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Self on Class-G audio power

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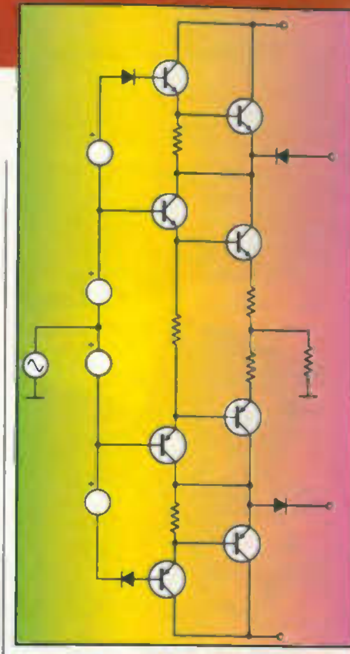
CONTENTS



Doug Self investigates Class-G audio power. It certainly offers benefits in terms of efficiency, but can it compete on the linearity front? Find out on page 900.

This 433.92MHz transceiver provides bidirectional wireless communication at ranges up to 200m and runs from a 3.3V supply. There's a full description of this and other essential new products starting on page 937...

Exclusive reader offer

Complementing our article on Anadigm's programmable analogue array is a special offer on the company's development kit. It's normal price is £349, but for a limited period, EW readers can obtain it for £249 (excluding VAT). Turn to pages 914 & 917 for more...

January issue on sale 6 December

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ISSN 0959-8332

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Igniting the spark and fanning the flames

One cold autumn afternoon, when I was five, I came within a cat's whisker of killing myself. I'd been kept home from school with a minor illness and was playing upstairs on my own.

There was a 2kW electric bar fire blazing away in my bedroom. By that age, I already knew that metals conducted electricity, but I didn't really understand the consequences of mains electricity. So as an experiment, I removed the fire's plug from the socket, wrapped a padlock chain around its pins, and rammed it back in.

If I close my eyes now, I can still see the dazzling blue, hemispherical flash, fringed with brilliant orange sparks. Still, I can hear the sound – a surprisingly soft popping noise, like a balloon bursting under a duvet.

Apart from a few stern words, I was never punished for my 'experiment'. In retrospect my parents probably thought I had punished myself enough. In any case, I suspect my father had a sneaking regard for such a bold experiment (as a boy in Poland he had blown the cellar door off its hinges making gunpowder).

I already had a love of science. My father was an engineer who encouraged my interest. Much of my early childhood was spent in my father's garage, amidst a plethora of tools, car batteries, bits of old televisions, model steam engines, dynamos, valves, tobacco tins crammed with nuts and bolts, bottles of mysterious and wonderful-smelling chemicals, electric motors, relays, bulbs, strange actuators filched from scrapped aircraft, timber, sheet metal, hardboard, paints, varnishes and Holt's Cataloy.

I would sit on the floor, playing with wires, while he made a rotary saw from an old school desk, a washing machine motor and a fan belt. The thing was lethal and made a hideous noise but it cut wood like a knife through butter. He taught me to solder when I was nine, and together we built my first short wave radio.

These early influences stood me in good stead during secondary school and university, when science got tougher and more and more maths crept in. Somehow I knew that beyond the slog, the essential wonder of science remained, and all the techniques that I found difficult were just so many tools in helping you achieve something really worthwhile.

In order to write this leader, I circulated a questionnaire to the students in my department here at UMIST. It asked them about their impressions of science and scientists while they were at school, and why they chose to pursue science or engineering at university.

Individual experiences aside, there was a considerable degree of conformity among the replies. First, most students chose science because they liked it – not necessarily because it improved their career prospects. Second, the reasons most people gave for students avoiding science and engineering was because it was perceived as being hard.

All agreed that teachers played a crucial role in promoting an individual's enthusiasms: almost all thought that while science was well-taught at school within the bounds of the national curriculum, not enough was being done during the early stages. And herein lies the kernel of my argument.

As any parent knows, the younger the child, the more readily he or she will respond to encouragement. The inverse applies equally – dishearten a child in the early years, and you will have a mountain to climb later.

I like to think this can be encapsulated in a simple equation that is no doubt inaccurate but which serves the point – 'effect equals encouragement divided by age'. It follows that if a child has a gift or spark for any subject, and encouragement is given, his or her enthusiasms will buoy them up through the turbulent waters of formal education.

Although there are clear implications here for our primary school system, the roots of the issue lie elsewhere – in the home. Perhaps some children, who would otherwise blossom as scientists and engineers, do not receive sufficient support in the domestic environment. This may be because the perceptions of science and scientists amongst the general public is at best ambiguous; science is certainly not considered trendy or fashionable – as indicated by the questionnaire.

Many view science with suspicion, regarding scientists as remote, aloof and even arrogant. It must be said, there is some truth in this argument. We, the scientific community, have some work to do to put our house in order.

Earlier this year, the Government commissioned an independent review into the supply of scientists and engineers, chaired by Sir Gareth Roberts FRS, published on June 21 (www.hm-treasury.gov.uk/docs/2001/scientists_2006.html). This review, a consultation document, was commissioned "in response to concerns that innovative businesses in the UK sometimes find it difficult to recruit the skilled researchers they need".

In reply to this, the Institution of Electrical Engineers (IEE) submitted a detailed reply to the Government (www.iee.org/Policy/Submis/s589.cfm) outlining the reasons it believed the problem existed. One of these was the desperate shortage of qualified and high quality maths and science teachers. Even more worrying, the UK has done poorly in comparison to other countries in attracting women into science careers. Although some progress was made, recent statistics show that the situation may have regressed.

These days my thrills come in the form of abstruse equations having a very real effect on the digital audio signals processed in our labs. But these connections are abstract, and do not appeal to a child's mind.

There is no question that schools play a vital role in addressing this issue, but so do parents, and probably more so. Maybe we need to take risks again, and show young people what science can really be like. I'm not suggesting that we take a party of tots down to the local swimming pool, sit them in the balcony and hurl a chunk of potassium into the water, but you know what I mean.

Science can be thrilling, and we need to prove that to our youngsters. ■

Patrick Gaydecki

Electronics World is published monthly. Orders, payments and general correspondence to **Jackie Lowe, Cumulus Business Media, Anne Boleyn House, 9-13 Ewell Road, Cheam, Surrey, SM3 8BZ**.
Newstrade: Distributed by COMAG, Tavistock Road, West Drayton, Middlesex, UB7 7QE Tel 01895 444055.
Subscriptions: Quadrant Subscription Services, Oakfield House Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 01444 475631. Please notify change of address.
Subscription rates 1 year UK £36.00 2 years £58.00 3 years £72.00. Europe/Eu 1 year £36.00 2 years £82.00 3 years £103.00 ROW 1 year £61.00 2 years £98.00 3 years £123
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, 475 Park Avenue South, 2nd Fl New York, NY 10016 Tel; (212) 679 8888 Fax; (212) 679 9455
USA mailing agents: Mercury Airfreight International Ltd Inc, 10(b) Englehard Ave, Avenel NJ 07001. Periodicals Postage Paid at Rahway NJ Postmaster. Send address changes to above.
Printed by Polestar (Colchester) Ltd, **Filmsetting by** JJ Typographics Ltd, Unit 4 Baron Court, Chandlers Way, Southend-on-Sea, Essex SS2 5SE.
Cumulus Business Media is a division of **Highbury Communications PLC**

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Buckyballs act as superconductors at –156°C

Pictured examining organic crystals are, left to right, Bell Labs scientists Hendrik Schon, Christian Kloc and Bertram Batlogg.

Buckyballs, football shaped carbon molecules C_{60}, can act as high-temperature superconductors, claim researchers from Bell Labs, when they trap other organic molecules.

Carbon C_{60} has shown superconducting behaviour before, but only at temperatures up to 52K, or –221°C.

Bell Labs has raised the critical temperature to a balmy 117K, or –156°C. This is significant, because transistors and other devices made using the materials can be cooled with liquid nitrogen, rather than much more expensive liquid helium.

"This shows that buckyballs may live up to their initial promise of being a material that will be very important to technology," said Federico Capasso, vice-president of physical research at Bell Labs.

To increase the superconducting temperature, physicist Hendrik Schon and his team inserted molecules of either chloroform or bromoform into the lattice structure of a buckyball crystal.

Increasing the spacing of the buckyball molecules lowers electrical and molecular attraction and increases the superconducting temperature.

"I'm surprised; I didn't expect the temperature to go up so much," said Professor Peter Littlewood, head of theory of condensed matter physics research at the University of Cambridge. "It's a very clean result."

The only materials that exceed 117K are copper oxides. However the physics for these materials is far more complex. Buckyballs could make cheaper transistors that are easier to work with.

"This result makes buckyballs infinitely more interesting to study," added Capasso.

Data hides in sound

Researchers in Cambridge are inaudibly hiding control data in sound, and toys may be the first application.

"We are negotiating for toys in 2003," said John Edgley co-founder of Intrasonics, the company spun out of Scientific Generics to exploit the technology. Toys could be in the shop in two years, he said.

The data-encoding technique allows data up to 20bit/s to be added to audio signals.

"It uses direct-sequence spread spectrum to make the data look like white noise and psycho-acoustic techniques to hide it," said Edgley.

Enabled toys will react to signals in the audio making them appear to dance to music for instance, or move in response to TV programmes.

According to Edgley, an extension of the technology, using four loudspeakers, enables a receiver with a single microphone to locate itself physically to within two inches in an encoded sound field.

This would enable, for instance, a whole audience equipped with enabled devices to be sent different data.

www.generics.co.uk

To work with its flexible displays, display company E Ink is developing flexible transistors at its new 900 square metre microelectronics technology group facility in Woburn, Massachusetts.

New software tests electronic kit remotely

Communications specialist Esgem from Bristol has developed test software for distributed network systems that can isolate problems when products are in the field.

By including extra hardware and software probes, a product can be accessed through the network when in use.

Called Assayer, the tool is aimed at products such as mobile phones, home entertainment networks and automotive systems.

"Testing is part of the whole life cycle of a product – including after you've shipped it," said Pete Moore, Esgem's product manager. "We think you need to be able to test when devices are in the field."

Moore cites a home entertainment system, which might be working perfectly until a new node, such as a DVD player is added. However, when the system crashes, the DVD machine might not be at fault. Instead another device, working close to the limit, was the cause.

Such problems need to be identified in the field to find the offending equipment.

Assayer consists of test-bed controller software that, through the network, interrogates the devices on the network.

"Testing is actually something that's part of the design flow, it's not something you do at the end," pointed out Stephen Maudsley, CEO of Esgem.

Via the Internet protocol network, or whatever connects the products, the probes could look at individual electrical signals, through to snooping a bus, or downloading application data.

Even if only one device in a network has been designed with Assayer probes, it can often figure out which other device is causing a problem, said Maudsley.

Will FPGAs kill off Asics?

Programmable logic firm Altera has formed a link with The MathWorks and its Matlab tools to improve digital signal processing in its field-programmable gate arrays.

Altera has developed a tool called DSP Builder, which can take designs from Matlab and Simulink and convert them into a form suitable for Altera's Quartus software.

"We want to make our design flow much more like the DSP designer is used to," said Paul Hollingworth, marketing director at Altera.

DSP Builder carries out the conversion, creating code in the VHDL hardware description language.

But, "designers need know nothing about VHDL or Verilog", claimed Hollingworth.

The move is driven by the increasing use of DSP in wireless communications systems. "The move from 2G to 2.5G to 3G adds a lot of data processing to voice. It's causing the DSP requirements to go up exponentially," Hollingworth said.

GSM requires about 400Mflops of processing, but full speed 3G will need over 12000Mflops. "That's a spectacular increase," said Hollingworth.

FPGAs are more suited to this processing than dedicated DSPs, he said, especially in terms of cost. "It's not to say we're going to take over the DSP business, but for many applications we can be more efficient," Hollingworth said.

Moreover, he reckons the recession is causing firms to shy away from Asics and the huge up front charges they bring.

"In five years' time the Asic market as we know it won't exist. We're just starting 0.13μm in the fab and non-recoverable engineering costs are three-quarters of a million dollars," he said.

This is a hand-cranked generator for charging mobile phones.

Developed by clockwork radio company Freeplay in conjunction with Motorola, it is expected to provide between three and six minutes of talk and several hours of standby time for each 45 second hand-cranking session, claims Freeplay.

Neither company wants to reveal exactly what is inside yet, although Electronics World has been told that there is no energy-storage spring.

Initially it will be aimed at Motorola phones. "A wireless phone user, using the new accessory, will never again be faced with a situation where he or she can't make a wireless phone call because their battery is dead and they have no access to an outlet," said Gary Brandt, business director in Motorola's Personal Communications Sector.

It should be on UK shelves before Christmas with the two companies sharing global distribution.

This gadget from Seiko may be for you if you think in diagrams and use a PDA. SmartPad hides a page-sized digitiser under the paper and copies sketches to your Palm or clone. For text entry there is an alpha-numeric keyboard to peck at under the pad.

Strong magnetic fields make rats walk in circles

Strong magnetic fields can alter rat behaviour and may affect humans, according to two Florida scientists after experiments on rodents.

Thomas Houpt and James Smith claimed the animals not only develop strong taste aversions after exposure but also walked in circles for a time.

The field used by Houpt and Smith was 9.4 teslas – not much more than that used in hospital MRI scanners, and exposure lasted 30 minutes.

The effects were that the animals, which enjoyed drinking sweetened water before the experiment, would not drink it after exposure. Control rats did.

In addition, rats oriented toward the south pole without exception walked in anti-clockwise circles and those facing the north pole turned clockwise. All settled down soon, and circling behaviour diminished to nothing after several exposures.

Medical MRI scanners used in hospitals generally work between 0.3 and 1.5T, with some operating at 4T. More powerful machines are proposed, at 17T the health of individual cells can theoretically be determined.

According to researchers, MRI-scanned patients don't complain of aversions to food after scanning – even at 3T, but there is some anecdotal evidence that research workers operating near powerful magnets have suffered nausea.

Subsequent analysis of test rat brains showed high neural activity in the stomach, intestines and inner ear control regions of the brain.

Low-cost InP chip provides alternative to erbium-doped fibre amps

An optical chip that amplifies multiple wavelengths of light has been developed by British start-up Kamelian.

Based on Kamelian's indium phosphide (InP) semiconductor optical amplifier technology, the optical linear amplifier (OLA) could replace expensive erbium-doped fibre amplifiers in certain applications

"There is an opportunity for a low-cost replacement for an EDFA in a metro environment," said Paul May, CEO of Kamelian.

A metropolitan area network will usually make use of fewer wavelengths of light than longer haul systems, perhaps eight or fewer. An OLA is smaller and cheaper to produce than an EDFA, the latter being more suited to long distance transmission.

Kamelian's OLA has variable gain, dependent on an injected current. Normally the amplifier would be designed into a closed loop system to maintain a consistent power output.

"We can vary the level of gain easily," said May. "Typically a system designer will want a certain output power."

The OLA will be produced at Kamelian's Oxford fabrication plant, which is close to coming on stream. "It's unique. Not many companies have growth, processing and packaging of indium phosphide," said May. "That puts us in a very strong position."

Paul May, Kamelian's CEO

Java for the masses

Croydon-based OneEighty Software has created a Java virtual machine (JVM) that can run on cheap $1 processors such as the venerable – yet popular – 8051.

The software, called Origin-J, is aimed at bringing Java to low cost systems such as smartcards, domestic appliances and Internet gateways.

"Origin-J is a very, very compact full implementation of Java," said Peter Dzwig, the firm's chief executive.

OneEighty has created a clean-room version of the JVM that can run on almost any processor, including the 8051.

"Origin-J on an 8051 comes in at between 40k and 50kbytes," said Russel Winder, the firm's chief technology officer and formerly professor of computing science at King's College, London. No other company, he believes, can run a complete JVM on so simple a processor.

"Performance is slightly slower than native C, but you would expect that," added Winder. "We're not trying to compete against ARM's Jazelle."

The software has already been ported to several processors including the 68HC12, ARM and i386. It also runs on Cyan Technology's 16-bit XAP processor.

Porting to other processors can take just hours for a simple von Neumann architecture or a couple of weeks for a more complex Harvard architecture.

Multi-threading allows several tasks to be run concurrently, without running multiple instances of the software.

***Artificial personality...** In an attempt to make an easy-to-use computer interface, Carnegie Mellon University is developing Vikia.*

'She' is an artificial personality being jointly created by the University's Robotics Department, the Human Computer Interaction Institute and Entertainment Technology Center. Vikia has, apparently, already been given her own personal history as the first robotic student at the university.

Voice recognition keeps tabs on young offenders

Young offenders released into the community will have a new security check in the future – voice recognition.

Rather than keep tabs on offenders with personal visits by parole officers and tagging, voice recognition will allow their whereabouts to be monitored by telephone.

Manchester-based systems firm On Guard Plus has integrated voice recognition technology into a security system for monitoring offenders in the community.

Called Vqvoice, the technology could be used, for example, to confirm that an offender has turned up for work, or is at home during curfew hours.

"Vqvoice is already being used by Securicor as part of the Youth Justice Board's intensive supervision and surveillance programme," said Stephen Freathy, business development manager at On Guard Plus.

Tagging is the traditional method of confirming location and is also used, but this only registers a presence near the transponder, usually at the home address.

"Voice verification allows the offender to be tested for his presence at various locations," said Freathy.

"It's not a replacement for tagging, but they could be used in conjunction."

Vqvoice is based on technology developed by Belgian firm Keyware. A biometric server enables multiple identification algorithms to run, such as fingerprint, voice or iris scan.

The voice recognition uses algorithms developed by Lernout & Hauspie, but any other algorithm could be included.

The error in the system, either falsely rejecting or passing, is about two per cent, so the password or phrase is normally repeated, cutting errors to 0.04 per cent.

This very humanoid robot has gone on sale in Japan. Developed by Fujitsu, the HOAP-1 weighs 6kg and stands half a metre tall. It is aimed at development of motion control applications such as two legged walking. Fujitsu hopes to sell a hundred of the robots in the next three years – with the price described as 'open'.

Words on wheels

Adflash, a Devon-based company, has spent five years developing WheelFX, which uses LEDs synchronised to the rotation of a wheel to display messages – from five to over 200mile/h said the company.

WheelFX has been adopted by film star Paul Newman and the Newman-Haas Racing Team to advertise on their car wheels and got its first outing in Britain this year at the Rockingham Motor Speedway in the CART FedEx Championship.

Formula 1 teams are said to be interested and taxi drivers need not be left out as Adflash is working on a version for them called 'Taxsee'.

Niobium looks set to replace tantalum for capacitors

Restrictions in the supply of tantalum, and environmental and human tragedies linked to its Congo sources, are turning capacitor makers to niobium as an alternative.

Technical difficulties have held up the introduction of niobium into commercial capacitors. The higher dielectric strength of its oxide should make for more compact capacitors, but leakage current and temperature stability are inferior to tantalum. And niobium capacitor production is not so well understood.

As it is, niobium capacitors tend to be slightly larger than the tantalums they replace.

The latest company to introduce niobium capacitors is Vishay which is sampling solid niobium capacitors in the industry-standard 293D, 292D, and 595D form factors, from 4 to 16V and 4.7 to 680µF.

New technique knocks spots off FFT

Pipeline Frequency Transform could soon be the expression on the lips of spectrum analyser makers and radar scientists.

Developed on the Isle of Wight by start-up company RF Engines, PFC as it is called is a hardware digital signal processing (DSP) technology that knocks spots off fast Fourier transforms and uses far less hardware than conventional multi-channel digital filters – or so claims its inventor John Lillington.

The technology "gives exactly the same output as a bank of filters", said Lillington, who is CEO of RF Engines, "but 16 384 filter modules are needed to filter 16k channels with conventional filtering. PFT needs only 14 modules".

PFT uses a series of filter modules one after the other, each one splitting the incoming spectrum into two half-width bands.

This architecture would seem to lead to a 'tree' structure with each layer having twice the number of filters as the previous layer, but this is not the case.

Lillington realised that, although twice the number of filters is needed, each filter has half the processing load of filters in the layer above.

This means the total processing load at each level is the same. One identical filter block can be for each layer, the lower ones time-multiplexed between channels.

The only thing that increases in the lower layers is memory, as swap files are needed for each channel.

On a demonstration board, using four Xilinx Virtex FPGAs, Lillington claims that he can handle in excess of 100MHz bandwidth signal at eight-bit resolution and extract 1024 channels with sharp (stop band rejection better than 75dB) filter characteristics.

With two XCV2000E FPGAs and some RAM, an entire 4MHz band can be monitored with better than 200Hz resolution, an update rate of 200Hz, a dynamic range of over 130dB and pass band ripple of 0.2dB.

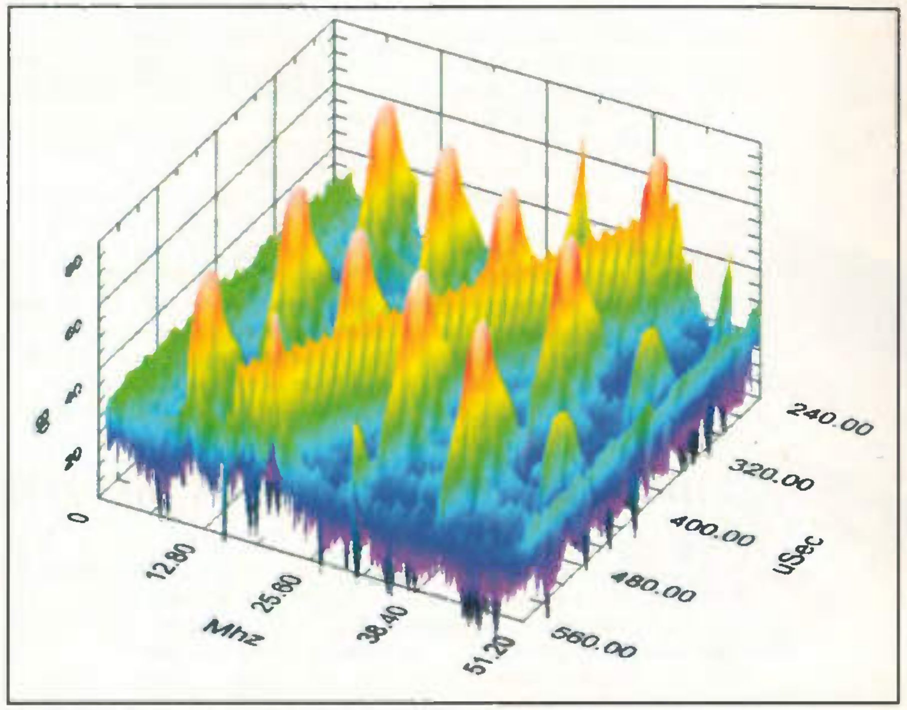
And this kind of performance is where the threat to spectrum analysers and warships comes from.

If these claims are correct, a few FPGAs executing a PFT can spilt a spectrum into narrow channels, with little hardware, at very high speed and in one go – with no need for sweeping filters.

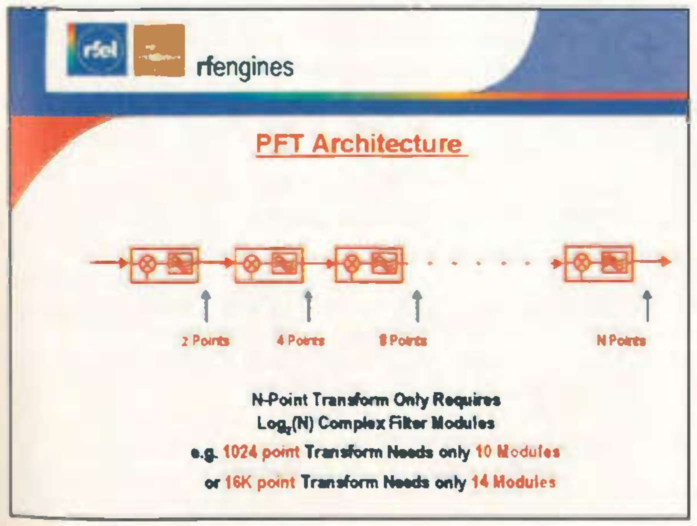
This performance makes good spectrum analysers look pretty silly and makes covert frequency-hopping military signals stand out clearly from surrounding noise.

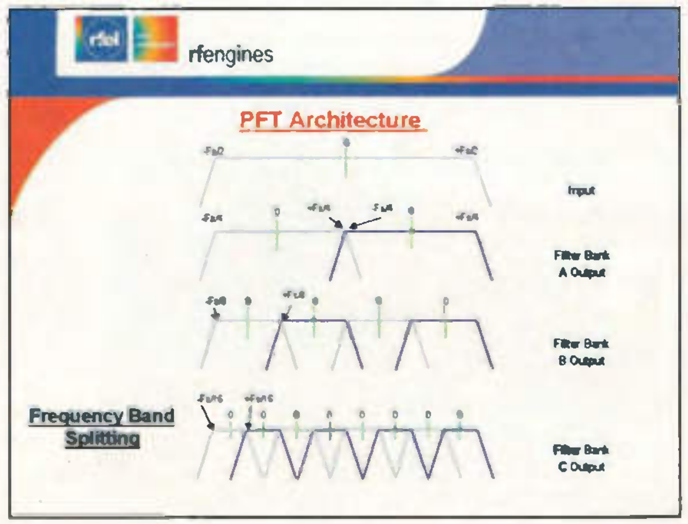
As well as looking at instrumentation and defence applications, RF Engines is looking at communications applications as PFT is suited to decoding several comms modulation schemes, particularly OFDM. This is the coding scheme used in Digital Audio Broadcasts and could well be the coding technique chosen for 4G mobile phones, said Lillington.

www.rfel.com



Extracting 30µs pulses switching between four frequencies and an 80GHz/s sweep signal would give most spectrum analysers a headache. A pipeline frequency transform can do it with very little hardware, claims its inventor.





Better and cheaper chip ESD protection

Improved circuitry for electrostatic discharge (ESD) protection could shrink I/O areas on chips by 30 per cent, according to developer Sarnoff.

The chip process technology firm developed the technology, called TakeCharge! at its European subsidiary in Belgium.

"We've found a way to shrink an area of the IC, the I/O and ESD protection area, that's been stuck at the same size for years. We reduce it at least 30 per cent, while improving device performance," said Koen Verhaege, technical director for device design at Sarnoff Europe.

"TakeCharge! typically allows $100 in extra parts to be placed on a wafer, and often several times that figure, with no additional masks or process line changes."

Part of the process is to get rid of the silicide blocking of transistors in the I/O area of the chip, a technique called 'back-end-ballast'. This reduces resistance in the diffusion and the gate, thereby increasing speeds.

An added bonus is that if this is the only use of silicide on the chip, then a whole mask step is saved, reducing costs further.

Two other changes help improve ESD by up to 60 per cent and reduce the area of the region by up to a factor of two to three.

In a 0.25μm process, the ESD circuitry held out against 8kV discharges for over 10000 repetitions, Sarnoff claimed.

Test devices have been manufactured in 0.25, 0.18 and 0.13μm CMOS, while 0.35 and 0.11μm are going through testing. Toshiba and Hynix are the first publicly named licensees of the technique.

Camera technology sends pictures direct to Internet

Cambridge Consultants has unveiled a digital camera concept product called 'SEE'. It has no memory to store images: instead it writes direct to the Internet, perhaps via a Bluetooth link and a mobile phone.

SEE includes both stills and video capability. A touch screen allows images to be edited and a pen allows messages to be added on top of captured images.

"The limits on file size and battery life can compromise the potential of digital cameras," said Donna Wilson, group leader with CCL's product definition team.

The camera will take up to one hundred stills or half an hour of video on one set of batteries, she said.

SEE uses a VideoCore chip from Alphamosaic, a recent spin-off from Cambridge Consultants.

Reconfigurable design tools aid pipeline pigs

Pipeline inspection firm PII is using reconfigurable hardware design tools in its 'intelligent pigs'. The pigs are autonomous inspection tools for pipelines that need processing capability of up to 6Mbyte/s.

The Northumberland company will use Celoxica's DK1 software and Altera FPGAs to replace obsolete, custom hardware in the pigs.

"We were using VHDL schematic entry tools but it was a struggle to maintain our in-house VHDL capability," said Gary Brayson, team leader for electronic design at PII.

"Each inspection tool, including circuits, drive elements, data capture and storage, costs the company around £6m to develop. Typically the return is realised over a 10 to 15 year term of service so component obsolescence between generations of inspection tool is a major issue," said Brayson.

New £300 000 prize for mathematical achievement

Mathematicians have long been denied the honour of being considered for the Nobel Prize, so the setting up of their own special award should be welcomed.

The Norwegian government has set up a prize for maths called the Abel, after Norway's most famous mathematician Niels Henrik Abel.

The award will be made annually with a cash prize of around £300 000.

"An international prize in mathematics dedicated to his [Abel's] name, is an expression of the importance of mathematics, and is intended to encourage students and researchers," said Norway's Prime Minister, Jens Stoltenberg.

Niels Henrik Abel lived until the age of just 26 at the beginning of the 19th century. His countrymen rate his achievements alongside those of Ibsen, Grieg and Munch, for their respective contributions to literature, music and art.

Surgeons in New York remove French woman's gall bladder – while she's still in France

A medical operation has been carried out by a remotely controlled robot.

Claimed to be the first use of Telemedicine, 'Operation Lindbergh' involved a team surgeons in New York and a patient in Strasbourg, France.

Surgeons in New York controlled a robotic system manufactured by Computer Motion to remove the woman's gall bladder. The link between the surgeons and the robotic system was an end-to-end high-speed fibre optic service provided by the France Telecom Group. The patient is said to be recovering without complications.

www.websurg.com, www.eits.org and www.computermotion.com

A new look at Class-G power

Douglas Self has been investigating one of the lesser-known classes of audio power amplifier – Class-G. His findings reveal that this class is considerably more power-efficient than Class-B when handling realistic signals. Class-G has a reputation for sacrificing linearity for efficiency, but the innovative design to be presented features a distortion figure that's lower than all but the very best of Class-B.

Class-G amplifiers have been around since 1976, but to the best of my knowledge, no constructional project for a Class-G amplifier has ever been published. This deficiency is now remedied.

In the field of audio it is easy to find amplifiers less efficient than Class-B; Class-AB is markedly less efficient at the low end of an amplifier's power capability, while it has to be faced that Class-A squanders huge amounts of energy. Finding something with higher efficiency is rather more difficult.

Class-D, using ultrasonic pulse-width modulation, promises high efficiency and sometimes delivers it, but it is inescapably an awkward technology. Class-D efficiency depends crucially on details of circuit design and device characteristics. The *LC* output filter will only give a flat response into one load impedance, and there are daunting EMC difficulties. It is not an attractive proposition for high-quality amplifiers that must work with separate speakers.

This pretty much leaves the Class-G principle, in which power is drawn from either high or low voltage supplies as the signal demands. This important technology is used extensively in very-high-power amplifiers for large PA systems, where the power savings can be crucial, and is now appearing in home theatre applications; if you have five amplifiers instead of two, their losses are significant.

Class-G has also recently begun to appear in powerful sub-woofers, and even in ADSL drivers. It has historically been used in high-power sonar transmitters, but since these tend to be fitted to nuclear submarines, details are scarce.

Basic principles of Class-G

It is the large peak-to-mean ratio of music that makes possible improved efficiency in Class-G. By large peak-to-mean ratio, I mean that most of the time the power output is much below the peak levels. Statistics on typical values for this ratio for various kinds of music are surprisingly hard to find, but it is generally accepted that the range between 10dB for compressed rock, and 30dB for classical material, covers most circumstances.

Clearly, the signal spends most of its time at low powers, if the peaks are much greater than the average level. As a result, a low-powered amplifier will be much more efficient. However, when the occasional high-power peaks do come along, they must be catered for by some mechanism that can draw high power, causing high internal dissipation, but only for brief periods.

In the usual Class-G configurations, there are two or three pairs of supply rails; for most of the time lower output levels are supplied from the lowest-voltage rails, with a low voltage-drop between rail and output, and correspondingly low dissipation. The infrequent peaks above the transition level are supplied from the high-voltage pair of rails.

Clearly the switching between rails is the heart of the matter, and anyone who has ever done any circuit design will immediately start thinking about how easy – or otherwise – it will be to make this happen cleanly with a 20kHz signal.

There are two main ways to arrange the dual-rail system; series and parallel, i.e. shunt. This article deals only with the series configuration, as it seems to have had the greatest application to hi-fi. The parallel version is more often used in high-power PA amplifiers.

Series-configured Class-G

A series-type Class-G output stage with two rail voltages is shown in **Fig. 1**. The inner devices are those that conduct continuously; those that perform the rail-switching are the outer devices.

In all cases in this article, the emitter-follower, or EF, type of output stage is used; the complementary-feedback pair, or CFP, configuration can be used for inner, outer, or both sets of output devices, but I fear I have insufficient space to deal properly with this issue here.

For maximum efficiency, the inner stage normally operates in Class-B, though there is no inherent reason why it could not run in Class-AB or even Class-A; more on this later. Therefore if the inner devices are in Class-B, and the outer ones conduct for much less than 50% of a cycle, the outer devices are effectively in Class-C.

According to the classification scheme I proposed in reference 1, this configuration is Class B+C. The plus sign indicates the series connection of the outer and inner devices. This basic configuration was developed by Hitachi with the explicit intention of reducing amplifier power dissipation[2,3].

Musical signals spend most of their time at low levels, due to the high peak/mean ratio. Dissipation is greatly reduced by running off the lower $\pm V_1$ supply rails at these times. The inner stage $Tr_{3,4}$ operates essentially in normal Class-B. Transistors $Tr_{1,2}$ are the usual drivers and R_1 their shared emitter resistor.

The usual temperature-compensated V_{bias} generator is required, shown here split in half to maintain circuit symmetry when the stage is simulated; note that it is the inner output devices whose temperature must be tracked.

Low-level supply power is drawn through diodes $D_{3,4}$. These are often called the commutating diodes, because of their rail-switching role. Using the word 'commutation' avoids confusion with the usual Class-B crossover at zero volts.

When the positive-going instantaneous signal level exceeds $+V_1$, D_1 conducts, Tr_5 and Tr_6 turn on. Diode D_3 turns off, so the entire output current is now drawn from the higher $+V_2$ rail, with the voltage-drop and hence the dissipation shared between $Tr_{3,6}$. Negative-going signals are handled the same way. **Figure** 2 shows how the collectors of the inner power devices retreat away from the output rail as it approaches the transition level.

Class-G is often said to have worse linearity than Class-B, the blame being placed on problems with diode commutation. As usual, received wisdom is only half of the story, and there are other linearity problems that are not due to sluggish diodes, as you will see shortly.

It is characteristic of Class-G that such glitches only

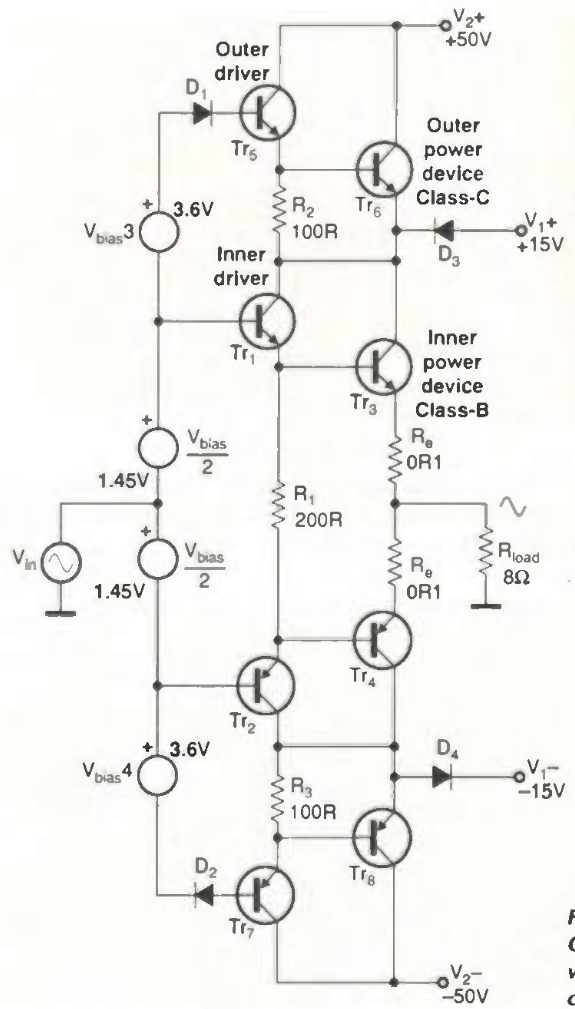


Fig. 1. A series Class-G output stage, alternatively Class B+C. Voltages and component values are typical. The inner stage is Class-B EF. Biasing by my method.[1]

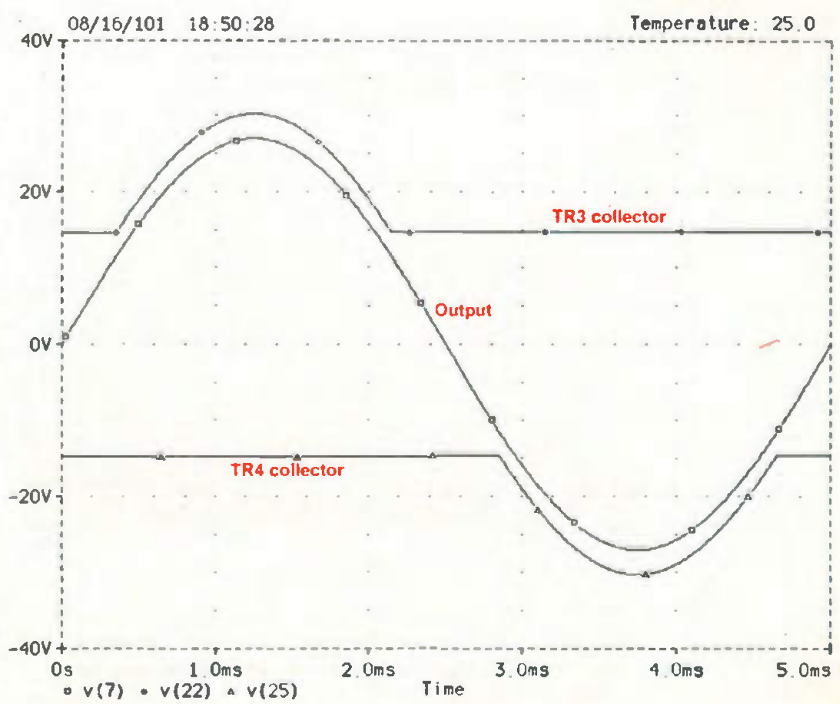


Fig. 2. Output of a Class-G stage and the voltages on the collectors of the inner output devices.

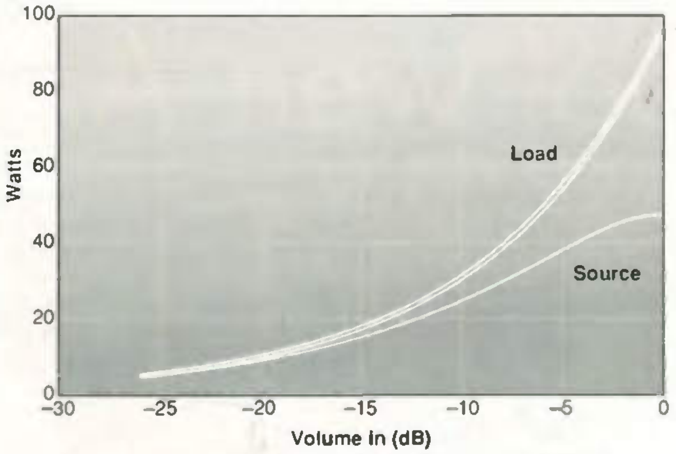


Fig. 3. Power partition diagram for a conventional Class-B amplifier handling a typical music signal with a triangular probability-density function. X-axis is volume.

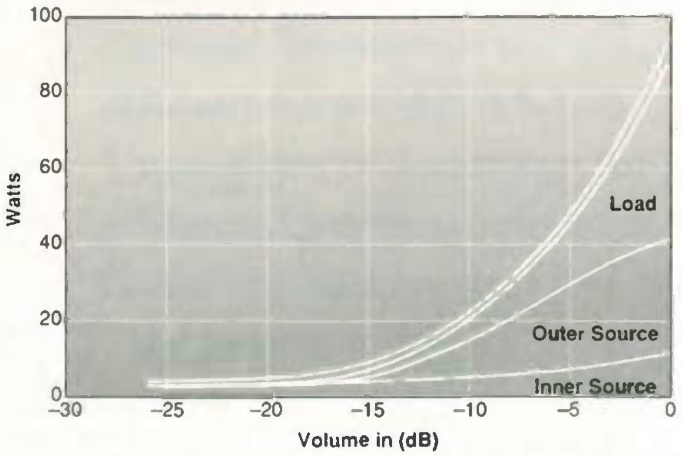


Fig. 4. Power partition diagram for Class-G with V1/V2=30%. Signal has a triangular PDF. X-axis is volume; outer devices dissipate nothing until –15 dB is reached.

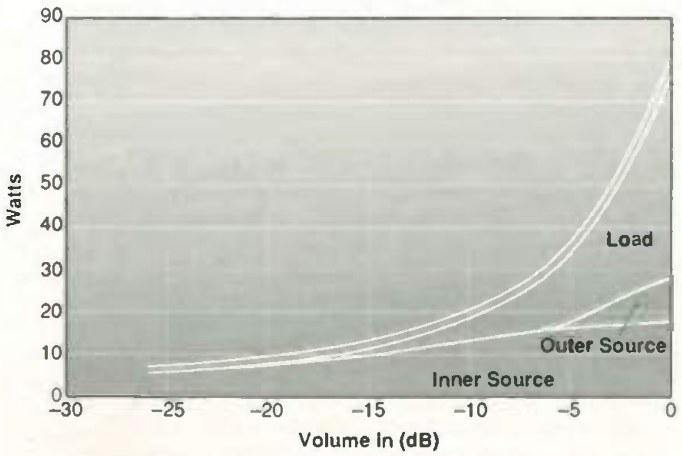


Fig. 5. Power partition diagram for Class-G with V1/V2=60%. Triangular PDF. Compared with Fig. 4, the inner devices dissipate more and the outer devices almost nothing except at maximum volume.

occur at moderate power or above, and they are well displaced away from the critical crossover region. A Class-G amplifier has a low-power region of true Class-B linearity, in much the same way as a Class-AB amplifier has a low-power region of true Class-A performance.

Class-G and efficiency

The standard mathematical derivation of Class-B efficiency with sinewave drive uses straightforward integration over a half cycle to calculate internal dissipation against voltage-fraction, i.e. the fraction of possible voltage swing.

As is well known, dissipation reaches about 40% of maximum output power, at a voltage-fraction of 63%, which also delivers 40% of maximum output power to the load. The mathematics is simple, because the waveforms do not vary with output level. Every possible idealisation is assumed, such as zero quiescent current, no emitter resistors, no $V_{ce(sat)}$ losses and so on.

In Class-G, the waveforms are a strong function of output level, requiring variable limits of integration... it all gets rather unwieldy. The simulation method detailed in reference 4 is much easier – if rather laborious – and can be used with any input waveform, to yield a power-partition diagram, or PPD. This diagram shows how the power being drawn from the supply is distributed between output device dissipations and useful load power.

It is recognised that sinewaves and the like are very poor simulations of music for this purpose. Their main advantage is that they allow direct comparison with the purely mathematical approach.

However, the whole *raison d'etre* of Class-G is power saving. With this technology in particular, the waveform used has a strong effect on the results. For this reason, I have concentrated on the PPD of an amplifier with real musical signals, or at any rate, their statistical representation. The details of the triangular probability distribution function, or PDF, approach are given in reference 5.

Figure 3 is the triangular PDF PPD for conventional Class-B EF, while **Fig. 4** is that for Class G with V_2=50V and V_1=15V, i.e. with V_1/V_2=30%. The PPD plots power dissipated in all four output devices, the load, and the total drawn from the supply rails. It shows how the input power is partitioned between the load and the output devices. The total sums to slightly less than the input power, the remainder being accounted for by the drivers and Re losses.

In Fig. 4, the lower area is the power in the inner devices and the larger area just above is that in the outer devices; there is only one area for each because in Class-B only one device is on at a time.

The outer device dissipation is zero below the rail-switching threshold at –15 dB ref maximum output. Total device dissipation at full output is reduced from 48W in Class-B to 40W. At first glance, this may not appear to be a good return for twice as many power transistors.

Figure 5 shows the same PPD but with V_2=50V and V_1=30V, i.e. with V_1/V_2=60%. Here, the low-dissipation region now extends up to –6dB, but inner device dissipation is higher due to the increased lower rail voltages. The overall result is that total device power at full output is reduced from 48W in Class-B to 34W, which is a definite improvement.

The efficiency is very sensitive to the ratio of rail voltages used. Very few domestic amplifiers are operated at full volume all the time. In real life, the lower V_1 voltage is likely to give lower general dissipation. I do not suggest

that V_1/V_2=30% is necessarily the optimum lower-rail voltage for all situations, but it looks about right for domestic hifi.

Practicalities

I have wrestled with many 'new and improved' output stages that proved to be anything but. My first thoughts on Fig. 1 ran something like: "Will this work in SPICE simulation?" It did. "And will this work for real at 1kHz?" It did – first time.

The second question is more subtle than it looks. It is all too easy to design complicated output stages that work beautifully in simulation but prove impossible to stabilise at HF. Some of the interesting output-triple configurations seem to suffer from this.

I also asked myself, "Will this work for real at 20kHz?" It will – and indeed at 50kHz too. I haven't pushed it further. This is a very different question from "Will this work for real at 1kHz?". It is quite possible to come up with a configuration that either just does not work at 20kHz or is provoked into oscillation that is not triggered by a 1kHz stimulus.

Having settled these points, I proceeded with the design.

The biasing chain

The biasing requirements are rather more complex than for Class-B. Two extra bias generators $V_{bias3,4}$ are required to ensure Tr_6 turns on before Tr_3 runs out of voltage. This voltage is not critical, so long as it does not fall too low, or become much too high. Fixed zener diodes of normal commercial tolerance are quite good enough.

If this bias is set too low, so that the outer devices turn on late, then the V_{ce} across Tr_3 falls too low, and its current capability declines; when evaluating this bear in mind the lowest impedance load you plan to drive. Alternatively, if the bias is too high, then the outer transistors turn on too early, and the dissipation in the inner devices is greater than it need be.

If the bias is too high, this is less of a problem. If you're in doubt, make this bias higher rather than lower. The original Hitachi circuit in reference 1 put zener diodes in series with the signal path to the inner drivers to set the output quiescent bias, **Fig. 6**. This effectively subtracted their voltage from the main bias generator, which was set up at 10V or so, much higher than usual. Simulation showed that Zeners in the forward path caused poor linearity, which is not exactly surprising.

There is also the problem that the quiescent conditions will be affected by changes in the zener voltage. Also, if the 10V bias generator is the usual V_{be}-multiplier, it will have much too high a temperature coefficient for proper thermal tracking.

To alleviate these problems, I rearranged the biasing as in **Figs 1 & 11**; the amplifier forward path now goes directly to the inner devices, and the extra bias voltages are

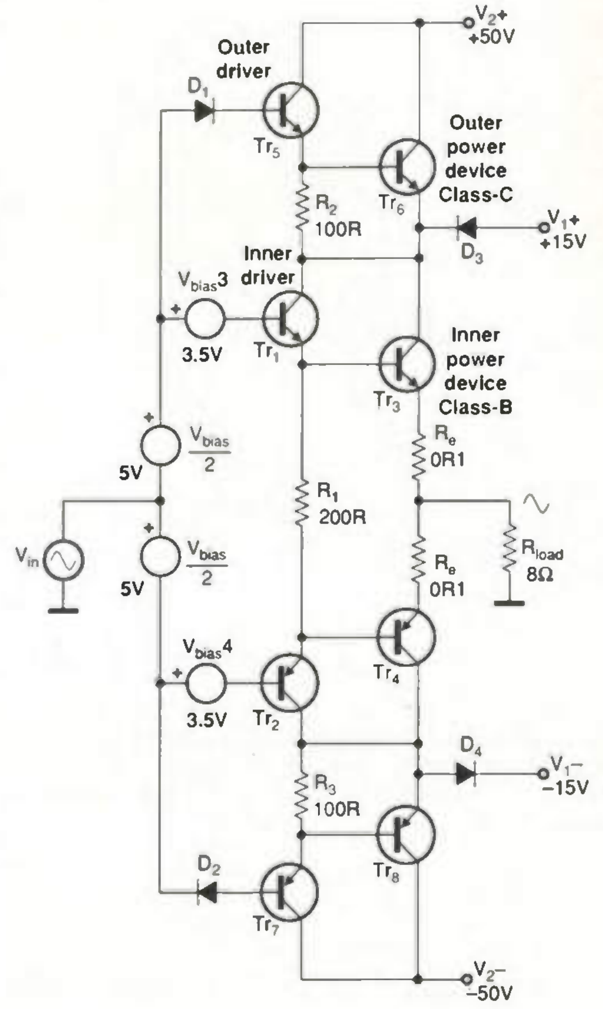


Fig. 6. The original Hitachi Class-G biasing system, with inner device bias derived by subtracting $V_{bias3,4}$ from the main bias generator.

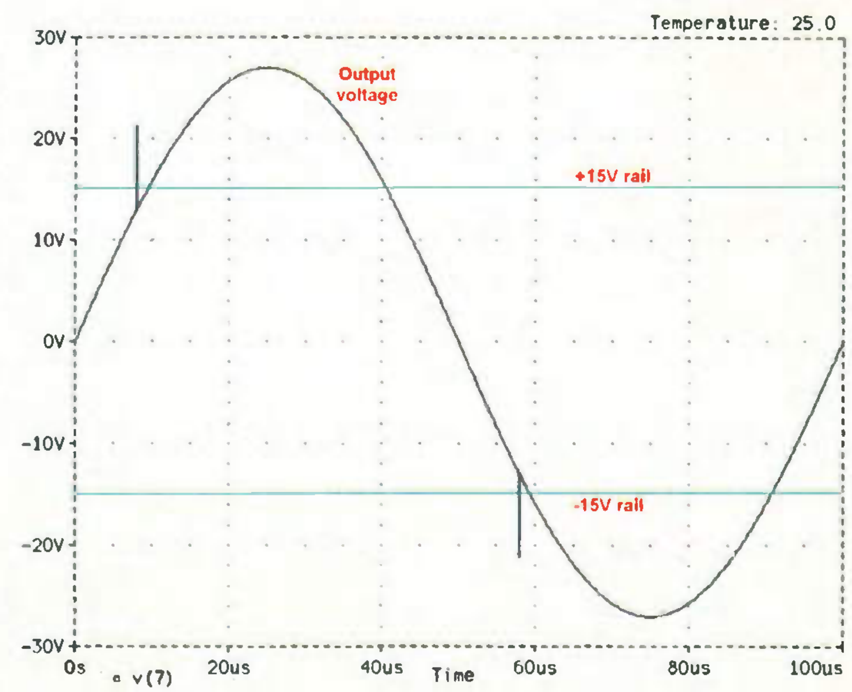


Fig. 7. Spikes due to charge storage of conventional diodes, simulated at 10kHz. They only occur when the diodes turn off, so there are only two per cycle. These spikes disappear completely when Schottky diodes are used in the SPICE model.

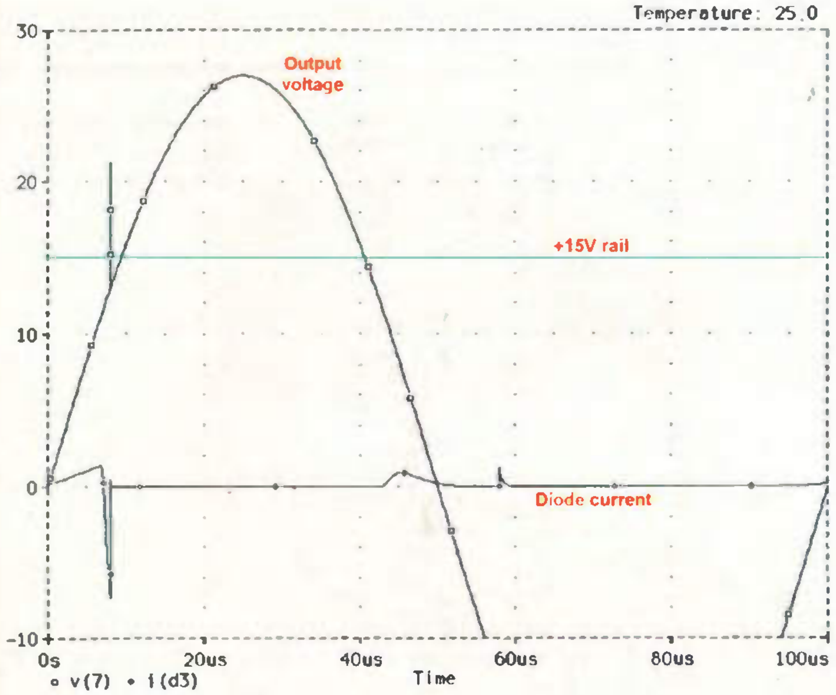


Fig. 8. Close-up of the diode transient. Diode current rises as output moves away from zero, then reverses abruptly as charge carriers are swept out by reverse-biasing. The spike on the output voltage is aligned with the sudden stop of the diode reverse current. Frequency 10kHz.

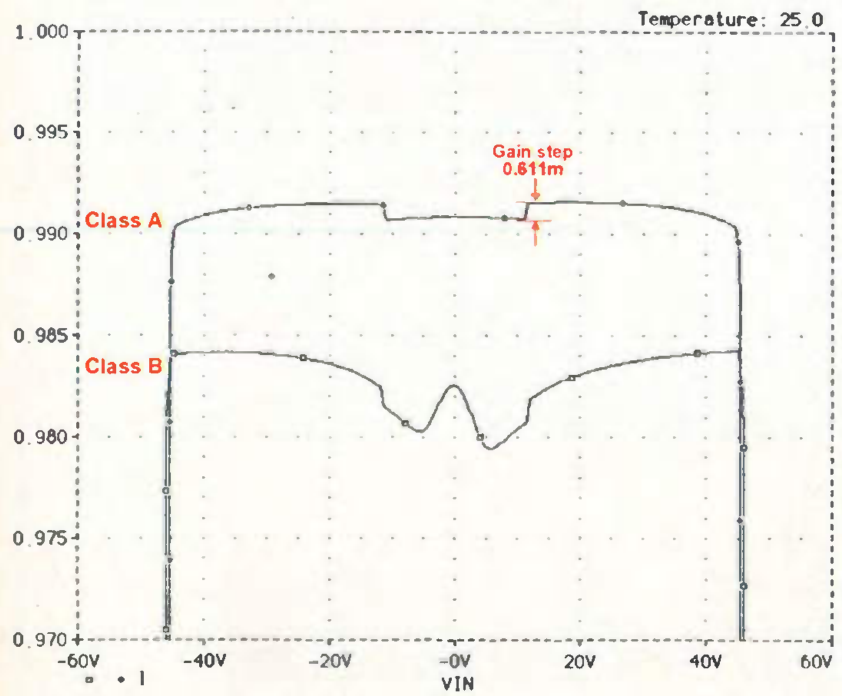


Fig. 9. SPICE simulation shows variations in the incremental gain of an EF-type Class-G series output stage. The gain-steps at transition – at ±16V – are due to Early effect in the transistors. The Class-A trace is the top one, with Class-B optimal below. For both, the inner driver collectors are connected to the switched inner rails, i.e. the inner power device collectors, as in Fig. 1.

in the path to the outer devices; since these do not control the output directly, the linearity of this path is of little importance.

The zener diodes are out of the forward path and the bias generator can be a standard type. It must be thermally coupled to the inner power devices; the outer ones have no effect on the quiescent conditions.

Linearity problems of series Class-G

Series Class-G has always had a question-mark against its linearity because of difficulties with the rail-switching. The diodes $D_{3,4}$ must be power devices capable of handling a dozen amps or so.

Conventional silicon rectifier diodes take a long time to turn off due to stored charge carriers. This has the following deleterious effect: when the voltage on the cathode of D_3 rises above V_1, the diode attempts to turn off abruptly. Its charge carriers though sustain a brief but large reverse current as they are swept from its junction. This current is supplied by Tr_6, attempting as an emitter-follower to keep its emitter up to the right potential. So far so good.

However, when the diode current ceases, Tr_6 is still conducting heavily, due to its own charge-carrier storage. The extra current it turned on to feed D_3 in reverse now goes through Tr_3's collector, which accepts it because of this transistor's low V_{ce}, and passes it onto the load via its emitter and emitter resistor.

This process is readily demonstrated by a SPICE commutation transient simulation, **Figs 7 & 8**. Note that there are only two of these events per cycle – not four – as they only occur when the diodes turn off. In the original Hitachi design, this problem was reportedly tackled by using fast transistors and relatively fast gold-doped diodes, but according to reference 2, this was only partially successful.

It is now easy to eradicate this problem. Schottky power diodes are now readily available – as they were not in 1976 – and they are much faster due to their lack of minority carriers and charge storage. They also have the added advantage of a much lower forward voltage drop at large 50A forward currents.

The only snag with Schottky power diodes is their relatively low reverse withstand voltage. Fortunately, the diodes are only exposed at worst to the difference between V_2 and V_1, and this only in the low power domain of operation.

Another good point about these components is that they appear to be reasonably tough; I have subjected 50A Motorola devices to 60A-plus repeatedly without a failure.

The spikes disappear completely from the SPICE plot if the commutating diodes are Schottky rectifiers. Motorola MBR5025L diodes capable of 50A and 25 PIV were used in simulation.

Static linearity

Spice simulation reveals in **Fig. 9** that the static linearity – i.e. that revealed by a DC analysis – is distinctly inferior to Class-B. There is the usual gain-wobble around the crossover region, exactly as for straight Class-B, but also there are now gain-steps at ±16V.

The result with the inner devices biased into push-pull Class-A is also shown, and proves that the gain-steps are not related to crossover distortion. Since this a DC analysis, the steps cannot be due to diode switching-speed or other dynamic phenomena, and Early effect was immediately suspected. Early effect is the increase in collector current when the collector voltage increases, even though the V_{be} remains constant.

When unexpected distortion intrudes into a SPICE simulation, and beta effects seem unlikely, a handy diagnostic technique is to turn off Early effect for each transistor in turn. In PSpice models the Early effect can be disabled by setting the parameter VAF to a much higher value than the default of 100. This quickly proved that the gain-steps were caused wholly by Early effect acting on both inner drivers and inner output devices: the gain-steps are completely abolished.

When Tr_6 begins to act, Tr_3's V_{ce} is no longer decreasing as the output moves positive, but substantially constant as the emitter of Tr_6 moves upwards at the same rate as the emitter of Tr_3. This has the effect of a sudden change in gain, which degrades linearity.

The effect appears to occur in both drivers and output devices in equal measure. It can be easily eliminated in the drivers by powering them from the outer rather than the inner supply rail. This prevents the sudden changes in the rate in which driver V_{ce} varies. The improvement in linearity is seen in **Fig. 10**, where the gain-steps are halved in size.

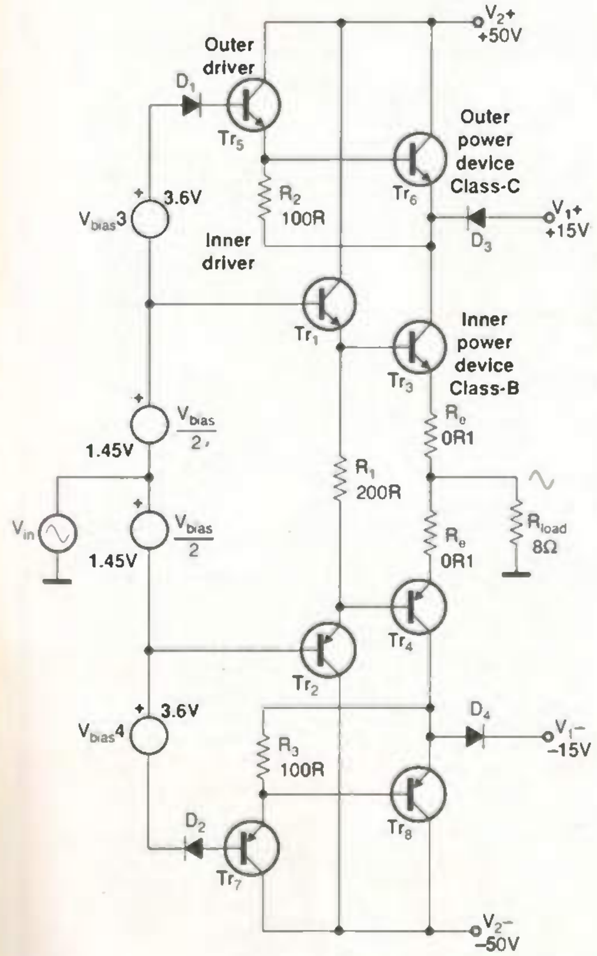


Fig. 11. Class-G output stage with inner drivers powered from outer supply rails.

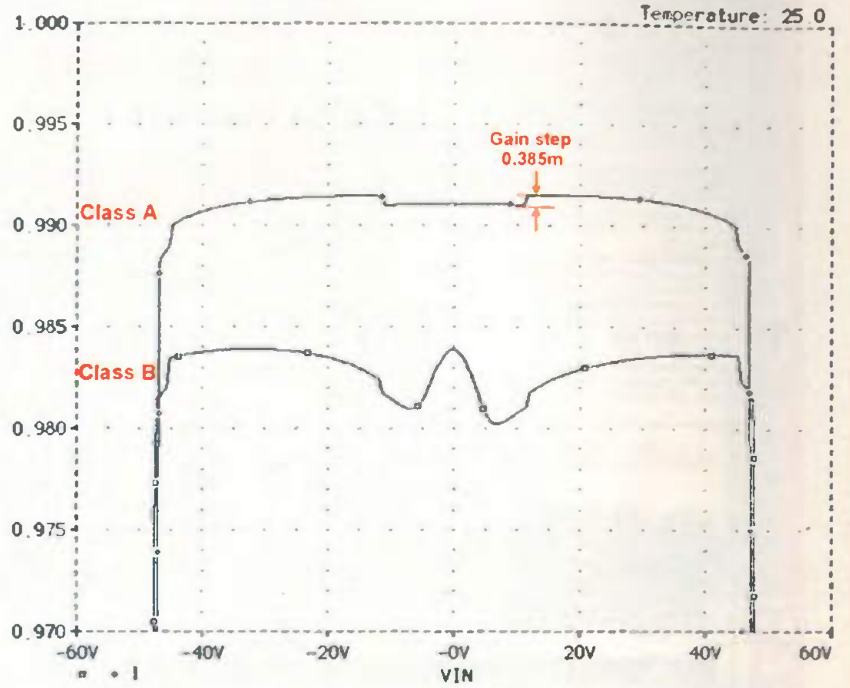


Fig. 10. Connecting the inner driver collectors to the outer V_2 rails reduces Early effect nonlinearities in them, and halves the transition gain-steps.

Figure 11 shows the resulting circuit. Driver power dissipation is naturally increased, but this is such a small part of the total that the overall efficiency is not significantly degraded. It is clearly not practicable to apply the same method to the output devices, because then the low-voltage rail is never used and it isn't Class-G any more. The small-signal stages before the drivers naturally have to work from the outer rails to generate the full voltage swing.

At this stage we have eliminated the diode glitches, and halved the size of the gain-steps. With these improvements it is now practical to design a Class-G amplifier with mid-band THD below 0.002%. My second article on this topic shows how to do just that, presenting a complete, working Class-G power amplifier design. ■

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Shielding and EMI - 1

Joe Carr explains the basics of one of the hottest topics in today's electronics design arena – EMI shielding. In this first article of a set of two, Joe gives practical tips on grounding the circuit and its shield, and on physical enclosure requirements.

It is almost an article of religion in electronics that shielding electronic circuits prevents EMI problems. A good shield will keep undesired signals inside in the case of a transmitter – or outside in the case of other forms of circuits.

All transmitters generate harmonics and other spurious signals. If they are radiated, then they will interfere with other services. Signals that go out through the antenna terminal usually pass through either tuning or filtering networks, which tend to clean up the emission. But if the circuits are not shielded, then direct radiation from the chassis will defeat the effects of the filtering.

In theory, shields are a good idea. Unfortunately though, many shields are essentially useless. In some cases, they may even cause more problems than they cure. The problem is not just on transmitters, or even just RF circuits in general, but on all electronic circuits. I once worked with medical and scientific electronic instruments that rarely used frequencies above 1000Hz, and they were subject to severe EMI. And where did it come from? It came from the 60Hz power line.

Let's look at shielding materials and methods. **Figure 1a)** shows a universal 'black box' circuit with three ports: A is the input, B is the output and C is the common. The term 'black box' means any form of electronic circuit. It is used to universalise the discussion so that ideas are not associated with any specific class of circuit.

What's inside the black box enclosure could be a circuit, or a system such as a transmitter, receiver, audio amplifier, or a medical electrocardiograph amplifier. It doesn't matter for our present purposes.

Shielding involves placing a metal screen or barrier around the circuit. In Fig. **1b)** the black box circuit is placed inside a shielded compartment, as indicated by the dotted lines. In addition, the input voltage V_{AC} and output voltage V_{BC} are shown: the subscript letter refer to the port designations.

Any time that two conductors are brought into close proximity to each other, but not touching, a capacitance exists between them. Sometimes, the capacitances are intentional. But in other cases, the capacitances are incidental to construction. An example of such incidental capacitance is an insulated wire laying on a chassis. In the case of Fig. 1b) there are three 'stray' capacitances represented: C_{AD}, C_{BD}, and C_{CD}.

Shielding of the sort shown in Fig. 1b) is not terribly effective. It can lead to instability – and outright oscillation under some circumstances. The feedback path that causes the problem is better seen in the re-drawn version of the circuit shown in Fig. **1c)**.

Capacitors C_{BD} and C_{CD} form a capacitive voltage divider, with the "output" connected through C_{AD} to the input terminal A of the black box. Under the right circumstances, this circuit can lead to very bad EMI/EMC consequences.

Shielding rule No 1

The solution to the problem is to apply shielding rule No 1: the shield must be connected to the zero signal reference point in the circuit being protected, i.e. the common line between output and input.

In some cases, the common might be a floating connection that is not earth grounded, i.e. a counterpoise ground plane. The common point may be at a non-zero voltage, but for the purposes of the input

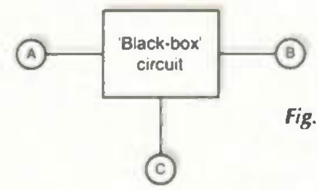


Fig. 1a)

Fig. 1. When considering shielding, it often helps to think of a circuit as a 'black box' with three terminals – input, output and common, as in a). In b), the black box is shown inside a shielding enclosure. But this method of shielding has drawbacks due to a feedback that is clearer to see in the redrawn version of the set-up in diagram c). In d) is a configuration wired with the shield connected to the zero signal reference point – as it should be for reliable shielding.

and output signals it is the zero-signal reference point. In most cases, the zero-signal point is, in fact, at a potential of zero volts.

Application of this rule is shown in Fig. 1d). The common, C, is connected to the shield D, effectively shorting out capacitance C_{CD}, and the common node of the feedback network evident in Fig. 1c). To restate the general rule: connect the shield and common signal line together.

In the case of Fig. 1, the 'black box' circuit is single-ended, so the common line of the internal circuit is connected directly to the shield.

Figure 2a) shows a situation that's a little more complex. In this case, some 'black box' circuit is inside a shielded enclosure, and supplies output signal to some sort of resistive load. The load connects to the shielded enclosure by some sort of shielded cable. Similarly, a shielded signal source V_{in} is connected to the input of the 'black box' by another length of shielded cable.

In this case, there could be too many grounds. Suppose that the common signal point inside the main shielded compartment is connected to the shield, and the shield is, in turn grounded at point A. The signal source is also grounded, but to a different point, namely point B.

If current I_G flows in the ground plane resistance R_G, then a voltage drop V_G will be formed across the resistance of the ground path. The current might be due to external circuits, or it may be due to a potential difference existing between two points in the circuitry inside the shielding.

Whatever the source, however, a difference of potential between points A and B gives rise to a spurious signal voltage V_G that is effectively in series with the actual signal voltage V_{IN}. Thus, the total signal seen as valid by the circuit is $V=V_{IN}+V_G$. This is the 'ground-loop' problem.

The key to solving the ground-loop problem is to connect the shield to the ground plane at the signal end, B, and not at any other points. An application of rule No 1 might say: "The shield and common of the internal circuitry should be connected together at the point where the signal source is grounded." In other words, break the

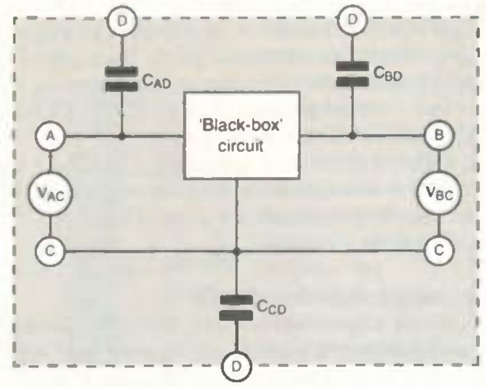


Fig. 1b)

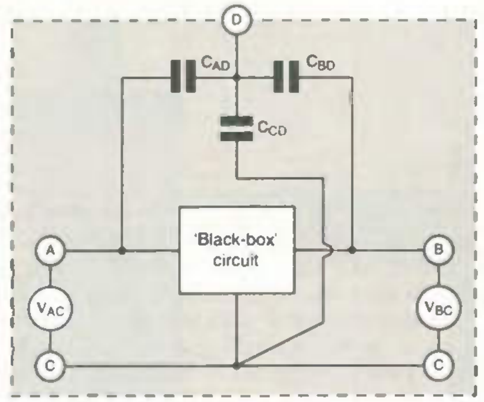


Fig. 1c)

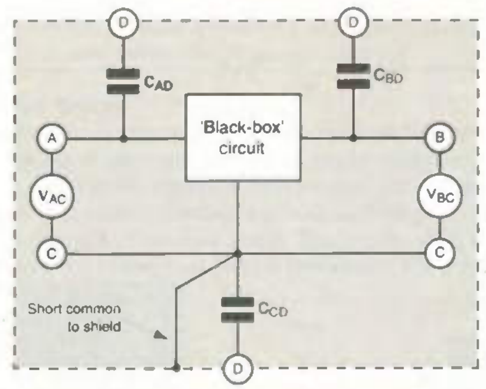


Fig. 1d)

connection at point A and rely instead on point B, as shown in Fig. **2b)**.

This class of problem is representative of a class of problems in which a common impedance – in this case a resistance, R_G – couples two segments of a circuit. If a voltage drop appears across the common impedance, then a problem will surface.

Two approaches to shielding

There are two basic approaches to shielding: absorption and reflection. These mechanisms often operate together.

Suppose a large external field is present. In the case of absorption, the field may penetrate the shield but is greatly attenuated. In the case of reflection, the field is turned back by the metal shield.

The absorptive method is usually used at frequencies below 1000kHz for magnetic fields. The types of materials tend to be the ferromagnetic materials such as steel and a special material used especially for magnetic shields called 'mu-metal' or μ-metal.

At higher frequencies – especially where the electric field is of more importance than the magnetic field – better shielding materials are copper, brass and aluminium.

Skin effect and skin depth

Alternating currents do not flow uniformly throughout the cross section of a conductor as is the case with direct currents. Due to skin effect, AC currents flow only near the surface of the conductor.

This effect creates a situation where the AC resistance of a conductor will be higher than the DC resistance. If the current density from the surface to the centre of a cylindrical conductor is graphed, then it will be found that the curve is a section of a parabola.

The critical depth for a cylindrical conductor is the depth at which the current density falls off to 0.368 times the surface current density. It is this current that we use to determine the AC resistance.

Sheets or plates of metal used for shielding also show a skin effect when currents flow in them. The skin depth, Fig. 3, is analogous to the critical depth in cylindrical conductors. In both cases, 63.2% of the current flows in the area between the surface and the skin depth, ∂. The skin depth is calculated from:

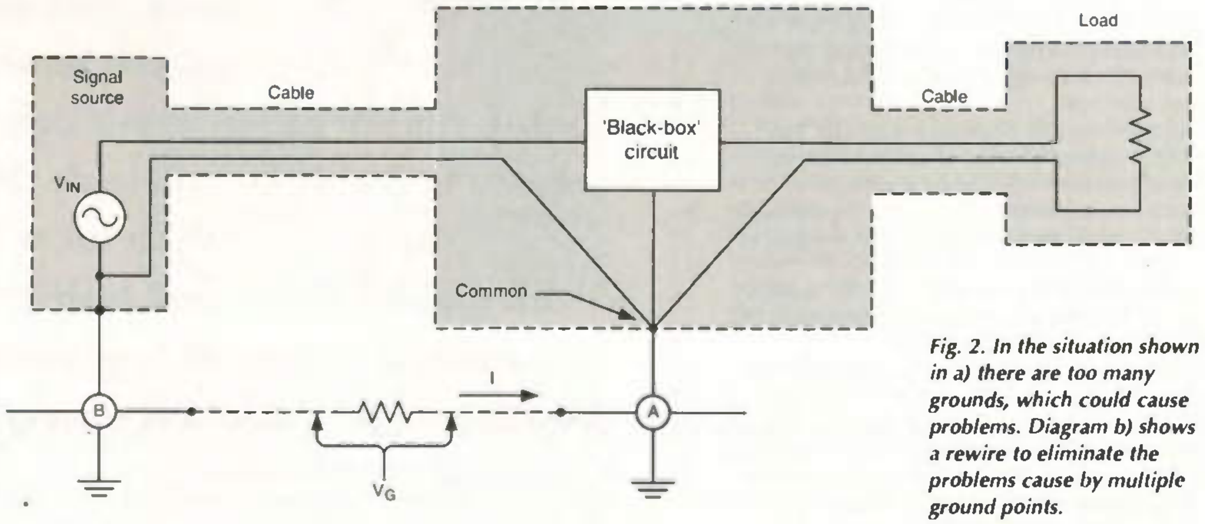


Fig. 2. In the situation shown in a) there are too many grounds, which could cause problems. Diagram b) shows a rewire to eliminate the problems cause by multiple ground points.

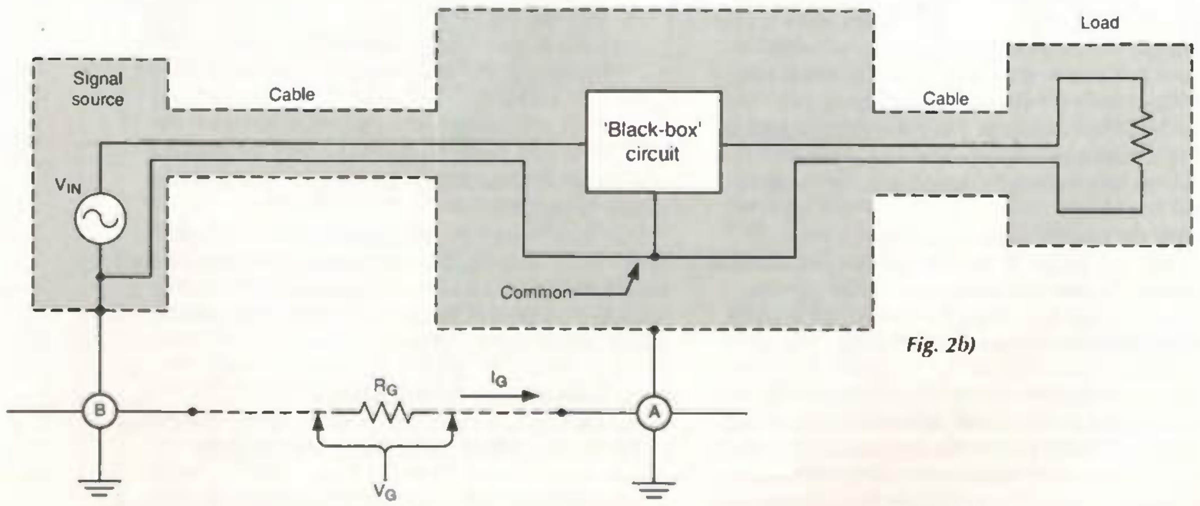


Fig. 2b)

$$\partial = \frac{2.602k}{\sqrt{F_{Hz}}} \qquad (1)$$

Where: ∂ is the skin depth in inches, F_{Hz} is the frequency in hertz, and constant k is 1.00 for copper, 1.234 for aluminium and 0.100 for steel.

Why is this important? In the case of absorptive loss, the attenuation is 8.7dB/∂. For example, at 60Hz a steel shield has a skin depth of 0.86mm (0.034in). If 1.6mm ($^1/_{16}$in) stock is used, the total depth is equivalent to 1.84∂, so the attenuation for magnetic fields would be

$$8.7\text{dB} \times 1.84 = 16\text{dB}.$$

To obtain maximum reflective loss at radio frequencies, the thickness of the shielding material should be about three to ten times the skin depth. The thicker the shield, the better the shielding – up to a point.

At 10MHz for example, aluminium has a skin depth of 0.0254mm (0.001 in), and copper has a skin depth of 0.020mm (0.0008in), so the shield thickness should be 0.254mm (0.010in) for aluminium and 0.2mm (0.008in) for copper or more. Given that common $^1/_{16}$in thick stock is 1.6mm thick, aluminium will be marginal while copper would be more than sufficient.

It is only fair to note that some textbooks say a shield should be at least three times the skin depth... but that is for minimal shielding.

Ground planes and wire size

The ground plane might be an actual earth ground. In most electronics circuits though, it will be either a printed circuit board or a chassis.

In the case of printed circuit boards, it is usually recommended in RF circuits to use a double-sided board with the top-side copper used as a ground plane, and possibly to carry DC power-supply lines.

In RF circuits, it is not advisable to use small wires or printed circuit tracks as ground lines. The AC resistance of cylindrical wire conductors is a function of both the wire diameter and the frequency. For any given wire size, the AC resistance:

$$R_{AC} = kR_{DC}\sqrt{F_{MHz}} \qquad (2)$$

The value of the *k* factor depends on the wire size:

Wire size (AWG)	*k*-factor
8	35
10	28
14	18
18	11
22	7

Thus, when you use #22 AWG solid hook-up wire to carry a 1MHz RF current, the AC resistance is seven times the DC resistance. If this wire is a ground, and carries a current, the AC resistance of the wire might be considerable, and create a nasty ground-loop voltage drop.

Even if the wire is large enough to reduce the effects of AC resistance at radio frequencies, the inductance might be a problem. The inductance of a straight length of #22 AWG wire is about 600μH/1000ft.

A one foot run of wire will have an inductance about 0.6μH. This inductance will not be noticed in an audio circuit, or even many low-frequency RF circuits, but as the frequency climbs it becomes significant. In the upper HF and lower VHF regions it is a significant portion of lumped inductances intentionally placed in the circuit.

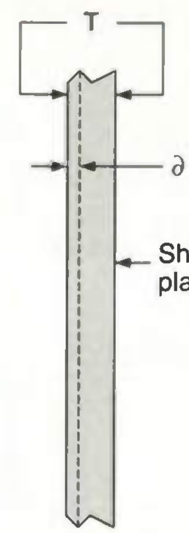


Fig. 3. In cylindrical conductors carrying AC, more current is carried in the conductor's 'skin' than in its core. This effect is also apparent in sheet metal and needs to be taken into account when you are designing an enclosure intended for EMI shielding.

If the wire is in a ground path, then it is a common impedance. Any RF voltage developed across its inductive reactance forms a valid signal, and may cause problems. The key to the problem is star grounding, i.e. grounding all circuit elements to the same point. If the signal source is grounded, then its ground connection ought to be used as the overall grounding point.

Shielded boxes

A number of manufacturers sell prefabricated shielded boxes. Some of them are quite good, while others are not very good at all. Figure 4 shows one of the poorer forms of aluminium shielded box from an EMI point of view. It consists of two half-shells. The bottom shell is bent into a 'U' channel shape (see end view). The top

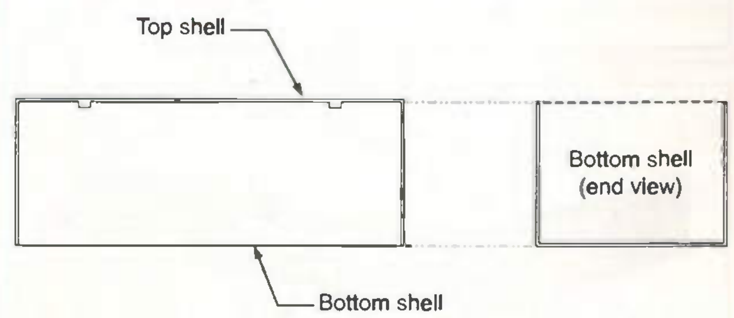


Fig. 4. Many prototype pressed aluminium and steel enclosures have a bottom with an open-ended channel structure and a lid with folded-over ends. Such an enclosure does not provide an effective screen – particularly at higher frequencies.

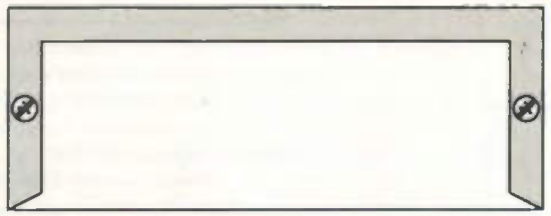


Fig. 5. This type of enclosure has the same basic construction as that of Fig. 4, but it has added lips, resulting in improved screening properties.

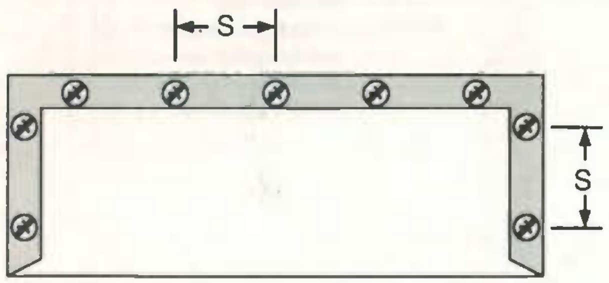


Fig. 6. At UHF, the improved enclosure in Fig. 5 also starts to leak because of the distance between the screws. Adding screws to bring the spacing down to at most 0.05λ of the highest frequency involved will solve the problem.

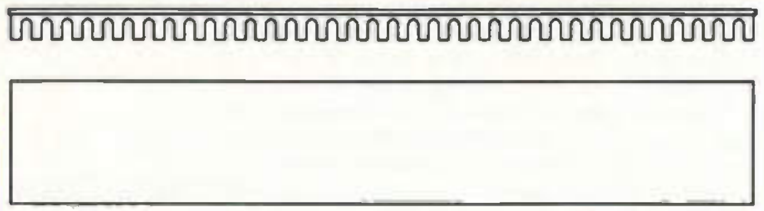
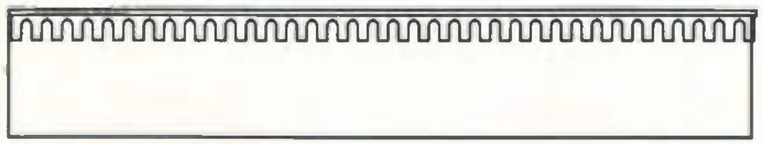


Fig. 7. An enclosure specifically designed for RF work may have many small lips in its lid, each of which will dig into the mating part of the enclosure and provide a low-resistance electrical path between the two parts.



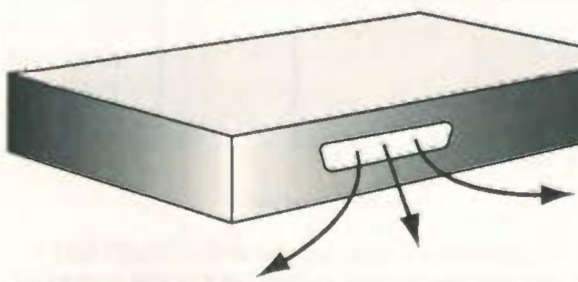


Fig. 8. Slots in an enclosure can provide very effective radiation paths. A slot for a standard 'D' connector is a good example. Such a slot will allow radiation even with the connector fitted.

shell is slightly larger, and designed to fit over the bottom shell.

A pair of tabs on each side of the top shell either overlap or fit into mating notches in the bottom shell. This type of box is suitable for low frequency – up to a few kilohertz – and DC applications that are not particularly sensitive to external EMI.

A somewhat better form of box is shown in **Fig. 5**. The bottom shell is essentially the same as in the other box, but the top shell is built using an overlapping lip rather than tab-and-notch construction. This form of construction is good up to several megahertz, but may fail in the VHF and up region.

Holes in shields

Ideally, a shield should contain no holes. In practice, this is impossible though. There are always some connections – input, output, power supplies – that must go in or out of the shielded enclosure.

In other cases, the circuitry may generate considerable heat so some holes are provided to ventilate the interior. The holes must be very small compared to the wavelength of the highest frequency signal being protected against.

The general rule is that screw or mounting holes should spaced not more than 1/20 wavelength (i.e. 0.05λ) apart at the highest frequency of operation. At 1MHz, this is not hard to meet, because $0.05\lambda>49$ feet. But at VHF and up it might be a bit tricky because the wavelengths are much shorter.

For example, spacing the screws that keep a shield firmly in place three inches apart may be sufficient for mechanical strength, and will shield at lower frequencies. But three inches is 0.05λ at 197MHz. Above 197MHz the shielding effect is therefore reduced.

In the shielded enclosure of Fig. 5 the screws are on the end portions of the flange. At UHF frequencies this can be a problem. A better solution is shown in **Fig. 6** where spacing S is less than 0.05λ.

The effects of wide spacing of mounting screws can be dramatic. I once saw a case where a mechanical engineer had 're-engineered' the specification for an RF enclosure without understanding the RF effects.

The electrical engineer who designed the original box demonstrated the effects to her by taking a well shielded pulsed RF transmitter and connecting it to a dummy load. He then used a spectrum analyser with a whip antenna on it to monitor the energy emitted from the RF box.

He started by removing every other screw. As soon as the first screw was loosened, the harmonics and spurs showing on the spectrum analyser display began to rise. He eventually reached the screw spacing recommended by the mechanical engineer... and at high frequencies the shielding was almost ineffective.

Another form of box is shown in **Fig. 7**. This type of box used a bottom shell that is enclosed on all sides but the top. A top cover with RF 'fingers' can be used to shield the top side. The fingers dig into the metal of the bottom shell, creating a tighter RF bond.

One popular form of this type of box is manufactured by SESCOM, and is made of tinned steel.

Slots in enclosures

Be really wary of slots in shielding enclosures. They are relatively efficient radiators – so much so that some microwave antennas are little more than arrays of slot apertures.

When the slot approaches $^1/_8$ wavelength or longer, then it may radiate rather effectively. This could occur when connectors such as the 'DB-*x*' type used for digital interfaces like RS-232C are mounted to the shielded enclosure, **Fig. 8**.

Connectors are not the only form of 'slot' found in some equipment. If covers or shell halves in aluminium project boxes are just butted together, as shown previously in Fig. 4, then the lack of a tight fit might form a radiating slot. The best solution is to use boxes with an overlapping 'lip' to join the halves together, Fig. 5.

Other accidental slots are created when internal shielding panels are put in place to create multiple shielded compartments, and the mechanical fit is not good. One reason to use copper or brass to make enclosures is that a bead of solder can be used to ensure these panels are firmly anchored to ground with no 'slotting' effects. ■

In the second article on this shielding, I will be looking at mounting connectors, multi-compartment enclosures, spray-on shielding, guard shielding and ground loops.

Anadigm is producing a programmable analogue chip that is quite different from Zetex's Trac device, featured in last month's issue. Described as a 'field-programmable analogue array', Anadigm's chip uses switched-capacitor technology and comprises a configurable matrix of 20 programmable cells. Claudia Colombini describes the technology behind this field-programmable analogue array, highlighting its design and cost benefits.

Field-programmable analogue array

Although a variety of programmable analogue arrays have been available on the market for about the past ten years, none of them has provided the reconfiguration flexibility that makes FPGAs such invaluable devices for digital designers. Now, by combining general-purpose analogue resources with static random access memory configuration logic, Anadigm has created a field-programmable analogue array, or FPAA, that looks set to revolutionise analogue system design.

The AN10E40 FPAA comprises a 20-cell op-amp array arranged in a 4-by-5 matrix, surrounded by a programmable interconnect and I/O structure. The device is packaged as an 80-pin $14mm^2$ QFP, requires a single 5V±5% DC supply, and has a typical power consumption of less than 13mW per active cell.

Many of the more common signal-conditioning functions such as rectifiers, gain stages, comparators and first-order filters can be implemented using just one cell. More complex functions such as high-order filters, oscillators, pulse-width modulators and equalisers can be implemented using two or more cells.

The frequency range of the FPAA depends on the circuit functions being implemented; the amplifiers have a bandwidth of 5MHz and the maximum switching clock rate is 1MHz. Typically, the entire array can handle signal frequencies from DC to 500kHz, making it ideal for filtering, instrumentation and control applications in industrial, medical, automotive and low-to-medium frequency communications markets.

The FPAA's circuit elements are dynamically configured each time the device powers-up, using data held in on-chip SRAM; the SRAM is loaded automatically direct from a low-cost serial EPROM during the power-up sequence. Digital field-programmable gate arrays are configured in a similar way.

Alternatively, the FPAA can be reconfigured on-the-fly by data from a microcontroller, making it an extremely versatile and space-saving component. Reconfiguration can be accomplished within 100μs. This is more than fast enough to allow, for example, several signal inputs to be multiplexed to a single analogue signal conditioning circuit. The ability to handle in-service changes to configuration – or even functionality – brings unprecedented flexibility to the world of analogue design.

Filter design made simple

Anadigm has recently added a filter synthesis tool to its free FPAA configuration software. Known as FilterDesigner, the new tool provides users with an extremely versatile means of creating high-order classical filters and then combining them with additional signal conditioning circuitry to implement complete, single-chip, analogue solutions.

The tool enables high-pass, low-pass, band-pass and band-stop filters to be created in minutes, using combinations of the standard bi-linear and bi-quad filter configurations from the AnadigmDesigner analogue function library.

Optimised technology

Each configurable analogue block, or CAB, comprises an amplifier surrounded by a switched capacitor feedback network, as shown in Fig. 1. These binary-weighted capacitor clusters can be set to any one of 256 different values. It is the use of this technology that is key to the FPAA's versatility, enabling highly stable *RC*-equivalent networks to be implemented using just switches and capacitors.

Figure 2 contrasts a conventional resistor-based circuit with a switched capacitor circuit; the charge that is transferred from node 1 to node 2 depends (to the first approximation) on the capacitor's value and the switching duty cycle, effectively making the FPAA an analogue sampled-data device.

As a point of interest, the AN10E40 is an all-CMOS device. Although it is difficult to fabricate semiconductor capacitors with accurate absolute values, it is relatively easy to ensure that all capacitors on the chip have exactly the same value – which is precisely what's required for this application.

The effects of temperature and ageing are alleviated by the fact that all components are on the same die and therefore track each other. The AN10E40 FPAA also includes 13 buffered analogue I/O cells, two uncommitted op-amps, four programmable clocks and a programmable voltage reference source.

Each configurable analogue block can connect to adjacent neighbours, and there are also 10 horizontal and 12 vertical routes for global connections; each block can drive up to eight adjacent blocks and an I/O buffer.

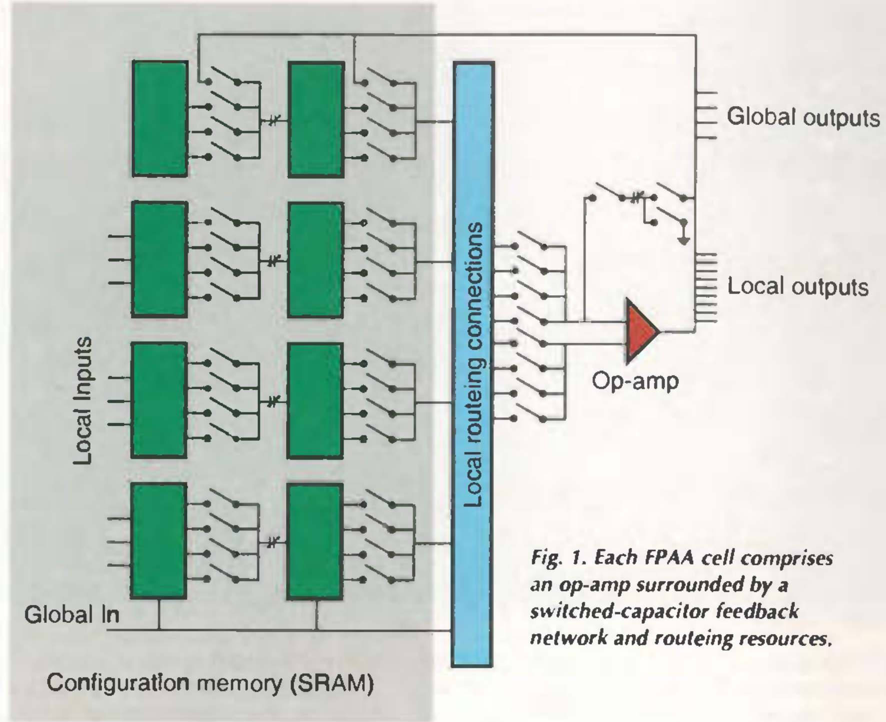


Fig. 1. Each FPAA cell comprises an op-amp surrounded by a switched-capacitor feedback network and routeing resources.

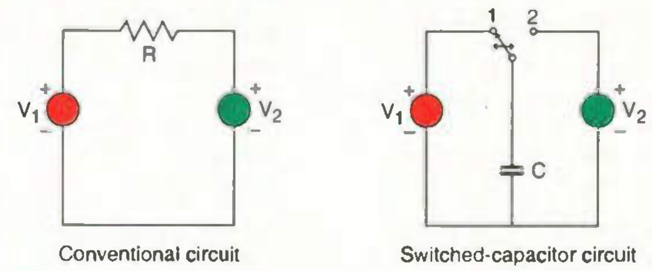


Fig. 2. Switched-capacitor technology substitutes capacitors and switches for resistors to enable digital control.

'Drag-and-drop' design ease

Designing an analogue system based on the AN10E40 FPAA demands minimal circuit knowledge, analogue simulation skills or maths abilities. A free CAD tool known as AnadigmDesigner enables the entire 'design', simulation and FPAA SRAM download process to be accomplished in as short a time as 10 minutes. You can download it from www.anadigm.com if you want to find out for yourself.

The software runs on a standard PC, and includes a library of more than 50 configurable analogue circuit functions that range from simple amplifiers, comparators, integrators and differentiators through to complex functions such as bi-quad filters. Most of the functions consume just one of the FPAA's 20 cells, and none takes more than three.

Building an analogue circuit is simply a case of selecting the appropriate analogue functions from the library. This is done by 'dragging and dropping' them onto the screen display of the complete array, and 'click-dragging' appropriate signal interconnects.

Performance characteristics of each function are specified via pop-up dialogue boxes. Should a chosen signal interconnect route not be available, the software automatically advises the user to choose an alternative.

The software also provides facilities for programming the array's clock generators and voltage reference source. There are also facilities for connecting or disconnecting a mid-rail voltage reference source, or VMR, to the array's analogue signal ground.

This approach to design means that users do not need to worry about the underlying circuit implementation. For example, to build a signal-conditioning chain, users simply select a summing amplifier and filter from the function library. They then specify the desired offset correction, gain and low-pass frequency parameters, and interconnect the array cells.

AnadigmDesigner includes a

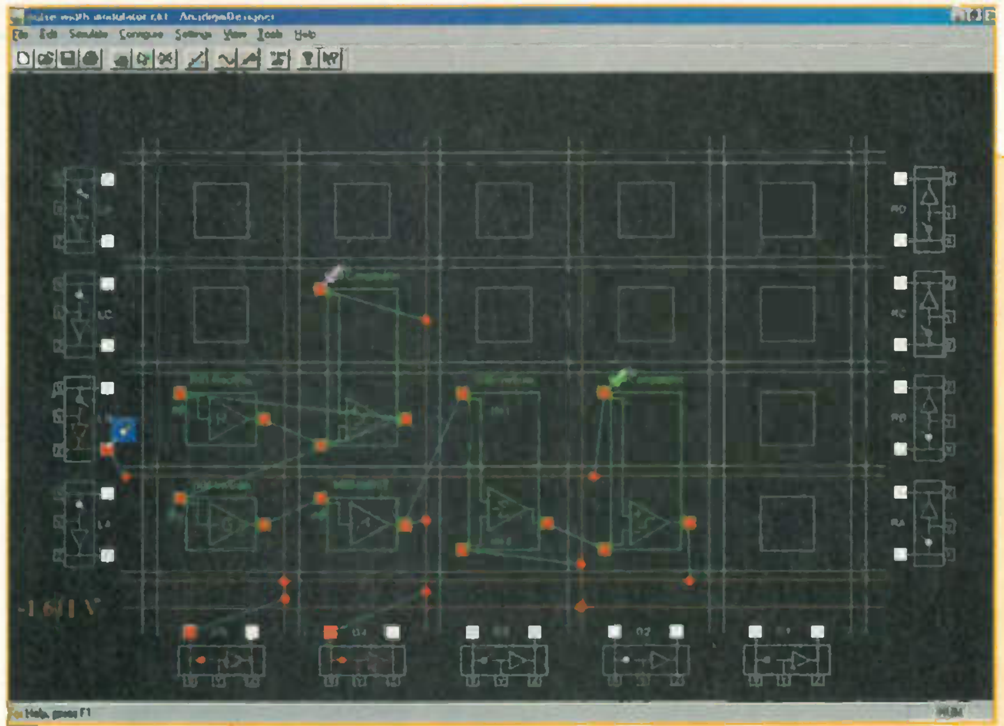


Fig. 3. Screen shot of the FPAA's configuration software.

built-in simulator that allows a virtual signal generator and oscilloscope to be placed strategically within the array, **Fig. 3**. The signal generator provides standard waveforms such as sine, square, triangle and ramp, as well as pulse outputs and arbitrary waveforms.

The software also accepts linear descriptions of arbitrary waveforms in numerical text-file format – just like many of the digital storage 'scopes on the market. It is capable of processing Windows' standard pulse-code-modulation encoded WAV-format files.

Once the design has been functionally checked with the simulator, the configuration data can be downloaded to the FPAA's SRAM and the circuit's 'real world' performance evaluated using standard bench test equipment. This 'electronic breadboard' approach to analogue design reduces development time and costs significantly, helping users to accelerate the introduction of their products to market.

Flexible development environment

Electronics engineers who want to evaluate FPAAs – especially if they are interested in exploring the concept of adapting FPAA functionality in the field – will probably choose to use the AN10DS40 FPAA development board, **Fig. 4**. This provides an AN10E40 FPAA and an RS-232 interface to facilitate downloading of configuration data from the PC running AnadigmDesigner software.

The board is equipped with numerous connectors, interfaces and status LEDs, and includes a socket for a serial-boot PROM, as well as an on-board 1MHz oscillator and a regulated +5V supply. It also incorporates a Motorola 68HC908 microcontroller to allow dynamic modification of FPAA functionality, and a standard peripheral interface which enables users to employ different microcontrollers or host systems if they wish.

In a second article in *Electronics World*, I will be taking a more detailed look at the filter-design tool. I will describe how to build an eighth-order band-pass filter and allied signal processing system to implement a universal single-chip programmable tone-detector circuit for telecoms applications. ■

Fig. 4. The field-programmable analogue array development board.

More information

For more information on the FPAA, have a look at www.anadigm.com

For details of how to order, contact: **Anadigm Ltd, 3-5** Mallard Court, Mallard Way, Crewe Business Park, Crewe CW1 6ZG, UK. Tel: +44 (0)1270 531 990, fax: +49 (0) 8142 4485840

Exclusive reader offer:

Own a development system for field programmable analogue arrays for £249+VAT – normal price £349

The subject of this month's exclusive reader offer is more than the average development board. Anadigm's development system for field-programmable analogue arrays includes a microcontroller to investigate programmable-analogue technology as a peripheral to an embedded system.

The regular cost of this designer's kit is £349. Anadigm will discount this by almost 30% to £249+VAT for *EW* readers – and include free postage and packing for anyone in the EU. (Interested readers from outside the EU please e-mail for shipping details rebecca.shade@anadigm.com).

This offer is valid until 31 December 2001. *Electronics World* dispatched by surface mail can take a while to reach its destination so the cut-off date for claims from outside the UK has been extended to 31 January 2002.

What do I get?

The kit includes the FPAA board, power supply, cable and manual. The FPAA configuration software is available as a free download from the web site.

The FPAA development board provides an environment for rethinking analogue electronics strategy. It introduces concept of analogue system as reconfigurable peripheral of a microcontroller.

Designated AN10DS40, the board provides a range of resources to help electronics engineers evaluate FPAA technology and develop working systems. It additionally incorporates an onboard microcontroller to demonstrate the way that FPAA functionality may be adapted in the field, to help users understand the technology's potential for radically reshaping the way electronics products function.

The board provides an FPAA – with its 20 configurable analogue cells – and a serial interface for PC connection. This interface allows programs created using the free AnadigmDesigner CAD package to be downloaded.

Also onboard are numerous connectors, interfaces and status LEDs to simplify development, interconnection and test, including stereo jacks for convenient interfacing in applications involving audio signals.

Users can develop two kinds of FPAA-based analogue systems. The first is a fixed-function FPAA to integrate discrete analogue component-based circuitry, which boots from a serial EEPROM – for which a socket is provided.

HC08 microcontroller

The AN10DS40 board's powerful HC08 microcontroller can dynamically modify FPAA functionality by reloading a new device configuration file – an operation taking just 100μs. This feature allows users to explore the concept of adapting analogue performance in a software-controlled, event-driven fashion.

Four pushbutton switches are provided to manually trigger interrupt-based reconfiguration, to simplify the real-world test of this innovative new capability.

A standard peripheral interface (SPI) is also provided on the AN10DS40 board to allow the FPAA to be controlled from an alternative microcontroller – and/or a user's own prototype hardware.

Dynamic reconfiguration can be used by an engineer to radically improve product performance, and lower costs – potentially replacing multiple PCBs with one chip.

For example, a general purpose data acquisition board could reconfigure its front-end signal conditioning for different sensors sequentially – as it scans channels – providing major savings in both PCB space and cost.

Alternatively, FPAA functionality could be modified according to operating conditions, such as a change in light level, providing analog designers with a practical low cost method of implementing real-time adaptive capability for the first time.

The EEPROM-based configuration method also provides designers with considerable flexibility, allowing one standard PCB to be configured for different applications at the end of the production or during installation.

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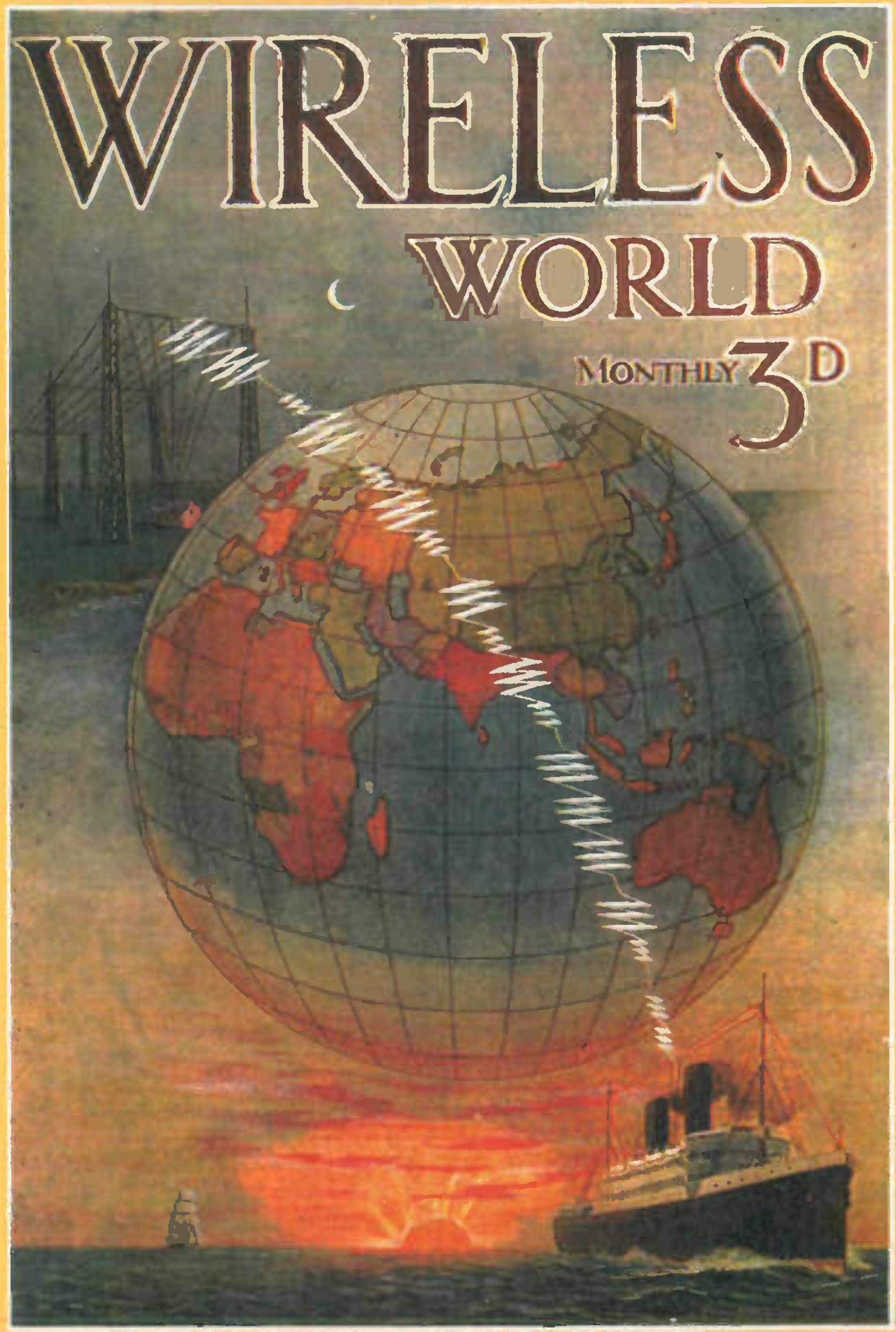
Wireless across the waves

Even as early as 1904, society was seeing the benefits of wireless. Anthony Hopwood has been looking at the first Cunard Daily Bulletin – a newspaper for transatlantic liner passengers that could not have existed, were it not for wireless communication.



Front cover from the second issue of Electronics World's predecessor.

In this centenary year of Marconi's first transmission across the Atlantic, it's interesting to look back 97 years to June 1904, when the first regular daily newspaper appeared aboard a transatlantic liner, showing how quickly the new medium had become established.

The ship was the Cunard liner RMS Campania, which left Liverpool on Saturday 4 June for New York with Marconi on board.

The first issue I have, No 32 (because earlier Bulletin numbers were single issues per trip) is dated Sunday 5 June and states:

> *" although it was not intended to publish the Cunard Daily Bulletin before Monday morning, the Cunard Company decided to print a limited number of Souvenir Bulletins for distribution to the Press..."*

Needless to say, these issues were jealously guarded by their recipients.

The eight-sided bulletins were printed via photogravure. Sunday's issue records the distance from the Poldhu station as 220 miles at 1pm. The news by Marconigram, 'direct to the ship' covers events in the Russo-Japanese war, including the use of carrier pigeons by the Russians under siege at Port Arthur. American news mentions floods in Kansas with all traffic on railways and tramways suspended.

Business in Wall street was 'stagnant and featureless'. Tabloid interest was met by a story that Frank J. Young, 'a well known bookmaker and horse owner' was fatally shot in a New York cab on his way to the White Star Line pier where he intended to join his wife to sail to Europe. His companion in the cab, a Mrs Nan Peterson told the coroner that the deceased 'had shot himself without taking the pistol from his pocket'.

The story develops in the 'Stop press' section – "The coroner has committed Peterson (ex-actress) to the Tombs without bail". One wonders if the story would have featured at all if Mr Young had been booked with Cunard!

Daily changes

Side six of the Bulletin also changes daily. Under the strapline, 'Smoke-room gossip', it seems to be a resting place for old music hall jokes and shaggy dog stories. Here's a sample:

> *"A man bought a horse at a horse fair, and found the dealer had short changed him by half a crown. He remonstrated with him and was told – 'Naw sir, we never gives money back, but you can have an extra horse if you like.'*

You get the idea.

On the 6th, radio communication was established with the Canada Marconi station at Cape Breton when the ship was still 2000 miles from New York. Apart from the Russo-Japanese war news, the positions of icebergs reported by other ships were given.

On the 7th, it was reported that striking miners at Cripple Creek, Colorado had gone on the rampage and killed 12 while placing dynamite under the railroad station. In a subsequent battle with State troops at least 22 were killed. The West was really wild in those days.

Aboard ship, things were quieter, with the 505 third class passengers' breakfast menu promising:

> Oatmeal porridge and milk – Golden syrup.
> Smoked red herrings. Beef steak and onions.
> Boiled jacket potatoes,
> Fresh bread and butter -
> Marmalade.
> Tea or Coffee.

In the high summer of Empire, cargo details that would be secret today were a matter of pride, and published in the daily Bulletin. Apart from 8000 packages of fine manufactured goods and 1650 mail bags, there were several parcels of diamonds and sapphires worth about £10 000, and £5000 worth of minted silver coin in 17 large cases. To deter opportunist thieves, "the precious stones and coin are securely stored in the ship's strong room."

On the 8th, Lord Inverclyde, Chairman of Cunard sent the following message to Marconi on board;

> *"I trust this message will reach you promptly and wish you continued success with your Wireless Telegraphy."*

During the 8th, while still some 1000 miles from New York the running log recorded that from 4am to 8pm, the sea temperature rose from 58 to 67°F as the ship entered the Gulf Stream.

On the 9th, the sporting news section reported that the amateur champion Travis was playing 'inferior golf' at the British Open, and subsequently retired.

Elsewhere, the Canadian Cape Breton Marconi station reported that Mary Virginia Rhodes, beneficiary in the Will of the late Cecil Rhodes, had been found at Washville, Carolina working as a missionary.

In a postscript to the first week's run of the Daily Bulletin, the Editor remarks that

> *"Wireless Telegraphy has indeed struck a staggering blow at the hitherto absolute power of Father Neptune, so that he is no longer in supreme command of his own domains."*

A sharp reminder that the old tyrant was still at work would come eight years later, when the White Star liner Titanic foundered on a clear cold calm night. Wireless was only able to save a third of those aboard after a Cunard liner, the Carpathia, scorched the paint off her funnel in a dash to rescue the survivors.

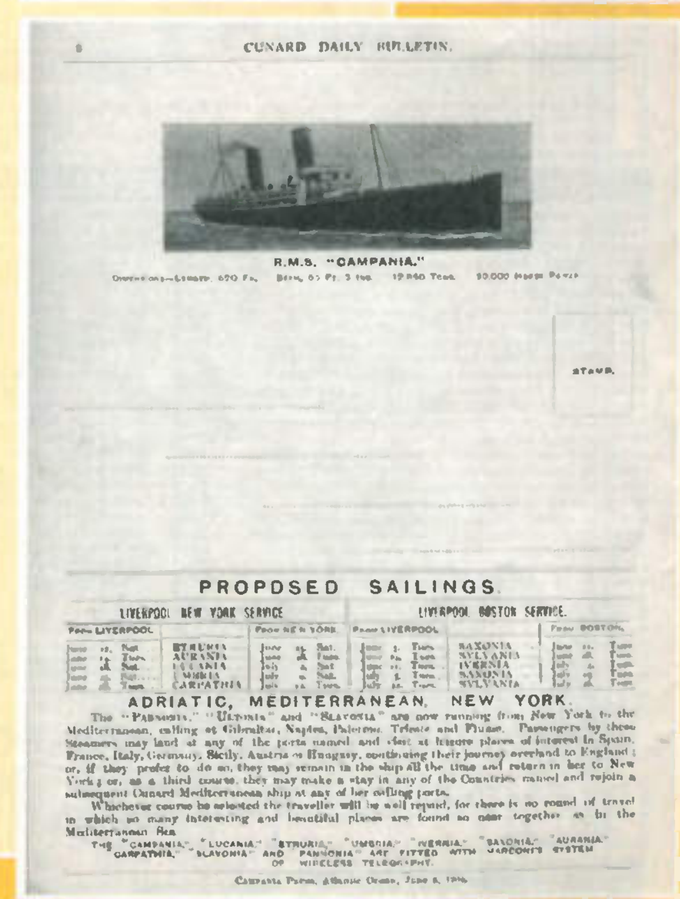
In the single 'Arrival Supplement' sheet on 11th, there is a less than prompt reply from Mr Marconi to Lord Inverclyde's message on the 8th.;

> *'Sincere thanks for your cordial wishes received Wednesday mid-ocean. Happy inform you Daily Bulletin entirely successful and greatly appreciated'.*

Even Marconi was watching call charges at 7/6d (38p) a word! ■


Cunard Daily Bulletin.

No. 34 R.M.S. "CAMPANIA," JUNE 8, 1904.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R.M.S. "CAMPANIA."

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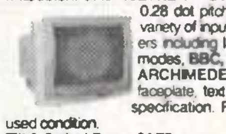

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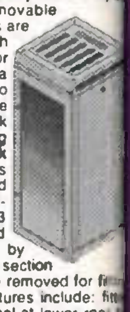

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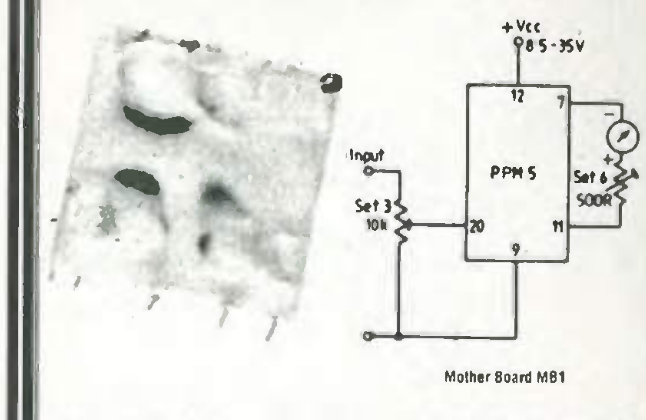

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Differential-in 100MHz scope probe

Designed for RF test and measurement, Cyril Bateman's differential-input scope probe system provides useful results at more than 100MHz and features switchable gain together with 4pF loading. As with Cyril's previous designs, this one follows the philosophy of high performance at low cost.

All voltage measurements take place between two points. Using a battery-powered multimeter to measure DC or low frequency AC, both measured points may differ from ground. However at higher frequencies – and especially using a conventional mains powered instrument – measurements are usually referred to one common point, electrical ground. Consequently, using a conventional oscilloscope to observe a voltage waveform, you are constrained to making ground-referred measurements.

Many two-channel oscilloscopes include a facility to add or subtract their 'A' and 'B' channel waveforms. This feature allows the voltage differences between two points to be displayed.

In many cases, if your oscilloscope can accurately subtract out the common voltage, you are able to display the difference voltage[1]. However this method has limitations:

- Both the 'A' and 'B' signals must be able to simultaneously 'fit' on the oscilloscope screen. This limits the maximum screen display sensitivity that can be used. If the required difference waveform is small relative to the 'A' or 'B' voltage waveforms, this difference waveform simply cannot be seen.
- Both oscilloscope input amplifier channels and any probes used must be carefully matched for gain and phase over frequency – certainly to better than 1%. However the thickness of most screen traces make these adjustments difficult – if not impossible. As a result, relatively small input amplifier gain variations can dramatically change your displayed waveform[2].

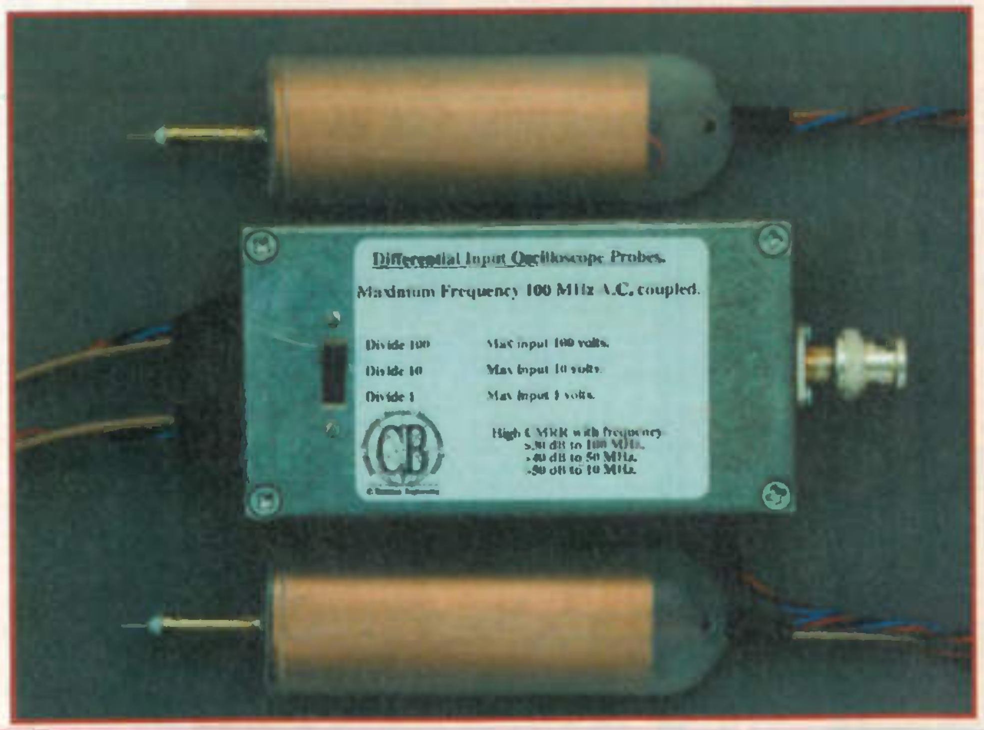
In an attempt to avoid these problems, many engineers disconnect their oscilloscope's earth, hoping that 'floating' the oscilloscope will suffice. Apart from causing a safety hazard, this imposes an ill-defined and unbalanced load on the test circuit.

The only satisfactory way to observe the difference waveform between two voltage points is by using a differential-input amplifier with phase-matched probes or a differential probe attachment.

Such equipment is available commercially, intended either for low-frequency, high voltage measurements or for high-frequency but lower voltage measurements. Both types can be extremely expensive.

This article describes the design and assembly of two 100MHz AC-coupled, high-impedance probes having less than 4pF capacitive loading. Each can be switched to accept maximum common-mode inputs of 10V or 100V. The article also describes a matching differential input, low-noise, low-distortion amplifier, having switched gains of 1 or 10, Fig. 1.

Fig. 1. Low-capacitance differential probes and amplifier, provide excellent performance from audio to 100MHz. When measuring single-ended signals using one probe only, a short ground lead should be clipped to the test prod body. For differential use, ground leads are not needed.



Design details

Each probe is designed to be hand held and is connected to the differential amplifier using a one metre length of RG179B/U coaxial cable.

Both probes are powered with ±5V obtained from the differential amplifier. Each includes a range-switching relay, controlled from a switch on the case of the differential input amplifier.

The differential-input amplifier is designed to mount directly onto the oscilloscope's front-panel BNC input connector. Powered from a nominal ±9V, it includes two 'three-leg' linear stabilisers that provide a stabilised ±5V for its own use and to power both probes.

The three relays used are each powered from the +9V input via 180Ω current limiting resistors. They are controlled by a switch to select between gains of ÷100, ÷10 or unity.

Common-mode rejection performance

This design ensures an excellent common-mode rejection. It is more than 70dB to 10MHz, 60dB to 30MHz, 50dB to 90MHz and 49dB to 120MHz. These figures are relative to a 1V input with the attenuator set to divide by ten.

At lower frequencies, CMR exceeds 80dB, being limited mostly by noise. Using a 3MHz bandwidth voltmeter, divide by 10 or 100 noise measured 0.07mV and 0.62mV in unity gain mode.

This low noise allows you to use an oscilloscope display sensitivity down to 10mV/cm. Small difference voltages can be displayed – even with common-mode voltages up to 100V.

With an input common mode voltage of 10V and the differential input amplifier set to unity gain, a 10mV peak difference voltage is clearly visible, **Table 1**.

Using the maths facility of a dual-channel oscilloscope with conventional probes, a 10mV difference signal in a common-mode voltage of 10V is completely invisible. In this case, the difference voltage would have to approach 0.5V for clarity.

Excellent CMR and the extremely flat frequency response allow a single probe to be used. This can replace a conventional scope probe, providing the benefit of a much lower capacitance loading on the test circuit.

Table 1

Measured performance of completed probes and differential input amplifier

Gain error by frequency using Probe 1 as input				CMR with differential probe input,		
Gain select	÷100	÷10	÷1		÷10	÷1
Frequency (Hz)	Input 10dBm	Input 10dBm	Input 0dBm	Frequency (Hz)	Input 1 volt	Input 1 volt
1k	0dB	0dB	0dB	1kHz	–86.4dB(5V)	–74.7dB
1M	0dB	0dB	0dB	1MHz	–87dB	–63dB
10M	0dB	0dB	0dB	10MHz	–73dB	–59dB
20M	–0.05dB	0dB	–0.1dB	20MHz	–66dB	–54.5dB
30M	–0.1dB	0dB	–0.1dB	30MHz	–62dB	–52dB
40M	–0.4dB	0.05dB	–0.1dB	40MHz	–58.5dB	–50dB
50M	–0.2dB	0.2dB	–0.2dB	50MHz	–61.5dB	–52dB
60M	–0.1dB	0.2dB	–0.25dB	60MHz	–56.5dB	–48.5dB
70M	–0.8dB	0.1dB	0dB	70MHz	–56.5dB	–45.5dB
80M	–1.2dB	0dB	–0.4dB	80MHz	–51dB	–41.5dB
90M	–1.8dB	–0.3dB	–0.5dB	90MHz	–51dB	–39.5dB
100M	–3.0dB	–1.2dB	–0.75dB	100MHz	–49.5dB	–37dB
110M	–4.0dB	–3.2dB	–4.0dB	110MHz	–49dB	–36dB
120M	–7.5dB	–5.5dB	–6.0dB	120MHz	–49dB	–36.5dB
Max safe input (volts)	100V	10V	2V		10V	10 V

Equipment used: HP331A voltmeter and the HP8405A vector voltmeter both have high impedance low capacitance inputs. The 331A meter is specified to 3MHz, the 8405A from 1MHz to 1GHz. To ensure consistent test signal voltage with frequency, the signal generator output was loaded with a 10dB attenuator and a Macom 50Ω through termination. The HP331A fitted with a divide-by-10 probe and the HP8405A were used to monitor test signals at this termination. See box 'Measurement equipment'

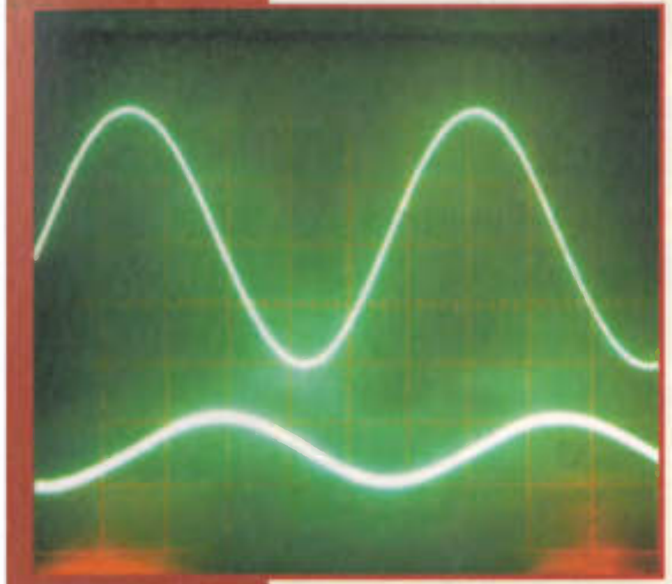


Fig. 2. Top trace, a Coline 250MHz divide 10 probe shows output at 0.1V/cm, from a TL071 voltage follower. Bottom trace at 0.01V/cm clearly shows a voltage difference between the signal input, pin 3 and output pin 6. This difference cannot be seen using the A-B facility on my scope.

Fig. 3. Using an identical set-up to that used for Fig. 2, but with the sinewave replaced by a triangular wave, shows a square wave input-output difference, as the op-amp struggles to follow this input signal. Even this difference waveform is not visible using the A-B facility on my 100MHz scope.

Applications of the probe

This differential design allows you to measure small voltage differences between two points on your circuit. It provides a simple and quick means of ascertaining in-circuit distortions and actual op-amp input output voltage differences.[3]

As an example, suppose you configure a unity gain voltage follower using a TL071 IC with a 1kΩ load resistor. Supply voltage is ±10V and input is a 4V pk-to-pk sinewave at 100kHz. Using my divide-by-10 Coline 250MHz scope probe with the 'A' channel's 'Y' gain set to 0.1V/cm, the output waveform in Fig. 2, top trace, is identical to the input waveform. Or is it?

The lower trace is the result of a differential voltage measurement between pins 3 and 6 of the TL071. For this trace, the differential probe was also set to divide by 10 but the 'B' channel's 'Y' gain has been set to 0.01V/cm. This clearly shows that a difference does exist between the input and output waveforms.

Even more dramatic results are obtained if the sinewave is replaced with a triangular waveform at the same amplitude and frequency, but leaving everything else as above, Fig. 3. Using a difference voltage measurement, you can quickly explore the effects of changing load impedances and waveforms on your circuits.

Throughout this article I have concentrated on using this differential probe with an oscilloscope, because that seems its most natural application. Because of its excellent single probe accuracy and differential CMR performance though, it can equally well be used as input to an AC voltmeter having with an oscilloscope equivalent input impedance of 1MΩ and 18 pF.

Differential probe design – ICs

Fundamental to this design was the choice of differential input amplifier to be used. In the past I have used the AD830 amplifier from Analog Devices[4]. The AD830 provides an excellent 100dB CMR at low frequencies, but by 10MHz this has reduced to 50dB.

To provide operation to 100MHz, I needed better high-frequency performance. It would also be preferable to have a gain of two to offset the loss incurred through using a back terminated cable between the probes and the amplifier.

I searched Internet and distributor catalogues and eventually settled on Maxim's MAX4144.[5] It is designed as a high-speed, low-distortion, differential line receiver with a gain-of-two bandwidth of 130MHz. It provides 70dB CMR at 10MHz, 50dB at 80MHz and more than 45dB at 100MHz. Its –90dBc spurious free dynamic range at 10kHz and some 12nV/√Hz noise at 1MHz looked worthy of further investigation.

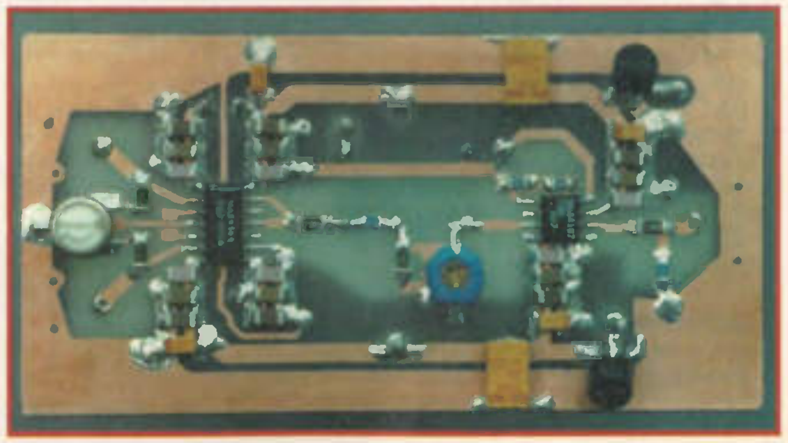
I now needed a 100MHz bandwidth, low-noise, low-distortion, output amplifier with a gain of 10 for an overall unity gain system, Fig. 4.

I choose Maxim's MAX4107 – an ultra low noise 300MHz op-amp. I decided to use it in a variation on the MAX4106 design that proved so successful in my unity gain, low capacitance 100MHz, active oscilloscope probe[6].

The 4106 is optimised for a gain of 5, while the 4107 is optimised for closed loop gains of 10 and above. With an input noise level of just 0.75nV/√Hz at 10kHz, a large spurious-free dynamic range, low distortion and 500V/µs slew rate, this IC looked a good starting point for my design.

The choice of IC for my ÷10, ÷100 probes was easy. Maxim's MAX4005 had already proved successful in my unity gain probe design[6].

Fig. 4. My second prototype PCB for the differential input amplifier. Note the extensive decoupling capacitance and enlarged power supply tracks, introduced after the failure of my first PCB. With the later addition of three small COG capacitors, this PCB was used for all performance test results.



Design – relays and trimmers

Choosing relays suitable for the gain switching proved quite difficult. As a first requirement the relay must be physically small in order to mount in a hand held probe and switch using a 5V supply.

More important however, to design a system capable of working to 100MHz, the effects of self inductance and self capacitances of the relay contacts on circuit performance must be minimised. Specifications for these parasitics were not available for any relay that I considered. But it was vital that I took these parasitics into account in my frequency domain simulations.

Contact capacitances at low frequency and resistance at DC were easily and quickly measured. Contact inductance was much more difficult to determine. In the event I had no choice but estimate contact inductance for my simulations. Using the relay's physical dimensions and

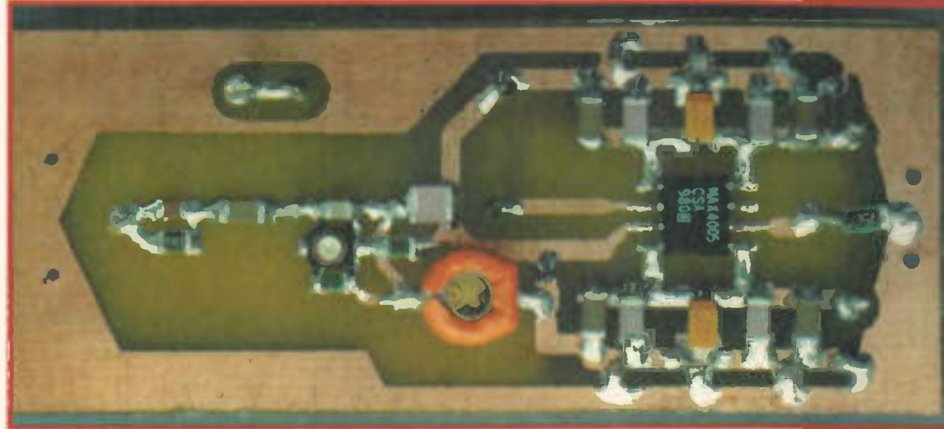


Fig. 5. This first probe PCB prototype was modified and used for the final build and measurements. Note the 150pF 1210 COG ceramic adjacent to the orange trimmer. Resulting from tests, the earth tracks from this capacitor were later 'improved'. These differences can be seen by comparing with Fig. 7.

Calibration

Both probes should be assembled then pre-calibrated separately as stand-alone items.

Solder a temporary 75Ω 0805 surface-mount resistor from each probe output to its ground plane and take all output voltage measurements across this resistor.

Apply ±5V power, taking care to note the location of the +5 and –5V terminals on the PCB. Set probe capacitor C_6 to mid value.

Input a known 1kHz signal not exceeding 10V to the test prod from a signal generator. This generator should be terminated in a through-load matching its output impedance. Measure and note the voltages across the temporary 75Ω resistor when the relay is energised from +5V and when de-energised.

With the same generator output voltage, increase signal frequency to 1MHz and energise the relay. Using a non metallic trimmer tool, slowly adjust C_4 to attain the same relay energised output voltage as noted for 1kHz. This is an extremely sensitive adjustment, tiny movements of C_4 will affect output at 1MHz and above.

De-energise the relay and adjust C_6 to attain the same de-energised relay output voltage as noted at 1kHz.

These trimmer adjustments affect high-frequency performance and can reduce the CMR at higher frequencies. Take care to match the 1kHz and 1MHz levels for each probe.

Since C_4 and C_6 interact, repeat the above 1MHz adjustments as needed.

Remove the temporary 75Ω resistor from each probe. This completes the probe pre-calibrations.

Connect both probe outputs, via matched lengths of 75Ω coaxial cable to the differential amplifier inputs. I used metre lengths of RG179B/U because its PTFE inner facilitated direct soldering to the PCB ground planes and simplified cable length matching. Matched coax lengths are needed to minimise phase differences between the probes at high frequency.

The coax cable from the signal generator should be terminated in a through load. Fit oscilloscope probe 'BNC' adapters to each probe's test prod. Attach a 'T' adapter to this load to provide two equal-length BNC output paths. Attach one probe with its BNC adapter to each output such that each probe 'sees' the identical test voltage and signal delays.

Apply a known 1kHz voltage not exceeding 10V, to both test probes. Energise the relays in both probes but do not energise the relay in the differential amplifier.

Using a suitable voltmeter, monitor the differential amplifier output. Carefully adjust VR_1 for best CMR, minimising the differential amplifier output. This should be much less than 1mV for a 10V input.

Remove one probe's test prod from the signal source and attach a 50Ω terminator to this prod.

With a 1V 1kHz signal applied to the second probe's test prod, check output while switching the relays. The differential amplifier should now output exactly 1V, 0.1V and 0.01V respectively.

Apply a known signal, up to 10V at 1MHz to the above probe. Energise the probe relays but not the differential amplifier relay. Note the voltage output from the differential amplifier.

Increase frequency to 30MHz. Adjust C_{15} in the differential amplifier for the same voltage as noted at 1MHz.

Reconnect both test prods to the 'T' adapter and input a known 1MHz signal to both probes. Energise both probe relays and monitor the differential amplifier output. Calculate CMR as $-20\log(V_{out}/V_{in})$.

If you can adequately measure this small output voltage, then very slowly and gently, slightly adjust C_4 in probe 2 only, to maximise CMR.

This completes all necessary calibration.

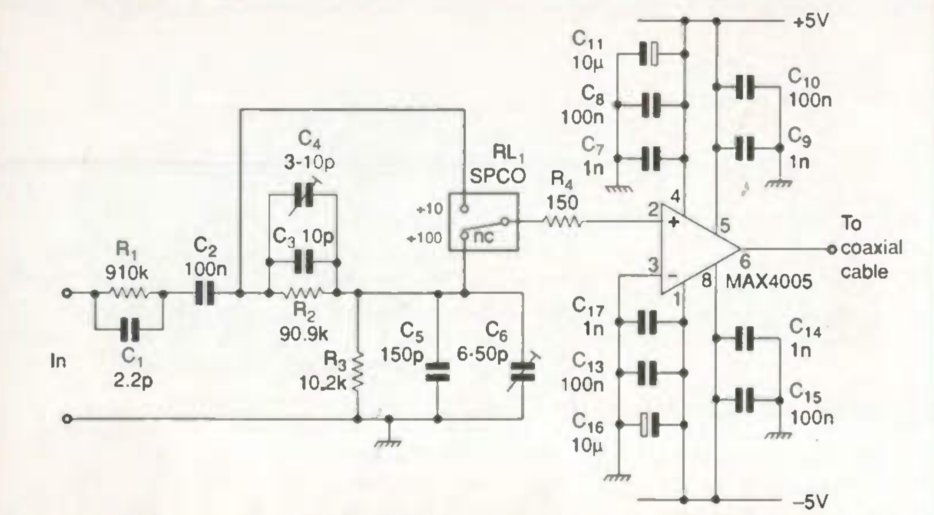


Fig. 6. Schematic for both test probes. Note the 150Ω resistor R_4 now inserted between the relay and input of the MAX4005. This resistor, combined with the IC input capacitance, damps out a very high frequency resonance between the relay contact inductance and this input capacitance. All signal path capacitors are COG ceramic.

assuming the inductance might be similar to a single turn coil, I estimated a value.

At least one trimmer capacitor would be needed for each probe and the differential amplifier. The differential amplifier would also need a pre-set resistor, to compensate for variation of probe gain and maximise low-frequency CMR.

Again I was unable to find any self-inductance specifications for these components, so I prepared estimates based on the component's physical sizes.

At high frequency, each component pad and track introduces capacitance and inductance. Each PCB layout or component change should be reflected in the simulation, **Fig. 5**.

Implementing the design

Developing a prototype proved extremely time-consuming, requiring repeated refining of PCB layout and re-simulation with amended circuit strays. This highlights only too clearly the difficulties of designing moderately high frequency circuits using an essentially time-domain Spice-based, low-frequency simulator.

For my needs, I cannot justify the cost of obtaining a 'proper' frequency domain or harmonic balance high-frequency simulator and related PCB design software. In the past I have used the ubiquitous 'Touchstone' and the 'MDS' microwave design systems.[7]

Such specialised CAD packages do much to facilitate high-frequency simulations and PCB layouts, but they still need to be fed with the correct modelling data. When it is not available from the component maker, data can be obtained from practical measurements, using perhaps a HP8753 vector network analyser.

To de-embed component data from measured values, a pre-calibrated test jig suited to the part being measured is essential. From my work measuring capacitors and EMC filters to 3GHz, I know that designing and calibrating high-performance test jigs, is extremely difficult and time consuming.

However repeated simulations and PCB layout refinements culminated in a prototype printed board assembly I believed was ready for performance testing.

Prototype PCBs for the probes

My original hope was to use close-tolerance capacitors and resistors, with just one trimmer capacitor to compensate the switched attenuators for stray capacitances. However during my simulations I realised this would not work out and two trimmer capacitors would be needed. Both trimmer values would then interact, **Fig. 6**.

By choosing suitable capacitor and resistor values I was able to arrange for simple calibration at two low frequencies only, 1kHz and 1MHz. At 1kHz these trimmer capacitors have no effect. Attenuation can be accurately measured at the probe output, for both relay settings, when subject to a known input voltage.

Increasing frequency to 1MHz with this same input voltage, the trimmer capacitors should be adjusted to attain output voltages identical to those measured at 1kHz. However because of their interaction, these trimmers will need re-adjusting in turn several times, as the relays are switched.

With the trimmer capacitors set correctly for 1MHz outputs, my initial probe PCB, size 63mm by 25mm worked well. It provided flat response to 50MHz. At 100MHz, a 1dB rise in output was measured when the system was set to divide by 100, Fig. 5. This rise was caused in part by an unnecessarily inductive PCB track associated with C_5.

Soldering in place a scrap of copper foil easily modified the board. This modified track is shown in the figure, but the photographs were taken prior to this modification, **Fig. 7**.

Prototype PCBs: differential amplifier

My initial differential-amplifier PCB required more substantial modifications, such that the first board was eventually scrapped. Following careful in-circuit probing using the HP8405 vector voltmeter and repeated simulations, the layout was improved. Revised boards measuring 68mm by 40 mm were assembled and used for the test results. Fig. 4.

The MAX4144 and MAX4107 together draw some 28mA from ±5V. When driven to produce ±2V output, the decoupling used on my original board was inadequate. Extra capacitance was needed and the 1mm wide tracks used to supply power to each IC had to be widened. Finally the impedance of the two-sided ground plane had to be reduced.

Decoupling capacitance and ground plane improvements were made on my original board. Copper shim was soldered around all four sides to bridge top and bottom ground planes together. The design then worked well, but looked decidedly 'modified'.

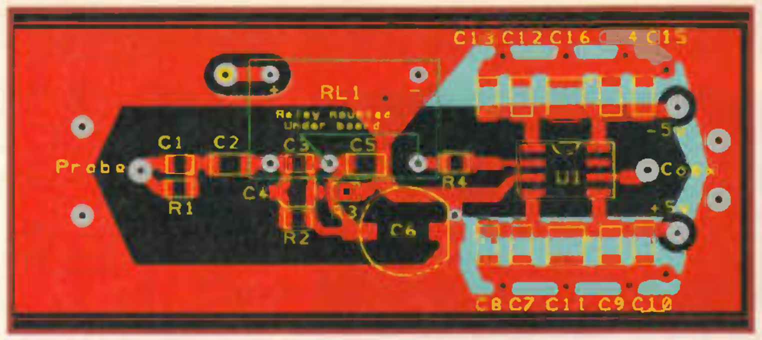


Fig. 7. Final double-sided PCB layout as used for the divide-by-10, divide-by-100 low-capacitance probes. The PCB track modifications with the addition of the 150Ω R_4 provided a flat gain and excellent consistency between switch ranges. At the highest frequencies, the self inductance of trimmer C_6 becomes significant.

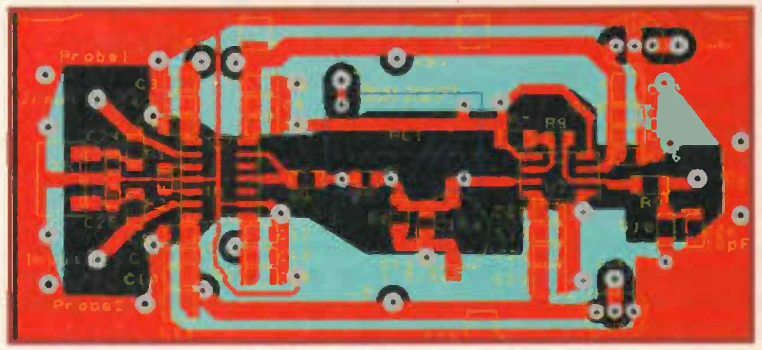


Fig. 8. Revised double-sided, differential amplifier PCB now shows the three capacitors introduced to reduce excess high-frequency gain and noise levels when viewed on an oscilloscope. The 18pF added to the output attenuator, enables operation with a wide range of capacitive loads, without compensating adjustments.

Frequency-domain simulation

The capacitor, resistor and inductor models built into Spice based simulators assume ideal loss-free components having a constant value regardless of frequency.

Some simulators include a relay model, but only for the switching mechanism's delay and contact bounce. This is no help at all with contact capacitance or self-inductance.

For transient or time-domain simulation, Spice automatically provides a facility for amplitude dependent changes for semiconductors, but not for passive components. Unfortunately, with real-life components, almost all parameters are frequency dependent.

The latest simulators still assume ideal passive components in their libraries. Some though, including Microcap 6, provide the facility to override the internal constant value model using a frequency dependent expression. Regrettably, as far as I am aware, suitable model libraries are not provided.

A restricted number of component models[9] can be downloaded from Intusoft, and are supplied with the company's simulators. These offer a limited choice so usually do not exactly fit ones needs. The modelling approach used was initiated in 1994, by John Prymak of Kemet.

Kemet[10] now offers a Spice based data sheet for their Ceramic and Tantalum capacitors as a free download. This software provides on-screen plots of capacitor behaviour with frequency, including capacitance, ESR, tanδ, inductance, impedance, series and parallel resonances.

For any one frequency of interest, the simulation circuit used and its component values can be displayed on screen. These can then be used in transient analysis and narrow-band frequency sweeps.

Unfortunately these simulation component values cannot easily be extracted for use in wide-band frequency domain analysis.

The main problem is that this frequency-dependent expression relates to an individual element. A resistor that possesses resistance, capacitance and inductance then requires three elements.

Capacitor models may be considerably more elaborate. It would be convenient if one could download manufacturers macro models for passive components – or better still S parameters – as has long been possible for many ICs.

Measurement equipment

The HP331A meter has an oscilloscope-equivalent input impedance and is specified for use to 3MHz. The HP8405 is a narrow bandwidth sampling vector voltmeter that uses permanently attached 100kΩ/5pF probes. It is specified for use from 1MHz to 1GHz.

To ensure consistent test signal voltage with frequency, the signal generator output was loaded with a 10dB attenuator and a Macom 50Ω through termination. The HP331A fitted with a divide-by-10 probe and the HP8405A were used to monitor test signals at this termination.

These instruments were then used to measure output from the differential amplifier. To compensate for the HP8405 probe's 5pF capacitance, additional capacitance was added to the differential probe output connector to produce an 18pF oscilloscope load. The 0.8% resistive load change, from the HP331A's 1MΩ to the HP8405A 100kΩ, when in parallel with the 910Ω output resistor R_{10}, was ignored.

My high-impedance RMS meter could also have been used.[8] However, the HP331A and HP8405A both provide direct readings in decibels relative to 0dBm. Using these instruments avoided a considerable number of calculations.

In the table, results for probe 2 are not listed. As you can see from the CMR results, they must be indistinguishable from those for probe 1. With differential measurements, CMR at higher frequencies is most important. Thus frequency performance was optimised for CMR and noise level, rather than for flattest response.

Using the HP331A voltmeter and with probe relays set to divide by 100 or divide by 10, noise level measured 0.07mV. Set to unity gain noise level measured 0.62mV.

With the relays set to divide by 10 or 100, less than 1mV peak noise is seen using a 100MHz oscilloscope. With gain set to unity, noise level is less than 5mV peak.

Because the maximum permitted input to the HP8405A vector voltmeter is 1V, no divide-by-100 CMR results could be sensibly measured at higher frequencies. But at 1kHz with 30V input, a divide by 100 differential measurement gave –109.5dB CMR.

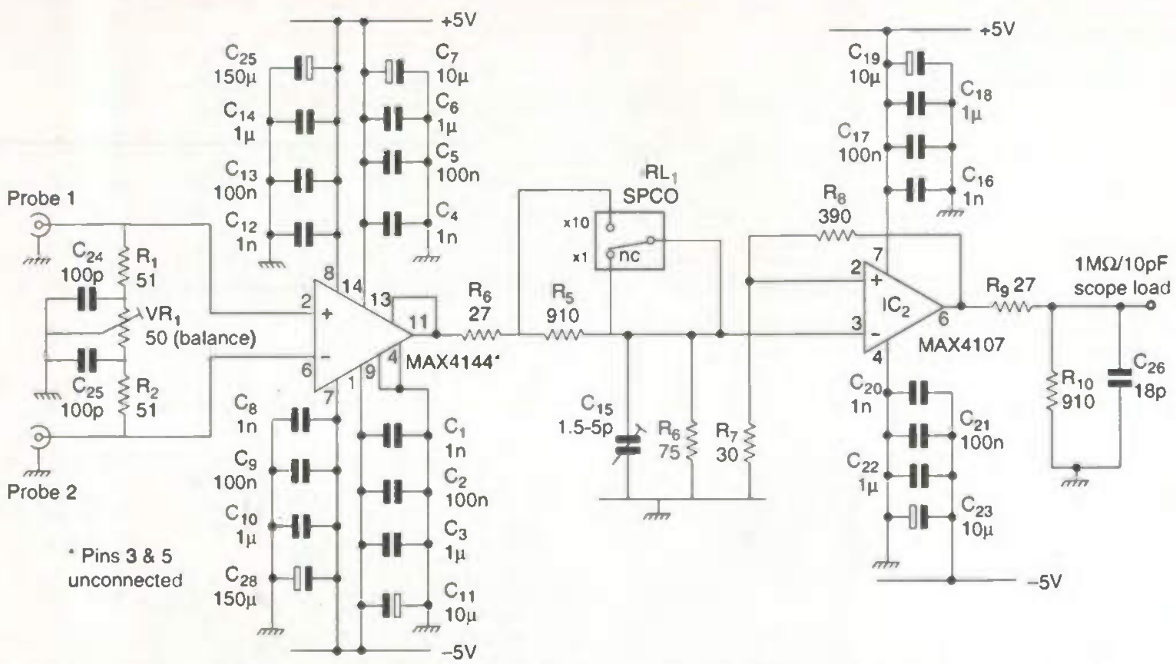


Fig. 9. Final schematic of the differential input amplifier as used for performance tests. Trimmer C_{15} is a slow adjustment that only affects high-frequency single probe gain linearity. This can be replaced by a fixed 2.2pF capacitor with little effect on differential measurements, should a suitable 30MHz test and measurement system not be available. Note that the two three-terminal regulators are not shown.

A revised PCB was assembled and used for final tests. This revised design is shown in the photographs and figures. With this design and using the ground plane through links as shown, copper shim soldered around the PCB edges was not needed, **Fig. 8**.

Interconnected to the finished probes using one metre of RG179B/U coaxial cable, the complete assembly worked extremely well. Common-mode rejection and frequency response were both excellent.

With no signal input and the relays set to unity gain, noise measured less than 1mV on a 3MHz bandwidth voltmeter. Noise output viewed on a 100MHz oscilloscope however was much higher than desired.

Following further measurement and simulations, noise output was found to peak near 100MHz. At frequencies below 60MHz, noise output was acceptably low.

Reducing noise

Careful measurements using my HP4805 vector voltmeter indicated an increasing voltage input to the differential amplifier. This rise was not found measuring the probe outputs terminated in 75Ω, or the differential amplifier with 75Ω input.

Investigations indicated this rise was probably due to inductance contributed by the trimmer resistor VR_1. To replicate this peak, my simulations needed a much higher self inductance than I originally estimated for this component.

Two small 100pF, 1% capacitors, C_{24} and C_{25}, were soldered between the VR_1 terminals and ground. They were 0805-sized COG types. At 100MHz these provided a 0.5dB reduction of signal input to the MAX4144 differential amplifier without compromising the CMR performance.

To further reduce high-frequency noise, I decided to roll off some output gain above 70MHz. This was achieved by adding an 18pF COG capacitor, C_{26}, in parallel with R_{14}. The differential amplifier can now be used with oscilloscope input capacitances ranging from 15pF to 22pF without compensating adjustments for changing load, Fig. 9.

These changes reduced the output noise viewed on my 100MHz oscilloscope. It is now less than 1mV peak with relays switched to divide by 10 or 100. Less than 5mV peak when switched for unity gain.

All the above changes were included for the results shown in Table 1, and are also shown in Figs 6 to 9. Note that these capacitors were added after I took the photographs.

Final performance results

I said earlier that this project required a considerable design effort and many repeated simulations caused by the lack of high-frequency data for components. However as you can see from the table, this very low cost, easily built design, provides excellent CMR and high frequency performance.

While obtaining these performance figures required a capability to measure small changes in low-level signals over a wide frequency range, much simpler facilities suffice to calibrate the three trimmer capacitors and the resistor.

To calibrate for accurate measurements to 10MHz, all that is needed are a signal source able to supply from 200mV to 1V at 1kHz and 1MHz when terminated in 50Ω, and the ability to accurately measure these voltages at both frequencies.

For higher-frequency use, ideally a source and measurement at 30MHz is also preferred to accurately adjust C_{15} in the differential amplifier. This is a 'slow' adjustment that affects single probe ÷10 and ÷100, high-frequency response.

However it has only a small effect on the accuracy of high-frequency differential measurements. For differential measurements, C_{15} could simply be replaced by a fixed 2.2pF COG ceramic capacitor, shown as C_{15A} on the PCB.

Clearly, C_{15} has a noticeable effect on the accuracy of single probe measurements above 50MHz, using an RF voltmeter. The amplitude accuracy of many – perhaps most – oscilloscopes at these frequencies will already have badly deteriorated. The effects of using a fixed value of 2.2pF C_{15} will then not be noticed. ■

Final assembly

Assembling the differential amplifier PCB and its housing requires little comment. With the exception of the two three-leg 100mA stabilisers, pre-set resistor VR_1, trimmer capacitor and relay, components used are designed for conventional surface mounting.

For pre-set resistor VR_1, I used a small single-turn Bourns trimmer, Farnell part 345-994. Its legs were bent to surface mount on the PCB pads.

The relay was an Omron CGE134P, through-hole mounted from the underside of the PCB. This relay, Farnell part 176-323, was similarly used for the test probes.

The 1-5pF trimmer capacitor was a Murata COG ceramic with its legs flattened and trimmed to suit the PCB pads. Farnell part number 108-218, **Fig. 10**.

For each probe PCB, the larger 6-50pF trimmer capacitor was a similarly modified Murata COG ceramic, Farnell part 108-222.

The very tiny 3-10pF trimmer was an AVX CTZ2 designed for surface mounting. Farnell part 578-370.

The PCBs were arranged as far as possible to accept either 0805 or 1206 size components. Resistors were all 1% or better and capacitors of 1nF or lower were COG ceramic types. Where possible, other values were X7R material with Z5U used for 1μF. The largest capacitors were surface mounting AVX type TAJ tantalum chips.

The most difficult part of final assembly was to find a suitable housing for the probes. I searched many catalogues with no success. However, a traveller's toothbrush case from the local supermarket proved to be an appropriate size and shape, Fig. 10.

To effect screening, I attached some copper foil to some thin Mylar insulation using two-sided adhesive tape. This laminate was cut and formed to fit tightly inside one half of the toothbrush case, and grounded to the PCB using a short flying lead.

The test prods for each probe were assembled from a short length of 4mm inside-diameter thin-walled brass modelling tube. The centre contact was a 0.7mm outside-diameter sewing needle with the point blunted. Two PTFE insulators, intended as PTFE stand offs, provided a tight fit into the brass tubing[6]. These test prods on their own measured just 1.5pF.

Fig. 10. Composite view of the prototype differential probes as tested. Both probes are housed in one half of a plastic traveller's toothbrush case. This has a wrap of copper foil inside to act as a Faraday shield. For ease of assembly the differential amplifier was housed in a 110 by 60 by 31 mm die-cast box. This PCB could easily fit in a much smaller box.

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10. Kspice185_32.exe. J. Prymak. http://www.kemet.com

PCB availability...

This design completes a series of four simple, self assembly, double sided PCB designs, targeted to assist high frequency measurements. While one-off single sided PCBs are easy to reproduce, one-off double sided boards are more difficult. If sufficient readers desire boards, I can arrange for the PCBs for this series to be professionally produced.

If you are interested, send a 220 by 110mm SAE to C. Bateman at Electronics World, Cumulus Business Media, Anne Boleyn House, 9-13 Ewell Road, Cheam, Surrey SM3 8BZ.

High-resolution PC voltmeter

Yongping Xia's PC add-on allows an 18-bit analogue-to-digital converter to be read via a printer port.

Maxim's MAX132 is an 18-bit plus sign, serial-output analogue-to-digital converter. Its multi-slope integration conversion method provides high resolution as well as speed.

The chip is designed to interface with many common microcontrollers through three inputs and a data output. One input selects the chip, one clocks the data and the third line carries the data.

With a few external components and ±5V power supply, the MAX132 can provide a high resolution a-to-d conversion with a –512mV to +512 mV measurement range.

A PC's printer port can be used for general-purpose input/output. With the circuitry shown, together with the C program, a PC can interface with MAX132 directly through its printer port.

In this design, outputs pin 4 to pin 6 are used to generate CS, D_{IN} and SCLK signals respectively. Serial a-to-d conversion data is read into PC through pin 13. Pin 18-25 are grounded.

As the MAX132 is 'bipolar', it needs both +5V and –5V power supplies. In order to simplify the design, –5V is generated by a charge-pump DC-to-DC converter built around a TC7660.

A reference voltage for the MAX132 is provided by a diode D_1 and dividers R_6, R_7 and R_8. The input range of the a-to-d converter is from –512mV to +512mV. The measured data shows on a PC's screen in two forms. One is the number of millivolts. The other gives an 'analogue' indication in the form of a horizontal bar whose length depends on the magnitude of the input.

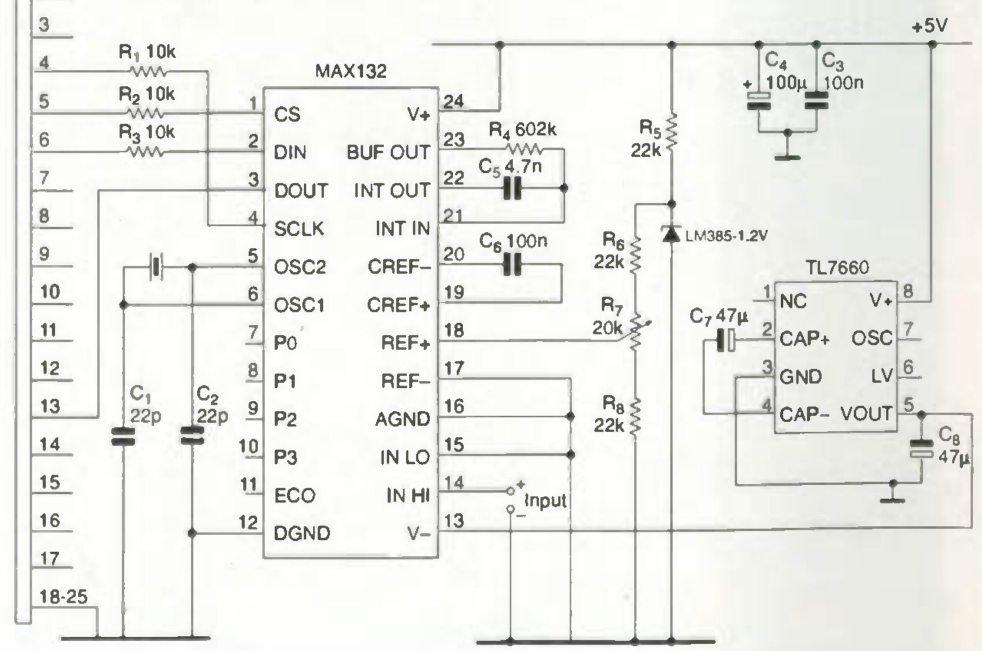
When input is shorted, the output of MAX132 should be zero. However, the circuit may not be perfect and the reading is likely to have some value.

Under software control, the MAX132 can short its input internally. This allows the zero value to be read without physically shorting the input. Once determined, the zero value can be subtracted from the reading later to compensate the offset error.

To reduce the error further, the zero error value and normal a-to-d conversion results are averaged by eight readings. ■

A-to-d converter input is displayed directly on screen in millivolts. Length of the horizontal bar depends on the input, giving an easy-to-assimilate analogue readout.

Printer port plug-in provides 18-bit a-to-d conversion for displaying voltages on screen.



Annotated C listing for reading the 18-bit analogue-to-digital converter interface through a PC's parallel LPT port. For an electronic copy of this listing, send a request via electronics.world@ntlworld.com with the subject heading '18-bit converter, December 2001'. Note that using this address for any other correspondence will result in disappointment!

```
#include <graphics.h>
#include <stdlib.h>
#include <stdio.h>
#include <conio.h>
#include <dos.h>
#include <bios.h>

#define CLOCK_LOW      0xfb
#define CLOCK_HIGH     0x04
#define DIN_LOW        0xef
#define DIN_HIGH       0x10
#define CS_LOW         0xf7
#define CS_HIGH        0x08
#define POWER_ON       0x01
#define POWER_OFF      0xfe

typedef unsigned int WORD;

int          out_port, in_port, out, power_boost_status=0;
char         msg[80];

float read_data(int measure)  /* read data from MAX132 */
{
    int i, j, test_bit, data[3], command[4];
    float data_dis;
    data[0]=0;
    data[1]=0;
    data[2]=0;
    if (measure==0)
             command[0]=0x92; /* set to read zero offset */
    else
             command[0]=0x82; /* set to read input voltage */
    command[1]=0x04;
    command[2]=0x00;
    command[3]=0x00;
    out|=CS_HIGH;
    outportb(out_port, out);
    for (j=0; j<4; j++)
             {
                      out&=CS_LOW             /* CS low*/
                      outportb(out_port, out);
                      delay(1);
                      test_bit=0x80;
                      for (i=0; i<8; i++)     /* start convert        */
    {
             if ((command[j]&test_bit)==0)
                      out&=DIN_LOW;
             else
                      out|=DIN_HIGH;
             outportb(out_port, out);
             delay(1);
             out|=CLOCK_HIGH; /* clock high  */
             outportb(out_port, out);
             delay(1);
             if (j!=0)
             data[j-1]=data[j-1]*2+(inportb(in_port)&0x10)/16;
             out&=CLOCK_LOW;
             outportb(out_port, out);        /* clock low   */
             delay(1);
             test_bit/=2;
    }
    out|=CS_HIGH;      /* CS high */
    outportb(out_port, out);
    delay(1);
    if (j==0)
    delay(100);                   /* wait 100ms  */
    }
data_dis=((float)data[2])/64+((float)data[1])*4+((float)(data[0]&0x07))/512;
if ((data[0]&0x08)==0)
    return (data_dis);
else
    return (data_dis-1024+1/512);
}

void display_data(float data_dis)
{
    int current_x;
    current_x=(int)(data_dis/2);
    setviewport(10, 200, 630, 220, 1);
    clearviewport();
    setfillstyle(1, YELLOW);
    bar(310, 0, (current_x+310), 10); /* show a bar on screen */
    setviewport(280, 240, 630, 300, 1);
    clearviewport();
    if (data_dis>512 || data_dis<-512)
             sprintf(msg, "Overrange!");
else
             sprintf(msg, "%.2f mV", data_dis);
outtext(msg);            /* show data on screen */
}

void init_screen(void) /* initialize display screen */
{
setbkcolor (BLUE);
setcolor(WHITE);
line(2,2,637,2);
line(5,4,634,4);
line(2,2,2,477);
line(5,4,5,475);
line(5,475,634,475);
line(2,477,637,477);
line(634,5,634,475);
line(637,2,637,477);
line(3,350,634,350);
setviewport(450,440,630,460,1);
sprintf(msg, "press any key to quit");
outtext(msg);
setviewport(0,0,639,479,1);
}

int init_graph(void)   /* initialize graphic mode */
{
    int gdriver = DETECT, gmode, errorcode;
    initgraph(&gdriver, &gmode, "");
errorcode = graphresult();
if (errorcode != grOk)         /* an error occurred */
{
    printf("Graphics error: %s\n", grapherrormsg(errorcode));
    printf("Press any key to halt:");
    getch();
    exit(1);           /* return with error code     */)
}

void find_port(void)   /* find printer port address */
{
out=0;
out_port=*(WORD far *)MK_FP(0x0040, 8);
in_port=out_port+1;
out|=POWER_ON;
outportb(out_port, out);
delay(100);
}

void clean_up(void)    /* close graphic mode */
{
closegraph();
out&=POWER_OFF;
outportb(out_port, out);       /* turn off power supply */
}

void main(void)
{
int channel, measure=0, k;
float tp, measured_data=0, zero_offset;
find_port();
init_graph();
init_screen();
delay(1000);
read_data(measure);
read_data(measure);
read_data(measure);
for (k=0; k<8; k++)
    measured_data+=read_data(measure);
zero_offset=measured_data/8;
measure=1;
do{
    measured_data=0;
    for (k=0; k<8; k++)
    {
    tp=read_data(measure);
    measured_data+=(tp-zero_offset);
    }
    display_data(measured_data/8);
    } while(!kbhit());             /* quit if hit any key */
clean_up();
}
```

More on the MAX123...

The MAX132 is a CMOS, 18-bit plus sign, serial-output, analogue-to-digital converter. Multi-slope integra-tion provides high-resolution conversions in less time than standard integrating a-to-d converters, allowing operation up to 100 conversions per second.

Low conversion noise provides guaranteed operation with ±512mV full-scale input range (2μV/LSB). A simple four-wire serial interface connects easily to all common microprocessors, and two's-complement output coding simplifies bipolar measurements.

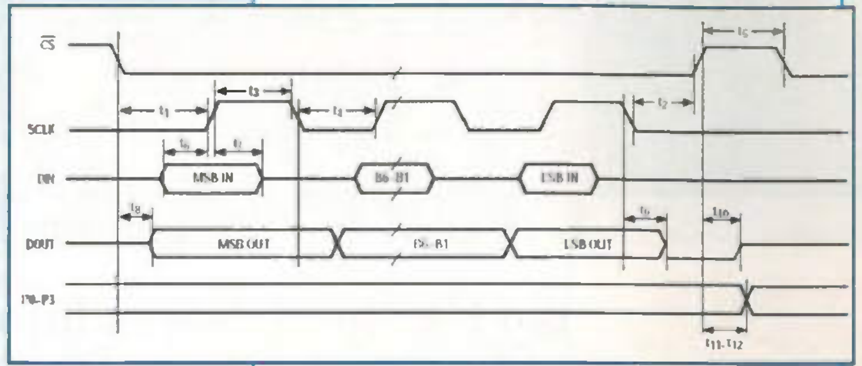
Typical supply current is 60μA, reducing to 1μA in sleep mode. Four serially-programmed digital outputs can be used to control an external multiplexer or programmable-gain amplifier.

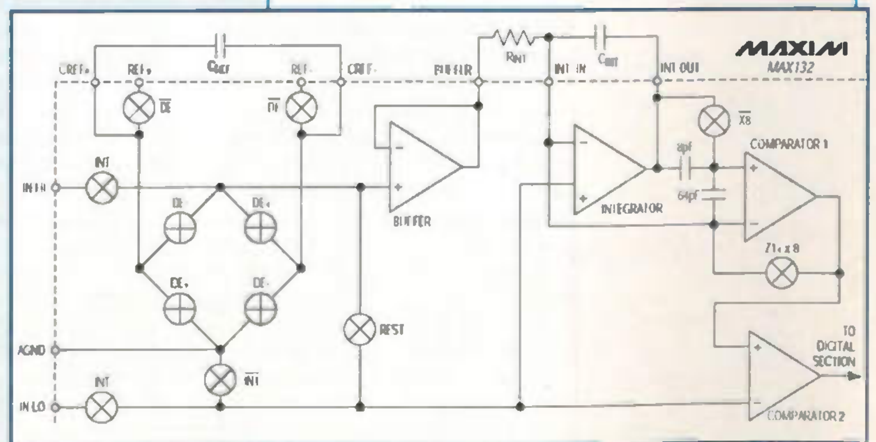
The MAX132 comes in 24-pin narrow DIP and wide SO packages. It's absolute maximum analogue input range is from the positive supply to the negative supply.

Feature summary

Supply Current: 60μA, normal operation
1μA, sleep-mode
±0.006% FSR Accuracy at 16 conv/s
15μV RMS noise
Serial I/O interface
Programmed output for MPX and PGA
Performs up to 100 conv/s
±2pA input current
50Hz/60Hz rejection

www.maxim-ic.com





Functional block diagram of the MAX123, above, and its serial timing flow, top.

Understanding Transformers

Ian Hickman delves into the inner workings of transformers.

In the August issue, I described a rudimentary mains transformer with a 1:1 ratio. The simplified equivalent diagram in Fig. 2 of that article shows the magnetising inductance in parallel with a perfect transformer.

On load, the efficiency will fall short of 100%, due to losses in the winding resistances. These have been bundled up together as an equivalent total winding resistance R_p in series with the primary winding. The loss in R_p will be $I_{pt}^2 \times R_p$ where the total primary current I_{pt} is the vector sum of the magnetising current and the current in the primary required to support the demanded secondary load current.

This all describes what happens when the transformer is supplying the load *in the steady state*. At switch-on though, things may be very different.

A transformer at switch on

Imagine the supply switch closes when the mains voltage is at its positive peak. Obviously, at the moment of contact, the current is zero, the same as a microsecond earlier. But a given voltage applied to an inductor causes the current through it, and the resultant flux, to increase, as happens here. And as the applied voltage falls from its peak to zero, the magnetising current and the flux increase from zero to their usual maximum values, lagging the primary voltage by a quarter of a cycle as shown in Fig. 4 and Fig. 5 of the earlier article. This is the nice scenario.

Now imagine that the switch closes when the mains voltage is zero, and positive going. When the voltage reaches the positive peak, the current and flux will have increased from zero to their normal maximum values. But the voltage won't be zero until a quarter of a cycle later, so to provide the back emf balancing the applied voltage, the flux will have to go on increasing, up to twice its normal maximum.

Unless the transformer has been very conservatively designed with a much larger core than usual, the core will run into saturation, to some degree. As the effective permeability drops, so the magnetising current will have to increase by leaps and bounds to provide the flux and hence back emf, to balance the applied voltage. This is the not so nice scenario, and explains why the fuse used with a transformer is often specified as of the 'slow-blow' variety.

If the relative permeability of the core dropped right down to unity – complete and utter saturation – the current would rise dramatically, being limited only by the primary winding resistance and the effective inductance of the primary without a core.

While in practice things are never quite that bad, if the switch closes at zero mains volts, the current drawn on the first half cycle will be much greater than usual. This will result in a significant volt drop across R_p, alleviating the peak applied voltage somewhat.

On the following negative half cycle, however, the drop across R_p will be much smaller. Thus the 'volt-second product' applied to the winding will be greater, and at the end of the first complete cycle the flux will have returned to, and passed through zero, reaching some negative value. The same sort of thing happens during the second cycle, and so over succeeding cycles, the positive flux peaks will decrease and the negative increase, until in the steady state they are equal.

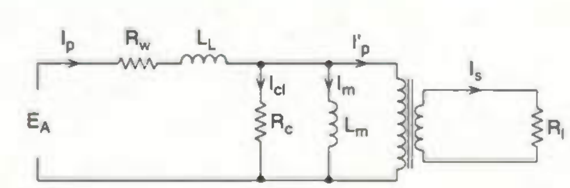


Fig. 1. In a transformer model, 'core loss' can be represented by a resistance R_c in parallel with the magnetising inductance L_m.

Volt-seconds

The volt-second product is a key concept when dealing with inductors of any sort; it is the thing that determines the change of flux between zeros of the applied voltage. A transformer designed for mains operation at 240V AC will see a peak voltage of 240×√2=339V. The volt-second product, with 50Hz mains, will be less than 339V×10ms by a factor 2÷π=0.637, giving a figure of 2.35Vs.

Some inverters, providing a 'mains' output from 12 or 24V batteries, produce what amounts to a lightly filtered squarewave. Exceeding the volt-second product of any transformers that may be supplied by the inverter would cause them to heat up and possibly fail. So such an inverter's peak voltage is limited to not much more than 235V, giving the required 2.35Vs product.

For some applications this is acceptable, but in others, it causes problems. For example, a piece of test equipment powered from supplies derived from its mains transformer may malfunction, due to the raw supply voltages to the stabilisers being inadequate.

The simplified treatment in the earlier article ignored the 'core loss', which in practice must be considered. It can be represented by a resistor R_c in parallel with the magnetising inductance L_m, as in Fig. 1.

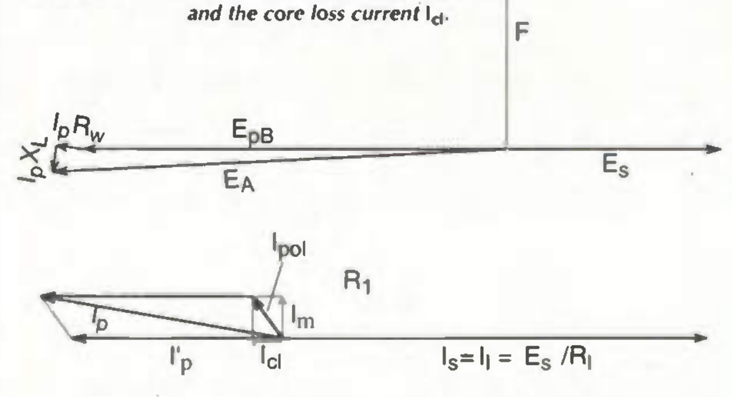
On full load, the core loss will actually reduce somewhat. This is because the voltage across it falls slightly due to the additional drop across the winding resistance R_w and the leakage inductance L_L.

Resistance R_c is a fiction, representing the heating in the core at rated mains voltage in the steady state. It is not a real component and as such has no absolute value; if the applied voltage is reduced to 90%, the core loss may differ from the expected 81% of the value at rated voltage. And of course R_c does not represent the state of affairs in the case of switch-on at a zero of the mains voltage.

'Magnetising current'

It is common parlance to talk loosely about the 'magnetising current' when what is really meant is the primary off-load current. The magnetising current is in phase with the flux Φ and 'mmfF' in Fig. 2, and in quadrature to the voltage across the primary of the ideal transformer, whereas the off-load current includes the in-phase core loss component.

Fig. 2. In a transformer with a 2:1 step down ratio, primary off-load current I_{pol} consists of the vector sum of the magnetising current I_m and the core loss current I_{cl}.



The figure describes a transformer with a 2:1 step down ratio, showing how the primary off-load current I_{pol} consists of the vector sum of the magnetising current I_m and the core loss current I_{cl}.

Note that the 2:1 ratio transformer in Fig. 1 supplies a full load secondary voltage which is slightly less than half the applied mains voltage E_A. If you want a 120V full load output from 240V mains, you don't want a 2:1 ratio transformer.

Transformer specifications

When designing a mains transformer, the requirements will specify the supply voltage, the desired secondary voltage(s), and the secondary current(s). The regulation may also be specified, especially if the transformer will experience a varying load.

Often, a transformer will be designed for maximum efficiency at full load, but not necessarily. For example, a local area mains distribution transformer will only experience full load, or something near it, at the peak demand hour. For much of the rest of the time, e.g. at night, it may be running almost off-load.

Clearly, the efficiency of any transformer is zero when off load, but the designer has the option to trade off full-load loss, primarily copper loss, against core loss, which is there all the time. So the said mains distribution transformer will probably be designed for very low core loss, at the expense of a little more copper loss on full load. On the other hand, a transformer designed for continuous operation at full load is generally designed so that the copper loss equals the 'iron loss'.

Current transformer

In fact, the first transformer I ever had to design was a current transformer. Here the requirements are very different, demanding a complete realignment of one's thoughts. As against primary voltage, secondary voltage and full load secondary current, for a current transformer the data are primary current, secondary current and full-load secondary voltage. One is in the topsy-turvy constant-current world.

Suppose, for example, it is desired

Vacuum leakage detector

Just occasionally, it is desirable to increase leakage reactance, rather than reduce it. The application I have in mind is in equipment for testing laboratory glass vacuum systems for leakage. I used an Edwards leakage tester when working on basic semiconductor research in the mid 1950s. I subsequently acquired a surplus leakage tester made by Ferranti, and the circuit diagram is shown in Fig. 3.

The mains transformer is wound on a three-limbed core using ordinary E and I laminations. However, the primary is wound on the centre limb, and the secondary on one of the outer limbs, the other being unused.

The secondary is connected across an adjustable spark gap. When this is open, no secondary current flows, the flux in the two outer limbs is equal and the secondary voltage large – I'm not sure what the turns ratio is.

When the gap is short circuited, no flux can pass through that limb, and the magnetising current merely doubles; the design is such that the open 'unused' limb does not saturate. When suitably adjusted, a spark is maintained at the gap, and the room rapidly fills with the smell of ozone.

Capacitor C_1 picks off the high-frequency energy from the spark. It actually consists of several high-voltage mica capacitors in series, with an effective capacity at a guess, of one or two hundred picofarads.

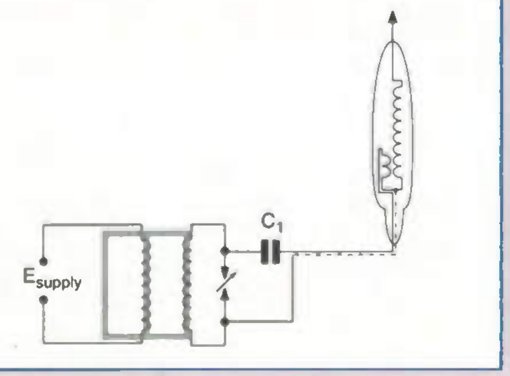
A screened lead conveys the high-frequency components to the primary of a small Tesla coil, fitted in the body of the hand-held business end. My sketch is not to scale; the lower part of the probe, the handle, is as long as the part containing the Tesla coil. In operation, whiskers of electrical discharge emanate from the tip.

A spark about an inch long can be drawn from the tip to the nearest object – test your CMOS here! The nearest object might be a finger. The high source impedance limits the current to a perfectly safe low value.

In use, the tip is waved all over the surface of a glass vacuum system, looking for pinhole leaks. Any such is betrayed by the much lower ionisation potential of any residual air, resulting in long purple streamers inside the glass piping, spreading out from the leak.

A grain of vacuum wax over the site of the leak, a touch with a warm under-run soldering iron and hey presto.

Fig. 3. Simplified diagram of the Ferranti leakage tester – a device capable of producing an inch-long spark and used for finding leaks in glass vacuum apparatus. The mains transformer is wound on a three-limbed core using ordinary E and I laminations but the primary is wound on the centre limb, and the secondary on one of the outer limbs, the other being unused.



to monitor a circuit that might carry up to 50A, with a 5mA full scale deflection meter. The primary would probably be just one turn, with a 10 000-turn secondary. If the meter circuit requires, say, 3V for fsd, that plus the number of secondary turns will determine the maximum flux required on the core.

As with any transformer, the primary current I'_p will balance the secondary current I_s in accordance with the turns ratio. There will also have to be some quadrature magnetising current I_m in the one-turn primary. This, plus I'_p, equals the 50A primary current.

Given the small volts per turn figure for the secondary, the flux required will turn out to be quite low, so the core size may be determined by the window area required to accommodate the 10 000-turn secondary. A 5000-turn secondary could be used, if the primary had but half a turn. This was achieved on the 10A range in the legendary AVO meter, by omitting the centre E from a stack of E and I laminations, and threading the tape shunt conductor through the gaps.

The 3V drop due to the meter circuit in the secondary will reflect in the primary as a 0.3mV drop. Note that the exact secondary voltage required is not critical; the magnetising current I_m, minuscule compared to 50A, will adjust itself to provide whatever voltage drop appears across the meter, at 5mA.

In fact, the transformer acts as a constant current generator, just as a conventional mains transformer acts as a constant voltage generator. And whereas for a conventional transformer the safe condition is secondary open circuit, for a current transformer, the safe condition is secondary short circuit. An open-circuit secondary is as potentially disastrous for a current transformer as is a secondary short circuit for a conventional transformer.

In the design of small mains transformers, leakage inductance and winding capacity do not present any real problems. But both are of prime importance in the design of wide-band transformers, where all possible steps are needed to minimise them.

Leakage inductance can be reduced by splitting primary and secondary into sections and interleaving them. Dividing into n sections will reduce the leakage inductance by a factor of n^2. This technique is used in the better class of output transformer for valve hi-fi power amplifiers, providing a ten octave bandwidth. Using interleaving, and the remarkable properties of thick mu-metal laminations, it is possible to produce a small-signal transformer covering 50Hz to 2MHz. ■

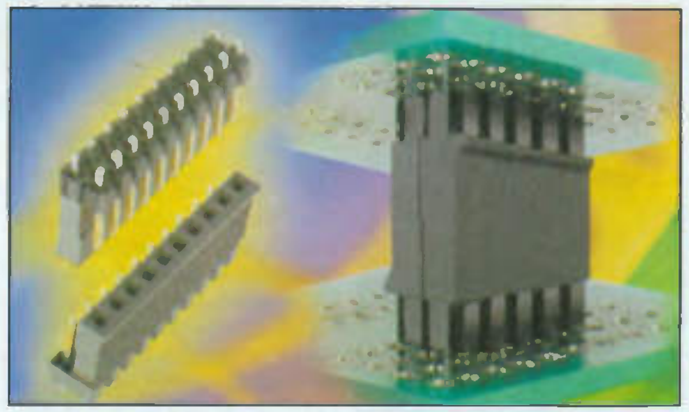
NEWPRODUCTS

Please quote *Electronics World* when seeking further information

Software for data logging

National Instruments has announced a software tool specifically designed for data logging. Engineers and scientists can acquire, log, view, and share data with this stand-alone, configuration-based software. Because it uses dialogue windows instead of traditional programming, almost anyone can create data logging applications. Engineers and scientists can use the logger for data logging applications, regardless of the number of channels the application requires. They can create a single-channel, temperature-logging application for product condition analysis using a variety of measurement hardware.

National Instruments
Tel: 01635 523 545
www.ni.com/uk



Combination socket/terminal strip

Samtec is offering as combination socket/terminal strip on 0.1in. (2.54mm) pitch. This is a self-mating system which is intended to eliminate the need to stock both a terminal and a socket in inventory. The locking feature on this connector increases the unmating force required for use in high-vibration applications and provides an audible 'click' upon proper mating. The inherent polarisation of this robust interconnect prevents incorrect mating. The LST series is a double row connector and is available with from 4 through 60 pins with a choice of common platings. This combination connector also features the firm's Tiger Buy contact designed for high retention. Currently, through-hole vertical applications are released. Surface mount, right-angle and other pitches are under development.

Samtec
Tel: 01236 739292
www.samtec.com

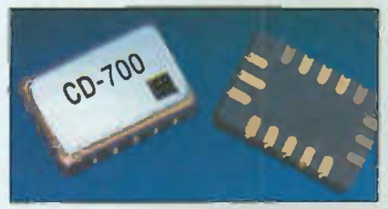
High-side optimised n-channel Mosfet

The latest n-channel Mosfets from Fairchild Semiconductor have been designed to achieve a reduction in die size. The FD56694, SO-8 packaged, 30V fast-switching trench Mosfet is designed for high-side applications in synchronous and conventional DC-to-DC converters for notebook computers, desktop computers, and other DC-to-DC power supply applications.

Fairchild Semiconductor
Tel: 01793 856856
www.fairchildsemi.com

Phase-locked loop on a chip

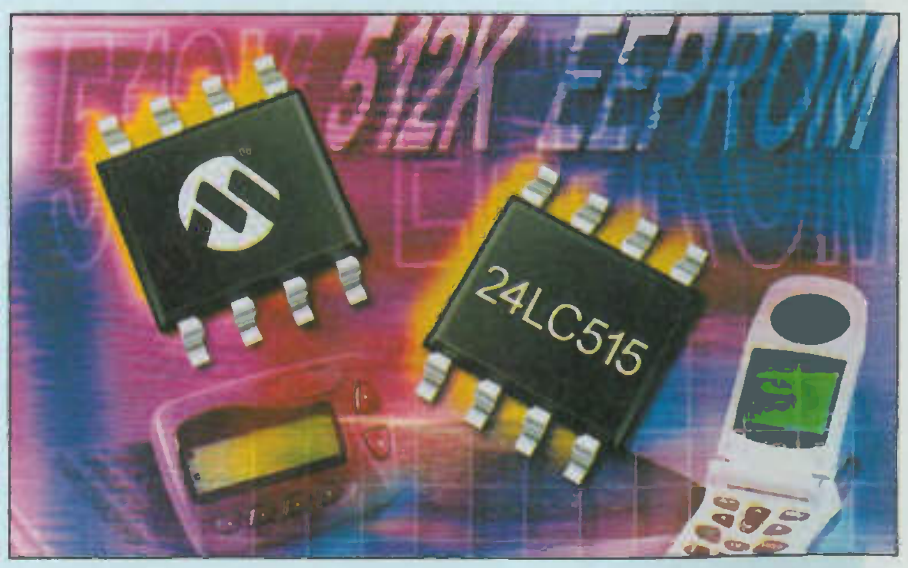
Vectron International has introduced a VCXO-based phase-locked loop (PLL) designed for clock recovery and data retiming, frequency translation, clock smoothing and clock switching applications. The CD-700 features a phase-locked loop Asic with a quartz stabilised VCXO for superior stability and jitter performance. Input data rates range from 8kbit/s to 65Mbit/s and the output has a tri-state option. In



Twin-die flash microcontroller

Microchip has announced its first 512kbit I²C serial EEPROM in a standard 8-lead SOIC package. With typical applications likely to include mobile phones, set-top boxes, pagers and data acquisition systems, the 24LC515 serial EEPROM has a page-write capability of up to 64 bytes of data and is capable of random reads up to the 512kbit boundary and sequential reads within each of its 256kbit blocks. Functional address lines allow up to four devices on the same bus, for up to 2Mbit total address space. Additional features include a clock rate of 400kHz, an operating voltage of 1.8V to 5.5V, and a temperature range of –40 to +85°C. A faster version of this chip (24FC515) is capable of operating at a bus speed of 1MHz over the same temperature range.

Microchip
Tel: 0118 921 5858
www.samtec.com



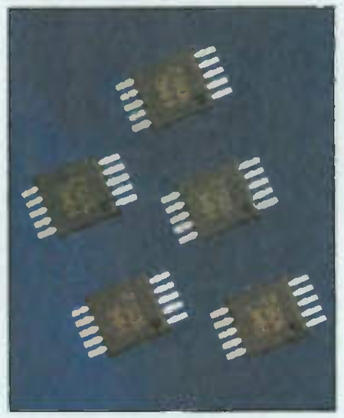
addition, the supply voltage can be either 3.3V or 5V. The device is hermetically sealed in a ceramic surface mount package measuring 5.0 by 7.5 by 2.0mm. Target applications include DWDM, Switching, Wireless Basestation, ATM, SONET/SDHa and xDSL.
Vectron International
Tel: 02380 766288
www.vectron.com

Audio design kit for games developers

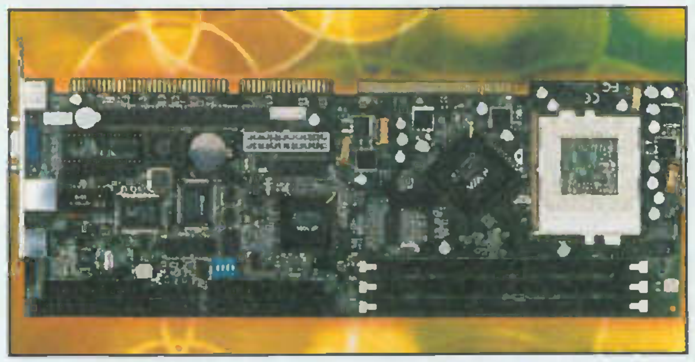
Analog Devices has extended its SoundMAX integrated PC audio design kit to software development kits for Windows PC and Sony PlayStation2 game titles. The toolkit enables game developers to add rendered sound effects by employing non-repetitive, interactive audio behaviours in place of the static sound effects typically found in games. According to the company, designers can increase the sophistication of their games' audio to that achieved in video. The toolkit provides complete API documentation, software libraries, sample code, and online help. In addition, the PC version features Mission Control III, a graphical user interface that allows audio designers to incorporate the firm's SPX (Sound Production eXtensions) game audio rendering technologies. Sounds such as footsteps, explosions, car engines and ambient sounds can be realistically produced in real-time, in direct response to the user's actions.
Analog Devices
Tel: 001 781 937 1622
www.staccatosys.com

Two-axis magnetic position sensor

Honeywell Solid State Electronics has developed a two-axis sensor on one chip for compassing and position sensing applications such as handheld wireless systems and GPS receivers. According to the



supplier, advantages of this patented design include nearly perfect orthogonal two-axis sensing in a 3mm by 3mm by 1mm, 10-pin miniature surface-mount package (MSOP). The HMC1052 has a sensitivity of 1mV/V/gauss, a field range up to ±6 gauss and can operate on a supply as low as 1.8V.
Honeywell
Tel: 0118 906 2600
www.honeywell.com

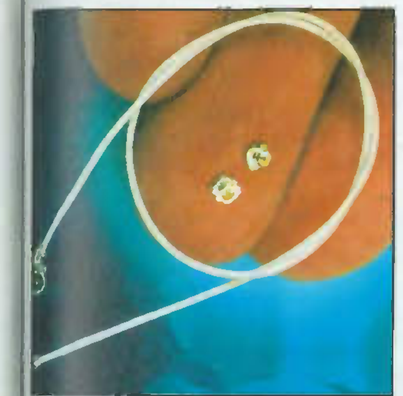


Socket 370 computer board for £240

Amplicon Liveline is offering its lowest-cost socket 370 single board computer based on the VIA chip set, which is priced at £240. Intended for applications in the areas of process control and manufacturing, the M5C3680 board can accept both Intel Pentium III and Celeron socket 370 processors up to 1GHz. On-board VGA and LAN are included to support CRT displays and Ethernet. According to the company, it can be used as a standalone computer, as the ATX connector obviates the need for a backplane. It also provides support for standard industrial PC configurations. Three 168-pin DIMM sockets with up to 768MByte SDRAM and ECC support, a bus mastered IDE with Ultra DME 33/66 controllers and a full array of ports are also included. For the more complex timing requirements in critical applications, the device has a 256-level watchdog timer. The onboard VGA controller supports CRT displays with up to 1600 by 1200 pixels, and hardware monitoring for CPU voltage, temperature and fan speed. System temperature and fan speed can also be monitored. Additional features include modem ring-on, wake on LAN, Windows 95/98 shut off and PC99 ACPI power management, all for the more demanding PC based industrial applications.
Amplicon
Tel: 01273 608331
www.amplicon.co.uk

Surface-mount coax connector is lightweight

The latest surface-mount coaxial connector from Flint Distribution is designed for 2.5mm high PCB-mounted connections, and according to the supplier is amongst the lightest available. The U.FL series connectors, which are suitable for signals up to 3GHz, feature a receptacle of mass 15.7mg with a choice of plug assemblies for applications using 0.8mm single-layer shielded or 1.32mm double layer shielded cables. The 0.8mm and 1.32mm plug and centre contact assemblies have a mass of 53.7mg and 59.1mg, respectively.
The connectors, which are manufactured by Hirose, are designed to produce a positive locking sensation despite their size. At 2.5mm, the coupling height is 0.7mm less than the



preceding E.FL miniature coaxial series, from which the new family is derived. They provide connections at right angles onto the PCB with an SMT receptacle. Potential uses include portable and mobile telephones, wireless communication devices, antenna connections for Bluetooth and other electronic measuring instruments and GPS receivers. The connectors are rated at 60V rms, with insulation resistance 500MΩ and 20mΩ maximum contact resistance at the centre contact. The outside contact resistance is 10mΩ maximum. The U.FL series achieves VSWR of 1.3 or less, enabling use with signals from DC to 3GHz.
Flint Distribution
Tel: 01530 510333
www.flint.co.uk

Low power radio transceiver

Available from Low Power Radio Solutions, the BiM 2 transceiver is a complete reworking of the first generation manufactured by Radiometrix. Operating in the harmonised pan-European SRD band at 433.92MHz, and fully compliant with both EN 300 220-1 and ETS 300 683, the BiM 2 integrates a SAW-controlled FM transmitter with a double-conversion superhet receiver to provide bidirectional wireless communication at ranges up to 200m. Radiometrix's original BiM was introduced eight years ago and has been used in applications such as handheld terminals, EPOS equipment, barcode scanners, belt clip printers and data loggers. With a transmitter output of 10mW and a receiver sensitivity of –100dBm, it is capable of data rates up to 64kbit/s. At reduced data throughput, the device has a useable range of at least 200m line-of-sight, and a respectable 50m indoors. The fully screened module shares the same 23 by 33mm footprint and pin-out as its predecessor, and so can be used as a plug-in upgrade in applications currently using the BiM-433-F. Two versions of the BiM 2 are available, for operation from nominal 5 or 3.3V single supplies. Power consumption for the 3.3V version is typically 8mA while transmitting and 14mA while receiving.
Low Power Radio Solutions
Tel: 01993 709 418
www.lprs.co.uk

Pocket-sized 1550nm laser source

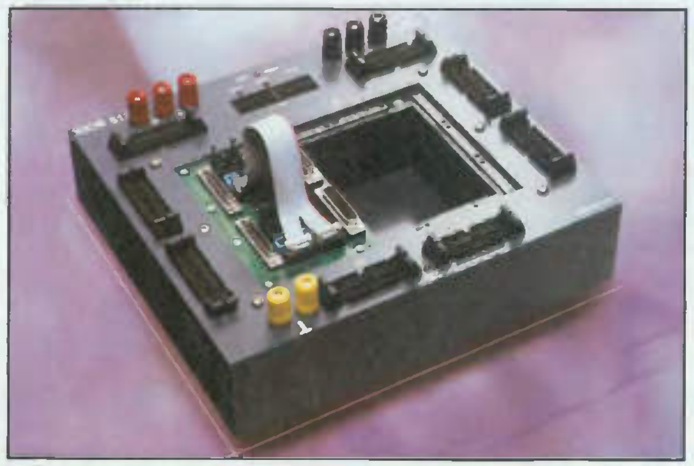
Acterna's OLS-6 light source is available in a 1550/1625nm version. It is a handheld light source with two optical fibre output ports. It is designed for use with an optical power meter, for measurement of loss in fibre networks. It is also claimed to be one of the first light sources of this kind which can be used to make loss measurements at 1625nm. This allows the detection and monitoring of such bending effects for optimisation of network performance, said the supplier. All the instruments in the OLS-6 range feature an internal launch fibre for stable power coupling between the port and the measurement cable.
Acterna
Tel: 001 301 3531560
www.acterna.com

EMI suppression

EPCOS has released the X2 series of compact EMI suppression capacitors. Offering rated AC voltages of 275V and 300V, the capacitance of the X2 series ranges from 10nF with the smallest capacitor, with dimensions of 4 by 9 by 13mm, to 4.7μF with dimensions of 20 by 39.5 by 41.5mm. The capacitors' polypropylene or polyester dielectric is encased in a tough plastic case to a high standard of insulation. They are available with tinned parallel wire leads in two standard lengths of 6mm and 26mm, with other length options available on request. The series has a self-healing system, preventing permanent dielectric breakdown in the event of sporadic voltage surges or overcurrent.
Epcos
Tel: 0990 550500
www.epcos.com

Boundary scan system with more I/O

JTAG Technologies has extended its line of boundary scan tools for testing complex printed circuit boards with the XIOS 512 extended I/O scan system



(XIOS). It provides additional auxiliary test channels used for fault detection on digital circuit boards using boundary scan (IEEE Std 1149.1) techniques. The XIOS 512 can be supplied with 128, 256, 384 or maximally 512 channels and may be upgraded by adding compatible 128 channel DIMM DIOS modules also supplied by JTAG Technologies. Each test channel may be individually programmed as input, output, bi-directional, or tri-state signals. The unit supports both 3.3V and 5V operation. The system is

16-bit flash microcontroller

Hitachi has extended its H8S/22xx 16-bit flash microcontroller series with a stripped-down, reduced cost device. The H8S/2214F is available in a chip-scale package (CSP) for volume production of applications such as portable audio, video and communications devices; bar code readers; portable medical equipment; and handheld measurement equipment. It offers 128kbyte embedded flash memory, 12kbyte RAM and a basic peripheral set to minimise cost in comparison with other members of Hitachi's H8S/22xx series. The device has a power consumption of 20mA (typical) at full speed and is pin compatible with Hitachi's H8S/2238F microcontroller. It offers 62.5ns minimum instruction cycle at 3.3V/16MHz. The 112-ball Chip Scale Package (TBP-112) measures 10mm per side with a ball pitch of 0.8mm. It can be processed using conventional industry soldering procedures and equipment. The H8S/2214F is also available in two Thin Quad Flat Pack (TQFP) packages. The device features a four channel DMA and a pseudo DMA function, which remove data movement tasks from the CPU. This allows the CPU to concentrate on its primary task of calculations. Other peripherals include three serial ports, one of which has a special high speed asynchronous mode, a three channel 16-bit timer module and a watchdog timer. The device has 72 I/O pins, some of which have schmitt trigger characteristics, interrupt capability and programmable pull-ups.
Hitachi
Tel: 01628 585 163
www.hitachi-eu.com

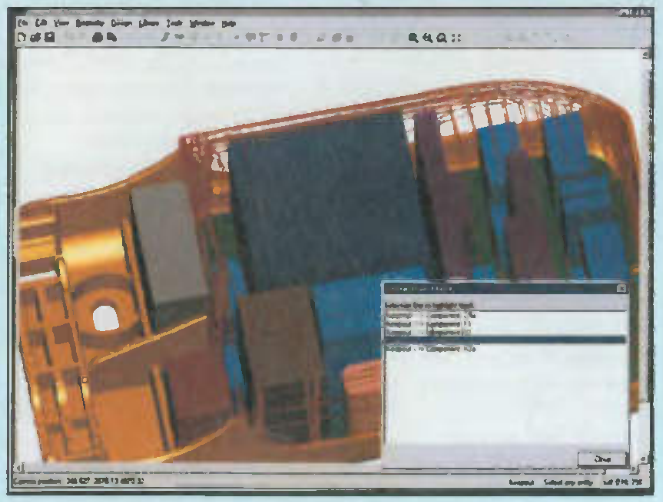
Please quote *Electronics World* when seeking further information

applicable in a variety of production test applications to enhance test coverage. For example, a target circuit board may contain elements that cannot be accessed directly by boundary scan such as board edge connectors or non-boundary-scan logic clusters. In such cases, the overall testability of the board is reduced, allowing some manufacturing faults to be undetected. The extended system tackles this problem, extending the reach of boundary scan to include testing of circuit board edge connectors and providing extra test points internal to the PCB to improve fault coverage. The system is compatible with the firm's DataBlaster family of boundary scan controllers. With a built in version of JTAG Technologies' test pod the XIOS plugs directly into a SCSI style connection from the DataBlaster.
JIAG Technologies
Tel: 01234 272226
www.jiag.com

Varistors have wide operating voltages

Retronix Europe has developed a range of zinc-oxide varistors with operating voltages from 12V to 1800V. The varistors feature a fast response to transient overvoltage and a large energy absorption capability.

They have a low clamping ratio and no following-on current, said the company. Applications will include transistors, diode, IC, thyristor and triac semiconductor protection. Other uses include surge protection in electronic products, and providing electrostatic discharge and noise spike suppression. Relay and electromagnetic valve surge absorption is another possible application area. In summary, applications cover any sensitive electronic system or any electronic product requiring noise protection.
Retronix Europe
Tel: 01635 874123
www.cabian.co.uk



Automatic 3D checking

Zuken's EM Checker is a design tool for Visula that provides automatic verification of a PCB design in its 3D environment. According to the company, this can remove the time consuming iterations between PCB and mechanical design groups, as enclosure and obstacle information can be loaded and modelled in the ECAD environment using the ISO-standard STEP AP203 format. Visualisation, measurement and design rule checking allow the electrical design to be checked against system requirements and any component placement changes made within EM Checker are automatically back-annotated into the PCB design. This ensures that mechanical design issues can be resolved early in the design cycle. Full 3D models provide true 3D analysis of the PCB, enclosures, substrates and other obstacles. Verification in 3D increases confidence in the design and avoids the wasted space created when only 2D design approximations are used, delivering the highest functionality products in the smallest volume. EM Checker uses the full PCB design representation from PCB Layout to ensure no loss of data, no changes in the data model and no re-entry of information between electrical and mechanical domains. As well as compatibility with the STEP AP203 format, 3D objects can also be transferred into EM Checker using STL or ACIS descriptions.
Zuken
Tel: 01454 207800
www.zuken.com

White LED driver IC

Toshiba's latest LED driver IC in a miniature, six-pin SOT23 package delivers an output power of 320mW, allowing it to drive up to six white LEDs in series with a minimum of external components, said the company. An automatic driving current temperature derating function that allows full current to be delivered at room temperature further minimises component count by reducing the number of LEDs needed for a given brightness level. Traditionally, said the company, this has not been possible as designers have

had to drive the LEDs with lower current levels to protect them against the effects of temperature increases. The TB62731FU incorporates an internal switching n-channel Mosfet with a typical on-resistance of 1.5Ω contributing to an overall device efficiency of up to 90% (pulse mode) and 85% (DC drive mode).
Toshiba
Tel: 01276 694730
www.toshiba-europe.com

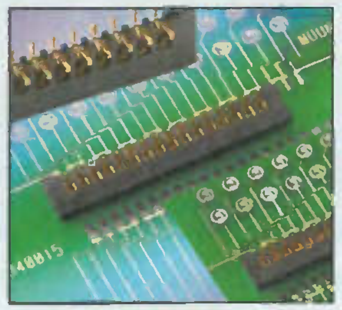
32-bit DSP with flash for control in C/C++

Two digital-signal processors (DSPs) for control from Texas Instruments are claimed to be the industry's first 32-bit control DSPs with on-board flash memory. With performance specified up to 150 million instructions per second (MIPS) target applications include industrial automation, optical networking and automotive control applications. The TMS320-F2810 DSP and the TMS320-F2812 DSP are based on the firm's code compatible TMS320C28x DSP core. This core is designed specifically for control applications and has extensions for up to 400 MIPS performance levels. The device's unified architecture that combines general-purpose processors and DSP capabilities allows both the system and math code to be developed completely in C/C++, reducing development time. The F2812 DSPs integrate 128kilowords (kw) of flash memory and the F2810 DSP 64 kw for reprogramming during development and in-field software updates. Acceleration technology allows code to be executed out of flash at 110 to 120 MIPs while time-critical code requiring 150 MIPs of performance can be executed directly out of the 18kw of on-chip RAM. In addition, the F2812 DSPs offer an external memory interface with an address reach of one megawords for systems requiring a larger memory model. Samples of the TMS320C28x DSPs are scheduled for availability during the first quarter of 2002, with volume production scheduled to follow in the second quarter. The F2810

DSP will be offered in a 128-lead LQFP, and the F2812 DSP in a 176-lead LQFP.
Texas Instruments
Tel: 01604 663399
www.ti.com

White LED driver IC

Linear Technology's LT1932 is a switch-mode, fixed-frequency, constant-current boost regulator for driving white LEDs. It has a fixed operating frequency of 1.2MHz, which the supplier claims allows the use of lower profile inductors and smaller ceramic capacitors, while reducing emitted noise. Typical efficiencies are above 80 per cent, compared with the 50 to 70 per cent efficiencies of charge-pumps. It uses a 36V switch allowing output voltages up to 35V, letting it drive up to eight white LEDs in series. It also uses a constant current topology to drive the LEDs in series ensuring there is a constant current source as the voltage drop across the LEDs varies with age and temperature. It can also drive 16 LEDs in two parallel strings of eight.
Linear Technology
Tel: 01276 677676
www.linear-tech.com



Low profile connector in a single piece

Hirose has developed single piece compression connectors that will let designers stack PCBs at 1.2 and 1.7mm. The surface mount connector is on a 1.1mm pitch and located onto the PCB using mounting bosses. The DF26 contacts connect with gold plated pads on the second board, thus freeing up PCB space. The gold plated contacts are rated at 0.5A at 50V AC. When mating with the second PCB, the contacts deflect 0.7mm and at the same time employ a wiping action to clean the mating surface, ensuring a gas tight connection. They are available in nine and 16 positions and supplied packaged in tape and reel for automatic placement.
Hirose
Tel: 01908 305400
www.hirose.co.uk

LCD controller kit

An analogue interface controller kit from Digital View is preconfigured for VGA, SVGA or XGA panel. Based on an ACG-1024 controller board, the kit includes cables, connectors, OSD switch mount, connection diagram and manual. It provides the components to run LCDs from LG, Mitsubishi, Samsung, Sharp and Toshiba. In a single-board format, the controller provides a connection to TFT LCD panels with resolutions of 640 by 480, 800 by 600 and 1024 by 768. It also provides full screen image expansion for XGA modes, with a daughter board for connectivity to TDMS and LVDS TFT panels.
Digital View
Tel: 0208 236 1112
www.digitalview.com

Capacitor range for audio design

Working with several loudspeaker manufacturers, capacitor maker Industrial Capacitors Wrexham has developed capacitors for audio applications. There are four ranges of metallised polypropylene film capacitors with a range of values and voltages. They are used in crossover units in hi-fi speakers and studio monitors. There are three axial, wrap and end-seal ranges – PW, PX and SA – and the ASMFP disc shaped capacitor. The ASMFP units are available in four voltages – 160, 250, 400 and 600V DC – with capacitance values ranging from 1 to 100µF.
ICW
Tel: 01978 853805

Single-supply op-amps for video markets

Maxim Integrated Products has introduced the Max4380 to Max4384 single, dual, triple and quad single-supply op-amps that are available in SC70, µMax, and TSSOP packages.
According to the supplier, the combination of rail-to-rail outputs, wide bandwidth and high impedance disable mode in small packages makes these op-amps suitable for multiplexing applications for the consumer video market.
They operate from a single +4.5 to +11V supply or from dual ±2.25V to ±5.5V supplies.
They require 5.5mA quiescent supply current while achieving a 210MHz –3dB bandwidth and a 485V/µs slew rate. Their input common-mode range extends beyond the negative power

Protocol tester handles latest Bluetooth specs

The PTW60 Bluetooth protocol tester from Rohde & Schwarz comes with more than 100 verified test cases that comply with version 1.1 of the short-range wireless specification. The test cases are of the prescribed qualification program that each product must pass before it can be brought into the market with the Bluetooth label. Providing test cases based on the SIG specifications makes it suitable for product development. Release 5.0 unit has test cases for the protocol layers baseband, link manager, L2Cap, service discovery protocol and generic access profile defined by the SIG. Signalling characteristics, link establishment, link disconnection and data transfer in the master or slave modes are tested. It is suitable for all makers of Bluetooth chip sets, protocol software and final products as well as for test houses and can be adapted to future versions of the Bluetooth specifications by updates.
Rohde & Schwarz
Tel: 01252 811377
www.rohde-schwarz.com

Please quote *Electronics World* when seeking further information

supply and their outputs swing rail-to-rail, making them suitable for high-gain applications. Large signal (2Vp-p) gain flatness of 0.1dB to 40MHz, low differential gain/phase of 0.02 per cent/0.08° and a spurious-free dynamic range (SFDR) of –65dBc at 5MHz are for standard and high-definition video applications.

Maxim
Tel: 0118930 3388
www.maxim-ic.com

High-side current sense monitor

In a five-pin package, the ZXCT1010 provides a technique for measuring load current with a sense resistor. Taking the voltage across a shunt resistor and translating it into a proportional output current, this current sense

monitor IC from Zetex uses a single scaling resistor to convert the current into a ground referenced output voltage and eliminates ground plane disruption. Taking less PCB space than alternative matched transistor pairs and offering an improved sensing accuracy of 1 per cent, the SOT23-5 packaged current monitor suits portable battery powered equipment.

With a quiescent current of 4μA, the ZXCT1010 operates over an input voltage range from 20V down to 2.5V. An enhanced version of the companion ZXCT1009 device, the ZXCT1010 high-side current monitor IC, features a separate pin to ground that improves the typical output offset from 500mV to 30μV.

Zetex
Tel: 0161 622 4444
www.zetex.com

Multi-chip package for mobiles

Fujitsu has announced a triple stacked multi-chip package (MCP), combining 64Mbit NOR-type flash memory and l6Mbit mobile fast-cycle random access memory (FCRAM) with an asynchronous SRAM-type interface, and 4Mbit SRAM. The expectation is that designs will use flash memory for program and data storage, high data-capacity FCRAM as working memory, and SRAM as cache memory for backup storage when downloading data or when the device is in standby mode. The MCP also achieves a reduction in the amount of power used.

The MB84VR5E3J1A1 offers an address access/program (one word) time of 85ns maximum, a standard NOR-type flash memory access time of 80ns, a mobile FCRAM random read access time of 90ns maximum and an SRAM read access time of 85ns maximum.

Fujitsu
Tel: 00 49 6103 690257
www.fme.fujitsu.com

Please quote *Electronics World* when seeking further information

4ns dual comparators

Elantec has introduced high-speed comparators. The EL5281C and EL5283C have a 6ns propagation delay and operate from 10mA of supply current per channel, and the EL5285C and EL5287C feature a 4ns propagation delay on 16mA of current per channel.
All have an input voltage range of 5 to 12V and have common mode input ranges that go below the lower rail. The output sections have their own supplies and can be used to interface directly with 3 or 5V logic.
Elantec
Tel: 0118 977 6020
www.elantec.com

Process technology for making ICs

Vitesse Semiconductor has announced its latest process technology for the manufacture of analogue and digital ICs for data transmission at rates in excess of 40Gbit/s.
The process is built around indium phosphide (InP) heterojunction bipolar transistors (HBTs). The first generation of the InP HBT process will be used to manufacture physical layer ICs for Sonet OC-768 applications and circuitry for 10Gbit/s systems that use RZ encoded data. Succeeding generations will provide ICs with up to 100Gbit/s levels of performance and integrated optical devices. This will provide the ability to make monolithic optical ICs.
The first generation InP HBTs use a vertical mesa isolated npn bipolar transistor.
Vitesse
www.vitesse.com

Direct stream digital-audio support

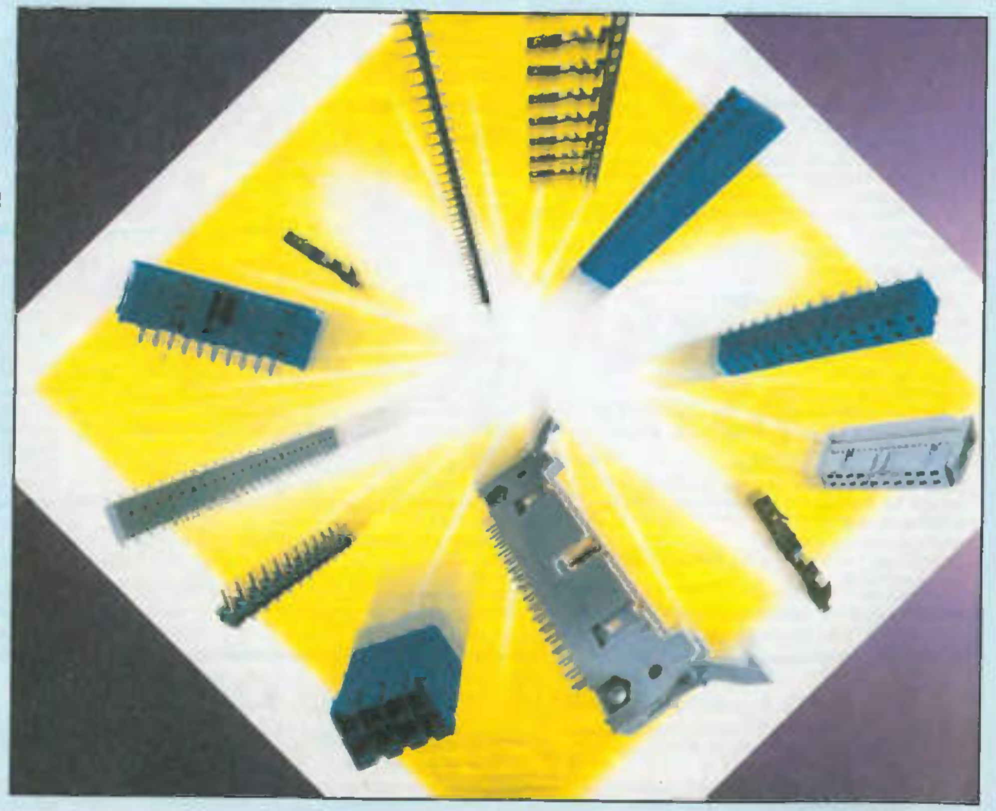
Wolfson Microelectronics has introduced a stereo d-to-a converter with support for the direct stream digital (DSD) audio data format. The DSD audio mode in the WM8728 extends the reach to cover existing PCM audio formats for CD and DVD players as well as the super audio CD (SACD) bit stream standard. DSD, the core recording technology behind SACD, enables the reproduction of sounds that are close to the original-source material. It uses a multi-bit sigma-delta converter architecture. Based on a 6-bit architecture, the 24-bit audio d-to-a converter integrates a DSP interface and the common industry audio formats, which let the user connect multiple d-to-a converters to a single time-multiplexed DSP interface. It provides a hardware or software interface for audio control. Data input word lengths in PCM format are from 16 to 24 bits with sampling rates up to 192kHz. In SACD mode, it provides two channels of 1-bit DSD with a sampling rate of 2.8224MHz.
Wolfson Microelectronics
Tel: 0131 6679386
www.wolfsonmicro.com

CPU board with up to 1Gbyte DRAM

Concurrent Technologies has released a single-slot 6U VME board. The CPU board has a choice of an 850 or 700MHz Pentium III CPU with up to 1Gbyte onboard DRAM within a single slot format. This means the VP CPI/Pxx can be used as the main CPU board in a VME system. The processor has 32kbyte of level one cache and 256kbyte level two cache. The board uses the 82440BX chipset, which supports the 100MHz front side bus. A heatsink is fitted to the CPU so no fan is used on the board. Up to 1Gbyte of DRAM can be installed on the board, using a combination of Sodimms and modules. This memory is accessible from the Pentium III, PCI bus, and VME bus.
Features include support for 10/100Mbit/s Ethernet interface via a front panel RJ45 connector, EIDE interface, AGP graphics with flat panel display support, two RS232 serial interfaces, floppy disc interface, real-time clock, parallel interface, two USB interfaces and a watchdog timer. There are also interfaces for keyboard and mouse. Hardware byte swapping and bus error detection are included. ■
Concurrent Technologies
Tel: 01206 752626
www.cc.co.uk

Board connectors get technical support

Distributor TTI is offering FCI's Basics+ connector product support package. The aim of the package is to offer engineers access to technical support, drawings, specifications and 24 hour samples. The core product range can be shipped within 48 hours, says the supplier.
The Du Box range of connectors offers board to board, and board to wire receptacles in a 2.54mm modular system. PCB or crimp versions are available as are polarised and anti-mis-match options.
Bergstik headers are also for 2.54mm use, and are available in single or double-throw types in PCB or SMT form. The Quickie is a flat cable, 2.54mm IDC system available to a standard specification of DIN 41651.
TTI
Tel: 01494 460000
www.ttieurope.com



The hysteretic regulator

Fernando Garcia argues that the ripple regulator – considered by many to be long obsolete – can offer benefits today due to advances in components. To demonstrate the point, he's produced a prototype that gives 3.3V at 11A from a 13.5V input with an efficiency of just under 90%.

Switch-mode power supplies, once the exclusive realm of military, aerospace or other ultra-high performance applications, have now become ubiquitous and indispensable. The main reason for the shift has been the reduced cost and complexity of the switch-mode power supply, or SMPS, thanks to the introduction of high performance, low cost SMPS-specific integrated circuits by most semiconductor vendors.

Most devices in this category have settled for a constant-frequency PWM architecture. Here, the error voltage obtained from the feedback and reference signals is compared against a constant, high-frequency triangle wave from an internal oscillator. The PWM signal thus generated is routed to protection and steering logic and to the drivers for the main power switch.

Introduced in pioneering devices such as the UC1524, this basic architecture has been substantially improved over the years. Refinements, such as current-mode control, have been added to the modern PWM-controller devices, but at heart they still maintain those same basic features. **Figure 1** shows a simplified schematic of this classic circuit topology.

Back in the days when a complex IC was anything with more than 20 transistors, there was another voltage regulator architecture that made use of those simpler devices. As a matter of fact, it predates the integrated circuits.

Although the earliest high-frequency, all-discrete circuit I've personally seen dates[1] to the mid-1960s, I'm sure that there must have been earlier realisations, employing SCRs or magnetic amplifiers. For the sake of clarity, note that in those days, high frequency meant almost anything higher than the 50/60Hz power line frequency – which effectively ruled out SCR phase-control circuits.

This architecture is the hysteretic regulator, also known by several other names like self-excited regulator, or simply a ripple regulator.

But why should we even bother discussing an 'obsolete' architecture? Well as the saying goes, "what goes around comes around". Present technology has added a new lease of life to many older circuits. Numerous engineers are discovering that these circuits married to newer technology may provide high

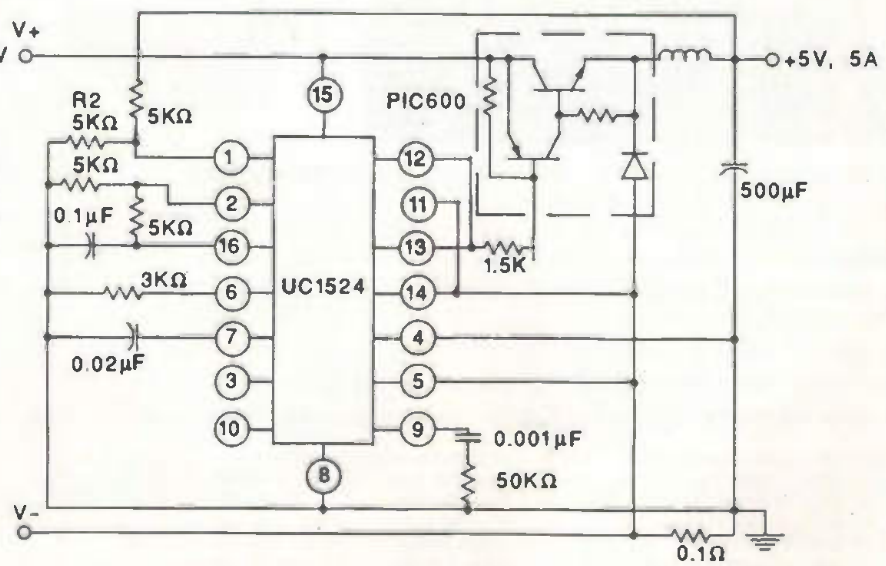


Fig. 1. The 1524 is a "classic" PWM controller IC.

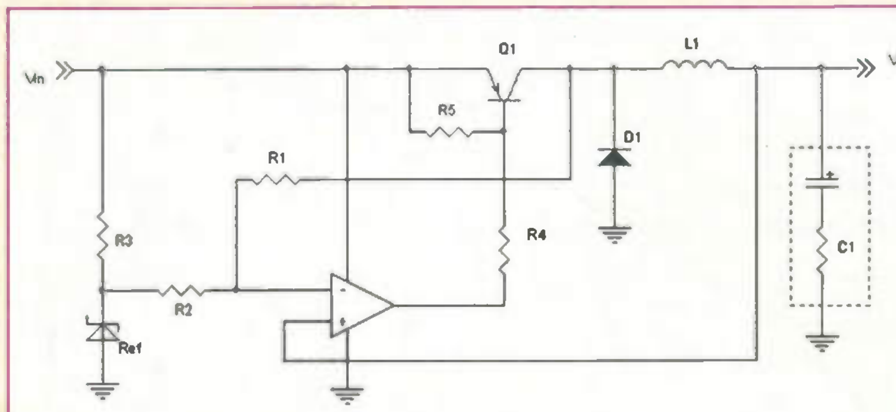


Fig. 2. Basic and crude implementation of the ripple regulator.

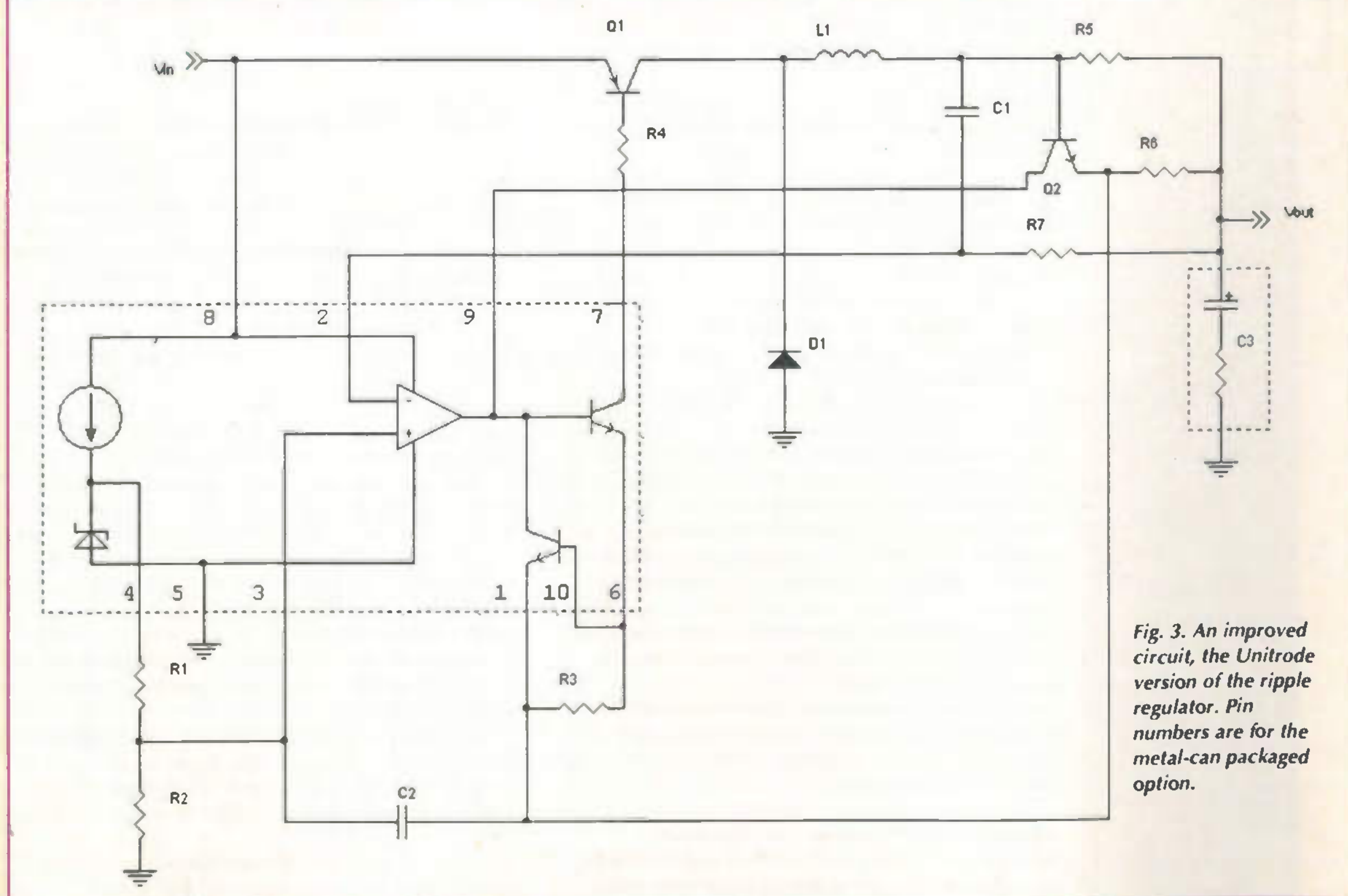


Fig. 3. An improved circuit, the Unitrode version of the ripple regulator. Pin numbers are for the metal-can packaged option.

Fig. 4. Normalised frequency change versus the input/output voltage ratio.

performance solutions to modern problems at a surprisingly low cost.

In many instances, the actual performance of older circuits was being impaired not by the topology itself, but by the intrinsic limitations of the then available electronic components. For instance, I distinctly recall the great pains an engineer had to endure to prevent secondary breakdown or thermal runaway in a power bipolar transistor.

Additionally, enhanced design tools such as CAD simulations or spreadsheets allow accurate optimisation of circuits rather than time consuming trial-and-error or cumbersome graphical techniques. Lastly, affordable and much improved laboratory instrumentation have allowed a more accurate assessment of the actual circuit operation. Remember the awkward and expensive CRT-storage scopes?

Step response to a load change

The solution we are looking for is load step response. The architecture shown in Fig. 1 includes a compensation network to provide closed-loop stability.

For many reasons beyond the scope of this article, the crossover frequency is set lower than half the switching frequency, which essentially means that the transient response will be slow.

The industry's response to the ever-increasing speed requirements for modern computing loads is to further increase the switching frequencies and to employ multiphase regulators. Such regulators are basically an array of parallel regulators with interleaved power pulses. This is a perfectly fine, although complex, approach to the problem.

On the other hand, several years back I built up a hysteretic regulator and I was pleasantly surprised by its simplicity and performance, especially with respect to transient performance.

This early circuit had several limitations, but I wondered, could it be improved with newer semiconductor devices? Texas Instruments thinks so, since it makes a dual combo regulator for fast DSP applications. One of its sections is built around a hysteretic regulator. One of its touted advantages is a fast transient response[2].

For many years. On-Semiconductor has offered yet another ripple regulator with a different control scheme (fixed on-time, variable off-time) which allows extreme simplicity[3]. My curiosity was thus awakened.

Self oscillation

A simplified, no-frills schematic of the ripple regulator is shown in Fig. 2. The heart of the circuit is a comparator, whose inverting input is connected to a reference voltage and the non-inverting input is fed back from the output.

At start-up, with the filter capacitor fully discharged, the comparator's output will swing low, biasing-on p-n-p transistor Q_1. It will start to pump current through L_1 and charge C_1.

The circuit will remain in this state until the voltage in the capacitor increases to a level that exceeds the reference voltage plus some hysteresis. At that point the comparator will swing high, cutting off the transistor and the associated charging current.

Inductor current, in a buck regulator fashion, is steered via the free-wheeling diode until it is depleted. At that point the output capacitor starts to discharge to a value equal to the reference minus the hysteresis voltage, where the comparator switches states again and the cycle repeats itself. The name given to the regulator becomes obvious: in steady state operation its output will oscillate between two voltage levels; the ripple voltage thus generated is required for proper regulator operation!

Ripple is not as bad as it seems; all switch-mode regulators generate ripple voltage. In this case, we can actually tame the ripple level to a known, reasonably small value, and one that's perfectly acceptable to the load.

If the load suddenly rises and the output voltage suffers a dip, the circuit turns on immediately,

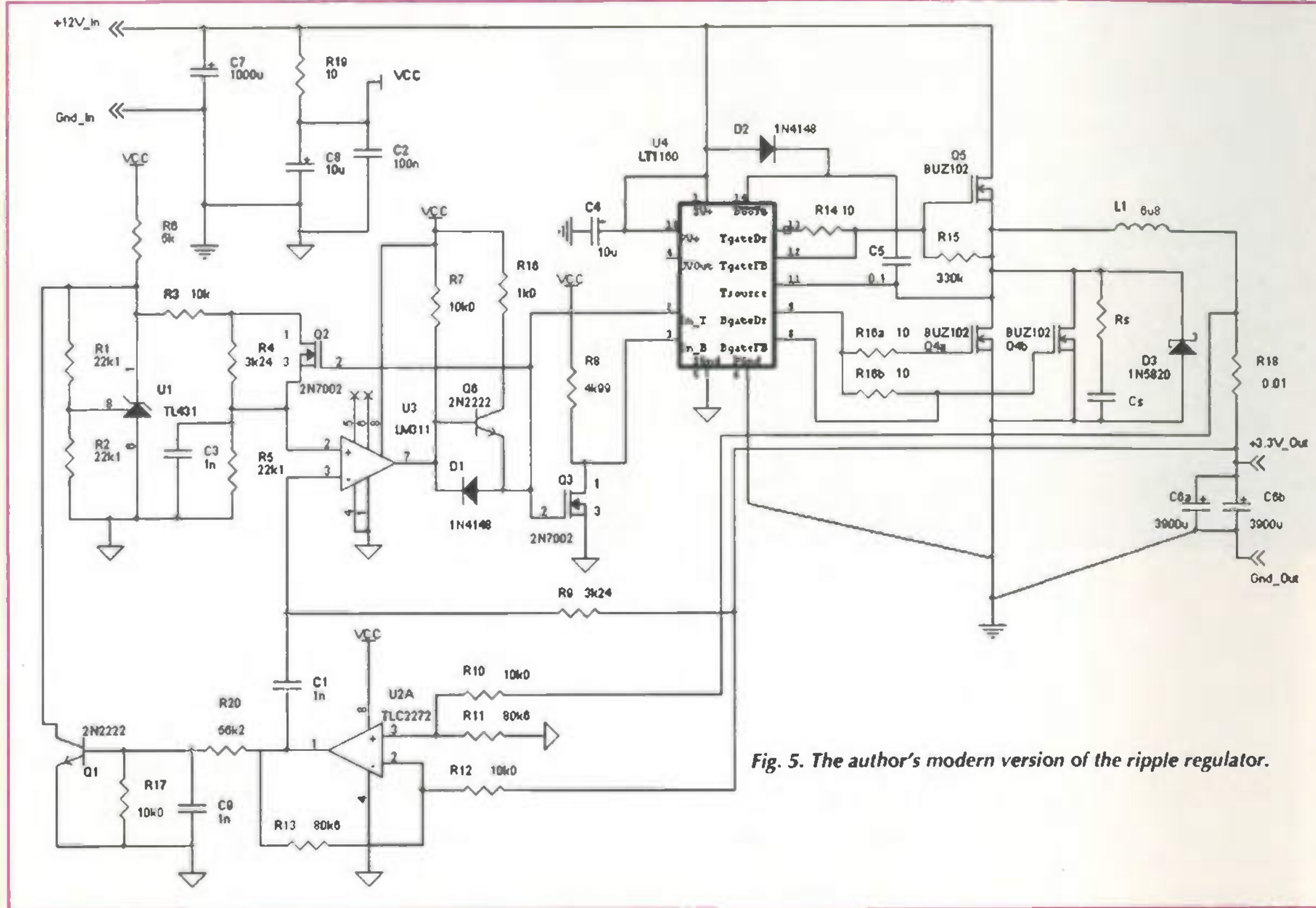


Fig. 5. The author's modern version of the ripple regulator.

replenishing the capacitor charge until the voltage has recovered. So much performance in such a simple circuit! But is it?

The short answer is: not quite. As every seasoned engineer knows, every circuit design is full of pitfalls and compromises. To gain an insight on the potential pitfalls consider the equations[4] that govern the turn-on and turn-off times:

$$t_{on} = \sqrt{\frac{2 * L * C * V_{ripple}}{V_{in} - V_{out}}} \qquad (1)$$

$$t_{off} = \sqrt{\frac{2 * L * C * V_{ripple}}{V_{out}}} \qquad (2)$$

$$f = \frac{1}{t_{on} + t_{off}} \qquad (3)$$

From this you can probably see that the switching frequency varies with the input voltage. This may be a drawback in some applications. However, there is more.

Consider the following: the ripple voltage will follow the comparator's thresholds. These, neglecting diode and transistor voltage drops, may be determined as follows:

$$V_{th+} = V_{ref}\left(\frac{R_1}{R_2 + R_1}\right) + V_{in}\left(\frac{R_2}{R_2 + R_1}\right) \qquad (4)$$

$$V_{th-} = V_{ref}\left(\frac{R_1}{R_2 + R_1}\right) \qquad (5)$$

Since the ripple voltage is equal to the difference in upper and lower voltage thresholds, then performing the subtraction of (4) and (5) gives:

$$V_{ripple} = V_{in}\left(\frac{R_2}{R_2 + R_1}\right) \qquad (6)$$

which means that the ripple voltage increases with increasing input voltages. This pitfall severely constrains the operating voltage range, which negates one of the advantages of a switch-mode power supply.

But wait; there's an additional *caveat* – a killer pitfall. This is the filter capacitor's effective series resistance (ESR), shown inside the dotted lines accompanying C_1 in Fig. 2.

The fact is that until now, only the ripple voltage caused by the charging/discharging of the capacitance itself has been considered. But as wary power supply engineers know too well, the impedance of a capacitor at typical switch-mode frequencies is dominated by ESR[5].

The end result is that the actual ripple voltage seen by the comparator has little resemblance to the actual capacitor charge and more with the product of the ripple current and ESR. The circuit becomes completely unstable unless the ripple current is minimised by an oversize inductor for the ripple voltage, or by large capacitors, or a combination of the two.

Those solutions not only increases the cost, size and weight of the supply, but they also slow down the load step response, negating the main advantage of this

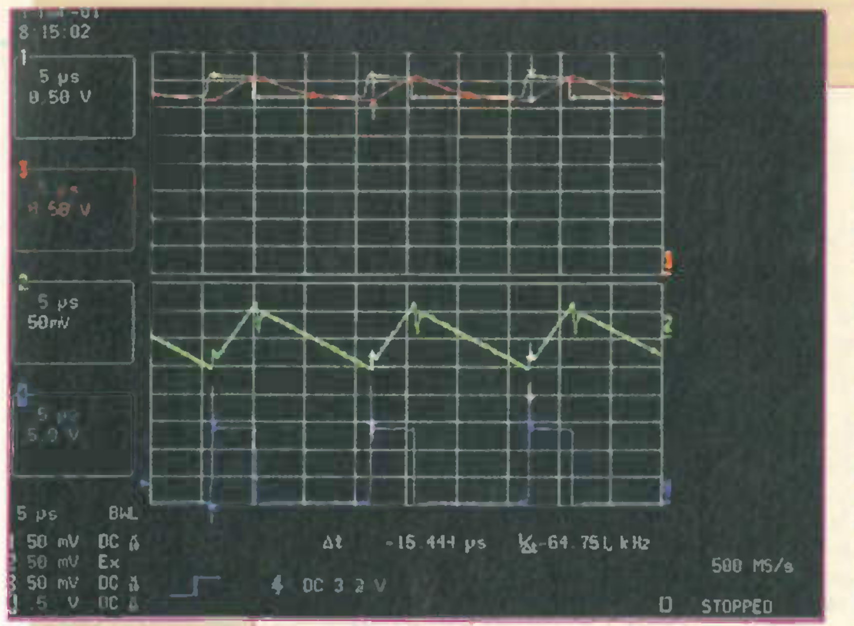


Fig. 6. Waveforms illustrating actual circuit operation.

type of regulator. For all the above constraints, plus the ones associated with the poor real world performance of early IC comparators and bipolar power transistors, this architecture found only limited applications.

An improved regulator

In the mid-1980s the Unitrode company was rapidly focusing itself on power components, both discrete and integrated. The company had introduced a new, high-performance power Darlington transistor. Like every company now and then, it would write an application note employing the new component.

Most application notes nowadays are mostly 'cookbook' recipes for the very specialised, high-integration ICs. However in the 1980s, the custom was to write very detailed application notes, with a full description of not only the part itself but on the associated circuit topologies. Many circuit tricks were discussed along with appropriate basic theory.

Although dedicated PWM IC controllers were becoming increasingly available, Unitrode wrote a landmark application note[6] describing a much improved ripple regulator. The proposed circuit employed one of the then ubiquitous voltage regulator ICs, such as the LM305 or the μA723.

An improved ripple regulator

Shown in **Fig. 3** is such a circuit built around a μA723; you may recognise it by the components enclosed inside the dashed box. The pin numbers correspond to the metal can packaging. I've shown the internal IC building blocks to aid with the following circuit description.

The improvement of this particular circuit over the previous one is that it has a DC feedback path and two AC feedback paths. The DC feedback works in the traditional fashion, which is to provide DC regulation. The first AC feedback, which follows through capacitor C_1 into the inverting input, has the following AC signal:

$$V_{inv(AC)} = (\Delta I_{L1} \times R_5) + \Delta V_{C3} \qquad (7)$$

Essentially, this means that the non-inverting input will see the output capacitor's ripple voltage plus the voltage drop across sampling resistor R_5 due to the inductor current swings.

On the other hand, the non-inverting input receives, via C_2, an AC voltage equal to:

$$V_{ninv(AC)} = (\Delta I_{drive} \times R_6) + \Delta V_{C3} \qquad (8)$$

This is the same output capacitor ripple voltage plus the voltage drop across R_6 due to the main transistor's base drive current.

On closer inspection, it becomes obvious that the capacitor voltage ripple, ΔV_{C3}, appears on both equations, so it is cancelled due to differential sensing.

This means that only the hysteresis voltage due to the transistor drive current is actually compared to the triangular voltage waveform developed by the inductor current swing through the sense resistor. You can equate (7) and (8) to obtain:

$$\Delta I_{drive} \times R_6 = \Delta I_{L1} \times R_5 \qquad (9)$$

So what's gained by doing this? Since the base drive current is regulated by the feedback action of R_3 through the transistor at pins 1 and 6 of the control IC, then the hysteresis voltage remains fairly constant. But most important is the cancellation of the capacitor's ripple voltage from the control equation. Assuming that the drive current and the resistor values to be constant, then the inductor's ripple current will be also maintained constant.

This is important, because the circuit's off time is governed by:

$$t_{off} = \frac{\Delta I_{L1} \times L_1}{V_{in}} \qquad (10)$$

Since the output voltage and the inductor ripple current are regulated, and assuming a constant inductance value, the off time remains fixed. It is independent of the input voltage and most importantly, from the output capacitor's capacitance or ESR values.

This circuit is still a buck regulator after all, its transfer function remains the same:

$$V_{out} = V_{in}\left(\frac{t_{on}}{t_{on} + t_{off}}\right) \qquad (11)$$

Since the off time remains constant, the circuit has to manipulate the on time in order to maintain regulation. The end result is that the frequency still changes with a changing input voltage.

Figure 4 shows a plot of normalised frequency versus the input-output ratio; if the maximum frequency is set at an input voltage of three times the output voltage, then at twice that value the frequency has already been reduced to 75% of its maximum value.

As the ratio is further reduced, the frequency approaches zero. Compare this against the behaviour of a fixed-frequency PWM controller.

Of course, several assumptions made above will not hold water in a real world circuit. For starters, the inductance value has a negative coefficient with respect to DC bias. This means that at low loads, the frequency may actually decrease.

I remember having built a prototype circuit and being pleasantly surprised with the high performance of such a simple design. However, the circuit would operate with difficulty at frequencies higher than 30kHz.

For starters, the μA723 regulator was never optimised for switch-mode operation. Its internal delays were not specified and unpredictable. They would vary significantly from vendor to vendor.

Then, the power bipolar transistors required a much more robust base drive to optimise speed than what could be achieved with the simple circuit. Baker clamps were often used, but there were many trade-offs between speed and efficiency.

Lastly, the current limiter that established the base drive had a soft gain and substantial temperature dependency. The end result was that the inductor ripple current, and all parameters that depend on it, drifted significantly during operation.

A second lease of life

With the advances in semiconductor processes, increasing integration became possible. Circuit simplicity was no longer an issue. Fixed-frequency, variable duty cycle PWM controllers became the norm. The hysteretic architecture was abandoned, save for a few isolated cases.

However, I was intrigued to see if, with newer, higher performance semiconductors, some of the limitations that plagued the earlier circuits could be overcome, while maintaining its basic strengths.

Several key limitations on the original circuit that could be improved upon were identified:

- Bipolar transistor power switches have been largely superseded by Mosfets. Although the obvious choice would be to employ a p-channel device to replace the p-n-p transistor, modern designs almost always employ n-channel devices. The difference in electron versus hole channel mobility makes the n-channel devices more efficient for a given die size. Fortunately, driver ICs that incorporate charge pumps that easily drive high-side n-channel devices have become widely available.

- Schottky freewheeling diodes were perfectly acceptable for 5V output supplies, but the efficiency penalty for lower voltage supplies becomes excessive. Synchronous rectification is now the norm. Semiconductor vendors, recognising the trend, produce combo driver ICs that satisfy the requirements for both the high side and synchronous-rectifier n-channel Mosfets. Linear Technologies' LT1160 device was selected since it provides fail-safe logic which maintains the upper Mosfet off if the bottom one is still on and *vice versa*, regardless of the control inputs. This is extremely important in totem pole power switches, to prevent shoot through conduction that may cause device failure.

- The heart of the circuit, the μA723 regulator itself, was state of the art 30 years ago but nowadays it leaves a lot to be desired. The internal comparator is extremely slow and has common-mode range

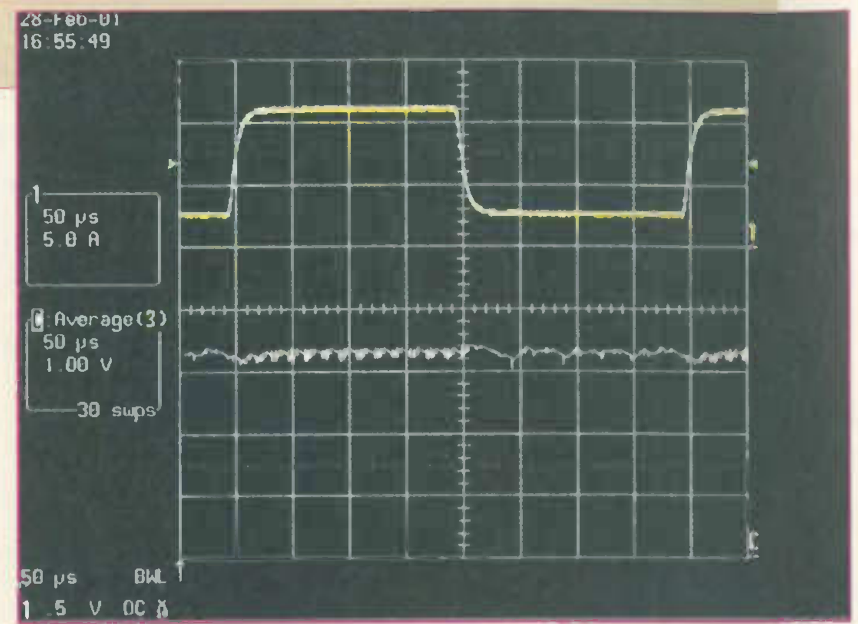


Fig. 7. Fast transient response is the circuit's main capability.

limitations. Its voltage reference has a wide tolerance, and the output transistor is not suited to drive Mosfets. The comparator section may be replaced with an improved, much faster device like the LM311 with reduced offset and wider common-mode range. There are many suitable candidates for the voltage reference, but the TL431 offers both economy and high performance, and tight tolerance versions are available from many vendors. Lastly, the drive requirements need be minimal as the power driver IC described above takes care of it.

- A current-limiting circuit that relied on the base-emitter voltage threshold was already a heavy efficiency burden for 5 volt supplies. With the lower voltages of 3.3 volts and below, the sampling resistor power loss becomes an unacceptable efficiency penalty. Therefore the sampling resistor in the new circuit drops a more manageable 110 millivolts at full load, which requires post-amplification to perform its task. Since this sampling resistor serves double duty as the inductor's current ripple feedback, the required amplifier has the triple task of low offset, high slew rate and a common-mode range that includes voltages close to ground. Fortunately, there are many newer op-amps that meet these requirements that won't cost you an arm or a leg.

- Finally, I wanted a comparator with a much more stable hysteresis voltage. Rather than employ the classic circuit in which positive feedback is taken from the comparator's output – with the input voltage variations affecting the hysteresis level – I used a variant of a circuit that I had previously published[7]. Here, a resistor that forms part of the reference voltage divider chain is shunted by a Mosfet toggled by the comparator output. The effect is such that the reference voltage at the non-inverting input increases when the output becomes high, much like the traditional positive feedback but much more repeatably.

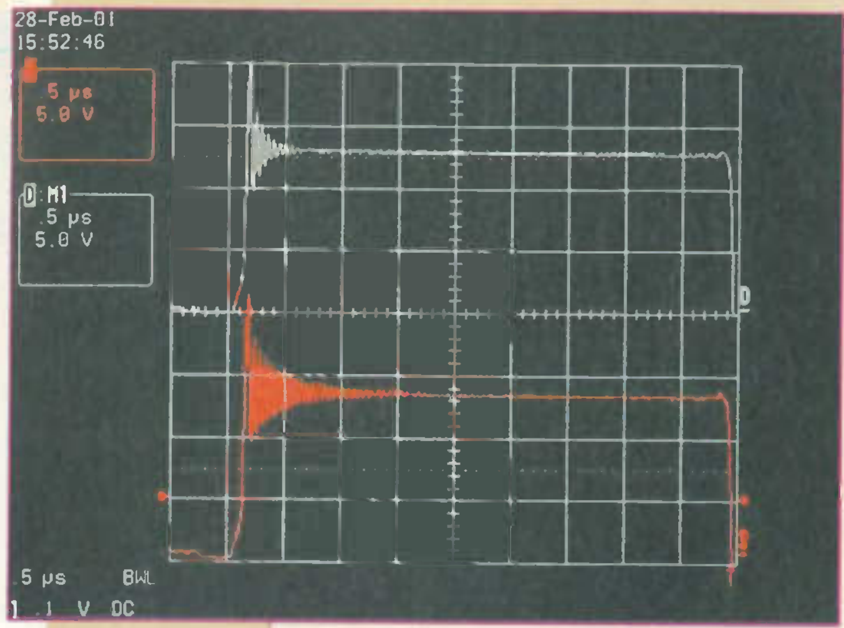


Fig. 8. Snubber reduces voltage ringing.

Practical circuit implementation

The completed prototype circuit is shown in **Fig. 5**. A stable voltage reference is provided via U_1, which is then fed to a resistive voltage divider string R_3, R_4 and R_5.

Transistor Q_2 shunts out resistor R_4 whenever its gate is driven high by U_3's output. Since this will in effect increase the reference voltage applied to the non-inverting terminal, positive feedback is generated. A stable value of hysteresis that is independent of the input voltage is thus achieved at the comparator's non-inverting input.

The comparator inverting input is fed from a DC feedback signal via R_9 and AC feedback via C_1. Feedback at DC provides the average DC output value. The AC feedback is a replica of the output inductor's current swing as previously discussed.

To maintain low losses, current sampling resistor R_{18} is kept at a very low value. Voltage thus developed requires amplification by U_2 and associated components. The operational amplifier used here offers low offset, reasonable slew rate capabilities and both input /output rail to rail capability.

The average DC component is low-pass filtered and applied to the base of Q_1. Whenever the maximum current limit is reached and the voltage is enough to bias Q_1 on, its collector will pull down the reference voltage.

Comparator U_3 is a mature yet quite fast responding device. However, having an open-collector output, its rise time can be significant which would add unnecessary delays and increase the output ripple, as will be illustrated below. Transistor Q_6 with D_1 assist in the pull-up current, allowing a much speedier transition. Transistor Q_3 inverts the comparator output. Thus complementary pulses are provided, necessary for the synchronous rectifier operation.

These pulses are applied to U_4, which is a Linear Technologies half-bridge driver. This is an interesting device, which offers the following features:

- TTL threshold-compatible inputs, which nevertheless may be pulled all the way up to the V_{CC} supply voltage without damage.
- Like most modern drivers, a bootstrap topology is employed. This works together with D_2 and C_5, and allows the use of n-channel Mosfets for the high-side power switch.
- Most important in a half-bridge configuration is the prevention of shoot-through currents. At best, these could cause an efficiency penalty and at worst they will result in total device failure. Different semiconductor manufacturers follow different approaches to prevent this, but the approach followed by LT is quite simple and effective: It senses the Mosfet's gate voltage, and until it has reached to a safe low level, it will not enable the drive for the opposite Mosfet. This regardless of the state of the 'top_drive' and 'bottom_drive' inputs.

The driver's top output is applied to the main power switch Q_5. The bottom output goes to Q_4, which works as the synchronous rectifier.

Resistor R_{15} keeps Q_5 off during power up, and resistors R_{14}, R_{16a} and R_{16b} are the gate limiting resistors that should be tailored for the rate of voltage rise.

Schottky diode D_2 conducts during the transition, where the top switch is turning off and the bottom one has not yet turned on due to the driver's anti shoot-through protection. Lastly, inductor L_1 and capacitors C_{6a} and C_{6b} form the energy storage elements in the classic buck topology.

Operation of the circuit may be better understood by looking at the waveforms of **Fig. 6**. Both top traces have identical vertical scales and offsets; they show the inverting and non-inverting comparator inputs.

The triangular trace is the amplified inductor current that has both an AC and DC component. It is compared against the reference voltage that has a squarewave hysteresis voltage. As the triangular current waveform crosses the appropriate thresholds, the comparator changes states.

The middle trace is the output ripple voltage. The bottom trace is displaying the voltage at the switching node – the Mosfet and inductor junction – showing the effective duty cycle for the power switches.

On closer inspection, you may observe that although the top trace has switched states, it takes the bottom trace about 1µs to respond. The reason for this is in the pull-up and driver delays. The net effect is that there will be a slight overshoot or undershoot in the current, and the overall ripple will be slightly higher than anticipated. Therefore, the emphasis to minimise delays with the active pull-up circuitry.

How about its touted advantage, the fast transient response? **Figure 7** shows the results. The top trace shows the load current jump from a minimum (2.5 amps) to a maximum (11 amps) and back. The electronic load was set for a slew rate of 5A/µs, although the actual slew rate was less due to wiring inductance.

However, you can see the minuscule voltage overshoot and undershoot, and the almost instantaneous settling time, in the bottom trace. This achieved with a very modest 65kHz switching frequency!

The only post-processing performed to the bottom trace is high-frequency averaging to remove the switching noise that would otherwise completely obscure the voltage change.

Increasing efficiency

Designing a power supply, like any other electronic circuit, is a succession of tradeoffs. These tradeoffs usually involve cost versus performance.

Of the performance, efficiency is usually one of the major considerations. The prototype circuit that I built achieved a 89.5% efficiency at a 13.5V input and a 3.3V output at 11A. You may achieve a larger or smaller efficiency figure depending on the tradeoffs as follows:

- Frequency: The first decision in designing a switch-mode supply is to determine the ballpark frequency that it will operate at. Everything else depends on that. A higher frequency is sought after if size is your primary concern. However, be prepared to start spending some serious money on components. For instance, aluminium electrolytic capacitors – even those low impedance types – start to rapidly run out of performance at frequencies higher than 100kHz. Specialty capacitors like solid tantalum or OsCon types are required, and multiple paralleled units must be used. Besides, electromagnetic interference becomes a concern. For that reason I chose a frequency between 60 and 70kHz.
- The energy storage inductor is a primary source of losses. Depending on the core material and geometry, wire, etc., the losses may vary greatly. My first attempt at using a ferrite rod core inductor generated a 4.5% efficiency penalty. The copper losses were excessive. Employing a powder-iron toroid shifted the losses from the copper to the core, as powder-iron core losses increase very rapidly above 50kHz. Employing a bifilar wound, Kool-Mu toroid[8], allowed me to achieve the quoted efficiency values.
- I have briefly touched the electrolytic filter capacitors. For cost reasons, I chose low-Z aluminium electrolytics. The key here is the parallelling of multiple devices.
- Most modern switch-mode supplies employ exclusively n-channel Mosfets. Semiconductor companies are always improving these devices with newer processes. At the time of this writing (early 2001) Infineon is marketing a 'Cool Mos' device with wonderful specifications and at a very attractive price. By the time you read this, its major competitors may have already introduced similar or improved devices.
- The duty cycle of a 3.3V supply operating from 13 or so volts will be approximately 1/3 of the time 'on', 2/3 of the time 'off'. Therefore the bottom Mosfet will be conducting most of the time so it makes sense to parallel two devices to reduce the static losses. Make sure that the Mosfet driver can handle the added load. If you don't, any gains in static losses will be more than made up by the dynamic losses.
- The icing on the cake is both the parallel schottky diode and the snubber network. As I mentioned before, there will be a brief instant where both the top and bottom Mosfets are not biased on. Although the body diodes may carry the inductor current, a better approach is to employ a paralleled schottky diode (D_2). Lastly, the snubber network resistors and capacitors need to be considered. Parasitic elements are an inescapable fact of life. They will create a severe voltage overshoot at the power switch node. How severe the overshoot depends on the actual components, component location, trace layout, etc., but it could be severe enough to damage your switching components. You may attempt to tame them by increasing the gate resistors and thus the rise time, but dynamic losses will increase. Sometimes the only solution is to add a small dissipative *RC* network.

Depending on how much energy you have to eliminate, the RC snubber will impact on the efficiency loss. In my particular case, the actual overshoot was not excessive, as shown in **Fig. 8**, bottom trace.

The snubber was only employed to damp the ringing, upper trace, and thus minimise interference. The efficiency penalty was a mere 0.5%.

In summary

The ripple regulator is not a universal panacea. In particular, the variable frequency operation and the large controller shifts as the inductor current becomes discontinuous may preclude its use in many applications. It is however, a simple, fast, almost fail-safe architecture that may appeal to certain solutions.

I would like to acknowledge the support from Martha Gonzalez during the development of this design. ■

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CIRCUIT IDEAS

Q-meter signal generator add-on

While developing RF circuits in the LF to HF range, a need arose for equipment to indicate the Q of resonant circuits, and the facility for trimming inductors for specific frequencies, i.e. a 'Q meter'.

The circuit diagram shows an RF generator add-on that measures Q directly. It can be used to determine the resonant frequency of a particular LC combination. While not a precision Q meter – there is some non-linearity due to the characteristic of diode D_1 – the small outlay is justified when compared to a commercial Q-meter.

Transistors Tr_1 and Tr_2 are required to drive the test circuit L_x and VC_1. The 50Ω or 75Ω modulated RF output of a signal generator is too high an impedance to series resonate the LC circuit.

Diode D_1 extracts AF modulation to drive the meter M_1. This modulation is amplified when the switch is in the 'SET' position, or buffered when 'Q' is selected.

The op-amp records the peak value of the recovered audio modulation. Capacitor VC_1 is calibrated in picofarads, either *in situ* using a capacitance bridge, or by substitution of close tolerance capacitors for C_x.

To calibrate M_1, it is necessary to have a tuned circuit of known Q; any RF coil can be used. With a 'scope or RF voltmeter connected across C_x and the inductor connected to terminals L_x, resonate the circuit by adjusting either the frequency of the generator or C_x for maximum indication. Transfer the probe to the anode of D_1. Measure the voltage in the SET and Q positions of SW_1:

1. Adjust the generator RF output for a reading of 20mV in SET and set RV_2 to read half scale, i.e. 100 on a 0-200 meter scale.
2. In position Q, if the resulting measured value is e.g. 1000mV, then the Q=50 (1000/20), so RV_1 should be adjusted to give a reading of 50 on the meter. The instrument is now calibrated to read Q directly on a scale of 0-200.

In use, if the SET reading is adjusted for half scale, the Q range is 0-200 as above, while if the SET reading is adjusted for full scale, then the Q range is 0-100. Other values of meter FSD and scales could be used, e.g. a meter scaled 0-10 and 0-30, could provide Q ranges of 0-10, 0-30 and 0-100 by selecting the appropriate SET indication on the scale.

Adjustment of the signal generator frequency and output should always be carried out after resonating the circuit as, with high Q coils, there may be some change in loading on the generator.

Depending on the available output level from the signal generator and FSD of M_1, R_7 may be required and the value of R_8 may need to be modified.

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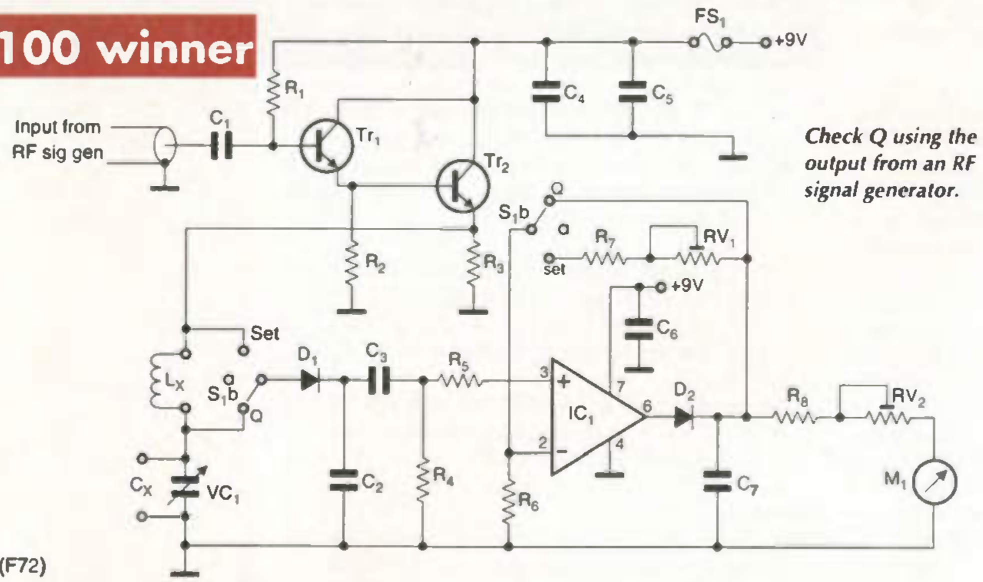
Circuit Ideas editor's note: if a silicon diode is used at D_1, a 680kΩ resistor should be connected in parallel with C_2.

£100 winner



(F72)

Check Q using the output from an RF signal generator.

Fact: most circuit ideas sent to *Electronics World* get published

The best circuit ideas are ones that save time or money, or stimulate the thought process. This includes the odd solution looking for a problem – provided it has a degree of ingenuity.

Your submissions are judged mainly on their originality and usefulness. Interesting modifications to existing circuits are strong contenders too – provided that you clearly acknowledge the circuit you have modified. Never send us anything that you believe has been published before though.

Don't forget to say why you think your idea is worthy.

Clear hand-written notes on paper are a minimum requirement: disks with separate drawing and text files in a popular form are best – but please label the disk clearly.

Send your ideas to: Jackie Lowe, Cumulus Business Media, Anne Boleyn House, 9-13 Ewell Road, Cheam, Surrey SM3 8BZ

Components

R_1	120kΩ	VC_1	500p max	Tr_1	2N2222
R_2	330Ω	C_1	6.8nF	Tr_2	BD135
R_3	20Ω (2×10Ω)	C_2	6.8nF	IC_1	CA3140
R_4	680kΩ	C_3	470nF	M_1	200μA
R_5	22kΩ	C_4	10nF	L_x	Test L
R_6	6.8kΩ	C_5	100μF	C_x	External C
R_7	SOT(0)	C_6	220μF	FS_1	500mA
R_8	4.7kΩ	C_7	10nF		
RV_1	220kΩ	D_1	Ge diode		
RV_2	500Ω	D_2	1N4148 or equiv.		

Battery life extender for torches and bike-lamps

Here, a CMOS 555 timer IC is connected as an astable driving a transistor switch. This ingenious circuit is arranged so that the on time increases as the battery voltage falls, thus maintaining a constant illumination. A 2-5V lamp is run at the equivalent of 1.8V increasing both lamp and battery life.

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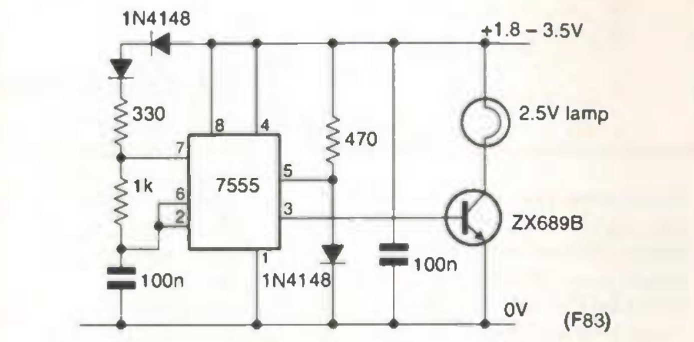
£50 winner



As the battery depletes, this regulator increases the PWM circuit's on time so the lamp stays roughly at the same brightness. Used correctly, it makes the battery last longer.

Non-locking push-buttons latch selection

Although the following circuit was designed to be used in a preamplifier, it can be used in any application where a selection has to be performed.

The circuit is designed for input selection with push-buttons rather than with a rotary selector switch. It can replace interlocked push-switches.

In the prototype, the inputs are selected with PCB-mounted relays on the same board as the input terminals. This has the advantage that the signals don't have to be routed through the whole amplifier to reach the selector switches at the front.

It is possible to use solid-state switching, using CMOS switches. In its present configuration it allows the selection of one input out of six signals, although it is easily changed to increase the number of channels.

The circuit is built around two 74HC238 ICs connected as a 1-of-3 selector. A feedback network is created for each IC, with D_4, D_9. This feedback network maintains the selected input after the push-button is released. Diodes D_{10}, D_{11} or D_{12} will disable Ic_2, when an input is selected with Ic_1; $D_{1\text{-}3}$ likewise when PB_3, PB_4 or PB_5 is pressed.

An extra feature is that if more than one button is depressed, all the inputs are switched off until only one button remains depressed, since only outputs D_1, D_2 and D_4 are used from Ic_1 and Ic_2. Thus if e.g. PB_0 and PB_1 are pressed, only the unused output D_3 is enabled.

An extra input is provided to disable the outputs during on and off switching of the amplifier. During, and after, switch-on no output is selected until one of the push-buttons is pressed. On power down, the inputs are switched off by the enable signal. A resistor divider is provided to attenuate the enable signal.

To drive the input selecting relays a ULN2803 driver IC is used, which also drives the LEDs that display the selected input.

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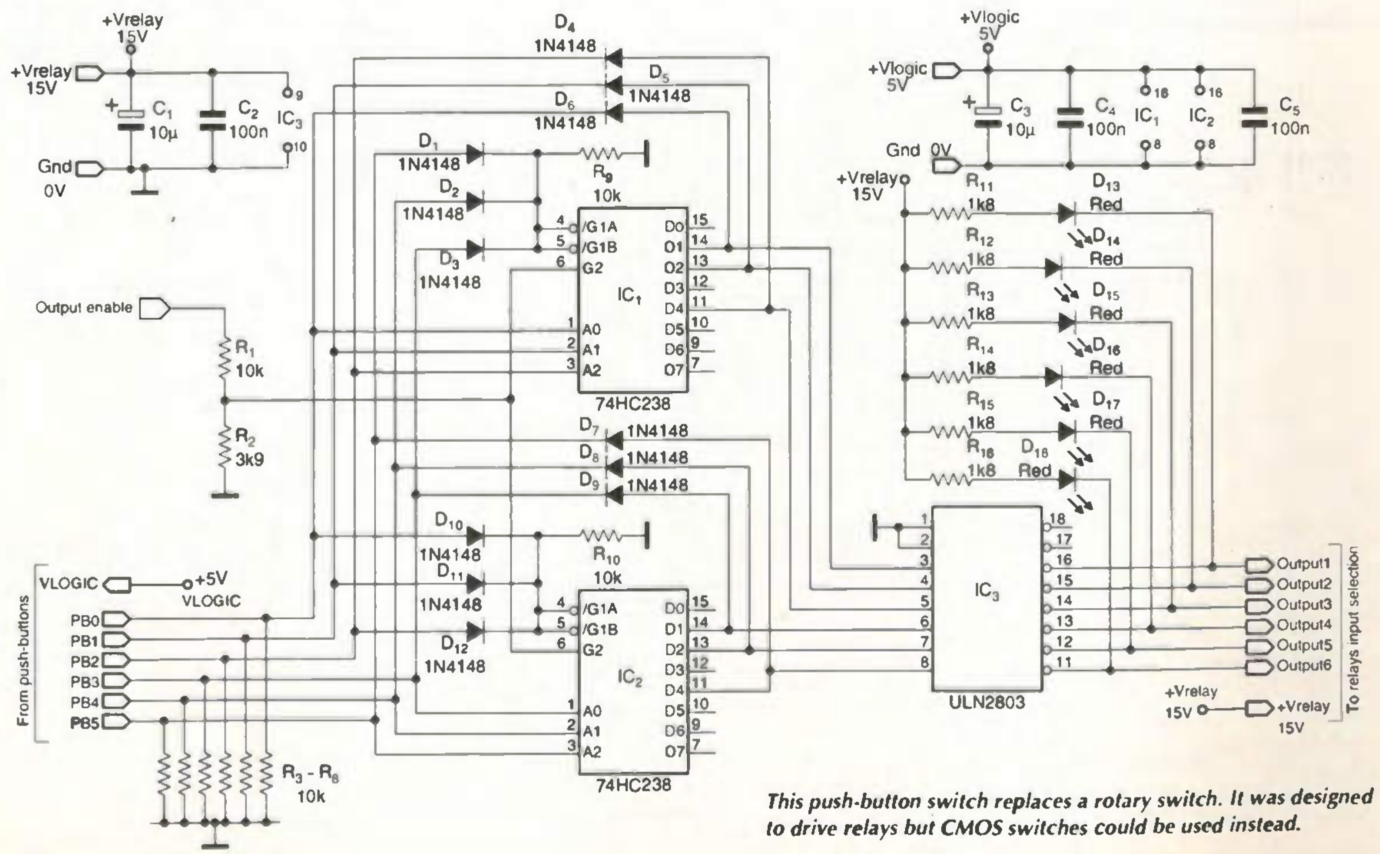

This push-button switch replaces a rotary switch. It was designed to drive relays but CMOS switches could be used instead.

Circuit Ideas editor's note: the following is a correction Jim Watson's 'Phase shifter for headphones', published on page 156 of the February 2001 issue. The non-inverting inputs of the two right hand op-amps should not be connected to 'Headphone ground (centre data)' as shown, but to the input ground at the left-hand side of the circuit diagram.

Audio level indicator

In the February 2001 issue, there was an interesting circuit for audio level indication using a tricolour LED by Graham Booth. However, I wanted a version that would operate from ±9V rails, such as from two PP3 batteries.

Using a TL074, I ran into headroom problems. Zeners such as specified by Graham are not very effective below about 4V, and anyway, I didn't have any in stock. I then thought of the crafty engineer's low-voltage reference, a green LED. This has a forward voltage of around 2V.

Substituting these for the original zeners, the indicator worked well down to a supply voltage of ±7V, as given by run down PP3s. A level control was fitted to the input so that the peak green illumination could be set for –6dBm, being standard level for much unbalanced audio equipment.

At +6dBm, the red LED is fully lit.

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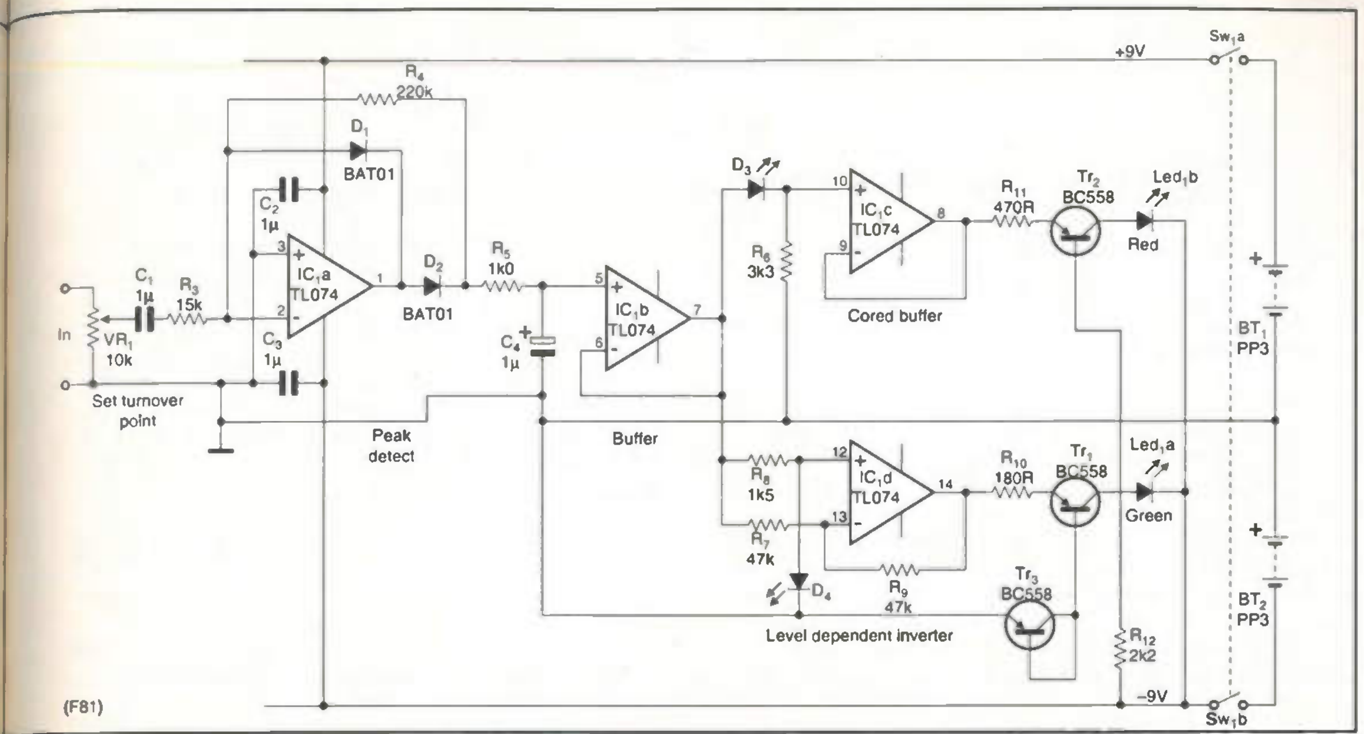

Three-phase sine generator

This generator can be used for example to control the velocity of rotation of a synchronous motor.

The circuit diagram of the generator is given in Fig. 1. Two phases – *R* and *S* – are generated by two counters. These are each followed by a fourth-order low-pass filter to produce a sinusoidal waveform; the third phase, *T*, is obtained by adding and inverting *R* and *S*.

For minimal distortion the output signals of the counters have to be symmetrical and to make it possible to achieve a phase-shift of 120° the number of count pulses has to be a multiple of three.

The first counter, IC_2, is a binary counter which is loaded with 2 by means of IC_{3A} when the output reaches 14, so the output Q_3 is symmetrical around 8 (6 count pulses high and low), **Fig. 2**.

Second counter IC_4, also binary, is loaded with 2 when the first counter reaches 6; this could be done by ANDing Q_1, Q_2 and inverted Q_1 of IC_2. That would require an extra inverter, so a variant is used; namely Q_1 and Q_2 of IC_2 and Q_2 of IC_4 are subjected to an AND function. After maximum two count cycles the desired loading of IC_4 is acquired.

Outputs Q_3 of IC_2 and Q_3 of IC_4 are fed to two fourth-order filters, IC_6 and IC_7, whose outputs are sinusoidal and form phases *R* and *S*. Phases *R* and *S* are added and inverted in IC_5 to obtain phase *T*.

The low-pass filters are clocked by a frequency 120 times higher than the output frequency. This gives a filter cut-off frequency that is 1.2 times higher than the output frequency to minimise undesirable phase-shifts, due to the low-pass filters.

As a result: the circuit delivers three sinusoidal signals of 1.5V pk-pk and the amount of clock ripple in the output is 40mV pk-pk; the frequency-range is 0-20kHz.

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Fig. 1. Digitally-generated three-phase square waves are converted to sine form by low-pass filters.

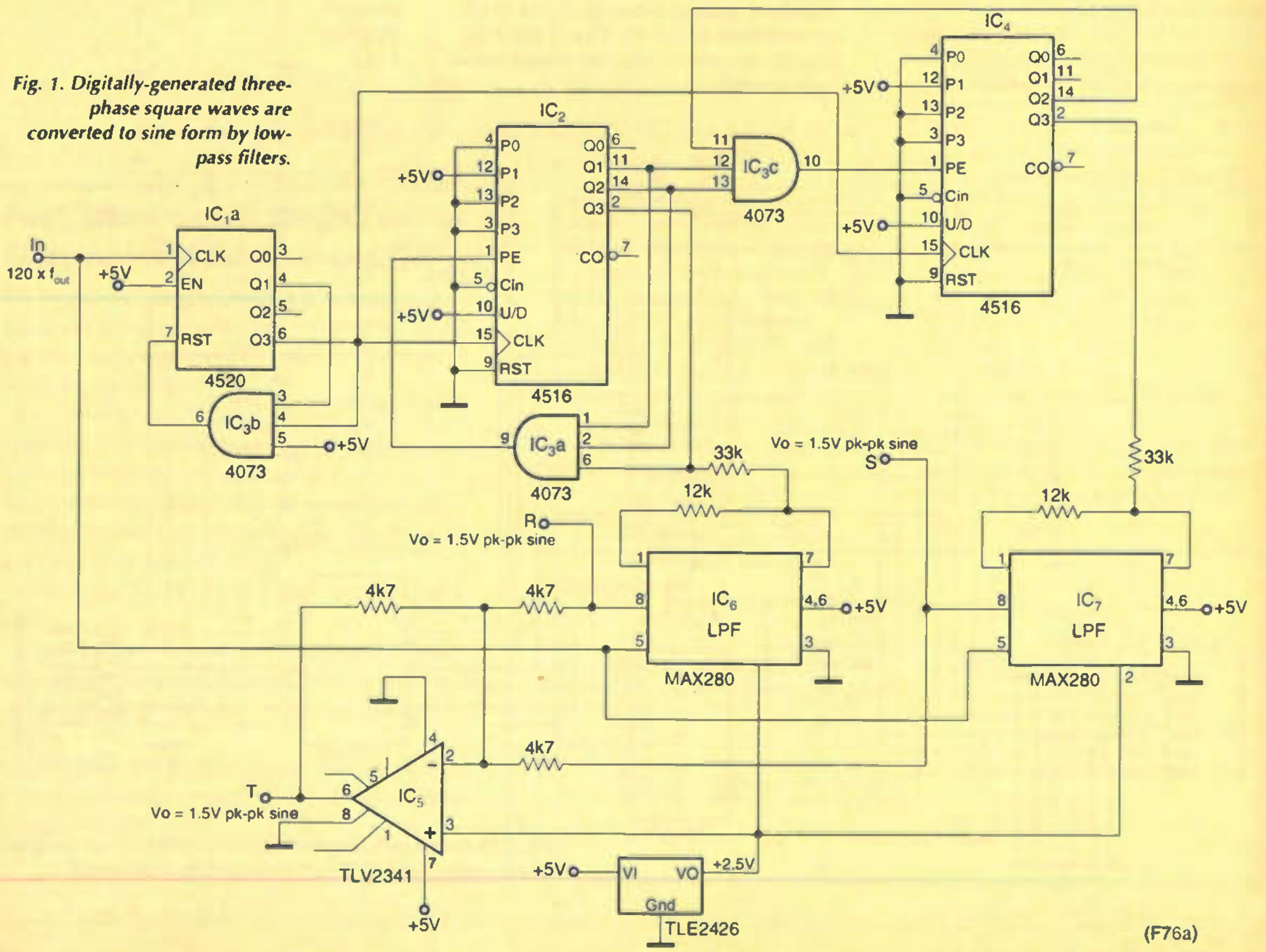
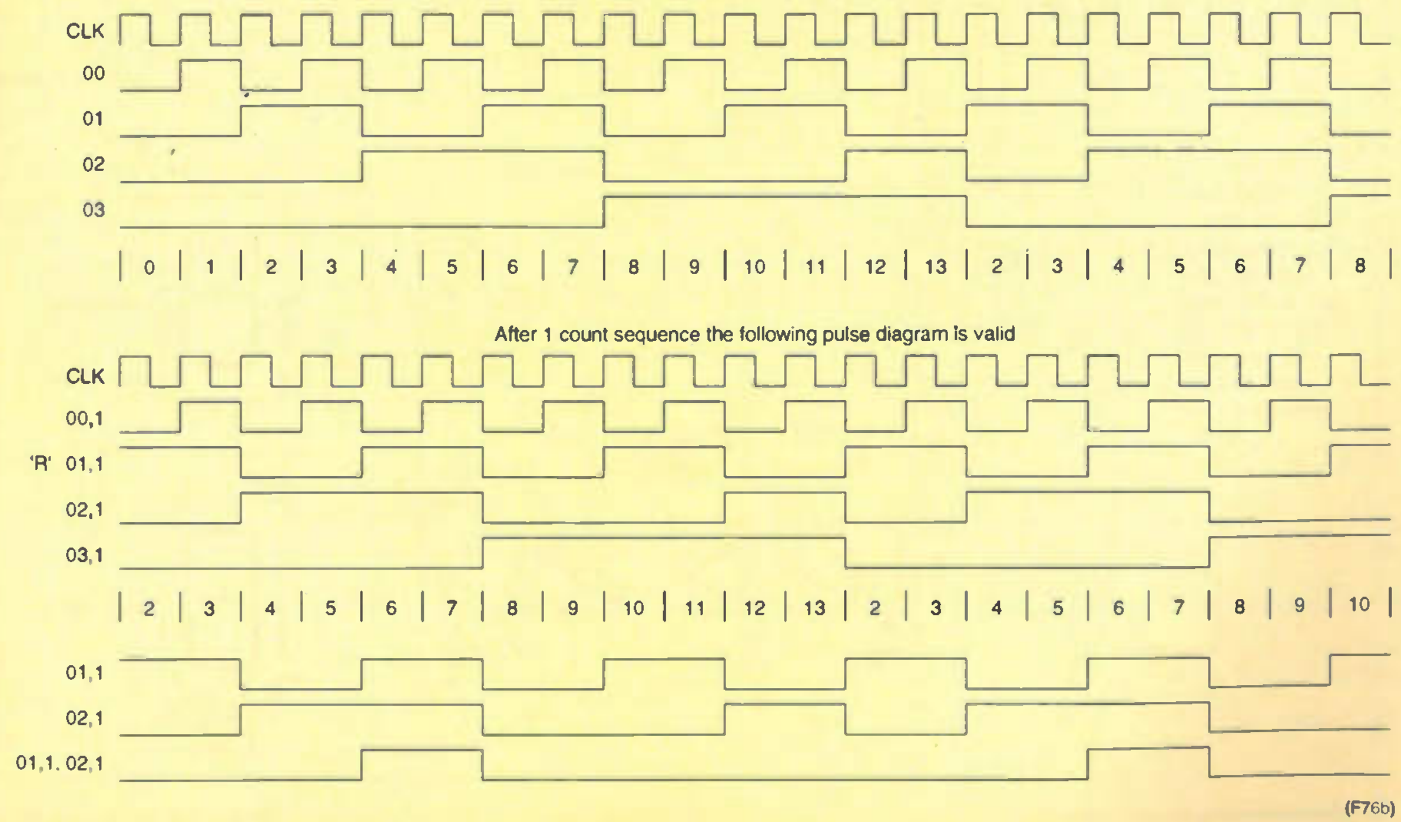


Fig. 2. Timing for the three-phase generator.



Simple three-state logic probe

The following logic probe has proved useful when teaching digital logic to undergraduate students.

While most logic probes use coloured LEDs to indicate different logic states, this one displays logic conditions on two seven-segment LED displays.

Shown in Fig. 1, the circuit uses two ICs, namely a CMOS 4049 inverter and a ULN2001A Darlington driver array. There are two common-anode LED displays – HDSP-3400s in the prototype – and seven resistors.

Output provided is as shown in Fig. 2, whereby the conditions of 'logic 0', 'logic 1' and 'open circuit' are displayed as 'LO', 'HI' and 'I I', respectively. Note that since the f and e segments of each LED display require to be illuminated for each of the three output conditions, they are connected through current limiting resistor R_6 to ground. They remain on continuously.

Resistors R_1, R_2 and R_3 form a potential divider such that with the values shown and the test probe connected to a 'logic 0' signal the voltage at point 'A' is approximately 1.5V while point 'B' will be close to ground potential.

Since CMOS technology nominally regards an input level of less than half of the supply voltage as 'logic 0', pin 2 of inverter (a) will be 'high' while pin 6 of inverter (c) will be 'low'. As a result Darlington driver (a) will sink current through segment d of the left LED display and current limiting resistor R_5. It will also sink current through segments a, b, c and d of the right display via resistor R_7. The segments thus illuminated combine to form the 'LO' output on the LED displays.

With the test probe connected to 'logic 1' point 'A' remains close to 5V, while the potential at point 'B' is approximately 4.7V. As a result pin 2 of inverter (a) will be 'low' preventing Darlington driver (a) from illuminating the LED segments which were required to form the 'LO' display.

Output pin 6 of inverter (c) will go 'high' allowing Darlington driver (b) to sink current through segments g, b, and c of the left LED display and resistor R_4. The display will read 'HI' in this case.

When the test probe is unconnected, the potential at point 'A' is approximately 2.7V, while the voltage at point 'B' is around 1.8V. This results in a 'low' being presented to the inputs of both Darlington drivers, effectively preventing them from sinking current through the displays. In this condition the display will show 'I I'.

Note that all unused CMOS inputs should be tied to ground or to the supply voltage.

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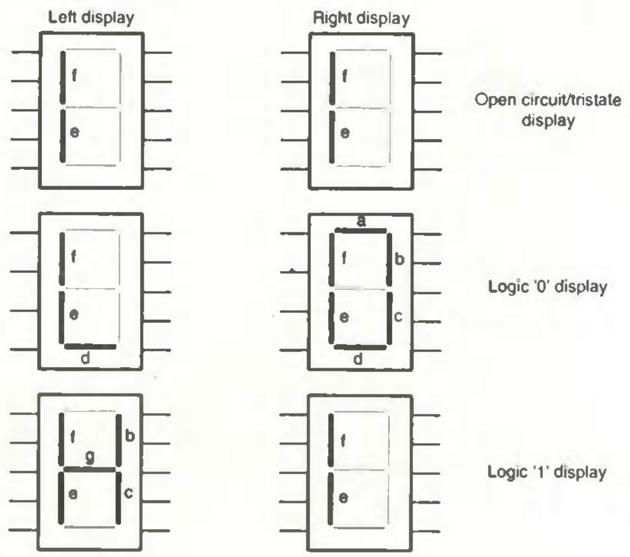
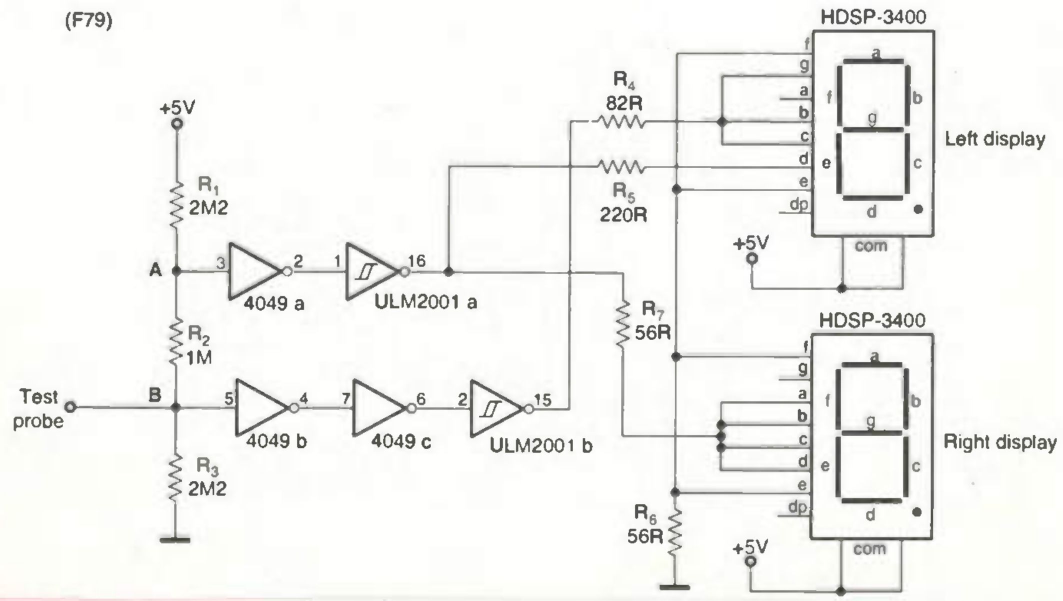


More user-friendly than a simple led indicator, this logic probe reads 'HI' or 'LO' for the two logic states and 'II' when there's no signal.

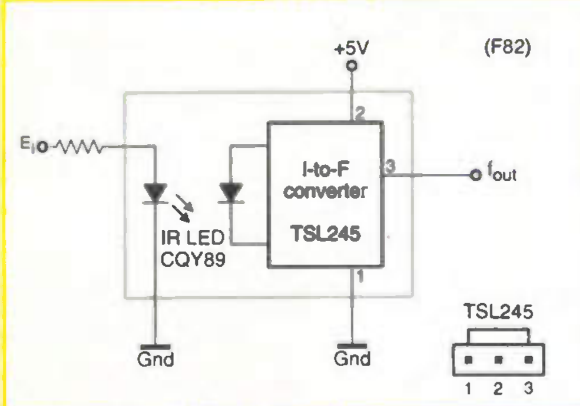


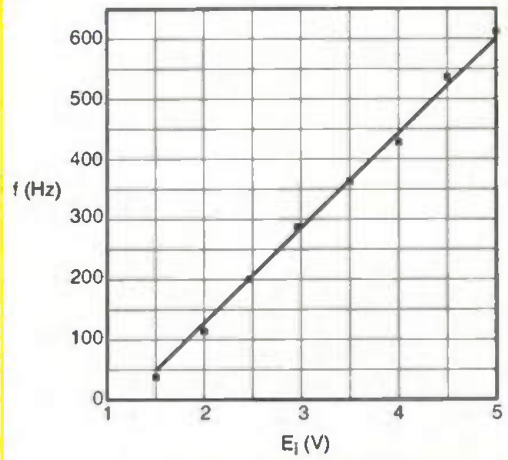
Simple VCO

In this simple VCO circuit, a CQY89 IR light-emitting diode and a TSL245 light-to-frequency converter are mounted close together in a small box, which is shielded against ambient light.

The relationship between input DC-voltage and output frequency is given in the graph. A higher output-frequency can be obtained by using more leds or by mounting the TSL245 closer to the CQY89.

W E C Dijkstra
Waalre
The Netherlands
F82





Easy to implement isolating voltage-to-frequency converter. Its frequency range can be tailored by altering the distance between the LED and detector or adding more LEDs.

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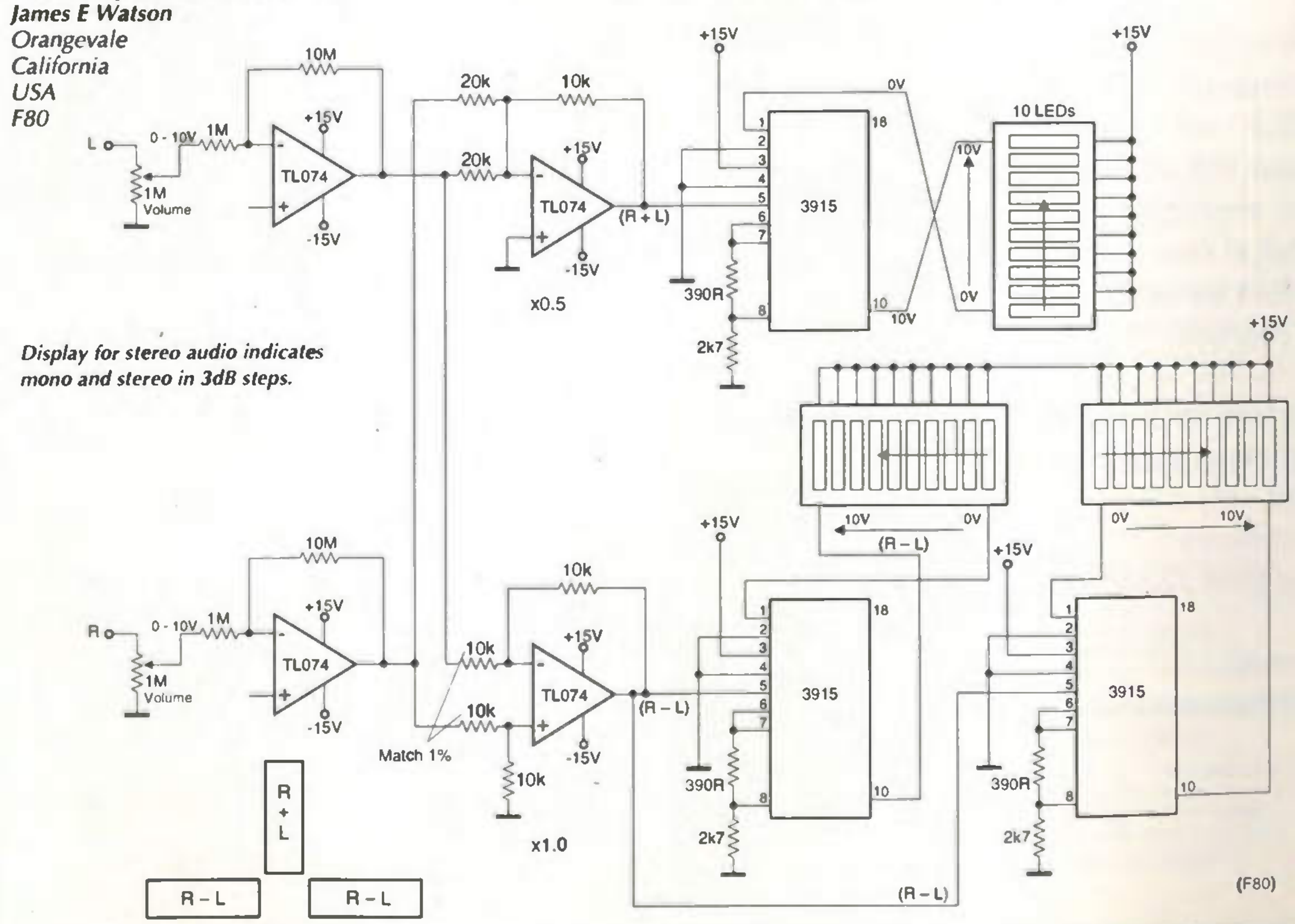
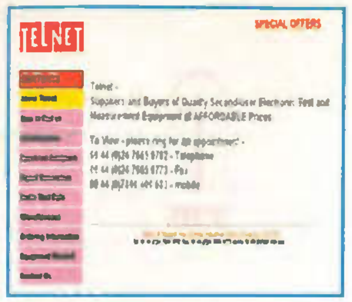
This matrix unit produces R+L and R–L displays. Monophonic power data is represented by (R+L) while (R–L) is the stereophonic data. Mono programme material indicates (R+L) variations in 3dB steps only. Stereo signals indicate (R+L) data on the centre display and (R–L) data on the side displays, in 3dB steps.

James E Watson
Orangevale
California
USA
F80



Display for stereo audio indicates mono and stereo in 3dB steps.



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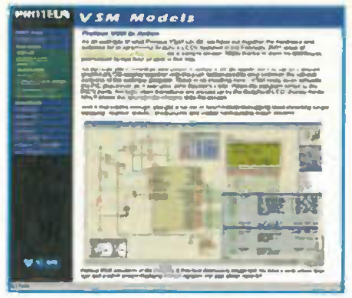

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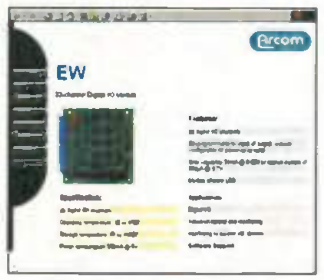

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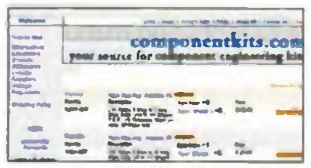

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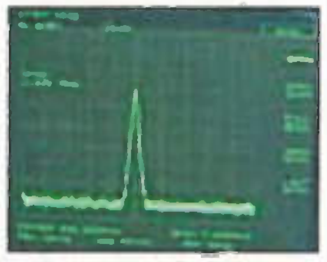

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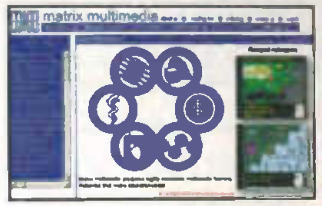

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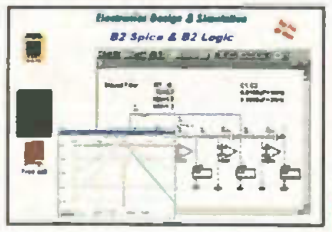

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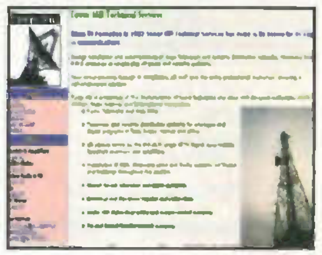

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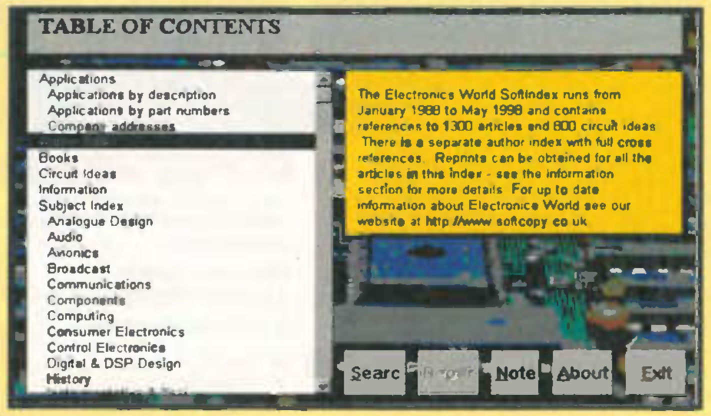

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AMPLIFIERS		
Amplifier Research 10W1000A 1GHz 10W Amplifier	3950	140
COMPONENT ANALYSERS		
HP 16097A Accessory Kit	1150	40
HP 4192A 5Hz-13MHz Impedance Analyser	4950	180
DATACOMMS		
Fluke DSP4000 Cat5/5e/6 Tester	2650	100
HP J3446C LAN Internet Advisor	8500	310
Microtest PentaScanner CAT 5 Cable Tester	1450	50
ELECTRICAL NOISE		
HP 8970B/020 2GHz Noise Meter	8950	330
HP 346B Noise Source	1250	50
ELECTRICAL TEST		
Megger FT6/12 Breakdown Tester (inc HV/LV Probes)	1250	65
Megger PAT1000S 240V Portable Appliance Tester	450	50
Megger PAT2000i 110 & 240V PAT	650	30
EMC		
Chase MN2050 15A Mains Network (LISN)	750	30
R & S EB 100 Miniport Receiver 20-1000MHz	3750	140
Schaffner NSG200E Mainframe	1350	50
Schaffner NSG222A Fast Transients Plug In	1150	40
Schaffner NSG225 Burst Simulator	1150	40
Schaffner NSG432 ESD Gun	2150	80
Schaffner NSG435 ESD Gun	3500	130
FREQUENCY COUNTERS		
EIP 578 26.5GHz Frequency Counter (Source Locking)	2950	110
HP 5342A 18GHz Frequency Counter	1850	100
HP 53181A RF Frequency Counter	895	65
Marconi CPM46 46GHz Counter Power Meter	4950	180
Racal 1992-55 1.3GHz Frequency Counter	1250	50
FUNCTION GENERATORS		
HP 33120A 15MHz Function Generator	950	50
HP 8116A 50MHz Function/Pulse Generator	2950	110
LOGIC ANALYSERS		
HP 1650A 80 Channel Logic Analyser	1150	45
HP 1663C 34 Channel Logic Analyser	2250	90
HP 16510A State/Timing Analysis Module	750	75
HP 16515A 1GHz Timing Card	750	75
HP 16500C Logic Analyser Mainframe	3750	140
NETWORK ANALYSERS		
Anritsu S331A-01 3.3GHz Scalar Network Analyser	5500	200
HP 3577A 200MHz Network Analyser	4950	185
HP 35677A S Parameter Test Set	1895	75
HP 85032B Calibration Kit	1150	40
HP 85032B/001 Calibration Kit	750	30
HP 85033D/K02 Precision Open/Short & Load	850	30
HP 85046A 3GHz S Parameter Test Set (50 Ohm)	3950	100
HP 8752A/003 3GHz Vector Network Analyser	7950	290
HP 8753A 3GHz Network Analyser	4500	160
HP 8753B/006 6GHz Vector Network Analyser	9950	340
HP 8753C 3GHz Vector Network Analyser	9500	340
OSCILLOSCOPES		
Fluke 97 50MHz 25MS/s 2 Channel Scope	750	30
HP 54111D 500MHz 2GS/s 2 Channel Digitising Scope	2950	110
HP 54501A 100MHz 10MS/s 4 Channel Digitising Scope	1350	55
HP 54502A 400MHz 400MS/s 2 Channel Digital Scope	1250	110
HP 54602B 150MHz 20MS/s 4 Channel Digital Scope	1750	90
HP 54603B 60MHz 20MS/s 2 Channel Digital Scope	1150	90
Tek 2465B 400MHz 4 Channel Analogue Scope	2500	110
Tek AM503/A6302/TM501 Current Probe Set	1750	60
Tek TDS320/14 100MHz 500MS/s 2 Channel Digitising Scope	1350	55
Tek TDS350 200MHz 1GS/s 2 Channel Digitising Scope	1850	70
Tek TDS380P/14 400MHz 2GS/s 2 Channel Digitising Scope	2750	100
POWER METERS		
HP 437B/002 RF Power Meter	1550	60
HP 438A Dual Channel Power Meter	2250	80
HP 8481A 18GHz Sensor	450	20
HP 84815A 18GHz Peak Power Sensor	950	30
HP 8990A Peak Power Analyser	9500	340
HP 8991A/004 Peak Power Analyser	8250	310
HP E4412A 18GHz Power Sensor	750	30
HP E4418A Single Channel Power Meter	1950	70
HP E4419A Dual Channel Power Meter	2850	100
Marconi 6960 RF Power Meter	750	30
POWER SUPPLIES		
HP 6011A 20V 120A DC PSU	1950	70
HP 66312A Dynamic Measurement Source 2A 20V	950	70
HP 6632A 20V 5A Power Supply	950	70
HP 6652A 20V 25A DC PSU	995	70
Hunting Hivolt Series 250 50kV, 5mA Power Supply	895	50
RF SWEEP GENERATORS		
HP 83620A/001/008 10MHz To 20GHz Synthesized Sweeper	16950	620
HP 8672A 2-18GHz Sweeper	4950	180
Marconi 6311 10MHz-20GHz Sweep Generator	4950	180
SIGNAL & SPECTRUM ANALYSERS		
Advantest R3361A 2.6GHz Spectrum Analyser inc T/Gen	6950	270
Advantest R4131D 3.5GHz Spectrum Analyser	4500	160
Advantest R9211A 100KHz Dual Channel FFT Analyser	3500	130
Anritsu MS2601B 2.2GHz Spectrum Analyser	5500	200
Anritsu MS2651B 3GHz Spectrum Analyser	4950	210
Anritsu MS2661A/1/2/3/4/5/6/7/9/11 3GHz Spectrum Analyser	5950	230
Anritsu MS2663A/1/2/3/5/7/9/11 9KHz-8GHz Spectrum Analyser	8950	330
Anritsu MS610B 2GHz Spectrum Analyser	2650	100
HP 4195A 500MHz Spectrum/Network Analyser	10500	385
HP 3561A 100KHz Dynamic Signal Analyser	3250	120
HP 3562A Dual Channel 100KHz Dynamic Signal Analyser	4250	300
HP 35660A 102.5KHz Dynamic Signal Analyser	3250	175
HP 70000 2.9GHz Spectrum Analyser System	9750	350
HP 85024A 3GHz Zero Loss Probe	1350	50
HP 8560A 2.9GHz Spectrum Analyser	9950	360
HP 8562A/001 22GHz Spectrum Analyser	15500	560
HP 8563A/026 26.5 GHz Spectrum Analyser	18500	670
HP 8566B 22GHz Spectrum Analyser	9950	365
HP 8568B 100Hz-1.5GHz Spectrum Analyser	4950	300
HP 8590B/021 1.8GHz Spectrum Analyser	2750	175
HP 8591A/001 1.8GHz Spectrum Analyser	5750	210
HP 8591E/004/041/101/102 1.8GHz Spectrum Analyser	7500	270
HP 8592B 22GHz Spectrum Analyser	9500	340
HP 8594E/004/041/101/105 2.9GHz Spectrum Analyser	7500	300
HP 8901A 1.3GHz Modulation Analyser	1250	50
HP 8903B/10/51 Audio Analyser with CCITT & 400HZ Filters	2750	100
HP 8903E Distortion Analyser	1450	85
Lindos LA100 Audio Test System	2500	100
Marconi 2380/2383 100Hz-4.2GHz Spectrum Analyser	5500	250
Tek 492P/01/03 21GHz Spectrum Analyser	4950	200
SIGNAL GENERATORS		
HP 83732B/1E1/1E5/1E8 10MHz-20GHz Synthesised Signal Generator	19500	700
HP 8642A/01 1GHz High Performance Signal Generator	2250	130
HP 8648B 2GHz Synthesised Signal Generator	3650	150
HP 8657A/001 1GHz Synthesised Signal Generator	1950	130
HP 8657B-001 2GHz Signal Generator	5500	200
HP 8683D 2.3-13GHz Signal Generator	5850	210
Marconi 2019A 1GHz Signal Generator	1000	100
Marconi 2031/001/002 2.7GHz Signal Generator	7250	260
R & S SMHU 4.3GHz Synthesised Signal Generator	13500	490
R & S SMY02 2GHz Synthesised Signal Generator	4850	170
SWITCHES & MULTIPLEXERS		
HP 44476A Microwave Switch	850	30
TELECOMS		
HP J2299B T1/ISDN Interface Module	1150	40
HP J2294C E1/ISDN SIM Balanced Interface Module	1150	40
HP J2300C WAN Internet Advisor	5950	210
HP J2912A OC-3c/STM-1 155Mb/s Interface Module	7500	270
HP J3446C LAN Internet Advisor (Fast Ethernet)	8950	320
Marconi 2840A 2Mb/s BERT Tester	1500	95
TTC FIREBERD 6000/7/B Communication Analyser	4950	180
W & G PFA-35 2MBPS BERT Tester	4500	170
TV & VIDEO		
Tektronix 1751 Waveform Vector Monitor	3950	150
Tektronix VM700A/1/11/20/30 PAL/NTSC Video Monitor	12500	450
WIRELESS		
HP 83220E/010 GSM/PCS/DCS Multiband Test Set	3500	130
HP 8922M/003/006/010 Multiband Test System	9500	340
HP 8922P/001/006 Radio Comms Test Set	11500	410
Marconi 2955B Radio Communications Test Set	3500	130
Marconi 2965 Radio Communications Test Set	6950	270
Racal 6103/001/002 GSM/DCS Test Set	7500	270
R & S CMS50-B33/B53/B55/B59 Radio Comms Test Set	3950	140
Stabilock 4015 1GHz Radio Comms Test Set	4500	160
Schlumberger 4031 1GHz Radio Comms Test Set	3500	130

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