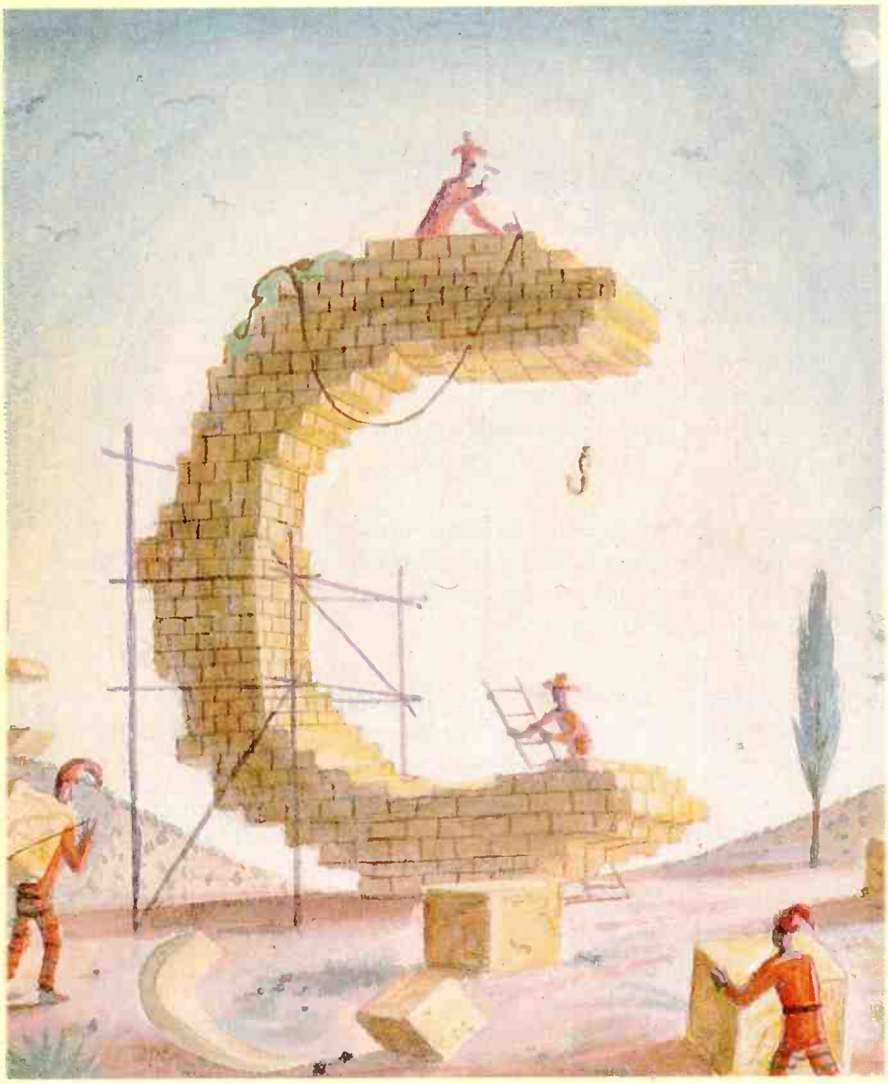

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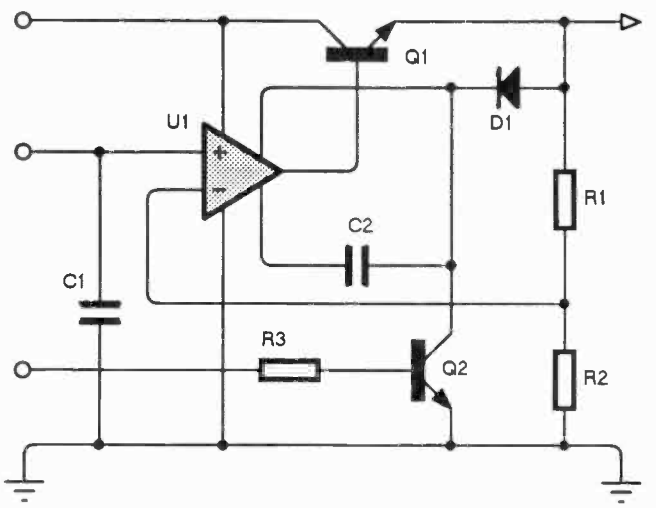

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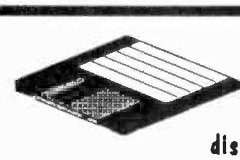

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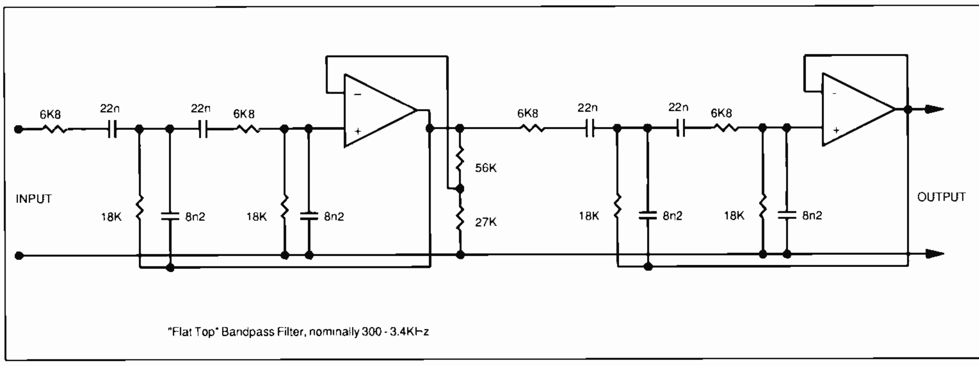


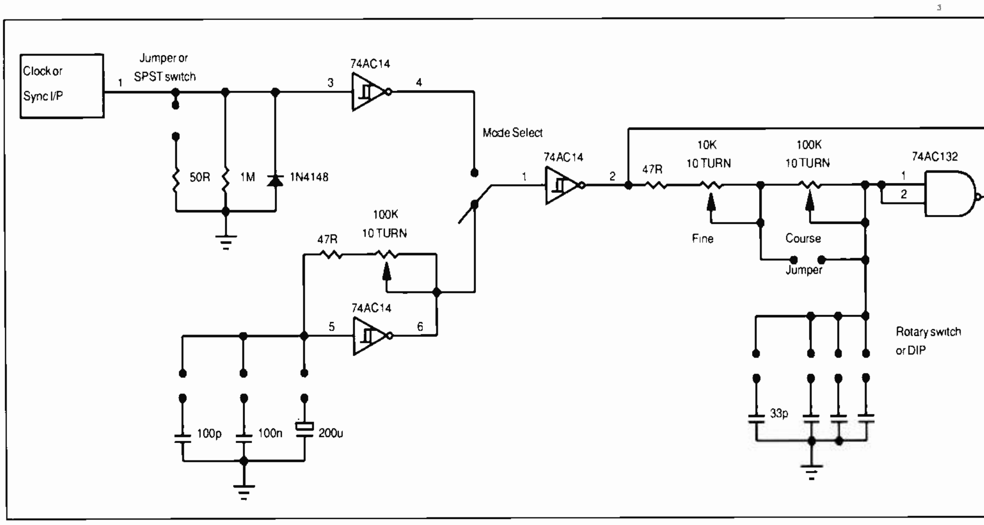
Exotic-looking band-pass filter for speech frequencies is a pair of combined equal-value Sallen and Key high and low-pass circuits, each section using the same op-amp.

Pulse generator

The pulse generator shown was designed for use in the testing of cable drivers and gives a pulse output of 5ns at up to 50MHz, or 10ns and 1MHz.

Clocking is by an external clock or sync. input, or by the internal clock oscillator. The two Nands and variable delay after the clock selector form the pulse generator,

Richard Payne
Wimbledon
London SW19



Comprehensive pulse generator providing variable-width pulses down to 5ns at up to 50MHz, or 1A 10ns pulses at 1MHz.

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PC-AT 386 20-DX FULLY LOADED FOR £499!

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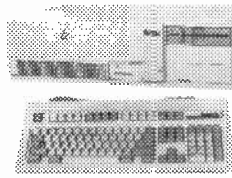
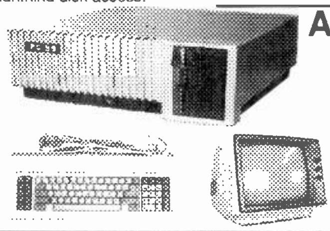


The MP386 quality made by Mitsubishi to last a lifetime! Brand new with all manuals and software plus Super PC-Quick Disk Accelerator for lightning disk access. Only............. **£499** (F)



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5.25" **Teac** FD-55(36). 360K half height.	£22.95(B)
5.25" **Teac** FD-55(72). 720K half height.	£24.95(B)

*** Data cable included in price.**

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THE *AMAZING* TELEBOX!

Converts your colour monitor into a QUALITY COLOUR TV!!

TV SOUND & VIDEO TUNER!

The TELEBOX consists of an attractive fully cased mains powered unit, containing all electronics ready to plug into a host of video monitors made by manufacturers such as MICROVITEC, ATARI, SANYO, SONY, COMMODORE, PHILIPS, TATUNG, AMSTRAD and many more. The composite video output will also plug directly into most video recorders, allowing reception of TV channels not normally receivable on most television receivers (TELEBOX MB). Push button controls on the front panel allow reception of 8 fully tuneable 'off air' UHF colour television or video channels. TELEBOX MB covers virtually all television frequencies VHF and UHF including the HYPERBAND as used by most cable TV operators. Composite and RGB video outputs are located on the rear panel for direct connection to most makes of monitor. For complete compatibility - even for monitors without sound - an integral 4 watt audio amplifier and low level Hi Fi audio output are provided as standard.

Telebox ST	for composite video input monitors	£32.95
Telebox STL	as ST but with integral speaker	£36.50
Telebox MB	as ST with Multiband tuner VHF-UHF-Cable. & hyperband For overseas PAL versions state 5.5 or 6mhz sound specification.	£69.95
Telebox RGB	for analogue RGB monitors (15khz)	£69.95

Shipping code on all Teleboxes is (B)

RGB Telebox also suitable for IBM multisync monitors with RGB analog and composite sync. Overseas versions VHF & UHF call. SECAM / NTSC not available.

No Break Uninterruptable PSU's

Brand new and boxed 230 volts uninterruptable power supplies from **Densei**. Model MUK 0565-AUAF is 0.5 kva and MUD 1085-AHBH is 1 kva. Both have sealed lead acid batteries. MUK are internal, MUD has them in a matching case. Times from interrupt are 5 and 15 minutes respectively. Complete with full operation manuals...........**MUK......£249** (F) **MUD......£525** (G)

POWER SUPPLIES

Power One SPL200-5200P 200 watt (250 w peak).Semi open frame giving +5v 35a, -5v 1.5a, +12v 4a (8a peak), -12v 1.5a, +24v 4a (6a peak). All outputs fully regulated with over voltage protection on the +5v output. AC input selectable for 110/240 vac. Dims13" x 5" x 2.5". Fully guaranteed RFE. **£85.00** (B)

Power One SPL130. 130 watts. Selectable for 12v (4A) or 24 v (2A). 5v @ 20A. ± 12v @ 1.5A. Switch mode. New. **£59.95**(B)

Astec AC-8151 40 watts. Switch mode. +5v @ 2.5a. +12v @ 2a. -12v @ 0.1a. 6-1/4" x 4" x 1-3/4".New **£22.95**(B)

Greendale 19AB0E 60 watts switch mode.+5v @ 6a,±12v @ 1a,+15v @ 1a. RFE and fully tested.11 x 20 x5.5cms. **£24.95**(C)

Conver AC130. 130 watt hi-grade VDE spec.Switch mode.+5v @ 15a,-5v @ 1a,±12v @ 6a.27 x 12.5 x 6.5cms.New. **£49.95**(C)

Boshert 13090.Switch mode.Ideal for drives & system. +5v@ 6a, +12v @ 2.5a, -12v @ 0.5a, -5v @ 0.5a. **£29.95**(B)

Farnell G6/40A. Switch mode. 5v @ 40a.Encased **£95.00**(C)

Farnell G24/5S. As above but 24v @ 5a. **£65.00**(C)

BBC Model B APM Board

£100 CASH FOR THE MOST NOVEL DEMONSTRATABLE APPLICATION!

BBC Model B type computer on a board. A major purchase allows us to offer you the **PROFESSIONAL** version of the BBC computer at a parts only price. Used as a front end graphics system on large networked systems the architecture of the BBC board has so many similarities to the regular BBC model B that we are sure that with a bit of experimentation and ingenuity many useful applications will be found for this board!! It is supplied complete with a connector panel which brings all the I/O to 'D' and BNC type connectors - all you have to do is provide +5 and ± 12 v DC. The APM consists of a single PCB with most major ic's socketed. The ic's are too numerous to list but include a 6502, RAM and an SAA5050 teletext chip. Three 27128 EPROMS contain the custom operating system on which we have no data. On application of DC power the system boots and provides diagnostic information on the video output. On board DIP switches and jumpers select the ECONET address and enable the four extra EPROM sockets for user software. Appx. dims: main board 13" x 10". I/O board 14" x 3". Supplied tested with circuit diagram, data and competition entry form.

Only £29.95 or 2 for £53 (B)

SPECIAL INTEREST

Trio 0-18 vdc bench PSU. 30 amps. New	£ 470
Fujitsu M3041 600 LPM band printer	£2950
DEC LS/02 CPU board	£ 150
Rhode & Schwarz SBUF TV test transmitter 25-1000mhz. Complete with SBTF2 Modulator	£6500
Calcomp 1036 large drum 3 pen plotter	£ 650
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HP 7580A A1 8 pen high speed drum plotter	£1850
Kenwood DA-3501 CD tester, laser pickup simulator	£ 350

BRAND NEW PRINTERS

Microline 183. NLQ 17x17 dot matrix. Full width.	£139 (D)
Hyundai HDP-920. NLQ 24x18 dot matrix full width.	£149 (D)
Qume LetterPro 20 daisy. Qume QS-3 interface.	£39.95 (D)
Centronics 152-2 9 x 7 dot matrix. Full width.	£149 (D)
Centronics 159-4 9 x 7 dot matrix.Serial. 9-1/2" width	£ 99 (D)

MONITORS

MONOCHROME MONITORS

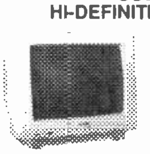
THIS MONTH'S SPECIAL!

There has never been a deal like this one! **Brand spanking new & boxed** monitors from NEC, normally selling at about £140! These are over-engineered for ultra reliability. 9" green screen composite input with etched non-glare screen plus switchable high/low impedance input and output for daisy-chaining. 3 front controls and 6 at rear. Standard BNC sockets. Beautiful high contrast screen and attractive case with carrying ledge. Perfect as a main or backup monitor and for quantity users! **£39.95 each** (D) **or 5 for £185**(G)

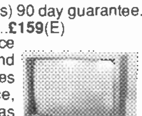
COLOUR MONITORS

HI-DEFINITION COLOUR MONITORS

SAVE £59 - ONLY £100 WHEN BOUGHT WITH THE 386 ABOVE!



14" Philips Model CM8873 VGA multisync all the way up to 34Khz with 640 x 480 resolution. This one has everything! Two switches enable you to select CGA, EGA or VGA and digital/analog. Unusual for a professional monitor, sound is also provided, with a volume control. There is also a special "Text" switch for word processing, spreadsheets and the like. Compatible with virtually all computers including IBM PC's, Amiga, Atari, BBC, Archimedes etc. Good used condition (possible minor screen burns) 90 day guarantee. 15" x 14" x 12". Only....................................£159(E)



KME 12" high definition colour monitors. Nice tight 0.28" dot pitch for superb clarity and modern two tone plastic styling. Operates from any 15.625 khz 15.625 khz RGB video source, with RGB analog and composite sync such as Atari, Commodore Amiga, Acorn Archimedes & BBC. Measures only 13.5" x 12" x 11". Also functions as quality TV with our RGB Telebox. Excellent used condition with 90 day guarantee. In nice two tone beige and brown case. Only ....£149 (E)

Brand new Centronic 14" monitor for IBM PC and compatibles at a lower than ever price! Completely CGA equivalent. Hi-res Mitsubishi 0.42 dot pitch giving 669 x 507 pixels. Big 28 Mhz bandwidth. A super monitor in attractive style moulded case.Full 90 day guarantee. Only .... £129 (E)

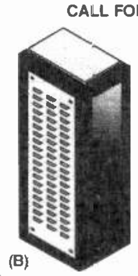
NEC CGA 12" IBM-PC compatible. High quality ex-equipment fully tested with a 90 day guarantee. In an attractive two tone ribbed grey plastic case measuring 15"L x 13"W x 12"H. The front cosmetic bezel has been removed for contractual reasons. Only.. .... **£79** (E)

20", 22" and 26" AV SPECIALS

Superbly made UK manufacture. PIL all solid state colour monitors, complete with composite video & sound inputs. Attractive teak style case. Perfect for Schools,Shops,Disco, Clubs. In EXCELLENT little used condition with full 90 day guarantee.

20"....£135 22"....£155 26"....£185 (F)

CALL FOR PRICING ON NTSC VERSIONS!



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Rack 1 Complete with removable side panels.......	£275.00 (G)
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80 mm	DC **5 v.** Papst 8105G 4w. 38mm. RFE.	£19.95(A)
92 mm	DC 12v. 18 mm thick.	£14.95(A)
4 inch	DC 12v. 12w 1½" thick	£12.50(B)
4 inch	DC 24v 8w. 1" thick.	£14.50(B)

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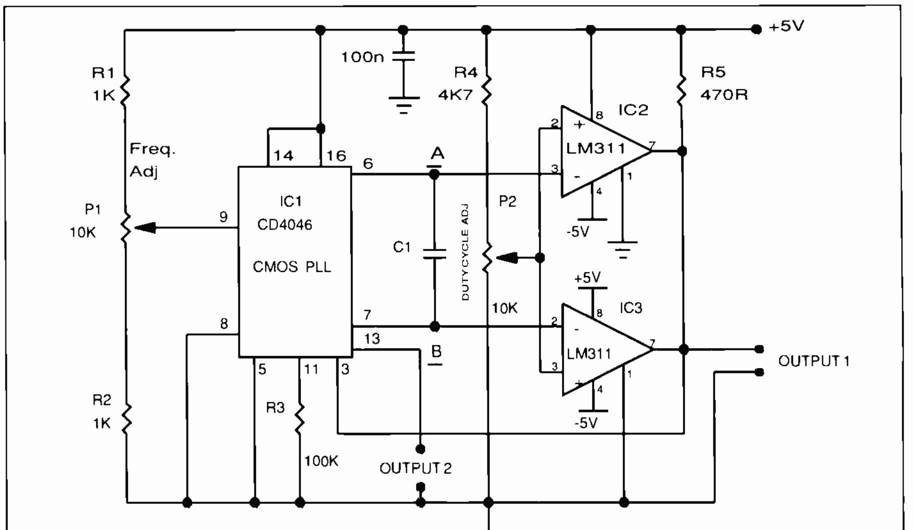
CIRCLE NO. 134 ON REPLY CARD

Fig. 1. This pulse generator offers independent control of frequency and duty cycle over the range 20Hz-200kHz and 0-100%.



C1	Freq. Range
100n	20Hz - 200Hz
10n	200Hz - 2kHz
1n	2kHz - 20kHz
100p	20kHz - 200kHz

Pulse generator with independent F and M:S setting

Using a *CD4046* phase-locked loop, this pulse generator accepts independent settings of frequency and duty cycle.

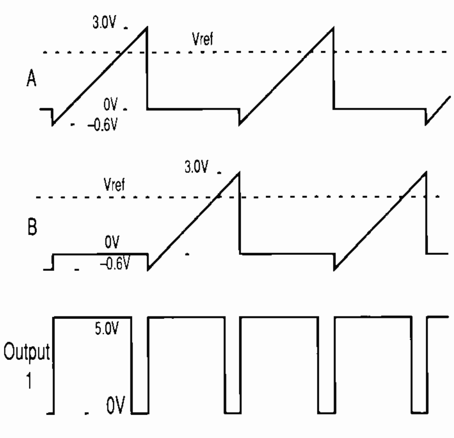
Initially, the VCO in the loop holds pin six, one side of C_1 in **Fig.1**, to ground, C_1 being charged by a constant current whose value depends on R_3 and the frequency-setting voltage at pin nine. When the resulting ramp at point *B* reaches the threshold of an internal inverter, an internal flip-flop changes state, whereupon point *B* is grounded and point *A* starts to ramp upwards. When this too reaches the threshold of an identical inverter, the flip-flop again changes state and the cycle repeats.

Since the internal inverters are identical, the ramps at points *A* and *B* are also identical. **Figure 2** shows the circuit action.

Comparison of the two ramps with a variable reference voltage in the two *LM311s* produces output 1; mark:space variation is by means of adjustment of V_{ref} and frequency setting by way of the control voltage on pin nine of the PLL. Output 2 is obtained by inverting output 1 in the PLL's phase comparator. Frequency is variable from 20Hz to 200kHz and duty cycle from 15% to 100% for output 1 and 0-85% in output 2 with a range of C_1 values from 0.1μF to 100pF.

M S Nagaraj
ISRO Satellite Centre
Bangalore
India

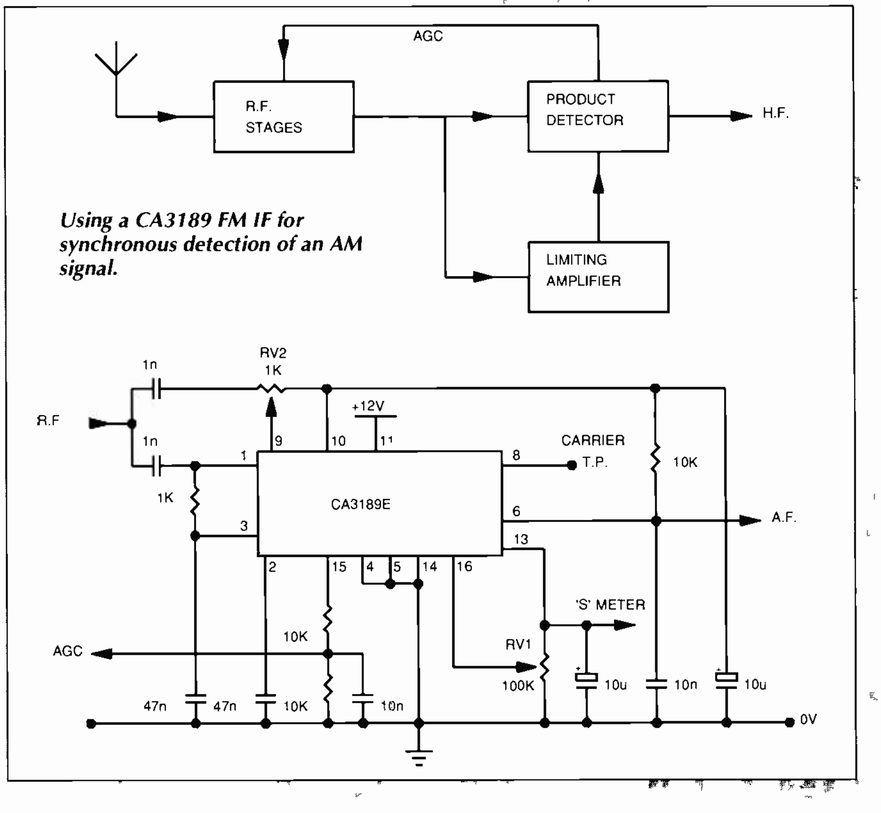
Fig.2. Waveforms in the pulse generator. Note that the ramp starts from -0.6V, not zero volts, giving a duty cycle range of 15%-100% in output 1 and 0-85% in output 2.



Product detector for AM

A CA3189 FM IF chip makes a good AM synchronous demodulator. The circuit shown provides AGC, an S-meter output and synchronous detection and only needs the RF input. Alternate half cycles of the input to pin nine are inverted by the switching waveform generated in the chip, so that unidirectional half cycles of the modulated carrier appear at pin six and simply need a filter to remove the high frequencies to leave the demodulated AM. Potentiometer RV_1 sets AGC to suit the RF stages used and RV_2 adjusts the input to avoid overload. To start with, adjust the DC at pin six to 3V on the strongest signal.

R Gough
Shenstone Staffordshire



Using a CA3189 FM IF for synchronous detection of an AM signal.

INSTRUMENTS TO BUY

Buy top quality instruments direct from **Electronics World + Wireless World** and avoid disappointment. If you are not satisfied, return the goods and we will refund the purchase price*.

*Goods must be returned within seven days of receipt: must be returned in their original packing; must not be tampered with in any way, and must be returned in the condition in which they were received.

FREQUENCY COUNTERS

MX1010F and **MX1100F** are 8-digit frequency counters offering a broad range of features.
MX1010F: 1Hz to 100MHz, sensitivity of 15mV and resolution to 0.1Hz, data auto set, 10:1 attenuator, high impedance input – £129.00 plus VAT (£151.58).
MX1100F: 1Hz to 1GHz, features as MX1010F except ranges 70MHz to 1GHz and 50Ω impedance. £160.00 plus VAT (£188.00).

MX1010F MX1100F

MULTIMETERS

The **180 series** of high performance multimeters provide advanced features and are supplied complete with probes, battery and rubber holster. The case is dust and splash proof making it ideal in most environments. Designed to meet IEC348 Class II safety standard.
183: $3^1/_2$ digit large LCD display, ACV, DCV, ACA, DCA, resistance, continuity buzzer, diode test, hold, basic accuracy 0.5%. £33.50 plus VAT (£39.36).
185: $3^1/_2$ digit LCD, bar graph, ACV, DCV, ACA, DCA, resistance, continuity buzzer, diode test, hold, temperature (–40°C to 1370°C), capacitance (1pF to 40µF), frequency (1Hz to 200kHz), max min, edit, %, compare, basic accuracy 0.3%. £68.50 plus VAT (£80.49)
185: $3^1/_2$ digit LCD, bar graph, ACV, DCV, ACA, DCA, resistance, continuity buzzer, diode test, hold, temperature (–40°C to 1370°C), capacitance (1pF to 40µF), frequency (1Hz to 200kHz), max min, edit, %, compare, basic accuracy 0.3%. £68.50 plus VAT (£80.49)
187: As 185 except auto ranging. £75.00 plus VAT (£88.13).
285: As 185 except $4^1/_2$ digit true rms, basic accuracy 0.05%. £89.50 plus VAT (£105.16).

THE 180 SERIES

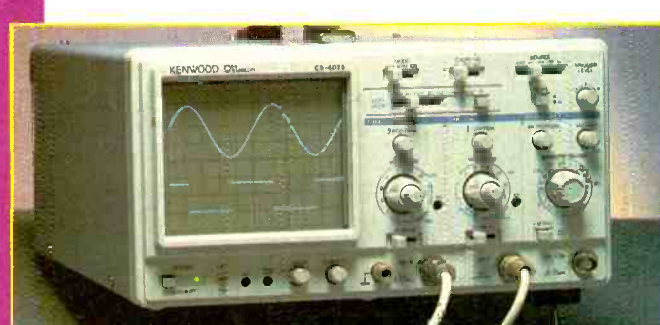
MULTIMETERS (2)

The **MX170B** and **MIC-6E** offer low cost measurement yet retain a large number of features. Supplied complete with probes.
MX170B: $3^1/_2$ digit LCD, compact size, ACV, DCV, DCA, resistance, diode test, low voltage battery test. £18.50 plus VAT (£21.74).
MIC-6E: $3^1/_2$ digit LCD, ACV, DCV, ACA, DCA, resistance, diode test, buzzer.
£33.50 plus VAT (£39.36).

MX170B MIC-6E

20MHz 2-CH OSCILLOSCOPE

The CS4025 20MHz dual trace oscilloscope offers a comprehensive range of facilities including a high sensitivity vertical amplifier providing from 1mV to 5V/div in CH1, ALT, CHOP, ADD, CH2 modes with inverse polarity on CH2. The horizontal timebase offers a sweep range of 0.5s/div to 0.5µs/div plus x10 sweep expansion and X-Y mode. Triggering can be auto or normal from vert, CH1, CH2, line or external sources with coupling provided for AC, TV-F and TV-L. The CS4025 is supplied complete with matching probes for £295.00 plus VAT (£346.62).



20MHz 2-CH OSCILLOSCOPE

PROGRAMMABLE POWER SUPPLIES

The PPS series of GPIB programmable DC power supplies offer high performance yet are extremely competitively priced using a 16 x 2 backlit LCD and 14 button keypad. All functions and conditions are easily selected and displayed. Overvoltage and overcurrent are selectable as is output enable/disable. Terminals for output and sense are provided on the front and rear to allow easy rack mounting.
PPS-1322: 0-32V 2A (GPIB) £375.00 plus VAT (£440.63)
PPS-2322: Dual 0-32V 2A (GPIB) £555.00 plus VAT (£652.13)

PROGRAMMABLE POWER SUPPLIES

MX2020

MX 9000

MIC-4070D

FUNCTION GENERATOR

The **MX2020** 0.02Hz – 2MHz sweep function generator with LED digital display offers a broad range of features. Output waveforms include sine, square, triangle, skewed sine, pulse and TTL. Lin and log sweeps are standard as is symmetry, DC offset and switchable output impedance from 50Ω to 600Ω. The digital display provides readout of the generators' frequency or can operate as separate 10MHz frequency counter. £175.00 plus VAT (£205.63).

LCR METER

The **MIC-4070D** LCD digital LCR meter provides capacitance, inductance, resistance and dissipation measurement. Capacitance ranges are from 0.1pF to 20,000µF plus dissipation. Inductance ranges from 0.1µH to 200H plus a digital readout of dissipation. Resistance ranges from 1mΩ to 20MΩ. Housed in a rugged ABS case with integral stand it is supplied complete with battery and probes at £85.00 plus VAT (£99.88)

FOUR INSTRUMENTS IN ONE

The **MX9000** combines four instruments to suit a broad range of applications in both education and industrial markets including development work stations where space is at a premium.

The instruments include:

1. A triple output power supply with LCD display offering 0-50V 0.5A, 15V 1A, 5V 2A with full overcurrent protection;
2. An 8-digit LED display 1Hz - 100MHz frequency counter with gating rates of 0.1Hz, 1Hz, 10Hz and 100Hz providing resolution to 0.1Hz plus attenuation inputs and data hold;
3. A 0.02Hz to 2MHz full featured sweep/function generator producing sine, square, triangle, skewed sine, pulse and a TTL output and linear or logarithmic sweep. Outputs of 50Ω and 600Ω impedance are standard features;
4. An auto/manual $3^1/_2$ digit LCD multimeter reading DCV, DCA, ACV, ACA, resistance, and relative measurement with data hold functions.

The MX9000 represents exceptionally good value at only £360.00 plus VAT (£423.00).

FG SERIES FUNCTION GENERATORS

The **FG500 series** sweep/function generators provide two powerful instruments in one package, a 6MHz or 13MHz sweep/function generator and an intelligent 100MHz frequency counter. The microprocessor based instruments offer sophisticated facilities yet remain extremely competitively priced. A menu driven display allows easy set up and operation. A 16 character by 2-line LCD display provides clear and unambiguous readout of generator output and frequency measurement.

FG-506: 2Hz to 6MHz sweep/function with 100MHz counter £325.00 plus VAT (£381.88)

FG-513: 2Hz to 13MHz sweep/function with 100MHz counter £482.00 plus VAT (£566.35)

Please send me the following instruments...

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I enclose a cheque/postal order/eurocheque to the value of £....................made payable to Reed Business Publishing Ltd or Please debit my Access/Visa/American Express/Diners Club with £.......................Card Number..

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Return to: Lorraine Spindler, Rm L333, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. Please allow 21 days for delivery.

CIRCLE NO. 138 ON REPLY CARD

Temperature transducer

Analog Devices's *AD592* is a two-terminal IC that puts out a current proportional to absolute temperature, at high impedance and with 1μA/K sensitivity. It needs no linearisation circuitry, voltage reference, bridge components or cold junction compensation and can be used over the same temperature range (-25°C to 105°C) as other kinds of transducer. Its high-impedance current output makes it invulnerable to the voltage drops and noise encountered on long lines and non-linearity is guaranteed less than ±0.1% over the whole range.

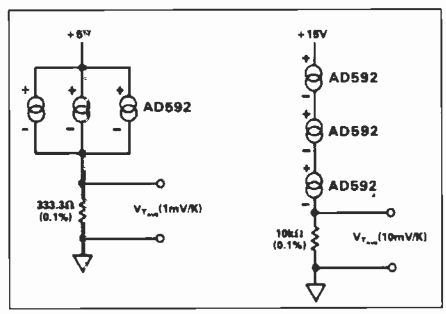
As an example of a very simple application, **Fig. 1** shows two circuits for measuring average and minimum temperatures of separated transducers. On the left, currents are added to give the average and on the right the current is proportional to the lowest temperature, since the coldest device limits current through the chain.

Figure 2 converts temperature to a 4-20mA current for use with 40V, 1kΩ systems. The *AD592* output is amplified to 1mA/°C and offset to give 4mA/20mA equivalent to 17°C/33°C. R_T is trimmed for correct reading at an intermediate temperature. The *AD581* is a 10V reference.

Using an *AD670* 8-bit A-to-D converter, a digital output for a microprocessor is obtained. **Figure 3** shows a circuit to resolve 1°C over the whole 130°C span for which the device is rated.

Figure 4 is a temperature controller to operate over the entire 130°C range: R_{high} and R_{low} determine the limits settable by R_{set}. The *AD581* keeps the set point constant and maintains 7V across the *AD592*. Resistor R_{hyst} and C introduce a measure of hysteresis to reduce the effects of noise.

Fig. 1. Analog Devices AD592 temperature transducers used to measure average (left) and minimum temperatures of a number of remote devices.



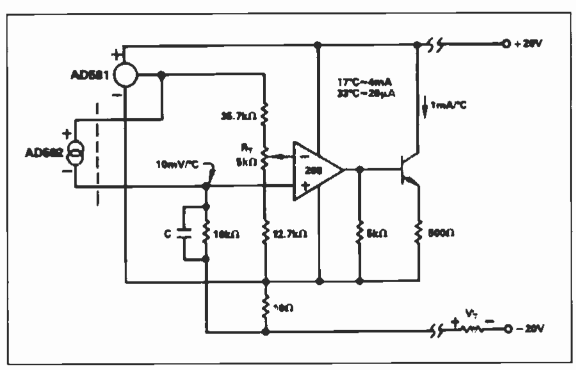


Fig. 2. Temperature measurement for a 4-20mA current-loop system, in this case between 17°C and 33°C, although the circuit will measure over the whole 130°C range of the AD592 if the 35.7kΩ and 12.7kΩ resistors are varied.

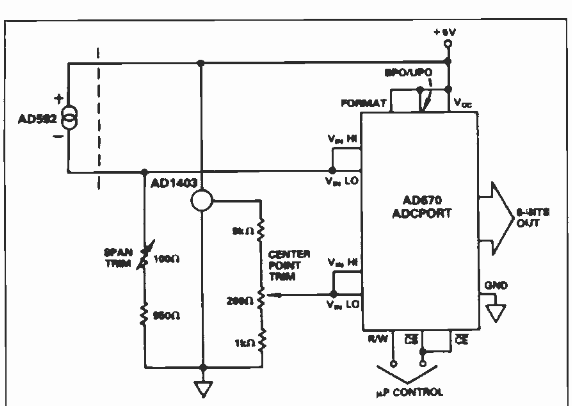
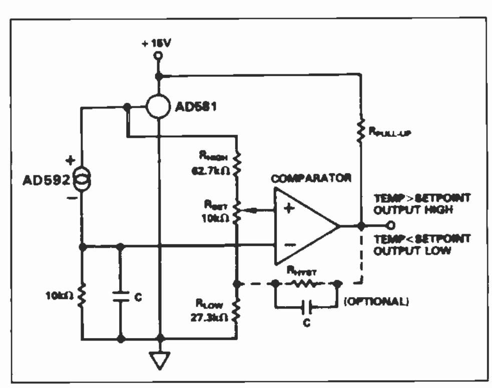


Fig. 3. If output to a microprocessor is needed, use this circuit, which resolves 1°C.

Fig. 4. A thermostat. The 10V AD581 reference keeps the bridge voltage constant. Resistor shown dotted may be needed to guard against noise.



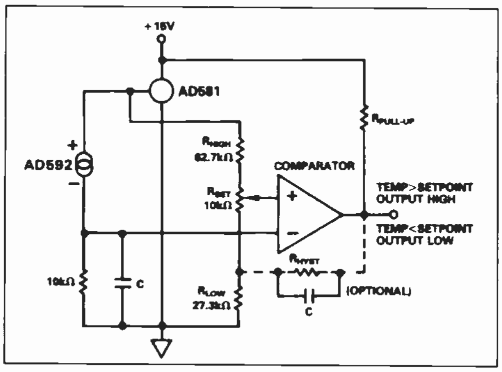


Fig. 5. For remote, multiple transducers, either a multiplexer or logic gates are usable, since the AD592's high output impedance tolerates their on resistance or output impedance.

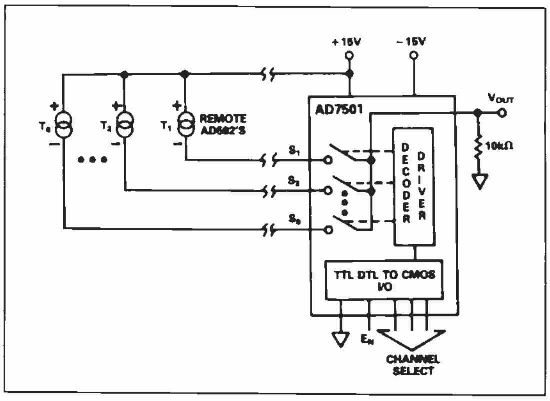


Fig. 6. When even more AD592s are to be read, this matrix will handle them and turns them off for idling.

If multiple, remote temperatures are needed, then the circuit in **Fig. 5** will do the job. The on resistance of a gate or a fet switch are not important, because of the high impedance of the *AD592*; the only important consideration is to keep 4V across it. Accordingly, an *AD7501* cmos multiplexer or a number of 5V logic gates can be used, but leakage currents in multiplexers and logic drives should be low. If a large number of transducers are needed, the matrix shown in **Fig. 6** is the answer. A decoder switches the supply to a column of transducers and a multiplexer looks at the rows; a 7-bit word determines which *AD592* is to be read. The multiplexer's enable input turns all transducers off.

***Analog Devices Ltd,** Station Avenue, Walton-on-Thames, Surrey KT12 1PF. Telephone 932 252320.*

Inexpensive radio data receiver

When the BBC's Radio 4 long-wave service moved from 200kHz to 198kHz, it was made to carry additional data in the form of phase modulation, to be used for remote time switching for lighting, for example, advertising display updates, radio clocks and even full-blown data transmission. Plessey, in its application note *AN86*, described a circuit to demodulate received data that can be built at a much lower cost than some available equipment.

Plessey's *SL6653* is a low-power (2.5mA at 2.5V-7.5V) IF/AF circuit for FM demodulation, containing mixer/oscillator, limiting IF amplifier and detector and was originally meant for cellular radio, cordless telephones and low-power radio. This and *TAB1043* quad op-amp, a fet and a bipolar transistor are the only active devices used.

Data rate is restricted to 25bit/s to avoid interference with the AM broadcast, and is therefore kept within a 0-50Hz band. Thirty blocks per minute are transmitted, each containing a cyclic redundancy check (CRC) error-checking code. Block 29 contains information on time, day, month, year, leap year and local offset. The BBC Research Department at Kingswood Warren, Surrey KT20 6NP has a publication on LF data.

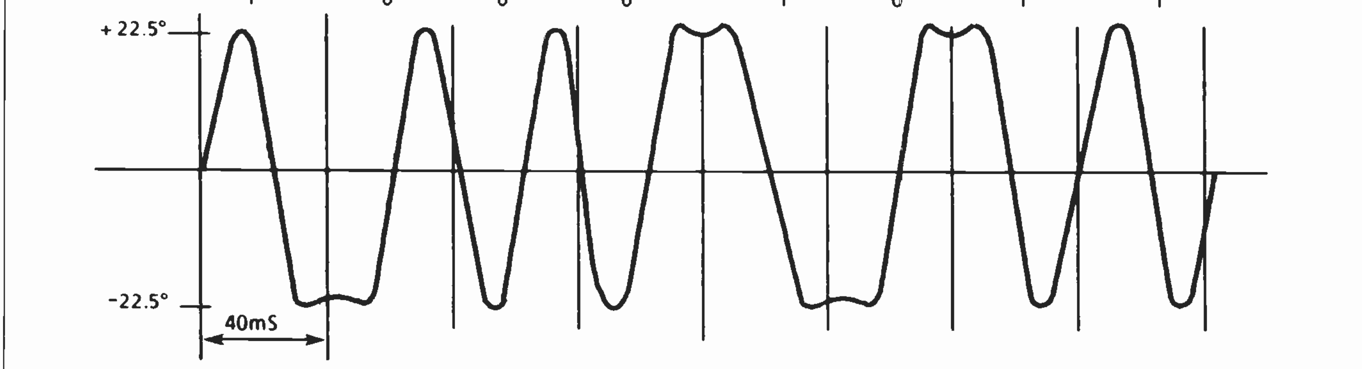
Figure 1 shows the form of the ±22.5° data modulation on Radio 4, from which a 1.8V pk-pk phase signal is to be obtained and **Fig. 2** is the radio part of the circuit.

Aerial signal goes to a fet buffer, which provides matching between the aerial and the following crystal filter, a special unit by AEL Crystals Ltd of Horley to reduce the level of interference from switched-mode power supplies in computer terminals and television receivers. Filtered input at 198kHz now goes to the *SL6653* limiter, from which it emerges as a square wave at pin 1, is phase-shifted by the quad circuit and goes back to the FM demodulator. Transistor T_3 buffers the demodulator output to drive the filter in **Fig. 3** and its test point *TP2* is also useful in adjusting the tuning of the quadrature circuit.

The first op-amp in Fig. 3 acts as a Sallen and Key low-pass filter to prevent trouble with AM or spikes affecting the phase output. This is followed by an amplifier with a gain of 45 to give a 1.3V-2V pk-pk phase signal at *TP4*.

Since frequency is a rate of phase change and the data signal is a phase shift, the output from the demodulator is differentiated and an integrator is needed to make a phase demodulator. The third quarter

Fig. 1. Long-wave radio data signal on BBC Radio 4 is ±22.5° phase modulation at 25bit/s.



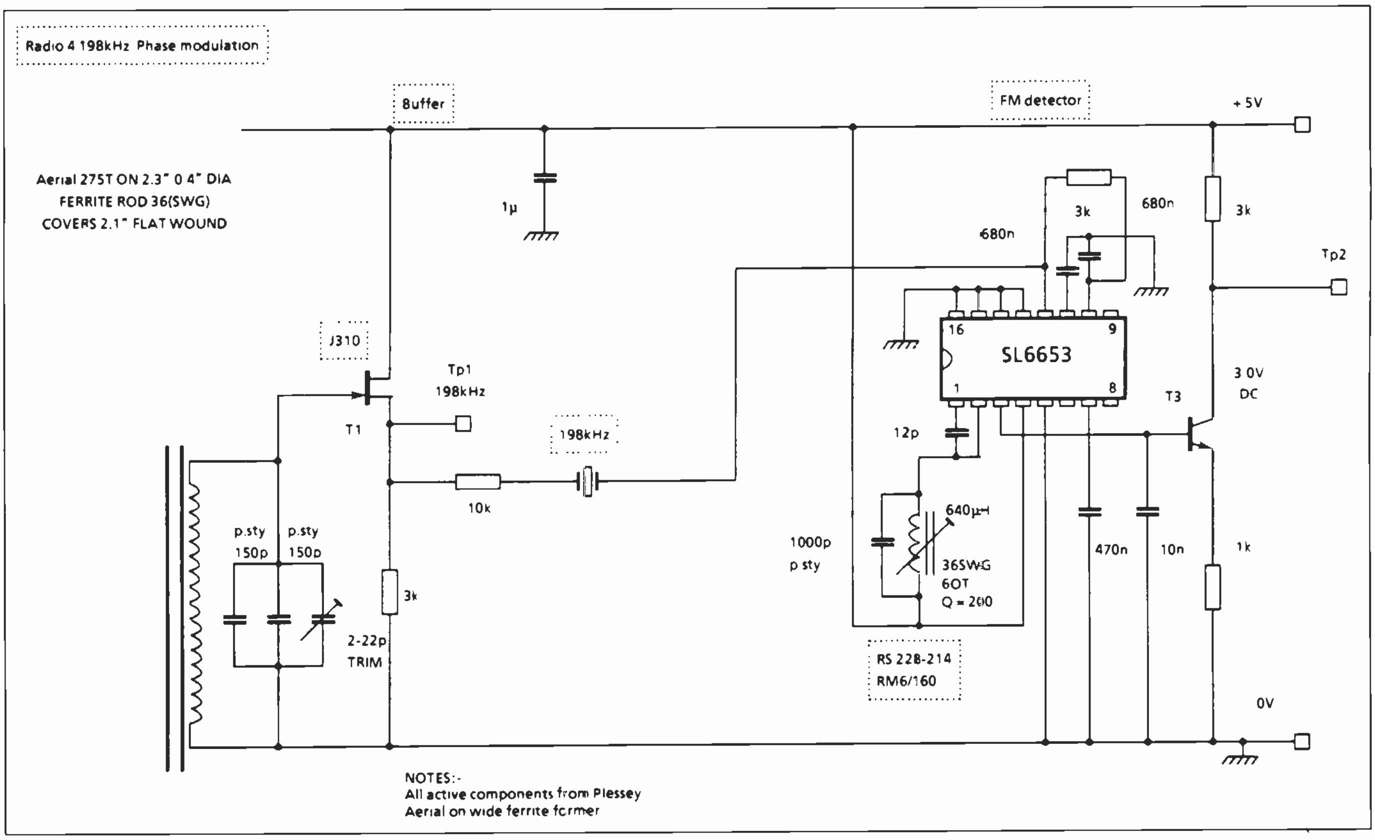


Fig. 2. Receiver section of data demodulator. Testpoint 1 is for aerial tuning and Tp2 for quad coil tuning.

of the *TAB1043* is therefore a simple integrator, its DC reference still being referred back to the quad detector and drift in tuning tracked by the integrator. From this op-amp, the output is the original signal seen in Fig.1.

A comparator now removes the varying DC component from the signal by comparing a smoothed version of it (27kΩ/6.8µF) with the raw input from the integrator. Output signals consist of two frequencies of 12.5Hz and 25Hz, from which clock and data are recoverable, preferably by using a microprocessor to sample the data to find an edge in the middle of each 40ms clock interval. A rising edge denotes a 1 and a falling edge a 0.

Plessey Semiconductors Ltd, *Cheney Manor, Swindon, Wiltshire SN2 2QW. Tel: 0793 518000.*

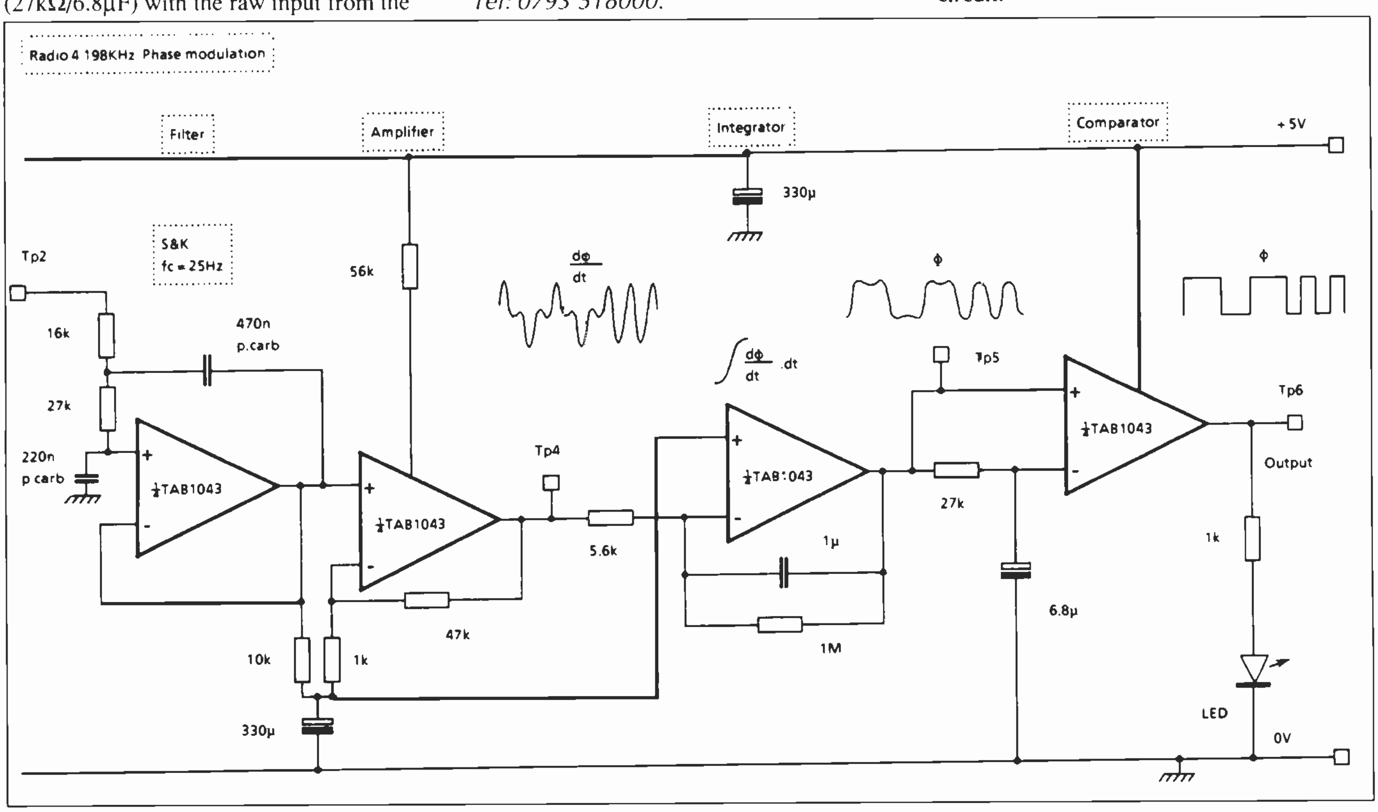


Fig. 3. Demodulator section. TAB1043 has adjustable open-loop gain and requires the 56Ω resistor. Adjust the amplifier's 47kΩ feedback resistor to allow for varying Q in the quadrature circuit.

CIRCLE NO. 139 ON REPLY CARD

ACTIVE

A-to-D & D-to-A converters

12-bit multiplying D-to-A. In one 24-pin narrow dip, Maxim's *Max514* holds four independent multiplying serial 12-bit converters and dissipates 10mW maximum. The converters have 1.5LSB gain accuracy, 1LSB relative accuracy, are monotonic and exhibit a settling time to 0.5LSB of less than 1µs. Each converter has its own reference input and a common 5V supply. Maxim Integrated Products, 0734 845255.

Quad D-to-A. Maxim's new *Max527* quad multiplying 12-bit voltage-output D-to-A converter for 5V systems replaces four single converters and four output amplifiers; it has guaranteed monotonic 12-bit performance with ±0.5LSB relative accuracy over the commercial, industrial or military temperature ranges for all outputs. With 50mW power consumption, the device offers a 5µs settling time and THD+N is less than 0.024% with 850mV reference signal up to 100kHz. Bandwidth to –3dB is 700kHz. Maxim Integrated Products UK, 0734 845255.

Discrete active devices

Micropower high-side driver. Linear's *LTC1156* quad high-side n-channel mosfet driver draws only 95µA from 5V (16µA when all four outputs are off). It has short-circuit protection and needs no extra components to drive four n-channel mosfet gates to 12V from the 5V supply. Each gate is independently controlled and ramped to eliminate interaction and to reduce RFI and EMI. The short protection comes on in 10µs, but can be delayed to handle difficult loads. Linear Technology (UK) Ltd, 0276 677676.

Small 1A mosfet. Zetex's *ZVN4306A* mosfet passes 1.3A continuous current but measures 4mm by 2.4mm by 4.7mm and is believed to be the world's smallest. On resistance is 0.22Ω at 3A and 10V gate drive. Peak current is 20A, breakdown 60V and dissipation around 1W. With turn-on and turn-off times of 8ns and 30ns, rise and fall times of 25ns and 16ns, the device is suitable for use in HF DC-to-DC converters. Zetex plc, 061 627 4963.

Transient suppression. Zener/avalanche diode fast-acting voltage clamps from Zetex, the *1.5KE* and *P6KE*, handle large pulse currents and up to 1.5kW and 600W respectively. Voltages available in the *1.5KE* series are 6.8-400V, while in the *P6KE* types the range is 6.8-200V. Both come in unidirectional or bidirectional form with 5% tolerance. *P6KE* devices operate in less than 1ps. Zetex plc, 061 627 4963.

Linear integrated circuits

Dual comparator. Common-mode range of -4V to 8V and propagation delay of 3ns are features of the Signal Processing Technologies *SPT9691* dual jfet-input comparators. Tracking bandwidth is 300MHz at -3dB, open-loop gain 60dB, input C 1pF, differential input voltage range ±10V and differential input R 2GΩ. No external buffering is needed Ambar Cascom Ltd, 0296 434141.

Dual op-amp. Input circuitry of Analog's *OP275* dual op-amp combines Jfet and bipolar techniques to obtain the advantages of both: low distortion and voltage noise from the bipolar circuitry; and fast slewing, low power and wide dynamic range from the jfet. THD plus noise is 0.0006%, with 6nV/ Hz voltage noise density. Current noise density is 1.5pA/ Hz and input offset voltage less than 200µV; slew rate is 22V/µs. Analog Devices, 0932 253320.

700MHz buffer amplifier. Harris's *HFA1110* uses the company's *UHF1* 8GHz bipolar process to produce a closed-loop buffer amplifier with a 700MHz bandwidth and selectable gain of +1, –1 or +2 with no extra components. Slew rate is 2500V/µs and settling time to 0.02% is 7ns. A gain error of 0.01V/V and gain flatness with frequency of 0.04dB up to 50MHz make the device very suitable for handling videol. Harris Semiconductor (UK), 0276 686886.

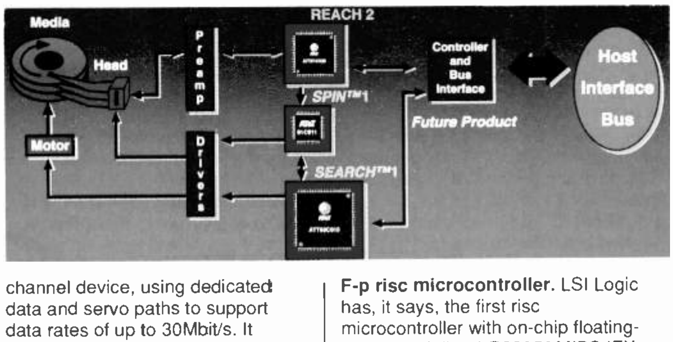
Voltage reference. *LM4431* is claimed by National Semiconductor to be the smallest ever voltage reference and is contained in an SOT-23 transistor package. No external capacitor is needed. Intended for high-volume manufacturing in areas such as disk drives and personal computers, it has a 2.5V fixed reverse voltage breakdown with ±2% accuracy, 0-70deg.C temperature range, 35µV RMS output noise and 100µA-15mA operating current. National Semiconductor, 0793-697466.

Logic building blocks

Elapsed-time counter. Dallas's *DS1603* elapsed-time counter module counts and records the time during which power is applied to it and the system in which it works. It also functions as a real-time clock and can be reset and used to measure equipment use during a set time. Accuracy is within ±2min/month and, since it has a 1Hz output, is usable as a clock for other functions Its three-wire serial link comprises data line, clock and reset. Dallas Semiconductor Corp, 021 782 2156.

PC i/o. *Super I/O II* by National Semiconductor is XT/AT-compatible and is meant for use in i/o-intensive applications. The low-power device offers floppy-disk control for four 2.88Mbyte drives, two uarts, a bidirectional parallel port, an IDE interface for two hard drives, XT/AT address decoding and configuration registers to select addresses, enable, disable or power-down. Super I/O III is a pin-compatible unit, but with more powerful uarts. National Semiconductor, 0793-697466.

Disk-drive chipset. Three chips from AT&T have the performance and low enough power consumption to put hard drives in 2.5in formats or smaller. No glue logic and few other components are needed. *REACH2* is the read-



channel device, using dedicated data and servo paths to support data rates of up to 30Mbit/s. It integrates read channel, synthesiser and servo demodulator. *Search1* is the positioning and spindle servo controller, including clock control and power down; and *Spin1* is the servo data converter, interfacing to either an 8-bit or 16-bit microcontroller. AT&T Microelectronics, 0732 460424.

Memory chips

Fastest 1Mbit eprom. Atmel claims its *AT27HC1024* 1Mbit erasable rom, with a read-access time of 45ns, to be the fastest available. Zero-wait-state operation is possible and the device offers the Atmel rapid-programming algorithm for accurate programming in 100µs/byte. Current consumption is 80mA at 10MHz (active) and 8mA (standby). It is organised as 64K by 16bit. Atmel (UK) Ltd, 0276 686677.

Microprocessors and controllers

High-speed correlator. Harris's *HSP45256* is a 33MHz, multiple configuration binary correlator, a device which compares a reference signal with an input and assigns a score based on the level of agreement between the two. Such a device is needed to detect, recognize or synchronize data among other data or noise in, for example, error detection and pattern recognition. There are 256 taps, which can be arranged in many different ways, from a single 256 tap to two independent 4 by 32 tap correlators, up to 16 devices being cascadable. Double registers allow new data input while correlation is taking place. Harris Semiconductor (UK), 0276 686886.

F-p risc microcontroller. LSI Logic has, it says, the first risc microcontroller with on-chip floating-point capability. *LR33050* MIPS IFX, the integer and floating-point accelerator, adds a MIPS *R3010*-compatible f-p accelerator to the *LR33000* self-embedding processor. It incorporates the glue logic of a true risc device and the f-p acceleration needed by embedded systems on one chip. Clock speeds available are 25, 33 and 40MHz and 5K caches are on-chip. LSI Logic Ltd, 0344 426544.

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

Upgradeable chipset. Computer system in which graphics and systems logic are merged, the *WinCHIPSet* from Chips & Technologies, allows users to see some sort of visible result from expensive upgrading exercises. When faster microprocessors or cache memories are added, this chipset keeps the graphics system in step with the increased power. It incorporates the Wingine Windows accelerator and the *CS4021* systems logic, which supports 386DX, Super 386, 486SX, 486DX and 486DX2 systems running at up to 50MHz. Thame Components Ltd, 0844 261188.

Opto sensor. An alternative CCD image sensors and photodiode arrays, TI's *TSL214* integrated opto sensor needs only a 5V supply and a pair of timing pulses. Output video reference and S/H circuitry are incorporated and the devices can be combined in parallel or serially. Data rate is from 10kHz to 500kHz. The imager consists of an addressed line of 64 charge-mode pixels, analogue and digital elements. There is also the *PC404* evaluation kit containing a sensor, drive board, lens and all clock and interface logic. Texas Instruments, 0234 223252.

Risc microprocessor. *ARM610*, a new member of VLSI's 32-bit risc micro family, gives 29K Drystones at 25MHz while dissipating less than 600mW. It consists of the 32bit *ARM6* risc macrocell, memory management, 4K cache, a write buffer and full boundary scan. Power consumption is quoted at 4.5mA/MHz, low enough for inexpensive battery-powered equipment. Its development software runs on Sun, dos and Mac systems. VLSI Technology Ltd, 0908 667595.

Mixed-signal ICs.
DC motor drive. A driver for the three-phase brushless motors in 5V Winchesters, the *A8902SLB* IC from Allegro is a 3-phase dmos back-EMF sensing circuit giving 1A, 1Ω outputs. Ground clamp and flyback diodes are included and there is thermal shutdown. No Hall-effect position sensors are needed, since internal circuitry provides start and run sequencing and internal linear current control used with external components allows frequency-locked speed control. A serial port enables a user to program reference frequency count number. Allegro Microsystems, 0932 253355.

Teletext processor. In addition to its capability of processing all teletext standards worldwide, ITT's *TPU2740* displays graphics from other sources. Its 3-bit RGB graphics enables display of 540 by 260 pixels and the high resolution enhances all screen functions and makes it suitable for multi-media use. Formats supported include WST, Caption and NABTS on pal, mac, NTSC and Secam and there is an optional 4Mbit storage capacity to store 200 pages, a 72Mips risc processor allowing real-time display. ITT Semiconductors, 0932 336116.

Speech chip. In addition to using the Oki ADPCM algorithm, the OKI *MSM6650* speech synthesiser accepts a linear 8-bit format, sampling at 4-32kHz. Filters are automatically adjusted and a total of 127 phrases are addressable. An internal 12-bit A-to-D converter and low-pass filter produce a good-quality output. In different versions, rom size varies from 288kbit to 2Mbit. Complete sentences can be coded in an "edit rom" to contain up to eight phrases, so that one address byte and one common byte give the position of the data in rom. Depending on rom size, a maximum of 130 seconds of speech can be generated, or up to 70 minutes with a rom-less version and 64Mbit of external memory. Oki Semiconductor (UK) Ltd, 0753 516577

Fast synthesisers. SSS has two new fast synthesisers; *SSSB138/139*, which work from 500MHz to 3.6GHz (*138*) and 2-4GHz (*139*). Each has a high frequency comparison frequency up to 50MHz for fast look-up and spectral purity from the VCO. All synthesiser functions are present, with the exception of charge pump, VCO and crystal oscillator. Division is controlled by cmos levels from 12 to 64. The devices are intended for multi-loop, fast-hopping synthesisers. Swindon Silicon Systems Ltd, 0793 614039.

Optical devices

High-power IR. Gallium aluminium arsenide infrared emitters from Opto Diode will put out up to 6.5mW. Three, six or nine chips in each package, working at 880nm, are on TO66 gold headers and beryllium oxide substrates for best heat dissipation. Hero offers a designers' data book. Hero Electronics Ltd, 0525 405015.

Transmissive sensors.
TCYS5201/6201 transmissive optical sensors from Telefunken Electronic have an aperture of 0.5mm for high resolution in paper positioning or as the sensors in shaft encoders. They work at a wavelength of 950nm and have TTL-compatible Schmitt open-collector outputs, which rise and fall in 50ns and 20ns at a maximum of 3kHz. The 5201 is a snap-fitting type, while 6201 is screw mounted. Transmitter and receiver are available separately as the TCZS 8100. Siliconix Ltd, 0635 30905.

Colour CCD. Sony's *ICX045BKA-6* is an interline transfer CCD image sensor for use in pal 1/3in colour video cameras. Separate yellow, cyan, magenta and green complementary colour mosaic filters give 6dB sensitivity improvement over the earlier unit and the use of Sony's hole accumulation diode (HAD) gives low smearing and high anti-blooming. Effective image area is 500 by 582 pixels. Field-integration read-out and a variable-speed electronic shutter. Sony Components, 0793 618492.

Programmable logic arrays

Fast, sub-micron gate array. *ATL80* gate arrays from Atmel are fabricated in 0.8micron cmos; 11 arrays in the family provide gate counts upt to 180,000 and up to 490 pins. Delay is 200ps and supplies from 3V to 5.5V, the devices still operating at 2.4V. Popular cad software is relevant and the current cell libraries for 1micron ATL designs are compatible. Atmel (UK) Ltd, 0276 686677.

2000-gate FPGA. QuickLogic claims that its *QL12*16* high-speed, 2000-usable-gate field-programmable gate array is the only such device available that will provide zero-wait-state support for 33-50MHz microprocessors. It can be used instead of masked arrays or to integrate high-speed pals. It is organised as a 12 by 16 array, each cell containing a flip-flop and enough logic to form two latches, resulting in 576 storage cells. QuickLogic, (USA) +1 408 987-2000

Power semiconductors

Power mosfet. Philips's *BUK101-50GL* is meant for use in car electronics and is a logic-level power mosfet, protected against overload and high temperature. Salient virtues are 50V off-state drain/source volts, continuous 20V/26A working, 75W dissipation, continuous 150deg.C junction temperature and a 60mΩ on resistance. Control and logic supplies are derived from the input. Gothic Crellon Ltd, 0734 788878.

Linear regulators. Three units in the Semtech range of 2A linear regulators produce outputs of 5V, 12V and 15V, the fourth device having an adjustable output from 4V to 30V; all the models have a grounded case. Temperature stability comes from the use of a band-gap reference and the power management circuitry limits internal dissipation up to the rated input voltage maximum of 40V and 20W dissipation. Semtech Ltd, 0592 773520.

PASSIVE

Passive components

SMT pots. A true surface-mounted potentiometer from Murata, the *POT0102W*, is a multi-turn type capable of withstanding wave soldering and cleaning. It measures 6.35 by 6.35 by 4.3mm and has a resistance range of 10Ω to 1MΩ ±10%. Units are rated at 0.25W and have a maximum wiper current of 100mA, working at 200V. Murata Electronics (UK) Ltd, 0252 811666.

Filters

Wide-range filter. Two-channel programmable filter from Kemo, the *VBF10*, offers a million-to-one range of cut-off frequencies: 0.1Hz-102.3kHz. Each channel is gain-programmable from −11dB to 70dB, can be connected in series or parallel and can be controlled by RS232 or GPIB. Additional units will expand the system to give up to eight pairs of channels and can be internally connected to provide bandpass, bandstop, lowpass and highpass characteristics. All functions are controlled from the front panel. Kemo Ltd, 081 658 3838.

Saw filters. Murata's first surface-mounted saw filters are known as the *SAFC* series, which measure 9.1 by 7.1 and 13.3 by 6.5mm. Centre frequencies are from 278 to 283MHz for use in pagers, and 71-183MHz for mobile telephones. They are suitable for reflow soldering and organic washing. Murata Electronics (UK) Ltd, 0252 811666.

RFI suppression. Corcom Q filters are meant to reduce RFI in equipment working at 10kHz and above; in particular, switching power supplies. Filters can be mounted on chassis or back panel, with or without IEC connectors. Maximum current rating is up to 6A, depending on the type. The list of approvals includes VDE, CSA, UL and Semco. Sterling Components Ltd, 0753 820753.

Hardware

Heat-sink attachment. A self-locking, solderable stand-off tag from Aavid gives a positive snap-in heat-sink attachment that retains heat sinks in position during flow soldering and allows the routeing of PCB tracks under the heat sink. Its bifurcated tip snaps into a 0.093in plated-through hole, where it protrudes less than 0.06in beyond a 0.0625in board. Aavid Engineering Ltd, 0279 520022.

Instrumentation

Digital wattmeter. Yokogawa's *2534* is a single-phase instrument for measurements at DC and in the range 10Hz-20kHz. Its sampling process renders it immune to waveform distortion, so that it can be used in circuitry such as switched-mode supplies to an accuracy within 0.5% of full scale. Three displays indicate voltage, current and power simultaneously, with the ability to integrate the displays to give active power, reactive power and energy consumption. Frequency measurement is available from 4Hz to 22kHz and the instrument has optional GPIB or RS-232C interfaces. Martron Instruments Ltd, 0494 459200.

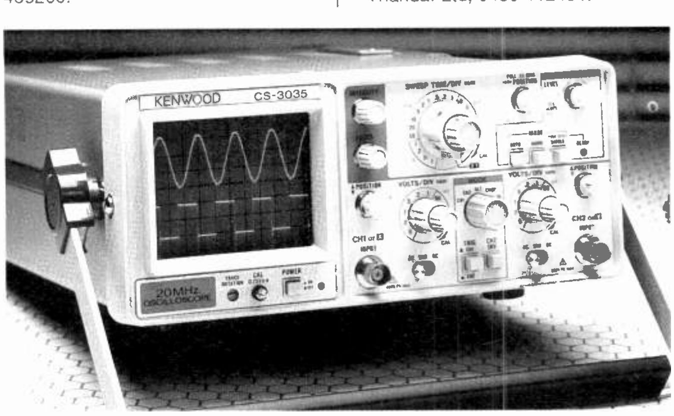
Four-in-one. *MX9000* from Saje contains four of the most frequently used instruments in one case: a power supply, a frequency meter, a function generator and a multimeter. The PSU provides 0-50V at 0.5A, 15V 1A and 5V 2A, with an LCD. An eight-digit counter measures 1Hz-100MHz with data hold. Sine, square, triangular, skewed sine, pulse and TTL outputs and a log/linear sweep into 50Ω or 600Ω from 0.02Hz to 2MHz are available and the multimeter measures direct and alternating voltage and current, resistance and relative measurements. Saje Electronics, 0223 425440.

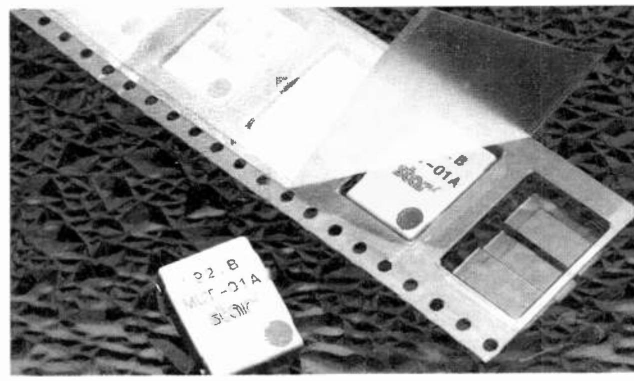
12in oscilloscope. Dual-trace oscilloscope from Thurlby-Thandar has a 310mm, medium-persistence tube and is meant for use with a sweeper to display frequency response. Model *12F* displays four traces on the 8kV tube, including two from the sweeper, and the medium trace persistence takes care of any flicker; a clamp holds the baseline stable. Sensitivity is 1mV/div. and –3dB bandwidth 10kHz. Thurlby-Thandar Ltd, 0480 412451.

Power supplies

High-voltage converters. Brandenburg 3.5kV DC-to-DC converters in the *390* series now have an output ripple less than 300mV pk-pk; 12V versions have been added to the original 24V types. Output is from 10V to 3.5kV, either positive or negative, 5W being available at the maximum voltage setting. Temperature coefficient is 100pp/°C and the units measure 95 by 50 by 19mm. Output is adjustable by potentiometer or by an external 0-10V potential. Astec High Voltage, 0384 440044.

20MHz oscilloscope. Mains/battery oscilloscope from Thurlby-Thandar measures only 8.5 by 3.5 by 11.7in, but provides 20MHz bandwidth, 1mV/division (6.35mm) sensitivity and a 0.2µs/div. sweep speed, expandable to 40ns/div. Vision-signal sync. is provided. Thurlby-Thandar Ltd, 0480 412451.





SM audio transducer. Star Micronics's *MUT-01A* transducer is claimed to be the world's first intended for automatic handling and surface mounting using IR reflow soldering. It is an electromagnetic device, measures 14 by 11 by 3mm and has its sound port on one end. From a 1.5V square wave, it produces an 85dB 3.2kHz tone. Roxburgh Electronics Ltd, 0274 281770.

Computer batteries. Designed for use in Compaq Deskpro 286/386 and 386 portables, Tadiran *TI5280* and *TL5283/C* user-replaceable batteries are fitted with 4A connectors and 200mm leads. Type *TI5242/W* meets real-time clock requirements and is usable with Toshiba, Tandy, Tatung, Wyse and Mitac computers. ESD Electronic Services, 0279 626777

DC-to-DC converters. A range of converters from FR starts with 1W single and dual output types working from 5V or 12V input and extends to 30W versions with single, dual and triple outputs and 4:1 input voltage ranges from 9V to 72V, each being fitted with an internal filter to reduce conducted noise. A number of cases are offered that can be supplied in black-coated copper to cut down on radiation. FR Electronics, 0202 897969.

DC-to-DC converters. Fixed-frequency converters in the *NFC10* series from XP give a 10W output in the ranges 9-18V, 18-36V and 36-72V DC, with single and dual outputs from ±5V to ±15V. The series meets EN60950 safety regulations and its MTBF is better than 1,000,000 hours. XP plc, 0734 845515.

Radio communications products

Stable microwave oscillator. Anglia's *KDO-150* 8-18GHz oscillator is meant for circuits needing great frequency stability and the facility of digital frequency control as well as the manual variety. Mechanical tuning gives ±25MHz adjustment. Anglia Microwaves Ltd, 0277 630000.

Transducers and sensors

Air-flow sensors. Sensor modules in the *TRMF* series by Murata use the change in resistance of a platinum element caused by air cooling to measure the flow velocity from 0 to 10m/s, a second sensor compensating for ambient temperature. Operating temperature range is 0-60°C in relative humidity up to 95%. Signal processing is incorporated. Murata Electronics (UK) Ltd, 0252 811666.

COMPUTER

Computer board level products

Transputer graphics. *IMS B437* is a credit-card sized transputer module giving graphical output, configurable to provide a variety of formats, from television standards with sync. and interlacing, to 1024 by 1024 displays. Machine vision and image-processing systems use the *B437* with the *B429* video image-processing module. The internal colour video controller is programmed by the transputer to generate the required timing and display resolution to make it compatible with any monitor. Support software is in Ansi C and Occam. Inmos, 0454 616616.

Computer systems Smallest PC.

MicroBox is an industrial PC chassis measuring 8.5in by 6in by 3.5in and is

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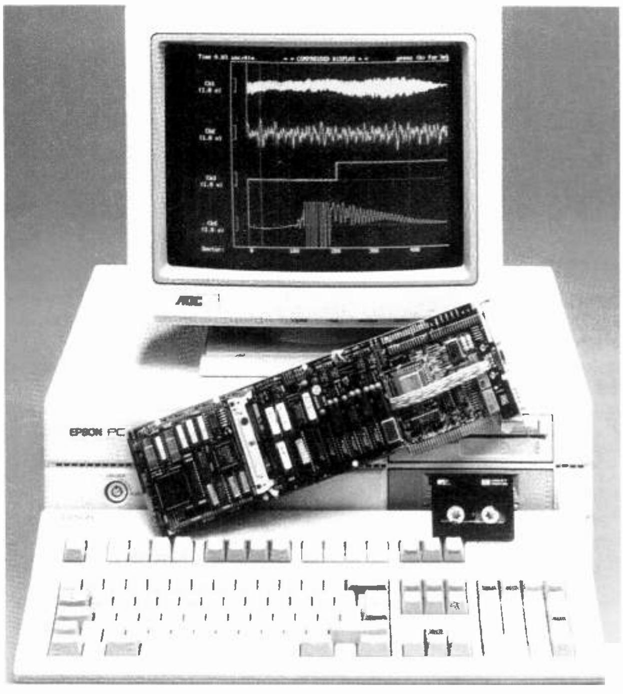


On and off-line DSP. For both real-time processing and off-line data reduction, Signalysys's *ACP-store* sends digitised signals direct to hard disks or tape and in parallel to a PC. Inmos standard modules provide up to eight channels of anti-aliased 16-bit data acquisition at 100ksample/s, synchronously sampled, *T800* transputers give real-time processing and the data reduction; the PC's keyboard and display being the only components in use. The whole system is contained inside the PC or in an expansion chassis. Signalysys Ltd, 0296 631306.

designed to accept three half-length cards in a passive AT backplane. 286 or 386SX cards are on offer, as is a range of silicon disk choices. External floppies or hard disks are connected via chassis sockets. Processors include serial and parallel ports, disk controllers, a watchdog timer and memory, two expansion slots being available. Fairchild Ltd, 0703 559090.

Development and evaluation

PIC16C5X programmer. Low-cost programming for the Arizona *PIC16C5X* range of microcontrollers is made possible by the ASL programmer, which costs £99. It consists of a free-standing PCB and needs an 18V-30V PSU, connecting to a PC via RS232. PC software generates pull-down menus for all the usual functions. Farnell Electronics Ltd, 0532 636311.

PC image compression. C-Cube's *CL550* JPEG processor forms the core of a still-image board development kit which allows the production of low-cost JPEG image-compression boards for PCs. Such boards provide up to 10 times the compression/decompression performance of software solutions at about three times the cost. The JPEG still-image board runs at 10MHz, supporting data compression rates of over 1Mbyte/s, handling a 24-bit, 640 by 480 image in around 0.7s when used on a 386SX PC. Kudos Thame Ltd, 0734 351010.

Graphics editor. For use in asic design, the *Iced-32* PC-based graphics editor from SMS reduces the cost of layout design by placing a polygon editor on the PC; there is also a design-rule check. CALMA and CIF i/o and up to 100 layers of cell nesting are provided, with wire and polygon editing, cut and merge commands and the use of text and lines for annotation. *Iced* is XT/AT-compatible and outputs to most of the standard laser, jet and dot printers. Silicon Microsystems Ltd, 0666 824844.

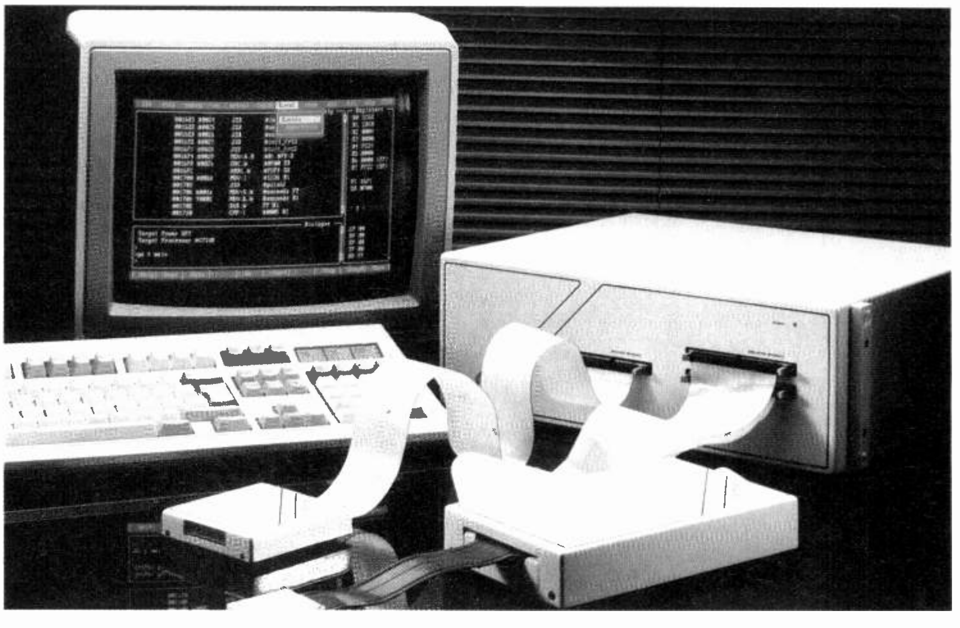
Software

Virtual instruments. *Amps* from Bores is a combined hardware and software package that provides a Windows-based approach to instrumentation and DSP hardware for analogue i/o and real-time processing such as filtering or spectral analysis. "Instruments" include two kinds of oscilloscope ("knobs" or "push-buttons"), a data recorder to disk and a spectrum analyser. Filter design incorporated allows one to make digital filters as required and to insert them into the application in use. Bores Signal Processing, 0483 740138.

Image compression. Fractal Transform image compression software from Iterated Systems can be integrated into users' dos or Windows applications by means of the SDK v.1.0 Software Compressor with a single function call. Decompression is integrated by way of an existing software development kit. The company also offers SDK v.1.0 Resolution Enhancement, which is also integrated into dos and Windows to enhance image resolution without compression, images being enlarged and endlessly zoomed without pixelation. Iterated Systems Ltd, 0734 880261.

Spox 1.4. Spox, the operating system for signal processors, has been upgraded to version 1.4 for *TMS320C30/31* processors; other processors are not affected. Improvements include support for Texas Instruments's latest compiler, which provides for code inlining of seldom-used functions and several i/o enhancements. Spox will run on PCs or Sun workstations, the versions being Spox/dos and Spox/Sun. Loughborough Sound Images Ltd, 0509 231843.

DSP development. Based on Free Software Foundation's GNU *C* compiler, Motorola's new software development toolkit for *DSP56000/1* digital signal processors enables high-level language access to DSP development, by-passing assembly-level programming. The GNU *C* compiler increases program execution speed by a factor of nearly two and supports floating-point emulation, eliminating the need to scale results. A *C* source-level debugger and documentation are included in the kit. Motorola Ltd, 0908 614614.



H8/330 emulator. Emulation for Hitachi's *H8/330* microcontroller is provided by the *EM8330*, which allows software and hardware debugging at up to 10MHz without using the microcontroller's internal resources. An unlimited number of breakpoints can be set in any memory location and four hardware event comparators look for combinations of address, data and user probes. There is 8K by 1024bit of trace memory to capture execution information in real time. Hitachi Europe Ltd, 0628 585000.

CIRCLE NO. 140 ON REPLY CARD

Bootstrap base to bridge building

High input impedances are required for bridge detector circuits used in measuring small capacitances. Ian Hickman bootstraps his way to a solution.

In a bridge detector both inputs of the detector amplifier should have such a high input impedance that even on extreme bridge ratios, for example when measuring very small capacitances, they do not load the bridge arms.

The detector amplifier should also have a very high CMRR (common mode rejection ratio). This is particularly important when the impedances of the lower arms of the bridge are much higher than those of the upper arms, since the difference signal that has to be detected rides on a much larger common mode component.

Bootstrapping is a good way to achieve the necessary high input impedance for such applications – given a suitable circuit design. Simply view a square-wave source via a high series resistance to obtain a quick guide as to whether the bootstrapping is effective over a range of frequencies.

A very high input impedance at 0Hz – a high input resistance – is provided by any jfet input or cmos op amp. For instance the RCA bimos *CA3130* op amp which has been around since the 1970s features a typical input resistance of 1.5TΩ or 1.5 10^{12}Ω. Input characteristics of some of the wide range of Texas Instruments op amps are as shown in **Table 1**.

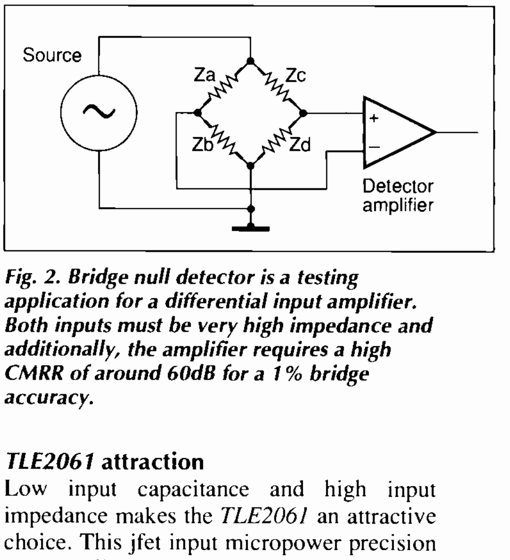


Fig. 2. Bridge null detector is a testing application for a differential input amplifier. Both inputs must be very high impedance and additionally, the amplifier requires a high CMRR of around 60dB for a 1% bridge accuracy.

TLE2061 attraction

Low input capacitance and high input impedance makes the *TLE2061* an attractive choice. This jfet input micropower precision op amp offers a high output drive capability of ±2.5V (min) into 100Ω on ±5V rails and ±12.5V (min) into 600Ω on ±15 volt rails, while drawing a quiescent current of only 290μA.

The device operates from V_{cc} supplies of ±3.5V to ±20V, with an input offset voltage as low as 500μV (/BC version), while its decom-

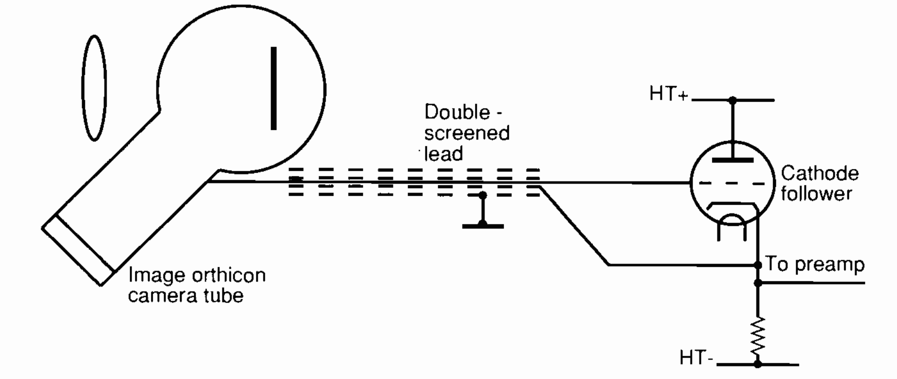


Fig. 1. An early application of bootstrapping. The camera signal is connected to its preamplifier via a double screened cable, the inner screen of which is driven by the output of the cathode follower buffer stage. Since the gain of the latter is very nearly unity, there is no ac voltage difference between the inner conductor and the inner of the two screens, so the signal does not 'see' any cable capacitance to ground.

Bootstrapping Blumlein

Bootstrapping is a powerful technique, long part of the circuit designer's armoury. AD Blumlein is often credited with its invention, in connection with his pre-war work at EMI laboratories developing the 405 line television system. The technique enabled the signal lead from the TV camera tube to its preamplifier to be screened, without adding so much stray capacitance as to reduce the signal's bandwidth.

Table 1. Input characteristics of TI op amps.

Device	C_{in} pF	I_{bias} (typ at 25K)	R_{in} typ
TLC27L9, TLC277	not quoted	0.6pA	1TΩ
TLC2201	not quoted	1pA	not quoted
TLE2021 (bipolar)	not quoted	25nA	not quoted
TLE2027, TLE2037 (*)	8pF	15nA	not quoted
TLE2061, TLE2161	4pF	3pA	1TΩ

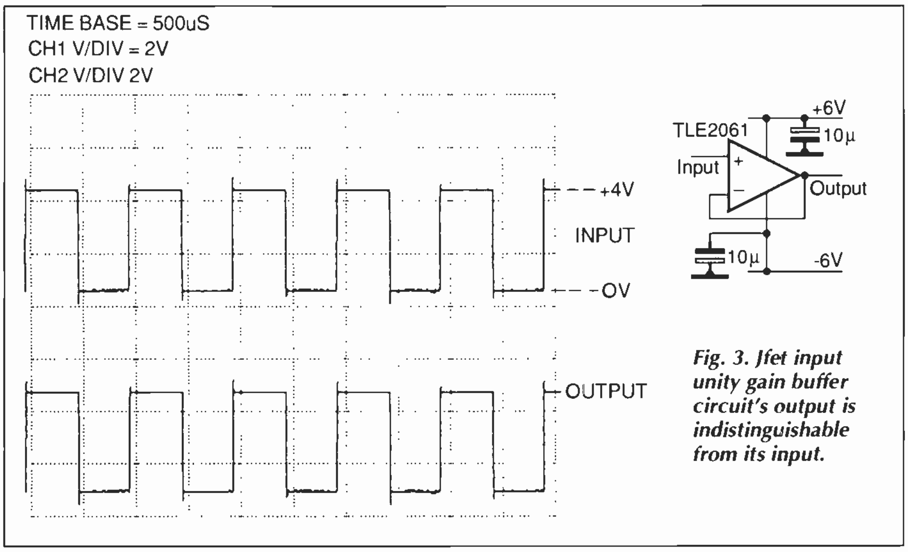


Fig. 3. Jfet input unity gain buffer circuit's output is indistinguishable from its input.

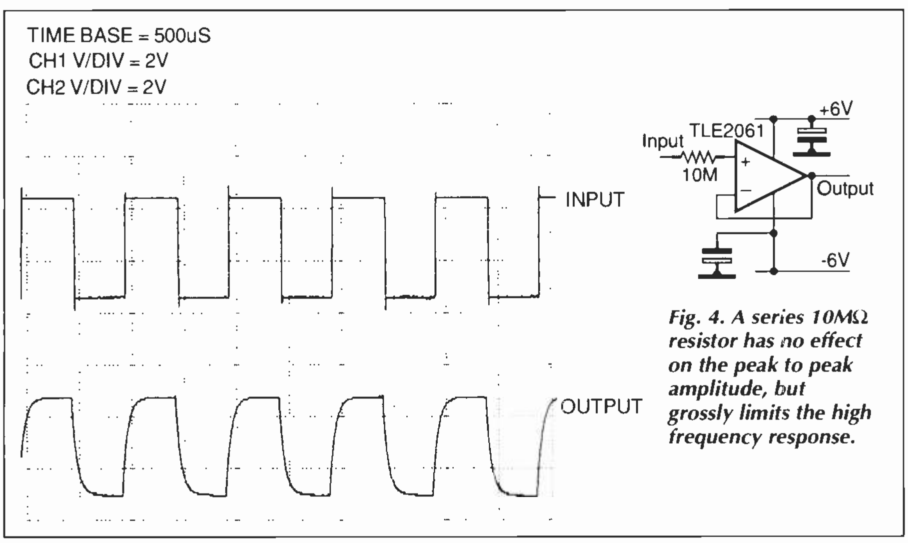


Fig. 4. A series 10MΩ resistor has no effect on the peak to peak amplitude, but grossly limits the high frequency response.

pensated cousin, the *TLE2161*, features an enhanced slew rate of 10V/µs for applications where the closed loop gain is x 5 or more.

Connect the *TLE2061* was as a unity gain non-inverting buffer (**Fig. 3**) and apply a 1kHz 0V to +4V square-wave input (upper trace). Spikes on the leading edges appear to be an artefact of the digital storage adapter's screen dump software, there being no trace of them on the oscilloscope trace. Allowing for that, the op amp's output (lower trace) is pretty well a perfect replica, as would be expected.

Next, place a 10MΩ resistor in series with the op amp's input (**Fig. 4**). The input and output then appear as in the upper and lower traces respectively. With the op amp's 4pF input capacitance and allowing 1pF for strays, the input circuit time constant comes to 50µs, and viewing the lower trace at a faster time-base speed shows that the time to 63% response was indeed just 50µs.

Clearly in this application, the influence of the input capacitance is far more significant than that of the input resistance. Use of guard rings, as recommended in the data sheet, will maintain the high input resistance and will minimise stray capacitance external to the op amp, (see **Fig. 5**; the similarity to Figure 1 is clear).

Bootstrapping cannot reduce the effect of the device's internal input capacitance, however. So precede the op amp with a discrete bipolar buffer stage, using a *BC108* (**Fig. 6**). The inadequate input resistance and lower than unity gain with this arrangement is evident on comparing the lower trace with the upper. But the high frequency response is better than in Fig. 4 – which is to be expected as the input capacitance is now only that of the transistor, mainly C_{cb} or C_{obo} approximately. The data sheets give this as 6pF max, although the *Transistor DATA Book* (Vol. 1, 1977) gives C_{cb} typical as 2.5pF. The input time constant is about half that in the circuit of Figure 4.

Bootstrapping boon

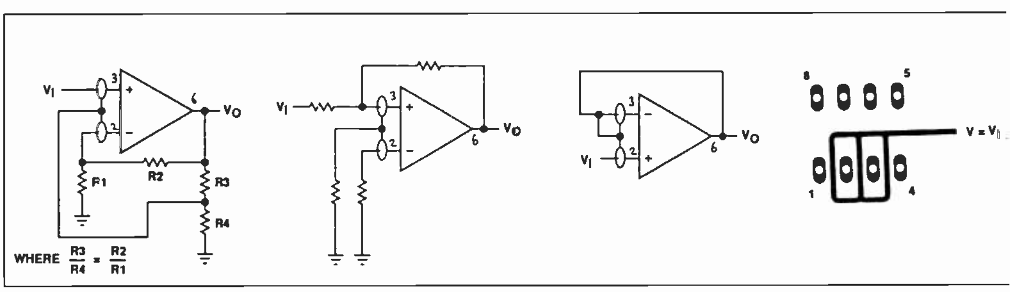
On the face of it, the result is hardly an improvement; slightly lower input capacitance has simply been traded for a much lower input resistance. This is where bootstrapping really comes into its own, hauling the input up by its own bootstraps.

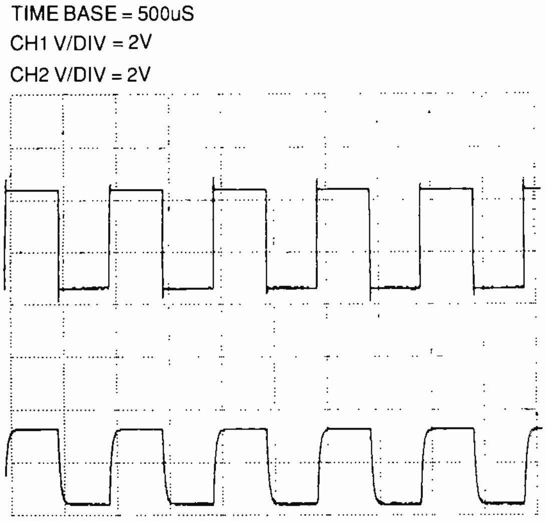
Stage 1 involves bootstrapping the *BC108*'s collector (**Fig. 7**), which is seen to be very effective in shortening the input time constant, though there is still a shortfall in low frequency gain.

The all-important point to note is that the bootstrapping of the collector only works because there is a separate stage following the emitter follower, providing current gain. The collector cannot be bootstrapped from the input emitter follower's own emitter, even though such an arrangement has previously been proposed in *EW + WW*.

Stage 2 extends the bootstrapping to the input emitter follower's emitter circuit (**Fig. 8**). Now the output (lower trace) is indistinguishable from the input. The improvement does not extend down to DC – the input resistance at 0Hz being unchanged – but only

Fig. 5. Guard rings around the input terminals minimise the effect of board leakage and capacitance by surrounding the input pins with copper track at the same potential as the input. There is thus no potential difference to force current through any leakage paths or through stray capacitance.





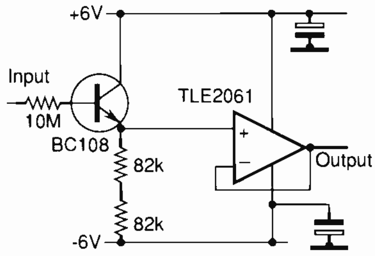
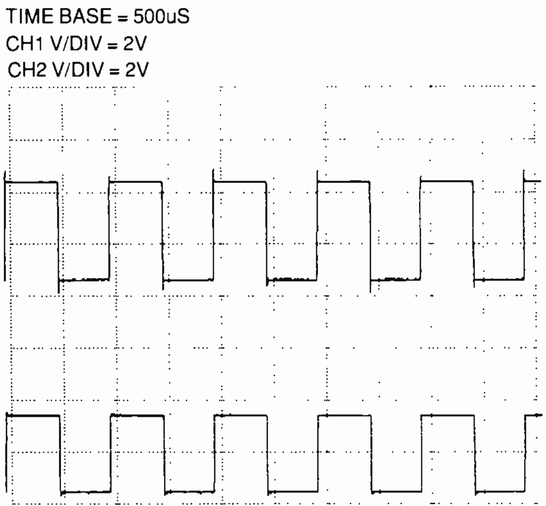


Fig. 6. A discrete emitter follower buffer ahead of the op amp improves the high frequency response by a factor of two, but low input resistance pulls down the peak to peak output.



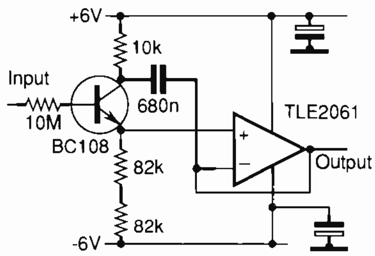
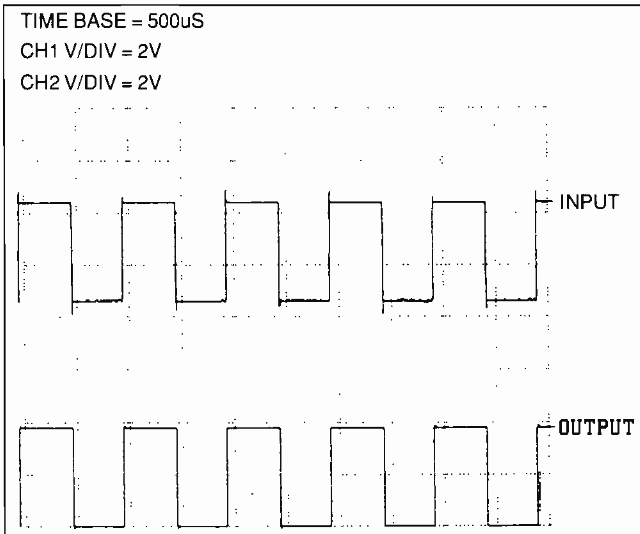


Fig. 7. Bootstrapping the emitter follower's collector shortens the input time constant to a negligible value, but the dc gain is still well below unity.



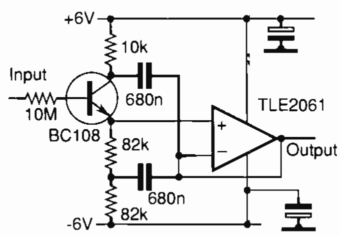


Fig. 8. Bootstrapping the emitter circuit as well results in an output indistinguishable

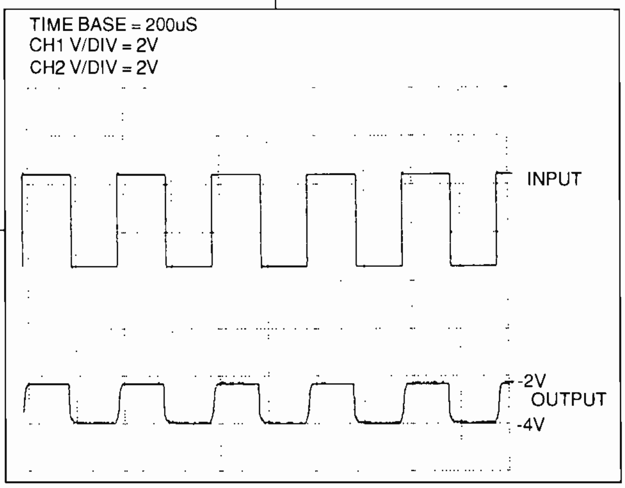


Fig. 9. As in Figure 8 but with the input resistor raised to 50MΩ. The poor performance is the fault of the designer, not the circuit.

down to a frequency at which the time constant of the emitter bootstrapping circuit starts to be significant.

To extend the boostrapping down to DC, the emitter circuit bootstrap capacitor would need to be replaced by a zener diode. To see how far it is now possible to push the circuit, replace the 10MΩ input resistor by a string of five 10MΩ resistors in series. The result is a substantially reduced output shown in **Fig. 9** – an unduly rapid collapse , bearing in mind how good the performance had been with 10MΩ series resistance.

Probing around the circuit shows that the emitter swings between –2V and –4V, due to the volt drop caused by the transistor's base current flowing through the 50MΩ resistor. So the emitter current is totally inadequate. Checking the DC conditions (the first recourse when a circuit is not behaving it ought) and raising the op amp supply rails to ±15V resulted in an output virtually as good as in **Fig. 8**.

Clearly, properly applied, bootstrapping can raise the input impedance at DC and up to a frequency determined by the op amp's performance, to such a high level that a 100MΩ source resistance results in no loss in amplitude – ie to an input impedance of 10GΩ or more.

As a matter of interest, the circuit of **Fig. 9** (with ±15V rails) can be modified by substituting a *BF244 N* channel small signal jfet for the *BC108*. Of course there is no volt drop across the 50MΩ input resistance, and the op amp output voltage sits at a positive level set by the fet's gate source reverse bias voltage at the source current defined by the two 82K resistors.

The fet's drain gate capacitance is the best part of 2pF, so the collector bootstrapping is still necessary. But the input resistance is so high that a source circuit bootstrapping capacitor is not needed.

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AC127	30p	BC516	22p	BD332	40p	BF371	17p	BU546	140p
AC128	35p	BC537	25p	BD361	60p	BF420	16p	BU608	150p
AC128K	40p	BC546	8p	BD362	60p	BF421	18p	BU626	150p
AC141K	45p	BC547	8p	BD370	30p	BF422	21p	BU636	150p
AC142K	45p	BC548	8p	BD371	30p	BF423	25p	BU801	80p
AC176	22p	BC549	8p	BD410	50p	BF450	20p	BU806	75p
AC176K	28p	BC550	8p	BD433	28p	BF455	12p	BU807	70p
AC187	25p	BC556	8p	BD434	30p	BF458	19p	BU902	130p
AC187K	40p	BC557	7p	BD435	31p	BF459	19p	BU903	130p
AC188	25p	BC558	8p	BD436	30p	BF462	50p	BU920	130p
AC188K	40p	BC559	8p	BD437	28p	BF469	30p	BU922	130p
ACY18	48p	BC560	8p	BD438	36p	BFR90	52p	BU930	130p
ACY19	48p	BC637	20p	BD439	40p	BFR91	99p	BUT11AF	70p
AD149	60p	BC638	20p	BD440	40p	BFT43	30p	BUT12	80p
AF125	50p	BC639	20p	BD441	40p	BFX29	20p	BUT56A	150p
AF127	50p	BC640	20p	BD533	50p	BFX84	20p	BUX80	180p
AF139	30p	BCY33	200p	BD534	38p	BFX85	20p	BUX84	50p
AF239	30p	BCY34	200p	BD535	38p	BFX87	15p	BUX85	50p
		BCY70	16p	BD536	38p	BFX88	15p	BU69A	200p
BA157	12p	BCY71	16p	BD537	40p	BFX89	60p	BUY71	300p
BB105B	18p	BCY72	16p	BD538	40p	BFY50	14p	BUZ71	75p
BB205B	24p	BD115	30p	BD643	50p	BFY51	14p		
BC107	8p	BD124P	50p	BD645	50p	BFY52	14p	MJ10012	300p
BC108	8p	BD131	25p	BD647	50p	BFY56	25p	MJ15001	325p
BC109	8p	BD132	25p	BD649	50p	BFY64	25p	MJ15002	300p
BC109C	10p	BD133	50p	BD675	40p	BFY90	45p	MJ15003	325p
BC140	20p	BD135	20p	BD676	40p	BLY48	85p	MJ15004	370p
BC141	20p	BD136	20p	BD677	38p	BR100	14p	MJ2501	110p
BC142	20p	BD137	20p	BD678	40p	BR103	37p	MJ2955	55p
BC143	20p	BD138	20p	BD679	40p	BR303	85p	MJ3000	115p
BC147	8p	BD139	20p	BD680	40p	BSX20	15p	MJ3001	115p
BC148	8p	BD140	20p	BD681	45p	BSX26	18p	MJE29A	30p
BC149	8p	BD144	90p	BD682	45p	BT106	180p	MJE30A	30p
BC157	8p	BD157	38p	BD705	50p	BT109	90p	MJE340	25p
BC159	8p	BD166	30p	BD707	50p	BU104	80p	MJE350	80p
BC160	30p	BD175	30p	BD709	50p	BU105	80p	MJE520	30p
BC171	10p	BD177	30p	BD711	50p	BU108	100p		
BC172	10p	BD179	32p	BD736	50p	BU109	80p	OC28	350p
BC177	14p	BD181	45p	BD826	50p	BU110	110p	OC29	250p
BC178	14p	BD182	60p	BD828	50p	BU111	140p	OC35	350p
BC179	14p	BD184	60p	BD897	50p	BU124	60p	OC36	250p
BC182	7p	BD187	30p	BD899	50p	BU126	65p	OC45	50p
BC182L	7p	BD201	33p	BD901	50p	BU180	150p		
BC183	7p	BD202	38p	BD977	50p	BU184	100p	TIP29	15p
BC183L	7p	BD203	42p	BDX32	100p	BU204	75p	TIP29A	22p
BC184	7p	BD204	42p	BDX33	60p	BU205	70p	TIP29C	25p
BC184L	7p	BD222	31p	BDX65	80p	BU206	100p	TIP30	25p
BC212	7p	BD225	31p	BDW23	55p	BU208	70p	TIP30C	30p
BC212L	7p	BD232	31p	BDW24	55p	BU208A	60p	TIP31A	22p
BC213	7p	BD233	30p	BDW93	50p	BU208D	80p	TIP31C	30p
BC213L	7p	BD234	32p	BDW94	50p	BU209	140p	TIP32	24p
BC214	7p	BD235	28p	BDY92	100p	BU225	190p	TIP32A	21p
BC214L	7p	BD236	30p	BF225	30p	BU226	190p	TIP32C	28p
BC237	7p	BD237	21p	BF240	16p	BU312	120p	TIP33	50p
BC238	7p	BD238	24p	BF245	25p	BU325	55p	TIP33C	60p
BC239	7p	BD239	30p	BF254	15p	BU326	75p	TIP34	50p
BC300	20p	BD240	40p	BF255	12p	BU406	70p	TIP34C	60p
BC301	20p	BD241A	40p	BF256	18p	BU406D	85p	TIP35C	65p
BC302	20p	BD243A	50p	BF257	18p	BU407	55p	TIP36C	65p
BC303	20p	BD244	50p	BF258	18p	BU407D	80p	TIP41A	20p
BC304	25p	BD245	50p	BF259	18p	BU408	75p	TIP41C	22p
BC327	7p	BD246A	50p	BF262	25p	BU408D	85p	TIP42A	20p
BC328	7p	BD265	45p	BF263	25p	BU409	95p	TIP42C	22p
BC337	7p	BD267	45p	BF270	18p	BU426A	70p	TIP47	40p
BC338	7p	BD269	45p	BF273	15p	BU500	100p	TIP48	40p
BC441	28p	BD278	50p	BF311	21p	BU508A	70p	TIP50	60p
BC446	8p	BD311	100p	BF336	20p	BU508AF	95p	TIP51	120p
BC449	15p	BD314	100p	BF337	20p	BU508D	80p	TIP52	120p
BC461	28p	BD315	150p	BF338	20p	BU508DF	115p	TIP53	120p

TIP54	140p	2N3773	100p	LM324	30p	6116	80p	Z80BPIO	150p
TIP105	65p			LM335Z	120p	6264-15	160p	Z80ACTC	140p
TIP106	65p	**VOLTAGE**		LM339	35p	6264-12	200p	Z80BCTC	200p
TIP107	65p	**REGULATORS**		LM348	50p	6264-10	210p	Z80ASIO	210p
TIP110	40p	7805	25p	LM358	45p	62256-12	300p	Z80ASIO-1	210p
TIP111	50p	7806	25p	LM380	80p	6502	300p	Z80ASIO-2	210p
TIP112	40p	7808	25p	LM381	150p	6502A	360p	Z80ADART	260p
TIP115	45p	7812	25p	LM382	130p	65C02	930p		
TIP116	45p	7815	25p	LM386	60p	6522	280p	**SIMMS**	
TIP117	50p	7818	25p	LM387	100p	6551	380p	256K×9-80	1000p
TIP120	37p	7824	25p	LM393	45p	6800	210p	256K×9-70	1100p
TIP121	46p	7905	25p	LM709DIL	30p	6802	220p	1MB×9-80	3000p
TIP122	47p	7906	30p	LM723	40p	6803	500p	1MB×9-70	3200p
TIP125	47p	7908	30p	LM741DIL	18p	6808	500p		
TIP126	56p	7912	30p	LM741MET	45p	6809	500p	**SIPS**	
TIP127	56p	7915	30p	LM747	55p	6810	150p	256K×9-80	1100p
TIP130	30p	7918	30p	LM748	30p	6818	380p	256K×9-70	1200p
TIP131	30p	7924	30p	LM1458	30p	6821	130p	1MB×9-80	3100p
TIP132	30p	78L05	24p	LM1889	300p	6840	290p	1MB×9-70	3200p
TIP141	90p	78L08	24p	LM3900	40p	6845	200p		
TIP142	90p	78L12	24p	LM3909	80p	6850	90p	**IC SOCKETS**	
TIP145	65p	78L15	24p	LM3911	160p	8080A	380p	8-pin	5p
TIP146	90p	78L18	24p	LM3914	160p	8085A	300p	14-pin	6p
TIP147	100p	78L24	24p	LM3915	160p	8086	500p	16-pin	7p
TIP150	90p	79L05	35p	LM3916	270p	8088	480p	18-pin	10p
TIP151	90p	79L08	35p			8156	300p	20-pin	12p
TIP2955	42p	79L12	35p	**COMPUTER**		8224	240p	22-pin	13p
TIP3054	45p	79L15	35p	**IC's**		8226	240p	24-pin	14p
TIP3055	42p	7818KC	100p	2114	150p	8243	250p	28-pin	16p
		7824KC	100p	2532	200p	8250	750p	40-pin	18p
2N3053	18p	LM309K	100p	2716	100p	8251	200p		
2N3054	40p	LM317K	200p	2732	200p	8253	160p	**ZENERS**	
2N3055	38p	LM317T	100p	2732A	220p	8255	160p	400mW	
2N3055H	50p	LM323K	350p	2764	150p	8256	700p	BZY88 range	
2N3440	58p	LM723	35p	27C64	200p	8257	220p	2V7 to 39V	5p
2N3442	85p	78H08KC	800p	27128	150p	8271	3400p	1.3W	
2N3585	120p	78H12KC	700p	27256-25	150p	8279	270p	BZX61 range	
2N3702	9p	79HGKC	800p	27512	300p	8283	400p	2V7 to 39V	9p
2N3703	9p					8284	440p		
2N3704	9p	**LINEAR IC's**		4116	40p	8287	260p	**LED's**	
2N3705	9p	LF347	110p	4164-15	80p	8288	650p	3mm	
2N3706	9p	LF351	45p	4164-12	90p	8748	700p	Red	5p
2N3707	9p	LF353	48p	4164-10	100p	8755	800p	Yellow	8p
2N3708	9p	LF355	60p	41256-15	80p			Green	8p
2N3710	12p	LF357	70p	41256-12	100p	Z80ACPU	100p	5mm	
2N3711	12p	LF398	300p	41256-10	110p	Z80BCPU	160p	Red	5p
2N3771	85p	LM301	25p	41464-12	150p	Z80ADMA	200p	Yellow	8p
2N3772	90p	LM311	35p	41464-10	200p	Z80APIO	150p	Green	8p

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CIRCLE NO. 142 ON REPLY CARD

Filtering out the origins of A-to-D aliasing

Filtering will remove the input components which cause A-to-D aliasing errors. David Mawdesly* explains why choice is a matter of compromise.

Presenting signals containing frequencies above a certain limit to an A-to-D convertor is going to result in aliasing errors in its output. Avoiding components in the input likely to cause aliases is vital – but even if a measured signal is known to have only frequency components below a certain value, spurious effects from the system might appear as higher frequency noise in the signal presented to the A-to-D. Sharp transitions in the measured signal can excite resonances, which may cause aliases, and general system noise and electrical pick-up all contribute to uncertainty.

One common misconception is that the problem can be eliminated with a fast sampling A-to-D card, but in practice this is not a viable option. Increasing the sampling rate causes severe problems due to memory limitations and the lower resolution convertors needed to run at the higher speeds. PCs are limited in their data throughput capacity too, and many data acquisition packages are unable to handle data at higher sampling rates.

Filtering after A-to-D conversion is also not possible, and filtering must be carried out before digitising, in the analogue domain.

Sampling rate

What is the appropriate sampling rate of an A-to-D convertor for measuring signals in the frequency range from 0Hz to a chosen maximum frequency $f_{(max)}$? Theory indicates that this sampling frequency must be at least twice the highest frequency component in the analogue signal. This sampling rate is generally referred to as the Nyquist rate. In turn, the frequency at half the sampling rate is called the Nyquist frequency, and any component above this frequency may appear as an alias. To avoid the possibility of aliases corrupting results of the measurements, an anti-alias filter must be used. A suitable low-pass analogue filter placed immediately before the A-to-D convertor can block all frequency components capable of causing aliases from reaching the convertor. Cut-off frequency $f_{(c)}$ of the filter is set to the highest frequency of interest so $f_{(c)} = f_{(max)}$. Then, in principle, the convertor should have a sampling rate of at least twice the cut-off frequency of the filter. In practice, the sampling rate needs to be somewhat higher then twice $f_{(c)}$.

Anti alias filters

An anti-alias filter reduces out-of-band alias-producing signals to less than the quantisation threshold of the A-to-D convertor – without introducing distortion of the signal's in-band components and without affecting overall measurement accuracy. The high frequency rejection requirements are determined by:

- Highest frequency of interest $f_{(max)}$;
- Sampling rate, and
- A-to-D convertor resolution.

The highest frequency of interest sets the cut-

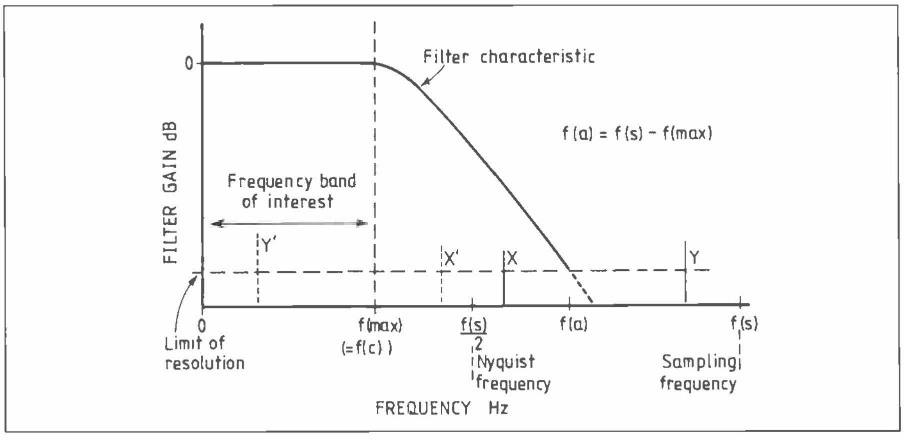

Fig. 1. Highest frequency of interest sets the cut-off frequency.

*Laplace Instruments

off frequency $f_{(c)}$ (**Fig. 1**). Potentially, any frequencies above the Nyquist frequency, $f_{(s)}/2$, could cause aliases. But only frequencies above $f_{(a)} = -f_{(max)}$ will be a problem because frequencies between the Nyquist frequency and $f_{(a)}$, although aliased, do not appear as an image in the frequency range of interest.

For instance, frequency component X appears as an alias at X', not within the range of interest. But component Y does require blocking because its alias appears at Y' where it will look like a real component.

There is a frequency band, $f_{(max)}$ to $f_{(a)}$, in which the filter must roll off the response so that at $f_{(a)}$ its attenuation at least equals the dynamic range of the A-to-D convertor (**Table 1**). If a 10-bit convertor is used, the dynamic range is 60dB calling for a minimum of 60dB attenuation above $f_{(a)}$.

A typical case might have a maximum frequency of interest of 5kHz, a sampling rate of 20kHz and an A-to-D convertor resolution of 14bit. Most filters quote their 3dB points as the cut-off frequency. Choosing 6kHz as the cut-off point gives $f_{(c)}$ as 6kHz and $f_{(s)}$ as 20kHz: $f_{(a)}$ = 14kHz.

The frequency band for roll off is 6-14kHz – ie slightly over one octave – and the required attenuation (= dynamic range of the A-to-D) = 85dB (from Table 1). So the filter is required to have a roll off of better than 80dB per octave.

Although standard anti-alias filters will deliver (and substantially exceed) the required attenuation performance, there is a price to pay in terms of signal accuracy and distortion.

The sharper the roll-off, the greater passband ripple and phase distortion introduced and so a compromise must be reached. For some applications, phase is not important and 1dB pass-band ripple may be acceptable. Where a problem remains, software compensation can sometimes be applied to correct the results in fixed installations.

Table 1. A-to-D characteristics

Dynamic range dB	resolution for A-to-D	ratio	2V full scale
0	–	1	–
6	1	2	1V
9.5	–	3	–
12	2	4	500mV
15.6	–	6	–
18	3	8	250mV
20	–	10	200mV
24	4	16	125mV
30	5	32	63mV
36	6	64	31mV
42	7	128	15mV
48	8	256	8mV
54	9	512	4mV
60	10	1024	2mV
72	12	4096	0.5mV
84	14	16384	125µV
96	16	65536	30µV
108	18	262144	8µV
120	20	1048576	2µV
132	22	4194304	0.5µV

Implementation

Even the most esoteric of anti-alias filters will probably consist of standard Sallen-Key or state variable filter stages cascaded to form the required number of poles: switched capacitor types are only a version of the state variable circuit implemented on an IC substrate.

Filter characteristics are determined by the value of the various R and C combinations. Estimation of these values can be most tedious. But fortunately tables (written in hardcopy and software) are now available for obtaining the values. Switched capacitor filters offering high performance, repeatability without tuning and the ability to move the cut-off frequency at will over several decades (typically 1Hz to 50kHz) are used for most applications. To design and produce an analogue equivalent is virtually impossible other than at very high cost. But these digital filters do have their drawbacks:

- As an input sampling circuit they introduce their own aliasing effects:
- Output is similar to that produced by a D-to-A convertor with a clock frequency typically 100 x $f_{(c)}$. The clock frequency component imposed on the output – known as clock breakthrough – may be quite substantial.

Filter characteristics

Filters can be grouped into four basic types of frequency characteristics, though all are compromises, tweaked to idealise one or other parameter. Main parameters are pass band gain accuracy, phase response, stop band attenuation and steepness of roll-off (**Fig. 2**).

Phase response is a function of group delay (the time taken for a signal at a given frequency to pass through the filter), with the ideal being when all frequencies take the same time – a linear phase characteristic. When achieved, a signal whose components all lie within the pass band will remain undistorted (as viewed by a scope) after passing through the filter. If the phase response is non-linear then we get all the right frequencies, but they may be in the wrong order!

At first consideration, the tendency is to select (**Table 2**) the characteristic giving the maximum roll-off rate, simply because that is the main characteristic needed from a filter.

But Cauer (elliptic) filters which maximise attenuation steepness also have significant stop-band ripple and a very non-linear group delay curve.

For a constant group delay, to give results that are undistorted in the time domain, the Bessel characteristic is the one to choose. Unfortunately, Bessel offers the slowest roll-off rate. If constant gain in the pass band (no pass-band ripple) is the target, the Butterworth characteristic offers just that, at the expense of indifferent performance in other respects.

Chebyschev characteristics offer a reasonable compromise, and can be set to give relatively low pass-band ripple and high cut-off rate.

Filter Considerations
Roll off rate
Minimum attenuation in the stop band
Maximum allowable ripple in the pass band
Range of cut-off frequencies
Resolution of cut-off frequency adjustment
Phase linearity
Filter characteristics
Noise performance
Clock breakthrough
Stand-alone or plug-in card
Programmability
Cost
Compatibility with A/D convertor (system/card)

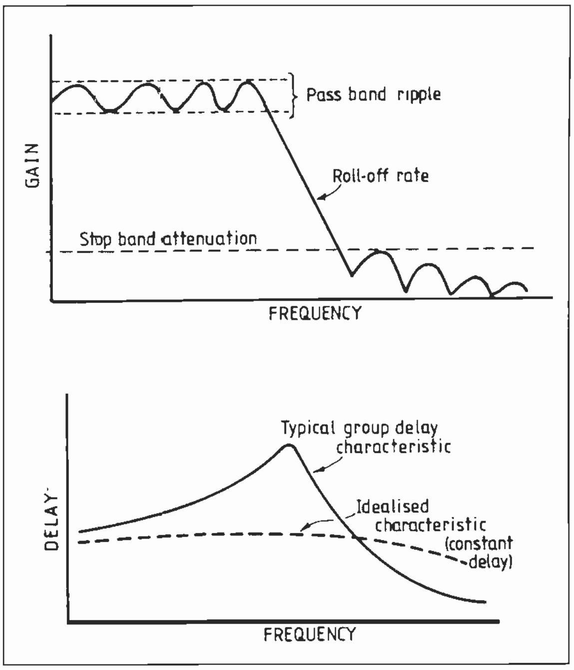

Fig. 2. Inter-relation of the main filter parameters. Selection of a filter always involves compromise.

Switched capacitor filters

Two fundamental types of filters are available – continuous (or conventional) and switched capacitor types – though both are basically

Table 2. Filter characteristics.

Characteristic	Best parameter	Worst parameter
Butterworth	constant pass band gain	slow roll-off
Chebyschev	compromise	pass band ripple
Cauer (elliptic)	roll-off rate stop band	attenuation
Bessel	phase response	very slow roll-off

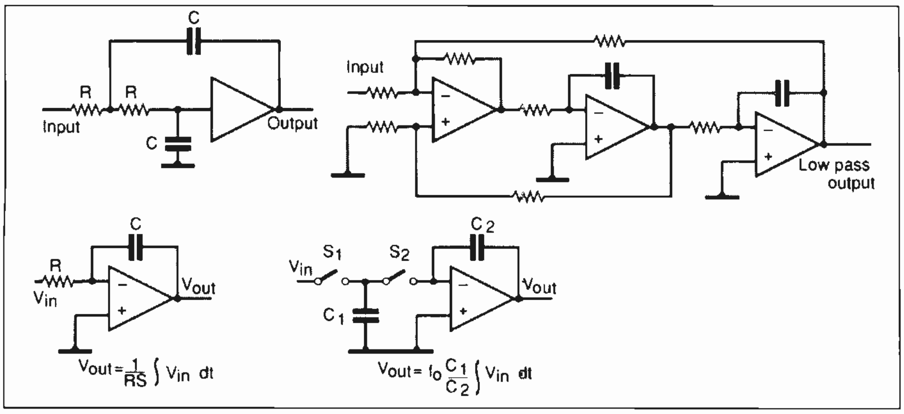

Fig. 3. Sallen and Key circuit (left) and biquad filter that can be configured to give a low-pass response of the type required for anti-alias requirements.

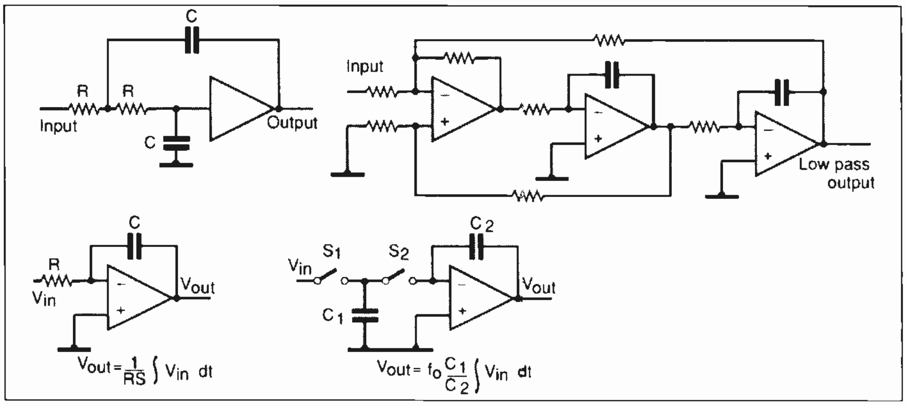

Fig. 4. Biquad design comprises several integrator stages (left) replaced by the circuit shown (right).

implementations of the same type of circuit.

"Continuous" is a fancy name for a standard circuit using op-amps, capacitors and resistors. **Figure 3** shows the well known Sallen and Key circuit **(a)** and a biquad filter **(b)** which can be configured to give a low-pass response of the type needed for anti-alias requirements. The problem with these, as many will know, is that the values of capacitors and resistors to be fitted are critical, especially for high performance implementations. Slight variations, either at assembly stage, or due to aging or temperature drift, can significantly affect and degrade performance. Also, once made, cut-off frequency is fixed.

Switched capacitor filters are an implementation of state variable, or biquad, circuits on an IC substrate using analogue switches to produce, in effect, components which can be varied in value. Note that the biquad design comprises several integrator stages as shown in **Fig. 4(a)**, replaced by the circuit shown in **Fig. 4(b)**.

The clock signal alternately closes S_1 and S_2, and while S_1 is closed, the capacitor C_1 charges to V_{in}, holding charge $Q=V_{in}C_1$. When the switches reverse, C_1 discharges to virtual earth, transferring the charge to C_2, causing the output voltage to change by an amount proportional to the input signal.

$$\Delta V_{out} = \Delta Q/C_2 = (C_1/C_2)V_{in}$$

If the clock frequency is fast relative to the input signal, then

$$V_{out} = \int_0 C_1/C_2 \int V_{in}\, dt$$

In other words, the integrator's gain is a function of the clock frequency. Because the characteristic frequency (cut-off frequency) of the biquad filter depends only on the integrator gain, a variable clock frequency can be used to produce a filter with a variable cut-off frequency. In addition, rather than discrete values of capacitors, the circuit uses capacitor *ratios* which are infinitely better in terms of stability and tracking – especially when integrated onto an IC substrate. As a result really high performance filters can be produced, in an IC outline, which can be programmed to any desired cut-off frequency within the range of operation of the device.

Unfortunately, the drawback to this approach is that the signal is now being clocked through the filter.

In effect we have inherited some of the negative aspects of sampled data systems. First, the clock will inevitably break through to the output signal, with the breakthrough typically of the order of 15mV, regardless of output signal level and corresponding to the clock frequency.

Normally, the clock frequency is 100 times the cut-off frequency, so that a relatively simple analogue (continuous) filter applied to the output signal will eliminate the effect. The problem is that this simplistic solution fails if the cut-off frequency is to be varied over a wide range. Secondly, aliasing can become a problem – precisely what we are trying to eliminate.

If cut-off frequency is 1kHz and the clock is 100kHz, simple aliasing theory shows that

Plug-in card filters

Various filter types are available as plug-in cards for PC computers. The following are available from Laplace Instruments*.

Cauer: The most commonly used filter type on the PC-based filter cards is the elliptic Cauer ("brick-wall") filter. This is a 7th or 8th order low-pass switched capacitor filter. Maximum cut-off frequency is 50kHz with an input clock frequency of 100 times the cut-off frequency. This filter provides a sharp cut-off, but maintains low wideband noise (typically 150µV RMS) and low harmonic distortion (0.03% or better). Pass-band ripple is typically +0.15% and stop-band attenuation at 1.5 times the cut-off frequency is 68dB minimum. Attenuation per octave is 120-150dB.

Butterworth: The Butterworth 8th order low-pass filter provides a maximally flat pass-band and is an alternative for high speed applications and for where ripple in the passband is unacceptable. Attenuation slope is 48dB/octave and stop-band attenuation over 80dB. Total wideband noise is typically better than 80µV RMS. With a THD of 0.03% it is suited to 16bit A-to-D cards. Maximum cut-off frequencies are 125kHz with a high speed version offering up to 250kHz.

Bessel: The 8th order monolithic lowpass Bessel filter provides a linear phase response over the whole pass-band. Wideband noise level is 60µV RMS.. Maximum cut-off frequency is 95kHz and attenuation is 1.5 times the cut-off frequency of approx 10dB.

Optional Cauer: A second Cauer filter type, with cut-off frequencies up to 100 kHz, is an 8th order low pass type with a lower pass-band ripple (0.1dB) than standard, but with a slower roll-off. Stop band attenuation is 80dB/octave and attenuation at 2.5 times the cut-off frequency is 92dB. Total wideband noise is 135µV RMS.

High speed Butterworth: For high speed applications a 6th order Butterworth has a clock to cut-off frequency ratio of either 25:1 or 50:1 giving maximum cut-off frequencies of 250kHz and 125kHz respectively. It maintains linear phase response and group delay in the passband. This filter exhibits steeper roll-off than the equivalent 8th order Bessel and the amplitude response approximates maximal flat passband.

aliases will be apparent if the input signal contains frequencies between 99 and 101kHz (and multiples thereof). To overcome this requires an anti-aliasing filter. Fortunately, this does not mean an ever-increasing spiral of problems. The pre-filtering requirement needs to reduce input signal frequencies around the clock frequency to below the dynamic range of the sampling system. As with the clock breakthrough problem, the solution is simple provided that the cut-off frequency is not to be varied over too wide a range.

Finally, because switched capacitor filters have clocks running through them, overall signal to noise ratio can never match that of quiet, well designed continuous filters. An effective limit is imposed on the dynamic range of any signal acquisition system using these filters and 90dB is normally considered to be the limit for well designed filters of this type. ■

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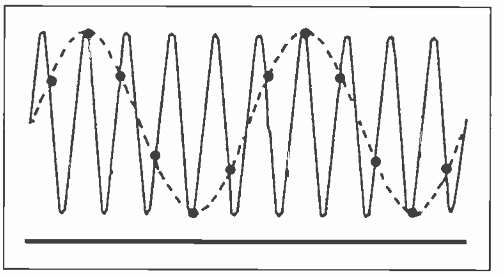
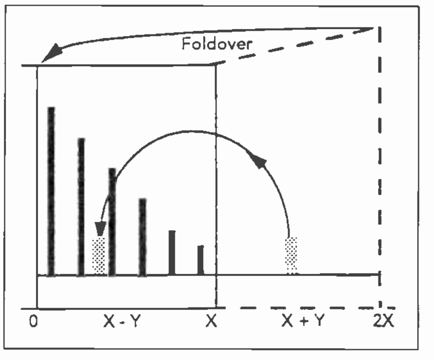
What is aliasing?
Aliasing is where a high frequency component takes on the identity of a lower frequency. Once created by the A-to-D process, the false signal component cannot be distinguished from a true signal. In some applications, FFT analysis for example, even the threat of aliasing can completely destroy the integrity of results.



Insufficient sampling of high-frequency sine-wave (solid line) leads to low-frequency aliasing (dashed line). To avoid it, sampling rate must be at least twice highest signal frequency.

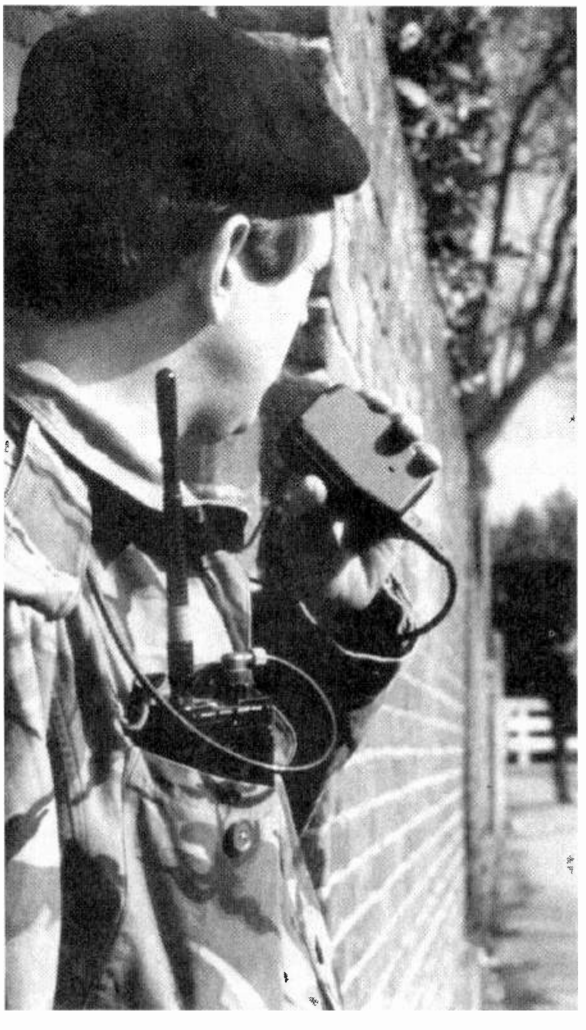


Relation to foldover. Aliasing occurs systematically as a result of foldover. If sampling rate is 2X, then foldover frequency is X, the limit set by the sampling theorem. Frequency components below X in the waveform being sampled will appear as they should. A component X+Y Hz will actually appear as the alias X-Y Hz below it. Hence the spectrum display is said to fold over at X.

The handhelds that endanger eyes

Possible health hazards resulting from long-term exposure to low levels of non-ionising RF radiation remains a controversial and unproven subject. As pointed out last year by Professor E H Grant (King's College, London and a member of the National Radiological Protection Board: "The evidence for the induction of malignant or other disease requiring a non-thermal mechanism for its explanation is, despite various reports, not hard enough to be taken account of when formulating a Guidance Document"[1].

On the other hand, there appears to be increasing recognition that the future widespread use of hand-held VHF/UHF transceivers – with their small antennas usually held within a few inches of users' eyes – requires more stringent guidelines than found, for example, in the widely-accepted Ansi recommendation that RF output of 7W or less is not hazardous. Although most handheld transceivers have significantly lower power than 7W, several amateur-radio transceivers are marketed with outputs of around 5W.



Vulnerable eyes

Eyes are particularly vulnerable to the thermal effects of RF radiation. As pointed out by Prof Grant: "A living organism is accustomed to receiving thermal stimuli and, provided these are not too large, the body can deal with them by the normal thermoregulatory mechanisms... More serious than these reversible effects are the irreversible thermal phenomena such as the induction of cataracts due to the irreversible denaturation of the (eye) lens proteins. This is a well known and well documented effect of non-ionising electromagnetic radiation, having been observed for furnace workers and glassblowers before the introduction of protective goggles. To produce it, the temperature of the lens needs to be elevated from the normal 37°C to around 42°C.

One reason why the lens is particularly vulnerable is because it has no blood supply and hence only poor thermal conduction paths. The other reason is the high content of water, both free and bound... Any protection guide must therefore recommend exposure levels which eliminate completely the possibility of irreversible effects and which limit the reversible effects to levels which are generally acceptable" [1].

Recent work, supported by the Swiss National Science Foundation, by Niels Kuster (Swiss Federal Institute of Technology) and Quirino Balzano (Motorola), underlines the need for a revision of part of Ansi C951-1982. This American standard (widely used in many parts of the globe) excludes all transceivers operating below 1.5GHz and radiating less than 7W from having to be assessed for compliance with basic safety limits based on spatial peak specific absorption rate (SAR). The SAR is the power (in watts) absorbed per kilogram of tissue.

Kuster and Balzano note that portable hand-held communication transceivers are becoming widely used consumer products, with cellular telephones and new digital systems (GSM, DECT etc) expected in the near future.

Revised regulations wanted

Their studies on the SAR of layered plane phantom in the near-field of antennas at frequencies above 300MHz have shown that the spatial peak SAR is related to the antenna current and not to input power. They conclude a detailed appraisal as follows:

"A consequence of this study is that the health safety regulations for hand-held communication equipment must be revised, because the 7W exclusion clause is not always consistent with the Ansi safety limits for the spatial local peak SAR recommended for the controlled environment (8mW/g). For the uncontrolled environment (1.6mW/g), the exclusion is in direct contradiction with the peak SAR limits, shown by the following example. Assume that the feedpoint current of a 7W 1.5GHz transceiver in 2.5cm distance from the eye tissue is increased to about 350mA due to feedpoint changes. The result is a spatial peak SAR, averaged over 1g of tissue, of over 40mW/g.

"Further, in the close near field, the SAR is not directly related to the input power but to the antenna current distribution."[2]

References

1. EH Grant, "Guidelines and standards for exposure in the frequency range 100kHz - 300GHz" *ERA Conference Volume*. "Electromagnetic Fields and Human Health", 1991, pages 3.2.1-3.2.6.

2. N Kuster and Q Balzano. "Energy absorption mechanism by biological bodies in the near field of dipole antennas above 300MHz", *IEEE Trans on Vehicular Technology*, Vol 41, No 1, February 1992.

RF exposure puzzles remain

For more than a decade, the extent of potential hazards arising from exposure to electromagnetic fields has been heatedly debated by those concerned with radio communications, broadcasting, radar and medical electronics. A fresh impetus was given to this subject in the 1980s with publication of epidemiological studies that appeared to reverse earlier work clearing RF radiation from any link with cancers. In turn, more recent studies have highlighted other, possibly more significant, correlations of the same rare cancers with the proximity of traffic and with concentrations of radon gas. Other studies have also shown odd and inexplicable correlations between childhood leukaemias and watching black-and-white (but not colour) television and with the use of hairdriers but not other domestic appliances!

A BBC research report by S Wakeling (RD1990/4), *Electromagnetic field exposure in broadcast environments*, underlines the renewed concern. The areas around broadcast antennas to which a person can safely gain access are being increasingly restricted. Broadcasters face a variety of problems when trying to ensure safe operational practices... The exposure guidelines have become more stringent each time they have been updated... International authorities are tending towards an "as low as reasonably achievable" principle in the light of inconclusive biological evidence which will create more problems in future...

Safety factors are being compounded by incorrect assumptions regarding exposure conditions. The derived field strengths of exposure standards are calculated assuming optimum coupling conditions and far-field, plane wave exposure whereas, in broadcasting, the potential for hazard is generally confined to the near field around the transmitting antennas. Hence the derived field-strength values are inappropriate for determining the SAR (specific absorption rate) levels and thus the actual level of potential hazard.

At an ERA Technology conference, *Electromagnetic Fields and Human Health*, Dr RD Saunders (NRPB) stressed that while acute exposure to sufficiently intense RF and microwave radiation will induce heating, resulting in detectable rises in tissue or body temperature, the vast majority of people are exposed to much lower field strengths.

There are several possible areas of biological interaction at low levels of exposure which may have important health implications and about which knowledge is limited. Mechanisms of interaction have been proposed but are not established. If there are such effects, then the evidence suggests that they are subtle and may well be masked by normal biological variation, claims Dr Saunders.

Reviewing human studies (including the auditory perception of pulse-modulated microwave radiation – an established phenomenon) and animal and cell studies including embryo and foetal development, Saunders stressed that: "the evidence from biological experiments that exposure will affect carcinogenesis in humans is far from convincing".

He concluded that: "There is some evidence from several research groups that responses to low level ELF fields or to pulse or amplitude-modulated RF or microwave fields only occur within specific frequency and amplitude windows. Some of these responses have been difficult to confirm, and their physiological consequences are not clear at present. In addition, mechanisms of interaction have not been convincingly established. It is important that these studies be continued and that the health implictions, if any, for exposed people are determined".

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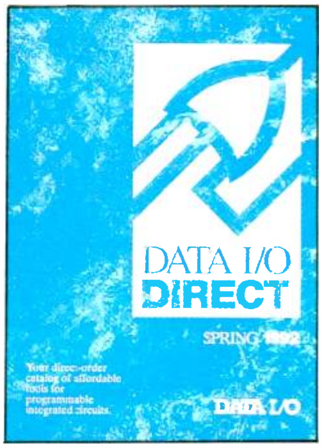

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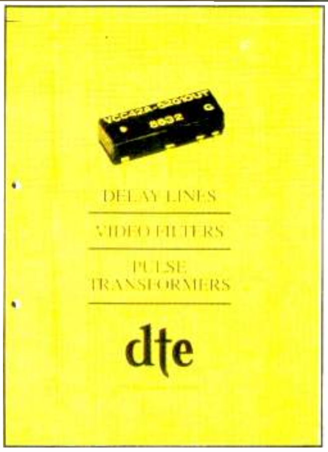

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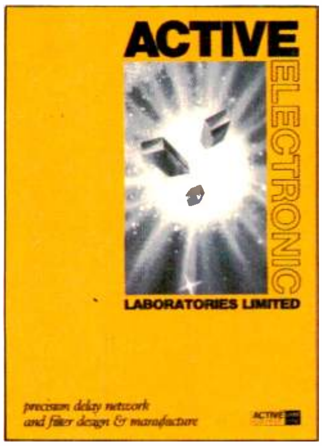

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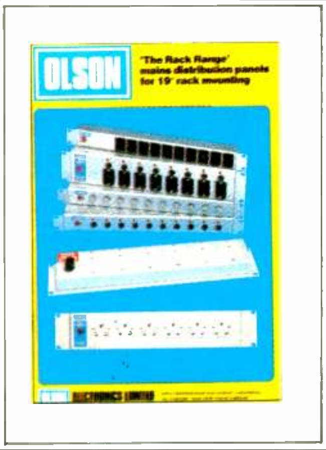

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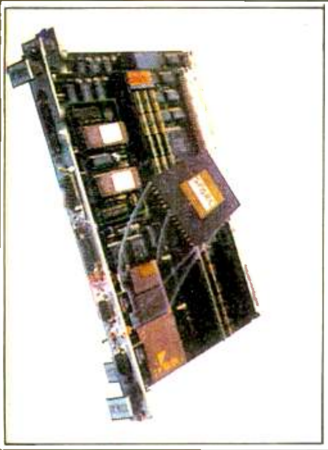

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