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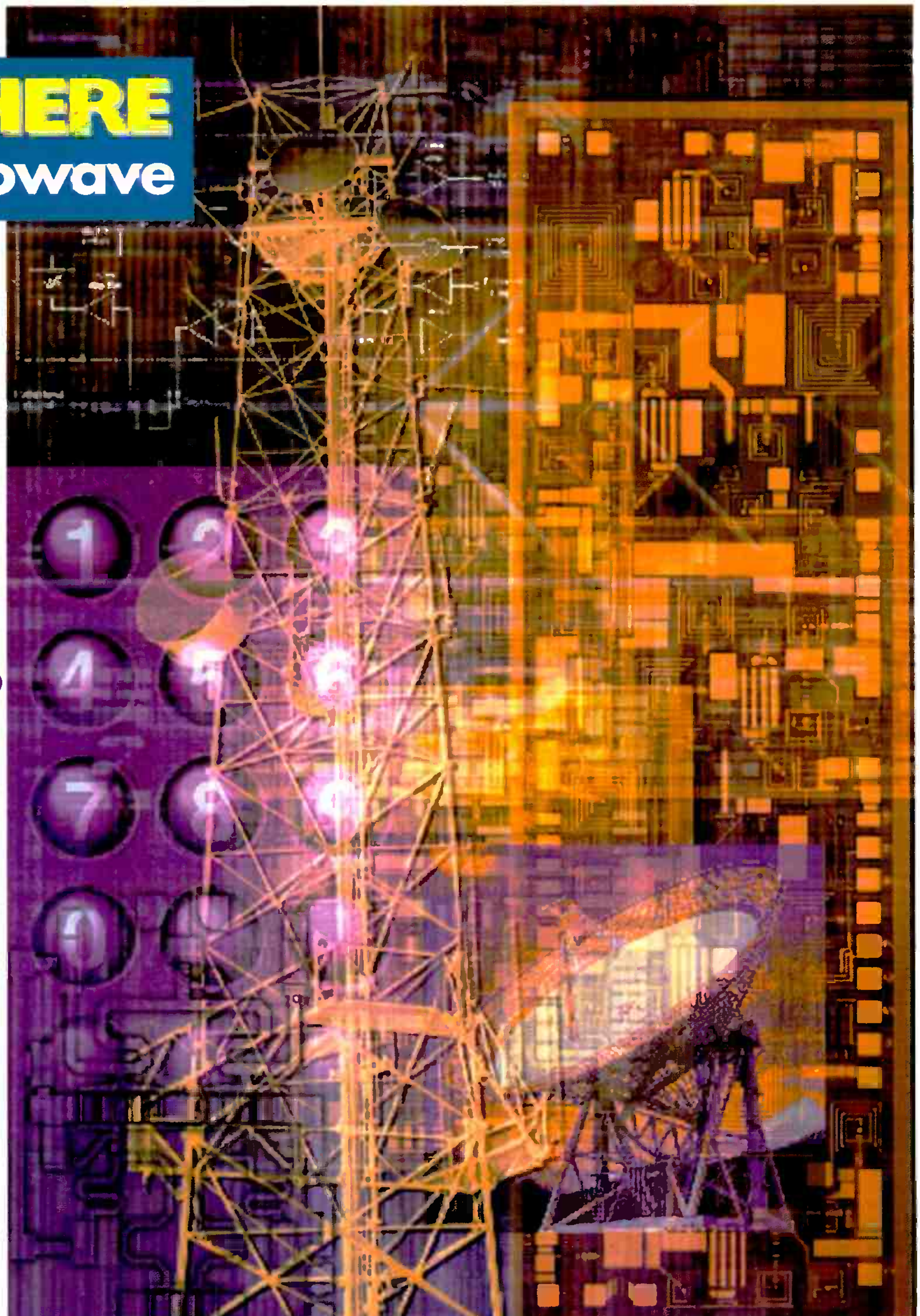

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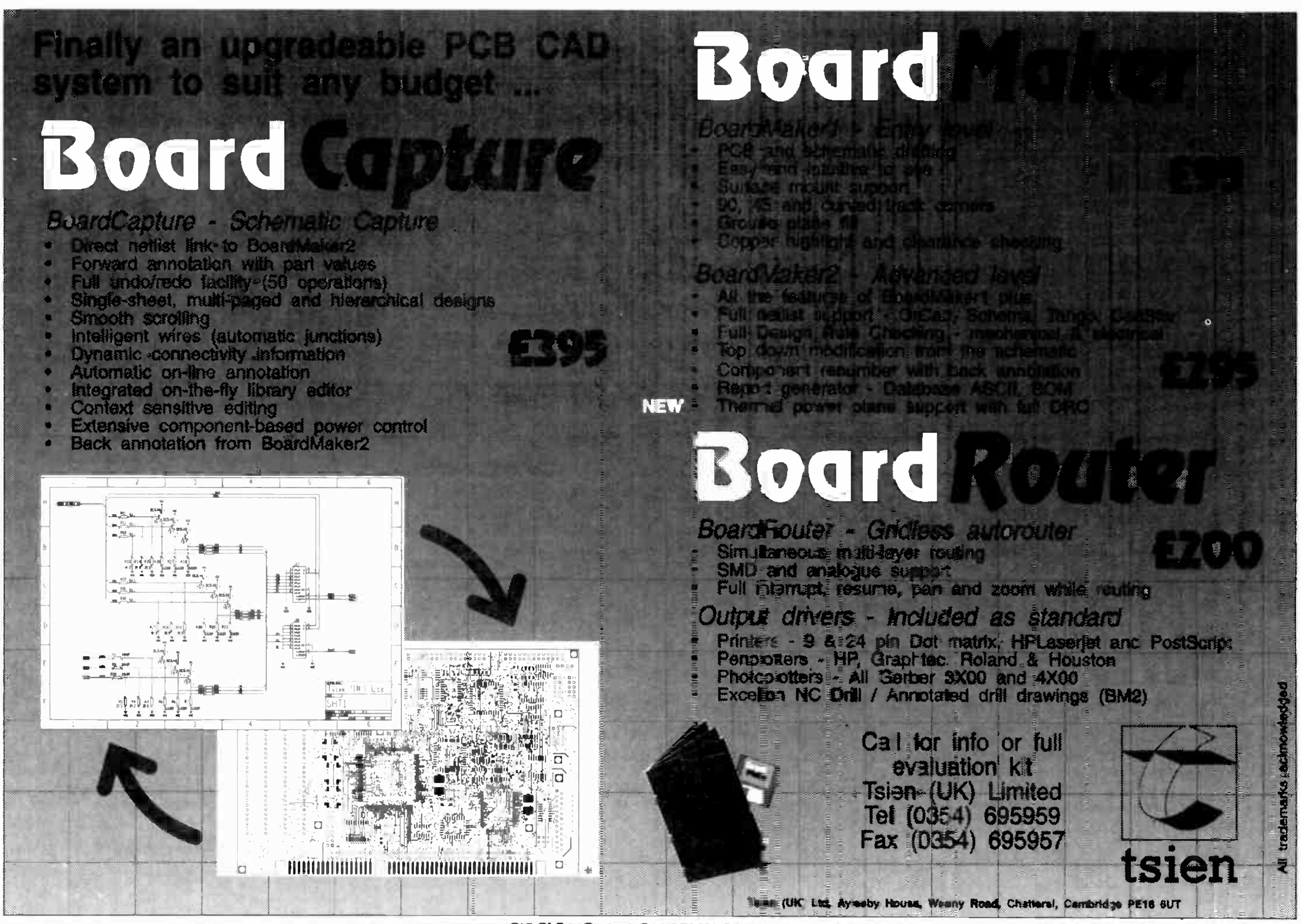

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Help with hypocrisy

I sit here writing this to the sounds of raging hypocrisy. Arms for Iraq, arms for Malaysia. Backhanders for everyone. Revolting! Disgusting!

"Nothing to do with me and its certainly nothing to do with you. So we will slap on a public interest immunity statement just in case."

It is actually concerns us all because prosperity is something that most people care about. Our society depends on technology to earn money. There is nothing else in the economy which can begin to compare with it. Most highly industrialised countries – with the exception of Japan and the new Pacific states – still greatly depend on defence to drive their research and development. Much of the relative prosperity in the UK can be traced back ultimately to our defence manufacturing activities. It is also certain that we would not have an indigenous electronics industry if it were not for military spending.

So, when we read in the papers that we have given Malaysia a £200m low technology dam in exchange for £1000m high technology defence contracts, we should be holding street rather than lynching parties. Congratulate the politicians who negotiated the deal rather than castigate them. I'm serious. It is only strong technology based industry which can deliver high per capita GNP. Arms, petrochemicals, nuclear technology and all the other politically incorrect things are what separate us from a rural peasant economy.

What I greatly object to is that our politicians also realise this but are too spineless to admit it.

I'm not suggesting for a moment that we shouldn't try to diversify from defence; I am a longtime supporter of the view that GEC and other major defence contractors have short-changed this country in all sorts of ways, but mostly by turning their collective backs on the civil hi-tech business which the Japanese and the Koreans have made their own. I sincerely hope that Lord Prior and his chums eventually get their come-uppance... but not before they have had a chance to take money off the Malaysians and others. Because if they don't, Aerospatiale and Dassault will. And if that means trade for "aid", so what? No one else makes a big deal about it and neither should we.

Wouldn't you rather see jobs and tax revenue in this country before any other? Of course. And politicians should stop equivocating and have the courage to admit it.

Frank Ogden

Electronics World + Wireless World is published monthly. By post, current issue £2.25, back issues (if available) £2.50. Orders, payments and general correspondence to **L333, Electronics World + Wireless World, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.** Telex:892984 REED BP G. Cheques should be made payable to Reed Business Publishing Group.
Newstrade: IPC Marketforce, 071 261-5108
Subscriptions: Quadrant Subscription Services, Oakfield House, Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 0444 445566. Please notify a change of address. Subscription rates 1 year (normal rate) £30 UK and £43 outside UK surface mail. Air mail rates are found on card bound into issue.
USA: $52.00 airmail. Reed Business Publishing (USA). Subscriptions office, 205 E. 42nd Street, NY 10117.
Overseas advertising agents: France and Belgium: Pierre Mussard, 18-20 Place de la Madeleine,Paris 75008. United States of America: Ray Barnes, Reed Business Publishing Ltd, 205 E. 42nd Street, NY 10117.Telephone (212) 867-2080. Telex 23827.
USA mailing agents: Mercury Airfreight International Ltd Inc, 10(b) Englehard Ave, Avenel NJ 07001. 2nd class postage paid at Rahway NJ Postmaster. Send address changes to above.
Printed by BPCC Magazines (Carlisle) Ltd, Newtown Trading Estate, Carlisle, Cumbria, CA2 7NR
Typeset by Marlin Graphics 2-4 Powerscroft Road, Sidcup, Kent DA14 5DT

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Low power stars at high power chip conference

The theme of this year's International Solid State Circuits Conference recently held in San Francisco was low power circuit and system techniques. Not only was there an entire session on low power technology, the low power trend carried over to notable watt-burners such as the latest generation microprocessors. Four of the seven papers in the microprocessor session discussed modifications to existing devices to slash their power consumption.

But some of the most innovative papers concerned analogue signal processing with two emerging themes. First, engineers are striving to develop analogue functions, principally the amplifier, which can provide high precision while operating from very low supply voltages – as low as 1V. Secondly, at these supply voltages – necessary for portable equipment – the incoming data has to be passed quickly into the digital domain to alleviate problems caused by noise.

Entire signal processing systems could be built on single chips operating from a 1V supply. This was the conclusion of a paper from Hitachi Central Research Laboratories. By developing two novel amplifiers, the firm's Tatsuji Matsuya showed how these could be applied to build analogue and digital systems operating from a 1.2V supply.

High-gain high-speed analogue amplifiers are difficult to design for low voltage operation. The traditional high speed cascode amplifier fails at low supply voltages because of its series connected cascode transistors. Non-cascode amplifiers do not provide sufficient voltage gain. Matsuya built a double feed-forward phase compensated amplifier to overcome this problem. This amplifier circuit was used to build a three-stage sample-and-hold amplifier for a 9-bit pipelined A-to-D converter. At 2MHz sampling rate the converterdissipated 4mW.

Matsuya also developed a self-current cut-off sense amplifier for use in low voltage srams. The intention was to reduce the DC power dissipated by ram sense amplifiers – a barrier to high-speed low voltage operation. In Matsuya's design the latching circuit automatically cuts off the DC after sensing, giving 5% of the consumption of conventional current-mirror amplifiers. The Hitachi engineers built a 16Kbit sram with a 32ns access time and 1.2V power supply. Dissipation was 2mW. **Simon Parry**, *Electronics Weekly*

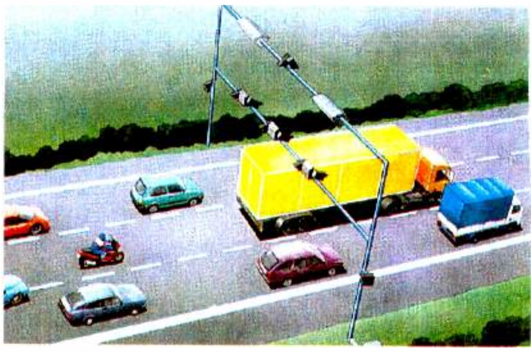
Road tolling in Euro-tussle

The Government is going ahead with plans for electronic tolling on Britain's motorways regardless of whether a European technical system specification can be drawn up by the end of the year. Manufacturers are being asked by the Government to submit proposals for electronic tolling systems that could be tried on UK roads this year.

But the absence of European standards for the transmission protocol and in-vehicle terminals has created uncertainty over which technology the UK will opt for. Europe already has a handful of non-compatible systems working in Italy, Austria, Sweden and Norway. The German government is considering no fewer than ten proposals, including GSM-based systems, infra-red, microwave and global positioning technology.

According to Philip Blythe at the transport operations research department at Newcastle University "everything is undecided".Two options for a short range system are a passive low cost in-vehicle tag with all accounting done at the roadside and an in-vehicle terminal with a smart card as demonstrated by Saab Combitech as part of the Adept EC transport research programme.

Peek plc is developing a UK version of Combitech, which is claimed to be Europe's only working multi-lane system. A representative for Peek expects the UK government to opt for a multi-lane non-stopping system using 5.8GHz microwave links between transceivers on overhead gantries and in-car smart card terminals.



According to Peek all that is left is the fine-tuning, whether there will be two or one microwave sensors per lane and whether cameras used to spot offending vehicles will be CCD based or infra-red.

But not all suppliers are so convinced. According to GEC Marconi, which has already delivered working systems in Italy and Singapore, maintains that the 5.8GHz frequency is not a certainty for the UK.

Blythe is more optimistic. He expects France, Germany and the UK to force the pace of a European standard. A short range system using 5.8GHz is most likely. It will provide communications link and an HDLC-based layer-two protocol between overhead gantries and vehicle.

The Adept multi-lane system supports 200Kbit/s data throughput across three lanes. This would be sufficient to offer full vehicle ID as well as debiting the smart card, if that is what the government has in mind. Each motorist would pay for mileage by buying a pre-paid smart card. Inserted in a dashboard terminal this would pick up the microwave signals from gantries across the motorway. **Richard Wilson**, *Electronics Weekly*

Happy birthday, capacitors: Next year is the 250th anniversary of the Leyden jar, the earliest form of capacitor, seen in action in this 1777 painting, The Electrical Experiment, by Charles Amédée Philippe Van Loo.

In the centre of the painting a woman stands on an insulating support with a rod in her left hand nearly touching a globe rotating against a leather cushion to produce a charge. Sparks jump between the globe and the rod.

Because the human body is a good conductor, the rod in her right hand is also charged. This is dipped into the water filled glass held by the boy. The glass is a Leyden jar with the glass acting as the dielectric and the water and boy's hand as the inner and outer electrodes.

The painting shows the capacitor being charged. The boy is going to bring his free hand closer to the rod, which will result in an electric spark jumping between the rod and his finger as the capacitor discharges through his body.

The painting is in the Arkhangelskoye Museum in Moscow and EW + WW would like to thank Curator Leonid Kryzhanovsky for loan of the photograph.

Technical doubt dogs fifth channel

It is now up to Peter Brooke and his Department of National Heritage to decide whether Britain leads Europe into the new age of digital television, or adopts old-fashioned analogue technology for a brand new tv service. The Independent Television Commission will not re-advertise the licence for a fifth tv channel, and has passed the buck to the DNH.

The BBC has publicly said that the analogue frequencies reserved for analogue Channel Five should now be used for digital tv. The ITC says it is "very doubtful" that technical tricks proposed by companies that want to start an analogue service will avoid interference. But the ITC has received too many similar proposals to be able to ignore them. "We are obliged under the Broadcasting Act 1990 to do all we can to establish a fifth terrestrial channel, and there is a strong case for re-advertising," says James Conway of the ITC. "Only the Government can release us from that obligation. We now need a clear steer on what the Government wants us to do".

It was hoped that so-called spare frequencies found by an alliance of MAI, Pearson and Time-Warner could have supplied the answer. These included channel 48 in Glasgow and channel 28 in London.

But the local oscillator of a TV set is 39.5MHz above the channel being watched. This causes spurious interference signals. British TV channels are 8MHz apart so the oscillator's interference is only 0.5MHz away from the programme that is five channels higher than the one being watched. This causes a herring-bone interference pattern on the screen.

In London, channel 28 would suffer interference from sets tuned to ITV on channel 23. In central Scotland, channel 48 is five channels up from ITV.

And five channels up from channel 28 is channel 33 used by BBC2. **Barry Fox.**

Secret problems: Westland Helicopters had interference problems with navigation equipment and it hit on the idea that it was caused by static on the blades discharging through the tail.

After doing loads of tests with an electric field meter rented from Livingstone Hire, they discovered the interference had nothing to do with static.

Chris Soundy, avionics systems engineer, said: "We did a whole set of readings and the theory wasn't valid. We now know what was wrong but I'm not allowed to say."

Fully static display could lead to electronic book

A new liquid crystal display for a portable computer plays a remarkable trick. When the power supply is switched off, the screen freezes and continues to display the last image, with no loss of clarity. This means that a battery-powered unit can run at least ten times longer on a single charge than any other unit currently available. The display's developers, Thorn EMI's Central Research Laboratories, believes that its *electronic paper* will finally make electronic books a viable competitor for printed paper.

CRL has already licensed the technology to a Japanese company with expertise in the mass production of LCDs. CRL has now built a 13cm screen to prove that the idea works and promises a product prototype in six months.

CRL's new screen is monochrome, and relies solely on reflected ambient light. So no power is wasted on backlighting. The image is pure black on white, like paper, so no power is needed for dithering. Unlike conventional LCDs, the screen can be viewed from a very wide angle. Also the screen consumes power only when the image is changing, eg when the user is scrolling through pages of text or data. This reduces battery drain by a factor of at least ten, and usually twenty. Quite magically, the prototype screen retains a clear, frozen image even when all power is disconnected and battery drain is zero. **Barry Fox**

Machine vision for people?

In an attempt to give sight to the blind, researchers at MIT and the Massachusetts Eye and Ear Infirmary are developing an ultra-small vision chip that could be placed inside an eye. The result of five years work, the chip is based on similar technology used in the electronic eyes of robots, which detect and process light. The vision chip will be used in patients that have lost sight due to retina degeneration.

The photosensitive chip grabs light via electrodes on the chip's surface and then sends out electrical impulses that replace the messages once sent by the rod and cone cells. The strength of the impulses fluctuates with the intensity of the light source.

According to Dr Joe Rizzo, a researcher at MEEI, the spatial density of the photosensitive elements in the chip are similar to that of the rods and cones in a human eye. Dr Rizzo said the chip will not give a high level of resolution, but would let people see light and dark spots and perhaps their location.

However, it is still not known how ganglion cells in the brain would cope with the electrical impulses. Ganglion cells conduct information from the retina to the brain and would have to perform the same function for the signals produced by the chip. The first test for the vision chip is set for this spring when the researchers will implant the chip in a rabbit. But it could be years before the chip is tested on a human.

686 chip in 1Gips PC?

Personal computers operating at 1Gips could be possible next year when Intel starts volume shipments of the 686 microprocessor, codenamed P6.

Vin Dham, vice-president of microprocessors at Intel, says the 686 successor to the *Pentium* (586) is a six million transistor, 300Mips microprocessor. It will be in volume production next year.

Starting with the second generation 0.6μm *Pentium*, announced a couple of months ago, special functions for linking microprocessors together will be a feature of future Intel processors. The 1Gips PC comes from linking four P6s together in one computer.

Dham said that 25% of the *Pentium's* production is for 66MHz devices. The new 0.6μm version will top 100MHz.

"Hi Vision HDTV is dead, long live digital"

It's official, well sort of. Japan's analogue HDTV system, the 1125 line *Hi Vision*, was obsolete even before its official launch as a public service in 1991. Although the director at Japan's Ministry of Post and Telecommunications withdrew the statement the next day, he still admitted that the rest of the world is going digital and Japanese companies will be making the products.

He caused uproar with the "*Hi Vision* is dead" announcement, but it was a fact well known but not acknowledged by the country's electronic and broadcasting industries.

For proof, just visit the Tokyo studio complex of the Japanese state broadcaster NHK where one floor is devoted to analogue 1125-line hdtv production. Visitors are shown demonstrations using high power projectors producing clear pictures whose quality approaches film.

But the downstairs lobby tells a more realistic story. Staff and visitors can watch hdtv, but only on sets with picture quality so poor that 525-line sets beside them look better.

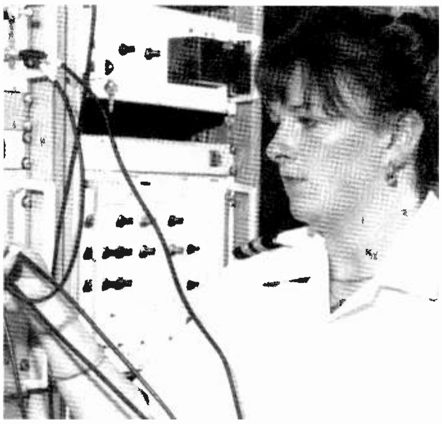
Hi Vision was first used commercially at the 1984 Olympics. Sony then developed a hdtv recording and editing system, which it loaned to Hollywood movie studios as a replacement for film. But Hollywood didn't buy, finding film cheaper and easier to use with better picture quality.

An alliance of US manufacturers is now close to finalising a standard for an all digital hdtv system for use in the US.

In Europe, analogue HD-Mac effectively died when Sky began broadcasting in old fashioned Pal in 1989. Japan was isolated in its commitment to an analogue future. The rest of the world was clearly going to go digital.

Since the service launch, NHK has transmitted eight hours a day of hdtv from a satellite. But the first year saw only 15,000 *Hi-Vision* sets sold. They were expensive (about £6500) and large enough to fill the average Japanese sitting room. And many of the wide screen sets convert the high definition signal into 525-line with definition no better than a conventional domestic TV set.

● Every delay makes an analogue Channel Five look less attractive. Last September, at a meeting in Bonn, European manufacturers and broadcasters signed a Memorandum of Understanding to create the Digital Video Broadcasting Project for Europe. The aim of the DVB forum is to draft European standards for digital television and there are now over 100 members from 12 countries. The DVB agreed its first draft at a meeting in Geneva in January. The target is a standard based on video compression technology and microchips already agreed by the Motion Picture Experts Group of the international standards organisations. This will speed receiver development and keep costs down. **Barry Fox**



1993 Young Woman Engineer of the Year: ***Helen Marsham on the job as third cable engineer for Cable & Wireless' marine division. Her work, which won her the joint IEEIE/EEF award, includes the operation, maintenance and calibration of submarine optical fibre transmission test equipment.***

Joint runners up were Kathryn Connor, manufacturing team leader with British Aerospace Defence Dynamics, and Deborah Hodges, a senior engineer at Westinghouse Signals.

The Mary George Memorial Prize for the entrant showing particular promise as an incorporated engineer went to Heather Arthur, a high voltage electrical diagnostician with the Ministry of Defence.

Solid state car lights

Hewlett-Packard has developed a light-emitting diode that burns up to four times brighter than conventional devices.

The diodes emit more light because their substrate structure is based on transparent aluminum iridium gallium phosphide (AlIrGaP), rather than light-absorbing gallium arsenide.

HP hopes the automotive industry will adopt the devices, using red-orange AlInGaP diodes in car's indicators and brake lights. Because they have a 10 million hour lifespan (the lifespan of incandescent lights is 2000 to 20,000 hours), the new leds will certainly out-live the car.

HP also hopes to attract cellular telephone makers with its amber and green leds. Because the new diodes consume just 5% of the power of standard backlights, HP claims they will greatly extend the battery life of cellular phones. The first telecommunication products with the new leds should be on the market within a year, while it may take two years for the automotive industry to adopt.

France plans optical highways

France's government is to spend billions of pounds over the next 20 years crisscrossing the country with highways of optical fibre cable. The decision to go ahead has been taken because of fears of future domination of Europe's information autoroutes by US companies.

The super highways will bring into every French office and home the full capabilities of interactive multimedia services. Asynchronous transfer mode technology will be used to transmit voice, data and video over the switched networks of optical fibre run into office and apartment blocks.

Initially, franchised operators of the services will have to link into the France Telecom network, for as long as that monopoly lasts. In making the commitment, the French government is pledging future governments to continue and finish the project.

Industry and telecommunications ministers have been lobbying the prime minister Edouard Balladur, using as a lever the success of the wired Minitel service, with six million terminals installed. But Minitel is outdated, because it is slow and cannot handle interactive video.

Jacques Toubon, one of the planners of the information highways, said: "There is no time to lose. In two years it will be too late. Already the US is putting itself in a dominant position. They will use our infrastructure to resell us cultural products made in the US and already paid for in their market."

Samsung tops chip chart

Korean chip maker Samsung overtook Toshiba to become the world's top mos memory chip maker, the first time in more than 12 years that a non-Japanese company has held the top spot. Market research firm Dataquest said Samsung grabbed the lead after an incredible 66% jump in sales from \$1.5bn in 1992 to \$2.5bn in 1993. Samsung ended a long Japanese reign that began in 1982 after Intel was toppled from the top of the mos memory market.

Radar reunion

Anyone involved with radar during WWII is invited to attend a reunion weekend in Blackpool from 20 to 22 May. In the three years the reunion event has been running, the organisers have found more than 3000 people who operated, maintained, installed, built and designed the ground and airborne radars. Those interested should contact Harry Jurd, 9 Chelmer Court, Basingstoke RG21 2DT. Tel: 0256 25980.

Optical systems "not immune to RFI"

Researchers have questioned the conventional wisdom that optical systems are immune to electromagnetic interference. Because light beams in photonic systems are typically modulated with electrical signals, the interfaces provide a way for electromagnetic energy to enter the system.

John Daher, a senior research engineer at GeorgiaTech, said: "These interfaces provide a path for external electromagnetic fields to disrupt operation of the photonic device, so it's a myth to say that installing a photonic system means you don't have to worry about electromagnetic interference."

Researchers tested electro-optic, acousto-optic, magneto-optic and charge coupled devices with interference from RF and microwave fields. They found that in some cases the interference seriously disrupted the operation of the equipment.

For magneto-optic devices, Daher said that "at high RF levels up to half the pixels would be switched to the wrong state and that would cause severe performance degradation."

A commonly used CCD sensor was affected by just 1μW of energy. Higher levels brought about complete loss of synchronisation.

In acousto-optic devices the interference caused the optical signal to spread out spatially, which could potentially cause interference in adjacent detectors. And in an electro-optic system, the rf energy caused the optical waveguides to heat up altering the refractive index and disrupting the operation.

In one case electromagnetic interference stopped a laser operating.

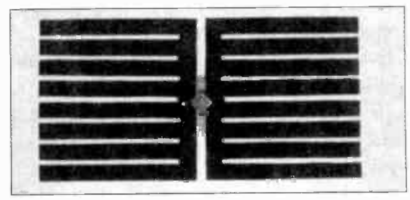
Daher said the results show engineers must still use hardening techniques to shield electronic components in photonic systems.

"One implication from this work is that you can't let your guard down just because you are using an optical system," he commented.

Shopping with RF: The tiresome business of unloading a supermarket trolley to pass the items individually over a barcode reader may become a thing of the past. RF interrogation developed by British Technology Group and the South African research organisation CSIR could allow the contents of a trolley to be read in seconds, and an entire store in minutes.

It works by putting a small chip, called *Supertag*, with a printed flat aerial on each item. External aerials flood the area with multiple polarised RF signals which are then modulated and returned on the tags afixed to the product. The readers are capable of multiple identification using an anti-clash communications function. So far they have achieved speeds of 50 items a second at a range of 4m.

CISR eventually hopes that the tags will cost just 1p each but most industry observers claim this to be unrealistic ■



Aerial array with chip in the middle: CISR needs to manufacture the mini RF transponders for just 1p.

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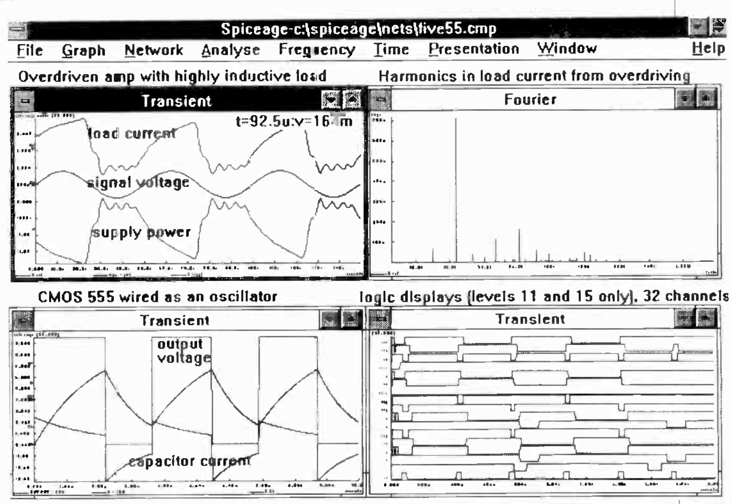

SpiceAge for Windows Level 1 is just £49.99 (half normal price). The manual, for those who need more than the extensive on-disk help systems, is an extra £45 including VAT. These prices are valid in the UK only and with Access, Visa or CWO payment. For other countries, please contact Those Engineers (Tel +44 81 905 0155, FAX +44 81 906 0969) for special details.

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Superconducting surge that is beyond belief

Two French superconductor teams have announced development of different materials that allow superconduction at dramatically increased temperatures. But the claimed temperature rises are so big, that nobody really believes them yet.

Ever since the discovery by Bednorz and Müller in 1986 of high temperature superconductors there has been a steady stream of papers claiming new materials with an ever higher critical temperature. Critical temperature T_c is the temperature below which the material loses electrical resistance and excludes magnetic fields.

All the high-temperature (relatively speaking) superconductors so far developed are brittle ceramic materials containing calcium, barium, yttrium, strontium, thallium or mercury combined with copper oxide. Until last December, most researchers accepted the existence of reproducible high-temperature superconductivity at around 133K (about −140°C). And it had taken five years to creep up to that point from 125K.

But the French breakthroughs appear to change everything.

The first (*Science*, Vol. 262, p. 1850), from a team led by Michel Laguës, claims a new and spectacular record for a material belonging to the BiSrCaCuO family. It has eight CuO_2 layers spaced 0.338nm apart and appears to become superconducting below 250K, which is roughly the temperature of a good domestic freezer. The snag – at least for any practical applications – is that the effect is only apparent for currents of around 10A.

The other French advance (*Phys Lett*, A184 pp. 215-217, 1994) by J. L. Tholence *et al* at Grenoble is to make a multi-layer compound based on the earlier record-breaking mercury cuprate superconducting at 235K.

At the time of writing, the physics community is still debating the significance of these results. No-one can doubt the credibility of the two research teams; both are highly reputable. But a sudden upward leap in record T_c after half a decade of slow steady progress seems hard to understand.

Commenting on the situation (*Nature*, Vol. 367 No 6458), Paul M. Grant of the Electric Power Research Institute in Palo Alto, California, points out that neither of the latest research groups has demonstrated reproducible electrical or magnetic behaviour over many temperature cycles without hysteresis. Measurement over repeated temperature cycles, he says, is a test most superconductivity experimentalist referees would demand.

But this year is certain to see publication, either of experimental confirmation of these results, or of yet more spectacular claims.

It is a field which Bednorz and Müller remind us is by its very nature empirical. It is also a field with huge commercial potential.

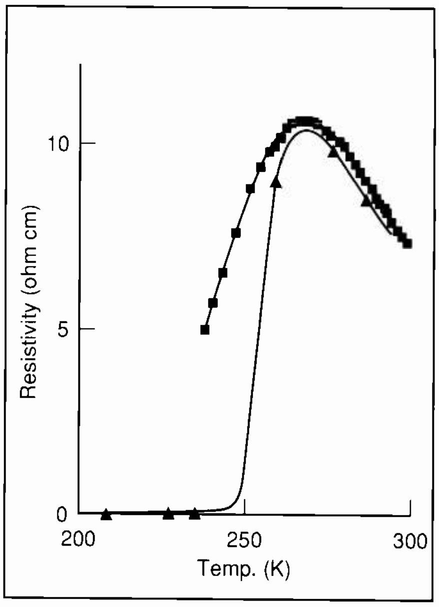


Fig. 1. Experimental results of Laguës' group indicate superconductivity beginning at the temperature of a good domestic freezer. Lower curve 10–9A, upper curve 10–8A.

Robots with attitude – science fiction or fact?

The suggestion that machines will one day be more intelligent than humans is a frightening one. But Professor Kevin Warwick, Head of the Department of Cybernetics at the University of Reading, is one of a growing body of respected scientists who believe not only that it will happen, but that it will happen in the next sixty years.

Machines displaying human levels of intelligence are generally looked on as pure science fiction. Even those working in artificial intelligence point to mathematical or theoretical reasons to explain why it will never happen.

But Professor Warwick has been increasingly arguing that artificial intelligence machines can already make decisions based on what they've learned, rather than on how they've been programmed. Robots working on this principle are currently operating in industry and, within certain limits, can already out-perform humans in terms of decision making.

Where machines have made less progress is in handling creativity and emotion. But that does not mean that computing systems will never paint masterpieces or weep when they make a mistake. Prof. Warwick acknowledges that AI systems do not yet appear very intelligent when simulating biological functions. He compares present-day machines with the intelligence of a slug or a fly – a far cry from human brains.

But a fly does have some intelligence, a lot more than, say, a motor car. For one thing it can take avoiding action when something is about to hit it.

Progress in artificial intelligence may appear slow, but Prof. Warwick argues that the most advanced systems are already remarkably intelligent. So will they ever bridge the gap between the performance of a fly's brain and that of Einstein or better?

Already there are plans in Japan to build an AI system that replicates the intelligence of a cat, with results expected, says Warwick, within the next five or six years. The programme is not intended merely to perform as a sort of advanced mouse-trap, but to embrace a whole range of feline behaviour. It could well be awkward, obstinate or aggressively territorial – a considerable advance on an artificial fly.

From feline intelligence, Warwick believes the step may not be too large to simulate the intelligence of an ape. Warwick argues persuasively that a machine that can interact intelligently with its environment is not just simulating intelligence; it is intelligent.

He goes further and says that many human functions such as emotion and self-awareness are probably less complex than we imagine. A machine may never behave in exactly the same way as a human brain, but it could well perform the same functions or feel the same feelings. Indeed, Warwick believes that machine intelligence will probably overtake human intelligence within our lifetime.

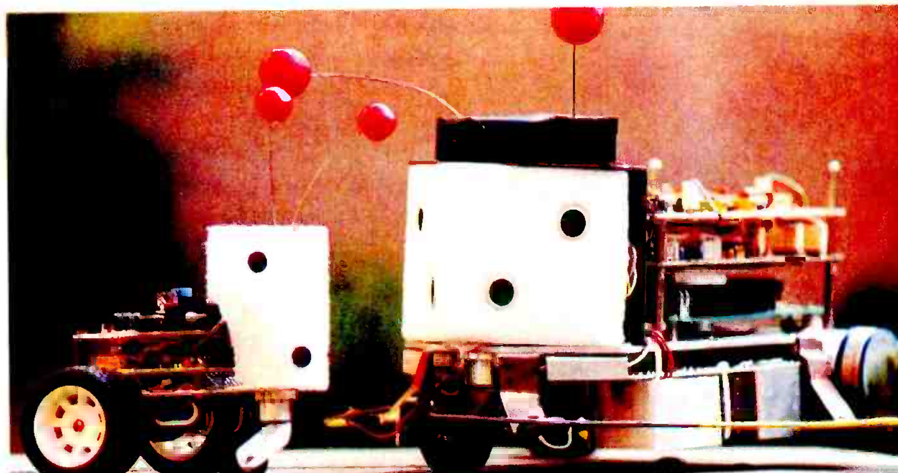
So will our lives in the future be ruled by machines? Will an army of computers take control of the world?

Professor Warwick says that the threat of nuclear war pales into insignificance compared with the threat from super-intelligent machines. He points out that highly intelligent animal species (including humans) have always tended to treat less intelligent ones badly. So there's no reason to expect that super-intelligent machines will act benignly towards our descendants.

The laws of Darwinian evolution and survival of the fittest may apply to machines just as much as they do in the jungle. So, as with subjects such as medical ethics, perhaps we ought now to be setting up committees to consider just how far ai researchers should be allowed to go. It could be just as important as deciding the limits for genetics researchers.

Professor Warwick believes that if we don't address this matter, our children almost certainly must.

One possible lesson from the jungle might be to avoid making machines that are too intelligent in too many activities at the same time. Highly specialised animals have not usually become too prolific.



Part of Reading University's research into machine intelligence, this robot has been taught two insect-like instincts – when switched on you must keep on moving, and second, when moving you must not bump into any obstacles. Japanese scientists say we could have robots with cat-level intelligence by the year 2000.

Optical technology brings microwaves to the motorway

German optical engineers have announced an advance that could bring closer the possibility of practical and cheap microwave communications systems.

Tomorrow's advanced microwave radio systems – whether for picocell telephony or for intelligent vehicle highway systems – will depend on units transmitting in the upper reaches of the microwave spectrum. Frequencies around 60GHz have been earmarked because of the bandwidth available and the atmospheric absorption that occurs. This enables the same frequencies to be available to other users relatively close by.

The snag is that any system that manages to keep in regular touch with a vehicle as it travels along a motorway needs a large number of low-power transmitters. Thousands of individual highly stable rf generators would obviously be expensive and would need lengthy power and feeder cables.

But it would be equally impracticable to generate one large signal at some central point and distribute it by a waveguide system to thousands of antennas.

A solution long contemplated for generation of multiple low-power microwave signals is to modulate a beam of light with the required rf and then send it along optical fibres to various transmitter sites. At each site it would be simultaneously demodulated and radiated.

The same source of light could feed any number of fibres, so it should be possible to have an equally unlimited number of precisely synchronised rf sources – ideal for distribution along a motorway.

The one problem is that dispersion along the optical fibre would degrade the signal. An optical signal of 1550nm (infra-red) modulated at 60GHz would degrade unacceptably after little more than 100m.

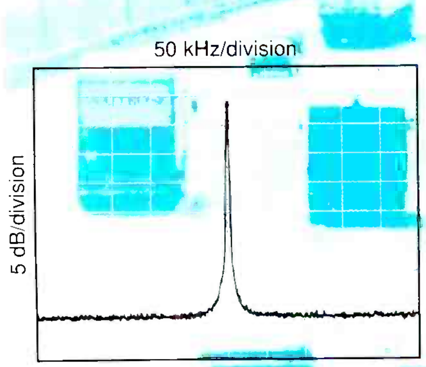
But a group from the Alcatel SEL AG Research Centre in Stuttgart look to have demonstrated (*Electronics Letters*, Vol. 30, No 1) how this can be overcome. They describe a system that will distribute a 60GHz signal to more than 1000 transmitting sites with a spectral line-width too narrow to measure.

The central equipment generates a 15GHz signal of high purity which is then frequency doubled to 30GHz. This rf signal then drives a lithium niobate Mach-Zehnder interferometer that amplitude-modulates the output of a 1550nm infra-red laser.

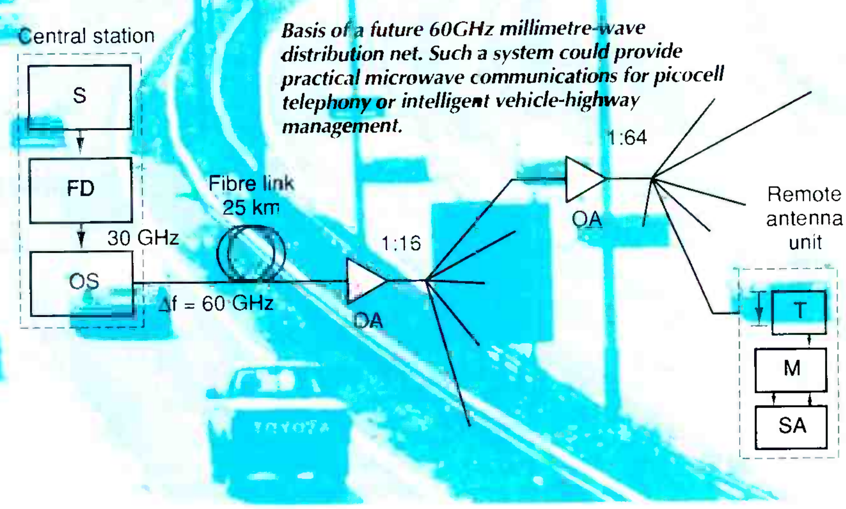
The resulting beam, consisting of two infra-red signals separated by 60GHz is distributed through a fibre-optic system where it is boosted along the way by erbium-doped fibre amplifiers. At the end of each fibre the two optical signals are mixed together to generate the required 60GHz rf.

As the 60GHz signal is a difference signal, the effects of fibre dispersion and laser jitter are completely eliminated.

In claiming that broadening of the spectral line width of the resulting rf is too small to measure, the Stuttgart team add that this is the first time that a 60GHz signal has been generated and distributed to more than a thousand different base stations.



60GHz millimetre-wave spectrum detected at a remote antenna unit showing that degradation is too small to measure.



Basis of a future 60GHz millimetre-wave distribution net. Such a system could provide practical microwave communications for picocell telephony or intelligent vehicle-highway management.

Red light means 'go' for blue laser

A cheap solid state laser that emits in the blue region of the spectrum is still the dream of many researchers. But until that dream becomes reality, an economical frequency doubling technique that allows infrared lasers to produce blue light could open up blue light applications – everything from colour copying and colour displays to optical storage systems such as cd-rom drives.

The attraction of blue light is that the shorter the wavelength, the more information a device can handle. Now a team led by Dr Thomas Penner at the Eastman Kodak Company in Rochester, NY, has developed a potentially cheap and efficient way of doubling frequency (*Nature*, Vol. 367 No. 6458).

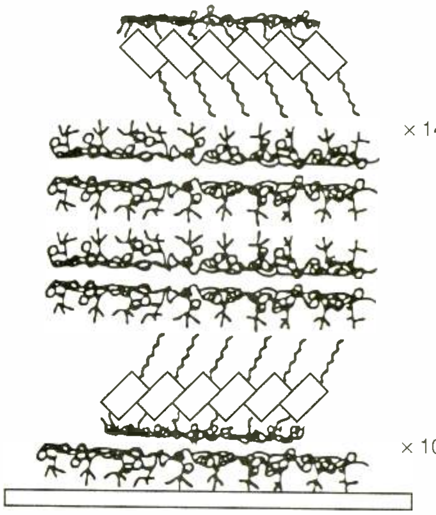
Frequency doubling in the optical domain is not of course new; there are many inorganic crystals, such as lithium niobate and potassium titanyl phosphate that possess the necessary non-linearity.

But they are inefficient because of low power confinement and a tendency to disperse the radiation.

As Penner puts it: "Bulk single crystals often require megawatt pulses from high powered lasers to produce milliwatts of harmonic light" – hardly the safest way to build a cd-rom drive!

Penner's team has adopted a new approach, hoping to overcome some of these problems by dispensing with inorganic crystals and making use of multiple Langmuir-Blodgett (monomolecular) films of organic polymers. The films not only provide the non-linearity, but also act as an optical waveguide to confine the radiation. The result is a long interaction length that makes possible the generation of harmonic light with an efficiency several orders of magnitude greater than previously possible.

Much work still remains to be done before a practical and cheap blue emitter reaches the market-place. But Penner and his team have proved that all the complex optical and physical requirements of an efficient second harmonic generator can be incorporated into a polymer film structure made from inexpensive materials.



Monomolecular films of organic polymers could be used to frequency double infra-red laser to produce a blue light beam. Such films provide not only the necessary non-linearity but also act as optical waveguides.

No space for a leak

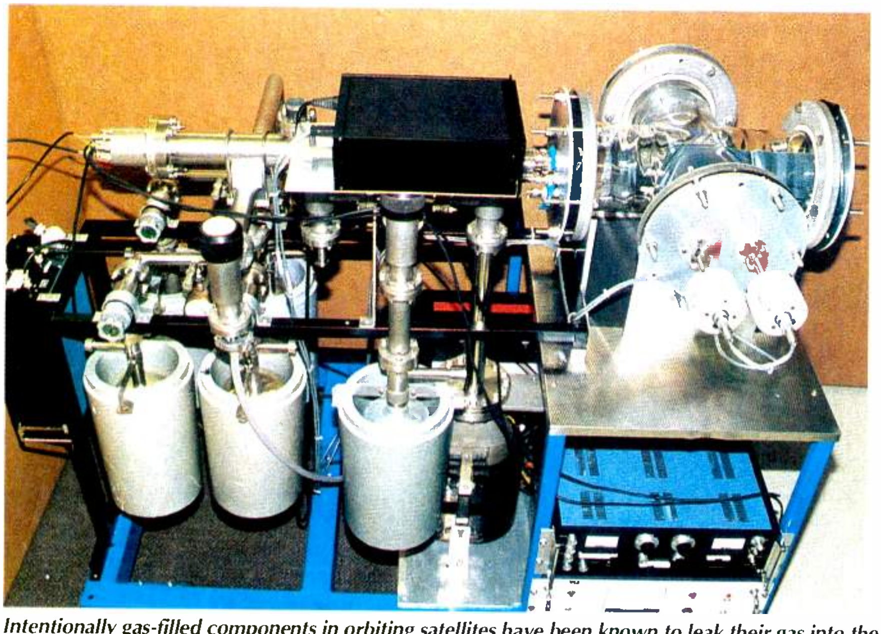
The dreaded blue glow that signalled the demise of a 'soft' valve was depressingly familiar back in the days when thermionics ruled. It happened when a faulty seal allowed air to enter the envelope and become ionised. So it is fascinating to hear today of what amounts to the opposite problem affecting the very modern world of space missions – components that fail when they lose their gas and become evacuated.

To tackle the fault, the European Space Agency (esa) has asked the Electronic Engineering Laboratory at the University of Kent to study the electrical breakdown of hermetically sealed space components that leak their intended gas filling into the vacuum of space. Working with the Belgian company Space Applications Services, the Kent project will work towards providing engineering guidelines that will avoid the failure which has caused the premature demise of equipment on several space missions.

The problem arises because many electronic and electro-mechanical components are encapsulated in hermetically sealed enclosures filled with inert gas at atmospheric pressure at the time of manufacture. This is done both to avoid contamination and also to prevent damaging discharges – the sort that occur in neon tubes, at pressures well below atmospheric.

But during space missions, components are exposed to a virtual vacuum for ten years or more. Over this period, if the hermetic sealing is inadequate, the original gas filling leaks away until the internal pressure in a component is low enough for discharge to take place, often destroying the component.

Perhaps the Kent group will have to re-invent the valve. ■



Intentionally gas-filled components in orbiting satellites have been known to leak their gas into the surrounding void, causing failure. Researchers at Kent University are using the vacuum equipment shown here to research this problem – the inverse of 'soft' thermionic valves.

Research Notes is written by John Wilson of the BBC World Service.

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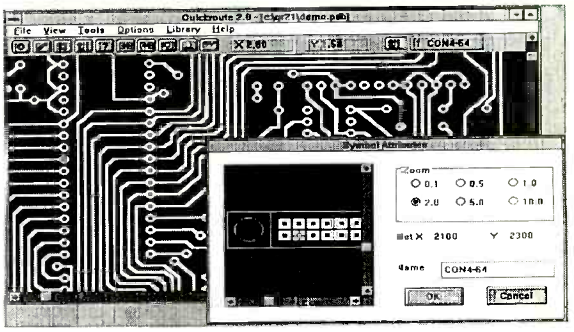

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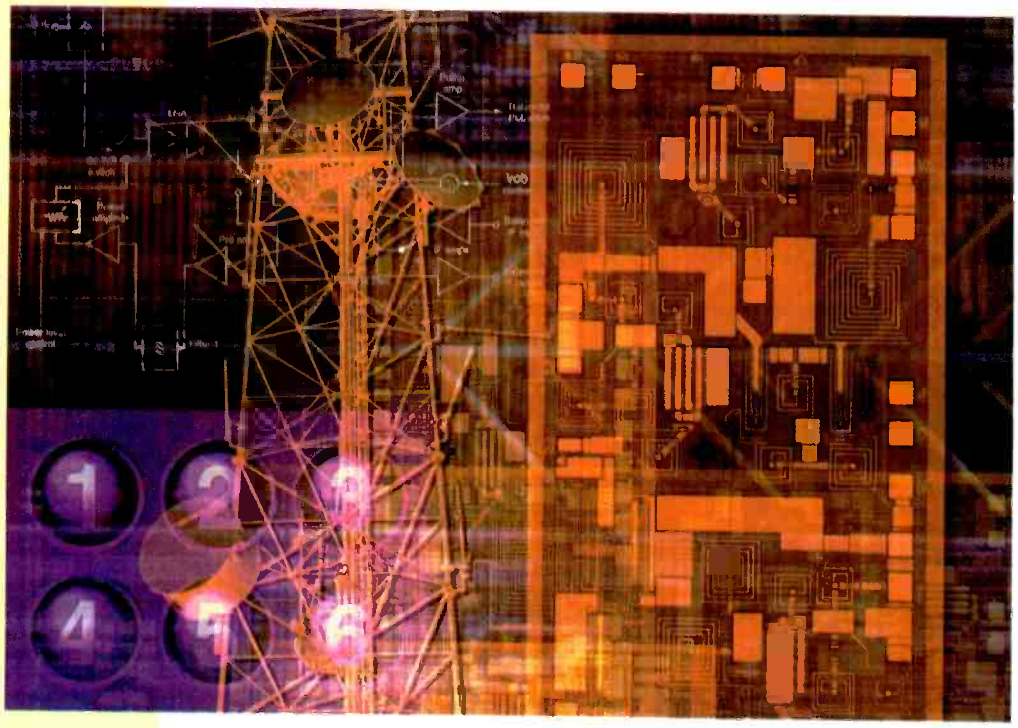

New wave MICROWAVES

Microwave engineering now finds itself within mainstream electronics design, mainly through the boom in wireless and personal communications. The semiconductor companies have responded with integrated components which take much of the rigour out of system design. Mike Hosking begins an extended series which explains the engineering behind the new microwave technology.

Mike Hosking is a lecturer in telecommunications and microwaves at the University of Portsmouth.

1: basic concepts, circuits and devices

The actual starting point of the microwave region varies somewhat in definition, but can conveniently be stated as commencing from about 1000MHz and extending to 30GHz. This equates to wavelengths from 30cm to 1cm. Millimetric waves. Sub-millimetre waves extend to 300GHz (1mm) and on to sub-millimetre bands, often referred to as far infra-red. This part of the spectrum supports applications such as inter-satellite communication links, atmospheric studies and astronomy. However, the scope of this series will be restricted to microwave and millimetre bands.

Microwave development in earlier times was mostly defence driven and GaAs integrated circuitry a defence technology. The basic semiconductor and tube devices are still around, although tube applications below the millimetre bands, except for high power, have now been mainly replaced by solid state. The GaAs fet now dominates the semiconductor field to above 100GHz. A derivative, the high electron mobility transistor (hemt) has revolutionised receiver noise figures. Development pressure has changed from defence to commercial markets for personal communications, mobile computing, traffic sensing and control and navigation systems.

Digital switching and modulation rates now extend to several GHz or Gbits. Thus high frequency circuit design must now be applied to circuit layout and impedance matching in these digital systems.

There are two main reasons for using microwaves: antenna considerations and wider bandwidths. Apart from general broadcast antennas and mobile communications, most other systems are concerned with focusing the microwave signal into a narrow beam for efficient transmission in a particular direction. This may only be achieved by increasing the number of wavelengths which will fit into the dimensions of the antenna. A small, directional antenna means high frequencies.

As an example, the half-power beamwidth of a low-sidelobe directive antenna is approximately $70\lambda/d$ in degrees, where λ is the wavelength and d is the aperture dimension in the plane of the beamwidth. So a domestic satellite TV receiver operating at 11.9GHz and having a dish antenna 600mm in diameter would have a beamwidth of about 3°. However, to achieve this same beamwidth at 100MHz would require an aerial dimension of 70m. On the other hand, a 600mm antenna for an inter-satellite link operating at 300GHz could deliver a beamwidth of only 0.12°.

Thus, practical systems requiring directive beams must operate at these microwave frequencies. One could carry this logic further and propose still smaller wavelengths in the infra-red bands. However, aiming such narrow beams would be a problem and infra-red does not penetrate clouds, rain, fog or smoke.

Basic concepts

Microwaves are the one region of the e-m spectrum where physical dimensions of the components and circuits are the same order of size as the wavelength itself, typically centimetres and millimetres in this case. Thus, as well as any intrinsic electrical properties which a component might possess, its actual size and shape also dictate the way in which it behaves in a circuit. If we reflect that, at lower frequencies, an 'inductive' or 'capacitive' effect is produced by the way in which an inductor or capacitor alters the phase difference between voltage and current, then a similar effect may be produced by perturbing the electric and magnetic fields of a microwave circuit. Thus, the actual shape and size of sections of circuit can make them appear as inductive or capacitive components.

The problem which has to be solved is that of relating transmission line geometry to an equivalent component value, and vice versa. Again, due to the wavelength size, it is no longer convenient to think in terms of voltage and current. These do not have unique values in a circuit: they are a function of position and are not directly measurable at these frequencies. Instead, power and impedance are the most usual circuit parameters to be determined, with circuit design and analysis being based on the determination of the electromagnetic field configuration at particular points in the circuit.

This has an interesting corollary. Lower-frequency electronics, where size and shape are not taken into account, is thus really an approximation (albeit a good one) to the more rigorous field analysis.

A further feature of high frequency circuits is the phenomenon of skin effect. At dc, the current in a conductor is distributed uniformly throughout its cross section but, with an alternating signal, there is a tendency for the current to concentrate towards the surface. This effect increases with frequency until, at microwave frequencies, the current is contained within a thin 'skin' at the conductor surface. Typically, skin depth is less than 1μm: it is important to ensure that conductor thicknesses are several times the skin depth (mainly applicable to integrated circuits) so as to avoid excessive conductor loss.

Lastly, the microwave signals have to be "guided" from point to point, either over a short distance within a microwave circuit or over a relatively long distance. Random shape, size of wire or circuit layout cannot be used for this purpose as, due to the short wavelength, the signal would radiate. Also, the wire or circuit would appear to the electromagnetic fields as some strange impedance inevitably causing mis-match, reflections and inefficient transmission. This applies both across a circuit board and from antenna to system front end. Hence, even in non-microwave circuits such as high-speed switching or computing, a microwave approach must be taken with the design and layout. Radiation and reflections will manifest themselves as unwanted coupling and interference.

As frequency increases, the alternating current in a conductor concentrates increasingly closer to the surface. The current density is a maximum at the surface and gradually decreases with penetration distance into the conductor. Skin depth is defined as that distance for which the current density has decreased to $^1/_e$ of its value at the surface, where e = 2.7182. It is a function of both conductivity and frequency and is given by:

$$\delta_s = \left(\frac{2}{\omega\mu\sigma}\right)^{\frac{1}{2}}$$

ω being the radial frequency = $2\pi f$; μ the permeability = $\mu_0\mu_r$ and σ the conductivity. **Fig.**

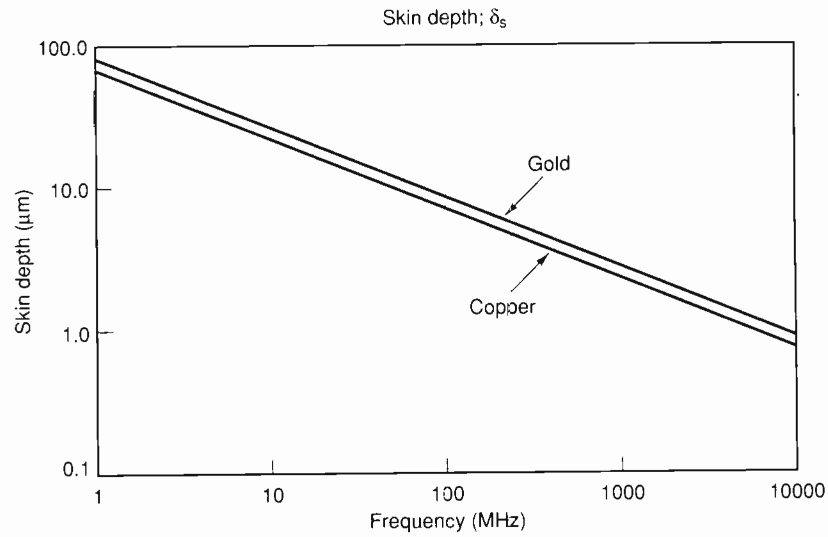

Fig. 1. Variation of skin depth with frequency.

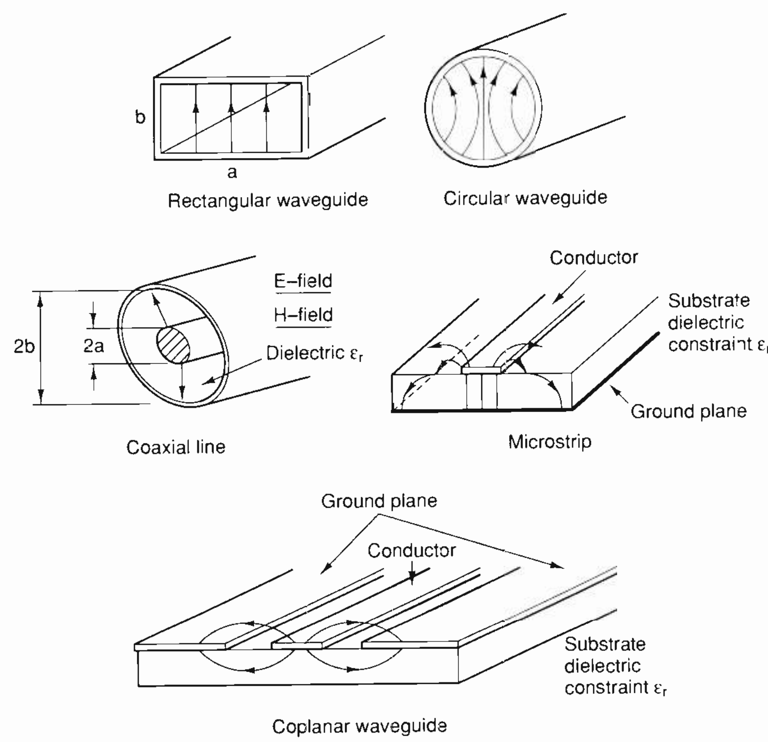

Fig. 2. E-field patterns of various forms of microwave transmission line are influenced by boundary conditions. In the rectangular guide, the E-field is paralleled to the narrow dimension for the dominant mode. Generally, a=2b. For the coaxial line, characteristic impedance is a function of ratio b/a. In microstrip and coplanar waveguides, phase velocity is modified by the proportion of field in air and dielectric. This leads to an effective dielectric constant different from ε_r.

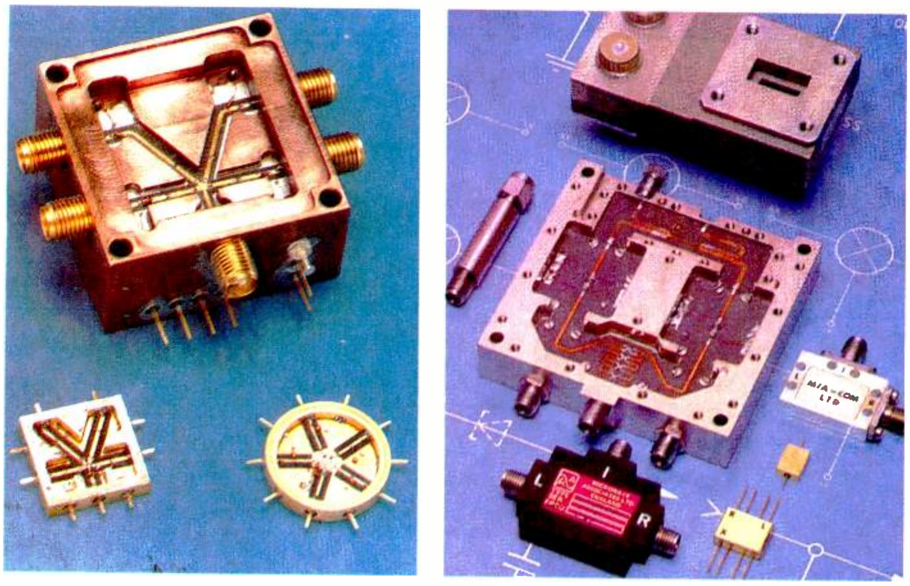

Fig. 3. There are various ways of designing microwave circuitry. For instance each circuit function may be treated as a separate component fabricated in microstrip, say, packaged in its own enclosure with input, output and supply connectors and with its individual microwave and environmental specification. In (a), left, are multi-throw switches in packaged microstrip, incorporating pin diode active devices. Photo (b) shows microwave mixer and detector modules. (M/A Com Ltd).

1 shows the variation of skin depth with frequency for two commonly used conductors of copper and gold.

In order to avoid excessive attenuation through conductor losses, the conductor thickness must not be less than the skin depth (δ_s) and, in practice, should be at least $5\delta_s$ thick. This isn't a problem with waveguides and coaxial lines, but may be significant with thin film or integrated circuits.

Historically, the rectangular and circular waveguide, together with coaxial line were the first types of transmission line and still have specialised application; their properties will be reviewed later in the series. Integrated circuitry in high-volume applications uses the planar forms of microstrip and coplanar waveguide (CPW). There are other variants of the types shown in **Fig. 2**, but they are much less commonly encountered. All of these transmission lines force the fields to conform to certain boundary conditions and patterns to propagate down the line. These field patterns are termed modes and, for any particular type of transmission line, are dependent on the detailed geometry and frequency of operation. Planar forms of microwave circuit use mainly microstrip but increasingly CPW at millimetre wavelengths.

Circuit building blocks

Transmitter applications, such as those for communications or radar systems would require the following circuit blocks: oscillators and/or up-converters to generate the microwave signal; phase, frequency, amplitude and electro-optic modulators; filters of different types; controllable attenuators and phase shifters; controllable switches for signal routing; isolators and circulators; low and intermediate power amplifiers. The receiving side might additionally include: low-noise amplifiers; mixers (down-converters) and detectors; limiters. In addition, many signal processing and switching applications operate well into microwave frequencies or bandwidths, to perform such functions as: IF amplification, delaying, on/off switching, correlation, matched filtering, spread spectrum.

There are a number of design approaches to microwave circuitry. For instance each circuit function may be treated as a separate component fabricated in microstrip (say), packaged in its own enclosure with input, output and supply connectors and with its individual microwave and environmental specification. **Fig. 3a** shows an example, a selection of single pole multi-throw switches using pin diodes as active devices.

Typically, the circuits may be fabricated on PTFE/ceramic substrates having a relative permittivity of about ten. The radial tracks are 50Ω transmission lines about 1.3mm wide. They incorporate PIN diode chips which are bonded in series-shunt configurations to the tacks. These diodes, the subject of a future article, can be switched between high and low impedance states. This is done by applying a dc bias current routed to the devices by the small air-cored chokes just visible.

The top component is a single-pole four-throw switch, completely packaged and tested as a self-contained component building block. They have SMA connectors for the microwave signals. Careful positioning of the diodes, together with design of the central junction is important. It affects matching, wide-band performance (often multi-octave, i.e. 2 to 18GHz), insertion loss and isolation. The lower two components are also multi-throw switches, but are designed to be used as drop-in modules in stripline circuits.

Fig. **3b** is a selection of mixer and detector modules in waveguide, coaxial and integrated circuit form. Normally, the waveguide mixer would cover the full band corresponding to its size, i.e. 8.2 to 12.4GHz. The coaxial detector could typically cover an eleven-octave band from 10MHz to 20GHz. Depending on the application, the diodes used are either silicon, gallium-arsenide Schottky barrier or point contact types.

These circuits use active devices bonded onto the microstrip circuit in either chip or packaged form. Such components are then referred to as hybrid ones. More complex subsystems may then be assembled by interconnecting appropriate individual modules. Below about 40GHz, semi-rigid, copper jacketed coaxial cable is often used as the connecting transmission line.

As an alternative, many individual functions may be located within a single package and possibly on the same microstrip substrate. **Fig. 4** shows a multistage fet amplifier operating in the 2 to 6GHz range, assembled from 16 indi-

Integration with a difference

Microwave circuit design uses changes in the dimensions of the transmission line to change the phase relationship between the E and H fields, thereby creating inductive and capacitive effects. Lengths of the line itself can also be used to perform impedance transformation. Thus, such circuit elements must be a significant fraction of a wavelength in extent (in fact, anything up to a half wavelength) and this places a lower limit on the size of an integrated circuit.

Microwave circuits confine and manipulate electromagnetic fields. Thus, adjacent components in an integrated circuit cannot be placed arbitrarily close together, or else unwanted coupling of the fields will occur. Typical separations might be two to three times substrate thickness. Again, this imposes a size constraint.

Impedance matching is important in microwave circuit design, no always for reasons of maximum power transfer, but often in order to avoid multiple reflections within a circuit. This is especially important, for example, in high-speed pulse or fast switching circuits.

Microwave integrated circuitry is almost always analogue in operation combining a range of different functions. Thus, there is no equivalent to the digital LSI incorporating thousands of identical elements with built-in redundancy.

The next article in this series will focus on particular techniques of microwave circuitry using microstrip and will show how many of the particular components are designed.

vidual microstrip circuits and presented as a single packaged component.

Most of the microstrip circuitry in each of the stages consists of input and output matching networks for the fets, together with biasing. Substrate material is alumina with a relative permittivity of about 9.8. The interdigital circuit elements are called Lange directional couplers and have the important property of being able to operate over a multi-octave bandwidth.

A more varied circuit such as a complete receiver front end could also be made in this way. It might include low noise pre-amplifier, filters, mixer, local oscillator and IF amplifier. Such integrated components are still hybrid in that they contain discrete active devices bonded into passive microstrip circuitry.

A higher level of integration requires a fully monolithic microwave integrated circuit (mmic). It happens that semi-insulating GaAs has good transmission line properties and so may be used as the microstrip substrate on which passive microwave circuit elements may be designed and fabricated. However, instead of incorporating discrete active devices as in hybrid circuits, the substrate itself may be selectively doped, etched and processed to produce active regions: fets and Schottky barrier diodes, for example. This construction is effective well into the millimetre wave region.

A design example is shown in **Fig. 5a** which is a complete transceiver on a GaAs substrate operating at 2.4GHz for wireless lan application. The circuit is approximately 5mm square on a 0.125mm thick substrate and includes the main building blocks, shown in **Fig. 5b**: high-power and low-noise amplifiers, filters, switches, oscillator, down converter, as well as numerous minor components.

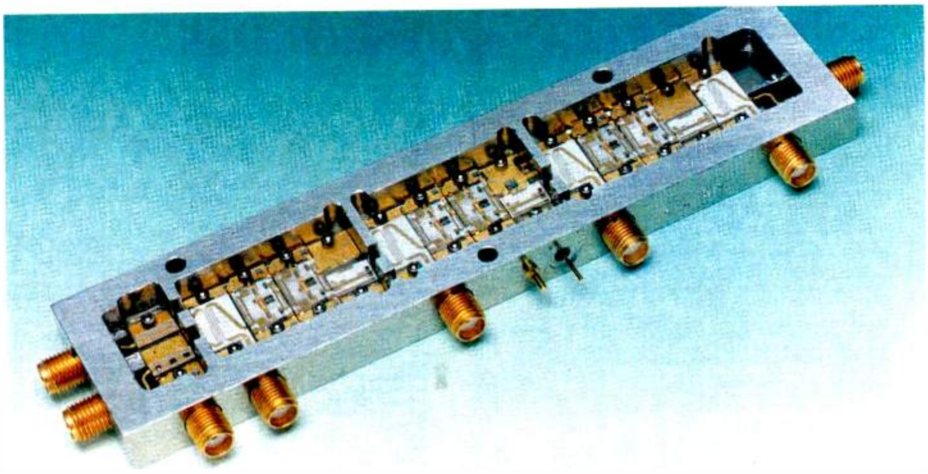

Fig. 4. Multi-stage hybrid amplifiers in ceramic microstrip transmission lines. (M/A Com Ltd)

Semi-insulating GaAs forms the substrate material, with the active devices being created by localised doping and processing. Lumped-elements are used for the reactive components in the form of small chip capacitors, air-bridges and printed spiral inductors. An on-chip fet local oscillator can be voltage tuned from 2 to 2.2GHz to track different input frequencies. A 5dB noise figure is achieved from the receiver. Output transmit power can be switched between 20dBm and 10dBm, or 100-10mW. Another advanced feature of the chip is the very low standby current of <0.5mA for the whole device.

So, which design approach to adopt? As usual, the answer is a trade-off between production quantity, yield and initial design investment. Many companies worldwide specialise in the design of discrete components suitable for small quantity applications, prototype systems or in cases where short-term changes to the design may be required.

The integrated approach is used where a larger production quantity is required, for which the performance requirements will remain stable. Such applications might typically apply to a system production run for a missile or satellite sub-system. Although GaAs mmic technology has existed for more than a decade and many successful devices have been produced for specialised systems, there have been few large-scale applications. Such applications are necessary in order to amortise the high manufacturing investment and relatively low yield of such circuits. They are beginning to come through with mobile and associated communications, satellite TV and global communication services. ■

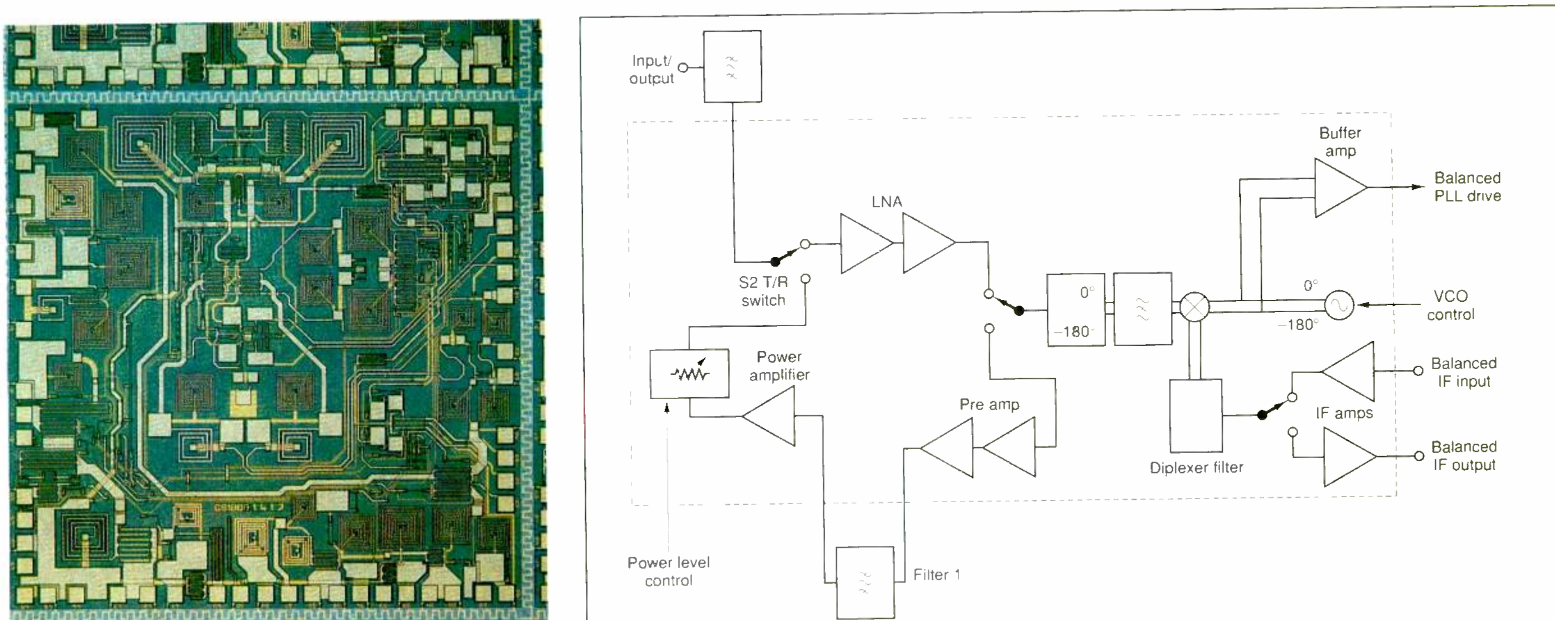

Fig. 5. Photo a, left, shows an example of an advanced GaAs mmic: a 2.4GHz transceiver for wireless lan application. Semi-insulating GaAs forms the substrate material, with the active devices being created by localised doping and processing. Lumped-elements are used for the reactive components in the form of small chip capacitors, air-bridges and printed spiral inductors. An on-chip fet local oscillator can be voltage tuned from 2 to 2.2GHz to track different input frequencies. A 5dB noise figure is achieved from the receiver. Diagram b shows main building blocks of the transceiver. (GEC-Marconi Materials Technology)

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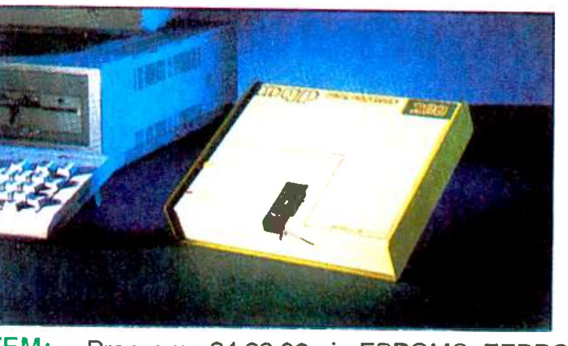

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Mac man

How curiously out of touch are many of the views and opinions expressed in *EW + WW* these days on the subject of computers. I am not sure I like being regarded as structured but lifeless (*Comment*, March), while the idea that programming is an underclass to electronics and that creativity, elegance and innovation are foreign to software is truly bizarre.

Then we have the notion that reverting to some incomprehensible command line interface just to save a few kilobytes or milliseconds is a great idea.

If exposure to graphical interfaces is limited to the rather inefficient, poorly structured and unattractive *Windows*, then perhaps these views are understandable. Fortunately, like dos, *Windows* is not long for this world.

For a proper gui with a user-centred approach that actually allows you to get your work done, buy a Macintosh – not a partial simulation.

All this nostalgia for lean, mean programs written in assembler completely ignores the fact that if written like this, software would take so long to develop and be so hard to debug that nothing would ever see the light of day.

Yes, I too remember the first personal computers and how small their application programs were. I also recall the narrow scope of their abilities. Memory is cheap, my time is not.

Lastly, there is just no room for more than two or perhaps three major players in desktop computing. The *Archimedes*, fine performer though it is, has missed the boat as have so many other platforms. Without a large body of first-rate software, constant innovation and global distribution there is no hope for the success of any computer, no matter how technically interesting.

Incidentally, I designed circuits long before I designed software.

Joel M Sciamma
Windlesham
Surrey

Window fan

The main complaint of May, Silson and Standen (*Letters*, December 1993), that Microsoft *Windows* is slower than dos is quite true. But they barely touch on how much easier it is to use *Windows* applications compared to similar dos packages. After all, double-clicking a mouse button on an icon is easier than changing to the directory containing the required program, and then trying to remember its name before typing it. The idea that *Windows* is only an aid for children is nothing more than intellectual snobbery.

All too often we engineers and programmers forget that we are no longer in the majority of pc users. Secretaries, clerks, accountants, managers and directors have no need – and often no desire – to learn the ins and outs of dos. For them, *Windows* is good because it is easy to use. If their only concern was for speed, multiprocessor systems would be much more popular than they are.

Rather than stand and criticise, engineers should provide software that people can use and understand readily, without dictating the terms of its use too greatly. If this means providing programs to run under *Windows* as well as ones to run under dos, we must do so or face an inevitable loss of sales and respect from our customers.

Jason W Ross
Gillingham
Kent

Mathcad not incomparable

The conclusion reached in your review of *Mathcad* ("*Mathcad 4* – 16 bits of difference", February, pp.127-129), that it is the best tool for small scale linear modelling, is unjustified. The article did not consider any other product. What about *Spice*, *Maple*, *Theorist*, and my favourite *Mathematica*? These are all important tools for engineers and should not be unfairly dismissed without comparison.

For example *Mathematica* with *Nodal*, can easily determine the characteristics of *LCR* circuits and solve differential equations.

Their cost is such that few have the luxury of purchasing more than one. So you should be very careful before you allow your pages to carry a "best buy" tag for any one product without a full and considered review. Or I fear that the credibility of *EW + WW* will suffer.

Richard Thomson
Edinburgh

Cable communication

Gareth Connor (*Letters*, *EW + WW*, March) makes a number of points about the "sound" of cables.

As a professional broadcast engineer, I used to be sceptical of the claims made for cables, believing there were a whole lot of other problems to be dealt with first. But I was recently given silver and ptfe to make up some cables and was most surprised to find they sounded superior to a commercial product – very little time and money was involved so I don't feel I have an axe to grind.

Connor asked: "If the signal has already passed through several thousand metres of cables, pcbs and connectors, what difference will a mere dozen or so metres of exotic cable make?"

Thousands? Hundreds, yes. But broadcast and recording studio signals are invariably balanced, and (excepting microphone cables) balanced at a far higher level than their domestic counterparts. Far greater immunity to rfi should be the result.

Professional cables also tend to be driven by proper line-sending amplifiers rather than a single emitter follower or op-amp. So the driving amplifier is less likely to suffer rfi driven back into its feedback loop.

Professional cables may be longer, but they are better driven and terminated.

On the subject of silver and ptfe, I cannot think of any reason why silver should be better, as the improvement in conductivity over copper is so small.

But the good physical reason for

Dab it...

Your comprehensive survey of dab (*Comment*, *EW + WW*, February 1993) reminds me of Cassandra, who was able to foretell the future and was also doomed to see it happen.

Anyone with a modicum of understanding of politics within Europe and the prevailing economic climate, can see that dab is down at the first fence and unlikely to get up.

The BBC is already thought to be having serious doubts, so the technology will certainly follow the fate of D2-MAC and RDS. Only three countries might be interested in funding dab – assuming industry is ready to provide the hardware: UK, France and Germany. The others will watch which way the wind blows.

Germany simply cannot afford it for at least three years and has said so. A commitment by the DBP to the established and technically-successful DSR has already been demonstrated, since it provides an excellent distribution vehicle for the extensive cable network in that country. Germans are already calling it *SuperRadio* since it is on a par with cd quality and recognised as such by the fidelity-conscious listener. Sixteen stations already function on one transponder and two more have now been designated for extension of the service.

France, as always, will continue to go its own way. No one but the French can receive the French digital radio service.

And, as always, the UK's policy will be left to an eventually-privatised BBC. We know what will happen then – Nothing. At the present rate, there won't be any engineers left to implement it.

Will we never learn? The public still resists even fm and will never be persuaded of the virtues or otherwise of dab.

That could be a blessing in disguise. The decision will be influenced by factors quite out of the control of those who think they have settled on Eureka 147 as a European standard.

Reg Williamson
Kidsgrove
Staffs

...Off

As one of the "few readers of *Hi-Fi News*" mentioned in Norman McLeod's excellent article ("Dab – delivery, delay or debacle?", *EW + WW*, February, pp.160-165) I want to point out that truncation of sound was introduced specifically for car users. The BBC's acknowledged first priority with dab is to get it right for mobiles, with the rest to follow.

The antidote (though not for cars) is the digital satellite service DSR.

This is a service that has been a secret over here until recently and is still under wraps as far as the national media are concerned – not to mention the BBC.

However it is a benchmark by which to judge others, regardless of its outmoded technology, wasteful use of bandwidth etc.

For instance, when R3 fm broadcasts a live European Broadcasting Union concert, it may be compared with up to three German dsr regional transmitters broadcasting simultaneously on the Bundespost's TVSat2. They are usually louder and clearer, especially on winter evenings when satellite reception improves after dark.

With due respect to the BBC, dab and musicam, we few static audio fans will sit in our carefully prepared dens praying that the Bundespost "bird" doesn't die before we do.

Hugh Haines
Sunderland

using ptfe is that it has a very low value of $d \approx 0.0005$, known to correlate strongly with dielectric absorption (read any literature on sample and hold circuits for a-to-d convertors).

Many cables use pvc as an insulator, which has a far worse value, $d \geq 0.01$ – you don't see many pvc capacitors around. This difference in d and consequent dielectric absorption could well result in smearing of transients.

I was glad to see that Connor felt that the claims about silver and ptfe should be investigated, until I realised he was being sarcastic. Nevertheless, I intend to conduct some (double) blind tests to see if I can find any physical explanations.

Nobody suggests that the colourful vocabulary used by wine connoisseurs to describe the subjective sensations of drinking a fine wine should be supplanted by gas spectrograph measurements. So why do sensible engineers get into such a tizzy when the topic changes to audio and music. It is a purely subjective and personal experience anyway?

Morgan Jones
Southampton

Sparks fly over early radio

I was interested to read George Pickworth's article, *The spark that gave radio to the world*, (*EW + WW*, November 1993, pp.937-942) on the technology of spark transmission.

I too have built scale size spark transmitters of all the main varieties and this combined with my fairly extensive collection of original documents, manuals and publications leads me to question several of his statements.

Firstly, the question of "arc versus spark" was indeed a matter of controversy at the time. But even when arcing did occur momentarily, it produced unwanted effects[1] – notably severe spark gap erosion and even more interference than usual.

A spark system relied fundamentally on the shock excitation of either an aerial earth system or coupled tuned circuits. So any actual arcing would cause the tuned circuit to be too tightly coupled to the supply and reduce the efficiency of the transmitter.

Early operators often noticed that highly-quenched systems caused less interference and in practice were easier to read. In fact the quenching referred to arcing rather than sparking: the circuit was shocked and then left to oscillate as freely as possible. This became a major selling point for the quenched gap German Telefunken system.

The next problem arises over the Slaby/Wien quenched gap. I must take issue with its description as a "magnetic" device. This was put forward at the time as a theory of operation. But a study of its origins and derivation[2] will show this to be unlikely. Magnetic effects are proportional to current and we have here a high voltage, low current device. My own feeling, based on theory and practice, is that thermal and electrostatic effects were much more important in the quenching process. The spark rose up to the outer groove because of the heat it generated. Model gaps showing increased erosion at the top of the electrodes strongly support this idea. Also the quenched gap is the most difficult type to scale due to local heating concentrating in the relatively small and semi-sealed spark gap.

Slaby and Wein[2] derived the quenched gap from work based on plain gaps in series. This alone produced a worthwhile improvement in efficiency. With such an arrangement, magnetic fields would have virtually no effect.

It would be fascinating to get an actual measurement of fields involved in the discharge, though a spark transmitter, like a car, is a very unfriendly place for modern electronics!

A further query concerns the various forms of the Marconi-type disk or rotary gap discharger. It appears that in the low frequency ac-excited stations of the period there was little attempt to synchronise spark gap with the supply frequency, as implied in text[2] and shown as oscilloscope-style drawings. But more importantly, experiment with the model transmitter shows a great increase both in efficiency and "purity" of received signal when the gap runs synchronously with the 50Hz ac supply. A higher supply voltage would somewhat reduce this effect, but the improvement in signal quality would remain.

The Marconi company claimed on several occasions that its larger stations were not "spark". But the only discharger ever used by them which could be anything else was the original plain disc discharger. With a dc supply some very unstable forms of arc could have resulted but with negligible contribution to rf output. Doubtless[4] Marconi reverted to stud electrodes to improve readability with magnetic detectors in particular. But the magnetic detector was a device working on signal peaks[5] and so even a relatively smoothly modulated signal would be less audible.

A study of true arc transmitters soon shows that generation of large amounts of rf energy means going to considerable lengths, including: a powerful transverse magnetic field; rotating electrodes at a constant distance, and a hydrogen or hydrocarbon atmosphere[6].

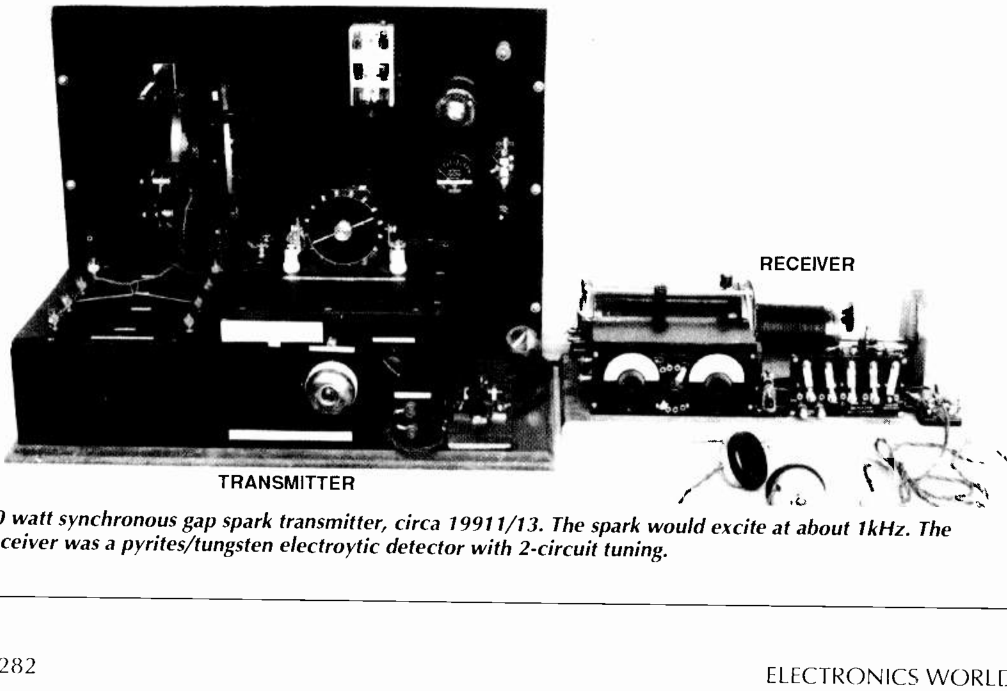
Not one of these, except perhaps a form of rotating electrodes was present in any spark transmitter. So, except possibly at very low frequencies, the contribution of arc generated signals to the output of a transmitter was vanishingly small. Certainly a system as used by Marconi at Clifden in 1907/8[7] was *not* an arc transmitter. The spark frequency was probably mainly determined by the time constants of the low frequency chokes and other components in the keying circuit. Normal practice was to relate these component values to the spark frequency. A model plain-disc discharger produced good power output but poor signal quality with 500Hz excitation – in fact showing the symptoms of poor quenching.

In the US a few years later, Federal Wireless faced problems[8] in scaling up the Poulsen arc that made clear that a metal to metal arc burning in air simply will not generate a worthwhile amount of power, even at 40kHz or so. Sustained oscillations above audio frequencies are not attained[9].

The fundamental point is that arc transmitters were *not* shock excited, rather they relied on the negative resistance characteristics of an electric arc.

In modern terms, the output waveform of Clifden (for example) would have been an overmodulated carrier but with lower peak signal levels than the studded gap. Stations using crystal or valve detectors would be quite effective but magnetic detectors would have been disadvantaged – hence the modifications carried out by Marconi.

C C Wright
Auckland
New Zealand



20 watt synchronous gap spark transmitter, circa 19911/13. The spark would excite at about 1kHz. The receiver was a pyrites/tungsten electroytic detector with 2-circuit tuning.

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Ham fist

Why should radio amateurs be expected to work at the same level as professional r&d labs ("New challenge for amateur radio", *Comment*, December 1993)? No one expects amateur pilots to fly 747s or weekend golfers to come in at nine under par.

Amateurs that DO work at this level will, as you pointed out, be professionals using the amateur licence as a flag of convenience and this is no more amateur radio than is the hobby of radio dx-ing.

The 1950s and 60s were, by your own admission, "boom periods" in amateur radio. But do you recall the thickness of the amateur call book in 1960? Thin, wasn't it. There are less radio amateurs now, with many who would have been radio hams in 1960 entering other branches of hobby electronics like computing and electronic music. So it stands to reason that most amateur transmitting licence holders today are *not* radio amateurs, they are radio dx-ers.

Radio dx-ing is, effectively, a collecting hobby with radio being secondary. The hobbyist scans the bands for new rare calls like the philatelist scans the markets for new rare stamps. It is this hobby, not amateur radio, that depends on the Japanese transceiver with its attendant retailers and service departments (whoever heard of a radio amateur taking a set to a dealer to be mended for heaven's sake?).

So what have the remaining radio amateurs been doing – amateur radio of course. Coaxing ex pmr sets to work on the vhf bands. Taking advantage of the cb boom and getting the cheap multi-mode transceivers going on 14-70MHz. Making straightforward, simple equipment and learning much for our own amusement, but hardly breaking into new frontiers.

What gives our hobby the right to survive? Purely the fact that it is instructive, harmless and fun. In a free country no other reason is needed and governments who expect us to fight their wars for them would to well to remember that!

Long live amateur radio, may it continue to be not a scrap of use to anybody.

Stephen Dyke
Sandy
Bedfordshire

I feel that we are both on the same side – Ed.

Reproducible losses

After upgrading my photocopier, I wanted to sell the old one – a Sharp *SF-740* in perfect working order – but was surprised to be told that toner for it was no longer available.

I was also told that this can happen with some machines when they are only five years old. Parts are not interchangeable and one retailer said they throw away three perfectly good machines each week as supplies were no longer available.

Maybe electronics hobbyists should visit their local copier shops and, for a few pounds, try to buy these machines for their motors, switches, power supplies and even microprocessors – not to mention lenses, prisms and mirrors.

All copiers work on the electrostatic principle, using a black fusible toner. Yet it appears that toner from one make of machine will not work in another. Would it be possible to modify machine A to work with toner intended for machine B by altering the electrostatic charges on the wires? Or an agency could be set up to supply toners for obsolete machines.

It does seems absurd that we are willing to discard perfectly good machines that have an environmental cost in their manufacture.

Perhaps some *EW + WW* readers have some ideas as to how this situation can be remedied.

John de Rivaz
Truro
Cornwall

Bass-is for upset neighbours

May I express my interest in your bass special *Bass from all angles*, pp. 100-122, *EW + WW*, February) as it is a subject I know well – up to and including a speaker cone that tore itself to strips before my eyes!

The result is a "beautiful sound" but very anti-social! Like Mr Macaulay, I finally abandoned my sub-woofer as I just couldn't use it with any pleasure. Not that there are many sources of 20Hz sounds unless you rely on oft-repeated discs and tapes.

So do be warned, you have to live with the consequences. I get a wry smile when I contemplate the front picture of that issue... I know just how he feels.

Ronald G Young
Peacehaven
East Sussex

Circuit analysis

Hyland's oscillator circuit (*Circuit Ideas £100 winner*, January) is certainly neat and interesting.

But with its "bottom" stage being a phase non-inverting common emitter, the title "cascode" is hardly appropriate. Rather it has some similarity to a long-tailed pair with input to base A, output from B.

Output impedance at the point shown is also not necessarily low at all. Essentially it is the parallel resonant impedance of L_1C_1 (not L_1C_2 as misprinted), depending on the LC ratio and Q factor of the tuned circuit, shunted directly by the output load, and to some extent by transistor saturation. The components shown, and a Q factor of 50, would give a tuned impedance of about 50Ω – hardly low.

On the other hand, the emitter of the lower transistor would provide perfectly good low impedance output, probably best via a resistor of a few kΩ. It may then be necessary to bleed some current from the +ve rail to the emitter to prevent the transistor being cut off during part of the cycle.

When operating at high frequencies, a few pF shunted across R_3 may be needed, of the order of 0.25 of C_{cb} of the lower transistor.

Gerald Pasch
London NW2

Arts and spacecraft

Artificial intelligence researchers have failed to shed light on the nature of intelligence by teaching computers to play chess. What they should be doing is building computers with hard-wired networks for the arts. Arts disciplines reflect the structure of the human body. Each transducer (ear, eye etc) has an associated art discipline (singing, drawing etc).

To process signals from the outside world the brain must have evolved a signal processing network for each transducer. A simple model of the brain is a set of arts processing networks (music, dancing network etc) together with a centre for emotion (providing motivation) and one for memory. If a network is stimulated by a set of regular – say symmetric – signals we experience beauty.

In contrast, science is not internal to the mind (subjective) but external (objective). So the brain has not evolved networks to process scientific disciplines.

The idea can be extended to imagine possible alien intelligences. On a planet with little light for example, an alien may have evolved sonar to help navigate (like bats). To process signals from its sonar receiver, it would need a sonar signal processing network inside its brain. So the alien's art disciplines could be sonar-based. Sonar opera for example.

John R Owen

Know?...

I am interested in transmitting and audio 1930-55 but am having difficulty finding relevant literature. Does any *EW + WW* reader have surplus literature, data books etc that I could use?

I also urgently need three VT4c/211 valves (mod spec) for a ww2 transmitter currently undergoing restoration. Any assistance would be much appreciated.

M J Evans
37 Foxwell Street
Worcester WR5 2ET

...Kan do?

I need a 30-4000Hz range simple pll oscillator circuit with an output power of 4W or more, either pulsed or harmonic, with a TTL reference signal.

For my requirement, there is no point buying a costly 0-2MHz commercial synthesiser with –70dB harmonic distortion or phase noise.

Can any *EW + WW* reader help me?

S Kan
13 bis rue de la Valle
St Remy les Chevreause
France

Light inconstant

Recent independent research in Australia and Russia indicates that the speed of light has undergone a precipitate decrease. Since the first measurements by Roemer in 1676, thousands of readings using sixteen different methods, have given consistently lower values, changes not due to observational error or limitations of equipment.

Michelson measured c between 1879 and 1926 and noted a steady decrease. In 1931 DeBray reported that: "invariably, new determinations give values that are lower than the last". Statistical regression analysis by Barry Setterfield and others has shown that many "constants", eg Planck's, the gyromagnetic ratio, Boltzmann's, Rydberg's etc have all displayed a trend that could be expected by a falling c.

The implications are far-reaching in astronomy, the red shift, geology, the rate of radioactive decay, the atomic clock, biology, physiology, the fossil record etc, and are part of increasing evidence indicating that the age of the universe may be thousands of years, not billions. (See "The facts of Life: Shattering the myth of Darwinism", Richard Milton, 1992 and "Science versus evolution", M Bowden, 1991.)

Amnon Goldberg
London

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Mainframe Plus 8552A IF Plug-In Plus 8556A RF Plug-In 20Hz – 300kHz Plus 8553B RF Plug-In 1kHz – 110Mc/s. Tested with instructions – £700.

Marconi TF2008 – AM-FM signal generator – also sweeper – 10Kc/s – 510Mc/s – from £250 – tested to £400 as new with manual – probe kit in wooden carrying box.
HP Frequency comb generator type B406A – £400.
HP Vector Voltmeter type 8405A – £400 to £600 – old or new colour.
HP Sweep Oscillators type 8690 A & B + plug-ins from 10Mc/s to 18GHz also 18-40GHz. P.O.R.
HP Network Analyzer type 8407A + 8412A + 8501A – 100Kc/s – 110Mc/s – £500 – £1000.
HP Amplifier type 8447A – 1-400Mc/s £200 – HP8447F. 1-1300Mc/s £400.
HP Frequency Counter type 5340A – 18GHz £1000 – rear output £800.
HP 8410 – A – B – C Network Analyzer 110Mc/s to 12GHz or 18GHz – plus most other units and displays used in this set-up – 8411a – 8412 – 8413 – 8414 – 8418 – 8740 – 8741 – 8742 – 8743 – 8746 – 8650. From £1000.
Racal/Dana 9301A – 9302 RF Millivoltmeter – 1.5-2GHz – £250-£400.
Racal/Dava Counters 9915M – 9916 – 9917 – 8921 – £1500 to £450. Fitted FX standards.
Racal/Dana Modulation Meter type 9009 – 8Mc/s – 1.5GHz – £250.
Marconi RCL Bridge type TF2700 – £150.
Marconi/Saunders Signal Sources type – 6058B – 6070A – 6055A – 6059A – 6057A – 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 PI – 18-26.5GHz or 6651 PI – 26.5-40GHz – £1000 or PI only £600.
Marconi distortion meter type TF2331 – £150. TF2331A – 200.
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.
Gould J3B test oscillator + manual – £200.
Tektronix Mainframes – 7603 – 7623A – 7613 – 7704A – 7844 – 7904 – TM501 – TM503 – TM506 – 7904 – 7834 – 7104.
Alltech 757 Spectrum Analyser – 001 22GHz – Digital storage + readout – £2000.
Marconi 6155A Signal Source – 1 to 2GHz – LED readout – £400.
Barr & Stroud Variable filter EF3 0.1Hz – 100kc/s + high pass + low pass – £150.
Marconi TF2163S attenuator – 1GHz. £200.
Farnell power unit H60/50 – £400 tested.
Racal/Dana 9300 RMS voltmeter – £250.
HP 8750A storage normalizer – £400 with lead.
Marconi TF2330 – or **TF2330A** wave analysers – £100-£150.
Racal/Dana signal generator 9082 – 1.5-520Mc/s – £500.
Racal/Dana signal generator 9082H – 1.5-520Mc/s – £600.
Tektronix – S12 – 7S14 – 7T11 – 7S11 – S1 – S52 – S53.
Marconi mod meters type TF2304 – £250.
HP 5065A rubidrum vapour FX standard – £2.5k.
Systron Donner counter type 6054B – 20Mc/s – 24GHz – LED readout – £1k.
Racal/Dana 9083 signal source – two tone – £250.
Systron Donner – signal generator 1702 – synthesized to 1GHz – AM/FM – £600.
Tektronix TM515 mainframe + **TM5006** mainframe.
Rhodes & Schwartz power signal generator SLRD-280 – 2750Mc/s – £250-£600.
Ball Efratom rubidrum standard PT256B-FRKL.
Farnall electronic load type RB1030-35.
Racal/Dana counters – 9904 – 9905 – 9906 – 9915 – 9916 – 9917 – 9921 – 50Mc/s – 3GHz – £100-£450 – all fitted with FX standards.
HP4815A RF vector impedance meter c/w probe – £500-£600.
Marconi TF2092 noise receiver. A, B or C plus filters.
Marconi TF2091 noise generator. A, B or C plus filters.
HP180TR, HP182T mainframes £300-£500.
Fluke 8506A thermal RMS digital multimeter. £400.
Philips panoramic receiver type PM7900 – 1 to 20GHz – £400.
Marconi 6700A sweep oscillator + 6730A – 1 to 2GHz – £500.
HP8505A network ANZ + 8503A S parameter test set + 8501A normalizer – £4k.
Racal/Dana VLF frequency standard equipment. Tracer receiver type 900A + difference meter type 527E + rubidium standard type 9475 – £2750.
HP signal generators type 626 – 628 – frequency 10GHz – 21GHz.
HP 432A – 435A or B – 436A – power meters + powerheads – Mc/s – 40GHz – £200-£1000.
Bradley oscilloscope calibrator type 192 – £600.
Barr & Stroud variable filter EF3 0.1Hz – 100Kc/s + high pass + low pass – £150.
Marconi TF2370 spectrum ANZ – 110Mc/s – £900.
Marconi TF2370 spectrum ANZ + **TK2375** FX extender 1250Mc/s + 1st gen – £1.5k.
HP8614A signal generator 800Mc/s – 2.4GHz, new colour £400.
HP8616A signal gen 1.8GHz – 4.5GHz, new colour £400.
HP 3325A syn function gen 20Mc/s – £1500.
HP 3336A or B syn level generator – £500-£600.
HP 3586B or C selective level meter – £750-£1000.
HP 3575A gain phase meter 1Hz – 13Mc/s – £400.
HP 8671A syn microwave 2 – 6.2GHz – £2k.
HP 8683D S/G microwave 2.3 – 13GHz – opt 001 – 003 – £4.5k.
HP 8660 A-B-C syn S/G. AM + FM + 10Kc/s to 110Mc/s Pl – 1Mc/s to 1300Mc/s – 1Mc/s to 2500Mc/s – £750-£2800.
HP 8640B S/G AM-FM 512Mc/s or 1024Mc/s. Opt 001 or 002 or 003 – £800-£1250.
HP 8656A S/G AM-FM 0.1 – 990Mc/s – £1500.
HP 8622B Sweep Pl 01 – 2.4GHz + ATT – £1750.
HP 86290B Sweep Pl 2 – 18.6GHz – £1250.
HP 86290B Sweep Pl 2 – 18GHz – £1000.
HP 86 Series Pl's in stock – splitband from 10Mc/s – 18.6GHz – £250-£1k.
HP 8620C Mainframe – £250. IEEE – £500.
HP 8615A Programmable signal source – 1MHz – 50Mc/s – opt 002 – £1k.
HP 8601A Sweep generator .1 – 110Mc/s – £300.
HP 4261A LCR meter + 16038A test leads – £400.
HP 4271B LCR meter 1MHz digital meter + 16063A test adaptor – £850.
HP 4342A Q meter 22kHz – 70Mc/s 16462A + qty of 10 inductors – £850.
HP 3488A HP – IB switch control unit – £500 + control modules various – £175 each.
HP 3561A Dynamic signal ANZ – £3k.
HP 8160A 50Mc/s programmable pulse generator – £1400.
HP 853A MF ANZ + 8558B – 0.1 – 1500Mc/s – £2500.
HP 8349A Microwave Amp 2 – 20GHz Solid state – P.O.R.
HP 3585A Analyser 20Hz – 40Mc/s – £4k.
HP 8569B Analyser .01 – 22GHz – £5k.
HP 3580A Analyser 5Hz – 50kHz – £1k.
HP 1980B Oscilloscope measurement system – £600.
HP 3455A Digital voltmeter – £500.
HP 3437A System voltmeter – £300.
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HP 3552A Transmission test set – £400.
HP 3763A Error detector – £500.
HP 3764A Digital transmission analyser – £600.
HP 3770A Amp delay distortion analyser – £400.
HP 3780A Pattern generator detector – £400.
HP 3781A Pattern generator – £400.
HP 3781B Pattern generator (bell) – £300.
HP 3782A Error detector – £400.
HP 3782B Error detector (bell) – £300.
HP 3785A Jitter generator + receiver.
HP 8006A Word generator – £100.
HP 8016A Word generator – £250.
HP 8170A Logic pattern generator – £500.
HP 59401A Bus system analyser – £350.
HP 59500A Multiprogrammer HP – IB – £300.
Phillips PM5390 RF syn – 0.1 – 1GHz – AM + FM – £1250.
Phillips PM5519 Colour T.V. pattern generator – £250.
S.A. Spectral Dynamics SD345 spectrascope 111 – LF ANZ – £2500.
Tektronix R7912 Transient waveform digitizer – programmable – £400.
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Tektronix 576 Curve tracer + adaptors – £900.
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Tektronix 7L5 LF analyser – 0 – 5Mc/s – £800. OPT 25 – £1000.
Tektronix AM503 Current probe + TM501 m/frame – £1000.
Tektronix 501 – SC502 – SC503 – SC504 oscilloscopes – £75-£350.
Tektronix 465 – 465B – 475 – 2213A – 2215 – 2225 – 2235 – 2245 – 2246 – £250-£1000.
Kikusui 100Mc/s Oscilloscope COS6100M – £350.
Farnell PSG520 Signal generator – £400.
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CIRCLE NO. 111 ON REPLY CARD

Pulse and bar generation in an instant

*The generator design described here produces a waveform at television line frequency having a sync pulse, a rectangular pulse, a sine-squared pulse, and a linear sawtooth. Applying the output of this simple circuit to a video link reveals major system deficiencies at a glance. By John Cronk**

The value of pulse and bar test waveforms are well established in television broadcast engineering. It is usual to transmit the pulse and bar and other hidden test signals continuously on certain lines during the field blanking period of the standard CCIR 625 line system I television waveform. A special oscilloscope with a line counting selector circuit is required to view this, and a special graticule is used to aid instant checks on the video quality.

The waveform generator described here will deliver the standard 1V peak to peak output across a 75Ω load that can be fed into amplifiers, equalisers, rf links, coax lines and so on. Because every line of this generator's waveform is similar, i.e. no half lines or frame pulses, the output can be viewed on an ordinary oscilloscope.

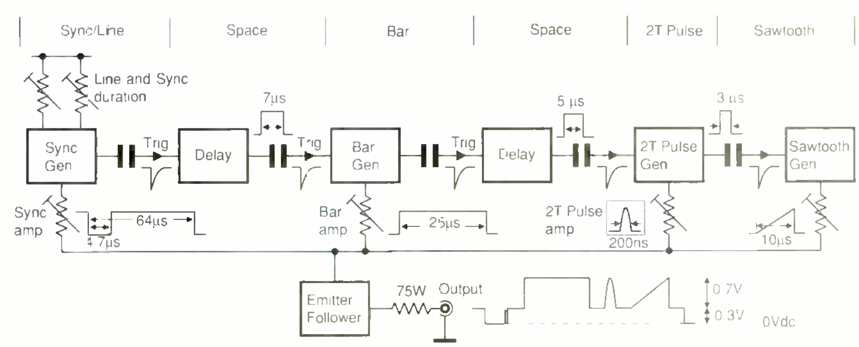
This simple circuit falls slightly short of regular pulse and bar generators, but the difference in cost will make it acceptable in many workshops. The output is dc coupled, but note – zero is sync bottom, rather than black level. As most equipment has an ac coupled input circuit after the termination resistor, this should not be troublesome. An output capacitor was not used as the additional time constant could interfere with the push button test. The 75Ω output is short circuit proof.

Linearity may be tested in a video system by accurate reproduction of the sawtooth. It should have the same amplitude as the bar and a straight line ramp to the point. Any amplitude limiting will be obvious.

The bar provides a peak white 700mV reference above the black level. Any tilt would indicate poor middle frequency response and corner rounding a poor HF response. Also reflections due to incorrectly terminated cables will be apparent.

It is well known that video circuits can exhibit faults with pulses that are not apparent

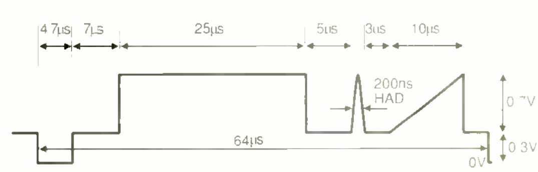
*John Cronk is GW3MFO



Block diagram of waveform generator. The complex waveform is built-up from individual components generated by a chain of 555 monostables. A couple of the devices provide a fixed delay. This scheme could have been implemented in many other, purely digital, schemes. The timer chain shown here works perfectly well… an example of simplest is best.

with sinewaves. It is for this reason that the sine-squared pulse proposed by Macdiarmid in 1951[1] is so useful. It is a carefully dimensioned pulse referred to as 2T that, for an ideal CCIR 625 line system I, should pass through an ideal 5.5MHz low pass filter without distortion. That is, without loss of amplitude, ringing or overshoot. The 2T sine-squared pulse represents the fastest transient that can occur in the video signal and which contains no redundant frequency components. Some distortion in a video system is permissible, typically, a 2% variation in the amplitude of the 2T pulse. This is specifically referred to a the *K* factor.

When first proposed, the sine-squared pulse was quite difficult to generate using valve technology; it required triggering a surprisingly powerful pulse which was then shaped in a multi-element filter. The circuit offered here is much simpler, but still produces an acceptable shape. This pulse was tested by using a double beam, 25MHz oscilloscope and

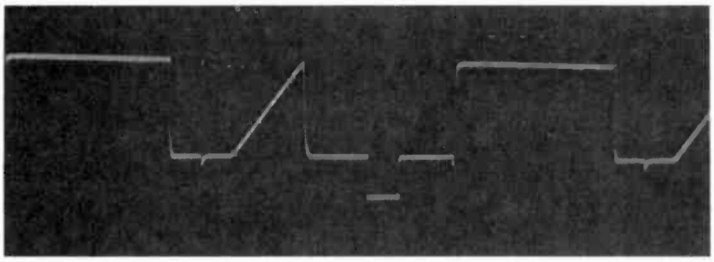


The complete waveform built up from individual components. Note that the narrow pulse section represents the fastest transient which should be able to pass through a standard PAL system without attenuation. Any reduction in amplitude relative to the long pulse indicates poor system HF response.

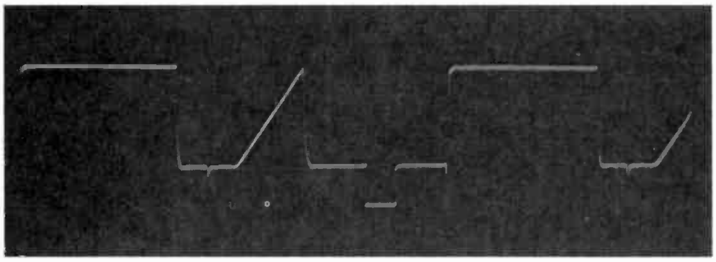
observing the pulse before and after a 75Ω CCIR recommended ten element low pass filter. There was no measurable difference before and after the filter.

When the whole waveform is displayed on an oscilloscope, the 2T pulse should have the same amplitude as the bar. Low amplitude indicates poor high frequency response, or excessive amplitude, overpeaking. In short, the waveform is ideal for rapidly setting up the gain and frequency response of video equipment.

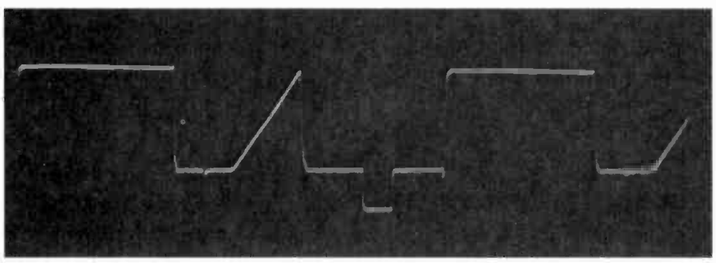
A push button inhibiting the bar from the waveform provides a useful test for dc restoration and line clamps. Suppressing the bar makes a significant change to the average



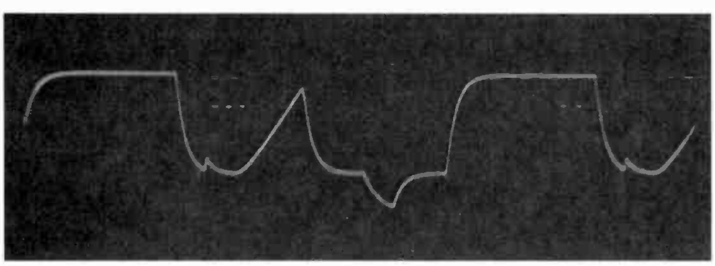
A: Comprising sync, rectangular and sine-squared pulses together with a linear sawtooth, this test waveform provides a wealth of information about video amp performance. It forms the source signal for the following tests.



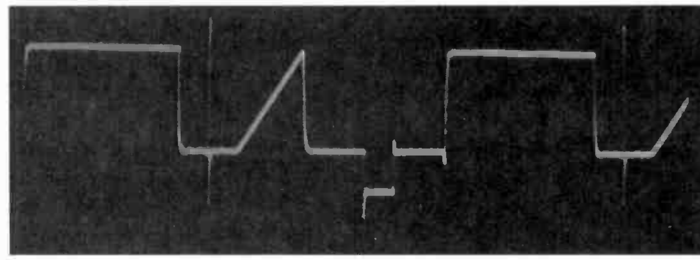
B: Output of a correctly terminated and adjusted amplifier, showing slight ringing on the left and right-hand plateaus.



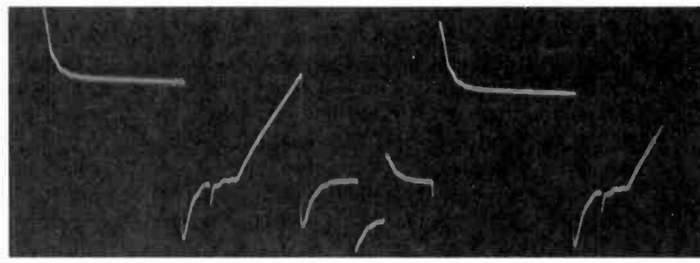
C: In a video amplifier with poor hf response, amplitude of the fast pulse is the first to suffer.



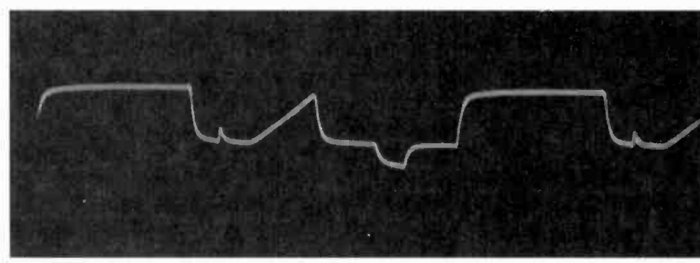
D: Video amplifiers with poor mid-frequency response cause the fast pulse to disappear and significant rounding on rising and falling edges.



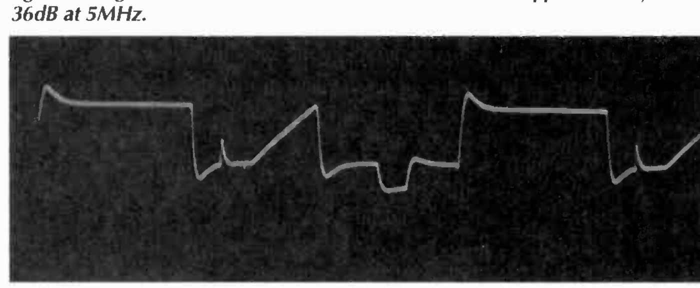
E: Output of a video amplifier with excessive hf gain shows some over and undershoot on the fast pulse.



F: An amplifier with excessive high and mid-frequency gain causes significant over and undershoot on all fast edges. Over and undershoot recovery is slow.



G: Distortion in the form of excessive hf loss caused by passing the video signal through 2km of RG58 coaxial cable. Attenuation is approximately -36dB at 5MHz.



H: partial correction with a simple video amplifier showing over-correction of mid-frequencies and under-correction of HF.

These photographs illustrate how useful the pulse and bar generator is for testing video amplifier performance. Test equipment used is a 25MHz oscilloscope with compensated ×10 probe.

video level. This can be demonstrated by observing the black level on the oscilloscope, first with ac coupling, then with dc coupling when the button is pressed.

Description of the circuit

Ubiquitous 555 timers produce the pulses in all stages. IC_1 is connected as an oscillator with adjustable mark and space. The 4.7µs off-period is the sync pulse and the on-period, the line duration, forms a DC platform which is the black level on which the bar and sawtooth are added.

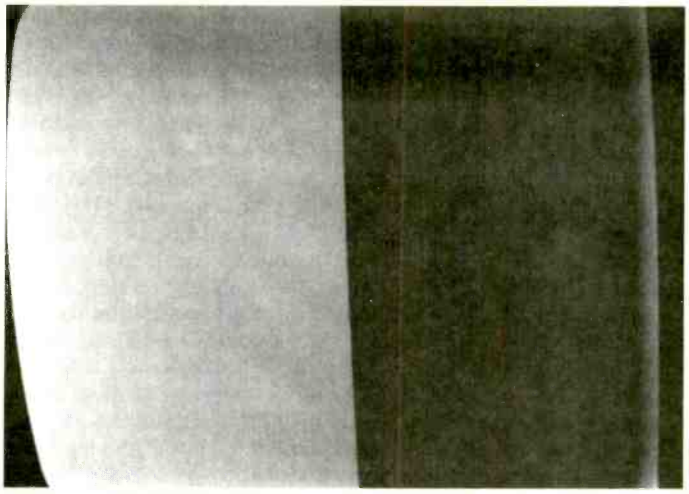
The next IC is a triggered monostable multivibrator used to delay the trigger pulse to IC_3 similarly configured which generates the 25µs pulse for the bar. The falling edge of this pulse triggers another short delay (IC_4) which then triggers IC_5, the 2T pulse generator. The falling edge of this pulse triggers IC_6, the timing capacitor of which is charged through a pnp transistor connected as a constant current regulator. This produces a linear ramp across the capacitor.

The wanted pulses are fed through amplitude adjusting trimmer potentiometers to the bases of the output emitter followers. The long pulse from IC_1 turns Tr_2 on, so it operates linearly during the bar and sawtooth period. A second emitter follower Tr_3 without bias, adds the sine-squared pulse to the output.

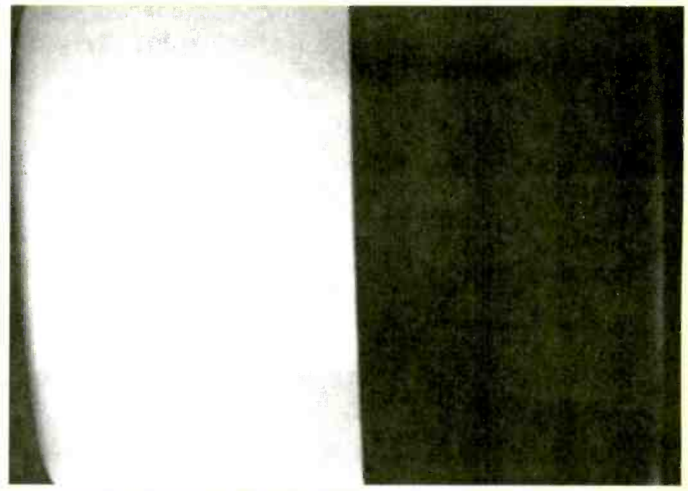
The fast rising edge of the 3µs pulse from IC_5 is applied to an L/C circuit which determines the shape and duration of the sine-squared pulse (200ns half amplitude duration, or HAD). A germanium diode absorbs the unwanted negative half of the sine-wave and the remaining low amplitude ringing is cleaned up by the unbiased emitter follower.

The 1kΩ resistor in the output circuit prevents the emitters of Tr_2 and Tr_3 running open circuit; normally the 75Ω source and external termination resistors provide the dc path.

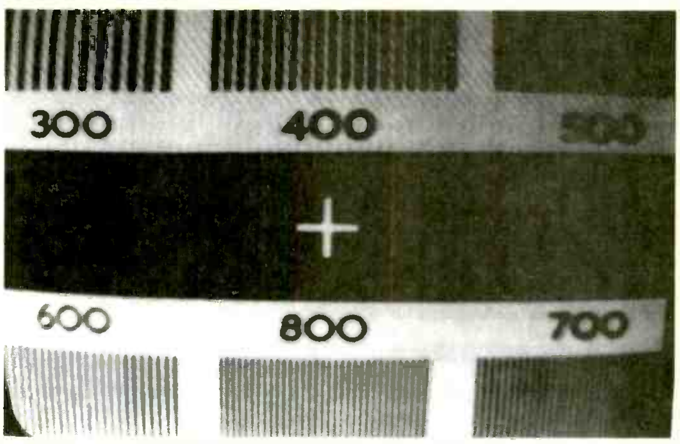
The pulses at the output of the 555s have risetime faster than required for the CCIR video specification. However the high value series resistors together with the stray capaci-



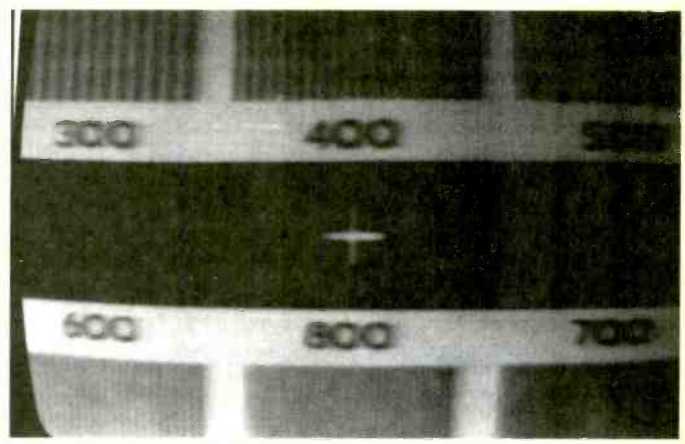
I: Results of the test signal displayed on a monitor via a short piece of coaxial cable. Effects of the fast pulse are evident as a vertical stripe just right of centre.



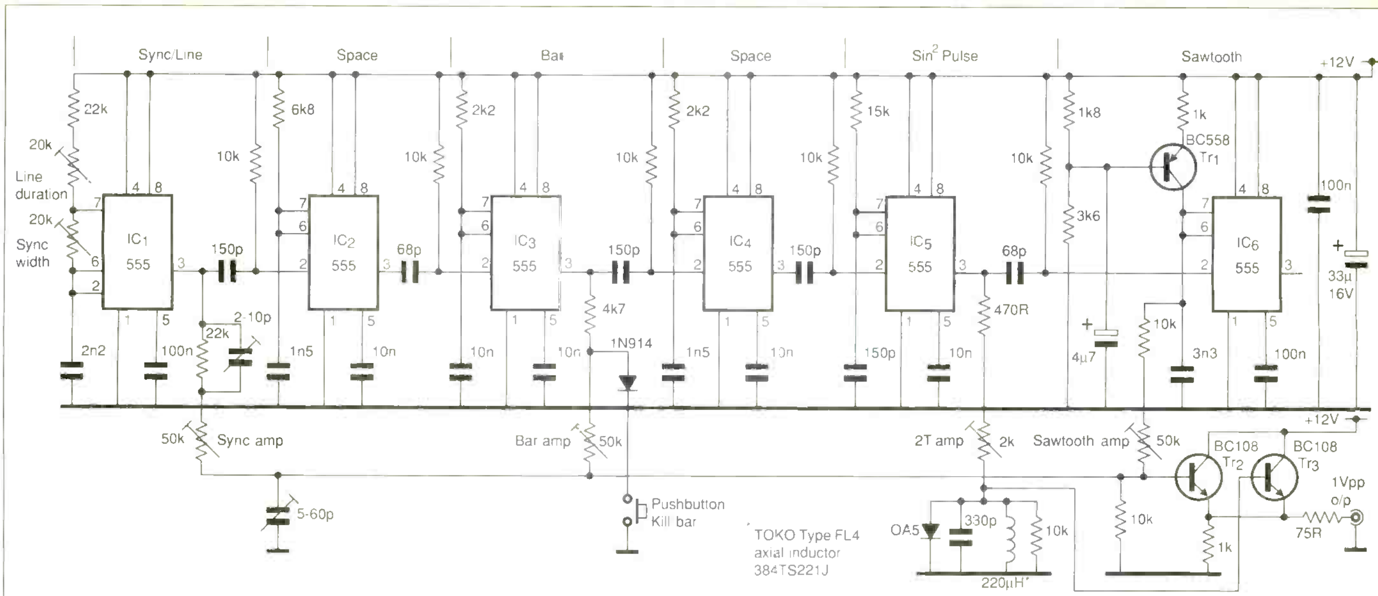
J: As you would expect, using 2km RG58 cable to feed the monitor significantly affects hf response. Definition is lost and the stripe has disappeared.



K: Monitor screen showing a display from a closed-circuit camera. The camera is connected via short cabling and hf response is adequate.



L: As in K, this monitor is displaying output from a camera but 2km of RG58 cable is degrading the signal. Use of the pulse and bar signal allows a correcting amplifier to be set up in minutes, without ambiguity.



Circuit diagram. Despite the critical role which the 555 timing capacitors play in the system operation, standard polystyrene components are acceptable. The parts which make up the pulse-forming network for the sine² pulse should be close tolerance however.

Broadcast television insertion signals

During the field blanking interval, lines 16-20 even frames, and 329 to 333 on odd frames, may contain identification, test and control signals. Lines 16 and 329 may contain international identification and control signals.

Lines 17, 18, 330 and 331 international test signals.

Lines 19, 20, 332 and 333 national test signals.

There are two usual national test signals. First line 19 and 332 comprise a 10μs white bar, a full amplitude (2T) negative pulse, (2T) positive pulse, a composite pulse (10Tc) containing chrominance and luminance information, and a five-riser staircase with colour sub-carrier at constant phase and amplitude.

The second test signal on lines 20 and 333 contains a half-amplitude luminance bar with sub-carrier superimposed, and an extended burst of sub-carrier for the second half of the line. The 2T pulse and 10μs bar enable the *K* rating to be obtained, while measurement of line-time non-linearity may be made by passing the staircase waveform through a filter. The 10Tc pulse permits assessment of chrominance-

tance of the circuit board integrate the pulses, slowing the transitions and causing slight rounding of the corners. Two small trimming capacitors are added to restore the pulse edges.

The power supply requirements are modest, a stabilised voltage between 9 and 12V at 100mA with a low impedance is sufficient

Construction notes

Care should be taken with routing the base connections to the output transistors which are at fairly high impedance and therefore prone to pick up unwanted pulses. Wide conductors should be used for the ground and power rails to minimise stray coupling due to common conductor impedance. The only high tolerance components required are the 330pF capacitor and 220μH inductor in the sine^2 pulse forming network; all other critical values are adjustable.

Miniature ceramic trimpots were chosen for stability. Multi-turn types, although not essential, would give a precision feel to setting up.

The suggested capacitors are polystyrene types, which are more stable than ceramic components. Incidentally, it was found that the decoupling capacitors on pin 5 of the ICs were not essential but they were fitted anyway.

The diode across the 2T pulse network must have a low V_f, if an alternative to the *OA5* germanium diode is to be used.

Commissioning

The duration of the sync pulse and line period should be set carefully, but there is more tolerance on the other timings determined by fixed components. ±2μs is allowable, and the excess can be accommodated by the 9μs spare at the end of the line period.

The important 200ns HAD of the sine-squared pulse is determined by the 220μH inductance and the 330pF capacitor and is not likely to require adjustment. The pulse amplitude trimpots should be set as accurately as possible. The aim is for 1V peak-to-peak with the standard 70:30 picture to sync ratio when the output is terminated with a 75Ω load.

Finally, the waveshape should be examined. The rise and fall times of the pulses at the output of the ICs are less than required. The amplitude control resistors and stray capacitance slow down these transits and the non-linearity of the output transistors should clean up the sync pulse bottoms. The two pulse shaping trimmers may be adjusted. The 5-60pF integrating trimmer will slow the transitions and round the corners of all the pulses except the 2T pulse, while the 2-10pF differentiating trimmer will sharpen the corners of the sync pulses. Adjust both trimmers for best waveshape with minimum overshoot. The specified rise and fall time for the line sync pulses is between 200 and 300ns. ■

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4. Wood A. *Pulse and Bar Testing*. CQ-TV Magazine No 142, P.42. (1988).

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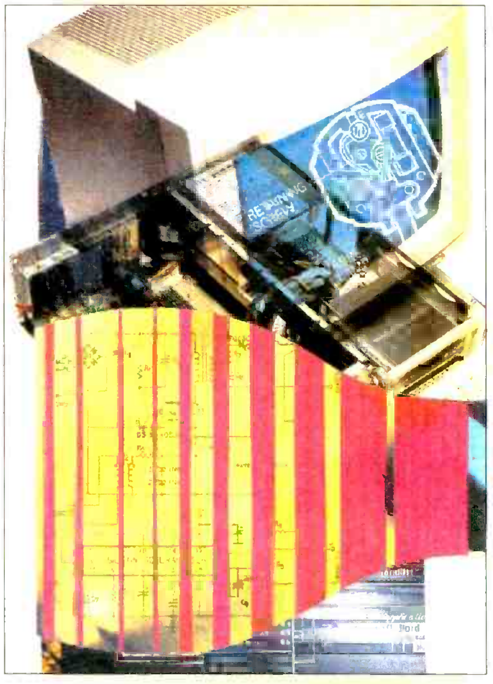

SMPS Design

2: Switching power converters can be designed to operate in discontinuous mode but most text books only discuss continuous designs. Duncan Smith discusses both here, showing that there are valid arguments for each.

Switched-mode supplies are usually designed to so that the inductor current never falls to zero, i.e. they operate in continuous mode. In discontinuous operation, current in the inductor falls to zero before the end of a switching cycle. The inductor is said to dry out, **Fig. 1**. This can occur by design, or when the load drops below a minimum level.

The essential difference between the two modes is that in continuous conduction the output voltage is always controlled by the feedback. In discontinuous mode, another state is introduced where there is no feedback directly controlling the output level. In both modes, output voltage is maintained by linearly varying the duty cycle.

Further, when either configuration has a low load current, the regulator can miss cycles to maintain the output level, causing interference. Pulse skipping occurs when the duty cycle can not fall any further to maintain the output voltage, **Fig. 2**.

Generally, because of switching losses, it is better to design a high output regulator to run continuously. Low output levels are best catered for with discontinuous operation. Complete converters on a chip, such as the *LT1073*, run continuously, but in bursts. This combines high efficiency and low output power capability.

Running in continuous mode often results in a large inductor value, producing a low input ripple current, but the inductor is physically large. In addition, the response to load tran-

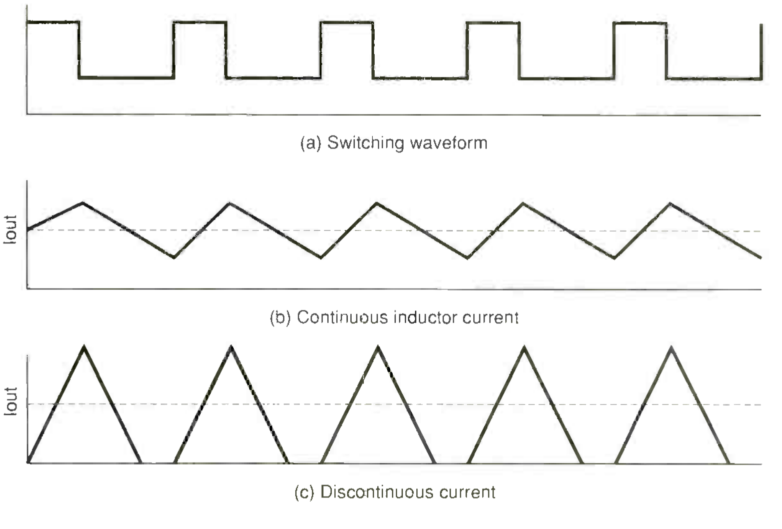

Fig. 1. Continuous and discontinuous mode switching waveforms compared. Discontinuous operation, (a), has a number of inherent disadvantages but even so, it can be the best choice for low-power designs.

Slope compensation

There is a conflict of interest when choosing the amount of slope compensation added to a current-mode controlled power supply. On the one hand, optimum current-mode control needs to be provided, on the other, stability problems in the inner current control loop need to be prevented.

In most current-mode controllers it is peak inductor current that is controlled, not the average output current. As a result, perfect current-mode control is not achieved. As the duty cycle alters, the difference between peak and average current varies. The average current increases with increasing duty cycle.

Slope compensation corrects for this by forcing average inductor current to be constant regardless of duty cycle. This occurs when the slope of the added ramp is equal to half the down slope of the inductor current. The inductor now acts as a voltage-controlled current source.

Severe regulation problems can be caused by an instability of the inner current control loop. This can arise with any current-mode circuit topography that allows for duty cycles more than 50%. Above this level, any action that changes the current through the inductor, such as a fall in input supply, result in the variation getting larger. As this is an open loop problem, neither the current nor voltage feedback paths can correct for it.

Addition of a decaying slope removes this instability by reducing the variation in the current each cycle, effectively damping the rising oscillation. To guarantee stability between 50% and 100% duty cycle, the amount of slope compensation needs to be greater than half the down slope of the inductor current.

Allied to the above is the problem of inductor current ringing. The circuit behaves as a series-resonant *RLC* circuit, peaking the gain at half the switching frequency. This may also lead to sub-harmonic oscillation.

Ringing is triggered by line and load transients, but can be damped by adding a ramp that falls at the same rate as the inductor current. Slope compensation at this level will reduce the current ringing within one cycle; this is similar to critically damping the *RLC* circuit.

Another problem is sub-harmonic oscillation of the voltage feedback loop. It is identifiable by significant differences in the duty cycle between successive cycles.

Two conditions cause the outer voltage feedback loop to oscillate at half the switching frequency. The first is the introduction of positive feedback due to gain peaking in the inner current feedback loop; the second is an excess of gain and phase shift in the error amplifier. The effect is as seen with ringing inductor current and open-loop instability.

Curing the problem is the same as before. At a compensation slope of m_c=$-0.5m_L$ the point of instability moves out to 100% duty cycle. Nevertheless, the fact that the amplifier gain is not zero at half the switching frequency will cause oscillations to occur long before 100% duty cycle is achieved. To guarantee stability the slope compensation must be increased to m_c=$-m_L$ so that loop gain becomes independent of the duty cycle.

The actual amount of slope compensation introduced is a compromise between perfect current-mode control and immunity from noise. When the ratio of ripple current to average current through the inductor is low, small quantities of noise can cause large pulse width variations. Effects of this noise can be reduced by increasing the amount of slope compensation.

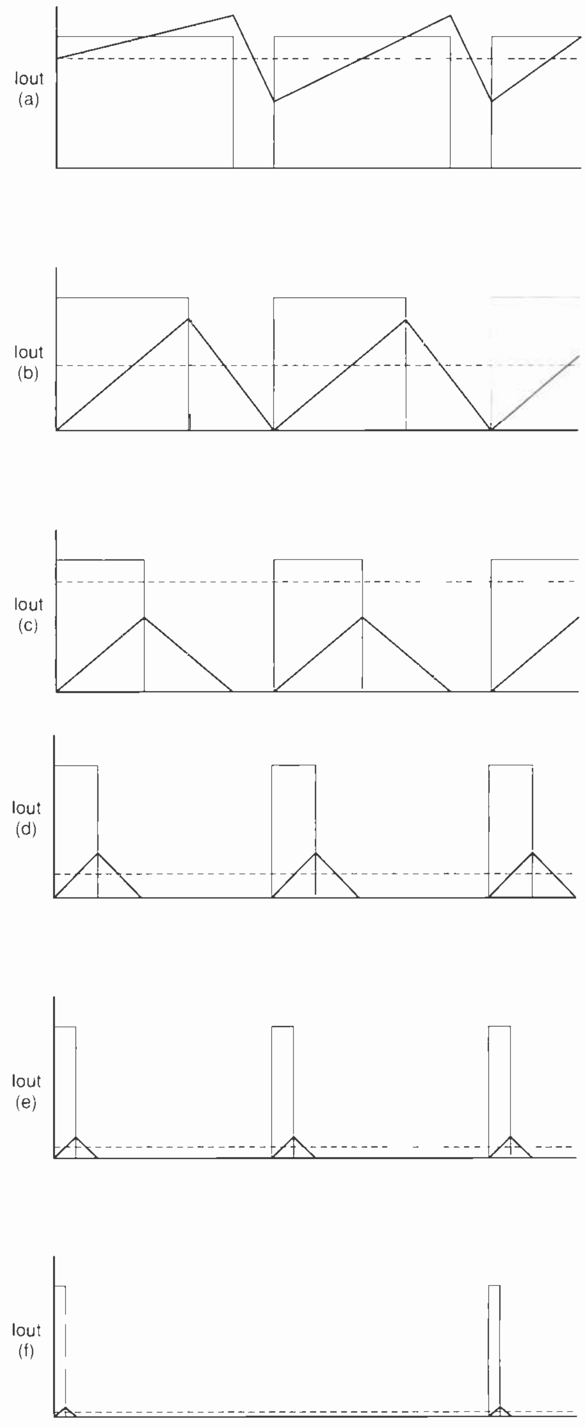

Fig. 2. Onset of pulse skipping. This happens when loading becomes so light that duty cycle cannot be made small enough to cope.

sients is impaired because of the lower slew rate of the inductor. If a low value inductor is chosen, as in the discontinuous case, currents are higher and maximum output power is limited. The converter enters discontinuous mode when the current falls below the minimum load. This may not always be desirable because of increased switching losses and ripple currents.

In continuous mode, maintaining stable operation over a wide range of loads and input voltage variations in continuous mode can sometimes prove to be difficult. This is due to the complexity of its transfer function, i.e. multiple poles.

Discontinuous mode converters involve high switching currents, due to their small inductor values. Usually, this results in lower efficiency as the switching losses are higher. Output power is frequently limited to keep switching currents down, hence the inductor can not transfer enough energy to meet high load requirements. Operation in discontinuous mode is ideal when low power levels are required, as then both output and switching currents have low values.

However, the SGS-Thomson *L4963* is a complete switched-mode regulator designed to operate in discontinuous mode. It can deliver up to 1.5A at 5V. Boost converters are some-

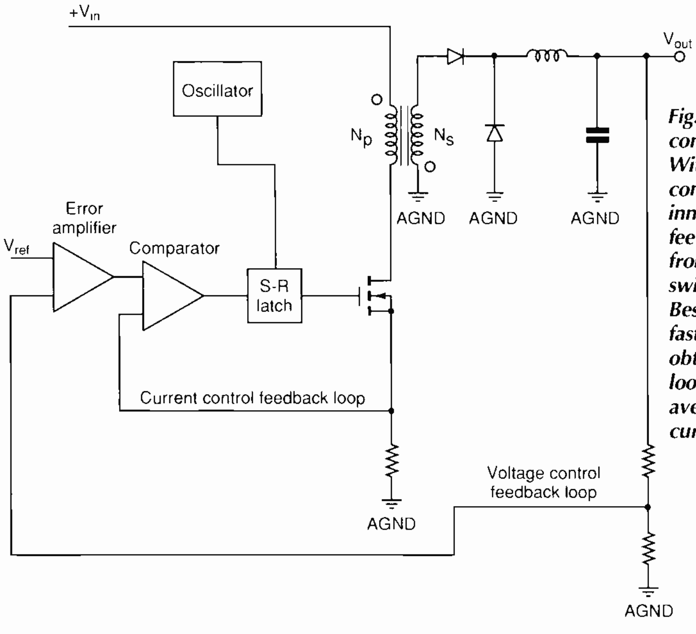

Fig. 3. Current-mode converter example. With this configuration, an inner loop is added, feeding back current from the main switching device. Best regulation and fastest response is obtained when the loop controls average inductor current.

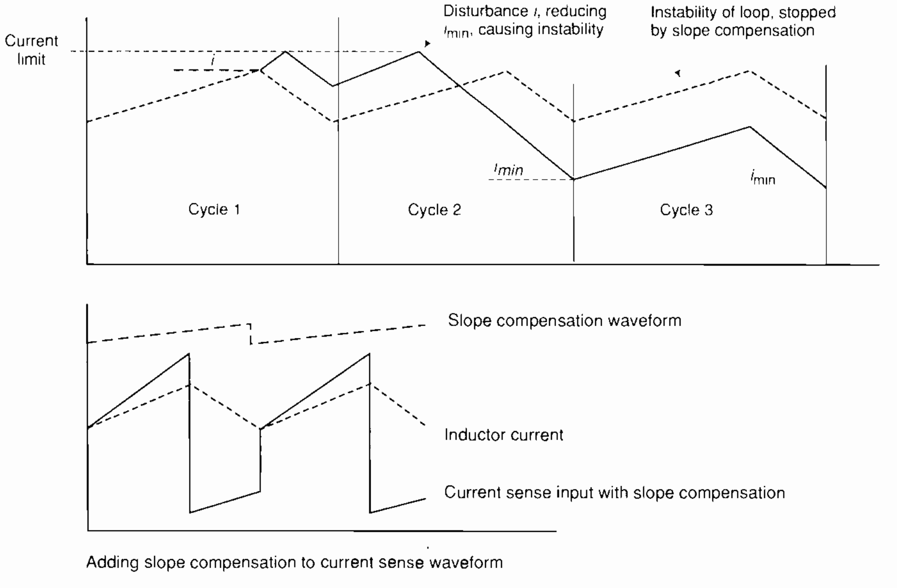

Fig. 4. Operating current-mode controllers with continuous inductor current causes a number of problems. Most of these can be removed by slope compensation, the effect of which is evident here.

Diode switching

During the time the diode takes to turn off, the device is conducting even though reverse biased. This leads to unwanted dissipation in the diode. The faster the turn-off time, the more dissipation will be shifted to the switching transistor.

A slow turn-on time can have more serious consequences than slow turn-off. A large amount of power can be dissipated in the diode if it switches on slowly. As the device turns on, not only will current be flowing, but also the forward voltage drop across the device will be high, at volts rather than millivolts. This can also lead to a high voltage turn-on spike in boost/flyback circuits, which appears across the switching transistor. It is thus in series with the output voltage.

Under such conditions, breakdown voltage of the transistor can be exceeded as it turns off, and the diode turns on. The size of the voltage spike is related to the final current value and the rate of current rise – i.e. the inductor value.

times designed to operate discontinuously for stability reasons. A pole is removed from the transfer function so there is less phase lag introduced, simplifying compensation and improving transient response.

Discontinuous mode is often ignored in text-books, while data sheets for switch-mode power ICs contain design examples. If the operation of the circuit in discontinuous mode is catered for in the design, and in the component ratings, then there is no reason not to run in this mode. In fact discontinuous mode can be an advantage if space is a problem, due to the smaller inductor.

Current-mode SMPS

Current-mode PWM regulators have an advantage over traditional SMPS designs. They give an immediate response to input line variations, can provide pulse-by-pulse current limiting, and are simpler to stabilise. Operation in both continuous and discontinuous modes is possible, though the former is more common.

As discussed earlier, the common pulse-width modulated converter compares an error signal with a ramping reference to control output voltage. In a current-mode PWM converter an additional inner feedback loop is added, **Fig. 3**. Current through the switching device is sensed. This sample is used as the reference signal by the PWM comparator, and is compared with error amplifier output.

Control voltage from the error amplifier now directly determines the peak inductor current. To obtain the best regulation and fastest response, the loop should control average current through the inductor. In practice, control of the peak current is simpler. It is analogous to controlling the average inductor current as it is directly proportional to it.

In continuous mode the inductor can be looked at as a voltage controlled current source, eliminating its pole in the response and reducing the transfer function control loop to a single pole. Now, the regulator is simpler to stabilise since the dominant pole consists of the output capacitor in parallel with the load. This can improve dynamic regulation of the output voltage by opening up the regulator bandwidth. This speeds up response to load variations. In addition, the method instantly corrects for variations in input voltage.

There are problems with fixed-frequency current-mode controllers operating with continuous inductor current. These problems are open-loop instability with duty cycles above 50%, **Fig. 4**, sub-harmonic oscillation, ringing inductor current, and sensitivity to noise.

Most of these problems can be overcome by adding slope compensation. This is achieved by adding a ramping voltage, derived from the switch control oscillator, to the current waveform. More commonly the ramping voltage is subtracted from the error amplifier signal. Optimum current mode control is achieved with a voltage that slopes at a rate equal to half the down slope of the inductor current. In discontinuous operation, no sub-harmonic oscillations can occur.

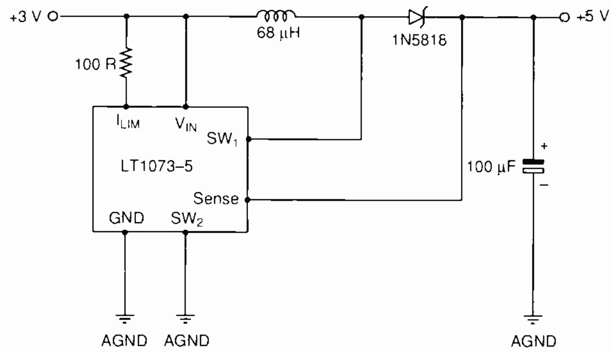

Fig. 5. Highly integrated ICs allow a switch-mode converter to be built using very few components, as this 3 to 5V step-up supply shows. Output current is 100mA maximum.

Soft-start circuits

On powering up into a load, both linear and switched-mode regulators take a large current surge from the input supply to charge capacitors on the line. This can cause the input supply to drop, or worse, collapse completely. In turn, this can lead to false microprocessor resets, failure to power-up – especially with partially full batteries – or even the system pulsing on and off. With switched-mode supplies this problem is made worse as large peak currents are taken to store energy in the inductor.

The purpose of the soft-start circuit is to slowly ramp up the converter output supply, producing a gradual increase in current demand. This results in an orderly system power up. A typical method is to clamp the error amplifier output to a slowly ramping voltage, from an RC circuit for example, until the amplifier output has reached a nominal value. The soft-start circuit then turns off, allowing the control loop to operate normally.

Component selection

Component selection is a very important part of the design of a switched-mode supply. When prototyping, it is advisable to use sturdier devices than envisaged in the final system. Down-grading of the parts to reduce cost can be undertaken once peak voltages and currents have been measured using the final PCB layout. In this way, stray and parasitic inductances can be accounted for.

A design will often work with devices found lying around in the laboratory but to obtain the best performance, using the correct components is important. That said, it is rare that efficiency is the overriding criterion. Often the ratio of cost to performance is more important. Similarly, rather than obtaining 98% efficiency, generating the required voltages from an input supply is usually more important – as long as your supply does not turn out to be unreliable due to component stressing. So what parameters should be considered?

Capacitor. The main output capacitor should be chosen to have a low equivalent series resistance, or ESR, and series inductance. This limits the voltage step at the output when delivering the stored power to the load and capacitor. Generally, the larger the physical size of the capacitor the lower the ESR. A good rule of thumb is that at least ten times the calculated capacitance is required to obtain an acceptable ESR. A good capacitor will not come cheap, but a large output voltage transient spike could do more costly damage. As an alternative an inductor and capacitor output filter can be used. This reduces the size of the main output capacitor.

Switching elements. The switching elements, i.e. the diodes and transistors, must be able to handle the highest voltages in the converter. This makes high breakdown voltages necessary. It is easy to get large voltage spikes that can destroy a device. The higher the breakdown voltage the less the likelihood of the device being damaged by a switching spike. The penalty is cost.

Any switching devices used must have fast rise and fall times, otherwise power will be lost during switching. Low on resistances for mosfets, or saturation voltages for diodes and bipolar transistors, are necessary to reduce the power losses. Mosfet I_D^2 is affected by $R_{DS(on)}$, diode I_F by V_F and bipolar transistor I_C by $V_{CE(sat)}$.

Diode. Important diode parameters are switching time, forward voltage, breakdown voltage, and leakage current. Forward voltage of the diode determines the loss in the device when it is fully conducting; the higher the current flow the higher the voltage drop. Further, using a diode with a greater current handling capability than necessary results in a lower forward voltage.

Diode reverse breakdown voltage is important in inverting or voltage boosting power supplies. The diode must handle the maximum output to input differential voltage, plus any voltage switching spikes. Diode reverse breakdown has a detrimental effect on regulation. Selecting a diode with a reverse voltage rating of at least twice the differential voltage is a good starting point.

General purpose rectifier diodes are not suitable for switching even moderate currents quickly. Fast, small signal diodes can be used with caution in low current designs. Schottky diodes have a low forward drop of typically 0.3V, and switch on and off quickly. However, they suffer from lower reverse breakdown voltages and higher leakage currents than conventional diodes. Leakage currents lower the efficiency of micro-power designs.

Mosfet driving

To drive a mosfet gate successfully, the total gate charge must be known. This is normally on the data sheet, quoted in nano-coulombs, nC. Take for example the *IRF730*. This has a maximum total gate charge, Q_g, of 35nC. Maximum drive current required at a frequency f is $i_{max}=Q_g.f$. For the *IRF730*, maximum drive current is ±3.5mA at 100kHz.

Total gate charge is derived from the equivalent input capacitance, which is non-linear. This varies with both V_{GS} and V_{DS}, and is made up of contributions from the gate-source capacitance and the non-linear Miller capacitance across the drain and gate, C_{rss}. On most data sheets, C_{rss} is called the reverse transfer capacitance.

Equivalent input capacitance is,

$$C_i=C_{iss}+(1+A_v)C_{rss}$$

where A_v is voltage gain. Effects of C_{rss} can therefore not be ignored. If only C_{iss} is considered then the drive current requirements will be grossly underestimated, derived from $Q_{iss}=C_{iss}V_{GS}$.

The closer C_{rss} is in value to C_{iss} then the higher the drive current requirement will be. If no value, or graph, is given for Q_g, or for switching times, then the device is possibly not intended for use in switching circuits. In this case it is better to select another type.

Emitter follower or common emitter stages are capable of driving the gate at low frequencies, say 10kHz or less. By 100kHz however, they are often unable to discharge gate capacitance, although they may still be able to charge it. This leads to unwanted dissipation and possible device damage due to the slow turn-off.

Push-pull gate drive stages become essential as the operating speed of the switcher increases. Obviously, depending upon the mosfet chosen and the operating speed, *HC* logic devices with totem-pole output stages can be used. A *74HC08* gate can source or sink 4mA, while *AC* series logic is capable of driving to ±20mA.

In boost/flyback converters, another capacitive problem can occur due to the capacitive coupling between the drain and gate, i.e. C_{rss}. As the inductor voltage rises at turn-off, gate voltage also increases due to capacitance between drain and gate. This can rise above the gate breakdown voltage, destroying the device, so a zener diode gate clamp is essential.

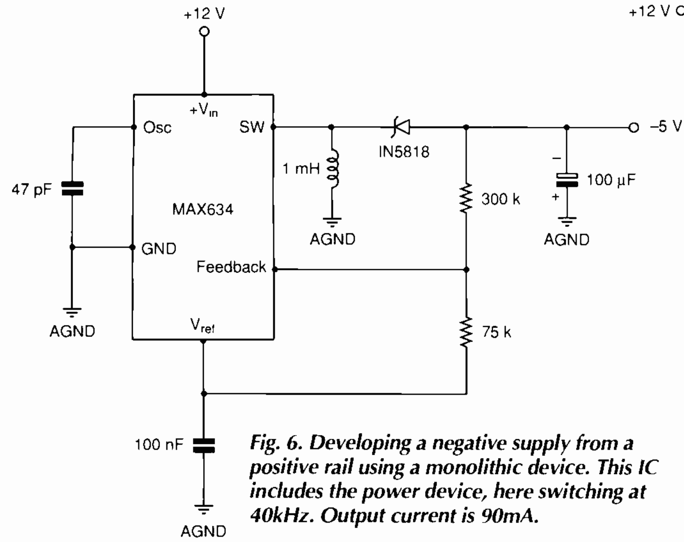

Fig. 6. Developing a negative supply from a positive rail using a monolithic device. This IC includes the power device, here switching at 40kHz. Output current is 90mA.

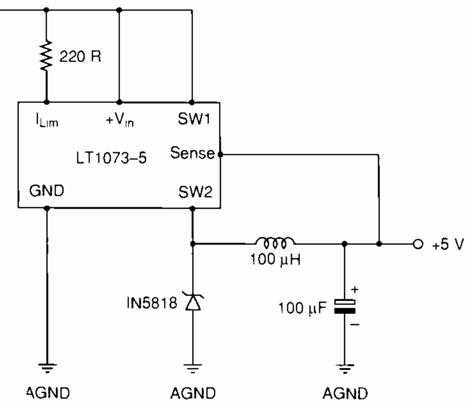

Fig. 7. Five-component 12V to 5V DC converter. Switching frequency is 19kHz while output current is 350mA.

Bipolar devices. Unless switching frequencies of a few tens of kilohertz and low currents are required it is not worth considering using bipolar devices for the main power switch. Mosfets do a better job. To achieve a reasonable saturation voltage, $V_{CE(sat)}$, base currents of at least a tenth of the maximum collector current need to be used. Efficiency of the design can be reduced dramatically. Power loss during the transistor switch on and off time occurs as a result of charge storage in the base region.

As an illustration, the first SMPS I designed while a junior engineer resulted in a once proud transistor being reduced to a charred lump of plastic. I didn't attempt to design another for five years! This was alleviated on the next incarnation by using a Baker clamp around the device to stop junctions being driven into saturation.

Mosfet. A large number of problems found with bi-polar transistors are overcome by using a mosfet. Its two main advantages are switching speed and ease of driving. Mosfets have drawbacks though. The gate-source junction is more easily damaged by high voltages than its bipolar counterpart and zener protection diode is essential.

Secondly, the quoted drain-source resistance is given for a specified gate-source voltage, which is often higher than you might think. A logic level mosfet may turn-on at 1.8V. To achieve the desired $R_{DS(on)}$ however, the gate must be driven to 4V. This is quite a contrast to the 0.6V or so needed for a bipolar device.

Another much quoted advantage of mosfets is that they are voltage driven and therefore require next to no gate current. This is not the complete story. Used as simple switches this is the case, i.e. dc drive. But at the speeds required of a SMPS this is not so, and ac drive current is required. A mosfet input is essentially capacitive, and this capacitor must be charged up. Equally importantly it must be discharged to make the device switch within the desired time (discussed further in one of the panels). The mosfet with the lowest C_{iss} and C_{rss}, and the highest forward transconductance, g_{fs}, should be chosen.

An integral diode across the drain and source is inherent in any mosfet structure. Normally it is not turned on and causes no problems. Its breakdown voltage and current capability are in general on par with those of the mosfet itself. However should the diode be turned on, its reverse recovery time is often an order of magnitude larger than the switching times of the mosfet. For example, the *IRF730* mosfet will switch in around 35ns, but the integral diode's reverse recovery time, t_{rr}, is 600ns, during which time it will be dissipating power.

Inductor. The inductor is the element of a switched mode supply that causes the most difficulties in both calculation and construction. Looking at the various IC manufacturer data sheets, the most common method used to define the inductor is to base its value on the desired current change – i.e. ripple current – through the inductor while the switching element is on. For continuous mode, this is usually set to around a fifth of the maximum allowable switch current. For a particular peak switching current, the regulator will give close to maximum output power. However, the main requirement of the inductor is that it must be able to store the required flux without saturating at the current level produced while the switch is on.

The following guide-lines are also useful when choosing an inductor. Inductance should be low enough to store the required energy at worst case conditions, i.e. minimum input voltage and switch on time. It should also be high enough so that the maximum current ratings of the switch and the inductor are not exceeded at worst case, during maximum input voltage, again at switch on time. In addition, the inductor must have a low resistance to limit power losses, which is achieved using thick wire.

Where electro-magnetic interference, EMI, is important, a toroidal or pot core should be used. With these core forms, most of the flux is contained in the core and does not leak out. There are many inductor manufacturers and it is generally easier and cheaper to buy a ready made device. Most manufacturers quote maximum allowed current, resistance and frequency.

Parasitics

Parasitic inductance and capacitance do not affect the input-to-output voltage transfer function of a converter, but lead to stressing of the switching elements. In general it is parasitic inductance rather than capacitance that causes problems, especially in flyback converters.

Stray and parasitic capacitances tend to increase turn off and on times, leading to power dissipation in the switching elements. However, junction capacitances of semiconductors can cause problems, as with C_{rss} in a mosfet. Compared with standard devices, Schottky diodes have a high capacitance.

A 2mm diameter wire in free space has an inductance of around 1.3µH per metre, or 0.4µH per foot. Looking at parasitic inductance, what sort of voltage level can be generated? If maximum current flowing is 5A, and the mosfet can switch off in say 50ns, then the rate of change of current, di/dt, is 100×10^6A/s. Any stray inductance around will convert this rate of change of current to a voltage spike. The voltage can be derived from the formulae for inductance,

$$V=L.\text{d}i/\text{d}t.$$

As a result, 13nH per centimetre of inductance (or 33nH per inch) times 100×10^6A/s produces a spike of 1.3V/cm (3.3V/in). This voltage adds to the output level and can cause breakdown of the switching element. In addition, stray inductance can render a fast switching diode useless as it will slow the turn-off speed of the device.

Similar problems occur in transformer-coupled designs where leakage inductances can create large voltage spikes, causing voltage breakdown in devices. Energy associated with these leakage inductances also has to be dissipated by the switching elements. These unwanted inductances can be minimised by maintaining tight magnetic coupling between the windings and the judicious use of *RC* snubbing networks.

Layout

Layout of a switched-mode power supply is very important, and is even more so at high frequencies and powers. The first point is to keep all tracks as short as possible. This is especially important around the high current switching circuitry.

Input supply should be bypassed with good quality electrolytic and ceramic capacitors placed as close as possible to the switching mosfet and inductor. Output decoupling/filtering capacitors should again be as close as possible to the switching circuitry.

While laying out the circuit it is worth while considering possible parasitic/stray inductances, and modifying it to try to minimise them. Likewise watch out for capacitive pick up due to adjacent tracks being too close together. This can be a problem when unwanted signals get back into the feedback loop. Heavy-current tracks should be kept as wide as possible to lower resistance, and it is worthwhile considering the use of board with a thicker copper layer.

Ground return paths can be a problem: there is little point mounting the decoupling capacitor next to the mosfet and having its ground connection some distance away. Further, it may be wise to separate signal and high current paths, and have a single point connection linking them. A complete ground plane is ideal. If a ground plane is not an option, the ground copper area should be maximised wherever possible. Ground returns should be linked at a single star connected grounding point.

Designing circuits

There is currently a variety of complete monolithic switched-mode converters available, making it relatively easy to design a switching power supply. Three simple examples are shown in **Figs 5** to 7. These illustrate how few external components are required.

In the first design, Fig. 11, a 3 to 5V step-up converter based on the Linear Technology *LT1073* is shown. The second, Fig. 12, is an inverting circuit based on the Maxim *MAX634*, while the third, Fig. 13, is a step-down converter. Other devices are available from companies such as SGS-Thomson, Texas Instruments and National Semiconductor – to name but a few. Cheap inductors suitable for low power circuits are available from Toko, and Coiltronics. ■

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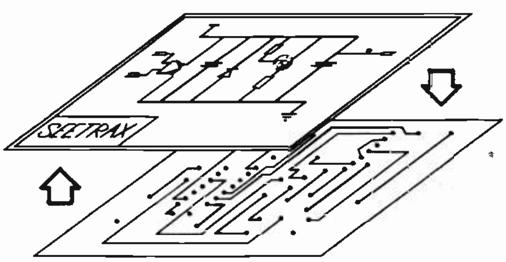

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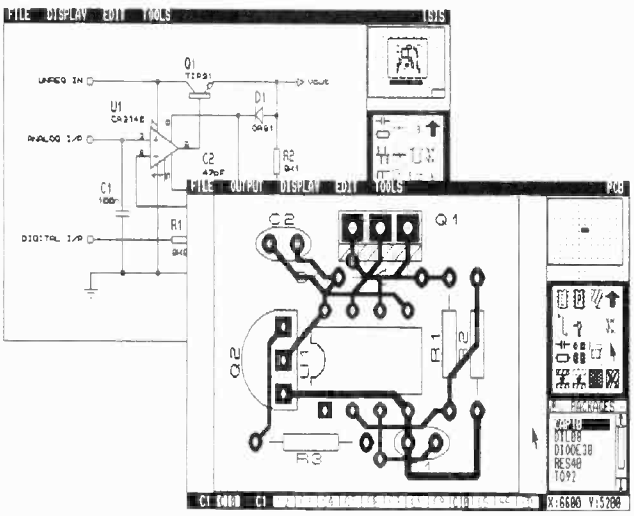

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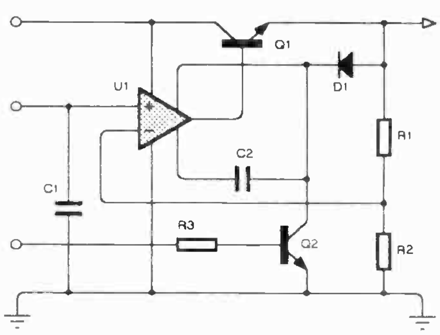

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CIRCLE NO. 115 ON REPLY CARD

Motor control via PC

Deriving its basic control signals from a PC with i/o card, this precise driver for a 180W dc motor involves a number of power control concepts including PWM, speed feedback and protection mechanisms. Being modular, it makes a useful teaching aid, as lecturer Peter Hale explains.*

Speed control of motors is used extensively in industry. Cranes, ship winches and conveyer belts for example all use motors whose speed is finely controlled.

Both ac and dc motors can be used in speed control but each has its own advantages. AC motors are inexpensive but their controllers – inverters – tend to be complex and expensive. DC motors on the other hand are relatively expensive but their controllers are inexpensive and simpler.

The main advantage of dc motors is that their speed is easy to control. In addition, speed changes are smoother and they offer better torque characteristics. Where only a portable dc supply is available and accurate speed control is needed, dc motors are the best alternative. An example is fork lift trucks. Although they operate on low voltage, the chopper control principle is still used.

Controlled power rectifiers make it possible to link a dc machine to an ac mains supply regardless of whether the supply comes from the national grid or on-site generators. To this end speed control packages are readily available. This article discusses one such package using pulse width modulation, PWM techniques.

I designed this system primarily as a teaching device for engineering undergraduates, to enable them to learn basic techniques of electronic design. Since it is a teaching aid, the ICs used in the control electronics are not dedicated motor-control solutions.

The entire system is closed loop. A microcomputer controls motor speed and receives back information about angular velocity and the temperature of the motor field windings. There are four distinct parts to the system. Two modules carry the power electronics signal processing sections while the remaining two comprise transducers and the controlling C program, **Figs 1, 2**.

Power electronics module

Elements of the the power module are presented in **Fig. 3**. A constant 220V is

Characteristics of dc motors

This curve and equations show that torque is directly proportional to armature current.

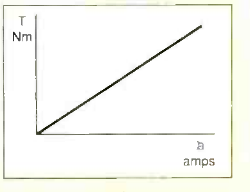
$$T \propto I_F I_A$$

or,

$$T \propto \Phi I_A \text{ viz. } T = K\Phi I_A$$

where I_F is field current and Φ is flux per pole in weber. $K\Phi$ is constant for constant current.

Torque as a function of armature current in a dc motor.



Speed of a dc motor falls predictably as torque increases according to these relationships.

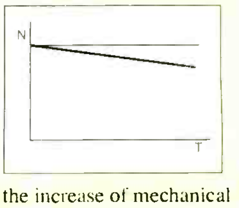
$$N = \frac{60}{2\pi}\left[\frac{V_A}{K\Phi} - \frac{R_A T}{K\Phi^2}\right]$$

where N is speed rev/min, $V_A/K\Phi$ is no-load speed and R_A armature resistance. This equation can be written as,

$$N = \frac{60}{2\pi}\left[\frac{V_A - I_A R_A}{K\Phi}\right]$$

$K\Phi$ is constant while $I_A R_A$ is small, typically <5% of V_A. Hence motor speed is approximately proportional to applied armature voltage V_A.

Speed falls approximately linearly with torque.

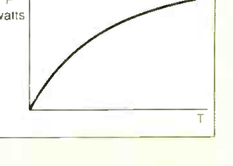


This equation shows the increase of mechanical output power with rising torque.

$$P = \left[\frac{V_A T}{K\Phi} - \frac{R_A T^2}{(K\Phi)^2}\right]$$

It assumes a separately excited field, i.e. a field with a voltage applied that remains constant regardless of armature voltage. A constant field voltage produces a constant field current and hence a constant $K\Phi$.

Mechanical output power against torque.



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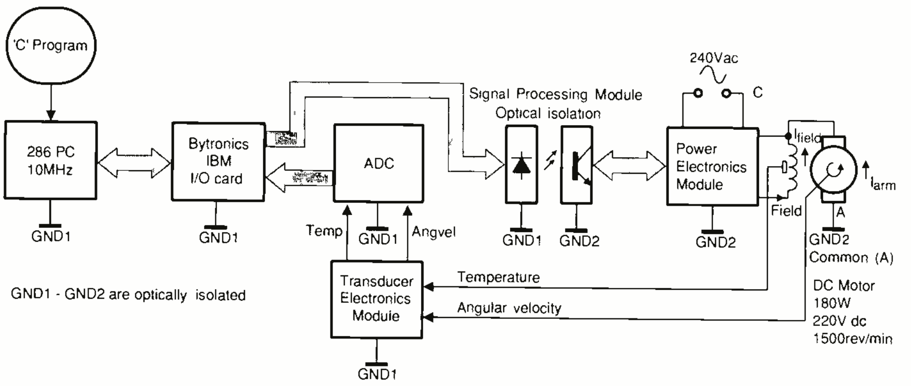


Fig. 1. Key elements of this modular controller for a 180W dc motor are a mosfet power interface, pulse-width modulator driver and transducer feedback. Being PC based, user interfacing is easy to develop and low cost.

applied to the field and the power mosfet supplies a chopped dc voltage across the armature. The maximum applied chopped armature voltage should be about 180V dc.

The relay's purpose is to remove the field voltage following a shut down. If the computer detects excessive field temperature or speed then a function named SHUTDOWN in the controlling C program is called. This routine removes armature voltage by switching off the mosfet. It then de-energises the relay to remove field voltage. It is important to note here that, for separately excited fields, the armature voltage should *never* be in situ when field voltage is absent.

The push switch provides a kick, necessary to energise the relay. In this way, power is provided to the signal processing electronics via the transformer and the four voltage regulators.

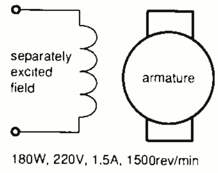


Fig. 2. Details of the 180W dc motor used. A constant supply of 220V dc is applied to the field while a variable voltage feeds the armature to control speed.

Power ground GND_2 is optically isolated from the computer ground GND_1 and must remain so. **Figure 4** shows the field voltage and armature voltage when the motor was running at 965rev/min.

Signal-processing module

There are many different ways to carry out the signal processing. I have chosen frequency-to-voltage conversion since is one of the techniques that is of interest to students.

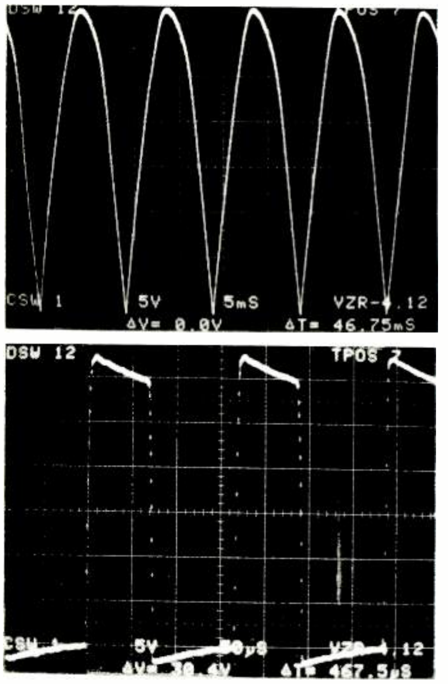


Fig. 4. Field and armature voltage for the 180W dc motor using the controller in Fig. 3. Speed, altered by varying the armature pulse width, is 965rev/min.

Figure 5 shows the signal-processing module. Control input, on the left, feeds a Bytronics card which is an i/o interface mounted inside a PC. Its timer-zero output produces 50% duty cycle pulses at a typical frequency of 100Hz. This pulse train is derived via PC software and transmitted to the power electronics via the opto-isolator.

Signal for input to pin 1 of the frequency-to-

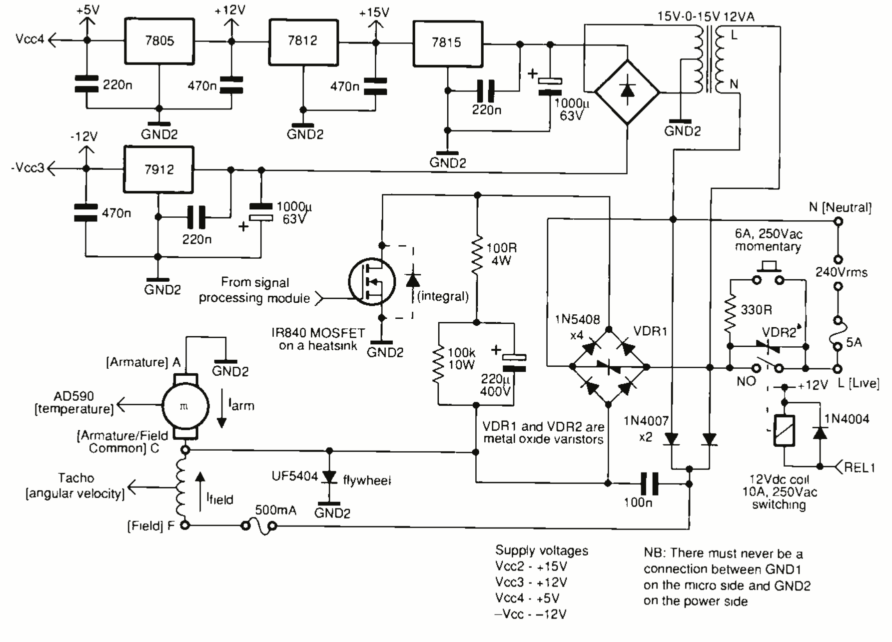


Fig. 3. Power control for a dc motor is greatly simplified by the use of a high-voltage mosfet. The relay is added to remove field voltage when the shut-down command is received.

Learning goals

Students taking on the task of reproducing this system have the potential to learn some important techniques in electronic design. Among them are,

- programming in C
- analog to digital conversion
- optical isolation concepts
- converting frequency to voltage
- voltage conversion (dc to dc)
- comparators
- oscillators
- amplifiers and buffers
- driving a power mosfet
- power supplies
- transducers
- dc machines

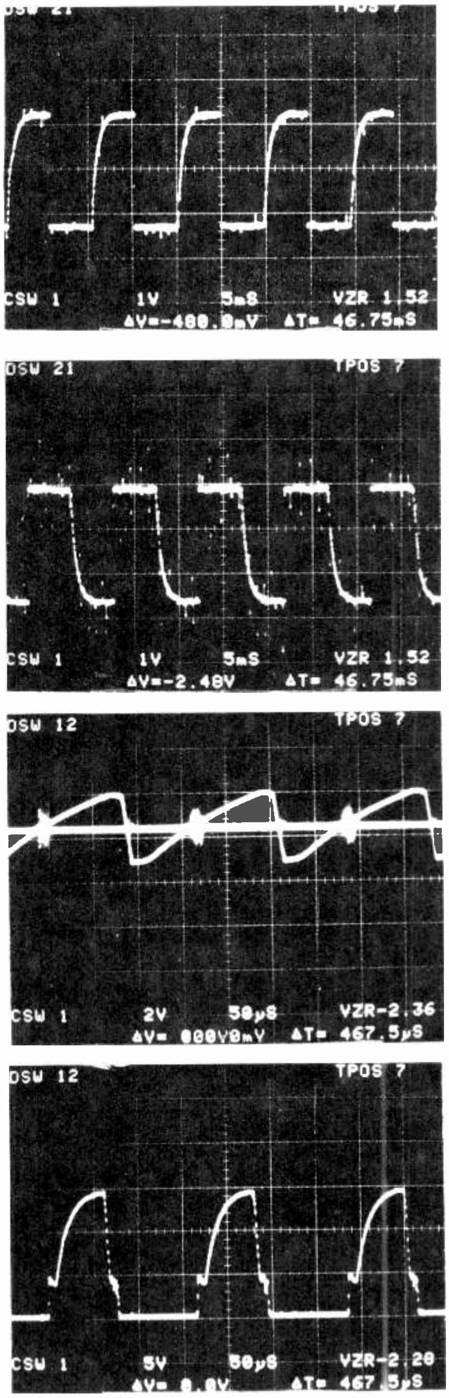


Fig. 6. Waveforms found in the signal processing section, Fig. 5. Pulses in the top two traces, from circuit points WF_1 and WF_2 respectively, are derived from the PC and changed to a dc level by a frequency-to-voltage converter. In turn, this level is used to control the width of pulses fed to power mosfet. Third photo down shows inputs to comparator IC_1, at around 6.4V dc, and bottom is final PWM output.

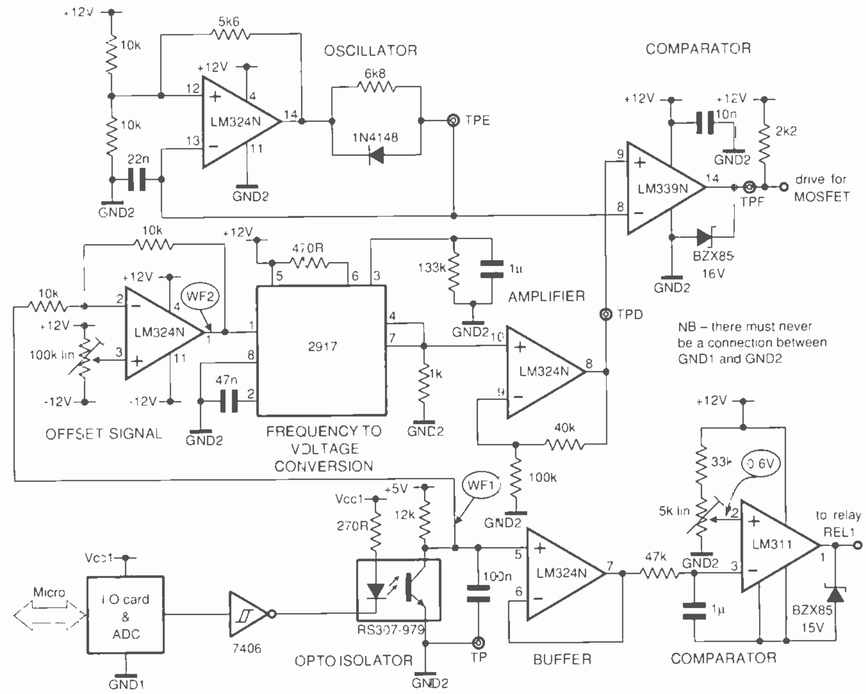


Fig. 5. Signal processing circuitry is optically isolated from the i/o card which is a PC plug-in. Under PC control, the upper part of the circuit develops PWM to drive the fet in the power section. Both lower op-amps control the shut-down relay.

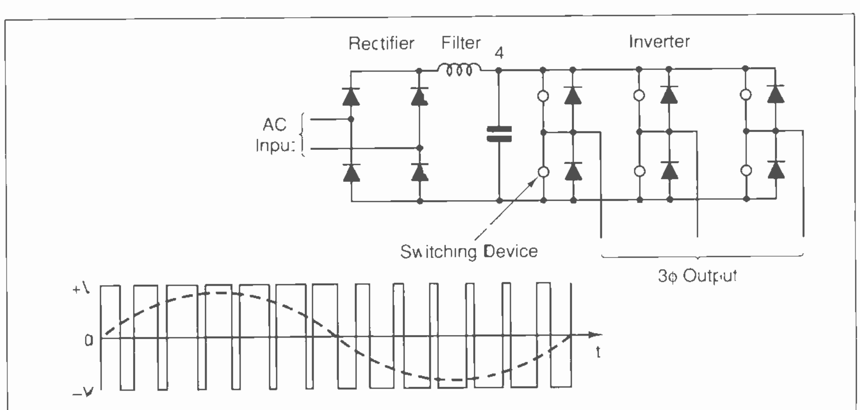
Further examples of pwm in action

Using pulse-width modulation, an inverter operating from a fixed-voltage dc supply can generate an ac output that is variable both in amplitude and frequency. In the diagram is a circuit for producing three-phase output from a single-phase ac supply. Load is switched alternately between the positive and negative rails of the dc supply. By controlling the switching instants of the power devices, an output waveform with the desired frequency and amplitude can be produced

Power mosfet switching speeds allow switching frequencies outside the audible range. Typically, the switching frequency is an order of magnitude higher than the output frequency. This makes output filtering a practical proposition so inverters can be made with virtually no harmonic and rfi problems. Audible acoustic noise is also eliminated, making such systems acceptable in domestic and office environments.

Traditionally, PWM has been generated by inputting a triangle wave and sinusoidal reference into a comparator. This method is gradually being replaced by digital waveform generation The digital method has the advantages of freedom from drift, absence of dc components in the output and perfect phase balance.

Digital techniques also allow non-sinusoidal waveforms to be produced. An example of the need for this is when output of the inverter needs to be increased by adding a third harmonic to the phase voltage waveform.



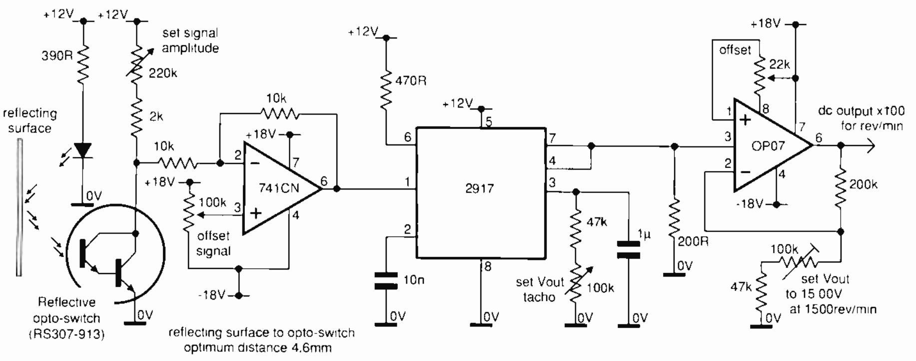


Fig. 7. Feedback used to determine the speed of the motor is processed by this tachometer circuit which turns frequency into a dc level. Pulses representing rotor speed are derived from a disk on the motor shaft whose black stripes are detected by a reflective opto sensor.

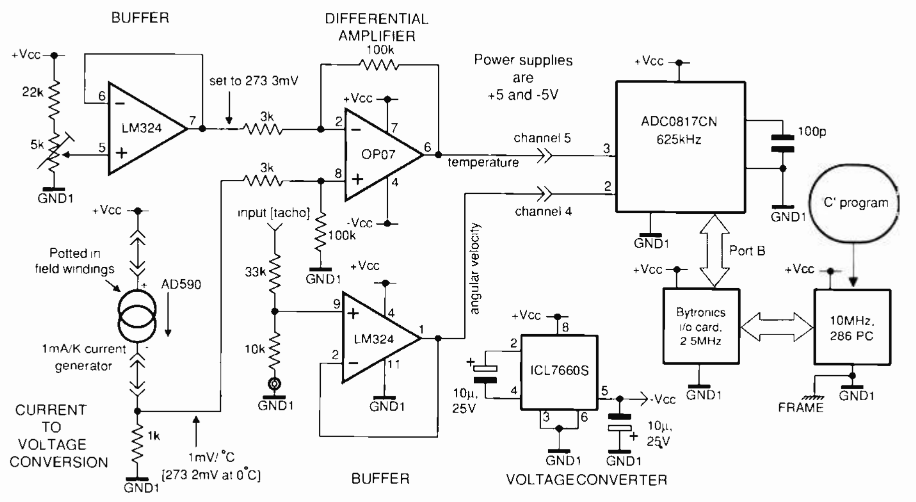


Fig. 8. Motor temperature, derived from an AD590 sensor potted into the motor field winding, is fed to one input of the PC i/o card. A second input receives speed feedback from the motor tachometer circuit.

voltage converter is offset by the mid-left op-amp. Output of the converter is a dc voltage directly proportional to the frequency of the timer output. A further op-amp amplifies this signal for driving the non-inverting input of the comparator.

Inverting input of the comparator comes from the oscillator, top left, so a PWM waveform is obtained at the output of the comparator. This signal drives the gate of the power mosfet. The oscillator is a relaxation type with two time constants, namely T_1 determined by the 22nF capacitor and 6.8kΩ resistor, and T_2 by the same capacitor and forward bias resistance of the diode. This produces a waveform with a periodic time of 160µs, i.e. 6.25kHz, which is the constant switching frequency of the power mosfet.

Figure 6 shows the waveforms of the signal processing module. Output from the comparator in Fig. 5 controls switching of the relay in Fig. 3. The first waveform is integrated by the RC combination before the comparator inverting input, giving about 1.2V dc at the input. A reference of about 0.6V is set at the comparator non-inverting input via the 5kΩ potentiometer. As a result, when this waveform is present, comparator output is low and the relay is energised. When the waveform is not present, i.e. when timer zero is disabled, comparator output goes high and the relay is de-energised. In this situation, the live line is opened, removing power from the entire system.

Transducers

Speed of the dc motor is monitored by the computer. The tachometer shown in **Fig. 7** is fed by a reflective opto-switch placed about 4.6mm away from a disc connected to a motor

Mosfet versus bipolar in motor drives*

Due to the high switching losses associated with bipolar transistors, mosfets have clear advantages in motor drives switching at 20kHz or more. However, for PWM drives operating at 1 to 2kHz, mosfets must be compared with bipolar transistors on the basis of conduction losses and cost.

Conduction losses in a mosfet can be reduced to any desired level by using a device with a large enough die, or by paralleling devices so the issue is mainly one of cost. Consider an application needing 400V transistors and where maximum load current is 5A. The nearest equivalent to a power mosfet is a Darlington transistor with an integral free-wheeling diode. A bipolar device rated at 8A will generally have a gain that is just adequate at 5A. Such a device would typically have a guaranteed maximum V_{CE} value of 2V when carrying 5A.

The *IRF350* is a 400V power mosfet with a drain-source on resistance of 300mΩ at 25°C. At 100°C, its resistance rises to 0.5Ω. This means that peak forward voltage drop at 5A and 100°C is 2.5V, which is roughly equivalent to that of the bipolar Darlington. However, switching and base-drive losses will be lower for the mosfet. Further, the Darlington may need a snubber to make sure that it operates within its safe operating area during switching.

Advantages offered by mosfets increase when base/gate drive requirements and the ability to withstand adverse operating conditions are considered. While bipolar transistors need both positive and negative base current to achieve respectable switching times, mosfets can be driven directly from buffered logic.

Additionally, surge current ratings for mosfets are approximately four times their average current rating. The *IRF350* for example can withstand 60A surges, representing a 1200% overload.

If a mosfet-based motor drive needs upgrading to deliver more power, it is usually a matter of simply replacing the mosfets. In a bipolar-based design, base-drive and snubber circuits will probably need altering to suit the new transistor. Mosfets such as the *IRF350* need no snubber.

*This information is derived from a comprehensive application note entitled *Using Hexfet III in pwm inverters for motor drives and ups systems*, from International Rectifier.

Features relevant to motor drives of mosfet versus bipolar transistors

IRF350	**Bipolar device**
Integral free-wheel diode	Some Darlingtons have diode
No snubber needed	Snubber often required
Low switching losses	High switching losses
Can be driven directly from buffered logic	New base drive and snubber designs needed for different power rating
High overload capability	Limited overload capability
Avalanche operation	Usually no avalanche capability

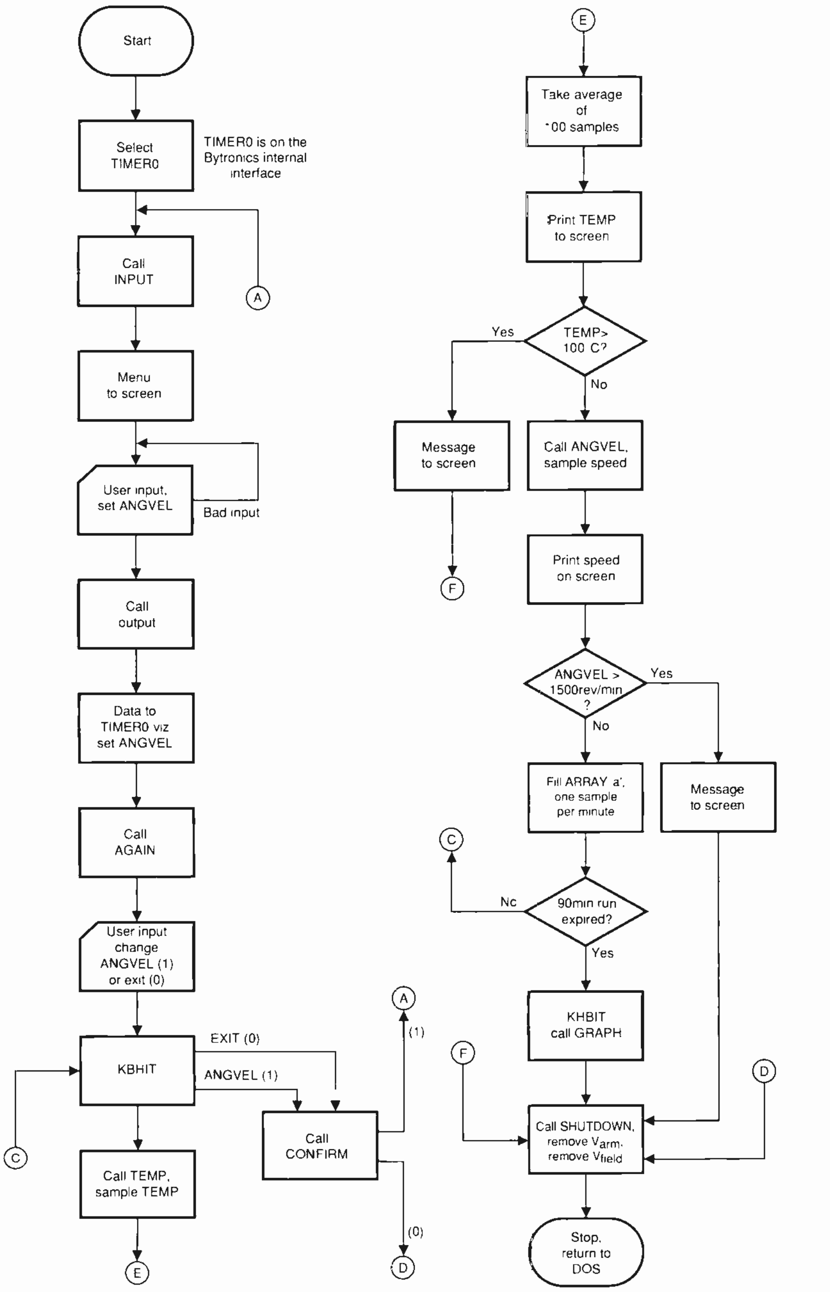


Fig. 9. This software, written in C, runs on a PC to control motor speed. As well as controlling motor speed using tacho feedback information, it monitors winding temperature and provides user interfacing.

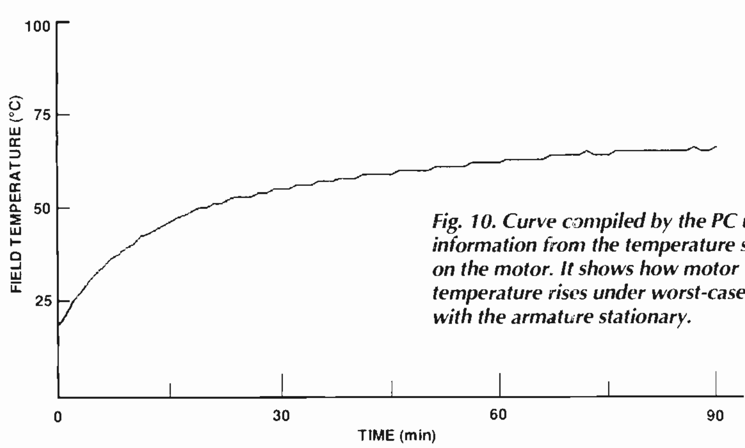


Fig. 10. Curve compiled by the PC using information from the temperature sensor mounted on the motor. It shows how motor field temperature rises under worst-case conditions, i.e. with the armature stationary.

Critical software timing

There are two global variables in the software, Fig. 9 – COUNT and *i*. Variable COUNT in line 183 of MIFV.C controls the sampling interval of the field temperature. I set this variable to 55, which gave one minute intervals. Variable COUNT is in the TEMP function.

Variable i, in line 194 of MIFV.C, controls the total run time of the motor. For example testing for 'if *i*>91' will produce a run cycle of 90 minutes. Variable *i* is also in the TEMP function.

Motor speed at which SHUTDOWN is called is limited by setting a variable called *av*. This local variable is in the angular-velocity ANGVEL function.

There is a further local variable called av in the TEMP function. Not to be confused with *av* in ANGVEL, this variable sets the upper limit of field winding temperature at which SHUTDOWN is called.

The software is set so that temperature is sampled once per minute and 90 samples are taken. You can set your own parameters by assigning values to global variables COUNT and *i*.

Data is stored in an array A[*i*] and a graph of field temperature as a function of time is drawn following a KBHIT after a 90 minute run.

shaft. There are 32 reflective strips on the disc printed on to a matt black background. Output of the tachometer is 15V dc for 1500rev/min.

In **Fig. 8** is the circuit for processing the transducer signals. Here, the 33kΩ/10kΩ resistors divide tachometer potential to give 3.5V at 1500rev/min. This voltage forms input to channel four of the *ADC0817* analogue-to-digital converter.

Temperature of the field windings is monitored by the computer. Temperature transducer *AD590* is a constant current source type with an output of 1µA per kelvin. It is potted in epoxy within the field windings of the motor. A precision 1kΩ resistor to ground effects current-to-voltage conversion from 1µA/K to 1mV/K.

Gain of the differential amplifier is 33.33. Level at the buffer output is set to 273.2mV by the 5kΩ potentiometer on the left. Output of the *AD590* is 273.2mV when temperature 0°C. Hence at 0°C, output of the differential amplifier is 0mV and at 100°C it is 3.33V. From differential amplifier, the temperature signal is fed to channel 5 of the A-to-D converter. A negative voltage rail, needed for the differential amplifier, is provided by the *ICL7660* voltage converter.

Control software in C

Called MIFV, the C control program has nine user defined functions. Its source code is about

11Kbyte and the compiled version, in .EXE form, is about 43Kbyte.

Figure 9 is a flowchart for the whole process. User-defined functions are,

MAIN
OUTPUT
INPUT
AGAIN
CONFIRM
ANGVEL
TEMP
SHUTDOWN
GRAPH

Very simply, the program works as follows. INPUT is called so that the desired motor speed can be entered by the operator via an on-screen menu.

Next, OUTPUT is called to set up the timer on the PC card. Routine AGAIN now invites the operator to change motor speed or exit. Should exit be requested, SHUTDOWN is called. Meanwhile temperature of the field windings is monitored together with motor speed. If either exceeds user defined limits then SHUTDOWN executes.

Field temperature and motor speed are constantly written to the screen. In my system, maximum field temperature is set at 100°C and maximum speed at 1500rev/min. Graph **Fig. 10** shows how field temperature rises in the dc motor for a stationary armature in a worst case condition.

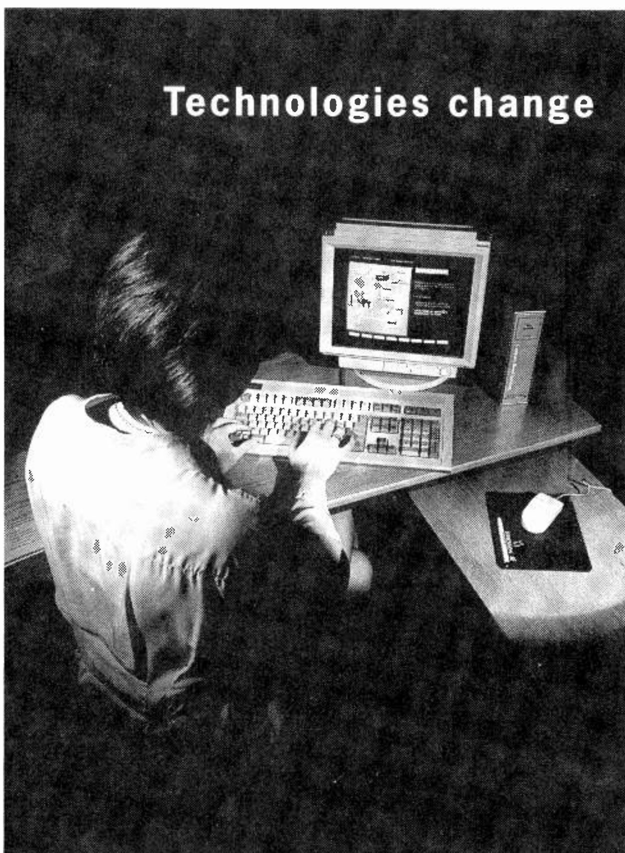
In summary the program first invites you to select any one of 40 discrete motor speeds. It then monitors field winding temperature and motor speed, closing down the system should temperature exceed, say 100°C or speed rise above 1500 rev/min. Shut down also occurs when 'exit' is requested from the menu. The SHUTDOWN routine removes armature voltage followed by field voltage. Finally, a graph is drawn and control returns to DOS.

Port B of the Bytronics i/o card carries data from the a-to-d converter. Port A is used for control with bit 7 feeding the converter output enable and bit 6 starting conversion. Bits 2 to 5 form address lines A to C respectively. These address lines are also needed to control the converter.

A '286-based PC with a 10MHz clock gives about 40 discrete motor speeds. Any computer used will clearly need graphics dump software. ■

Motor control software on disk

A disk is available containing the C motor control program covered in this article. It includes a 10Kbyte C source file, 6K of object code and a 42K execution file. A copy of the disk can be obtained by sending £10 to *EW&WW*, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.



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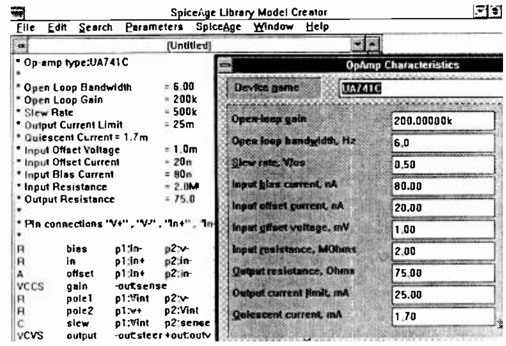

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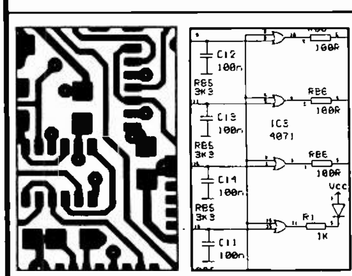

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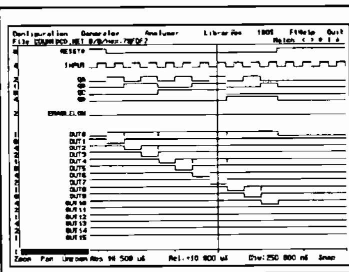

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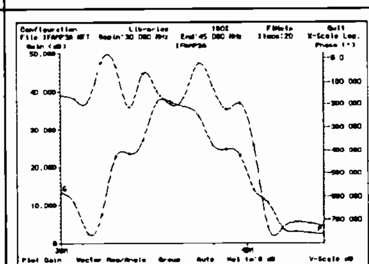

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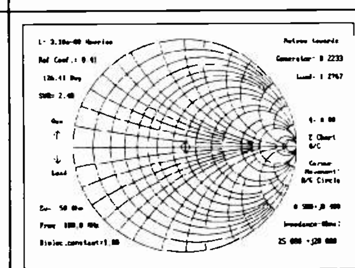

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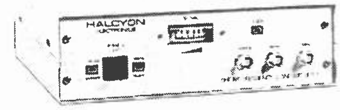

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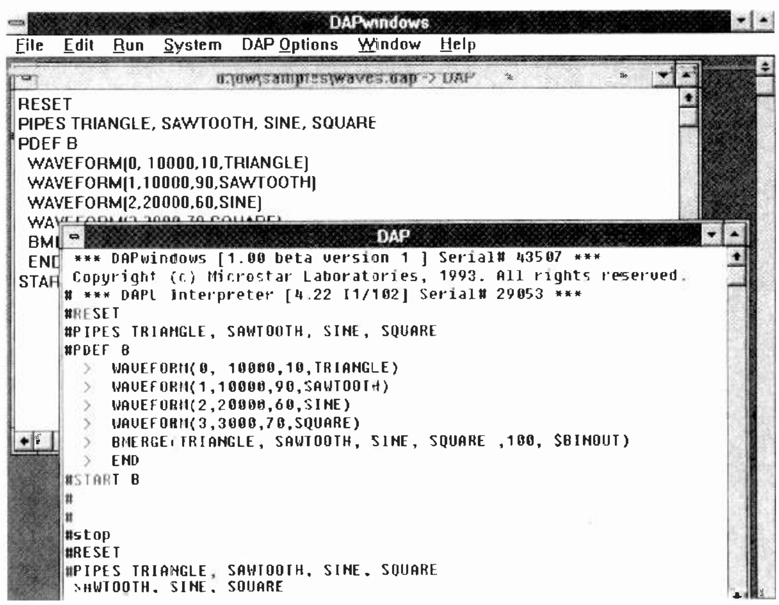
The right acquisition for PC data gathering?

Amplicon's DAP series of data acquisition hardware and software provides the complete development environment for specific data acquisition and processing applications. Allen Brown offers a developer's guide to a developer's system.

One of the principal uses of the PC in the engineering laboratory is to acquire signals from systems. The variety of analogue acquisition expansion cards, which are commercially available, indicates healthy growth. There are many companies which supply hardware and software for this market and the new entrant to the subject will probably be overwhelmed by the variety offered.

Two types of users of PC data acquisition systems can be identified. The first type of user would expect to purchase an analogue acquisition card, complete with operating software and have the whole thing up and running within half an hour. This is equivalent to an off-the-shelf product which performs a range of general purpose functions.

The second type of user would be interested in a tailor made product they would customise from a set of proprietory tools. This user would find a splendid product available from Microstar Laboratories (supplied by Amplicon) which has the collective title of the *DAP Series* (Data Acquisition Processor). The product consists of PC expansion card hardware and a comprehensive suite of software development tools.



An example of the DAP programming language (DAPL) written with the DAPview for Windows option.

DAP expansion cards

When using a traditional expansion card with analogue and i/o facilities only, all processing must be performed by the CPU in the PC. Not an ideal state especially if the PC is executing several concurrent tasks. An alternative approach is to transfer the intelligent processing from the PC onto the expansion card itself. This design principle underlies the *DAP Series* of products for intelligent data acquisition.

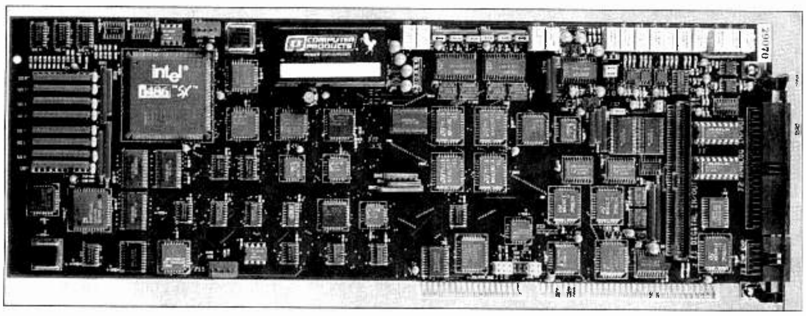
The hardware includes a range of intelligent expansion cards which have digital and analogue i/o facilities together with a microprocessor controller. The cheapest card (*DAP800/1*) has a sample rate of 75kHz with an Intel 80188 whereas the top of the range card (*DAP3200e/102*) has a sample rate of 312kHz and features a 80486SX 32-bit microprocessor. All the cards have Intel processors apart from the *DAP2400e* which hosts the Motorola 56001 digital signal processor.

Many data processing tasks can be greatly speeded up by using a DSP processor and for such applications the *DAP2400e* is worth considering. However it must be recognised that for intensive numerical processing requirements this family of expansion cards is up against some formidable opposition... Burr Brown and Loughborough Sound Images for example.

Each DAP expansion card is supported by its own onboard memory. For example, the *DAP3200e* is furnished with 4Kbyte of dynamic ram which can be used for either data or program storage. As expected from a range of data acquisition cards, each has several multiplexed analogue i/o and digital i/o channels. One of the recognised problems of using expansion cards for data acquisition is the physical positioning of the i/o pins which are hidden at the back of the PC and are not always easy to access. To alleviate this problem the expansion card can be connected to a bench top printed circuit board via a ribbon cable. The printed circuit board allows easy access to all the i/o lines with room to add load resistors as required. It also has a small patch area for the user to add extra gates which can be powered from the PC. The choice of DAP expansion cards represents a comprehensive range of options allowing the potential user to tailor their requirement with a specific board.

Software development tools

Having an extensive range of intelligent PC expansion cards would be of little value without suitable software development support products. Each *DAP* expansion card can host a real-time, multitasking operating system called *DAPL*. Being



Top of the range performance. The 32-bit DAP3200e/102 samples at 312kHz. But it carries a top of the range price: £3379.

resident on the card, this system may be accessed directly by downloading ascii command files from the PC to the expansion card. The programming of the expansion card is therefore performed by using the programming language. Like any other, *DAPL* can be structured with conditionals and constructs with the provision for error checking. It has a vast range of high level instructions referred to as the command list. The list consists of five command groupings:

- data processing
- input configuration
- output configuration
- function definition
- system control.

The data processing group has over eighty commands ranging from FFTs and proportional integral differential (PID) controllers to signal generators. The user's program would therefore consist of a sequence of instructions from the command list. However the user has an alternative method of using the *DAPL* instructions. Microstar Laboratories also supply real-time plotting software called *DAPview* which allows interactive communication between keyboard, screen and expansion card. When *DAPview* is up and running the user can enter, via the keyboard, instructions from the command list which are transferred to the expansion card (where appropriate) and executed. By this means the user has an easy method of changing process parameters such as gain, sampling rate and output control signals. In fact it is quite possible to configure the expansion card purely from within *DAPview* which is obviously attractive for the casual user. The software offers real-time graphics in various modes (line, bar and scatter plots), control over grids, screen colours, log or linear scaling and a number of other functions.

Microstar Laboratories has take the product development one step further by supplying an enhanced version called *DAPview Plus*. This version contains an on line text editor for generating command files to control the *DAP* expansion card. To further the appeal of *DAPview PLUS* it is possible to implement a turnkey solution which will operate without user interaction.

PRICES

DAP3200e/102 32-bit main board	£3379
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DAP and Windows

A supplier of software to run on a PC today must take into account the potential demand for the product to run under Windows. Microstar has addressed this in a number of ways. Firstly, by providing a Windows toolkit and secondly by providing DaisyLab – Windows application software which will be available in a few months.

The Toolkit makes use of the now infamous Dynamic Link Library (DLL). This means that the resident *DAP* expansion card can be accessed by using any Windows development language such as Visual Basic, Visual C^{++} and Excel. Microstar Laboratories also supply an advanced development toolkit which integrates with Microsoft C. However with the shifting emphasis towards C^{++} using object oriented design, it would probably be more appropriate for Microstar Laboratories to offer a choice of development toolkits.

One of the main problems of any software running under Windows is the time consuming interrupt latency (time lost as the processor switches from task to task) an ever present burden in multitasking processing. However by using the DLLs the processing can be off loaded to the expansion card. Therefore custom realtime data acquisition and control software can be developed in run very efficiently under Windows.

The user that gets on well with *DAPview* will also appreciate a dedicated Windows version – *DAPview for Windows*. It operates similarly to the dos version except in a Window. This provides the opportunity of multitasking processes within the Windows environment.

DaisyLab, the other Windows software for *DAP*, is aimed primarily at engineers who are not strongly inclined towards programming. It is designed around a graphical user interface which enables the user, with relative ease, to access the *DAP* expansion cards directly from Windows through the use of icons. A separate review will be published in a future edition of *EW+WW*.

User manuals

The series is well supported by a set of user's manuals, no less than five with the above products. The *DAP* manual contains a detailed discussion of all the entries in the command list and individual examples of use. There is a glossary defining many of the words which are peculiar to the product range. For example a *Pipe* is a first in, first out dynamic data buffer between tasks. It needs to be dynamic in size because the data in and data out rates may vary; it would be quite difficult to glean this definition from other parts of the manual. The systems manual gives a detailed account of *DAPview* and *DAPview plus* together with examples of use when programming in C, QuickBasic, Pascal and Fortran.

The hardware manual discusses the detail of *DAP* expansion cards and the applications manual provides examples of the programming language *DAPL*. Finally, the Windows toolkit manual discusses in detail how to integrate the control of DAP expansion cards (using DLLs) into either Visual Basic, C/C^{++}, Turbo Pascal or Excel programs.

The *DAP* product series is a well engineered concept aimed at custom data acquisition systems. It will appeal directly to the engineer involved in designing specific purpose systems. All the tools are present for constructing closed loop control systems using PID control methods and multichannel data acquisition. It is however not for casual use but more for a minimum system for specific applications. After all, minimum systems are easier to test, maintain and modify by other designers if necessary. *DAP* is well suited for this purpose. ■

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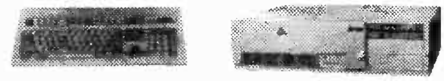

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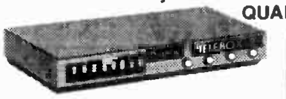

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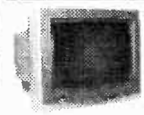

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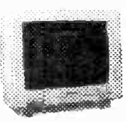

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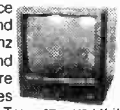

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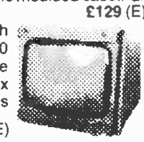

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Easy path to 8-bit radio links

Radio data transmission needn't be a complex business. Steve Winder describes a simple encoding/decoding system based on standard logic parts.

Serial data transmission over radio links, such as the Radiometrix devices described by Ian Hickman (Feb 93) is easily achieved with cheap, dedicated devices. The Motorola *MC145026* encoder and the *MC145027* decoder are one such chipset. With an *MC145026* in the transmitter and an *MC145027* in the receiver it is possible to transmit a 4-bit word, complete with error checking and a 5-bit tri-state address key.

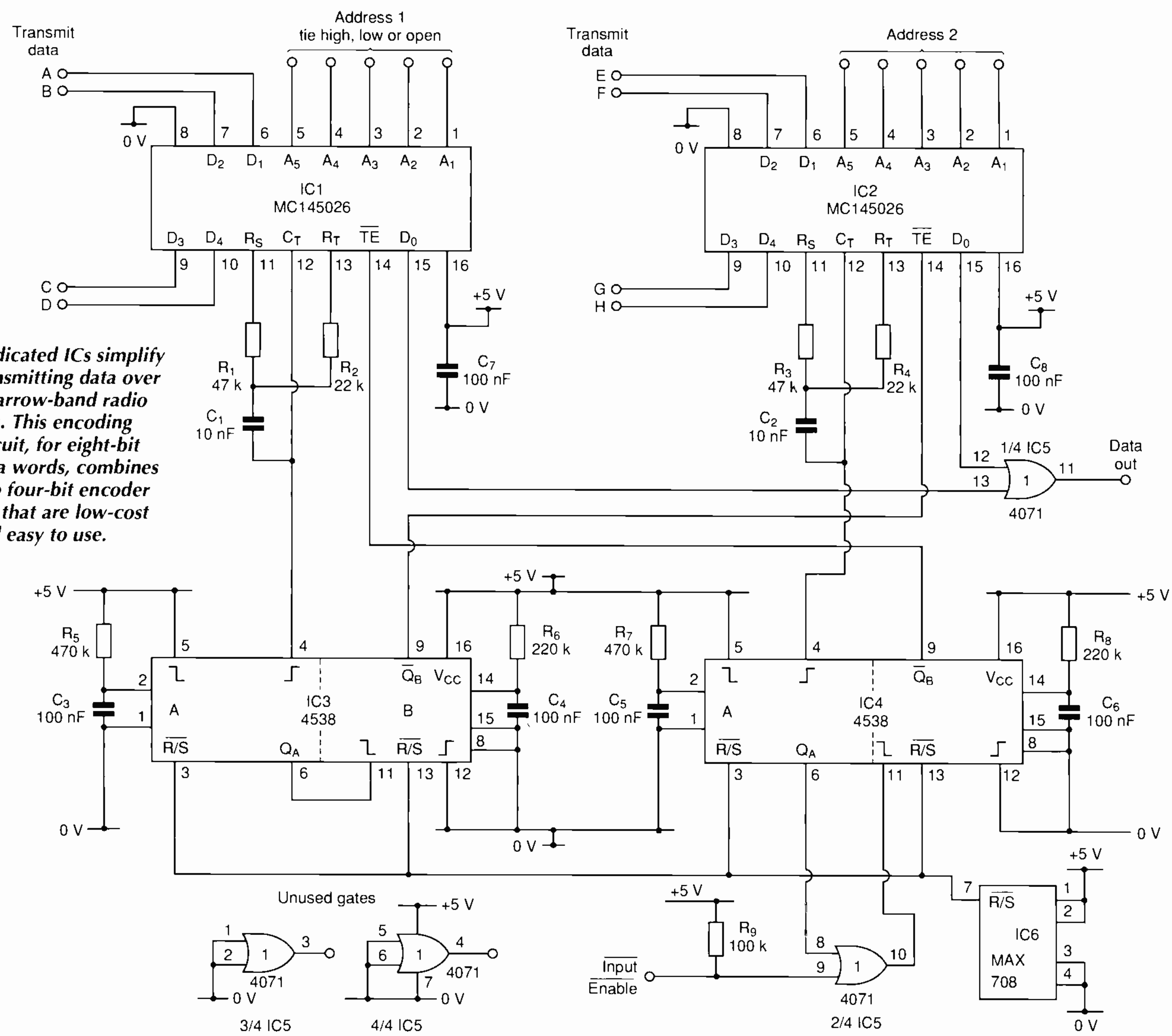
Applications exist for eight-bit word transmission, for instance sending data from an analogue-to-digital converter. While it is possible to use a microprocessor to generate codes for transmission, this requires development time and capital investment. A solution using standard logic, combined with the Motorola chipset, was chosen.

The encoder IC will transmit the address and data word twice if its enable pin is momentarily taken low, but will continually transmit data if its enable pin is held low. The decoder IC will output a VALID DATA pulse after it has received two correctly addressed messages that contain the same data. The circuits described in this note were required to behave in an identical manner to the stand-alone chipset, except that eight bits are transmitted. It is possible to extend these circuits to provide transmission of more bits.

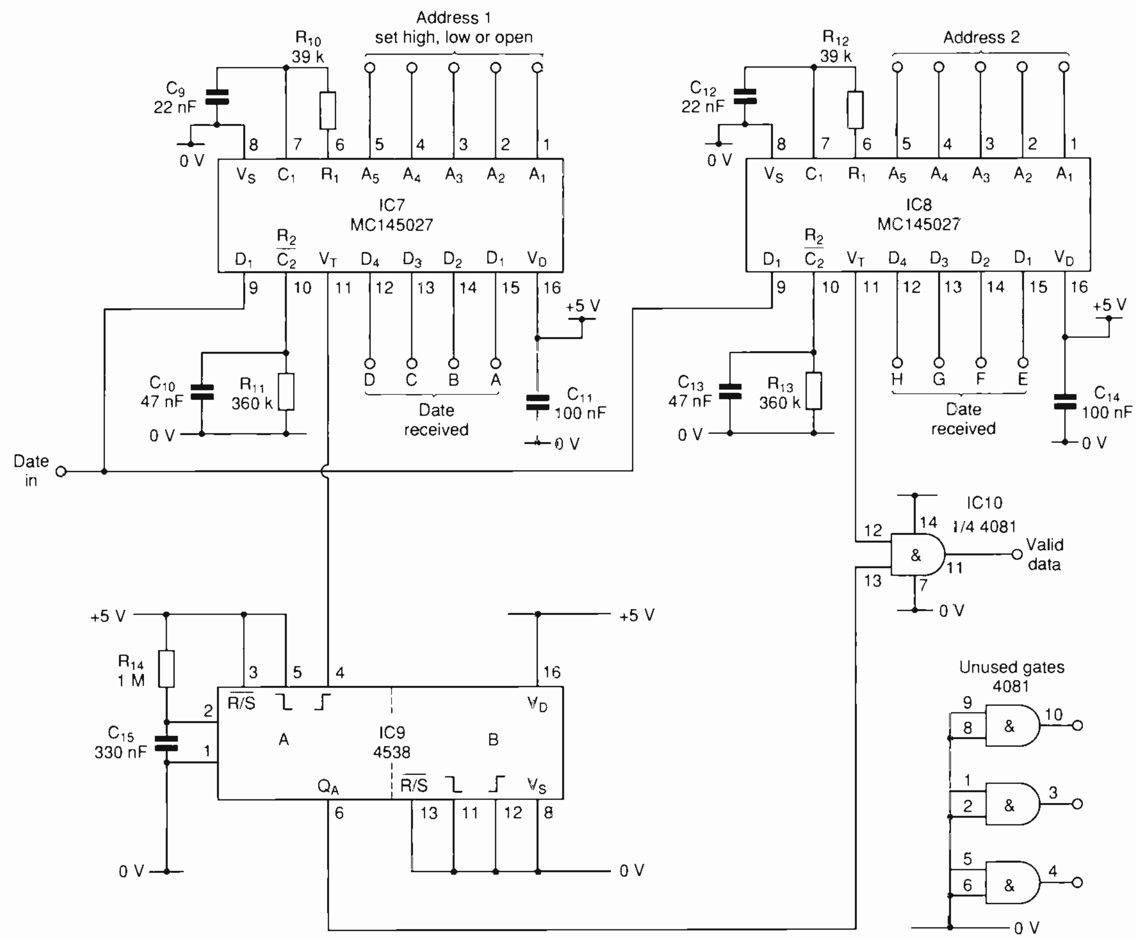
Transmitter circuit

The circuit given in **Fig. 1** uses two *MC145026* encoders. Transmission begins when the INPUT ENABLE pin, connected to OR gate $^2/_4$ IC_5, is taken low. The other OR gate input is low, so its output falls to logic 0 and triggers the monostable, IC_{4b}. This monostable then produces a logic 0 condition for 22ms which enables transmission from the encoder IC_1. This encoder activates its internal clock generator, and then transmits an address and data word twice.

While the first encoder is transmitting data its clock generator continues to operate. The clock pulses appear at pin 12. These pulses are used to trigger the monostable IC_{3a}. The monostable is continually re-triggered and its output remains high while clock pulses are present at its input. When the encoder has finished its transmission cycle it stops generating clock pulses. This allows the monostable out-



Dedicated ICs simplify transmitting data over a narrow-band radio link. This encoding circuit, for eight-bit data words, combines two four-bit encoder ICs that are low-cost and easy to use.



In common with its mating encoder, Fig. 1, this decoder for radio-link data is designed for eight-bit words but is easily expandable to any word length in multiples of four bits.

put to return low after a period of 47ms, set by C_3 and R_5.

Monostable IC_{3b} is triggered by the falling edge of IC_{3a} output, 47ms after the first encoder IC_1 has finished its transmission. The output of IC_{3b} goes low for 22ms which enables transmission of data from the second encoder, IC_2.

The second encoder generates clock pulses while transmitting data. Using similar monostable circuits as described for the first stage, IC_{4a} maintains its output at logic 1 until 47ms after the clock pulses stop. The output of IC_{4a} is connected to the input of an OR gate, $^2/_4$ IC_5, to which the INPUT ENABLE is also connected. The output of this OR gate is used to trigger the monostable IC_{4b}. If the input enable pin has returned high, the monostable will not be re-triggered. If the input enable pin is low, the pulse from the second stage counter will cause the first monostable to re-trigger and the full eight bits will be re-transmitted. This makes the overall circuit behave in an identical manner to a single encoder.

The encoder clock rate is set by resistors R_1 to R_4 and capacitors C_1 and C_2. The values chosen give a 1.9kHz clock. This gives reliable data transmission over a narrow band radio channel. The clock rate could be increased to 4kHz by changing the capacitor values to C_1 and C_2 = 4.7nF. Resistor values are unchanged.

The monostable pulse width periods could be shortened. The encoder IC transmit enable pulse is set by the 22ms monostable period, this must not be reduced below 65ns. Reducing the transmit enable pulse width will have no effect on the transmit time; the encoder clock starts as soon as the pulse goes low. The time between transmissions should be at least 27 clock periods long, i.e. three data bits. This is set by the monostable period of 47ms. It should not be reduced below 15ms at the 1.9kHz clock rate. Using a 4kHz clock rate, this period could be reduced to a minimum of 7ms.

A microprocessor supervisory IC, the *MAX708*, is used to reset the monostables on power-up. This is to ensure that no invalid data transmission occurs at the start of circuit operation.

Unused gates should have their inputs tied low.

Receiver circuit

The receiver circuit is simpler, see **Fig. 2**. Two *MC145027* decoders are connected in parallel to the output of the receiver. Four messages are received from the transmitter, two addressed to one decoder followed by two addressed to the other. Each decoder is set to an address that matches the corresponding encoder at the transmitter end.

When the first decoder IC_7 receives a valid address, the following data stream is decoded. After receiving a second identical message the data is delivered, together with a valid data pulse. This pulse triggers a monostable, IC_{9a}, so that the circuit remembers that the first word is valid.

The second decoder IC then decodes data from the third and fourth messages. If these messages are identical, the data and a valid data pulse are produced. This VALID DATA pulse is logic and-ed with the output from the monostable. A valid data state will be present when both words have been received correctly. The decoder timing components are set for a 1.9kHz encoder clock rate. If the clock rate is increased to 4kHz these components should be changed to: R_{10}=R_{12}=27k; R_{11}=R_{13}=240k; C_9=C_{12}=15nF and C_{10}=C_{13}=33nF. The monostable period of IC_{9a} is 330ms which may be decreased to 100ms if the period between data streams is reduced below 47ms. Reducing the value of C_{15} to 100nF will accomplish this.

Extending the circuit

It is possible to extend the circuit to transmit a greater number of bits, in multiples of four.

In the encoder circuit, IC_1 and IC_3 form a module which can be repeated, each monostable 'B' output triggering the next encoder. Circuits using IC_2 and IC_4 would remain at the end of the encoder chain, feeding back a trigger pulse to the first encoder if the INPUT ENABLE pin is held low. The spare OR gates could be used to combine up to two more encoder stage outputs.

The decoder circuit could be modified by adding further decoder ICs and monostables as required. Only one half of a monostable IC is required for each decoder IC added to the circuit. Unused AND gates could be used to combine the additional monostable outputs, for a valid data indication. ■

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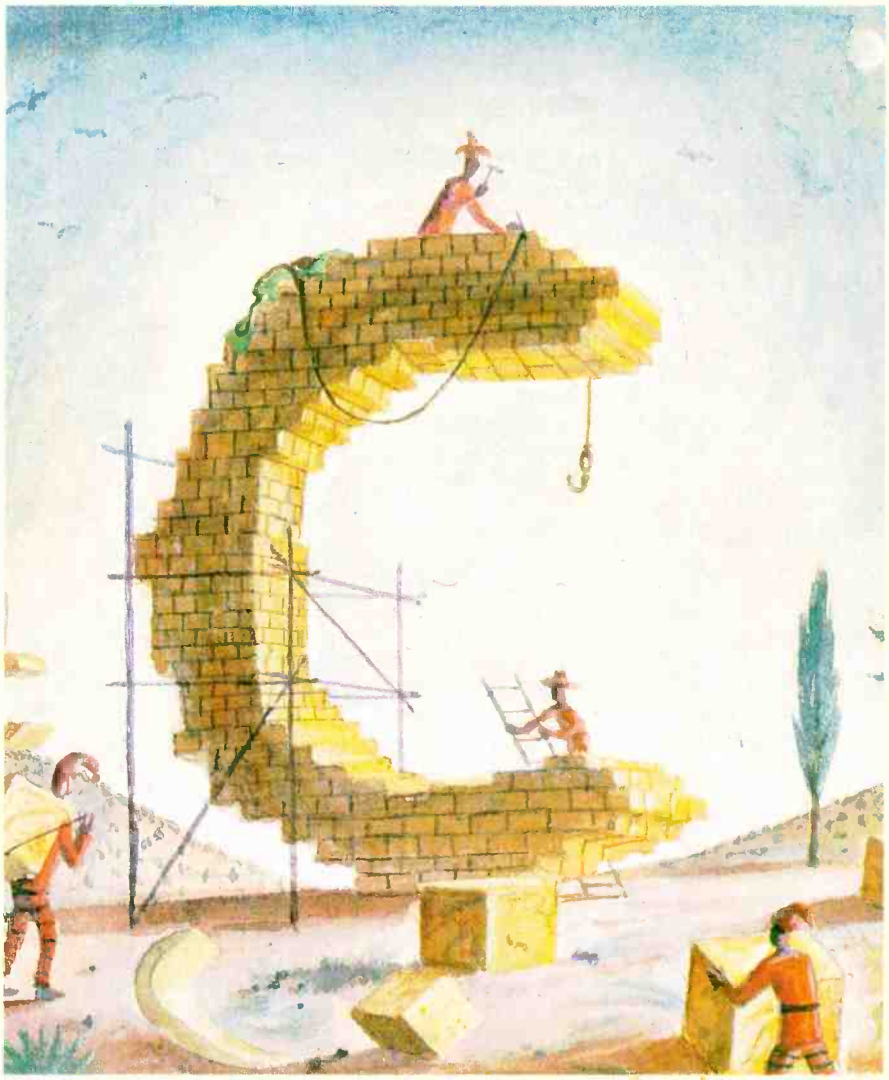

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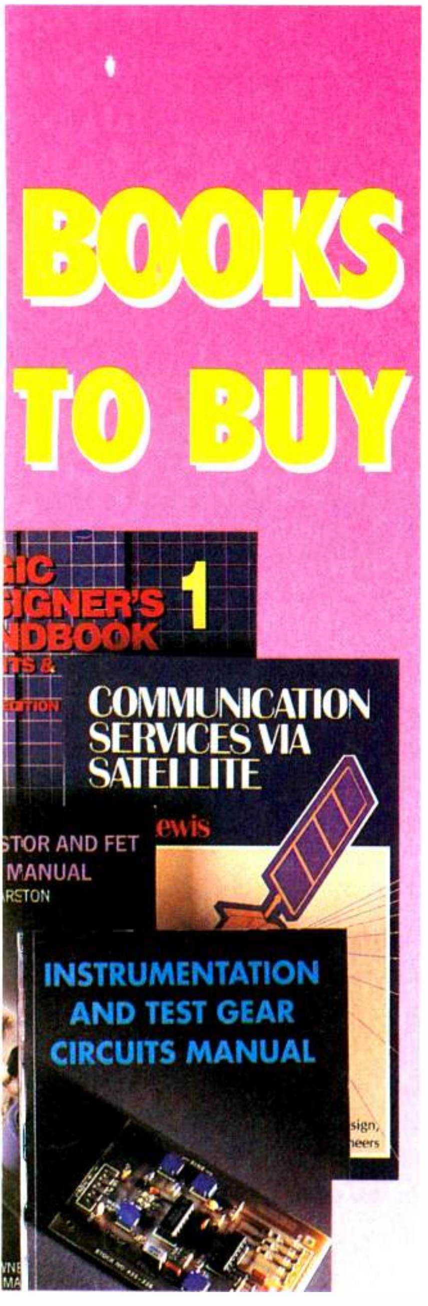

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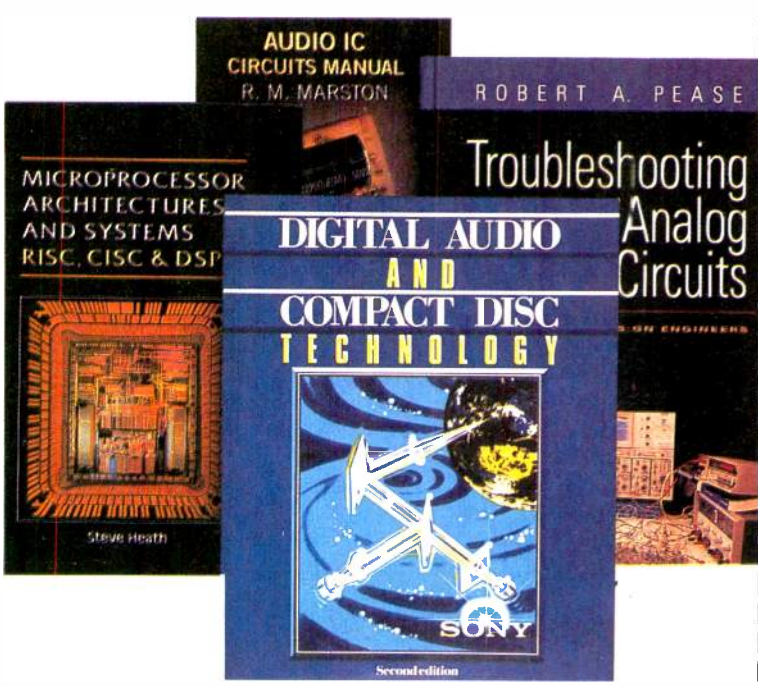

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USING RF TRANSISTORS

Single-ended, parallel, or push-pull power?

How do you select the best power amp configuration and circuit components, and design appropriate matching networks? Norm Dye and Helge Granberg show how frequency spectrum, band-width and power level influence choice, and discuss the pros and cons of different component strategies. From the book RF Transistors: principles and practical applications.

Norm Dye is Motorol 's product planning manager in the Semiconductor Products Sector, and Helge Granberg is Member of Technical Staff Radio Frequency Power Group (Semiconductor Products) at Motorola. Their rf transistors book includes practical examples from the frequency spectrum from 2MHz to microwaves, with special emphasis on the uhf frequencies .

RF Transistors: Principles and practical applications is available by postal application to room ***L333 EW+WW, Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS.***

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288pp HARDBACK 07506 9059 3 Cost £19.95 + Postage £2.50

Optimum performance of a device is usually produced by single-ended narrow band amplifiers. Such circuits are ideal when power gain or other information needs to be compiled for a specific application or where an amplifier is to be used for a single frequency.

Up to about 500MHz, lumped constant matching networks can be used, while strip-line designs are common at 1000 to 2000MHz and higher. Strip-line concepts are actually the most practical at uhf and microwave frequencies. For vhf and lower uhf, etched air-line inductors – resembling lumped-constant elements – may be the best choice for inductance, particularly for production repeatability. Proper techniques make possible bandwidths of an octave or more.

Higher power levels, not possible with single transistors, can be reached by paralleling – a technique widely used at microwaves where push-pull designs for higher power levels become too critical. But many problems can be encountered too (such as extremely low impedance levels and uneven power sharing if the devices are not closely matched) and so the technique is not usually recommended without special guidelines.

At low band and up to uhf, push-pull circuits offer certain advantages over the other two circuit configurations. The most important is probably suppression of even-order harmonics, though that depends on the matching of the two devices. Other advantages are wider bandwidths, higher input/output impedances, and less critical bypassing – especially in the output circuitry.

Lumped circuits in single-ended rf amps

If a single-frequency or relatively narrow-band rf amplifier is the choice, one designed with lumped constant *LC* elements is probably the most economical and easiest to design – especially since the capacitances, and in some cases the inductances, can be made variable.

Circuits using lumped-constant elements for impedance matching the device's input and output to 50Ω are widely used for transistor test circuits up to about 300MHz (or up to 900-1000MHz for low power designs). The variable elements allow both adjustment for optimum performance and compensation for transistor parameter tolerances occurring from unit to unit.

Good emitter grounding is essential (**Fig. 1a**), and a lower ground plane should be provided, at least in the immediate area of the transistor mounting. A continuous ground plane may also be a good idea, depending on the exact circuit layout, just to provide low inductance grounding points for the capacitors, through feed-throughs to the top of the printed circuit board.

Foil pads usually provide appropriate locations for element interconnections to be soldered down. But since all *L*s and *C*s are surface mounted, at higher power levels and at continuous operation some circuit elements could heat up enough to melt the solder – a definite disadvantage in rf power amplifier design, though it may be lessened with suitable air flow.

Distributed circuits

In a typical uhf or microwave common-base amplifier circuit (**Fig. 1b**) impedance matching can be achieved completely with microstrip transmission lines.

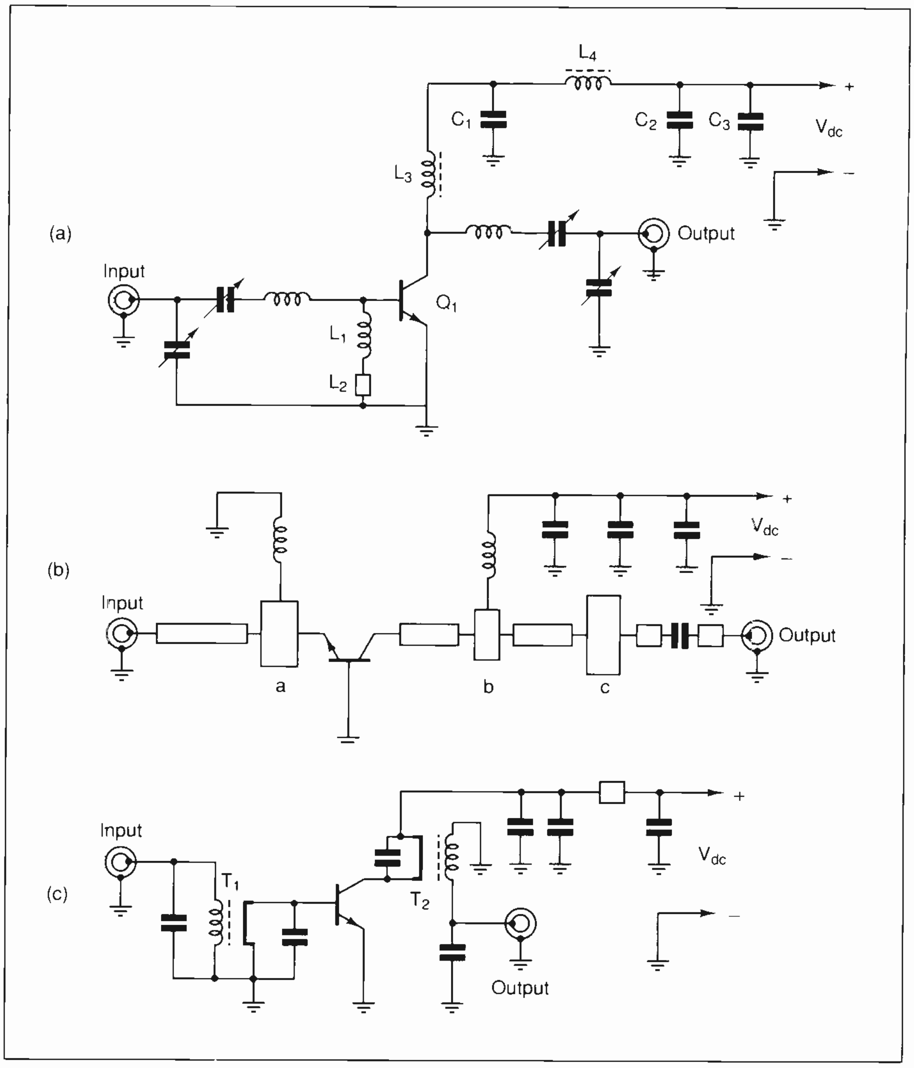
Design of these (and common emitter) circuits, requires that the exact dielectric constant (ε_r) of the substrate material be known.

Maximum ripple tolerable within a specified bandwidth determines the number of reactive elements (n) needed to match the input and output to 50Ω. Inductors are formed by stripline of specified widths and lengths, and capacitors by open stubs at specific points on the lines. A 3-4dB difference is experienced in ripple between $n = 1$ and $n = 4$, but after this there is only 0.5dB change up to $n = \infty$.

Manually, the reactive elements can be designed as Chebyscheff lumped-constant matching networks, which are then converted to microstrip format. But numerous computer programs are now available to calculate the line and stub dimensions directly.

Prototype line and stub dimensions can be modified by 'cutting off' metal or adding copper foil with its specially-developed conductive adhesive backing. Unlike lumped

Fig. 1. Single-ended rf amplifier circuit configurations.
In (1a), lumped constant matching limits use to relatively narrow band applications and frequencies up to vhf or low uhf. Although shown as a class C configuration, with proper biasing arrangements it can be biased to classes A, AB, or B as well.
(1b) is a typical uhf or microwave common-base circuit where impedance matching is achieved wholly with microstrip.
(1c) represents a wideband amplifier circuit where transformers are used for impedance matching. It is usable up to uhf in small signal designs.



circuit lower-frequency designs, feed-throughs to the circuit board bottom ground plane are not needed since the capacitors are formed with stubs (shown in the schematic Fig. 1b as *a*, *b* and *c*). Exceptions are shorted-stubs and ground returns for the input rf choke and collector supply by-pass capacitors.

For stability reasons, the base-to-ground inductance must be at its minimum. Fortunately, modern transistors make this relatively easy. A package configuration should be chosen with base leads connected directly to the mounting flange which in turn is grounded to the heat sink along with the circuit board ground. Common-emitter amplifier configurations also usually call for minimum emitter-to-ground inductance, but for a different reason: preventing loss of gain. Again dual-emitter packages (such as SOE) and/or packages with 'wrap around' emitter metallisation (no external emitter leads) will minimise the effect of common element inductance.

Quasi-lumped elements

Wide band amplifier circuits using transformers for impedance matching (**Fig. 1c**) are best suited to low frequency operation up to power levels of 50W. Or perhaps up to 500MHz in small signal use (100-200mW) where the impedance levels are high.

The conventional wide-band transformers, (T_1 and T_2) are limited in bandwidth compared with transmission line types. But this is not the main problem. 50-70W amplifiers up to 30MHz have been designed using this configuration.

However, good bypass of the transformer ground-return is difficult, even with multiple capacitors of mixed values used in parallel. Impedance levels are extremely low and rf currents particularly high at these points.

Circuits biased to any class calling for a positive base voltage will face the same problems on the input side. The input transformer ground-returns cannot be de-coupled and would have to be similarly bypassed to ground.

Good quality chip capacitors will improve the situation. But then connections to a solid ground become even more important than in other circuits (e.g. Fig. 1a).

High power paralleling

Higher power outputs than can be obtained from a single transistor can be achieved by

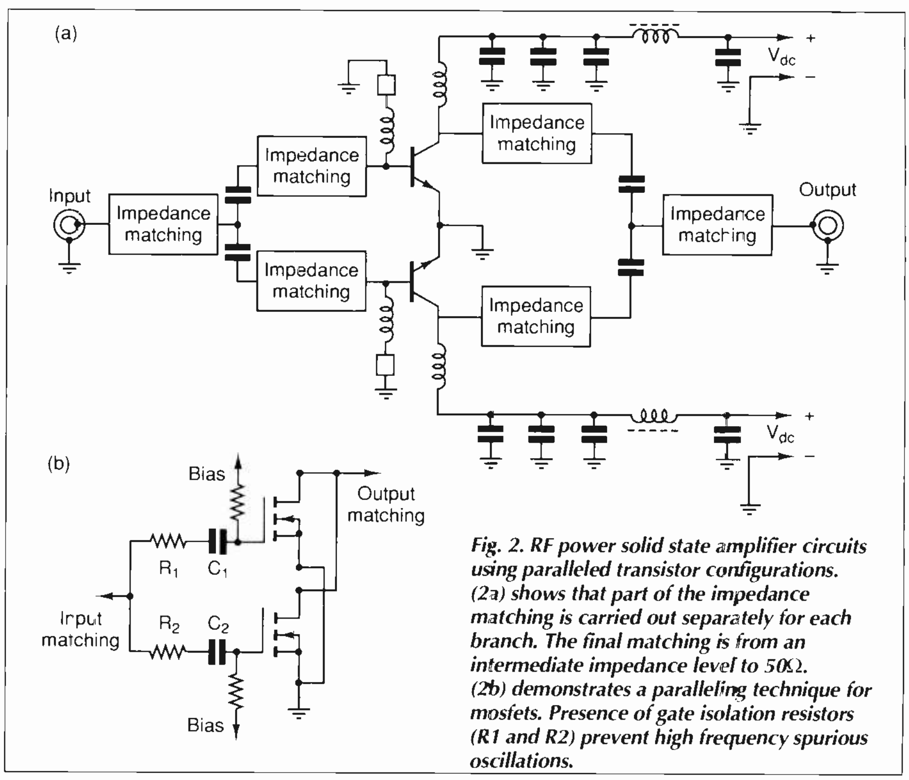


Fig. 2. RF power solid state amplifier circuits using paralleled transistor configurations.
(2a) shows that part of the impedance matching is carried out separately for each branch. The final matching is from an intermediate impedance level to 50Ω.
(2b) demonstrates a paralleling technique for mosfets. Presence of gate isolation resistors (R1 and R2) prevent high frequency spurious oscillations.

paralleling transistors in rf power amplifiers. Paralleling is usually done with the highest power devices available for a given application, otherwise it would be cheaper and simpler to select a higher power single device.

Impedance levels (especially at the input) become extremely low if the devices are directly paralleled. To avoid the resultant lossy matching-networks into 50Ω and difficult designs, the customary first step is to perform an impedance transformation to an intermediate level such as 10-25Ω. These intermediate impedance points for each device are then paralleled and the resultant transformed to 50Ω by additional matching networks (**Fig. 2a**) Paralleling more than two devices is rarely attempted.

The larger the number of transistors paralleled, the more impracticable the situation. In addition to the intermediate impedance dropping lower, all transistors must be closely matched in power gain and output capacitance. Also, for class A or AB, V_{BE}*(forward)* and h_{FE} must be matched unless the devices are individually biased.

Intermediate impedances for each device must be identical too, which is difficult to achieve except in microstrip designs.

But paralleling many transistors in low power applications can be feasible: for example if the desired power output is moderately low (say 2-5W) and the aim is to use inexpensive 1W devices in a TO-39 or similar header.

Transistor paralleling can be used for low-band applications up to microwaves and is commonly seen in L-band radar equipment.

Paralleling mosfets

Many designers who have tried to parallel mosfets have, to their surprise, experienced some unusual and seemingly inexplicable behaviour. Devices can'blow' when biased to a low idle current, or if not biased, when rf drive is applied.

But the explanation is that the parallel configuration forms an oscillator comparable to the emitter-coupled multi-vibrator known from bipolar circuit technology.

Mosfets have a high enough unity-gain frequency that the inductance formed by the gate/source bonding wires, the leads, and their external connection together with the device's internal capacitances form a resonant circuit which permits oscillations to occur. Oscillations are usually at a resonant frequency beyond the pass band of the intended amplifier, as high as 400-500MHz for higher power devices and up to 1000-1500MHz for lower power ones. High currents can flow at the oscillating frequency resulting in the destruction of the device.

Unless the designer accidentally detects the oscillations (usually with a spectrum analyser) and takes corrective action, devices can be lost and many headaches experienced.

Mosfets can be paralleled, but their gates must be isolated and the Q value of the resonant circuit lowered, with resistors (**Fig. 2b**) or comparable values of low-Q inductive reactances. Obviously, either method affects the device's high frequency performance. An RC or LC low-pass filter is formed between the outside input terminal and the gates because the C is the device's C_{iss}. This limits frequency of operation of the configuration to vhf at best, where the input impedance levels are still relatively high even with the isolation components added.

Fets have higher input/output impedance values, so the impedance-matching procedure described for bipolar transistors is not needed (Fig. 2a). Intermediate-matching may provide the gate-isolation necessary, though this has not been pursued by the authors.

The isolation scheme limits the high frequency performance of an amplifier. For example, with devices rated for a power output of 150W, a resistance or comparable value of low Q inductive reactance of 3-5Ω would be required at the gate of each transistor, limiting the maximum frequency of operation to below 100MHz. With smaller devices (30 to 40W), these resistance or reactance values would be of the order of 10 to 20Ω.

Mosfets are suitable really only for applications up to low vhf, and gate isolation is only applicable to push-pull circuits.

Push-pull amplifiers

Push-pull circuits offer certain advantages over single-ended and parallel transistor designs. They can be designed as a narrow band system using lumped constant elements or using microstrip techniques at higher frequencies. But such designs are rather critical, calling for extreme symmetry between each side.

Wide band designs, using transformers for impedance matching, are much more tolerable because of their ability to have 'floating' centre taps. The floating tap, whether in the input or output transformer, means that a physical centre tap is not needed – there is a 180° phase shift across the transformer winding in either case.

In a centre-tapped design the ground reference is well defined. But any imbalance in the two winding halves will be reflected to the transistors, resulting in an amplitude difference in the drive signals to each side of

Power amp pros and cons

Lumped circuits
Advantages: Adjustable components allow a transistor's best performance to be reached at a specific frequency; no special components required; relatively inexpensive.
Disadvantages: Not suitable for continuous operation at high power levels; limited power output capability; limited frequency range; poor repeatability for mass production.

Distributed circuits
Advantages: Easy design procedure; low number of components; good repeatability for mass production.
Disadvantages: Limited to a narrow frequency spectrum without redesign; expensive substrate material; power handling capability limited at cw (used mostly for pulsed operation at higher power levels).

Quasi-lumped elements
Advantages: Broad band performance (three-five octaves); inexpensive; good repeatability for mass production; can be designed to achieve peak performance from the transistors.
Disadvantages: Limited power output capability; critical circuit layout for optimum performance (large signal); limited frequency range; requirement for special components.

Paralleling bipolar devices
Advantages: High power output is possible by using two or more devices; circuit repeatability is good in microstrip designs; does not require 180° phase shift in input and output.
Disadvantages: Requires closely matched devices and tight passive component tolerances; no even order harmonic suppression; at lower frequencies the design is more critical than push-pull.

Paralleling mosfets
Advantages: High power output is possible by using two or more devices; no intermediate impedance matching required; excellent power sharing even with poorly matched devices.
Disadvantages: Limited high frequency operation due to necessary gate isolation.

Push-pull amplifier configurations
Advantages: Even order harmonic suppression; easier input/output matching over single-ended and parallel designs due to higher impedance levels; emitter/source grounding and collector/drain by-passing less critical; automatically combines the power outputs of two devices.
Disadvantages: Requires matched devices; creating the required 180° phase shift becomes more critical at increasing frequencies, making the configuration impractical at high frequencies such as microwaves.

the balanced circuit or unequal loads on the transistors in the output.

In the input of a push-pull amplifier, a transformer with a floating (or physically non-existent) centre tap provides a much more balanced drive to the two transistor inputs. The return ground path for the 'on' transistor, with a floating transformer, is created by the input capacitance of the 'off' unit.

Assuming the input capacitances of both devices are equal, and since the rf voltage amplitude across the whole winding is twice that from one side to a centre tap, amplitudes to both the on and off transistors are equal in each case. No change in the input return loss should occur either.

The same conditions exist in the output, except we do not rely on the output capacitance of the off transistor, which is at the power supply voltage potential (dc). The on unit is at ground so a voltage close to the dc supply is always superimposed by the rf voltage swing across the output transformer primary.

Peak voltages as high as five times the dc supply across the transformer winding are common. This voltage (collector-collector or drain-drain) is twice the peak rf voltage from the collector/drain of a single device to ground, representing a 4:1 difference in impedance.

Symmetry is more important in the output matching of a push-pull circuit than in the input matching. In addition to the well known suppression of even harmonics, the balance affects the amplifier's stability, efficiency, and susceptibility to mismatched loads. One of the best ways to reach a good balanced condition in wide-band transistor output matching is with a separate collector/drain structure.

Push-pull circuits with only lumped constant elements are not really feasible because creation of the exact 180° phase shift becomes too critical and every unit would have to be individually adjusted in production. A hybrid design (**Fig. 3a**) is a much better choice. Initial matching to an intermediate impedance is achieved with *LC* networks (as in Fig. 2a) while the 180° phase shift is produced with simple and reliable 4:1 and 1:4 transmission line transformers.

The intermediate matching networks can also be microstrips, such designs being common in uhf amplifiers.

Another possibility is to bring the impedances of each device directly to 50Ω, so that a 1:1 balun could provide the phase shift.

Figure 3b shows an amplifier circuit best suited to low frequency applications up to 50-100MHz. The upper frequency limit is determined by the types of transformers used. For conventional types, the upper frequency limit is usually 30-50MHz, though some conventional rf transformers will perform up to 200-300MHz.

Both circuits shown in Figs. 3a and 3b are for class C, though with proper base forward biasing they can be converted to linear amplifiers (class A or AB).

Transmission line transformers used in the circuit of Fig. 3b would extend the bandwidth. But the circuit would become fairly complex because impedance ratios such as 16:1 and 25:1 would be required – especially in high power and low voltage applications.

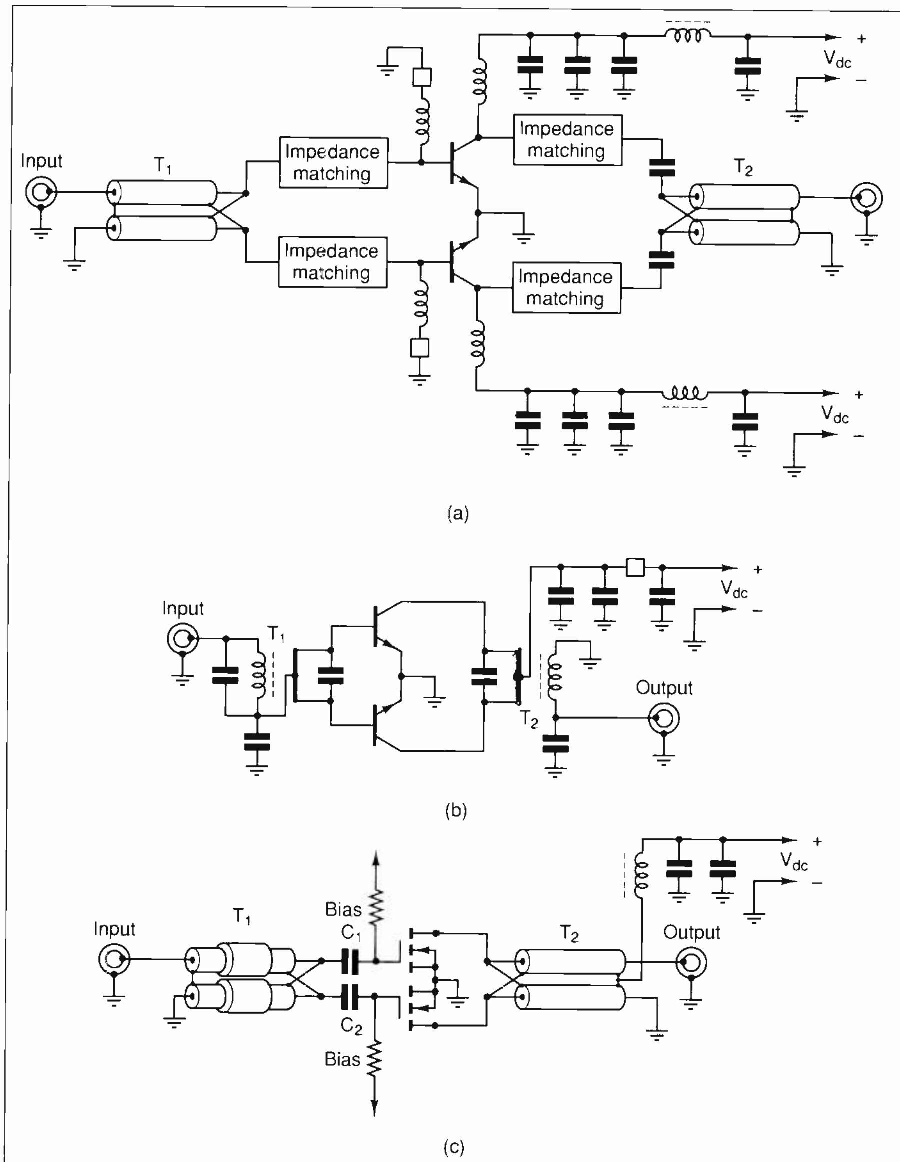


Fig. 3. Push-pull circuit configurations.
(a) shows how part of the impedance matching is achieved with individual matching networks such as microstrips. Transformers are used for further impedance matching and generation of a 180° phase shift.
(b) represents a straight-forward push-pull circuit using conventional transformers to provide the phase shift and input/output impedance matching.
in (c), 4:1 and 1:4 transmission line transformers are used to provide the functions above. With the impedance ratios shown, the circuit is useful for operation from high supply voltages where the output impedance is relatively high.

In a typical push-pull amplifier designed with mosfets (**Fig. 3c**), since their impedances are higher than those of bjts in general (at least up to uhf), impedance matching is easier. The configuration shown is directly adaptable for a 200W vhf amplifier, although the capacitors compensating for leakage inductance in the transformer have been omitted. Even without low frequency operation, transformer T_1 should be loaded with suitable magnetic material to provide the isolation necessary between the fet gates.

h_{FE}/g_{FS} should be matched in all push-pull circuits, and in mosfet circuits all the $V_{gs}(th)$s must be matched too if biased from a single voltage source. For devices with g_{FS}s of 4-6mhos, a difference of 50mV is acceptable. Since the drain idle current is directly related to g_{FS} vs $V_{gs}(th)$, matching becomes less critical with lower power devices (30-40W) where values of g_{FS} are in the 1mho range. In these cases differences of 100-150mV between the $V_{gs}(th)$s can be tolerated.

Impedances and matching networks

Many designers of rf equipment, used to vacuum tubes or solid state small signal circuits, are not familiar with solid state rf power designs and the importance of various

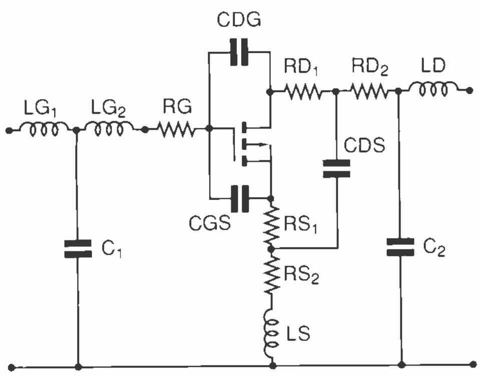


Fig. 4. A typical model of a mosfet that could be used for Spice parameter extraction. The model would be very similar for bipolar transistors. Most of the values must be obtained from the transistor manufacturer and must be derived according to geometry and process profile.

LG1 – Package lead inductance.
LG2 – Wire bound inductance.
C1 – Package lead capacitance.
RG – Poly gate sheet resistance plus gate metal runners to bond pads.
RS1 – Metal finger and material sheet resistance.
RS2 – Wire bond pad and metal runner resistance.
CGS – Source plus channel capacitance under poly gate.
CDG – Gate metal to drain MOS capacitance.
CDS – Drain – source and intrinsic diode capacitance.
LS – Source wire bond and lead inductance.
RD1 – Resistance of epitaxial layer.
RD2 – Resistance of substrate material.
C2 – Package and lead capacitance.
LD – Package lead inductance.

factors in developing suitable hardware.

At rf power, design guidelines are far more critical than those for low power and small signal design. The same rules apply in each case, but the physical layout of rf power circuits is much more sensitive because of the low input and output impedance levels involved. The important factors are device-impedance dependence on operating frequency, voltage, and power level.

As a rule, for a given voltage of operation and power level, the normalised input impedances and output impedances of unmatched devices are divided by a factor of approximately two with every octave of increasing frequency.

But inductive reactances increase at the same rate, making impedance matching much more complex at the higher frequencies. The effect applies to both bjts and fets, except that the input impedance of the fet is higher by an order of magnitude: at very low frequencies it approaches infinity since the gate represents only a pure capacitance.

On the other hand, a bipolar transistor has a base-emitter diode junction which must be forward biased to turn the transistor on. So the base input impedance, even at low frequencies, depends on the conduction angle and base forward bias.

When rf transistors are characterised by the manufacturer – with impedance values measured along with other parameters such as power gain, linearity and efficiency – the unit is usually inserted into an optimised test circuit (see box characterising rf values in practice). Various methods can be used to determine transistor impedances, but the indirect method is by far the most common.

In some instances, especially at uhf and microwaves, load pull contours are plotted on a Smith chart to indicate the device's behaviour at multiple frequency points. All Smith chart data of impedances given in the data sheets are in serial form but it is often advantageous (especially for low frequency designs) to convert it to parallel form to determine the actual resistive and reactive components. This is carried out with formulas:

$$R_p = R_s[1 + (X_s/R_s)^2] \text{ and } X_p = R_p/(X_s/R_s)$$

where R is the resistive component and X is the reactive one. If X is not very large compared to R, in most cases a fairly accurate composite impedance (Z) can be obtained:

$$(R_s^2 + X_s^2)^{0.5} \text{ or } (R_p^2 + X_p^2)^{0.5}$$

S-parameters are standard with small-signal class A devices, and sometimes with class A power devices. But power devices are seldom characterised with *S*-parameters because most experts question their accuracy and usefulness under large signal conditions, except for class A and stability calculations.

Spice parameter modelling – a newer approach to describing rf transistor behaviour using a model suitable for use with computer aided design (cad) programs – is claimed to give more accurate results.

Again, this is more likely to be true for linear operation rather than for non-linear operation. The Gummel-Poon model of a bipolar transistor (used in Berkeley *Spice*) is a linear model and would not be applicable to large signal, non-linear bipolar transistors. It also does not include package parasitics. A macromodel can be created for high power non-linear parts but the problem is determining a model that would apply for more than one set of operating conditions.

A mosfet model (**Fig. 4**) draws together data involving die parameters, package stray inductances and capacitances and wire bond inductances. Data are generated by the device die designers working with applications engineers who characterise the device.

Building the model is time consuming, and its accuracy in multiple applications is questionable – explaining why such data are not presently included in most device data sheets.

For the output, impedance levels are more or less dictated by supply voltage and power output. Output impedances with each type of device are capacitive at lower frequencies, but turn inductive when the wire bond inductances become dominant – determined by the device's output capacitance.

Output impedance matching into 50Ω is usually easier than input impedance matching due to its usually higher level. It also remains capacitive up to higher frequencies than input impedance. At low frequencies the output impedance can be determined with a fair accuracy as

$$(V_{CC}-V_{SAT})^2/P_{OUT} \qquad \text{(bipolars)}$$

or

$$[V_{DD}-V_{DS}(ON)]^2/P_{OUT} \qquad \text{(fets)}$$

But beyond 100MHz or so – depending on the device's electrical size – the complex impedance values must be taken into account. The nature of the output impedance and its matching is more critical than the input impedance since it also determines the overall efficiency of operation. Input matching only relates to the input return loss.

One of the problems facing a circuit designer is design of high frequency matching networks. Developing networks that will accomplish the required matching, harmonic

Characterising rf values in practice

Connections to the transistor are made with as short lead lengths as possible and clamping is used for temporary device mounting, allowing a number of units to be tested with easy insertion and removal.

The test circuit should have the necessary elements for fine adjustment of the input return loss to its maximum and power output to a specified level, while maintaining desired efficiency. After the test and adjustments, the transistor is removed from the test circuit and resistive terminations are connected to the circuit's input and output. A special probe consisting of a lead frame of the same type as the transistor is clamped in the circuit in place of the transistor. Connections are made with short lengths of precision coaxial cable – such as the semi-rigid type – from the probe to a network analyser. The numbers obtained from the analyser are the conjugates of those given in data sheets.

Several devices are usually tested and measured to ensure consistency of parameters. Since the values measured may be extremely low, errors in the form of stray inductances may limit accuracy to about ±20% and in most cases are only guidelines for a designer.

Practical design example

Assume an amplifier with an output power, P_{out}, of 125W at 100MHz is to be designed. Supply voltage is 28V, and the power gain required is 40dB. Browsing through various device data books shows that the 40dB gain requirement at 100MHz can be met with two stages if mosfets are used. For example, the *MRF174* has an indicated power gain of 14dB at a frequency of 100MHz, meaning that 5W are required to drive it to a P_{out} of 125W.

As a driver, the *MRF134* will do nicely with its 27dB power gain at 100MHz. So these two stages should satisfy the 40dB gain requirement.

Next, we need to select the matching network configuration. From the data sheets we find that the *MRF134* output impedance is 20.1–j46.7Ω (P_{out}=5W) and input impedance of the *MRF174* is 1.33–j2.98Ω at 100MHz. Converted to parallel form these are 130–j55 (Z_p=140) and 8.0–j3.6 (Z_p=8.8)Ω respectively.

Output impedance of the driver is relatively high so the most suitable networks are **Figs. 5b** and **5e**. Figure 5b may prove to be the more versatile since C_1 and C_2 can both be made variable elements, whereas low value *Ls* are difficult to make adjustable except with a very limited range.

The Q is defined as R_1/X_{C1} and since R_1=130Ω, Capacitor C_1 would need to become extremely small if the Q is to be kept low.

On the other hand, a high Q means instability problems may be encountered and the bandwidth narrows.

A Q of 5 will give X_{C1} as 26Ω or C_1 as 60pF. As established earlier, X_{Cout}=55Ω (28pF), R_L=8.8Ω. Then:

$$X_{C2} = 8.8\sqrt{\frac{130/8.8}{(5^2+1)-(130/8.8)}}$$

=10.1Ω, or 158pF,

and

$$X_{L1} = \frac{5\times130+(130\times8.8/10.1)}{26}$$

=29.4Ω, or 47nH.

We now have the values for all three elements as

$$C_1(\text{actual}) = C_1 - C_{out}$$

=32pF.

Capacitor C_2 is 158pF and inductor L_1 is 47nH.

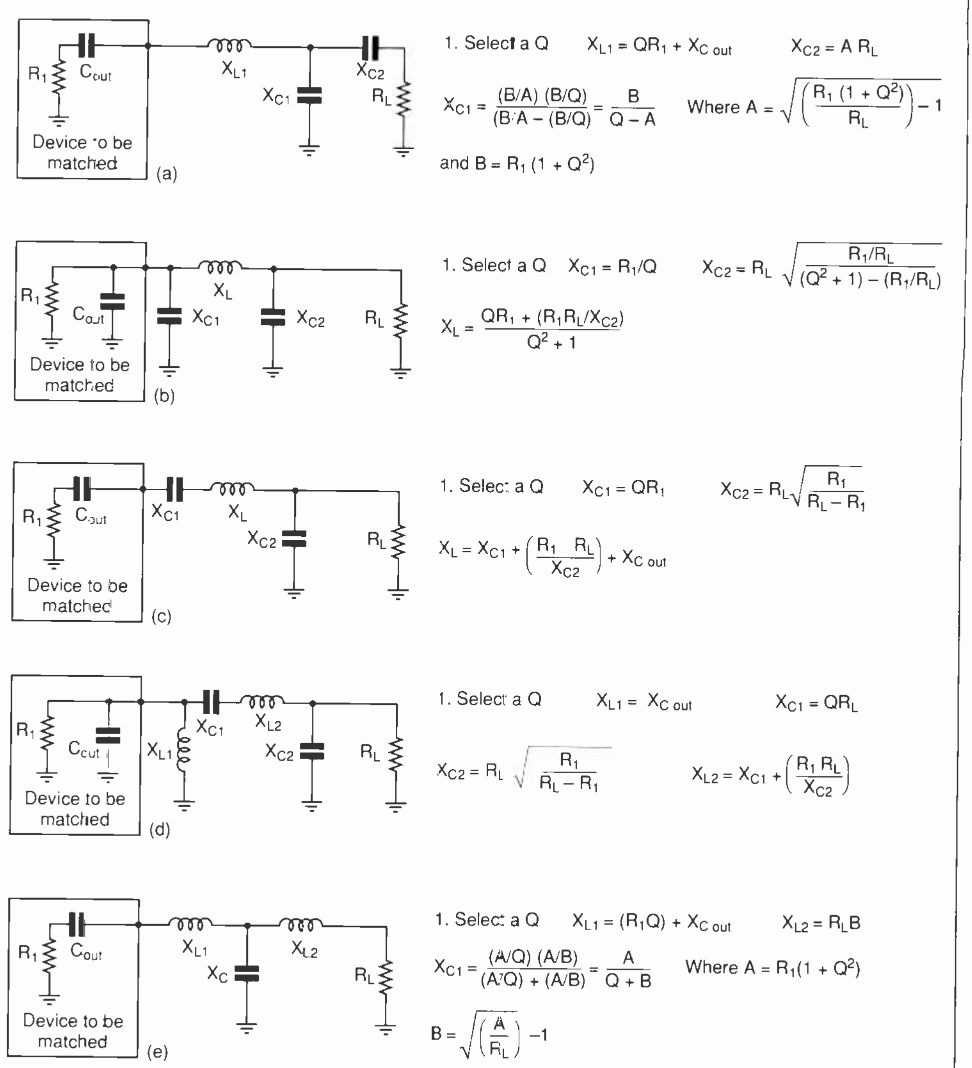


Fig. 5. Narrow band LC matching networks applicable up to uhf. Although shown for output matching only, all configurations can be reversed. They can also be used for interstage matching depending in the impedances involved. In certain cases, the component values may be physically or electrically impracticable.
Network (a) is applicable only when R_1 is less than 50Ω. When R_1 approaches 50Ω, the reactance of C_1 approaches infinity.
Network (b) is the P_i network widely used for matching at higher impedance levels and may be impracticable where R_1 is small. For values of R_1 less than 50Ω, the inductance of L becomes impracticably small while the capacitances of both C_1 and C_2 become very large.
Networks (c) and (d) have very similar characteristics. In both, R_1 must be less than 50Ω. But these configurations often yield the most practical component values where low values of R_1 must be matched.
Network (e) is a "T" network, useful in matching impedances of greater than 50Ω and, is especially applicable to small signal circuit design. Laboratory tests show that this network yields very high efficiencies when used for output matching in transistor rf power amplifiers.

suppression, bandwidth, using components having realisable values can take many hours of design time without a cad capability. Design of matching networks involves an infinite number of possibilities, and tabulation of all possible network solutions would be virtually impossible. But a small number of commonly used matching networks (**Fig. 5**) can be used for matching in transistor rf power amplifier circuits, with a wide variety of source and load impedances.

All networks are reversible, but in general the R_L side is more suitable for matching into a higher impedance. They can also be used for interstage matching, but with modifications since R_L becomes a complex function. Another approach is to use the intermediate impedance values discussed earlier.

Finally, the component values must be examined to see if they remain within practical limits. R_1 and C_{out} represent the complex input or output impedance of a transistor. They are shown in series form for **Fig. 5(a)**, **(c)** and **(e)** and in parallel for **(b)** and **(d)**. But each network can be converted to either form for a particular application. ■

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DESIGN BRIEF

Differentiating op-amps

Demonstrating how different op-amps can be, Ian Hickman investigates two high-performance devices – one optimised for DC and low frequencies, the other a high-speed device with a 1.7GHz gain-bandwidth product.

In an ideal op-amp, bandwidth, open-loop gain, differential and common mode input impedance, CMRR and PSRR would all be infinite. Input offset voltage, input bias current and output impedance on the other hand would all be zero. In practice however, designers have to select from a wide variety of op-amps all with their parameters optimised differently

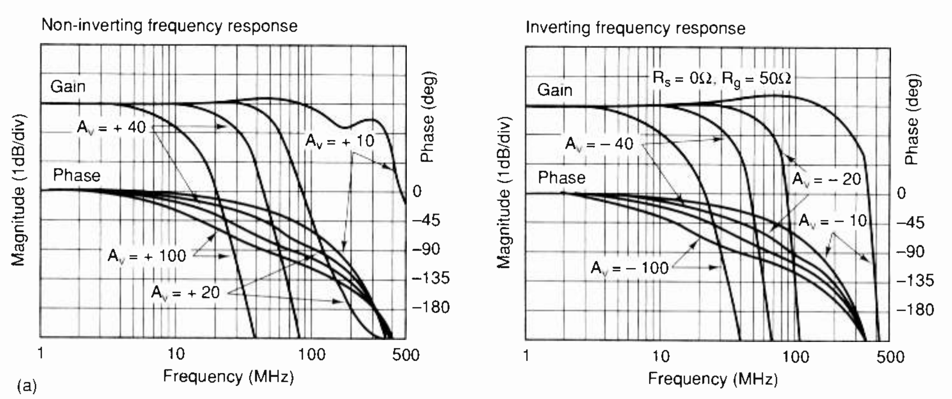
The following contrasts two bipolar op-amps. Both have input-offset voltage temperature coefficients in the microvolt range. Otherwise they are dissimilar and suited to very different tasks.

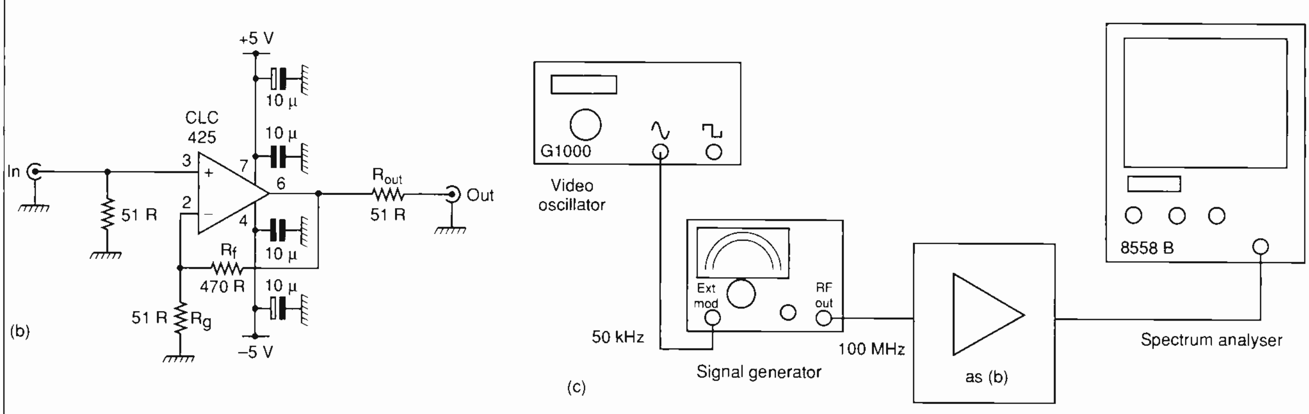
Op-amp excels at rf

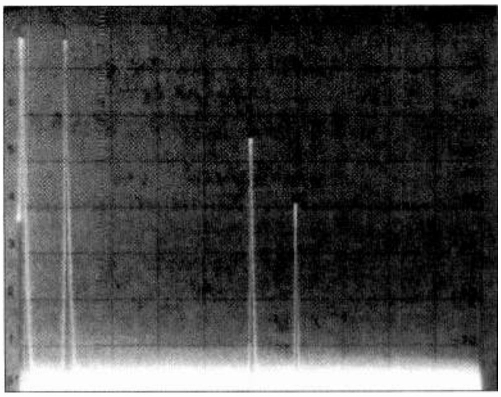
Features of the Comlinear *CLC425* op-amp are a 1.7GHz gain bandwidth product, a typical input offset voltage of 100μV, and a 360V/μs slew rate. It also has a low input noise of 1.05nV/√Hz, 1.6pA/√Hz, although its 1/*f* corner frequency is around 500Hz.

Among the uses of the device are ultrasound and

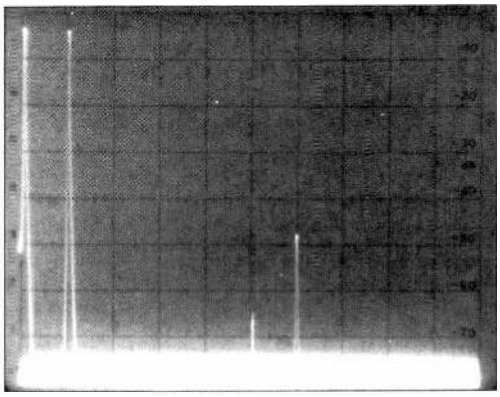
Fig. 1. Gain versus frequency (a) for the CLC425 in inverting and non-inverting modes. In (b) is the circuit used for evaluating of the CLC425 low-noise wideband bipolar op-amp. The device produces a voltage gain of 10, but as the input is terminated with 51Ω and a 51Ω source resistance Rout is included in series with the output, insertion gain is 5 or just 14dB. Diagram (c) shows the test set-up.







(a)



(b)

Fig 2. Right-hand trace in (a) is a 100MHz –40dBm test signal applied directly to a spectrum analyser. Horizontal scaling is 20MHz/div., 0Hz is one div in from left-hand side. Vertical scaling is 10dB/div. referred level 0dBm, IF bandwidth 300kHz. No video filtering is used. The left-hand trace is as above but the signal is fed via the amplifier board, demonstrating an insertion gain of just over 14dB. In the right-hand trace of (b), parameters are as above but amplifier ±5V supplies are turned off and the external attenuator set to 0dB instead of 30dB. This shows 38dB of powered-down isolation. In the left-hand trace, an external attenuator is set to 0dB instead of 30dB with the amplifier powered up but input and output ports interchanged. This shows 55dB reverse isolation and a 70dB ratio of forward to reverse gain.

instrumentation sense amplifiers and magnetic tape and disk pre-amplifiers. Applications extend to rf instrumentation since the device has a non-inverting gain of ten, ±1dB, up to 350MHz, **Fig. 1a**. In fact, while its ac performance is reminiscent of a current feedback op-amp, it works with voltage feedback, providing vastly improved dc characteristics. These include a 2μV/°C input offset-voltage temperature coefficient.

This amplifier is a 'decompensated' type. While not *uncompensated*, it is not as heavily internally compensated as would be necessary for use at closed loop gains down to unity. The minimum gain recommended for stability is ten.

I mounted a sample of the device on the company's evaluation board. This is a well laid out, plated-through board measuring 4cm square. It is designed to take PC mounting SMA sockets and leaded components. However, I used 10nF chip decoupling capacitors in parallel with tantalum types as these could be mounted closer to the device pins.

Resistors R_f and R_g were 470Ω and 51Ω respectively, Fig. **1b**, setting a demanded gain of 521/51, which is 20.2dB. This gives an insertion gain, board in to board out, of 6dB less than this. Figure **1c** shows the test set-up used for subsequent measurements.

Figure 2a is a double exposure, the right hand trace showing a –40dBm 100MHz test signal applied directly to a spectrum analyser. The corresponding zero hertz marker is one division in from the left-hand side. In the second trace, which is offset one division to the left, the same signal is applied via the amplifier on the evaluation board indicating a gain of a shade over 14dB. Zero hertz is at extreme left-hand side of the display.

Figure **2b** shows another double exposure. On the right is output from the amplifier when the +5V and –5V supplies were turned off, the external attenuator having been set to 0dB instead of 30dB. There is 38dB of isolation at 100MHz through the amplifier when powered down.

Even more interesting is the left-hand trace. Here the amplifier is powered up but has had its input and output ports interchanged. Again the external attenuator was set to 0dB instead of 30dB. This indicates a reverse isolation of about 55dB and a ratio of forward to reverse gain of around 70dB.

An obvious application for the *CLC425*, with its low noise and flat gain versus frequency, is as a preamplifier to extend the input sensitivity of a spectrum analyser. Spectrum analysers are designed to cope linearly with a welter of frequencies at their input, so as to display them all faithfully while adding the minimum of additional spurious signals due to intermodulation products. Consequently they usually employ a straight-into-the-first-mixer architecture which, while maximising linearity, results in a noise figure in the range 20 to 25dB.

When a signal applied to the analyser is known to be free from large unwanted signals, the sensitivity of the analyser can be extended by adding a low noise preamplifier at the input. To evaluate the effect of adding the circuit shown in Fig. 1, a low deviation fm test signal was applied. The 100MHz test signal was frequency modulated with a 50kHz sinewave, with a modulation index of 0.14. This gave a peak frequency deviation of 8kHz and a peak phase deviation of only 8°. At this low modulation index, the second-order fm sidebands of the signal are almost 50dB down on the

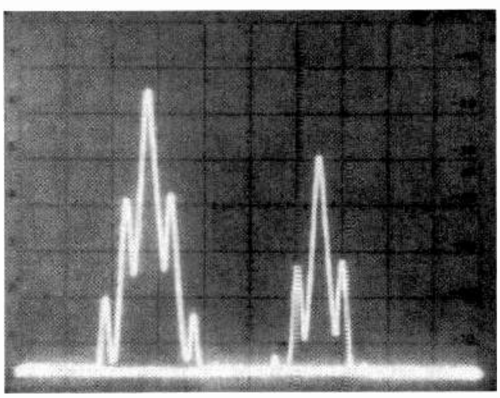


Fig. 3. At not much over –110dBm in the right-hand trace, second order fm sidebands of the 100MHz –60dBm test signal (see text) are barely visible above noise. Horizontal scaling is 100kHz/div., vertical is 10dB/div referred to –30dBm. Video filtering is set to maximum and IF bandwidth is 10kHz. A 50kHz modulating frequency is applied. In the left-hand trace, via the 14dB gain of the CLC425 amplifier board, second FM sidebands are clearly visible and their level readily measured.

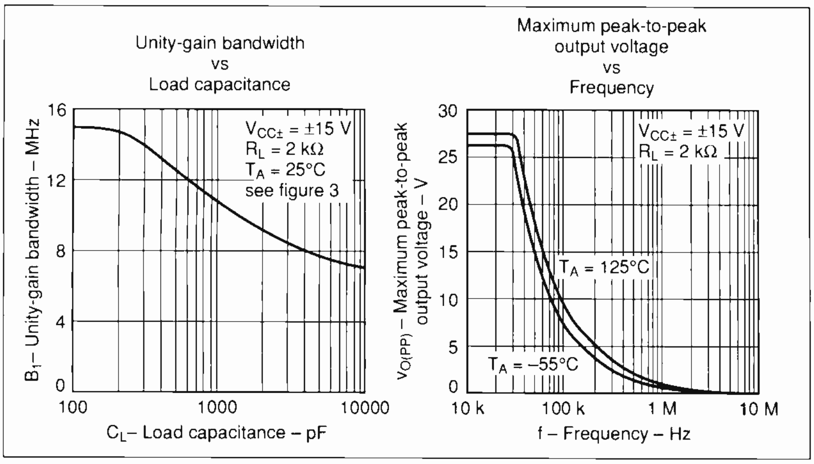


Fig. 4. Unity-gain bandwidth of the TLE2027 versus capacitive load, (a), and full power bandwidth (b) showing that the device covers the audio band and beyond.

carrier, the amplitude of which is virtually unchanged from the unmodulated condition. At not much over –110dBm, the second fm sidebands in **Fig. 3**, right-hand trace, are barely visible above the analyser's noise floor. Note also that there is smoothing of the noise floor by the video filter, which takes it almost half a division above the graticule base-line.

Insertion of the amplifier board between the external attenuator and the analyser's input, left-hand trace, rescues the signal and makes measurement easy. A bespoke design using discrete transistors could doubtless achieve an even lower noise figure. And the analyser's noise figure is so high that a second preamplifier stage similar to Fig. 1 could be usefully cascaded. But for an amplifier with 50Ω input and output, high reverse isolation and a small-signal flat response from dc up to a few hundred megahertz, the circuit shown provides an instant solution.

Operating from ±5V supply rails, the device's maximum output swing for a THD of less than 1% is maintained at over 7V peak-to-peak up to 10MHz.

Settling time to 0.1% is a shade over 20ns.

Penalties for enhanced dc performance

The second op-amp in question is the Texas *TLE2027*, which has a higher-specification counterpart, the *TLE2027A*. Made using the company's *Excalibur* technology, this device contains no fewer than 62 transistors – all bipolar devices except for one fet.

Internally, this device is compensated for gains down to unity .It features an open-loop gain of 153dB (45V/μV) which, with its 15MHz unity gain frequency, corresponds to an open loop gain roll-off starting at below 1Hz. Contrast this with a gain roll-off in the other device starting at around 100kHz.

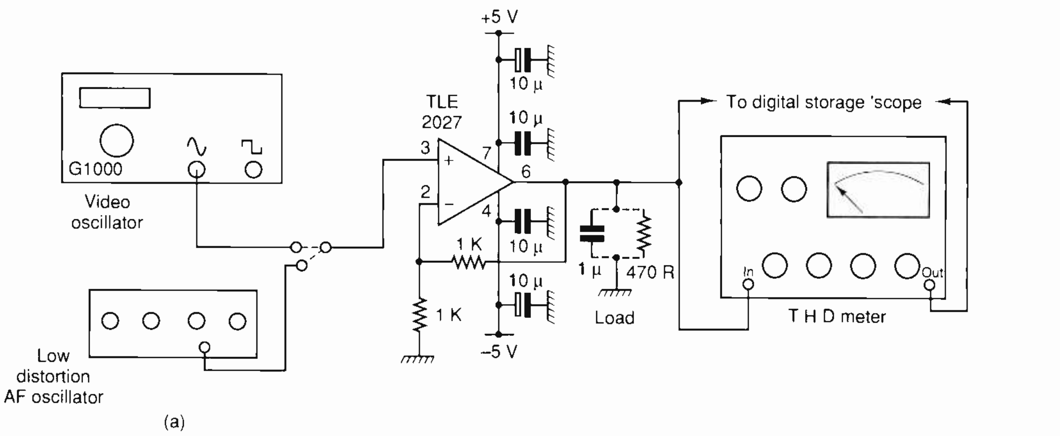
Direct-current characteristics are excellent, the typical input offset, temperature coefficient and drift being respectively 10μV, 0.4μV/°C and 0.006μV/month for the premium 'A' version. As far as ac characteristics are concerned, the *TLE2027* is clearly aimed more at audio frequencies.

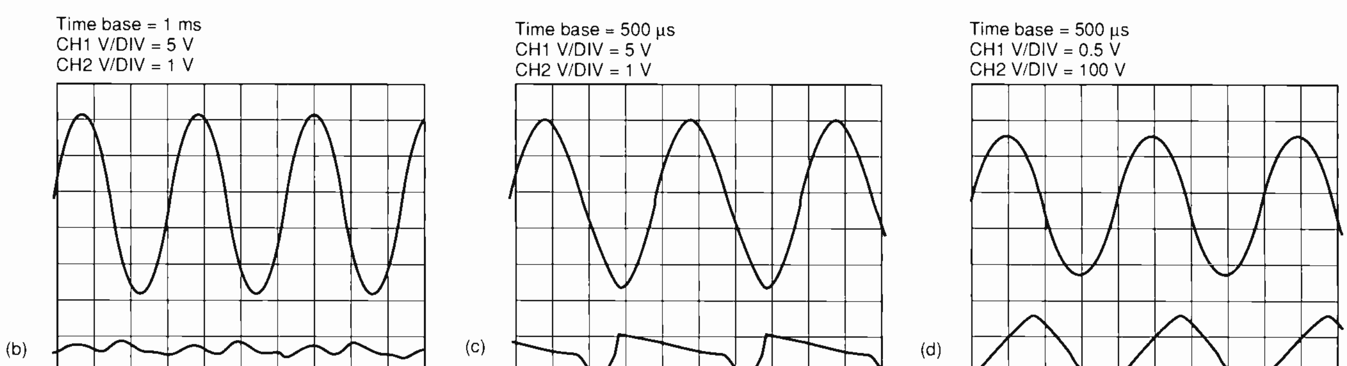
For a source resistance of 50Ω, equivalent input-voltage noise of the *CLC425* may be reduced by increasing the power consumption of the IC slightly, to below 1nV/√Hz (the device has a bias pin). With a 100Ω source resistance, noise of the *TLE2027* is 2.5nV/√Hz. However, with its 1/*f* corner frequency of 500Hz, the *CLC425*'s noise has already risen to 8nV/√Hz at 100Hz as against only 3.3nV/√Hz at 10Hz for the *TLE2027*.

An unusual feature of the *TLE2027* is its wide phase margin, enabling it to tolerate large capacitive loads, **Fig. 4a**. Reactance of 10nF at 7MHz is a mere –j2.3Ω, and the device's internal current limit is of the order of ±35mA! In fact with ±15V supplies, the maximum output voltage of a little over 25V peak to peak into a load resistance of 2kΩ is maintained over the full audio band and beyond, to 30kHz – Fig. **4b**.

With capacitive loading, the peak-to-peak swing

Fig 5. Circuit used to test the TLE2027 driving capacitive loads (a). In the upper trace of scan (b), the device is driving 23V pk-pk into 1μF at 318Hz. Lower trace shows THD is 0.06% – but due to what? In (c), upper trace, the op-amp is trying to drive 23V pk-pk into 1μF at 500Hz. In the lower trace, is the effect of the op-amp's internal current limiting circuitry. Upper trace, (d) is the TLE2027 driving 23V pk-pk into a 380Ω resistor at 500Hz. Even at 0.005%, lower trace, residual is still mainly fundamental.





available will clearly depend upon the load reactance, i.e. upon both the size of the capacitance and the frequency of operation. I was particularly interested in the device's ability to drive large capacitances, which turned out to be amazing, when a year or so ago I designed an *RCL* bridge.

The design was based on the transformer-ratio-arm principle but the circuit was realised with op-amps in place of transformers. This principle has a unique advantage over any other parameter measurement technique. The arrangement needed an op-amp that would drive capacitive loads of 1μF, and even up to 10μF albeit with a small voltage swing.

Most op-amps I looked at would only cope with up to 100pF loading. Even when driving that capacitance, some need a series 47Ω resistor between the output and load. But with Fig. 4a showing the *TLE2027* driving up to 10nF, this device proved suitable.

Figure 5a, upper trace, shows the *TLE2027* driving 23V pk-pk into 1μF at 318Hz – twice 10^3 rad/s – in the circuit of Fig. **5b**. The lower trace indicates residual total harmonic distortion, which visibly contains both second and third harmonic components. It amounted to 0.07%.

Frequency was raised to 500Hz while leaving amplitude unchanged. At this frequency impedance of the 1μF capacitor falls to 320Ω, Fig. **5c**, upper trace. This resulted in the op-amp current limiting over the negative-going flank of the sinewave. Current-sink limiting of the device is 35mA against a 40mA maximum current source capability.

Current limit does not occur at the peak voltage, as would be the case with a 320Ω resistive load, but over the part of the waveform where the slope dV/dT is greatest. This is due to the current drawn by a capacitor being 90° phase advanced.

In the lower trace of Fig. 5c, the THD meter residual has been adjusted to show that response is perfectly linear over that part of the cycle where the op-amp is not in current limit. This residual indicated 10% distortion, but clearly the waveform has a component at the fundamental. When this was completely nulled out, the true distortion was measured as 6.3%.

I repeated the 500Hz measurement with the capacitance replaced with a 470Ω resistor. In this case, current limiting was not quite reached. Residual distortion is not shown as it was identical to that in Fig. 5b. It was not only identical in amplitude and waveform shape, but also in *phasing* – a fact that started alarm bells ringing, since the phase of the load current was different.

The video generator supplying the drive – the same one used to provide the 50kHz modulation in Fig. 3 – was replaced with an oscillator whose THD at 500Hz is less than 0.0005%. Voltage across the resistor is shown in Fig. **5d**, upper trace. The 'residual' shown in the lower trace was 0.005%. This residual is largely fundamental.

There was insufficient resolution on the THD meter's wire-wound phase and quadrature trimmers to achieve a complete null of the fundamental. This suggests that distortion in the *TLE2027* op-amp is little if any greater than the residual distortion of the distortion meter itself. From other tests, I know this to be about 0.0016%.

This is a remarkable performance for an op-amp supplying 23V pk-pk at a load current not much below the current limit value – given the 2kΩ gain setting chain at the non-inverting input in parallel with the 470Ω load. ■

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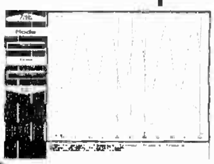

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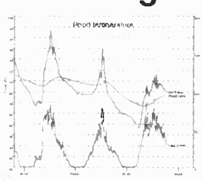

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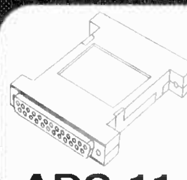

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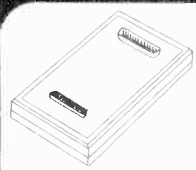

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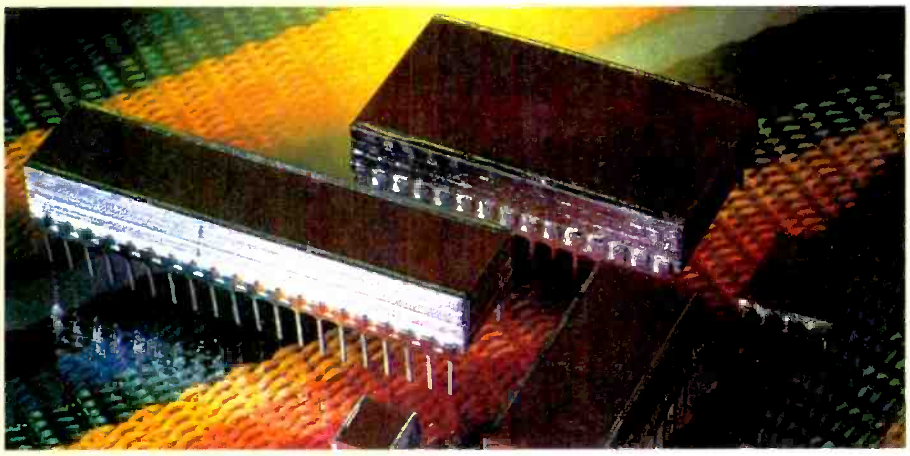
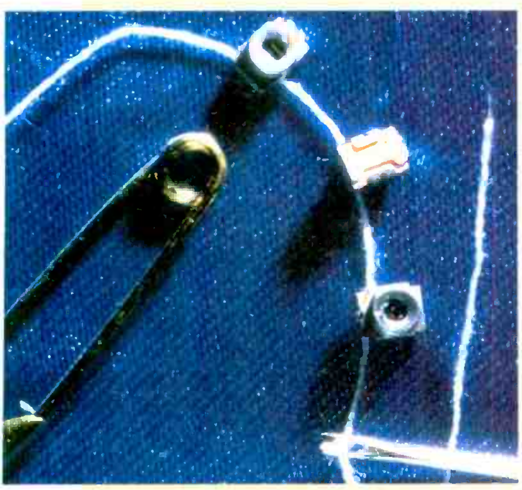


Ceramic capacitors deliver high values. These Murata low ESR, self-healing types, optimised for filtering in DC/DC converter and SMPS applications, have values to 100µF and voltage ratings of 50 or 100V depending on capacitance.

CHOOSING CAPACITORS

For some applications the choice of capacitor technology is irrelevant but in most cases there are benefits to taking a more careful look at dielectric characteristics. Tony Wong discusses features and disadvantages of the most popular dielectrics.



World's smallest trimmers. These devices, with values to 20pF, measure 2.3 by 3.2mm and are under 1.5mm high. Having a ceramic rather than plastic stator enhances long-term stability, according to Murata.

Capacitors – simply devices capable of storing electrical energy – are widely used throughout the electronics industry. There is a wide variety of different types available today, each with their own combination of features and drawbacks. This article presents a round-up of capacitor technologies looked at from a design-engineer's viewpoint.

Capacitors generally need to be as small as possible. Since most needs are for low-voltage devices, much of the emphasis in capacitor manufacture is on producing thin and uniform dielectric.

Capacitor technologies currently in widespread use fall into three categories, depending on their loss. Mica, glass, ceramic and polystyrene types form the first. These types exhibit low loss and good capacitance stability. Paper, plastic film and high-Q ceramic types, the second category, exhibit medium loss and medium stability.

Aluminium and tantalum electrolyte types have the best capacitance per unit volume characteristics. This class of capacitor involves extremely thin anodic oxide layers with a high dielectric constant.

Liquid electrolyte usually connects the oxide layer. Solid tantalum types however use manganese-dioxide semiconductor.

Electrolytic capacitors

These devices provide high capacitance in the smallest space at the lowest cost. Electrolytic capacitors are polarised. If connected incorrectly, the insulating oxide film is not formed and there is no capacitance. Reverse connection eventually causes overheating and then failure.

Electrolytic capacitors divide into two main categories – namely aluminium and tantalum.

Aluminium electrolytic capacitors. These capacitors can withstand a reverse voltage of up to 1.5V without noticeable effect on their operating characteristics. Excess voltages applied for short periods will cause some change in capacitance but will not lead to failure.

Aluminium electrolyte capacitors are made with a built-in pressure relief mechanism. This is designed to open and slowly release gas pressure that may build up if the device overheats during operation. A further disadvantage of aluminium electrolyte capacitors is their relatively high leakage current. This is cause by the oxide film not being a perfect insulator.

Tantalum capacitors. Tantalum capacitors tend to have as much as three times better capacitance/volume efficiency than aluminium electrolytic capacitors. They fall into the following three categories.

There are two main types of solid electrolyte tantalum capacitors in common use – hermetically sealed and epoxy dipped. Hermetically-sealed types have dc voltage ratings up to 125V. They are mainly used when low leakage current, low dissipation factor, reliability and stability with time and temperature are required. Epoxy-dipped tantalum capacitors have dc working ratings up to 50V. Applications are mainly in commercial and industrial equipment where low cost, small size, high stability, low dc leakage and dissipation factor are important.

Liquid-electrolyte tantalum capacitors, also called 'wet-slug' tantalum capacitors, have dc working ratings up to 150V. Their main application areas are industrial and military equip-

Table 1. EIA Class I dielectrics for ceramic capacitors – what the codes mean.

Significant figure		Multiplier		Tolerance	
code	value	code	value	code	value
C	0.0	0	–1	G	±30
M	1.0	1	–10	H	±60
P	1.5	2	–100	J	±120
R	2.2	3	–1000	K	±250
S	3.3	4	–10000	L	±500
T	4.7	5	+1	M	±1000
U	7.5	6	+10	N	±2500
		7	+100		
		8	+1000		
		9	+10000		

Example: for M7G dielectric is P100 ±30ppm/°C: for P3K dielectric is N1500 ±ppm/250°C.

Table 2. Code descriptions for EIA Class II dielectrics.

Lower limit °C		Upper limit °C		Maximum capacitance change Change, ΔC (%)	
code	value	code	value	code	value
X	–55	2	45	A	±10
Y	–30	4	65	B	±1.5
Z	10	5	85	C	±2.2
		6	105	D	±3.3
		7	125	E	±4.7
				F	±7.5
				P	±10
				R	±15
				S	±22
				T	22/–33
				U	22/–56
				V	22/–82

Example: for Z5U, ΔC is 22% to –56% from +10 to +85°C: for Y5V, ΔC is 22% to –82%, from –30 to +85°C.

Table 3. Characteristics of the three most common dielectrics used in ceramic capacitors.

	NP0/COG	X7U (mid K)	Z5U (high K)
Dissipation factor	<0.1%	<2.5%	4%
Temperature coefficient	<±30ppm/°C	0±0.15% ΔCmax	22/–56% ΔCmax
	–55 to 125°C	–55 to 125°C	10 to 85°C
Insulation resistance	100GΩ	100GΩ	10GΩ
Dielectric withstand	2.5xrated voltage	2.5xrated voltage	2.5xrated voltage

Tantalum chip capacitors with resin coatings, like these from Matsuo, are said to offer improved mechanical protection combined with enhanced moisture and solder heat resistance.

ment where small size, stable electrical characteristics and long service life are important.

Foil-based tantalum designs have the lowest capacitance per unit volume. They are also the least encountered, being best suited to the high voltages primarily found in older equipment. Some types are capable of handling dc working voltages up to 300V. Since this type of capacitor requires more manufacturing operations than the two other types, it is relatively expensive. The operating temperature range is –55°C to +125°C.

In summary, the circuit designer uses foil tantalum capacitors in circuits involving high voltages or where there may be a substantial reverse voltage.

Wet-slug tantalum capacitors generally exhibit the lowest dc leakage. However, the conventional silver-can design is not tolerant to reverse voltage. This type is of capacitor is commonly found in aerospace applications. It can withstand reverse voltage up to 3V and will operate under higher ripple currents and temperature up to 200°C.

Solid-electrolyte designs are least expensive. Often used in consumer equipment, they are physically small and capable of withstanding 15% of rated dc working voltage in reverse direction. They also exhibit good low temperature performance and are free from corrosive electrolytes.

Ceramic capacitors

Produced in discoidal and tubular form, ceramic capacitors have very high dielectric constant values. Disc ceramic types feature capacitance values up to 4.7μF and take up much less space than paper capacitors. For tubular ceramics, with values of 1 to 500pF, applications are the same as for mica equivalents but again they take up less board space.

Ceramic capacitors are often used for temperature compensation. The temperature coefficient for these components is classified into three most popular temperature characteristics. Differences between these characteristics and the technique of how to read the specification is discussed later.

Mica capacitors

Mica for capacitor use is expensive and values are normally limited to about 0.1μF. A modified form of mica capacitor is the silvered mica type. Such capacitors have a constant capacitance with change of temperature and are less likely to change in value with age. They generally have a high working voltage rating and perform well in high frequency circuits.

Paper capacitors

Disadvantages of paper capacitors include variation in capacitance with temperature change and a shorter service life relative to most other types. Paper capacitors are often used for medium capacitance values of approximately 0.001 to 1μF.

Polymer types

There are two types of polymer dielectric used for capacitors – polar and non-polar. Generally, non-polar types produce devices that are frequency independent and polar types *vice versa*.

Polystyrene film types fall into the non-polar category. They have excellent electrical characteristics. The polymer has an oriented crystalline structure, making it possible to manufacture the film in very thin and flexible strips. Solvents affect the film so sealed encapsulation is needed.

Power factor of polystyrene film capacitors is low over a wide frequency range but the electrodes degrade performance with higher values as frequency increases. In a typical example, at 1MHz, values between 10 and 100nF might have a power factor between 0.005 and 0.02 whereas at 1kHz, the power factor will be as low as 0.0003. A device under 1nF will typically rise from the same starting power factor at 1kHz to only 0.001 at 1MHz.

Two other types of polymer film dielectrics are in common use – polyethylene-terephthalate and polycarbonate. All round, these two are similar in performance but polycarbonate has a lower loss angle and smaller change of capacitance with temperature. It is also possible to manufacture polycarbonate in a thinner film so devices are generally smaller.

Plastic film capacitors with metallised electrodes have superseded metallised paper types. Capacitor versus volume ratios are roughly the same but film types offer better electrical characteristics and higher, more stable insulation resistance.

Electrical properties

In order to understand the characteristics of a capacitor, its electrical properties need to be considered.

Dissipation factor. This is the tangent of the angle by which the current lags relative to the 90° vector of voltage over the capacitor.

Normally, measurements are made at 25°C and 1kHz with a test voltage 1V rms or less. Power factor is the cosine of the phase angle between the voltage and current vector, cosθ. For low-loss dielectrics, tanδ and cosθ are approximately equal, and can be used to express dielectric loss. In a 'low-loss' capacitor, the dissipation has a small value.

Quality factor, *Q*. The *Q* factor is the reciprocal of the loss factor and is dependent on frequency. A near-perfect capacitor has a *Q* approaching infinity. At 25°C with a 1MHz test voltage of 0.1 to 3V rms, *Q* should be greater than 1000 for general purpose capacitors.

Rated voltage. The maximum working voltage of a capacitor is the sum of the DC voltage plus the AC peak voltage which may be applied continuously to its terminals. Operating a capacitor at a voltage lower than its maximum working value extend its life.

Surge voltage. There is a maximum safe voltage to which a capacitor can be subjected under any combination of circumstances over a short period of time. This is the dc surge voltage rating. Normally, testing for surge voltage involves applying a signal several volts above the rated working voltage. It is applied via a 1kΩ series resistor in repeated cycles of 0.5 minutes on and 5 minutes off.

Leakage current. A relatively small direct current flows through a capacitor when a voltage is impressed across it. Generally, the maximum leakage current should not exceed the value given by these equations for capacitors with ratings between 16 and 100V dc.

$I=0.01CV$ μA (max.) for $CV \le 100{,}000$μFV

$I=3\sqrt{(CV)}$ μA (max.) for $CV > 100{,}000$μFV

Insulation resistance. This is a measure of the ability of the charged capacitor to withstand leakage of dc current. For capacitors of less than 0.01μF the insulation resistance should normally exceed 100GΩ.

Ripple current. The sum of dc and ac current that may be applied to an electrolytic capacitor is termed ripple current. The capacitor should be able to withstand this ripple current at 120Hz up to 85°C. In general, the higher the ripple current, the shorter the life of the capacitor.

Equivalent series resistance. For the purposes of calculation, all internal series resistances of a capacitor are lumped into one to represent the termination losses and dissipation in the dielectric. Equivalent series resistance, *ESR*, can be obtained from:

$$ESR = \frac{\tan\delta}{2\pi fC}$$

where *f* is measurement frequency in hertz and *C* is measured capacitance in farads.

Temperature coefficient. For ceramic dielectric capacitors, temperature coefficient defines the deviation in capacitance that occurs over a given temperature range. Dielectrics are classified into three most popular types based on temperature coefficient. These are designated by the Electronics Industry Association (EIA). The coefficient of higher K (ferroelectric) dielectrics is most often expressed as percentage capacitance change versus temperature. For linear dielectrics, the temperature coefficient is expressed in ppm/°C.

Class I ceramic dielectric has a linear temperature coefficient. Linear dielectrics, i.e. those with a K of less than 150, are defined in EIA and MIL specifications. Class I capacitors are used in circuits requiring stability and low loss (high *Q*) over the full temperature range. Typical examples of Class I dielectric EIA designations and their equivalent material coefficient are:

EIA code	material coefficient
C0	NP0
P2	N150
P3	N1500
R2	N220
R3	N2200
S1	N030
S2	N330
S3	N3300
T2	N470
T3	N4700

EIA coding comprises three digits, the first two showing the temperature coefficient of the material and the last digit indicating the tolerance of that coefficient. Combinations of the three digits are shown in **Table 1**.

Dielectrics with a K of more than 500 fall into the Class II category. These exhibit a non-linear temperature coefficient and are less stable all round. Both capacitance and dissipation factor are time and frequency dependent. Applications of Class II capacitors are filtering, dc blockage, by-pass and coupling components in circuitry where stability is not an important criterion.

This class can be further divided into two groups. These are the 'mid-K' stable Class II dielectrics, which include X7R of the EIA standard, and 'high-K' dielectrics. The high-K dielectric exhibits high capacitance density and is best for applications where space is at a premium. Types such as the EIA Z5U type can be used as replacements for tantalum capacitors. **Table 2** lists EIA Class II dielectrics at a 25°C reference temperature.

Compact size and large capacitance characterizes ceramic capacitors in Class III. These are used for coupling and by-pass applications. Compared with Class II, Class III dielectric has lower performance. Capacitance tolerance is larger, for example ±20%, and dissipation factor is higher, at around ±5%. In addition, insulation resistance is lower at 1GΩ minimum for example.

Characteristics of the three most commonly used dielectrics are contrasted in **Table 3**. ■

Charge and dielectric

Charge on a capacitor, *Q*, is proportional to the applied voltage *V*, ie. $Q \propto V$ or $Q=CV$, where *C* is capacitance. For a single plate device capacitance is,

$$C = \frac{Q}{V} = K \times \frac{A}{t} (farad)$$

where K is relative permittivity, *A* is area of the electrodes and *t* is dielectric thickness.

For monolithic multilayer constructions, capacitance, *C*, becomes,

$$C = \frac{K \times A \times N}{t}$$

where *N* is the number of dielectric layers.

Commercial capacitors are generally classified according to their dielectric. Most common electrolytes are aluminium, tantalum, ceramic, mica, paper and air. Their dielectric constants, represented by K, are,

Dielectric	K
Air or vacuum	1
Paper	2-6
Mica	5.4-8.7
Aluminium oxide	8.4
Tantalum pentoxide	26
Ceramic	12-400,000

Dielectric constant is highly significant. Charge activity is performed inside the dielectric material, causing a polarization effect. This effect arises from four mechanisms – electronic displacement, ionic displacement, permanent dipole orientation, and space-charge polarization. As a result, capacitors experience a variation in dielectric constant, and hence capacitance, under differing environments and test conditions. Characteristics of different types of capacitors and their related dielectrics are discussed in the main text.

Further reading

Reference data for radio engineers, Howard W. Sams/ITT.

TIME VIA MODEM

With the aid of a modem, you can now access highly accurate time information via your PC. Head of the NPL's time and frequency services, John Chambers, describes the new service – Truetime.

The National Physical Laboratory maintains and distributes the national time scale. Known as UTC(NPL), this time scale is made using a hydrogen maser. It is held within 1µs of the scale of Coordinated Universal Time (UTC) which is calculated from many atomic clocks worldwide by the Bureau International des Poids et Mesures just outside Paris.

The organisation is responsible for the 60kHz standard time and frequency transmissions from the Rugby radio station MSF. This station is continuously monitored against UTC. The MSF signal serves the whole country and is used to control clocks in public places as well as in commercial and domestic premises. It has the added advantage of automatic change to and from summer time. More accurate methods of time measurement, using satellite systems and even the physical transfer of clocks, are available to specialist users.

Audible methods of distributing time referred to UTC(NPL) include the telephone *Timeline*, operated by BT. Since February 1990 the Greenwich time signal has been originated by the BBC. Several broadcast signals are capable of carrying time information in coded form. These include the Radio Data System (RDS) carried on vhf/fm, the lf radio-data system carried on 198kHz, teletext, digital audio broadcasting (DAB) and the MAC satellite systems.

As a service to computer users, NPL now offers *Truetime*, giving time and date via a dial-up modem. Anyone with a V22 modem calling into the service will receive, every UTC second, a 77-character message describing the next second, followed by a three-character seconds marker. Relative timing is indicated in the drawing. Part of the message is designed for direct display on a screen, and other parts are intended for interpretation by software in the receiving system as required.

The accuracy of this method of time dissemination is limited by the uncertainty of the propagation delays through the telephone network and through the receiving modem. Normally it will be better than 20ms within the UK. The specification for the data line is that devised and already in use in Austria, Italy and Sweden.

Whenever it is necessary to demonstrate that a clock is traceable to the national standard of time to within 20ms, *Truetime* can be used as it is directly related to UTC(NPL). The service is suitable for any system where a clock needs resetting when equipment is switched on, or following a power failure.

UTC, GMT, and other time scales

In the UK, legal time is Greenwich Mean Time (GMT), offset by one hour in summer to give British Summer Time (BST). Historically GMT was determined so that, averaged over the year, the sun crosses the Greenwich meridian (longitude zero) at noon. It is more formally known as Universal Time (UT), and it is now derived by mathematical conversion from Greenwich Mean Sidereal Time (GMST), which measures the earth's rotation with respect to the stars.

Since 1967 the SI (International System) second has been defined in terms of the frequency of a particular radiation of the Caesium-133 atom. There are no gear wheels connecting it to the solar system, so a time scale based on the SI second, known as International Atomic Time (TAI), gradually gets out of step with GMT.

From 1972 a time scale known as Coordinated Universal Time (UTC) has been used as the basis for all time signals. UTC combines the accuracy of the atomic second with most of the convenience of GMT. Seconds ticks in UTC are the same as those of TAI, but occasional extra 'leap' seconds are inserted in the UTC time scale in order to keep it within ±0.9s of GMT. There have so far been 18 leap seconds, all positive and another is imminent. The most recent was at 0100 BST on 1 July 1993 and the 19th will be at 0100 BST on 1 July, 1994.

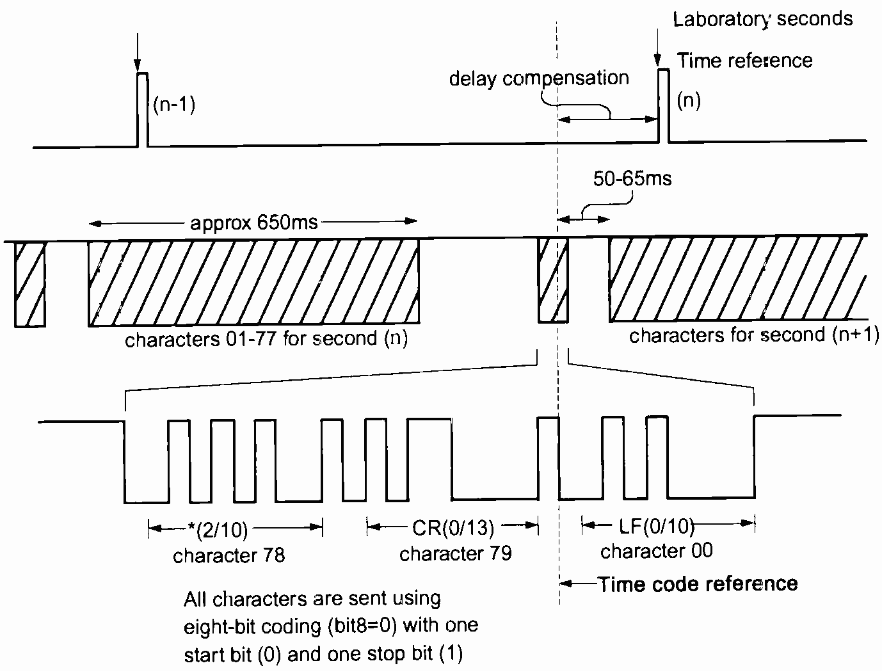
In GMT, BST, UT and TAI there are still always 60 seconds in every minute. In UTC there are 61 seconds in the minutes which contain positive leap seconds.

An automatic process could call the service and set the clock overnight. A regular call to the service will ensure that a clock in a critical application, such as logging financial transactions or in a security system, has not been altered. *Truetime* supplies advance notice of future summer/winter time changes and leap seconds. In applications where it is required to avoid the occasional 61-second minutes of UTC, the DUT1 code (carried by characters 55-56 in the data line) can be used to generate GMT/BST more closely.

To allow the date and time of two events to be subtracted to give a time difference, it is necessary to allow for the calendar, and for step changes of one hour and of one second. The modified Julian date, MJD, and UTC can be used to eliminate the first two of these problems, the last can be eliminated by knowledge of the cumulative number of leap seconds at each event. Although this information is not included in the data line, *Truetime* may include in the message sequence a six-character sequence *CKLSnn*, where nn is the cumulative known leap seconds (summed algebraically), modulo 100. This *CKLS* parameter would be updated with every new announcement of a leap second, and it relates to the situation after that leap second.

Truetime seconds markers and times are based on UTC. The first part of the message is intended to give British legal time and date. When the UK is on GMT, UTC+0 is sent and, when the UK is on BST, UTC+1 is sent. In the event of UK time being two hours ahead of GMT, UTC+2 would be sent.

The difference between UTC and GMT, or between UTC+1 and BST, is given to within one-tenth of a second by the DUT1 code, which is a signed integer in the range ±8. At



the start of 1994 the DUT1 code was +2, indicating that GMT was about 200ms ahead of UTC. To avoid the ambiguity due to the repeated hour 0100–0200 at the end of summer time the hours, minutes and seconds are separated by 'A' when describing the first hour and 'B' for the second. At all other times this separator is a colon.

The service uses a premium-rate telephone number to cover its costs, but all the information apart from the full message sequence can be obtained within the time allowed for the first unit of charge.

Anyone interested in the service is invited to write their name, address, telephone (and fax, if available) number on one side of postcard, together with brief details of their application. Send it to Time and Frequency Services (*Truetime*), National Physical Laboratory, Teddington, Middlesex TW11 0LW. We will then tell you how to call the service. *Truetime* is available 24 hours a day. ■

Further reading

Esson, L., *Leap seconds – story of the transfer from astronomical to atomic time, Wireless World*, July 1981.

CCIR Recommendation 460-4, *Standard-frequency and time- signal emissions, Geneva*, 1986.

Truetime data line format

Data is sent using the V22 1200 baud standard, with eight data bits, no parity, and one stop bit. Characters following each <carriage return> <line feed> are numbered 01-77 and listed below. Coding is ASCII. The seconds time reference is the leading edge of the start bit of the <line feed>, as shown in the diagram. Information in each line relates to the next following second.

01-10 date, changing at local midnight *YYYY-MM-DD*
11 space
12-19 local time in the format *hh:mm:ss* except when 'putting the clocks back' when the letters A and B are used instead of colons for the first and second repeated hour. Note that, in the case of a positive leap second, *ss* can take the value 60.
20 space
21-25 a name for the local time, currently 'UTC+0' during GMT and 'UTC+1' during BST
26 day of week, from 1=Monday to 7=Sunday
27-28 week of the year, from 01 to 52 or 53 (week 01 contains the first Thursday of the year)
29-31 day of year, from 001 to 365 or 366
32-37 month, day and hour (in local time) of next one-hour time change (*MMDDhh*)
38-49 year, month, day, hour and minute in UTC time scale (*YYYYMMDDhhmm*)
50-54 Modified Julian Date (MJD) – a five-digit decimal day count incrementing at 0000 UTC. As a current example, 1994 March 01 corresponds to MJD 49412
55-56 DUT1, the difference in tenths of a second between UT1 (in effect, GMT) and UTC in the range +8 through +0 to 8. A positive figure means GMT is 'ahead' of UTC.
57-59 a currently announced positive or negative leap second, at the end of the UTC month *MM*, is indicated by *+MM* or *MM*. 000 indicates 'no announcement'.
60-62 the number of milliseconds by which the next time reference is advanced with respect to UTC seconds markers.
63 message sequence number (see below) in the range 0-9
64-77 a 14-character field for a message line, up to ten different lines may be used in sequence, they are labelled by the message sequence number
transmission pause
78 this character is usually *, but it changes to # if the code is deliberately advanced in order to anticipate an estimated delay
79 <carriage return>
00 <line feed> - the leading edge of the start bit is the reference time for the second just described
transmission pause

CIRCUIT IDEAS

SEND YOUR CIRCUIT IDEAS TO THE EDITOR, ELECTRONICS WORLD, QUADRANT HOUSE, THE QUADRANT, SUTTON, SURREY SM2 5AS

DO YOU HAVE A £100 CIRCUIT? EACH MONTH'S TOP CIRCUIT IDEA AUTHOR WILL RECEIVE £100. ALL OTHER PUBLISHED IDEAS WILL BE WORTH £25. WE ARE LOOKING FOR INGENUITY AND ORIGINALITY IN THE USE OF MODERN COMPONENTS

Switched-gain amplifier minimises dc shift

£100 WINNER

Having its gain switched to one of two settings, this current-feedback amplifier presents excellent DC and high-frequency performance.

Bandwidth of current-feedback amplifiers is virtually independent of gain, but parasitics introduced by semiconductor gain-setting switches cause the bandwidths at different gains to be unequal; miniature relays can be used, but only at low speed. Here, an *EL2071* and an *EL2070* with gains of 20 and 2.8 are in parallel, their output disable pins ensuring that only one of them is in circuit at a time. Conveniently, the disable pins are complementary, so a single gain-control line switches amplifiers.

Since the DC performance of such amplifiers is poor, a number of circuit arrangements have been used to improve it. In this case, a dc amplifier is used as an error amplifier, comparing the input with part of the output. The output of the dc amplifier, which does not appear in the signal path, corrects offset in the current-feedback circuit. Correctly attenuated output signals are selected by the *DG419* switch and set by the potentiometers, the circuit arrangement being such that dc amplifier balance is unaffected by output load, which is nominally 50Ω.

The DC error amplifier corrects current amplifier offset by injecting more offset current into the inverting inputs and, since its swing must be 1V to 4V for the *EL2070* and –1V to –4V for the *EL2071*, diodes can be used to steer the correction to the relevant amplifier input. Adjust the circuit by injecting a 10kHz square wave at the input and setting the gain-balance presets for minimum AC at the error amplifier output.

Most bifet and bipolar amplifiers are suitable for the error amplifier, which determines drift; an *OP77* was used.

B Vojnovic and RA Orchard
Gray Laboratory
Mt Vernon Hospital
Northwood
Middlesex

Reference
Application note OA-7. Current-feedback Op-amp Applications Circuit Guide. Comlinear Corporation Data Book 1993-94 edition, pp 11.19-11.26.

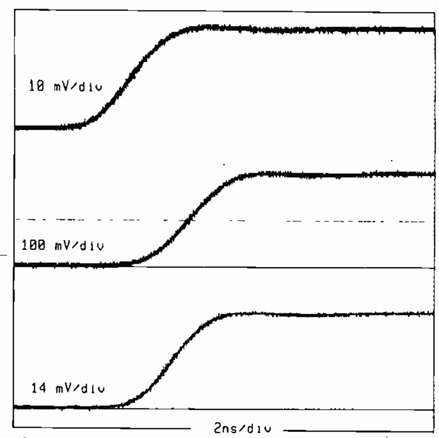
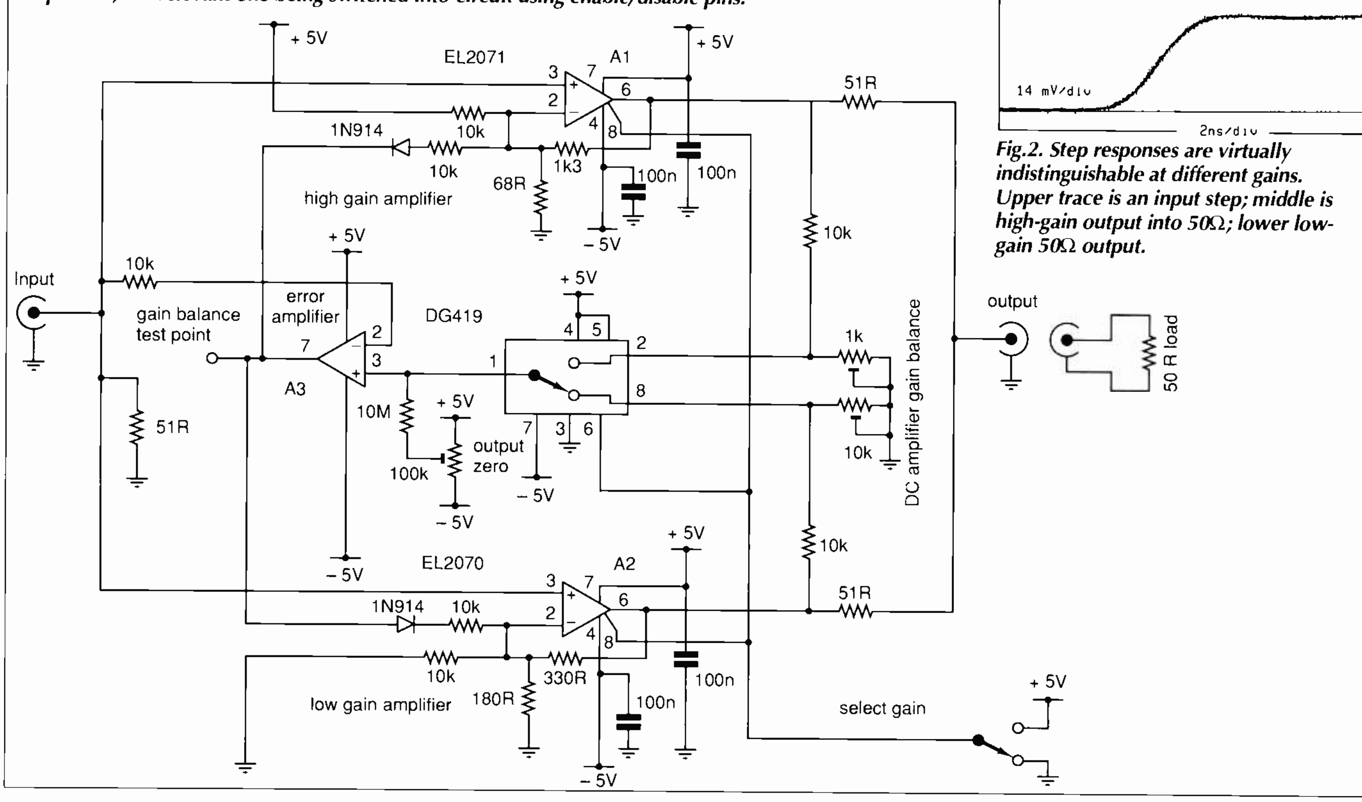


Fig.2. Step responses are virtually indistinguishable at different gains. Upper trace is an input step; middle is high-gain output into 50Ω; lower low-gain 50Ω output.

Fig.1. Switched-gain amplifier uses two current-feedback amplifiers with differing gains in parallel, the relevant one being switched into circuit using enable/disable pins.



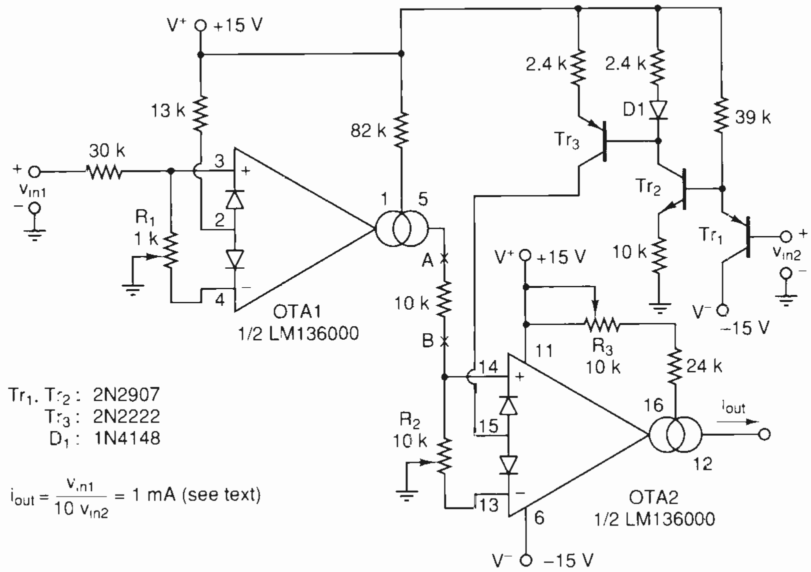
OTA analogue divider

National Semiconductor's *LM13600* is a dual operational transconductance amplifier with linearising diodes, the bias current of which may be varied to vary the gain of the amplifier. The circuit shows a simple and accurate analogue divider using the principle. Output current is $i_{out} = (v_{in}/10v_{in2})$mA. OTA_1 and the three transistors convert the signal voltage inputs to currents used as source and diode bias.

To adjust the circuit precisely, set R_1 with $v_{in} = 0$ so that $v_{AB} = 0$. Apply 10V to v_{in2} and adjust R_2 to give zero i_{out}. Then, with equal inputs to v_{in1} and v_{in2}, set R_3 to give $i_{out} = 0.1$mA.

Signal voltage at v_{in2} must be greater than zero and $|v_{in1}|$ less than $10v_{in2}$. Ideally, v_{in1} should lie between −10V and 10V, and v_{in2} should be 1V–10V. The circuit works over the 0-100kHz range.

Alexandru Ciubotaru
University of Texas at Arlington
Arlington
USA



In addition to the amplifier bias input, varying diode bias on National's LM13600 transconductance amplifiers varies the gain. This circuit uses this input to make an analogue divider, giving an output current proportional to V_{in1}/V_{in2}.

Pulse-width sequencer

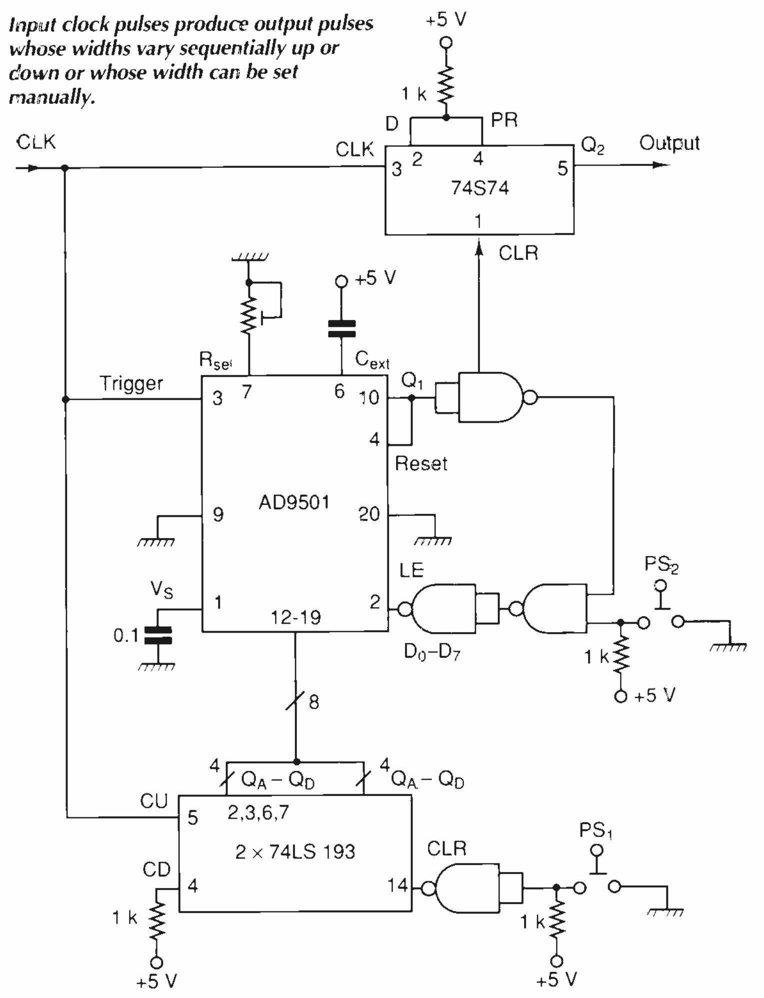
Pulse trains with sequentially varied widths, and digitally programmed pulses are both generated by this circuit or a variant.

Based on a digital delay generator *AD9501* (or the faster *AD9500*), the circuit sequentially increments or decrements the width of pulses, maximum width being set by R_{set} and C_{ext}. Input clock pulses set the flip-flop, increment the two *74LS193* up-down 4-stage binary counters and trigger the delay generator, which delays the output pulse by a period determined by the output from the counters; initially, delay is simply the propagation delay of the DGG. Also, the counter outputs are latched into the DGG to provide the new delay for the next cycle. As pulses appear at the clock input at a fixed frequency, the output pulse widths vary from minimum to maximum, whereupon pulse width returns to minimum as the counter returns to zero. Alternatively, the *193*s count down and the pulse widths decrease instead of increasing.

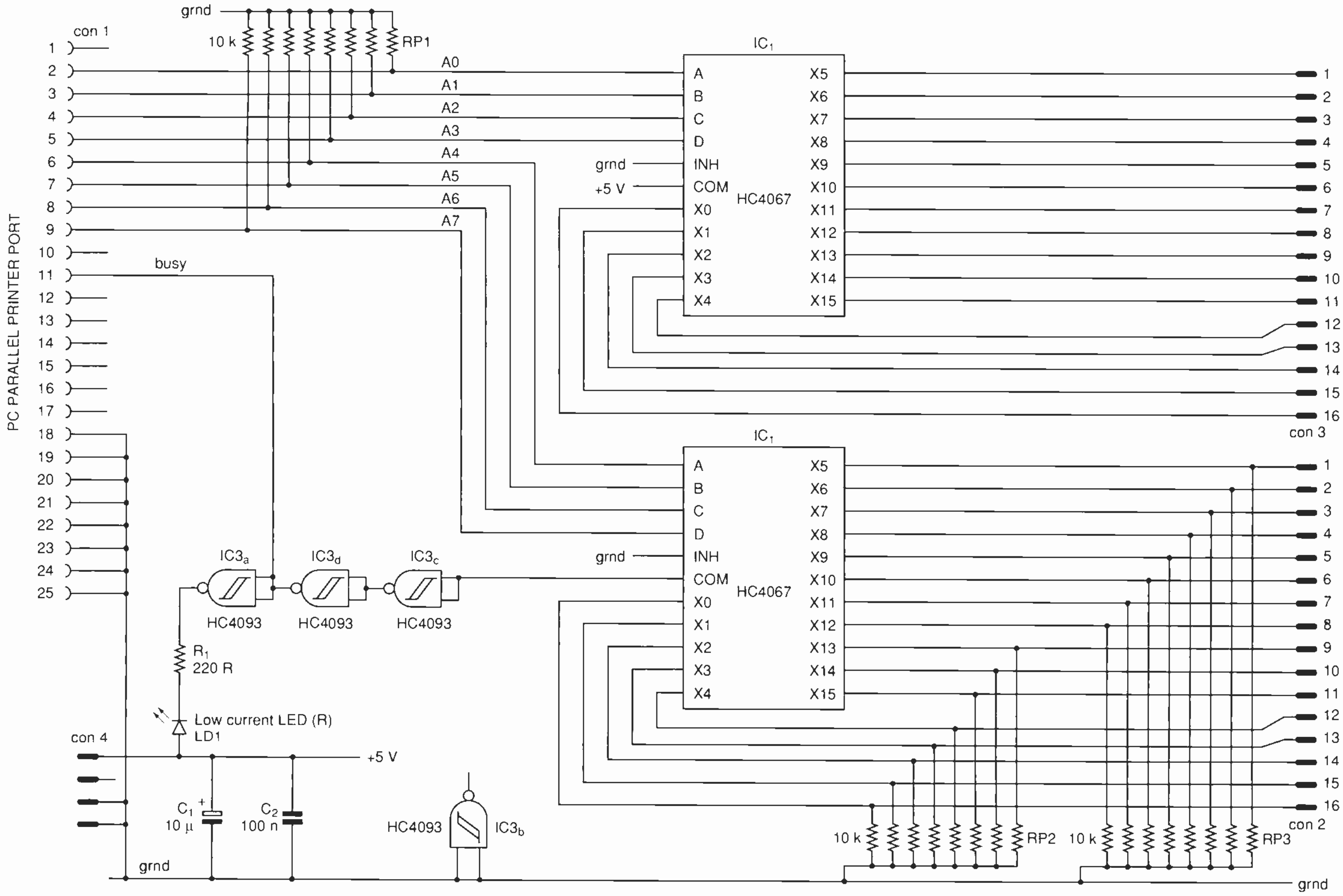
If dip switches replace the counters, or the counters are inhibited at the required count, the circuit becomes a digitally programmed pulse-width generator.

SR Kaul, IK Kaul and R Koul,
Bhabha Atomic Research Centre
Bombay
India

Input clock pulses produce output pulses whose widths vary sequentially up or down or whose width can be set manually.



Will Mr Paul Lovell, whose wide-range ceramic vfo appeared in the March issue, please contact *EW+WW*'s editorial office on 081 652 3614?



Automatic cable and connector tester

Connected to the parallel printer port of a PC, this device tests for shorts and open circuits in connectors and cable assemblies with up to 16 ways.

Two *4067* multiplexers take a sequence of addresses from the PC port, the top half of the 8-bit bus addressing IC_1, which switches a 5V test voltage to each of the 16 output lines in turn. While each of these lines is selected, the lower half of the bus addresses IC_2, which scans the 16 inputs for a voltage. Absence of voltage indicates open circuit, while voltage on more than one line shows short circuit.

When the 5V is found on an input, the printer port Busy line goes high and the led is lit. The top 4-bit nibble of the 8-bit address bus is used in the software to generate a column address, the lower nibble providing a row address to a 16 by 16 matrix, in which the BUSY bit acts as data; any lines out of sequence are thereby shown.

Tester connected to a PC parallel port automatically indicates open-circuits, shorts and wrong connections in connectors and cable assemblies.

A G Birkett
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Cheaper, low-voltage ultrasonic microphones

Small electret microphone inserts are sensitive to at least 90kHz, are inexpensive and only need a 1.5V to 9V supply; they can often replace the more expensive capacitor microphones which need a polarising voltage.

In the diagram, the supply comes via a resistor of 1kΩ, typically. The coupling capacitor is chosen to give a low-end roll-off in the lower ultrasonic region to avoid amplifier overload in the audio band and to provide a rising characteristic to counteract the drop in sensitivity at higher frequencies. Any further frequency-response shaping can come after the low-noise first-stage amplifier. ■

Les May
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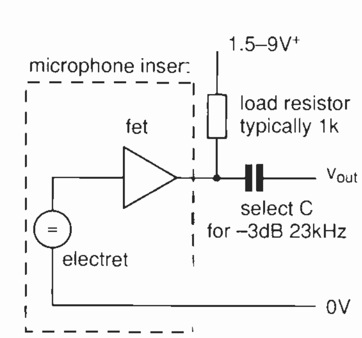
Lancashire



Consider using an electret microphone insert to replace the more expensive capacitor type, which needs a higher voltage supply, for ultrasonic work.

NEW PRODUCTS CLASSIFIED

ACTIVE

Asics

Non-volatile asic memory. Provision of non-volatile memory on asic without extra cost or complexity is offered by AMS. Where non-volatility is needed, but not reprogrammability, this one-time programmable technology avoids the use of eeproms, with their extra cost. Memory cells use polysilicon strip fuses, whose resistance changes from 50Ω to 20Ω when programmed in, eliminating re-growth. Alternatively, metal fuses are used in some low read-current roms, which also avoid re-growth problems. Austria Mikro Systeme Intnl, 0276 23399.

600MHz semi-custom array. GPS's *DX* series of arrays combine analogue and digital circuitry on one chip and is designed for high-current driving; bus, line and transformer interfacing; and analogue signal processing with high speed and low skew. AGC bandwidth is 300MHz with a control range of 20dB and the phase-lock loop operates at 300MHz. There are seven arrays in the series, from 630 equivalent gates and 26 analogue cells to 9405 gates/120 cells. GEC Plessey Semiconductors, 0793 518510.

0.5μ 3V asics. VLSI's 0.5μ 3.3V cmos gate array clocks at 200MHz – more than 30% faster than 0.6μ 5V arrays. Two cell-based libraries are available, offering up to 800,000 gates. These chips need only half the power of 5V types and are in ball grid array and Tab packages with up to 672 usable pins. VLSI Technology Ltd, 0908 667595.

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

Fast D-to-As. Qualcomm *Q2500* digital-to-analogue converters are 10-bit and 12-bit devices working at high speed and are intended for use in direct digital synthesis systems. *Q2510* is a 10-bit type offering an update speed of 100msps, settling time of 4.5ns to ±0.5lsb and low glitch impulse of 1.5pV/s. The 12-bit *Q2520* updates in 80ns and has a settling time of 27ns. Both types are ttl-compatible. Chronos Technology Ltd, 0989 85471.

Linear integrated circuits

AC bridge interface. Needing only a few passives to set frequency and gain, Analog's *AD698* is a complete signal-conditioning facility for linear variable differential transformers or any AC bridge transducer, output being scaled direct voltage. Gain and offset drifts are 20 and 5ppm/°C and, since both primary and secondary amplitudes are measured, the ratio determined and scaled, drift in the primary drive voltage is eliminated. A drive voltage of 20Hz-20kHz is provided at up to 24Vrms. Analog Devices Ltd. 0932 253320.

Low-noise op-amp. Comlinear claims its *CLC425* 1.7GHz op-amp to have the lowest input noise available at 10.5nV/√Hz and 1.6pA/√Hz. It is a voltage-feedback device with balanced inputs and low drift, and gives 96dB of open-loop gain, 100dB common-mode and 95dB power-supply rejection ratios. Slew rate is 350V/μs. Comlinear Europe Ltd, 0203 422958.

Microwave diodes. RF and microwave components by M/A-Com include silicon and gallium arsenide diodes, switches, chip capacitors, inductors and tuning varactors in a variety of package styles. There is also a schottky diode crossover ring quad in sot-143. M/A-Com, 0344 869595.

Quad spst switches. All three devices in Maxim's *MAX351/352/353* quad analogue switch series offer an on resistance of less than 35Ω, guaranteed to match within 2Ω between channels and to remain flat within 3Ω over the signal range. The switches operate with a 10V-30V single supply or with ±4.5V±20V split rails, while retaining cmos compatibility. Maxim Integrated Products Ltd, 0734 845255.

Fast, precision op-amp. Offering 100MHz bandwidth and 2000V/μs slewing, coupled with a 3mV offset and 5μA and 3μA input bias on inverting and non-inverting inputs respectively, National's *LM6182* dual current-feedback op-amp is claimed to be the best in its class. It also handles 100mA output current, driving a 2V signal into 50Ω and exhibits 0.05% differential gain and 0.04° differential phase. Settling time is 50ns to 0.1%. National Semiconductor, 01049 8141103499.

Wide-band vhf/uhf amplifier. A range of laser-trimmed, thin-film wide-band amplifiers from Philips, the *OM208X* family are for use as general-purpose amplifiers for vhf and uhf. Three basic modules, *OM2081/2/3* have one, two and three stages respectively and each type comes in two versions, covering either 40-600MHz or 40-860MHz. All take a +2V supply, put out 105dBμV for second-order and 1-5dBμV for third-order at –60dB intermodulation distortion, and exhibit a noise figure of 7.5dB. Gain figures are 10-12dB, 19dB and 29-30dB for the three types. A further type, the *OM956/1* covers 860-2050MHz. Philips Semiconductors, 010 31 40722091.

LCD controller. Everything needed to drive two-line by 24-character or four-line by 12-character dot-matrix lcds is contained in Philips's *PCF2116X* display controller. It has on-chip lcd negative-supply generation with all bias voltages, an 80-character display ram, rom and ram-based character generators and display-multiplexing logic and row and column output drivers. The multiplex clock is also internal, so that the only external components needed are two 100nF capacitors. Philips Semiconductors, 010 31 40722091.

Logic building blocks

PLL clock driver. Compatible in all ways with the Motorola *88915*, IDT's *IDT54/74FCT88915* is a minimal-skew clock driver for use up to 133MHz, having a built-in phase-locked loop to reduce duty-cycle distortion, eliminate delays and provide multiples or fractions of the input clock. Eight outputs are available. Either of two sources is pin-selected and there is a frequency-select pin for further divide-by-two of the output frequency. IDT Europe Ltd, 0372 363734.

Memory chips

4M srams. 25ns, 4M srams by Fujitsu, the *MB82201* and *MB82208*, are rated at 715mW and 825mW respectively at full speed, both having an automatic switch-to-standby feature to give a consumption of 137.5mW or less. *MB82201* is pin-switchable from x 1 to x 4 organisation, while the *MB82208* is organised as 512K x 8, with the centre-pin supply and ground for improved noise performance. Hawke Components Ltd, 0256 880800.

1Mbit drams. Available in 1Mb by 16 and 1Mb by 16 organisations, Toshiba's 3.3V and 5V multibit drams come in asymmetrical and symmetrical address options, offering refresh cycles of 1024 cycles/16ms and 4096cycles/64ms respectively. Access times lie between 60ns and 80ns, depending on voltage and address mode chosen. Toshiba Electronics (UK) Ltd, 0276 694600.

Microprocessors and controllers

Display controllers. From Mitsubishi, the *38000* series of 8-bit microcontrollers includes general-

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3V clock oscillator. Measuring 7.5mm by 5mm by 2.3mm, IQD's *IQXO-66* is a 3V surface-mounted crystal oscillator covering the 1.5-50MHz range and taking a supply current of 8mA (12mA over 40MHz). The devices are compatible with ATE equipment. IQD Ltd, 0460 77155.

purpose devices as well as those intended to control fluorescent and liquid-crystal displays. Rom, ram, i/o and control registers are all within the same address space to allow data transfer and operation to be performed by common instructions; in expanded and external memory modes, i/o ports become address, data and control lines. All devices have low-power modes from a 3V supply. Display controllers interface directly with fluorescent and lc displays. Highland Electronics Ltd, 0444 236000.

Eight bits at low-power. Hitachi's *300L* family of microcontrollers perform at power levels associated with 4-bit types. The CPU clocks at up to 5MHz, giving a 400ns instruction cycle time, but operates at 32kHz at low power. The *H8/3814* is a reduced peripheral set, low-cost version of the *H8/3834*, offering a choice of 32K, 24K or 16K of mask rom and 512byte of on-chip ram, together with an lcd controller/driver and a 12-channel, 8-bit A-to-D converter. Operating speed at 4V is 0.5μs and at 2.7V, 1μs. C software tools include a fuzzy logic compiler. Hitachi Europe Ltd, 0628 585000.

Processor with standard cells. Motorola's *68307*, a processor based on the *68EC000* core processor, is based on standard cells and is highly integrated. The technique will, in future, allow users to design their own processors using a core from the *68000* family. The device has three bus interfaces: a dynamically sized *68000* bus; an *8051*-type bus; and a Motorola I^{2}C-compatible M-bus. It interfaces without glue logic to boot rom, sram, and memory-mapped peripherals. Operating power is 150mW, reducing to 0.5mW in power-down. Motorola, 0908 614614.

Microcontroller for large memories. *TLCS-90* microcontrollers from Toshiba are C-based 8-bit devices, capable of addressing 8Mbyte of program and data. *TMP90C05IF* has a real-time clock and two general-purpose serial interfaces to be used as uart or clock-synchronous sio. Two dms channels handling 2byte transfer, 0.88Mbyte at 16MHz, are provided and two of the six general-purpose ports can be used for phase/pattern generation. There are 163 instructions, including arithmetic and 16-bit manipulative operations. Development tools are available. Toshiba Electronics (UK) Ltd, 0276 694600.

Mixed-signal ics

Frequency synthesisers. Three dual-frequency synthesiser ics from Philips, *UMA1015M/1018M/1020M*, provide low-power, low-voltage operation for all the newest mobile telephone systems, including *CT1*, *Amps/Tacs*, *GSM*, *DCS1800* and *Dect* covering the 400-2400MHz range. Two independent synthesisers control carrier frequency generation and IF or offset loop frequency and there is an output port or digital-to-analogue converter to control other handset functions. Philips Semiconductors, 010 31 40722091.

Single-chip audio Mpeg. A single ic supporting all Mpeg sampling and data rates, including free format, Texas's *TMS320AV110* Mpeg audio decoder will decode mono, dual stereo and joint stereo modes, producing decompressed audio output in either 16-bit or 18-bit serial pcm for input to most D-to-A converters. MPEG layers one and two are supported. If buffering of the audio stream is needed, there is provision for a 256K by 4 external dram and there is also a microprocessor interface. Polar Electronics Ltd, 0525 377093.

Optical devices

Wide-body optocouplers. Three ttl-compatible optocouplers from H-P are suited to high-speed logic interfaces, i/o buffers and line receiving in the presence of very high ground or induced noise. *CNW137*, *CNW2601* and *CNW2611* use an AlGaAs diode and high-gain detector and are contained in a 0.4in wide 8-pin dip in the interests of safety. Speed is 10Mbd and maximum working insulation is 1kV rms. Hewlett-Packard Ltd, 0344 362867.

Photodetectors. Infrared InGaAs photodetectors by Centronic have active areas from 0.25mm^2 to 100mm^2, providing detection in the range 800-1700nm at 0.998 quantum efficiency. Centronic is also offering overlaid Si and InGaAs detectors offering 350-1700nm in one package with no possibility of non-uniform spatial response. Optilas Ltd, 0908 221123

Oscillators

Crystal stability with no oven. Avoiding the need for a current hungry crystal oven and a protracted warming period, GEC Plessey has brought out the *OD9301* crystal oscillator, which uses a digital method of temperature compensation. Frequency stability of the 36mm by 27mm by 12mm device is ±0.3ppm from –10° to 70°C or ±0.5ppm from –40°C to 85°C. It is virtually immune to supply and load variation. GEC Plessey Semiconductors, 0793 518510.

Power semiconductors

Low-current regulators. Seiko's *S-8435/6* voltage regulators feature 0.9V minimum operation and operating and shutdown currents of 5μA and 0.2μA. Fixed output voltages between 1.5V and 12V are produced; external *L* and *C* convert the devices to step-up switching regulators. The *S-8436* series uses an external transistor to boost current. Amega Electronics Ltd, 0256 843166.

Dual regulator. The Cherry *CS8147* dual-output regulator gives 10V at 500mA and 5V at 40mA, with protection against reverse battery connection, 60V load dump, 50V reverse transient, shorts and thermal overload. It is provided with an enable input to put it into sleep mode, in which condition it takes 70μA. The device needs no external capacitor. Clere Electronics Ltd, 0635 299910.

Power-saving regulators. Texas Instruments *TL75* voltage regulator series includes a disable facility to conserve battery life. Use of a 4.85V regulator dropping out at 5.05V and producing a minimum 4.75V is an improvement on using a 5V regulator which drops out at 5.2V, extending the life of a 6V battery. The range covers 4.85V-10.22V. Flint Distribution Ltd, 0530 510333

70A, n-channel mosfet. Claimed to show the lowest on-resistance of any 30V mosfet in a TO-220 package – 10mΩ at 70A – the Harris *RFPON03* turns on in 80ns and off in 125ns. Salient features include 30V drain/source breakdown voltage, 175°C junction temperature, 1μA zero gate-voltage drain current and 3300pF input capacitance. A *PSpice* model for pc compatibles is available. Harris Semiconductor (UK), 0276 686886.

1200V diodes. *Hexfred* is a family of ultra-fast, soft-recovery diodes, rated at 1200V, which enable power switches to run cooler and gives lower rfi and emi and a reduction in snubber components. The diodes come in single or dual forms and are rated at 6-30A forward current with reverse recovery in 26-30ns. International Rectifier, 0883 714234.

Step-down converters. Maxim's *MAX730A* family of pwm step-down dc-to-dc converters deliver up to 50% more output current than the previous *MAX730* series and are up to 94% efficient. They feature fixed-frequency operation at around 180kHz and there is no noise below that frequency. Guaranteed limits of 159-212.5kHz avoid any risk of interference with the 455kHz IF band. Several models in the range have 5V outputs, others being adjustable from 1.25V to the input voltage. Output currents up to 750mA are available. Maxim Integrated Products Ltd, 0734 845255.

Step-up converters. 30mA at 5V or 60mA at 3.3V from one 1.1V cell can be obtained from the *MAX777/778/779* family of dc-to-dc converters. Guaranteed start-up is 1V on a 10mA load. For 1.5V input, output current is 150mA at 5V and 250mA at 3.3V. The only external components are two capacitors and a 22μH inductor. Output voltages are: 5V (*777*); 3V or 3.3V (*778*); and 2.7V-6V (*779*). Maxim Integrated Products Ltd, 0734 845255.

PASSIVE

Passive components

Chip inductors. Beckman's *BML* series of surface-mounted, multi-layer chip inductors are magnetically shielded to enable close mounting. Inductance range is 0.047-220μH in 43 standard values, with tolerances of ±20% (to 0.068μH) and ±10%. Beckman Industrial, 0384 442393.

High-voltage trimmers. Variable capacitors with breakdown voltages of over 30kV in values up to 200pF, Jackson's *770* series are said to be up to 60% less expensive than the vacuum-dielectric types. They are cylindrical and are varied by the rotation of a shaft which alters the position of a plunger inside the body. Rotor connection is via the single-hole mounting bush and a wire-end connection for the stator. 100pF and 200pF models have a non-rotating piston and constant shaft length for motorised operation. Jackson Brothers Ltd, 081-681 2754/7.

Smaller capacitors. *FA* series radial-lead capacitors from Panasonic are 40% smaller than previous types from the company, covering the 68μF to 15,000μF range at rated working voltages of 6.3V-63V, from temperatures of –55°C to 105°C. Panasonic Industrial Europe, 0344 853827.

Snap-fit capacitors. Two new families of electrolytic capacitor from RS have snap-fit terminals for easier soldering. A general-purpose range handles voltages from 16V to 450V in values from 68μF to 22,000μF, working up to 85°C, or from –25°C to 105°C in a reduced working voltage of 400V. The compact range spans 68μF to 68,000μF with the same voltages and temperature ratings. RS Components Ltd, 0536 201234.

Displays

Plasma displays. A range of plasma display modules from Okaya are based on lanthanum boride cathode technology, producing high secondary electron radiation to give brightness levels up to 700cd/m. Having no mercury, the units cope with a –20°C to 75°C temperature range. They are based on a 5 by 7 matrix, and displays range from 16-character by four-line displays to 40-character by 12-line types. Dot matrix panels are also available in sizes from 96 by 48 to 512 by 160 elements. Drive and control circuitry are incorporated to interface with ttl or a cpu bus. Highland Electronics Ltd, 0444 236000.

Captioned panel lamps. Panel-mounted led indicators in Oxley's *MIL80* range have alphanumeric characters and come in red, yellow and green, in addition to super-bright AlGaAs red. Oxley says it can match house styles. Other types in the range include wide-angle or focused, sunlight-visible models and flashing types. Oxley Developments Co. Ltd, 0229 582621.

Filters

Variable filters. New versions of Kemo's *VBF3/4* variable filter instruments offer a variable slope of between 6dB and 24dB per octave in 6dB/octave steps. They are both two-channel units with continuously adjustable cut-off frequency and with high-pass or low-pass response on each channel. *VBF3* covers 0.1Hz-10kHz in five ranges, while *VBF4* covers 1Hz-100kHz. Connecting both channels in series forms high-pass or low-pass filters with slopes of up to 48dB/octave, while a parallel connection allows band-stop and band-splitting arrangements. Characteristics are accurate representations of the Butterworth response. Kemo Ltd, 081 658 3838.

Hardware

Dip pin adaptor. *Correct-A-Chip* is an adaptor that accepts soic, tssop and tsop packages and is soldered onto a through-hole board in standard dip format. All three device types fit the same adaptor, including the 'skinny' type. Models are available with 4-24 pins. Aries Electronics (Europe), 0908 260007.

Instrumentation

PC pcb fault diagnosis. Providing control and display on a pc, ABI's *System 8* printed-board fault finder locates dry joints, broken tracks, shorts and malfunctioning ics. The software presents point-and-click menus and a graphical display, interrogation of the board via an ic clip resulting in a circuit map, 'transparent' to the user. Analysis shows expected and actual response and the PCB signature can be developed and saved for later comparison. Software tools include VI node and connections analysis, node voltage sensing and shorts detection. Complete package can include a pc or the required hardware and software for installation. ABI Electronics Ltd, 0226 350145.

100MHz oscilloscopes. Hitachi Denshi points out that its established *V-1065A* and *V-1085* real-time oscilloscopes are good value for money at £1495 and £1966 respectively. The two and four channel instruments incorporate sweep-time autoranging, crt readout, cursor readout, auto trigger level, selectable signal output, frequency counting and trigger lock for viewing irregular pulse trains. Delayed sweep is provided. Hitachi Denshi (UK) Ltd, 081-202 4311.

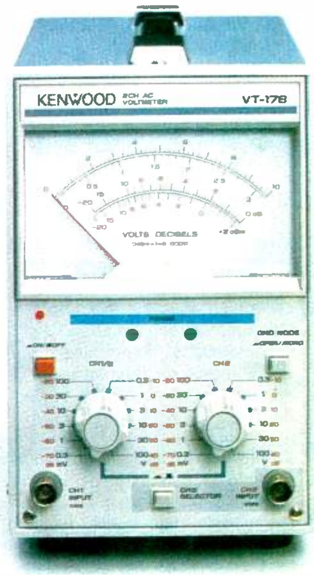
Stereo Rx and mod analyser. R&S says its *FMAS* is the first instrument to combine universal modulation analysis with fm stereo/television dual sound reception in the 5-1000MHz range. Input levels range from 10μV to 7V and, since the receiver can be switched off, the modulation analyser can be used over the 50kHz-1360MHz frequency range. Rhode & Schwarz UK Ltd, 0252 811377.

Oscilloscope probe. *PMK* oscilloscope probes can be used at bandwidths up to 500MHz. A new model has a retractable, spring-loaded tip which will not slip onto the board and, being sharply pointed, reaches into densely packed areas. It uses reed switches, hybrid circuitry and a laser-trimmed ceramic resistor network for accuracy. The standard *PMS221* probe is usable up to 150MHz. Test and Measurement Mktg, 043 879 513.

Colour pattern generator. For both uhf and vhf bands, Kenwood's *CG-922* portable colour television pattern generator provides video and rf outputs, with a 1kHz superimposed audio intercarrier. Outputs include red, green and blue rasters, with dot and cross-hatch patterns. Trio-Kenwood UK Ltd, 0923 816444.

Literature

PCbus catalogue. In 136 pages, Arcom describes processor and i/o boards, enclosures and software to enable pcs to be used in industrial control and data acquisition, all using *PCbus* or *PCATbus*. Arcom Control Systems Ltd, 0223 411200.



Voltmeter. A two-channel instrument, Kenwood's *VT176* is an analogue ac voltmeter, its large display indicating 0.3mV-100V rms in 12 ranges. Input impedance is 10MΩ and 45pF and frequency response is 5Hz-1MHz. Linked and independent outputs are switch selected. Trio-Kenwood UK Ltd, 0923 816444.

DMM/oscilloscope. Three models in Tektronix's *TekMeter* range combine a 5MHz oscilloscope with lcd readout and a true-rms digital multimeter. By containing most of the oscilloscope circuitry in one ic, size and cost is reduced, although input circuitry is separate to cope with surges up to 6kV. Full X and Y auto-setup handles a changing waveform, so that the oscilloscope is capable of virtually hands-free operation. The lcd allows a grey-scale, to simulate the varying brightness of a crt with frequency of painting. Single and dual channel models are available. Tektronix UK Ltd, 0628 486000.

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LF measurements guide. Fluke's practical guide to dc and lf measurement, *Calibration: philosophy in practice*, is in its second edition. In 544 pages, it covers standards, calibration, statistics, laboratory management and 'practical considerations'. There is a glossary and a section on metrology resources. Fluke (UK) Ltd, 0923 240511.

Power mosfet guide. IR's designer's guide to the use of hexfet power mosfets describes, in 1600 pages, all the company's devices and has appendices covering test circuitry, package outlines and tape and reel data. Application notes describe the practicalities of using hexfets. International Rectifier, 0883 714234.

EMC Directive infopack. An information pack on the European EMC Directive contains 32 pages of comment and reference material. It explains how to show compliance with the Directive and reviews existing, proposed and revised standards. There is a list of sources of further information. Free. Radio Frequency Investigations, 0256 851193.

Materials

PCB material. *RO3003* from Rogers Corporation is a new material for the fabrication of printed boards that offers uniform dielectric constant, low loss, electrical and dimensional stability with temperature and good plated-through hole reliability. It is available in 18in by 24in sheets and in thicknesses of 0.01in, 0.02in and 0.03in, with 0.5oz or 1oz copper. Rogers says it costs 25-30% less than Teflon glass-fibre material. Rogers Corporation, 010 32/9 2353611.

Digital potentiometer. *DP100BT* digital encoder by Control Transducers is based on an optical encoder, producing 100 ttl-level square waves per revolution, equivalent to 400 2-bit codes. Other resolutions are available. Free designer's guide. Control Transducers, 0234 217704.

Production equipment

Soldering station. Vector Components *Stiron* soldering station offers rapid heat recovery to cope with components that behave like heat sinks. Iron temperature is variable between 100°C and 400°C by means of a panel-mounted potentiometer, control being of the zero-switching type for low interference. A range of iron-clad bits includes a conical type and a 5mm screwdriver bit. Rendar Ltd, 0243 866741.

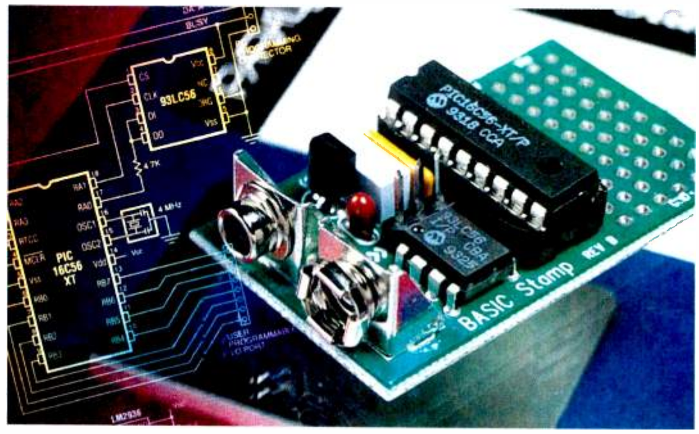
Power supplies

100W dc-to-dc converters. *Pico P* series high-power dc-to-dc converters are now available from Ginsbury. The range covers 3.3V-100V output and 18-380V dc input. Dual isolated outputs of 100W. Full protection is incorporated and fixed-frequency operation allows parallel operation. Ginsbury (UK) Ltd, 0634 290903.

Bench power supplies. New power supplies from H-P include the *HPE3630A* triple-output unit and the *HPE3616A* 0-35V supply, both in the low-cost range – less than £500. *E3630A* provides 0-6V at 2.5A, 1-20V and 0 to –20V at 0.5A, all with low noise. Also announced are the *HP6682A*, *6683A* and *6684A* 5kW switching supplies, exhibiting 10mV pk-pk normal-mode noise and 20mA pk-pk common-mode current. A down-programmer is provided to discharge the units, in between 475ms and 650ms. Hewlett-Packard Ltd, 0344 362867.

Supervisory module. Power-supply monitoring and battery backup are the functions of Maxim's *MAX1691* microprocessor supervisory module, with its lithium battery. Features include battery switch-over, memory write protection and a watchdog function which monitors software execution by asserting reset if there is no activity on a selected i/o line. Quiescent operating current is 35μA, 1μA in standby. Maxim Integrated Products Ltd, 0734 845255.

Rechargeable batteries. Yuasa has developed a high energy-density nickel-metal hydride battery that combines a nickel positive electrode with a negative electrode containing a Mischmetal-based hydrogen storage alloy. It has around twice the capacity of conventional nicad batteries. Terminal voltage is 1.2V. No cadmium or other pollutants are used. Yuasa Battery Sales (UK) Ltd, 0793 612723.



Radio communications products

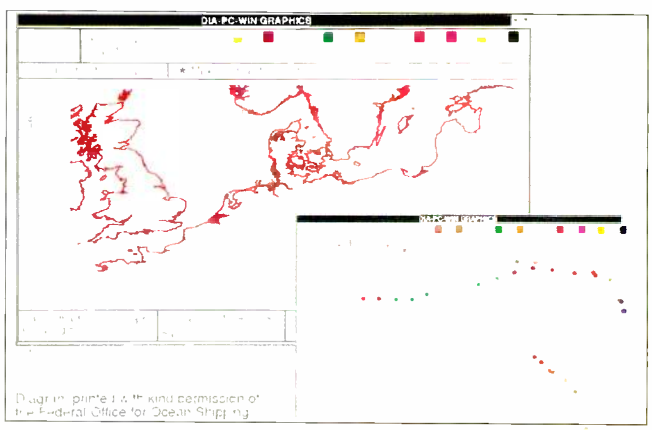
Telephone power amplifier. Iwatsu's *HAB* 900MHz power amplifier for portable telephones puts out 30dBm from 7dBm of drive, both impedances being 50Ω. Four ranges are available in the 824-940MHz band. Requirements are 4.7V at 370mA for rated power gain. Maximum supply voltage is 9V. Advanced Crystal Technology, 0635 528520.

18GHz dpdt switches. From KDI, the *SWM-1100* single-pole, single-throw and *SWM-1200* double-pole, st switches operate from 0.25GHz to 18GHz with a switching speed of less than 80ns. They have removable sma connectors and are sealed to mil std 883. Below 10GHz, isolation is over 61dB and more than 52dB up to 18GHz, with insertion loss 1.5dB/2.8dB. VSWR is 2:1 and rf power 1W with a peak of 10W. Anglia Microwaves Ltd, 0277 630000.

Broadband mixer. Covering the 200-3000MHz frequency band, the Mini-Circuits *RMS-30* mixer is contained in a 0.25in by 0.31in by 0.275in ceramic surface-mounted package. It

PIC controller development. A development kit for the Arizona Microchip *PIC16Cxx* microcontroller family is based on Parallax's *Basic Stamp* single-board computer and includes pc software for program editing and down-loading, with all cables and documents. Highland Electronics Ltd, 0444 236000.

Data in maps. *DataMap DIA-PC* from SciTech is a pc-based data evaluation package that links measured data with maps, plans and charts on a pc. The picture shows a North Sea map derived from an hpgl file, converted by *DIA-PC* to latitudes and longitudes. In this case radioactivity levels and the position of a ship are shown, part of the display being zoomed. The software runs under *Windows 3* or dos. SciTech, 0734 758857.



responds down to dc, being used for up and down conversion, bi-phase, qpsk and I and Q modulators. Mini-Circuits Europe, 0252 835094.

Telemetry transceiver. *BiM-418-F* 418MHz data telemetry and telecommand transceiver module by Quantelec contains a low-power uhf fm transmitter and superhet receiver with data recovery, and tx/rx antenna changeover circuits in one pcb-mounted package measuring 23 by 33 by 10mm. It provides a low-cost bi-directional datalink capable of half-duplex 19.2Kb/s data rates over 30m in a building or 120m over open ground. Quantelec Ltd, 0993 776488.

Switches and relays

Six-pole relay. Three versions of the LRE six-pole, 10A relay are available: *F670* is for electrical vehicles and trains, switching 1A resistive at 72V with a life of 1 million operations; *F600*, the military version carrying 1400A overload under 10A circuit-breaker protection; and *FD670*, a three-pole, double-break type, switching 3A at 72V dc or 1A at 110V dc. Several mounting options are on offer. LRE Relays + Electronics Ltd, 0962 734433.

Cmos-compatible relays. *116C* and *136C* two-pole changeover relays by Teledyne come in a hermetically sealed can and accept a cmos logic signal with no external buffering. Teledyne Electronic Technologies, 081-571 9596.

Transducers and sensors

Displacement-sensing processor. *Fastar* displacement sensors by Control Transducers now have a processor board, designed for rack-mounting or cased. Output is unidirectional at 0-20Vdc or bidirectional at ±10Vdc, with adjustable zero and span. A temperature compensation circuit reduces sensor errors and allows operation over a temperature range –50°C to 125°C. Sensor range is 19-609mm with errors down to ±0.1%. Control Transducers, 0234 217704.

Current transducers. Potted in a self-extinguishing case, Lem Heme's new range of Hall-effect current transducers handle up to 1000A, with a bandwidth of 25kHz and response time of 3µs. All models have a 25,000ampere-turn overload capacity, 10mA output burden, ±4V output sensitivity and ±1% accuracy. Dielectric strength is 3kV rms for a minute. Zero offset is ±10mV, drifting ±1mV/°C and gain variation ±0.05% of reading/°C. Lem Heme Ltd, 0695 20535.

Angular displacement sensor. Philips's *KMA10/70* contactless sensor is fully encapsulated, needs no adjustment and is intended for use in automotive and industrial application such as active suspensions and accelerator pedal position sensing. It is based on a magneto-resistive element, whose resistance changes in response to a moving field are detected by a hybrid signal conditioner producing a temperature-compensated 4-20mA output. Angular range is ±35°, covered at 20°/ms. Philips Semiconductors, 010 31 40722091.

COMPUTER

Development and evaluation

Logic synthesis. *PLSyn* is part of MicroSim's *Design Center* and offers device-independent logic synthesis integrated with schematic capture and min/max timing simulation, with analogue devices on the same schematic. The user can describe programmable logic on the schematic with a synthesis language, logic symbols or both – *PLSyn* searching a device database for the best devices for the design. Simulation of the design proceeds before and after implementation is selected. When the design is complete, the logic is automatically partitioned and fitted into the correct devices for speed, price or power consumption. MicroSim Corporation, (USA)800 245-302

Software

Filter design and test. SigCad claims its *Network Designer* to be the first integrated filter design and test package. Filter approximations included are Butterworth, Tchebychev, Inverse Tchebychev and elliptic in low-pass, high-pass, band-pass and band-stop types. Practical implementations include Sallen and Key and biquad. Specifications are entered using a gui, the software fitting a response between points set by the engineer, which is then turned into a circuit, edited and toleranced. SigCad Ltd, 0203 597797. ■

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High-voltage bridge driving

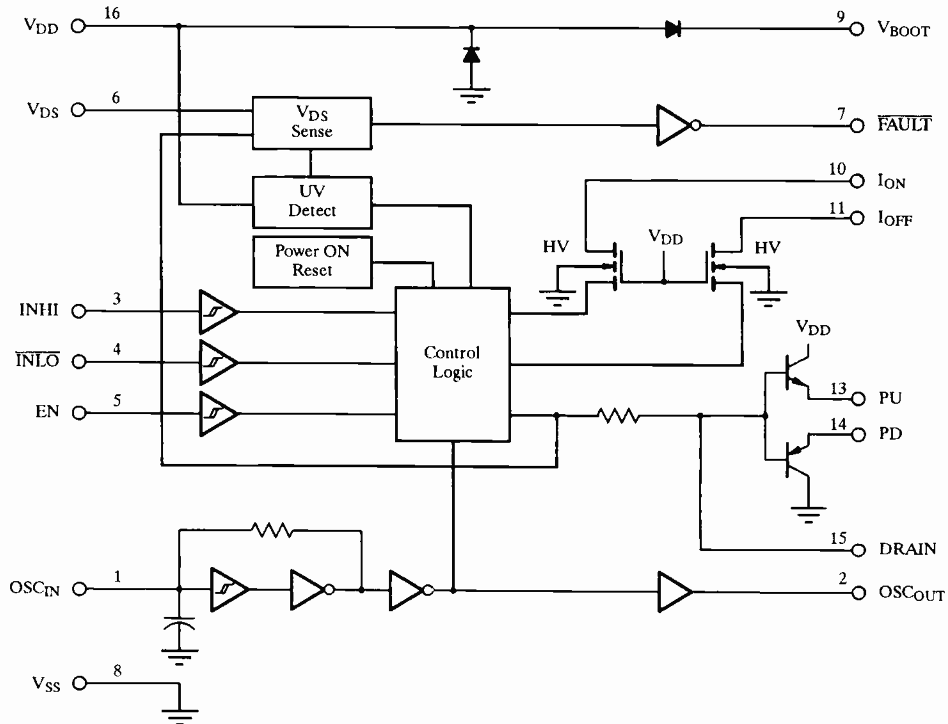
In motor control, power mosfet half bridges can provide efficient switching and low motor conduction losses. But to drive the top mosfet, whose gate may be at a potential hundreds of volts away from the gate on the low-side fet, either significant level shifting or isolation is needed.

Many designers use the high-side voltage to provide a bootstrap supply. According to Siliconix note *AN700*, this method is simple but has limitations. Level shifters need few components but suffer from noise sensitivity and need high-voltage components.

Addressing these problems, Siliconix has introduced an IC pair specifically for driving both high and low-side transistors in a power mosfet half bridge. Via special inputs and outputs, the two ICs can be connected directly. As a result, only one low voltage supply is needed apart from the high-voltage motor drive rail. Supply for the high-side IC is easily derived from the high-voltage rail.

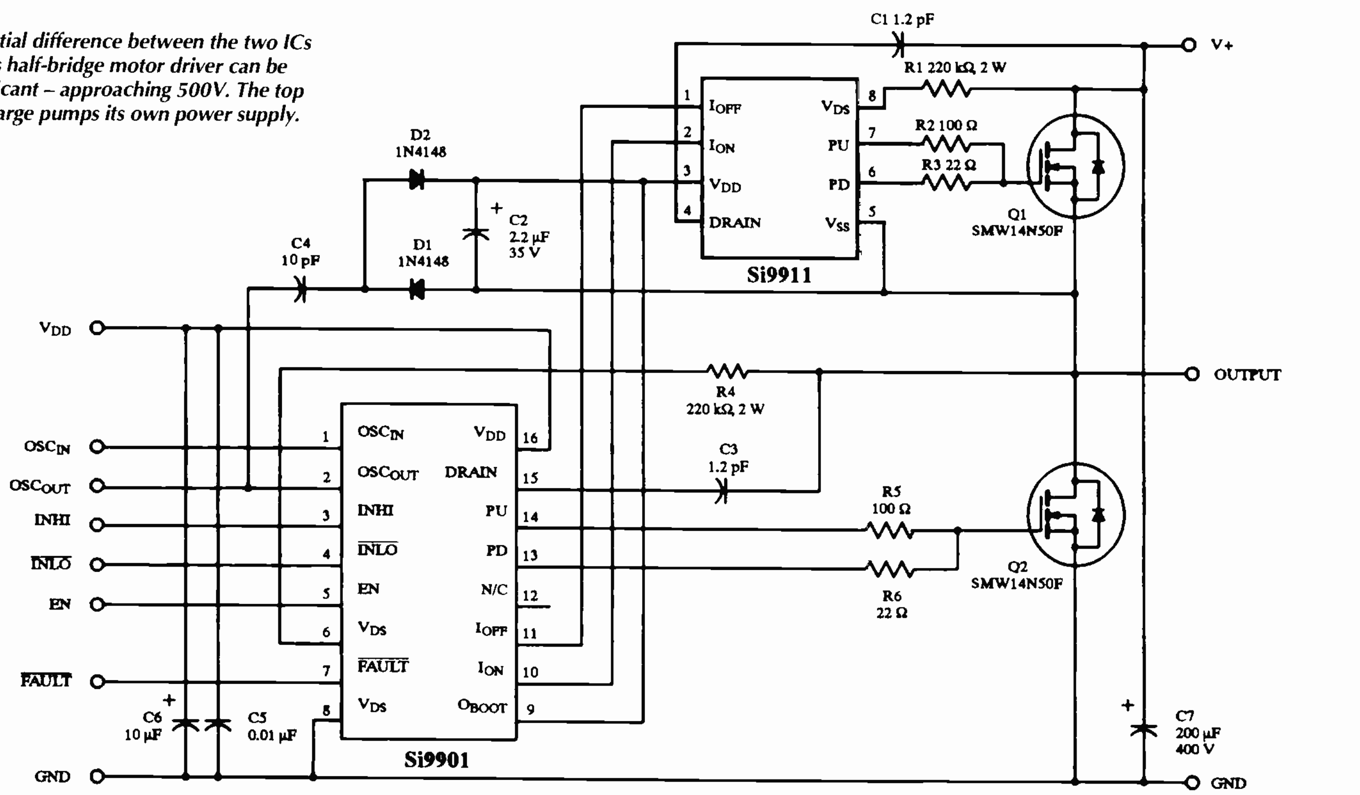
The two ICs can withstand a potential difference of 500V and their level shifters are said to offer enhanced noise immunity. Final output capability for driving the motor depends on the mosfets but the *9901* can source –0.5A on its PU output and sink 1A at PD. Drive capabilities of the *9911* are the same. In this instance the mosfets are 14A, 500V devices.

Control is based on a state machine driven



Functions of the low-side IC in the mosfet bridge-driving pair shown below. It has 500V level shifting and contains the 9901 oscillator for the high-side IC charge pump. Inputs are cmos compatible.

Potential difference between the two ICs in this half-bridge motor driver can be significant – approaching 500V. The top IC charge pumps its own power supply.



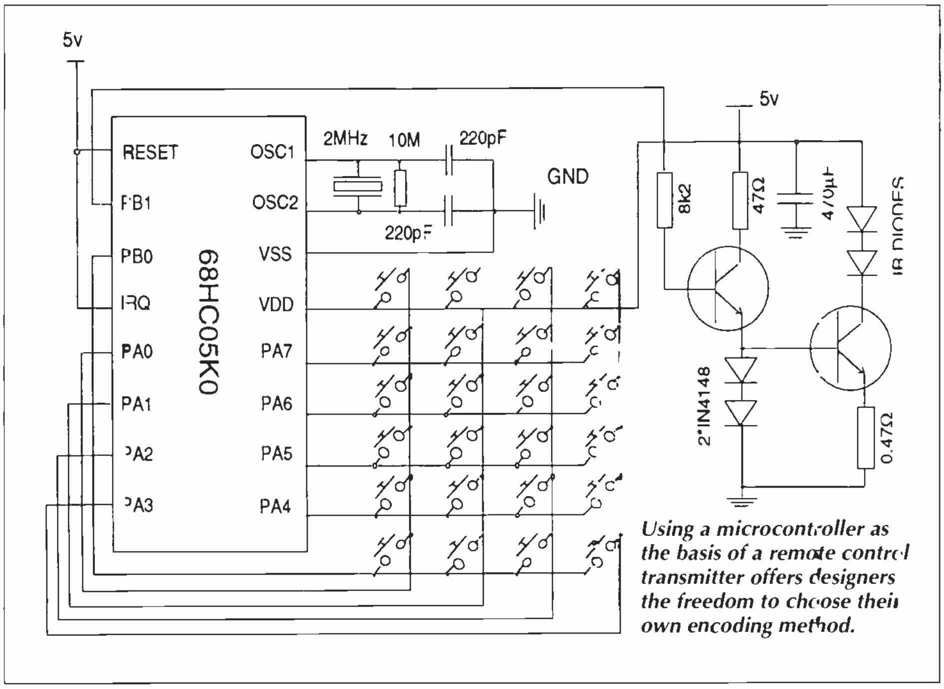
Remote controller via micro

Internal pull-downs and interrupt drivers on port pins result in a low-component count when using the *68HC05K0* microcontroller as the basis of a remote controller. Since the device has a software-controllable off switch and is cmos, battery drain is low.

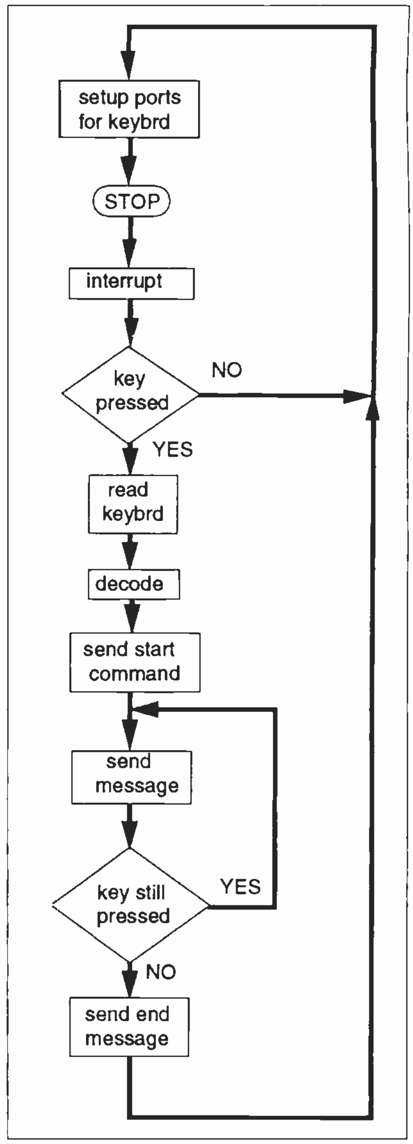
Detailed in Motorola's *AN463*, this remote control transmitter needs no components for keyboard interfacing and only a buffer for driving the infra-red diodes. By using the V_{dd} line to drive an extra row, the total number of keys available is 24 even though there are only nine input pins.

Full software details and listings are included in the note.

***Motorola European Literature Centre**, Tanners Drive, Blakelands, Milton Keynes MK14 5BP. Telephone 0908 614 614, fax 0908 618 650.*



Using a microcontroller as the basis of a remote control transmitter offers designers the freedom to choose their own encoding method.



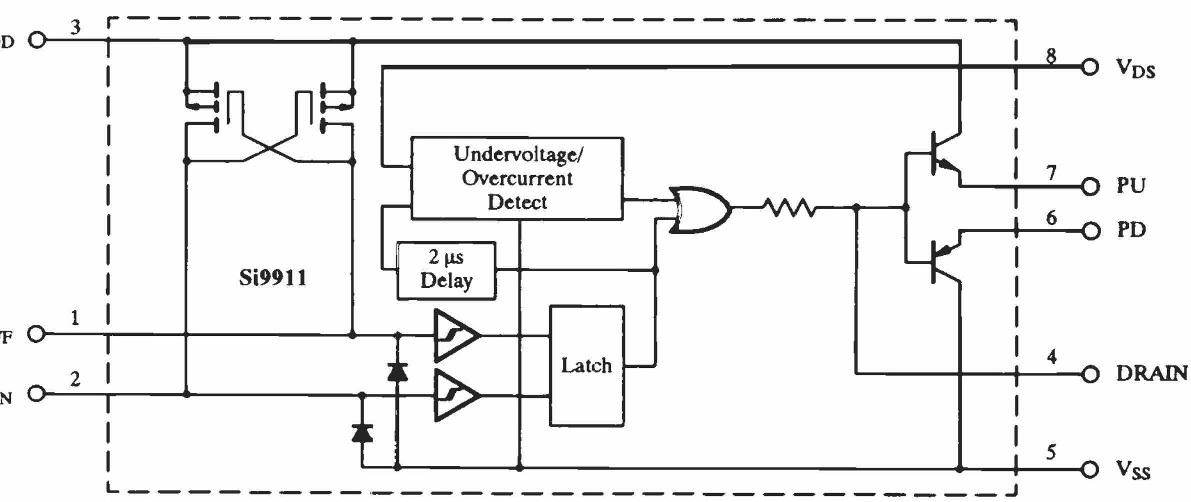
by a clock of typically 1MHz. All switching occurs on clock rising edges. Input signals IN_{HI}, IN_{LO} control the half bridge high and low-side mosfets respectively in conjunction with enable signal, EN. This structure facilitates PWM signals for motor direction and magnitude control in full bridges.

There is a second high-side drive option, the *9911* which needs external charge-pump diodes. All three devices have dv/dt control and in-built protection circuits against under voltage and shorts. Inputs of the low-side driver are cmos compatible.

Also included in the note are notes on selecting the right mosfet, calculating the charge-pump coupling and bootstrap capacitors in the high-side IC supply and determining the gate and sense resistors.

***Siliconix**, Easthampstead Road, Bracknell, Berkshire RG12 1LX. Telephone 0344 485 747, fax 0344 304213.*

9911 High-side driver shematic shows latching inputs and charge-pump circuuitry for generating power supply. Both high and low-side drivers have undervoltage lockout to ensure sufficient mosfet gate drive.



Audio noise reduction without encoding

The universal appeal of companders to noise-reduction circuit designers is the amount of noise they can suppress. However, one of the main reasons why Dolby B dominates the consumer market is that it offers only 10dB of reduction.

Although sufficient to lower cassette noise to an acceptable level, this reduction not high enough to spoil the sound of Dolby B tapes played back on a non-Dolby player. In fact, the sound of the hf boost from Dolby B encoded tapes is often preferred by listeners using medium-quality speakers.

The above statement comes from *AN386 – A non-complementary audio noise reduction system* in National Semiconductor's latest Linear applications manual. Noise reducers such as Dolby are said to be complementary, i.e. encoding is used for recording and complementary decoding is needed for playback. This note describes reducing noise by non-complementary means – namely dynamic noise reduction – for use with television and radio broadcasts, video tape audio and non-Dolby recordings

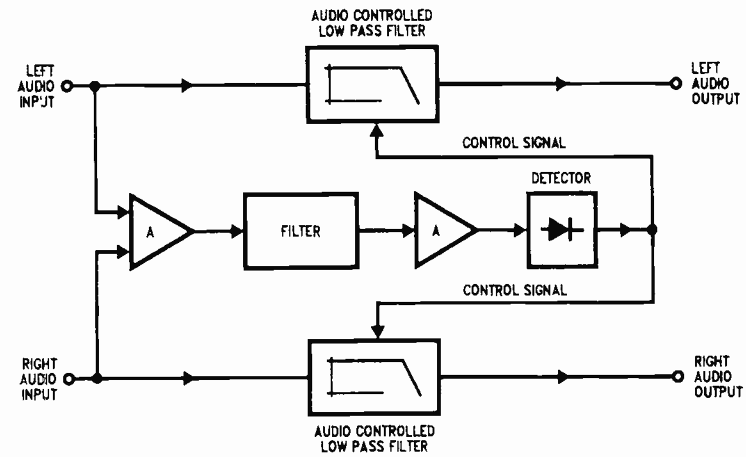
There are two National ICs for implementing noise reduction, the *LM1894* and the *LM832*. Both of these can offer between 10 and 14dB of reduction on stereo source material. Only the *1894* is dealt with in the paper

Two basic assumptions are involved. Firstly, noise is proportional to system band width and secondly, the desired program material is capable of 'masking' the noise when the signal-to-noise ratio is sufficiently high. Dynamic noise reduction automatically and continuously changes the system bandwidth in response to the amplitude and frequency content of the program.

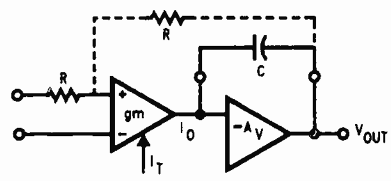
In the block diagram is a stereo noise reducer with separate controllable cut-off low-pass filters in the audio paths. Cut-off frequency control range is 800 to 35kHz. Both audio inputs contribute to producing the control signal via a peak detector. Using one control signal for both channels ensures that the stereo image remains stable.

The filter, simplified, is shown together with curves for open and closed-loop response. This topology is similar to the pole-splitting frequency compensation used in many op-amps. In it, a variable transconductance op-amp drives a further amplifier acting as an integrator.

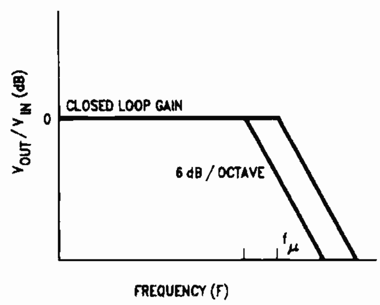
With a fixed capacitor, unity-gain frequency changes with transconductance as in (b). Putting dc feedback around both stages for unity closed-loop gain (c) results in flat amplitude response – unity gain – until f_μ is reached. At this point, gain



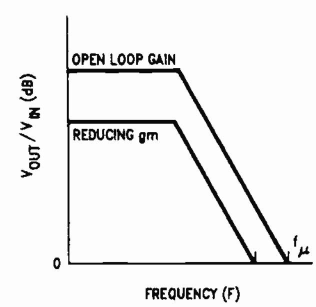
Unlike complementary systems like Dolby B, this dynamic noise reducer has the potential to improve any noisy audio signal.



(a) Variable Lowpass Filter

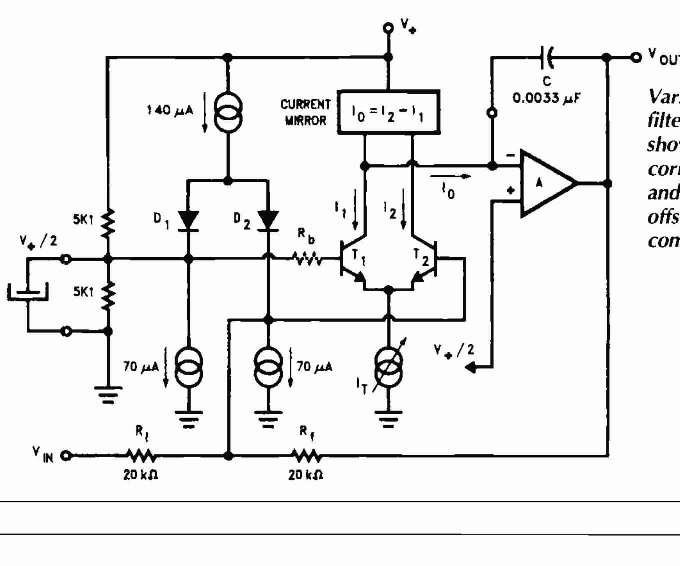


(c) Closed Loop Response



(b) Open Loop Response

Audio-controlled low-pass filter comprising transconductance amplifier and integrator. Closing the loop with R causes gain to remain constant, despite change in g_{mv} until the roll-off frequency is reached.



Variable low-pass filter schematic shows distortion-correcting diodes and control-voltage offset compensation.

follows the open-loop curve, falling at 6dB/octave.

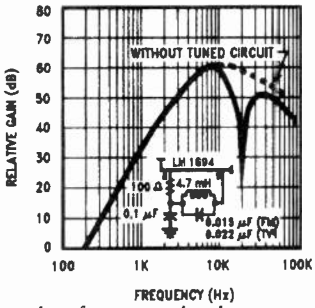
A more detailed schematic of this section is also shown, where R_f and R_i are providing the dc feedback. Current mirrors replace the more conventional load resistors in the differential pair forming the transconductance amplifier.

For the values given in the schematic, output voltage swing at the cut-off frequency is about 1V rms. At this input voltage level, the IC exhibits 3% THD. However, this is the condition for minimum bandwidth when noise only is presented at the input. When signals are present, bandwidth extends out to 35kHz. Transconductance-stage current is over 1mA, allowing theoretical signal swings of over 34V rms. In practice, the device handles over 4V rms.

Although there are other ways of forming a variable-cut-off filter, this design is said to offer advantages, especially in terms of control feed through.

Control path

The job of the control signal is to ensure that audio bandwidth is wide enough for the signal yet decreased rapidly when the signal falls. To do this, the control path must recognise the masking qualities of the signal

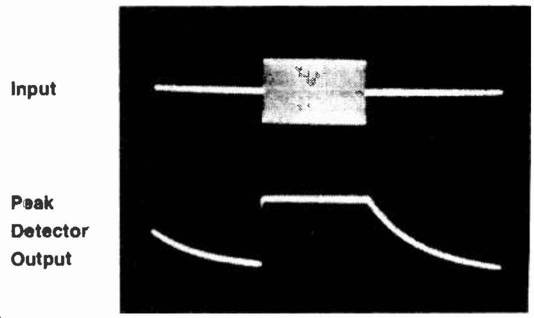


When no low-frequency signals are present, high-amplitude unwanted signals can cause bandwidth to open. Adding a notch filter prevents this.

source. In addition the detector must be able to take advantage of characteristics of the ear to prevent audible distortion.

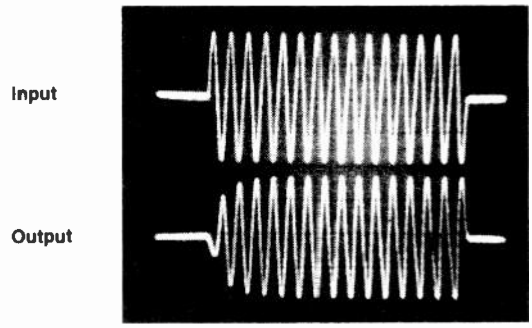
Left and right channels are summed and buffered. Noise levels can differ – cassette tapes are between –55 and 65dB depending on whether Dolby B is used, while fm broadcast noise is –45 to –75dB depending on signal strength. As a result, the control path is adjusted so that a noise input is capable of just increasing the audio bandwidth from its minimum value. In this way, any program material above the noise level increases bandwidth so that it is passed without distortion.

Combined with the control potentiometer is a 0.1μF capacitor that helps prevent high-



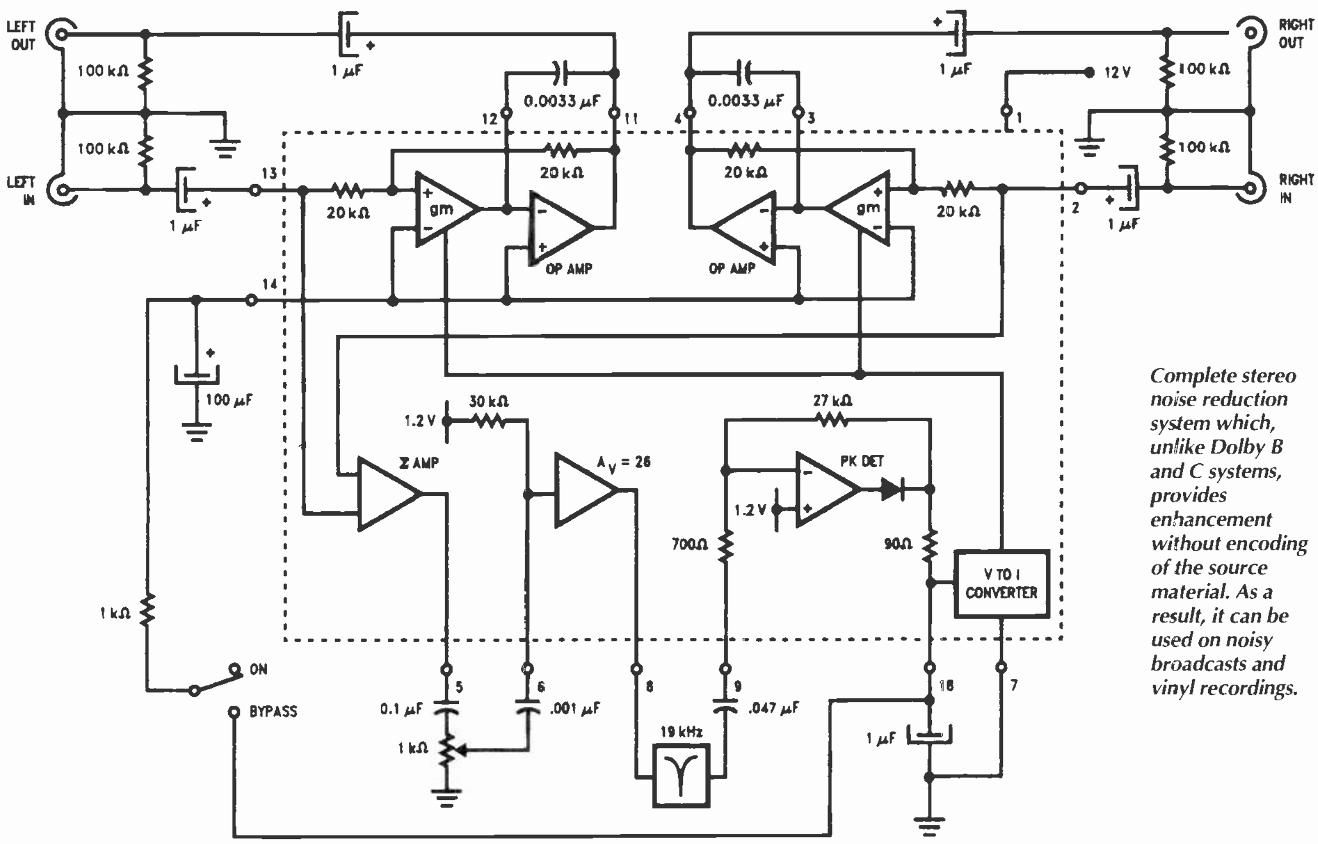
TIME: 20 ms/DIV

Peak Detector Response, 500 mV/Div



TIME: 0.5 ms/DIV

Audio Output Response, 10 kHz Tone Burst



Complete stereo noise reduction system which, unlike Dolby B and C systems, provides enhancement without encoding of the source material. As a result, it can be used on noisy broadcasts and vinyl recordings.

amplitude signals without content above 1kHz from opening up the bandwidth. For signals that do have high frequencies, control path sensitivity is increased at 12dB/octave. This rapid gain in sensitivity is important since harmonic content of program material falls off quickly with increasing frequency.

In some unusual circumstances, the signal can contain only high-level, high-frequency components. These can be caused for example by the line-scan signal of a tv or the fm stereo subcarrier. Although inaudible, such signals can affect the control path and cause noise. For this purpose the notch function shown in the control-path frequency-response diagram can be optionally inserted.

Detector

Both a negative peak detector and voltage-to-current converter are included in the detector. Input resistance of this stage, together with the coupling capacitor, forms part of the control-path filter. Similarly, output resistance of the detector and the gain-setting feedback resistor help determine the detector time constants.

An interesting contrast between this type of circuit and a companding one can be demonstrated with a 10kHz tone burst. Since the detector only responds to negative peaks, it takes about four input cycles to reach 90% of the final voltage on the detector capacitor – corresponding to a 500μs time constant.

After the first two cycles, audio bandwidth will have already increased past 10kHz. A comparison of input and output tone bursts causes only a slight loss in amplitude in the initial cycles as illustrated.

In a compandor however, fast detector time constants are normally avoided. This is because rapid changes in system gain that occur while recording transient signals could cause modulation products that may not be treated complementarily on playback. As a result there is a time lag before the system gain can change. This may be the maximum signal compression – anywhere up to 30dB.

Failure to compress at the onset of the tone burst results in overshoot up to 30dB higher than the final amplitude. To prevent this overshoot from overloading subsequent amplifiers, clippers are incorporated. These limit dynamic range.

A further advantage of this method of noise reduction is that the system can be bypassed. One obvious way of doing this is using a switch but the note describes more elegant methods. In addition there are expansions on most of the topics covered here and a section on undesirable but avoidable aspects of noise reducers.

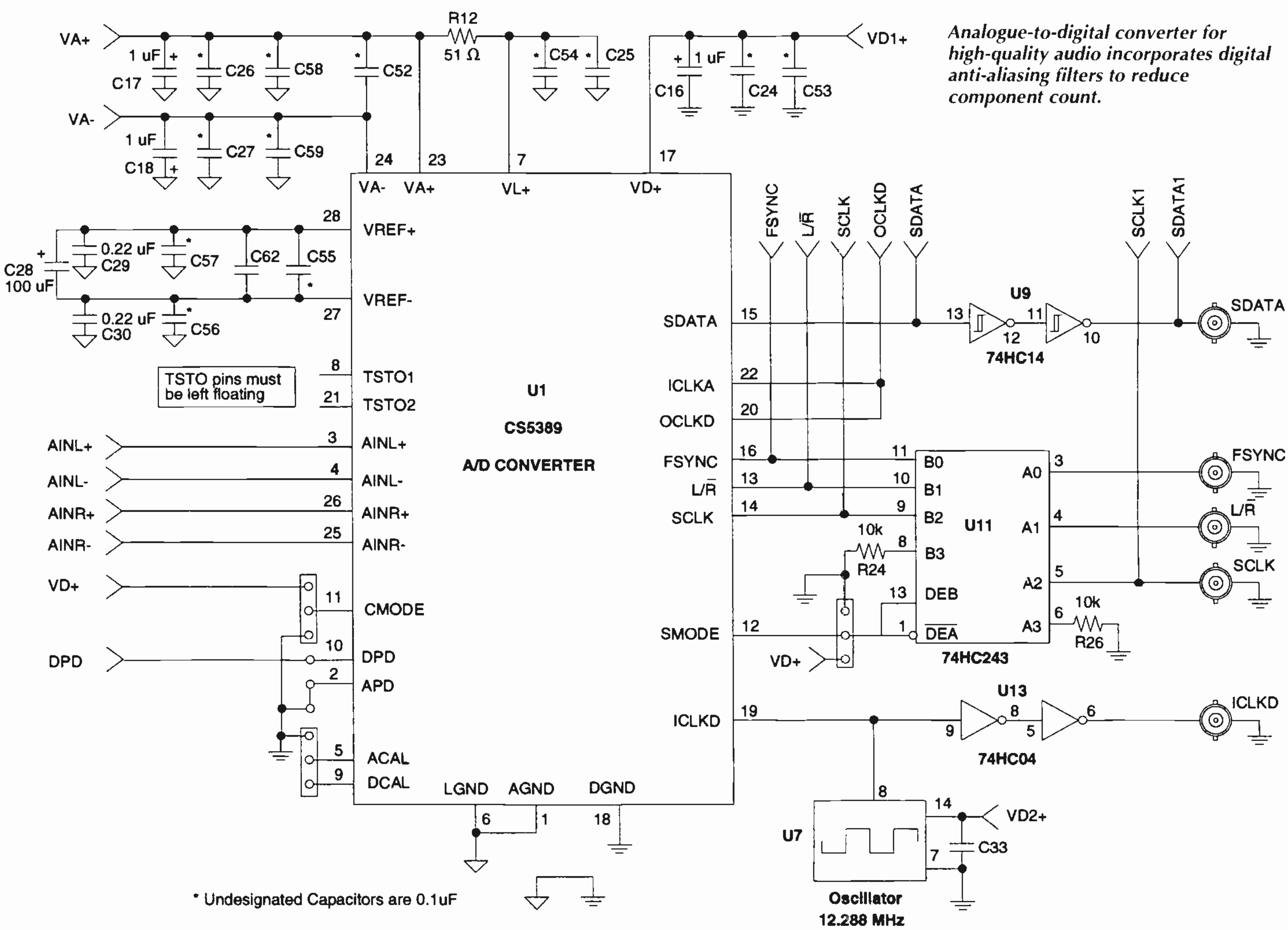
National Semiconductor, The Maple, Kembrey Park, Swindon, Wiltshire SN2 6UT. Telephone 0793 614141, fax 697650.

A-to-D converter for high-quality audio

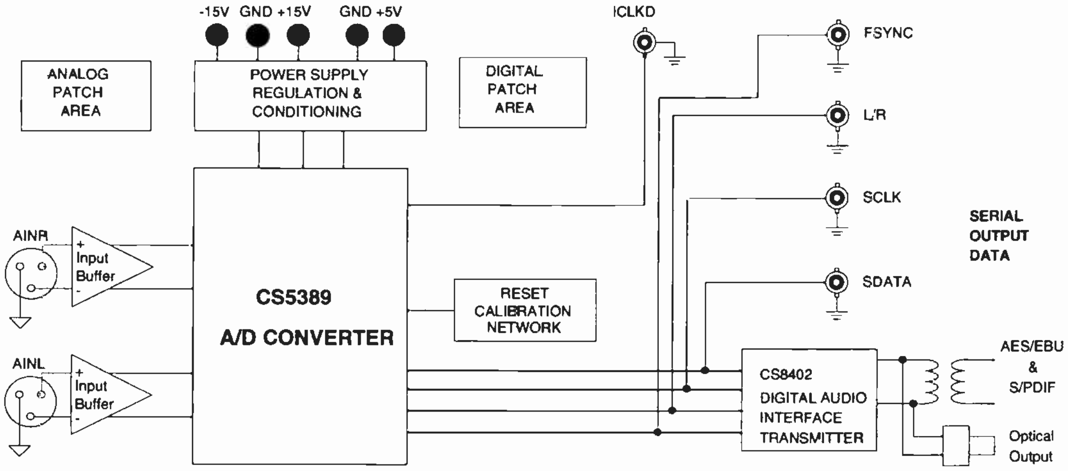
For engineers wanting to evaluate 18bit stereo a-to-d conversion, Crystal has included a complete sampling board in its *CS5389* data sheet.

The device is an integrated device with dual delta-sigma converters, digital linear-phase anti-aliasing, sample-and-hold inputs and a voltage reference. It has an A-weighted dynamic range of 107dB together with a combined noise and distortion figure of 100dB.

Serial data from the two 18bit audio



Analogue-to-digital converter for high-quality audio incorporates digital anti-aliasing filters to reduce component count.

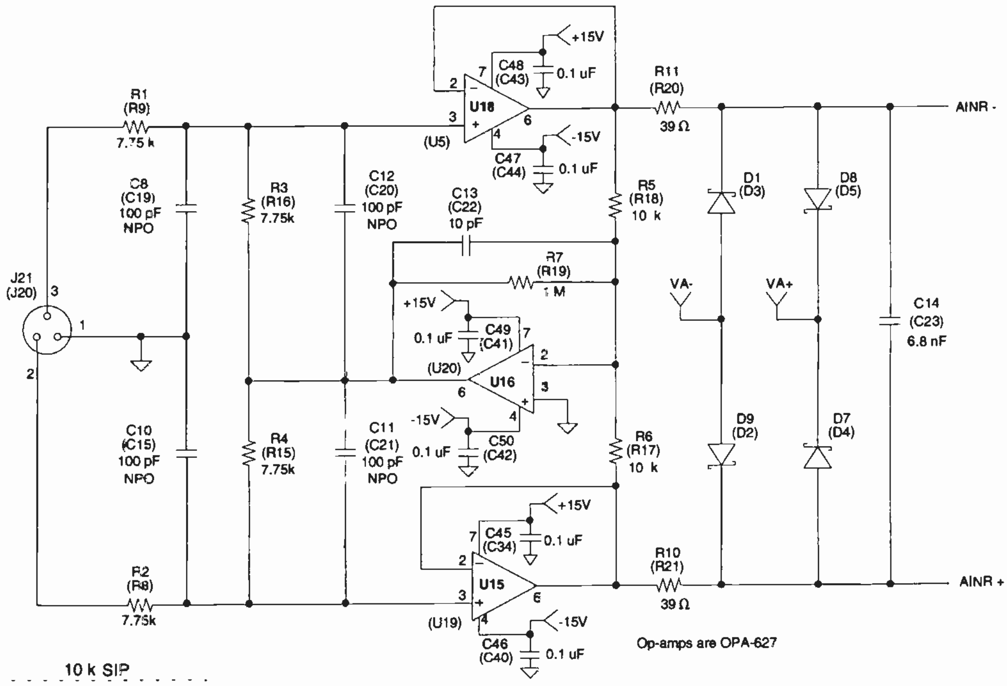


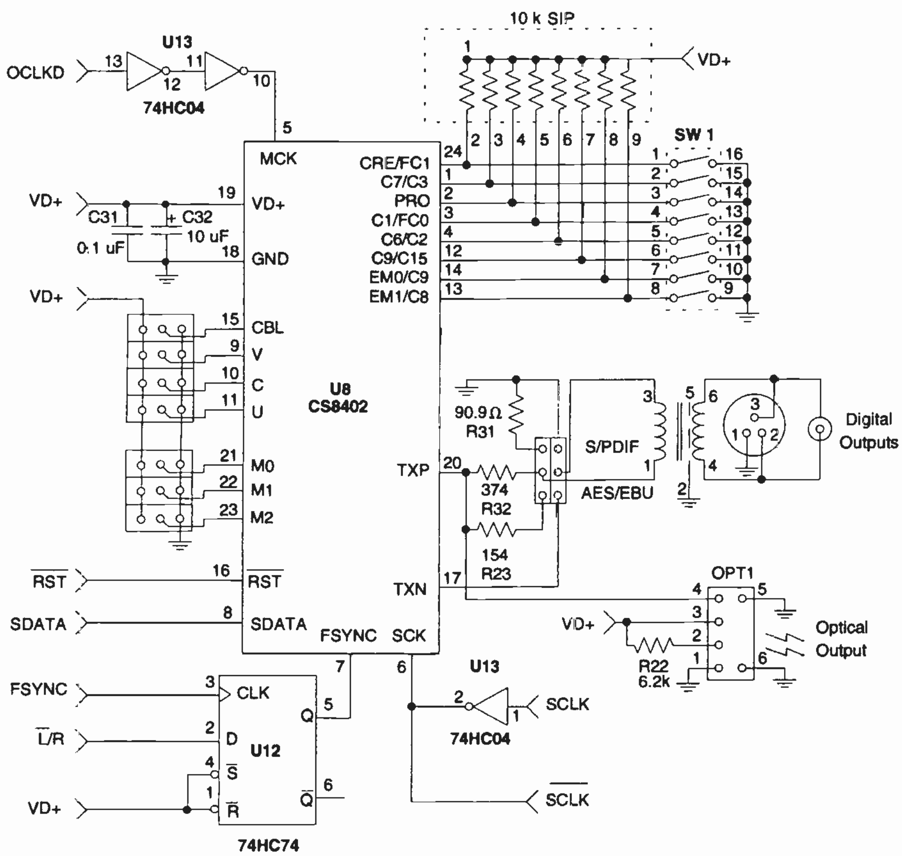
Elements of an evaluation board for the CS5389 18-bit A-to-D converter. A digital audio interface transmitter is included for AES/EBU, S/PDIF and EIJ CP340 compatible audio data.

channels can be output at up to 50kHz. Included on the evaluation board is a *74HC595*-based serial-to-parallel converter designed for interfacing to signal processing systems.

Digital audio interfacing on the board is carried out by a *CS8402*, providing AES/EBU, S/PIF, and EIAJ CP-340 compatible output. There is a pcb layout in the note, accommodating output via XLR, RCA phono and optical connectors.

Input buffering circuitry designed for the board is of general interest. In addition to providing protection, it accepts a differential or single-ended signal of either polarity, and provides a differential output regardless. It also produces the 6dB of attenuation required for scaling to professional audio levels to the device inputs. A nominal input level of 20dBV to the *CS5389* achieves full-





Above is an audio input buffer that turns differential or single-ended inputs into differential output. Below is a digital audio interface compatible with AES/EBU, S/PDIF and EIJ CP340 formats.

scale full-scale digital code.

Common-mode rejection of the system is limited by mismatch of the input buffer components. Resistors $R_{10,11}$ and C_{14} provide anti alias filtering and optimum source impedance for the converter, one channel only. Protection is afforded by the diodes. ■

Crystal Semiconductor, Lymes House, 2 Station Road, Frimley, Surrey GU16 5HF. Telephone 0276 685761, fax 0276 691090.

Looking into real-time

Because of the need for real-time multi-tasking, industrial control systems are often VME based. Implementing this complex bus is usually too costly for students and experimenters but here, William Dickinson* describes a 68020-based OS-9 evaluator providing VME-grade computing power without the usual price penalty.

Motorola's *MC68020* is a high-performance 32bit processor with a 32bit data path. It has an internal instruction cache which is a feature primariiy found in engineering workstations and industrial control systems. Processors from the *68000* family are used in the Apple Macintosh, Atari ST, Commodore Amiga, Nintendo, Sega, and many other commercial systems. They are also at the heart of most real-time industrial control systems.

In cost terms, the combination of *OS-9* and *68000*-family processors competes with many 8 and 16-bit processor alternatives. With a powerful processor at the heart of the system, the advanced features of *OS-9* will reduce the software development costs of many embedded control applications without unduly taxing available CPU resources.

This article describes an open-architecture *68020/OS-9* evaluation platform in the form of a single board computer, or SBC. It is supplied complete with *OS-9/68020* Professional and features an *MC68020* running with one wait-state at 25MHz. The system can be used as a *68020/OS-9* evaluation platform to demonstrate the processing potential of this powerful combination against VME systems at a fraction of the normal costs.

Called *Ωmega-II* this SBC enables you to evaluate the performance of the *MC68020* and *OS-9* within an open-architecture hardware environment. This architecture is free from the restrictions of modern VME designs which are notorious for asics and custom devices.

With the aid of an appropriate porting package, the *Ωmega-II* computer can be used to port *OS-9* to your embedded controller. When mounted in its purpose-designed enclosure, the board complies with BS6527 restrictions regarding EMI.

Building the system

The combined SBC/*OS-9* package is intended

**Windrush Micro Systems*

PC-based OS-9 development for under £700

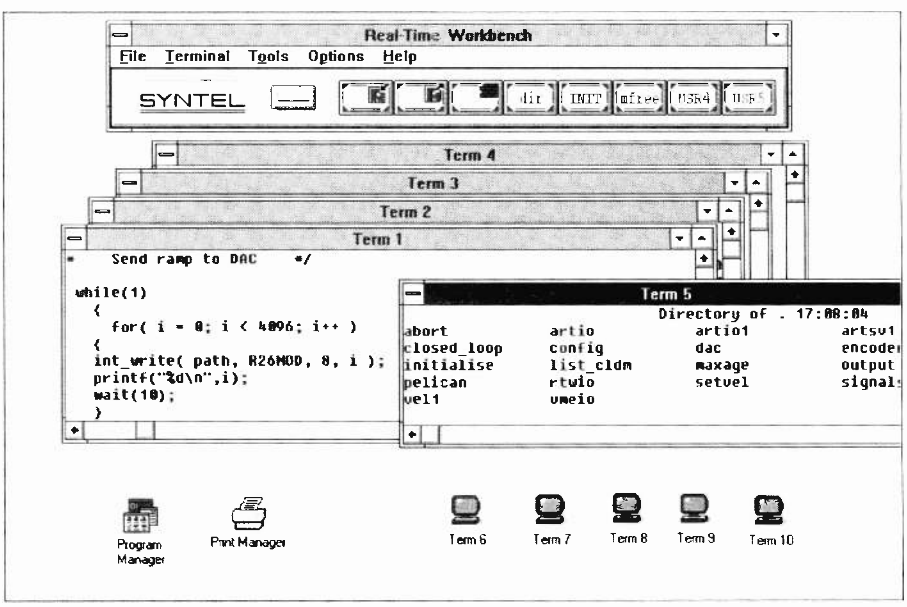
Realising the appeal of Windows as a low-cost user interface, industrial-control specialist Syntel has produced a real-time, multi-tasking system allowing *OS-9* development on the PC.

Communicating with *OS-9* system via Windows has two significant benefits. There is no need to learn how to use special *OS-9*-based tools such as editors and assemblers. In addition, user interfacing hardware and software – the PC and Windows – is cheap and readily available.

In the Syntel system, a *68302*-based processor card with 1Mbyte of ram plugs into the PC. This 'context' card talks to Windows on one side and, via *OS-9*, to the outside world in real-time on the other. It has a multi-protocol networking interface. This reconfigurable interface is set up for Topaz, which is an industrial-control network designed especially for real-time, multi-tasking applications.

Further, the context card has an RS232 interface and supports a range of industry-standard i/o modules. An entry-level development system comprising the *68302* context card with 1Mbyte ram, *OS-9 Professional*, C compiler and Windows *Real-Time Workbench* is £695 – including all manuals. This is not a special price so there are no restrictions on the package. Quantity discounts are available to educational users.

GSM-Syntel is at Victoria Works, Queens Mill Road, Huddersfield HD1 3PG, telephone 0484 535101, fax 519363.



Developing OS-9 on a PC. Without the multi-tasking of Windows, displaying all these terminals simultaneously would require a lot of extra hardware.

for use by hobbyists, students and commercial organisations for personal use or evaluation. For programming and monitoring, the SBC needs a VDU with an RS232 connection or a computer with terminal emulation software. It also needs at least one 3.25in 720K floppy disk drive. Via its SCSI interface, the board supports a wide range of hard disks and tape drives.

Applications ranging from small dedicated embedded controllers to networked factory automation systems can be implemented in *OS-9*. The Professional version of *OS-9* supplied for use with the SBC includes a C compiler, Basic interpreter and Pascal compiler. The *Sculptor* database generator, *Dynacalc* spreadsheet, and *Stylograph* word processor are also part of the package. You can commit the entire operating system and many of the utilities to the 512Kbyte of on-board 32bit wide rom if you wish.

Features

At the heart of the system is a 25MHz *MC68020* MPU with an optional 25MHz *MC68882* floating point co-processor (FPCP). The floating point co-processor is closely coupled to the *'020* and executes its instructions transparently.

Attached to this dual-processor core is 1Mbyte of battery-backed static cmos ram which is 32bit wide. This memory may be expanded up to a 4Mbyte limit. It requires one wait-state, resulting in a 160ns bus cycle which fully supports read-modify-write bus cycles.

External circuitry monitors the power supply rail. When the supply falls below a preset threshold an unmaskable interrupt is generated. Access to the ram is inhibited several milliseconds later. This prevents the random signals from the processor on power failure from corrupting the memory. It also makes it possible for memory to be used for a battery-backed ram disk, among other things.

Rom in the system may take one of two forms. The first is a simple eight-bit wide format offering ease of use at the expense of performance and a 128Kbyte upper limit. Alternatively four 128Kbyte roms may be used in a 32bit wide configuration providing a total of 512Kbyte. This option illustrates the dynamic bus sizing capabilities of the *'020* which allow it to operate with 8bit, 16-bit and 32-bit wide data paths on a cycle by cycle basis.

Most of the memory decoding is carried out by four pals. The first of these deals with function codes and decodes the co-processor and interrupt acknowledge cycles produced by the main processor. A second decodes the memory map, a third the 'byte write' resources of the processor. The fourth takes care of data size acknowledge to terminate bus cycles.

I/O resources

All of the resources required for a multi-user *OS-9* development system are provided by *Ωmega-II*. Included are five RS-232 serial ports, a parallel printer port, a high speed parallel port, a floppy disk controller and a SCSI controller for hard disks.

An *MC68901* multi-function peripheral device is used. It provides a serial port with programmable bit rate generators and several timers. There are also eight inputs that can generate vectored interrupts on positive or negative edges. The device is used for the standard *OS-9* terminal port and generates vectored interrupts for the floppy and SCSI disk controllers. Additionally, it provides the system 'tick' used by the pre-emptive *OS-9* scheduler and system clock. Up to 16 vectored interrupts at level six can be set, depending on the source.

A second pair of RS-232 serial ports is derived from an *MC68681* dual asynchronous receiver/transmitter. These ports are polled in the interrupt handler to determine the source and nature of the interrupt, for example which channel and whether it was a receive or transmit interrupt.

Serial ports three and four are also derived from an *MC68681* dual receiver/transmitter. Port four may be configured as an RS-485 multi-drop transceiver using a twisted pair. The single vectored interrupt via the multi-function peripheral device is then 'polled' by the interrupt handler as for the other port pair.

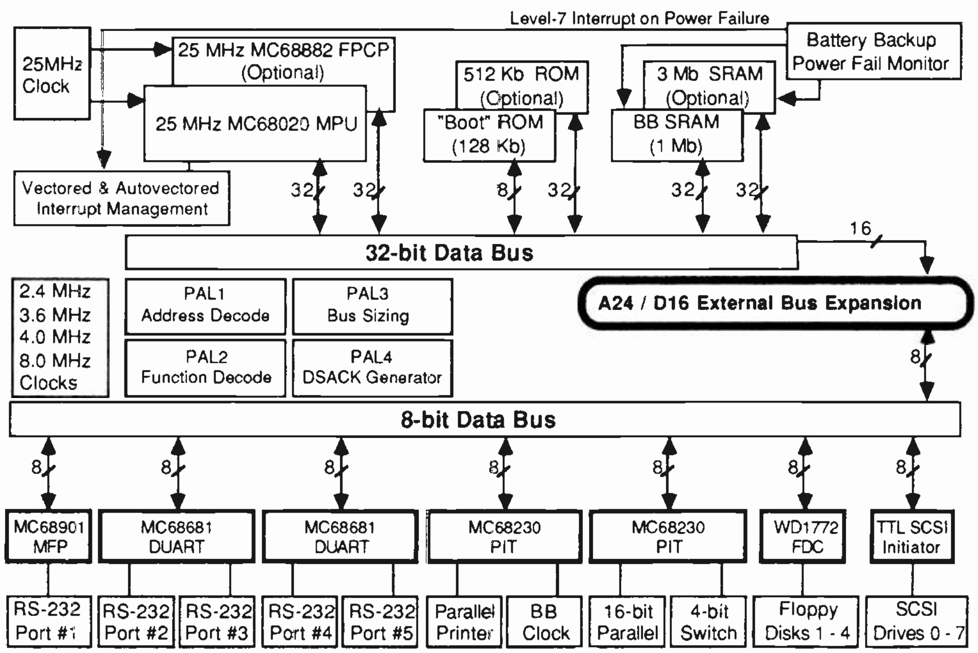
A 16-bit high-speed parallel interface complete with four edge sensitive handshakes for generating vectored interrupts is provided by an *MC68230* parallel interface/timer. The parallel port can interconnect two computer systems or used to implement a simple i/o system comprising lamps and switches. Alternatively, it allows experimentation with interrupt handlers within *OS-9* device drivers. A four-bit configuration switch interfaced via this device selects up to five vectored interrupts.

A further bidirectional parallel interface with strobe and acknowledge allows a Centronics

Ωmega-II specifications

- 25 MHz *MC68020RP25* processor
- optional 25 MHz *MC68882RC25* floating-point co-processor
- 8bit, 128Kbyte rom system monitor (*OS-9* boot rom from Cumana)
- Expansion to 32bit path and 512Kbyte optional
- 32bit wide ram (1Mbyte battery-backed ram, one wait state)
- Expansion capability to 4Mbyte
- Vectored interrupts from all i/o devices.
- Power-up/down sequencer to protect battery-backed ram and clock
- Level seven, non-maskable interrupt, generated 5ms before memory shutdown
- Battery backed clock calendar
- Floppy-disk controller for up to four drives
- Simple TTL hardware SCSI initiator for optional hard disks and tape drives
- Five RS-232 serial ports for VDUs (users) or printers
- Centronics parallel printer port
- Bidirectional 16bit parallel port with four handshakes
- CPU readable 4-position dil switch to control system start up, etc.
- A_{23}/D_{16} and two interrupt levels available for prototyping i/o hardware designs
- *OS-9/68020 Professional* V2.3, with C compiler, editor, assembler, linker, debugger, Basic, *Pascal, Sculptor, Stylograph and Dynacalc.*

Elements of the Ωmega-II development and evaluation platform. Within the price range of universities and experimenters, this board combines the industrial control power of the 68020 and OS-9 but avoids the cost penalty of VME.



Real-time, multi-tasking *OS-9*

Specifically for 68000 family microprocessors, *OS-9* is a high performance multi-tasking operating system. It combines new operating system concepts and real-time capabilities with the overall architecture of the popular Unix operating system.

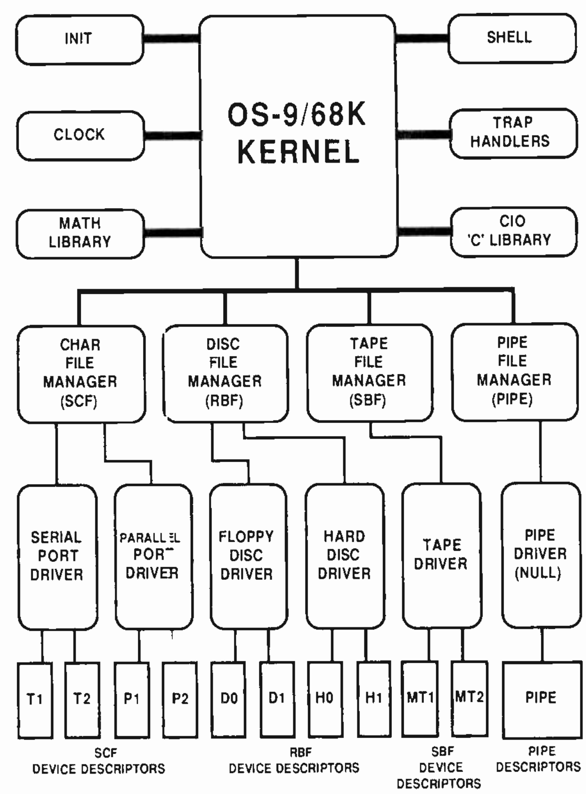
However, *OS-9* is much smaller and far more efficient. It provides key Unix features such as a tree-structured file system, device independent i/o and full multitasking facilities. In fact, most Unix applications software written in C can be easily ported to the OS-9 environment.

All speed critical portions of *OS-9*, including the kernel, file managers and device drivers, were hand-coded in *68000* assembly language. As a result, system performance does not suffer from compiler-generated code. This translates to greater timesharing user capacity and faster real-time response.

Standard system-wide library modules used by programming languages and applications software are provided by *OS-9*. The maths module includes long-integer maths, IEEE format single and double precision floating point maths, radix conversion functions and a complete set of transcendental and trigonometric functions.

OS-9 comprises several independent self-linking named objects called memory modules. A directory containing the name and address of each module is automatically maintained by the system executive. This is coordinated with the memory management system so multiple tasks can automatically share common program or data modules which vastly improves memory utilization. For example, two users on a timesharing system who are both using Basic would automatically 'share' a single copy of Basic rather than wasting memory with redundant copies.

Thanks to *OS-9*'s modular structure the system can be reconfigured without access to source code by simply adding or removing modules as desired. For example, if you need to add a new type of disk drive to the system you only have to add the appropriate driver and descriptor. Similarly if you want to build a rom-based system just omit the disk file manager, disk driver modules, and other unnecessary modules.



Organization of OS-9's major component modules. This real-time, multi-tasking operating system is reconfigured by simply adding or removing modules.

printer to be connected. Battery backed clock calendar information is provided by an OKI *MSM58321* while floppy disk interfacing is carried out by a Western Digital *WD1772-02*. Hard disk drives and tapes connect via a TTL-based SCSI initiator.

There are seven interrupts. Levels one to six are vectored by the i/o devices. Interrupt level seven has autovectoring and is used for the floppy disk controller and power fail signals. There is also a fully buffered bus expansion port for user hardware and experiments via a DIN connector. Several i/o boards including nine serial ports, colour graphics adaptor, ARCnet, IEEE488 and SCSI-2 are available from Windrush.

The version of with *OS-9* supplied for use with *Ωmega-II* is fully configured for each of the five serial ports, Centronics parallel port, clock calendar, floppy and SCSI disk controllers. Sources of all device descriptors are supplied and sources of all device drivers are available separately.

Product sourcing

A six layer SBC printed circuit board and complete kit of parts for *Ωmega-II*, excluding VDU, printer, floppy disks, hard disks, tape drives and *OS-9*, is available for £750 excluding postage and VAT from Windrush Micro Systems.

Comprehensive hardware documentation, schematics, PAL source equations, parts list, etc., are supplied with the kit. The SBC, its documentation pack and the PALs are available separately. A fully-finished metal enclosure is also available.

Measuring approximately 12 by 14in, the board requires an external power supply providing +5V, +12V and −12V plus whatever is needed for disk drives. A 55W switch mode PSU should suffice. The Ωmega-2 hardware is available from Windrush Micro Systems, Station Road, Worstead, Nr Walsham, Norfolk NR28 9SA, Telephone 0692 404086, Fax 404091.

The software package mentioned is available from Cumana for £98 excluding shipping and VAT. Due to its low price, the software is not eligible for hot-line support nor is it eligible for upgrade to current or future releases. Cumana is at Pines Trading Estate, Broad Street, Guildford, Surrey GU3 3BH, Telephone 0483 503121, Fax 0483 503326.

Floppy and hard disk drives are also available from Cumana. Microware UK is at Leylands Farm, Nobs Crook, Colden Common, Winchester, Hampshire, SO2 1TH. Telephone 0703 601990, Fax 601991. ■

Further reading

MC68020 & MC68EC020 microprocessor user's manual, first Edition 1992, publication M68020UM/AD, £6.60 from Motorola European Literature Centre, Milton Keynes.
MC68881/882 floating point coprocessor user's manual, second edition, publication MC68881UM/AD/REV-2 1989, £16.61 from Motorola European Literature Centre.
The OS-9 catalogue, 1992/3 edition, Microware.
OS-9/68000 technical manual, publication OST68NA68MO, price £64 from Microware.
P. S. Dyan, The OS-9 Guru, ISBN 0-9519228-0-7, publication OS9GURUNAMO, price £27.00 from Microware.
P. Dibble, OS-9 Insights, ISBN 0-918035-03-1, publication INS68SE68MO, price £50.00 from Microware.

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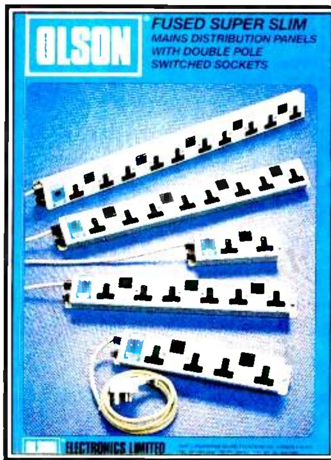

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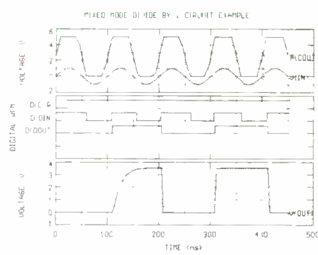

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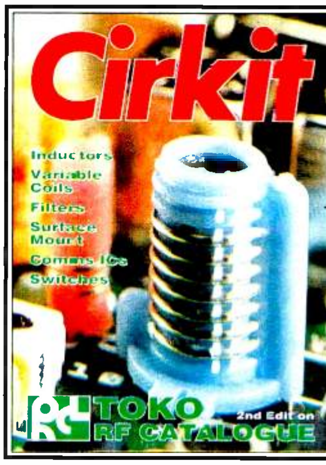

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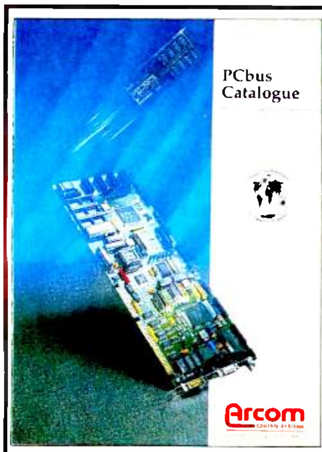

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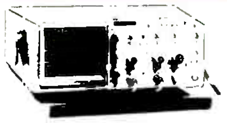

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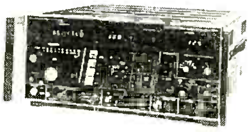

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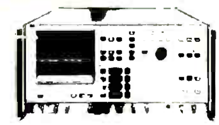

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PLEASE NOTE: ALL OUR EQUIPMENT IS NOW OPERATION-VERIFICATION TESTED BEFORE DESPATCH BY INDEPENDENT LABORATORY

We would be pleased to handle all grades of calibration or NAMAS certification by same laboratory at cost price. All items covered by our 90-day parts and labour guarantee and 7-day 'Right to Refuse' (money back) warranty.

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CIRCLE NO. 130 ON REPLY CARD

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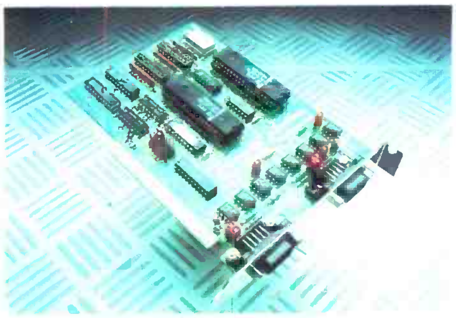
Dual port serial communications for your PC



The PC 47,48 and 49AT boards each provide two independent serial ports for any PC/XT/AT(ISA) computer. RS232, RS422 and RS485 standards are supported and all combinations are possible with this range of low power boards.

9 pin D connectors are provided for the RS422 and 485 ports. RS232 can be connected via 9 or 25 way D connectors and charge pump circuitry ensures signal levels in excess of 7V are transmitted.

Interrupt and base addresses are independently selectable for each port, full, half duplex and multidrop communications are fully supported and all boards use the industry standard 82C450, UART.

Designed and manufactured by Amplicon Liveline each board is supplied with a comprehensive technical manual and interrupt driven device driver software is available.

Write in number 1

750kHz professional data acquisition with Windows support

PC226 from Amplicon Liveline provides 16 true differential 12 bit analog inputs each with dynamically programmable gain. PC226 has sample rates up to 400kHz multichannel and 750kHz single channel, programmable scanning hardware, flexible triggers and a 2048 sample FIFO to ease programming in high speed applications.

PC 226 is supplied with menu driven software for DOS and Windows, LabTech Notebook drivers and comprehensive well documented libraries for most popular high level languages.

Optional software support for Microsoft Windows includes a complete Dynamic Link Library (DLL) and two icon driven packages, Signal Centre for signal capture and analysis and TRACS for process monitoring, visualisation and control.

Write in number 2

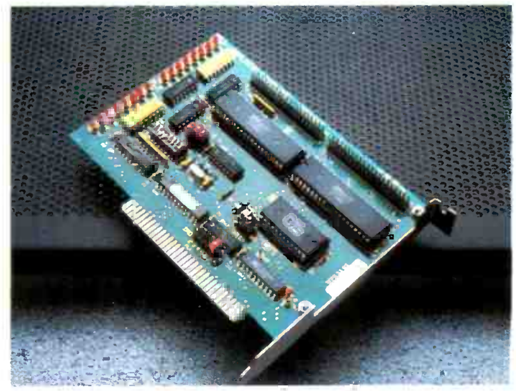
48 lines TTL I/O with 16bit counter timers

PC 14AT available from Amplicon Liveline is a high quality general purpose interface board for PC/XT/AT(ISA) computers.

The board has 48 lines of TTL compatible digital I/O provided as six 8 bit ports four of which can be programmed to be all inputs or all outputs and two which can be split to be 4 inputs and 4 outputs if required. There are also three 16bit counter timers which can be used to generate and measure pulses at up to 5MHz.

An on board 4MHz crystal oscillator is also included on PC 14AT along with LEDs showing the status of a selection of the I/O ports, these can be used as programming and debugging aids.

PC 14AT is suitable for a wide variety of interface, monitoring and control applications



and an LP (low power) version PC 14LP is available for installation in laptop computers, both boards have flexible base address and interrupt support. **Write in number 3**

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