Self on Class-G audio power

Differential RF probe
Ripple regulator revisited
EMI and shielding
Transformer tips
Programmable analogue chip

Circuit ideas:
Simple Q meter
3-phase sine waves
Logic probe
Stereo indicator
Simple VCO
Audio level indicator
Bike battery saver
Pushbutton latch
igniting the spark and fanning the flames

on a cold autumn afternoon, when I was five, I came upon a wicker cradle, killing myself. I'd been left alone in a forest school with a minor illness and was playing upstairs on my own.

There was a 24V electric box five feet away in my bedroom. By that age, I already knew that metals conduct electricity, but I didn't really understand the consequences of mains electricity. So as an experiment, I removed the fuse plug from the socket, wrapped a paddock chain around it, and pinned it back in.

If I close my eyes now, I can still see the dancing blue, herringbone flash, fringed with brilliant orange sparks.

Stil, I can hear the sound - a surprisingly soft, muffled noise - of a helium balloon bursting under a ceiling.

Apart from a few stern words, I was never punished for my "experiment." In retrospect my parents probably thought I had polluted myself enough. Any time, I suspect my father had a second-guessing for such a bold experiment (as a boy in Poland he had blown the cellar door off his mother making bread)

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, aiding a plethora of tools, car batteries, old of televisions, model steam engines, dynamos, valves, tuba time-cranked with nuts and bolts, bottles of mysterious and wonderful smelling chemicals, electric mowers, relays, bulbs, strange attractors flitted from scavenged aircraft, timber, sheet metal, hardboard, pipes, varnishes and Hol's catalog.

I would sit on the floor, playing with wires, while my brother sat by the old school desk, a washing machine motor and a fan belt. The thing was lethal and made a horrible noise but it cut a knife through butter. He taught me not to hide when I was five, and together we built my first short wave radio.

The radio influenced me in good stead during secondary school and university, when science got tougher and more and more maths creep in. Sometimes I knew, at least I thought I knew, 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.

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

In an informal telephone survey, there was a considerable degree of conformity among the replies. First, most students chose science because they liked it - not necessarily because it was important (myself, as a second). Second, the majority of 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 any individual's enthusiasm: almost all thought that science was well taught at school within the bounds of the national curriculum, not even 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 reverse applies equally to a misbehaving 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, early, and encouragement is given, his or her enthusiasm will buoy him up through the turbulent waters of formal education.

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

Many view science with suspicion, regarding scientists as thew, obsessed and even anemic. 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.bio-.
treaty.gov.uk/docs/2005/science_2006.html). This review, a consultation document, was commissioned "in response to concerns that inappropriate business in the UK sometimes find it difficult to recruit the skilled marketers they need."

In reply to this, the Institution of Electrical Engineers (IET) submitted a detailed reply to the Government's Science and Technology Committee's (SCT09) criticizing 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 reports show that the situation may have regressed.

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

There is no question that there is a vital role in addressing this issue, but so the parents, and probably more so. Maybe we need to take risks again, and show young people what science really could be in the most suggestive that we take a party of kids down to the local swimming pool, sit them in the water, and harass 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

"Do justice to the new millennium" Elektor 06/01

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Buckyballs act as superconductors at \(-156^\circ C\)

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 \(-221^\circ C\).

Bell Labs has raised the critical temperature to a balmy 117K, or \(-150^\circ C\). 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 Fedeleo 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 to 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 invisibly 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 Genesics to exploit the technology. Toys could be in the shops in two years, he said. The data-recording technique allows data up to 20bits 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 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.genesics.co.uk

**SALE!**

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 Exegem 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 problem 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, Exegem'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 Exegem. 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 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.

OSM requires about 400MIPS of processing, but full speed 3G will need over 1200MIPS. "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 Asian market as we know it won’t exist. We’re just starting 0.13um 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 42 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 digitizer under the paper and copies sketches to your Pains 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 ferociously in anti-clockwise circles. All settled down soon, and circling behaviour diminished to nothing after several exposures.

Medical problems scanners used in hospitals generally work between 0.3 and 1.5T, with some operating at 4T.

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 wave-lengths 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, depending 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."

Java for the masses

Croydon-based OneEighty Software has created a Java virtual machine (JVM) that can run on cheap 8-bit processors such as the venerable — yet popular — 8051.

"The system, 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 Dewig, 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 Russell Winder, the firm’s chief technology officer and 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 8051, ARM and 486. 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 Vikla.

"She is an artificial personality being jointly created by the University’s Robotics Department, the Human Computer Interaction Institute and Entertainment Technology Center. Vikla 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 Vovoice, the technology could be used, for example, to confirm that an offender has turned up for work, or is at home during curfew hours.

"Vovoice is already being used by Securcom 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.”

Vovoice 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.

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 200 miles it 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 2930, 2920, and 5920 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 16 314 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 (step 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 0.28Hz.

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

If these claims are correct, a few FPGAs executing a PFT can split a spectrum into narrow channels, with little hardware, at very high speed and in one go — with so 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 covert 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
**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 Samoff.

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 Ken Verhagen, technical director for device design at Samoff 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 B&V discharges for over 10,000 repetitions, Samoff claimed.

Test devices have been manufactured in 0.25, 0.18 and 0.13µm CMOS, while 0.35 and 0.1µm 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 PPI is using reconfigurable hardware design tools in its "intelligent pig". The pigs are autonomous inspection tools for pipelines that need processing capability of up to 640 MIPS.

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

"We were using VHDL entry tools but it was a struggle to maintain our 8-hour VHDL capability," said Gary Brayson, team leader for electronic design at PPI.

"Each inspection tool, including circuits, drive elements, data capture and storage, costs the company around £0.5m 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 Telemedecine, "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.

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 commercial 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 hi-fi 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 radio 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 radio 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 four 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 consequently 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.

If there are two main ways to arrange the dual-rail system, series and parallel, i.e. that 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-typed 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 once 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+CF. 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.

Musical signals spend most of their time at low levels, due to the high peak-to-mean ratio. Dissipation is greatly reduced by running off the lower ±V1 supply rails at these times. The inner stage +Tr1a operates essentially in normal Class-B. Transistors +Tr2 and the usual drivers and ±R1 then act as a shared emitter resistor.

The usual temperature-compensated Vbe 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 D2d. These are often called the commutating diodes, because of their role-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 +V1, +Di conducts, +Tr1a and +Tr turn on. Diode D2 turns off, so the entire output current is now drawn from the higher +V2 rail, with the voltage-drop and hence the dissipation shared between +Tr1a. Negative-going signals are handled in the same way. Figure 2 shows how the collectors of the inner power devices retract 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
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 sine-wave 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 a peak 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 voltage losses and so on.

In Class-G, the waveforms are a strong function of output level, requiring variable limits of integration... 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'être of Class-G is power saving. With this technology in particular, the waveforms used have 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 is given in reference 5.

Figure 3 is the triangular PDF PPD for conventional Class-B EF, while Figure 4 is that for Class-G with V1=30V and V2=15V, i.e. V1/V2=2/3. The PDF 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 area sums to the input power, the remainder being accounted for by the drivers and losses.

In Figure 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 switching threshold at 15 dB below maximum output. Total device dissipation at full output is reduced from 48% in Class-B to 40%. At first glance, this may not appear to be a good return for twice as many power transistors...

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 V1 voltage is likely to give lower general dissipation. I do not suggest that V1/V2=30% is necessarily the optimum lower-rail voltage for all situations, but it looks about right for domestic hi-fi.

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 too easy to design complicated output stages that work beautifully in simulation but prove impossible to stabilise at HP. Some of the interesting output-trip 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 Vbias,F are required to ensure TR4 turns on before TR7 turns off. This voltage is not critical, so long as it does not fall too low, or become too high. Fixed zener diodes of normal commercial tolerance are quite good enough.

If this bias is too low, the outer devices turn on late, then the V+/V− across TR7 falls too low, and its current capability declines; when evaluating this bear in mind the lowest impedance load you wish 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 doubts, 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 Zener 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+/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. 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 step of the diode reverse current. Frequency 10kHz.

Fig. 9. SPICE simulation shows variations in the incremental gain of an EF-type Class-G circuit. Class-G output stage. The gain-steps at transition – at 1.45V – are due to Early effect in the transistors. The Class-A trace is the top one, with Class-B optional below. For both, the inner power collector is connected to the switched inner rails, i.e., the inner power device collector, 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 zero 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 D3 must be power devices capable of handling a dozen amp 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 D3 rises above Vb, 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 Tr4 attempting as an emitter-follower to keep its emitter up to the right potential. So far so good. However, when the diode current ceases, Tr4 is still conducting heavily, due to its own charge-carrier storage. The extra current it turned on to feed D3 in reverse now goes through Tr4’s collector, which accepts it because of this transistor’s low Vbe and passes it onto the load via its emitter and emitter resistor.

This process is readily demonstrated by a SPICE com-

puter transient simulation. Fig. 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 Hinchei 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 not easy to eradicate this problem. Silicon 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 outer railings can cause Schottky diode’s relative lower reverse withstanding voltage. Fortunately, the diodes are only exposed at worst to the difference between Vb and Vbe, and this 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 of the commuting diodes are Schottky rectifiers. Motorola MBRS25L diodes capable of 50A and 250V 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 1.45V.

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 cur-

rent when the collector voltage increases, even though the Vbe remains constant.

When unexpected distortion intrudes into a SPICE simulation, betas 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 Tr4 begins to act, Tr5’s Vbe is no longer decreasing as the output moves positive, but substantially constant as the emitter of Tr5 moves upwards, at the same rate as the emitter of Tr4. 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 Vbe varies. The improvement in linearity is seen in Fig. 10, where the gain steps are halved in size.

References
Shielding and EMI

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 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 to that idea 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 b) 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 \( V_{OC} \) are shown: the subscript letter refers 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 many 'stray' capacitance represented: \( C_{AB}, C_{BD}, \) 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 redrawn version of the circuit shown in Fig. 1c.

Capacitors \( C_{AB} \) and \( C_{BD} \) form a capacitive voltage divider, with the output connected through \( C_{BD} \) 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 compensating ground plane. The common point may be at a non-zero voltage, but for the purposes of the input and output signals it is the zero-signal reference point. In most cases, the zero-signal reference 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 \( C_{BD} \) 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. 1e), 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 \( V_{AC} \) 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 impedance 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 shield.

Whatever the source, however, a difference of potential between points A and B gives rise to a spurious signal voltage \( V_{AC} \) that is effectively in series with the actual signal voltage \( V_{AC} \). Thus, the total signal seen as output is \( V_{AC} + V_{G} \). 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 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
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
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
1000kHz 

The effect of shielding absorptive

It is this current that we use to determine the AC resistance.

Shapers 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, δ. The skin depth is calculated from:

\[
\delta = \frac{2 \cdot 602k}{\sqrt{\mu_0}}
\]

Where: δ is the skin depth in inches, \( F_{XY} \) is the 
frequency in hertz, and constant k is 1.00 for copper, 1.234 for aluminium and 0.199 for steel.

Why is this important? In the case of absorptive loss, 
the attenuation is 8.76dB. For example, at 60Hz a steel shield has a skin depth of 0.86mm (0.034in). If 1.6mm 
(0.063in) skin is used, the total depth is equivalent to 
1.84δ, so the attenuation for magnetic fields would be 
8.76δ x 1.84 = 16dB.

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 (0.001in) 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 \( V_{th} \) skin 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 
recommeneded 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:

\[
R_{AC} = k R_{DC} \sqrt{f_m}
\]

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

Wire size (AWG) \( k \) factor
8 35
10 28
12 18
14 11
16 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 60μH/100ft.

A one foot run of wire will have an inductance about 
0.3μH. 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. 

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 
beaten into a "U" channel shape (see end view). The top
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 0.05A of the highest frequency involved will resolve 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 fitting (D) connector is a good example. Such a slot will allow radiation even with the connector fitted.

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

**Valve Radio and Audio Repair Handbook**

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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 FPAs 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 FPA, that looks set to revolutionise analogue system design. The AN10840 FPA 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 14mm² QFP, requires a single 3V±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 configuration of the FPA 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 50kHz, making it ideal for filtering, instrumentation and control applications in industrial, medical, automotive and low- and 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 FPA 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. This ability to handle in-service changes to configuration or even functionality brings unprecedented flexibility to the world of analogue design.

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 arrays 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 IC-equivalent networks to be implemented using just switches and capacitors.

Figure 2 contrasts a conventional analog-to-digital circuit with a switched-capacitor circuit, the charge that is transferred from node 1 to node 2 depends on the first approximation on the capacitor’s value and the switching duty cycle, effectively making the FPA an analogue sampled-data device. As a point of interest, Anadigm 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 AN10840 FPA also includes 13 buffered analogue I/O cells, two unconnected op-amps, four programmable clocks and a programmable reference voltage 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.

'Drag-and-drop' design ease

Designing an analogue system based on the AN10840 FPAA requires minimal circuit knowledge, analogue simulation skills or maths abilities. A free CAD tool known as AnalogDesigner enables the entire ‘design’, simulation and FPAA SRAM down-load 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-quadrature 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 library function. They then specify the desired offset correction, gain and low-pass frequency parameters, and interconnected the array cells. AnalogDesigner includes a

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 AnalogDesigner analogue function library.

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

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

Filter design made simple

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

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

Electronics engineers who want to evaluate FPAA functionality in the field — will probably choose to use the AN10DS40 FPAA development board, Fig. 4. This provides an AN10DS40 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 MHz oscillator and a regulated +5V supply. It also incorporates a Motorola 68HC008 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 time-detector circuit for telecoms applications.

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

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 6ZQ, UK. Tel (01270) 531999; fax +44 (0) 8142 446540

Exclusive reader offer:

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 evaluate 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 E&W 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 www.anadigm.com website.

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

Designed 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
This AN10DS40 board’s powerful HC08 microcontroller can dynamically modify FPAA functionality by reloading a new device configuration file — an operation taking just 100ms. 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 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 channel, providing major savings in both PCB space and cost.

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Alternatively, FPAA functionally 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.

Ordering details
Please send the coupon below, together with a cheque, to Anadigm Ltd, 3-5 Mallard Way, Crewe Business Park, Crewe CW1 6ZQ, UK. For queries or for corporate purchasing enquiries, please call +44 (0) 1270 531 990 or fax +44 (0) 1270 531 999
Even as early as 1904, society was seeing the benefits of wireless. Anthony Hopwood was looking at the first Cunard Daily Bulletin—a newspaper for transatlantic liner passengers that could not have existed, were it not for wireless communication.

Wireless across the waves

The 100th anniversary 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 photography on Sunday's issue records the distance from the Poldhu station at 220 miles at 1 PM. The news by Marconi was "detect to the ship" covering 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." Tabled interest was met by a story that Frank J Young, a well-known bookmaker and hotel owner, was fatally shot in New York 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 Petersen 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 signature, "Smoke-room gossip," it seems to be a resting place for old music hall jokes and shaggy dog stories. Here's one sample:

"A man bought a horse at a horse fair and found the dealer had short changed him by half a crown. He returned with him and was told — 'Now 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 Canadian 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 owners 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 democracy under railroad stations. 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 scarce today were a matter of pride, and published in the daily Bulletin. Apart from 38000 packages of fine manufactured goods and 1650 mail bags, there were several parcel of diamonds and sapphires worth about £10000, and $25000 worth of minted silver coin in 17 large cases. To deter opportunists thieves, "the precious stones and coin are securely stored in the ship's strong room."

On the 8th, Lord Laverydike, 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 8am to 9pm, the sea temperature rose from 58° F 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 gold" 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 ministeress. In a preface 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 domain."

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, scooped the pain of her funnel in a dash to rescue the survivors.

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

'Since thanks for your cordial wishes received' Wednesday 20th. Happy inform you Daily Bulletin entirely successful and greatly appreciated."

Even Marconi was watching call charges at 7/6 (38p) a word!
Differential-in
100MHz scope probe

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-referenced 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.

However this method has limitations:

- Both oscilloscope input amplifier channels and any probes used must be carefully matched for gain and phase across 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.

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-impedance measurements or for high-frequency but lower voltage measurements.

Both types can be extremely expensive.

This article describes the design and assembly of two low-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.

Design details

Each probe is designed to be held hand and is connected to the differential amplifier using a one metre length of RG179B/U coaxial cable.
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. 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-pk sine wave at 100kHz. Using my divide-by-10 Coline 25MHz 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. Oh 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 when the sine wave 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 discover 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 CMRR performance though, it can equally well be used as input to an AC voltmeter having 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 analog amplifier Devices. The AD830 provides an excellent 100kΩ CMRR at low frequencies, but by 100MHz this has reduced to 50kΩ.

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 MAX4414. It is designed as a high-speed, low-distortion, differential line receiver with a gain-of-two bandwidth of 130MHz. It provides 300kΩ CMRR at 100MHz, 50kΩ at 100MHz and more than 45dB at 100MHz.

I thought it would be useful to modify the AD830 to 300kΩ CMRR at 100MHz. I removed resistor R1 and put 12V±5V power supply 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 chose Maxim's MAX4410 - an ultra low noise 300MHz op amp. I decided to use it in a variation on the MAX4416 design that proved so popular in the previous example, low capacitance 100MHz active oscilloscope probe. The 4410 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 as a unity gain probe design.

Design - relays and trimmers

Choosing relays suitable for the gain switching proved quite difficult. As a final test, the relay must be physically small in order to mount in a 2 ½ inch 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 parasites 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 its performance, the parasitic capacitance and inductance were simply calculated and input into the design.

Calibration

Both probes should be assembled then pre-calibrated separately as stand-alone items. Solder a temporary BNC rotor surface-mount resistor from each probe output to its ground plane and take all output voltage measurements across this resistor. Apply ±5V power, take care to note the location of the +5 and -5V terminals on the PCB. Set probe capacitor C7 to mid-value and apply a known 1kHz signal not exceeding 10V to the test point 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 1kHz and energise the relay. Using a non-metallic trimmer tool, slowly adjust C7 to attain the same relay energised output voltage as noted for 1kHz. This is an extremely sensitive adjustment, tiny movements of C7 will affect output at 1MHz and above.

De-energise the relay and adjust C7 to attain the same de-energised relay output voltage as noted at 1kHz. These trimmer adjustments affect high-frequency performance and can reduce the CMRR at higher frequencies. Take care to match the 1kHz and 1MHz levels for each probe. Since C7 and C9 interact, repeat the above 1kHz 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 input. I used lengths of RG179/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" adaptors to each probe's test prod. Attach a "T" adapter to this test load to provide two equal-length BNC output paths. Attach one probe with its BNC adaptor to each output such that each probe 'sees' the identical test voltage and signal delay.

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 output signal. Carefully adjust V4 for best CMRR, minimising the differential amplifier output. This should be much less than 1mV for a 1kHz input.

Remove one probe's test prod from the signal source and attach a 50Ω terminator to this probe. With a 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. Energise the probe relays but not the differential amplifier relay. Note the voltage output from the differential amplifier. Increase frequency to 30MHz. Adjust C7 in the differential amplifier for the same voltage as noted at 1MHz. Reconnect both test probes to the "T" adapter and input a known 1kHz signal to both probes. Energise both probe relays and monitor the differential amplifier output. Calculate CMRR as:

\[ \text{CMRR} = \frac{2 \times |V_{\text{out}}|}{V_{\text{in}}} \]

If you can adequately measure this small output voltage, then very slowly and gently, slightly adjust C7 in probe 2 only, to maximise CMRR. This completes all necessary calibration.
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 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 simulators, including Microcap 6, provide the facility to override the internal constant value model using frequency-dependent expressions. However, as far as I am aware, suitable model libraries are not provided.

A restricted number of component models 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 \(^\text{10}\) now offers a Spice based data sheet for their Ceramic and Tantalum capacitors as a free download. This software produces 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 component. A series inductance, 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 programs - as has long been possible for many ICs.

The Spice-based, low-frequency simulator.

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

Some specialised CAD packages do much to facilitate high-frequency simulations and PCB layout, 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 both difficult and time-consuming.

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

Prototype PCBs for the probes My original thought was to use close tolerance capacitors and resistors, with just one trimmer capacitor to compensate for 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 as 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 obtain output voltages identical to those measured at 1kHz. However because of their interaction, the 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 C5.

Nevertheless 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 amplifiers.

My initial differential-amplifier PCB required more substantial modifications, such that the first board was eventually scrapped. Following careful in-circuit probing using the HP8404 vector voltmeter and repeated simulations, the layout was improved. Repeated PCBs 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 1nF 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. A mineral fibre strip was used to bridge top and bottom ground planes together. The design then worked well, but looked decidedly 'modified'.

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

To ensure consistent test signal voltage with frequency, the signal generator output was loaded with a 1Ω/Hz attenuator and a Vacom 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 5% capacitance, additional capacitance was added to the differential probe output terminal to produce an 18pF oscilloscope load. The 0.8% resistive load change, from the HP331A’s 1Ω/Hz to the HP8405A 10Ω/Hz, when parallel with the 90Ω/Hz output attenuator, was ignored.

My high-impedance R5V meter could also have been used. However, the HP331A and HP8405A both provide direct readings in decibels relative to 0dBm. Using these instruments avoids 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 flatness 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 100Ω/Hz 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 -10.5dB CMR.
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.

Insulated inserted to the finished probes using one metre of RG179/PU 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 1 mV on a 3 MHz bandwidth voltmeter. Noise output viewed on a 100 MHz oscilloscope however was much higher than desired.

Following further measurement and simulations, noise output was found to peak near 100 MHz. At frequencies below 50 MHz noise output was acceptably low.

Reducing noise

Careful measurements using my HP4195B vector voltmeter indicated an increasing voltage input to the differential amplifier with 75Ω input. Investigations indicated this rise was due to inductance contributed by the trimmer resistor VR. To replicate this peak, my simulations needed a much higher self inductance than I originally estimated for this component.

Two small 100Ω, 1% capacitors, C10 and C11, were soldered between the VR terminals and ground. They were 0.05Ω sized COG types. At 1000 Hz these provided a 0.5% reduction of signal input to the MAX1441 differential amplifier without compromising the CMR performance.

To further reduce high-frequency noise, I decided to roll off some output gain above 70 MHz. This was achieved by adding an 18pF COG capacitor, C12, in parallel with R10. The differential amplifier can now be used with oscilloscope input capacitances ranging from 15pF to 22pF without compensating adjustments for charging lead, Fig. 9.

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 probe required a considerable design effort and many repeated simulations to accurately achieve 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.

Assembling the differential amplifier PCB and its housing requires little comment. With the exception of the two three-leg 100mA stabilisers, pre-set resistor VR1, trimmer capacitor and relay, components used are designed for conventional surface mounting.

For pre-set resistor VR1, 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 176-333, was similarly used for the test probes.

The 1.5 pF 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, the larger 6.5 pF trimmer capacitor was a similarly modified Murata COG ceramic, Farnell part 108-222.

The very tiny 3-10pF trimmer was an AVX CT22 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 for 1pF. 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 tightly fit one half of the toothbrush case, and grounded to the PCB using a short flying lead.

The test probes 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 TPF3 insulators, intended as TPF3 stand offs, provided a tight fit into the brass tubing. These test probes on their own measured just 1.5 pF.

> PCB availability...

This design completes a series of four simple, self assembly, double sided PCB designs, targeted to assist high frequency measurement. 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 A46 to C. Bateman at Electronics World, Cumulus Business Media, Anne Boleyn House, 9-13 Ewell Road, Cheam, Surrey SM3 1BZ.
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.

Xia’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 circuit is designed as an 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 an 5V power supply, the MAX132 can provide a high resolution a-to-d conversion with a ±512mV to ±512mV 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, D7 to D0 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 ‘bi-polar’, 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 TLC7660.

A reference voltage for the MAX132 is provided by a diode D1 and dividers R1, R2 and R3. The input range of the a-to-d converter is from ±512mV to ±512mV. The measured data shown 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 shortened, 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 show its range 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.
```c
#include <stdio.h>
#include <conio.h>
#include <dos.h>
#include <bios.h>
#include <graphics.h>

#define CLOCK_LOW 0xFB
#define CLOCK_HIGH 0xF4
#define DIN_LOW 0xFB
#define DIN_HIGH 0xF4
#define CS_LOW 0x27
#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[10];
    data[0]=0;
    data[1]=0;
    data[2]=0;
    if (measure==0)
    {
        command[0]=0x82;
        outportb(out_port, command[0]-2);
    }
    else
    {
        command[0]=0x08;
        outportb(out_port, command[0]-8);
    }
    for (j=0; j<4; j++)
    {
        if (command[j] & test_bit)
            outportb(out_port, test_bit);
        test_bit=0x80;
        for (i=0; i<4; i++) /* start convert */
        {
            if (command[j] & test_bit)
                outportb(out_port, test_bit);
            test_bit=0x80;
        }
    }

    out=CS_LOW; /* CS low */
    outportb(out_port, delay1);
    test_bit=0x80;
    if (j==0)
    {
        outportb(out_port, out);
        out=CS_HIGH; /* clock high */
        outportb(out_port, delay1);
    }
    out=CS_LOW; /* clock low */
    outportb(out_port, delay1);
    if (j==0)
    {
        data[0]=data[1];
        data[1]=read_data[10][16];
    }
    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+101); /* show a bar on screen */
    setviewport(280, 240, 630, 310, 1);
    clearviewport();
    if (data_dis%12 || data_dis%12)
        sprintf(msg, "%d, %d", data_dis, data_dis);
    else
        sprintf(msg, "%.2f", data_dis);
    outtext(msg);
    return;
}

int init_graph(void) /* initialize graphic mode */
{
    int gdriver = DETECT, gmode, errorcode;
    if (gdriver & DETECT)
    {
        gmode = GRAPHICS
        if (gdriver & DETECT)
        {
            gmode = GRAPHICS
            if (gdriver & DETECT)
            {
                gmode = GRAPHICS
            }
        }
        errorcode = graphresult();
        if (errorcode != 0)
        {
            printf("Graphics error: %s.", get LOC(msg));
            exit(1);
        }
    }
    else
    {
        for (j=0; j<128; j++)
            outportb((out_port+j), out&=DIN_HIGH); /* set power supply */
    }
    return (errorcode & errorcode);
}

int init_screen(void) /* initialize display screen */
{
    /* load BIOS ROM drive, gmode, */
    errorcode = graphresult();
    if (errorcode != 0)
    {
        printf("Graphics error: %s.", get LOC(msg));
        exit(1);
    }
    else
    {
        for (j=0; j<128; j++)
            outportb((out_port+j), out&=DIN_HIGH); /* set power supply */
    }
    return (errorcode & errorcode);
}

void find_port(void) /* find printer port address */
{
    out=0;
    outportb((out_port+2), (WORD)(0x0040, 8);
    in_port=(out_port+1);
    outportb(out_port, delay1);
}

void clean_up(void) /* close graphic mode */
{
    closegraph();
    out=POWER_ON;
    outportb(out_port, delay1);
    return;
}

void main(void)
{
    /* load BIOS ROM drive, gmode, */
    errorcode = graphresult();
    if (errorcode != 0)
    {
        printf("Graphics error: %s.", get LOC(msg));
        exit(1);
    }
    else
    {
        for (j=0; j<128; j++)
            outportb((out_port+j), out&=DIN_HIGH); /* set power supply */
    }
    return;
}
```

More on the MAX123...

The MAX123 is a CMOS, 18-bit plus sign, serial-output, analog-to-digital converter. Multi-slope integration 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 (2μVLSB). 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 multipllexer or programmable-gain amplifier.

The MAX123 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
±0.005% FSR Accuracy at 16 convs
15μVRMS noise
Serial I/O interface
Programmed output for MPX and PGA
Performs up to 100 convs
22μA input current
50Hz/60Hz rejection

www.maxim-ic.com
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 windings and core. These have been bundled together akin to an equivalent total winding resistance \( R_w \) in series with the primary winding. The loss in \( R_w \) will be \( I^2 R_w \) where the total primary current \( I_p \) 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, things may be very different.

A transformer at switch on

Imagine the supply switch closed 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 worst scenario.

Now imagine the switch closed 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 e.m.f. 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 e.m.f. 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 in the first half cycle will be much greater than usual. This will result in a significant volt drop across \( R_p \), altering 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.

**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 240 V AC will see a peak voltage of 240 \times 2= 480 V. The volt-second product, with 50Hz mains, will be less than 339 \times 10ms by a factor 2 + e0.637, 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 no 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 mains transformers 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 \( R_p \) 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 \( L_d \).

Resistance \( R_p \) is a fiction, representing the heating in the core at mains voltage in the steady state. It is not a real component and is 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 \( R_p \) 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 \( \Phi \) and 'mmbf' 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.

The figure describes a transformer with a 2:1 step-down ratio, showing how the primary off-load current \( I_p \) consists of the vector sum of the magnetising current \( I_m \) and the core loss current \( I_k \).

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 \( E_m \). 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 to 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 realisation 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 whiskers

Vacuum leakage detector

Just occasionally, it is desirable to increase leakage reactance, rather than reduce it. The application I have in mind is 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-phase core and 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, 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, whistles 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 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 was 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-phase core and ordinary E and I laminations but the secondary is wound on the centre limb, and the secondary on one of the outer limbs, the other being unused.

to measure 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 full deflection, then plus the number of turns in the secondary 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 $I_q$ 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 I0A range in the legendary AVO meter, by shorting the centre E from a stack of E and I laminations, and threading the tap 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 $I_q$, 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 sections will reduce the leakage inductance by a factor of $\sqrt{n}$. 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.
NEW PRODUCTS
Please quote Electronics World when seeking further information

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 PlayStation 2 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 and realism of their 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 renderer 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: 0171 807 1602
www.analog.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 HMC502 has a sensitivity of 1mV/gauss, a field range up to 56 gauss and can operate on a supply as low as 1.8V.

Honeywell Tel: 0116 250 2600
www.honeywell.com

Socket 370 computer board for £240

Amplon Livevision 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 M36C680 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 768MB SDRAM and ECC support, a bus mastered IDE with Ultra DMA 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 on-board VGA controller supports CRT displays up to 1600x1200 pixels, and hardware monitoring for CPU voltage, temperature and fan speed. System temperature and fan speed can also be monitor.

Amplon
Tel: 01757 608331
www.amplon.co.uk

Surface-mount coax connector is lightweight

The latest surface-mount coaxial connector from Flipt Distribution is designed for 2.5mm high pin PCB-mounted connectors, and according to the supplier is amongst the lightest available. The F U.L series connectors, which are suitable for signals up to 50GHz, 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

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 10µF with the smallest capacitor, with dimensions of 6 by 9 by 13mm, to 47µF with dimensions of 20 by 35 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 inlined parallel wire leads in two standards, from 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: 0900 550500
www.epcos.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 10µF with the smallest capacitor, with dimensions of 6 by 9 by 13mm, to 47µF with dimensions of 20 by 35 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 inlined parallel wire leads in two standards, from 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: 0900 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.
NEWPRODUCTS

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 functionality 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 test monitors. With a built-in version of JTAG Technologies' test pod the XIOS plugs directly into a S100I style connection from the DataBlaster.

JAG Technologies
Tel: 01234 270228
www.jag.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, 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 transformers, dice, IC, thyristor and triac semiconductor protection. Other uses include surge protection in electronic products, and providing electronic discharge and noise spike suppression. Rely 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.eanab.co.uk

White LED driver IC

Toshiba’s latest LED driver IC in a miniature, six-pin SOT23 package delivers an output power of 325mW, allowing it to drive up to six white LEDs in series with a maximum 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, 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 TB6273FU incorporates an internal N-channel MOSFET with a typical on-resistance of 10mΩ, contributing to an overall device efficiency of up to 90% (pwm 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 TMS320F2810 DSP and the TMS320F2812 DSP are based on the firm’s code compatible TMS320C38x DSP core. This 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 micro code to be developed completely in C/C++-reducing development time. The F2812 DSPs integrate 128kB/256kB (kW) of flash memory and the F2810 DSP 64 kW for reprogramming during development and in field. 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 megaword for systems requiring a larger memory model. Samples of the TMS320F28x DSPs are scheduled for availability during the first quarter of 2002, with volume production scheduled to follow in the second quarter. The F2810

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Supplied either bare or pre-fitted with one of our Eze-Use, Serial interface boards with options such as software buzzer control, backlight control, operator interface, start-up message, big character generation, and PC-AT Keypad /mouse interface - all at very competitive pricing. We also stock Serial Interface Graphic modules and front panel bezels for all our LCDs. Please call (01977) 683665 or check our web site (www.milinst.com) for full details and pricing.

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South Milford
LEEDS LS25 9AQ
Tel 01977 683665
Fax 01977 461465
www.milinst.com
DSP will be offered in a 128-lead LQFP, and the P212 DSP in a 176-lead LQFP. Texas Instruments
Tel 01604 663899 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 3V 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 at 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 677576 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 1.3 and 1.7mm. The surface mount connector is on a 1.0mm pitch and located onto the PCB using mounting bosses. The DP260 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

Protocol tester handles latest Bluetooth specs
The PT6901 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 onto the market with the Bluetooth label. Providing test cases based on the SIG specifications makes it suitable for product development. Release 5.00 has test cases for the protocol layers handsharacle, 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 01262 811377 www.rohde-schwarz.com

LCD controller kit
An analogue interface controller kit from Digital View is preconfigured for VGA, SVGA or XGA panel. Based on an ACS-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 256 1112 www.digitview.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 600 DC - with capacitance values ranging from 1 to 10μF.
GWS
Tel 01978 656305

Single-supply op-amps for video markets
Maxim Integrated Products has introduced the Max383B to Max384B single, dual, triple and quad single-supply op-amps that are available in SOT23, μMAX, and TSOP packages. According to the supplier, the combinations 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 +7.4V supply or from dual ±2.5V supplies. They require 5.5nA quiescent supply current while achieving a 210MHz -3dB bandwidth and a 48V/μs slew rate. Their input common-mode range extends beyond the negative power

 TiePieScope HS801 PORTABLE MOST

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SPECTRUM ANALYZER-
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• The sophisticated cursor read outs have 21 possible read outs. Besides the usual read outs, like voltage and time, also quantities like rise time and frequency are displayed.

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• The HS801 has an 8 bit resolution and a maximum sampling speed of 100 MHz. The input range is 0.1 volt full scale to 80 volt full scale. The record length is 32K/64K samples. The AWG has a 10 bit resolution and a sample speed of 25 MHz. The HS801 is connected to the parallel printer port of a computer.

• The minimum system requirement is a PC with a 486 processor and 8 Mbyte RAM available. The software runs in Windows 3 or 95/98 or Windows NT and DOS 3.3 or higher.

• TiePie engineering (UK), 28 Stephenson Road, Industrial Estate, St. Ives, Cambridgeshire, PE17 4WJ, UK. Tel: 01480-460028; Fax: 01480-460340

• TiePie engineering (NL), Koperslagerstraat 37, 8601 WL SNEEK The Netherlands. Tel: +31 515 415 416; Fax: +31 515 418 819

Web: http://www.tiepie.nl
High-side current sensor monitor

In a five-pin package, the ZXT1010 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 sensor monitor IC from Zetex uses a single scaling resistor to convert the current into a ground referenced output voltage and provides high accuracy and low hysteresis. Taking less PCB space than alternative matched transistor pairs and offering an improved sensing accuracy of 1 per cent, the STO3-5 packaged current sensor suits portable battery powered equipment. The ZXT1010 operates over an input voltage range from 20V down to 2.5V. An enhanced version of the companion ZXT1000 device, the ZXT1010 high side current monitor IC, features a separate pin to ground that improves the typical output offset from 500µV to 30µV.

Zetex
Tel: 0161 622 4444 www.zetex.com

Multi-chip package for mobiles

Fujiitsu has announced a triple stacked multi-chip package (MCP), combining 64Mb NOR-type flash memory and 66Mb mobile fast cycle random access memory (FCRAM) with an asynchronous SRAM-type interface, and 4Mb 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 MB64V8RE3J1A 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
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Ian Sinclair’s Build Your Own books have established themselves as authoritative and highly practical guides for home PC users and advanced hobbyists alike. All aspects of building and upgrading a PC are covered, making this book the computer retailer don’t want you to read! By getting so grips with the world of PC hardware you can avoid the built-in obsolescence that seems to be part and parcel of the fast moving world of PCs, and escape the need to buy a new PC every year. You can also have a PC that keeps pace with the ever-increasing demands that new software applications place on your system.

The new edition of this book is based around building and upgrading to the latest systems such as Pentium III and dual processor Celeron motherboards running Windows 95/98 or Windows 2000. As well as guiding you round the inside of your CPU Ian Sinclair also covers monitors, printers, high capacity disk and tape systems, DVD drives, parallel port accessories...

CONTENTS: Preface; Preliminaries, fundamentals and buying guide; Case, motherboard and keyboard; About disk drives, Monitors, standards and graphics cards, Ports, Setting up, Upgrading, Modems and other connections, Windows, Printers and modems, Getting more, Index.
NEWPRODUCTS

Please quote Electronics World when seeking further information.

4ns dual comparators
Elantec has introduced high-speed comparators. The EL5281C and EL5282C 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 40Gb/s.
The process is built around indium phosphate (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 10Gb/s systems that use RZ encoded data. Succeeding generations will provide ICs with up to 100Gb/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. Please visit www.vitesse.com.

Direct stream digital-audio support
Wolfson Microelectronics has introduced a stereo d-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-a to converter integrates a DSP interface and the industry audio industry audio formats, which let the user connect multiple d-a to converters to a single time-multiplexed DSP interface. It provides a hardware or software interface for audio control. Data rates and 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 6679286
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 CPU/Px can be used as the main CPU board in a VME system. The processor has 32byte of level one cache and 256byte level two cache. The board uses the E2440BX chipset, which supports the 100MHz front side bus. A heatink is fitted to the CPU so no fan is used on the board. Up to 1GBYTE of DRAM can be installed, 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. EDIIE 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 762626
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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 to the mid-1960s, I’m sure that there must have been earlier realizations, 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.**
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 that govern the turn-on and turn-off times:

\[ t_{\text{on}} = \frac{2L + L}{V_{\text{in}}} \]

\[ t_{\text{off}} = \frac{2L + L}{V_{\text{in}}} \]

\[ f = \frac{1}{t_{\text{on}} + t_{\text{off}}} \]

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 follow:

\[ V_{\text{ripple}} = V_{\text{in}} - V_{\text{in}} \left( \frac{R}{R + R} \right) \]

\[ V_{\text{ripple}} = V_{\text{in}} - V_{\text{in}} \left( \frac{R}{R + R} \right) \]

\[ V_{\text{ripple}} = V_{\text{in}} - V_{\text{in}} \left( \frac{R}{R + R} \right) \]

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

\[ V_{\text{ripple}} = V_{\text{in}} \left( \frac{R}{R + R} \right) \]

which means that the ripple voltage increases with increasing input voltages. This pitfall severely constrains the operating voltage range, which negates some 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 \(C_f\) 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. 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 over-size 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 applications. One of its sections is built around a hysteretic regulator. One of its touted advantages is a fast transient response. For many years, on semiconductor technology has offered another ripple regulator with a different control scheme (fixed on-time, variable off-time) which allows extreme simplicity. 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-p-n 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 stored 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.
ESSENTIALLY, this means that the non-inverting input will see the output capacitor's ripple voltage plus the voltage drop across sampling resistor \( R_s \) due to the inductor current swings.

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

\[
V_{\text{in+}} = (\Delta v_{\text{cap}} + R_s \Delta i_L + \Delta v_{\text{in+}})
\]

(8)

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

On closer inspection, it becomes obvious that the capacitor voltage ripple, \( \Delta v_{\text{cap}} \), 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 waveform developed by the inductor current swing through the sense resistor. You can equate (7) and (8) to obtain:

\[
\Delta v_{\text{cap}} \times R_s = \Delta v_{\text{in+}} \times R_s
\]

(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 maintained constant.

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

\[
T = \frac{V_{\text{ref}}}{I_{\text{ref}}}
\]

(10)

Since the output voltage and the inductor ripple current are regulated, and assuming a constant inductance value, the frequency remains fixed. It is independent of the input voltage and most importantly, from the output capacitor's capacitance or ESR values. This circuit is stable independent of how it was designed, and if all, its transfer function remains the same:

\[
V_{\text{out}} = \frac{V_{\text{ref}}}{I_{\text{ref}}} + \frac{V_{\text{in-}}}{I_{\text{ref}}} + A
\]

(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, therefore, this matches 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 influence respecting 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 80kHz.

For starters, the \( \mu \)A723 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 earlier switchmode 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 switch design has 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 allow for the use of n-channel devices. The difference in voltage versus hole channel mobility makes the n-channel devices more efficient for a given size. Fortunately, power Mosfets that are designed to drive high-side n-channel devices have become widely available.

- Schottky free-wheeling diodes were perfectly acceptable for Mosfets. Although the efficiency penalty for lower voltage supplies becomes excessive. Synchronous rectification is now the norm. Synchronous rectifiers, recognising the trend, produce combo driver ICs that satisfy the requirements for both the high side and synchronous rectifier n-channel Mosfets. Linear Technology's LT1660 device was selected since it provides fast-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 todays power supplies, to prevent shorting through conduction that may cause device failure.

- The heart of the circuit, the \( \mu \)A723 regulator itself, was state of the art 30 years ago but nowadays it leaves much to be desired. The internal comparator is extremely slow and has common-mode range limitations. Its voltage reference has a wide tolerance, and the output transistor is not suited to drive Mosfets. The comparator 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 to 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 3 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 amplifiers have the 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! 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.
Practical circuit implementation

The completed prototype circuit is shown in Fig. 5. A stable voltage reference is provided via \( V_{ref} \), which is then fed to a resistive voltage divider string \( R_3 \), \( R_4 \) and \( R_5 \). Transistor \( Q_2 \) shorts out resistor \( R_5 \) whenever its gate is driven high by \( U_3 \)'s output. Since this will in effect raise 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 \( R_3 \) and AC feedback via \( C_2 \). Feedback at DC provides the average DC output value. The AC feedback is a fraction of the output inductor's current swing as previously discussed.

To maintain low losses, current sampling resistor \( R_9 \) 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 \( Q_1 \), its collector will pull down the reference voltage.

Comparator \( U_2 \) 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 \( Q_2 \) with \( D \), assist in the pull-up current, allowing a much specified transition. Transistor \( Q_2 \) inverts the comparator output. This complementary pull-ups are provided, necessary for the synchronous rectifier operation.

These pulses are applied to \( Q_6 \) which is a Linear Technologies half-bridge driver. This is an interesting device, which offers the following features:

- TTL threshold-compatible inputs, which are guaranteed to be pulled all the way up to the \( V_{CC} \) supply voltage without damage.
- Like most modern drivers, a undocumented topology is employed. This works together with \( D_2 \) and \( C_6 \) and allows the use of n-channel Mosfets for the high-side power switch.
- Most important in a half-bridge configuration in the prevention of shoot-through currents. At best, these could cause an efficiency penalty and at worse 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 \( Q_6 \). The bottom output goes to \( Q_6 \) which works as the synchronous rectifier.

Bias

The AC feedback is a fraction of the output inductor's current swing as previously discussed.

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CIRCUIT IDEAS

Q-meter signal generator add-on

While developing RF circuits in the LF to RF range, a need arose for equipment to indicate the Q of resonant circuits, and the facility for trimming inductions 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 divide; D1 – the small cutout is justified when compared with a commercial Q-meter.

Transistors T1 and T2 are required to drive the test circuit L1 and V1. The 560Ω or 752Ω modulated RF output of a signal generator is too high an impedance to series resonate the LC circuit. Diode D1 extracts AF modulation to drive the meter M1. This terminals L1, resonate the circuit by adjusting the frequency of the generator or C1 for maximum indication. Transfer the probe to the anode of D1. Measure the voltage in the SET and Q positions of SW.

1. Adjust the generator RF output for a reading of 20mV in SET and set RV1 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 RV1 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 PSD and scales could be used, e.g. a meter scaled 0-10 or 0-30, could provide Q ranges of 0-10, 0-30, or 0-100 by selecting the appropriate SET indication on the scale.

Adjustment of the signal generator frequency and output should always be carried out before reassembling 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 PSD of M1, R1 may be required and the value of R2 may need to be modified.

D W Dennis Brown Southampton F72

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.

C Stanforh

Winney Oxfordshire F83

£50 winner

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 selector 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 selected output at the front.

It is possible to use solid-state switches, 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 D3 and D4. This feedback network maintains the selected input after the push-button is released. Diodes D5 and D6 will disable IC2 when an input is selected with IC1. Diodes D7 and D8 are connected to IC1 and IC2. When IC1 is selected, D7 and D8 are used, while D5 and D6 are pressed, only the unused output DI is enabled.

An extra input is provided to disable the outputs during on and off switching of the amplifier. During and after, switch-on or output is selected until one of the push-buttons is pressed. On power down, the inputs are switched off by the enable signal.

To drive the input selecting relays a ULN2030 driver IC is used, which also drives the LEDs that display the selected input.

Bernard Van den Abeele Evergem Belgium F78

£100 winner

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, 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 as it is 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: discs with separate drawing and text files in a popular form are best – but please label the disc clearly.

Send your ideas to: Jackie Lowe, Cumulus Business Media, Anne Boleyn House, G-13 Ewell Road, Cheam, Surrey SM3 8BT.
Audio level indicator

In the February 2001 issue, there was an interesting circuit for audio level indication using a tri-colour LED by Graham Booth. However, I wanted a version that would operate from ±5V 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.

Michael Cox, CEng, FIEE
Twickenham
FB1

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, IC2, is a binary counter which is loaded with 2 by means of IC3a when the output reaches 14, so the output Q3 is symmetrical around 8 (6 count pulses high and low). Fig. 2.

Second counter IC4, also binary, is loaded with 2 when the first counter reaches 6; this could be done by ANDing Q6, Q7 and inverted Q1 of IC4. That would require an extra inverter, so a variant is used; namely Q4 and Q7 of IC2 and Q2 of IC3 are subjected to an AND function. After maximum two count cycles the desired loading of IC4 is acquired.

Outputs Q2 of IC2 and Q3 of IC3 are fed to two fourth-order filters, IC6 and IC7, whose outputs are sinusoidal and form phases R and S. Phases R and S are added and inverted in IC8 to obtain phase T.

The low-pass filters are clocked by a frequency 20 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 ±3V pk-pk and the amount of clock ripple in the output is 40mV pk-pk; the frequency-range is 0.10kHz—20kHz.

W F C Dijkstra
Waalre
The Netherlands
FB6

Fig. 1. Digitally-generated three-phase square waves are converted to sine forms 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 colour-coded 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 – HDSPIEEE 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’ respectively. Note that since the 1 and 0 segments of each LED display require to be illuminated for each of the three output conditions, they are connected through current-limiting resistor Rb to ground. They remain on continuously.

Resistors R1, R2 and R3 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 Rb. It will also sink current through segments a, b, c and d of the right display via resistor R7. 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 a, b, c and d of the left LED display and resistor R6. 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’.

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

Frank Kelly
Sterling
F79

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
FR2

Stereo Indicator

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
FB0

Display for stereo audio indicators mono and stereo in 3dB steps.
TELFORD ELECTRONICS
Old Officers Mess, Hoo Farm, Humbers Lane, Horton, Telford, Shropshire TF6 6DJ, UK
Tel: (0944) 01952 605611 / 01952 670778
Fax: (0944) 01952 677997
E-mail: telfordelectronics@btinternet.com / marc.007@btinternet.com / annie.007@btinternet.com
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Ten year index: new update

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## TEST EQUIPMENT SOLUTIONS

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<td>HP 66320A Dynamic Measurement Source 2A 20V</td>
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