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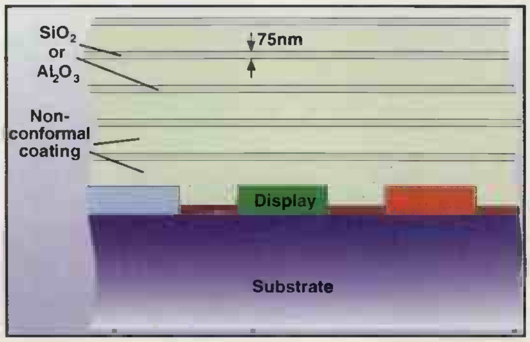
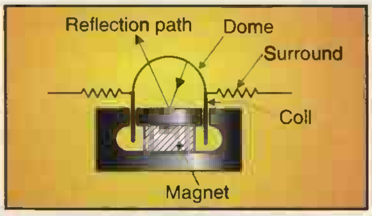


Organic LEDs that can be printed onto a flexible polymer layer have many uses, particularly if they can be produced in matrix form. Steve Bush reports on the current state-of-the-art on page 86.

Cover photography Mark Swallow

Unusually, David Ponting's accurate milliohm meter – shown here measuring a 50mΩ 1% resistor – does not involve driving a high current through the resistance. Turn to page 114 to find out how it works.



If you've just invested in an expensive hi-fi incorporating dome loudspeakers, please avoid page 130.

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No laughing matter

Lately this magazine's leader columns have taken a depressing and even hectoring turn pointing out the inadequacies in the way that UK plc and Britain.com are managed. Far be for me to break the trend, although this month's sermon (or is it harangue?) does at least target a different failing.

First of all, though, a word of praise. Technologists – the kind of people who read this magazine and turn electronic concepts in to reality – can take a bow. They are the people who consistently come up with new ideas and deliver working products and services, often within budget and timescale. Well done!

It's funny then – in an un-funny way – that many business leaders are far less adept in exploiting those developments, particularly in the field of electronic business. There's overwhelming evidence that e-business is set to revolutionise supply chains as online market-makers begin to demonstrate savings arising from using the Internet to make purchases. Survival, profitability and opportunities for future cost savings and growth all depend on its adoption, with trading on the Internet delivering greater market penetration, increased responsiveness to customers, more flexibility and lower costs.

It's a concept already adopted by businesses in the USA and a message that British business would do well to heed, according to the Engineering Council, which issued a rousing wake-up call to UK plc in December.

Many valuable insights arose from this top-level forum, in particular that UK businesses would do well to accept, as the Americans do, that change is the only constant in modern business. The moral is to embrace this 'change culture' with the enthusiasm and vigour of US businesses.

It is better to take the risk and invest in transacting either directly with customers or through e-marketplaces, than not to take the risk and exclude any chance of business-to-business success.

While uptake of e-business offers many benefits, it also entails significant challenges. Many of these challenges are a mixture of technological, cultural and legal issues, all of which should be seen in the context of the business they are designed to support. Serving that business and its customers must be the key objective.

The biggest threat of e-business will be that arising from inaction – being left behind – when given the opportunity to take part in what may become a key e-market place for a relevant industry sector.

Many of those who best understand the speed and impact of e-business are young. Many of those who make the strategic decision to invest in exploiting e-business are experienced and probably not so young. This is one issue in which the experienced could benefit from being mentored by the young.

Engineers with change-management skills are well placed to manage programmes to develop and implement e-business systems.

Crucial to success in e-business is the customer-supplier interface. Much more effort must be made here. "Customers are overwhelmed and confused about adopting e-business for their needs," declared David Smith of ICL to forum delegates. "E-business solutions are not a straight replacement for existing systems and processes, although many purport to be that. If e-business solutions are to be successful then traditional methods and processes need to be forsaken to allow a new operation to emerge."

Considerable improvement is needed also for websites, often used as the portal to e-business activity. Simple is best here, even though far too many websites exist purely as a monument to the extreme virtuosity of their designers or the corporate vanity of the organisations that commissioned them.

"There are 20 million sites on the Web and nearly all of them are awful," declared American website design and usability guru Jakob Nielsen, on a visit to London a week previously. "When people design websites, they rely too much on their own perceptions of style and aesthetics, without giving enough thought to how the average user, dialling up on a slow connection and lacking the sophisticated aesthetic sensibilities of a design professional, will explore the site. As a result," he continues, "users lose interest trying to fight their way through the artistic graphics to find what they really want and immediately develop a jaundiced perception of sites that fail to come up with the goods.

"If a site has not delivered the desired information within a few seconds, one click and they're off to a competitor – meaning lost profits to the website owner."

Those who bear no responsibility for these shortcomings may smirk, but it's no laughing matter. They may argue too that their motivation for reading this magazine is to learn about new technologies in electronics and how to apply them – not to be harangued on the incompetence of their marketing colleagues.

All the same, the issue is highly relevant, since technologists will starve if their employers fail to sell the products they create. For each and every person who earns a living from electronics, the writing is on the wall: embrace e-business or be left behind at the mercy of your competitors. ■

Andrew Emmerson

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Pan European R&D initiative:
UK firms excluded because of DTI rules

British electronics firms are to be cut out of a new pan-European research initiative because of rules imposed by the UK Department of Trade and Industry (DTI). Exclusion will affect UK companies' competitiveness.

MEDEA (Microelectronics for European Development and Applications) has been extended for another eight years under the title 'MEDEA+'. MEDEA+ was accepted as a Eureka project in June, and approved its first 18 R&D projects towards the end of 2000.

More projects in the ¤4bn programme are expected to be approved throughout December.

"The UK is not participating in MEDEA+," said MEDEA chairman Dr Jürgen Knorr, "but it is considering the situation and maybe they'll come to a decision to join later on."

Participation by UK companies is restricted under DTI rules which totally ban big companies from joining, and make it onerous for small companies to participate. The DTI's attitude persists, despite the fact that one of the biggest recent UK technology successes – Bookham Technology – got much of its early funding from a European research project.

In the original MEDEA, a total of 9400 man-years of research was completed. The French accounted for 3512 man-years, the Dutch for 1839, the Germans for 1614, the Belgians for 1056, the Italians for 1111, the Finns for 89, the Swedes for 31, and the British for 16.

MEDEA has had some notable past successes such as GSM, xDSL, automotive and smartcard technologies in all of which European companies are world leaders.

Companies in the United Kingdom complain that by being excluded, they get to exploit the results of R&D later than companies in mainland Europe.

Telecoms operators are avoiding xDSL technology

Telecoms network operators are finding they prefer to install the 15-year-old ISDN technology for upgrading telephone lines, rather than the latest, and much more powerful, xDSL technology.

"ISDN is how the telcos can stabilise revenues," said Christian Wolff, v-p and general manager of Infineon Technologies' carrier access division, speaking at Electronica 2000 last November .

"ISDN provides two channels, and allows voice calls. It is very easy to install and subscribers get two numbers," he added.

This means that telecoms operators or telcos can still get lucrative, time-monitored voice revenues rather than having those wiped out by xDSL technology which allows voice over Internet Protocol (VoIP) calls which are most conveniently paid for by a flat fee for 'always on', broadband Internet connection.

"ADSL has the flat-fee problem. ISDN is more strategic for telcos," said Wolff. So attractive do US telcos find the ISDN alternative, that they are marketing it over there as 'iDSL'.

Wolff is finding particularly lucrative markets for ISDN chip-sets in China. "China installs 30 million new analogue lines a year. ISDN is a solid business which fills the fabs and creates a stable revenue flow."

Nonetheless, Wolff does not believe that telcos can resist broadband xDSL for ever. "I can't imagine the network will stay analogue, it has to be upgraded to digital. If the telcos don't install broadband then cable modem will take over and they'll lose out anyway."

Wolff added: "Infineon's xDSL strategy is to solve the deployment issues so people can install DSLAM's (DSL access modules) more quickly. We will introduce such products early next year."

Infineon's new DSLAMs will be 'multi service access platforms' which include ADSL and the existing telephone network, known as POTS, in one box and so reduce the number of connections.

"The POTS splitter is included in the chip-set," said Wolff, "and ADSL is integrated with it, it's just like installing a new analogue line. The vision is to enable mass deployment."

David Manners

Touch-screen feature added to flat loudspeakers

Flat speaker specialist NXT is adding a touch-screen capability to its loudspeakers.

Called TouchSound, it fits in front of an LCD or any other flat screen, just like SoundVu, the firm's loudspeaker-in-a-screen system.

"SoundVu has little exciters to resonate the panel. If we also feed an ultrasonic signal through these exciters we know exactly what to expect at the receiver," said a spokesperson for the company.

When a finger or pen presses on the screen, it disrupts the complex pattern of resonant modes on the screen. The sound at the receiver changes, and NXT has figured how to calculate the position.

But does it affect the sound when the screen is pressed?

"In theory yes, in practice no. You can't even hear any difference," said the spokesperson. He claimed the resolution and accuracy are as good as other touchscreen technologies using capacitive or resistive techniques.

Path cleared for digital AM broadcast

The way is clear for digital radio to move onto the AM wavebands following the ITU's adoption of a worldwide standard on digital sound broadcasting.

The standard adopted was submitted by the DRM (Digital Radio Mondiale) consortium which has been working towards digital broadcasting in the AM bands below 30MHz since 1998.

Digital AM will operate alongside the Eureka 147 digital audio broadcasting (DAB) standard but will be unable to offer the data services of DAB. This should lead to cheaper sets.

"With digital AM, you're restricted in bandwidth and you're not going to be able to do much more than convert what you've got to digital," said Peter Florence, managing director of RadioScape. "But it does take it into the digital age so you should get improvements in audio quality which will be significant."

Software house RadioScape has been tracking the progress of digital AM and has already carried out preliminary work on it.

The final digital AM standard is expected to be available towards the end of 2001. Pilot transmissions will start in 2002 with the official launch in 2003.

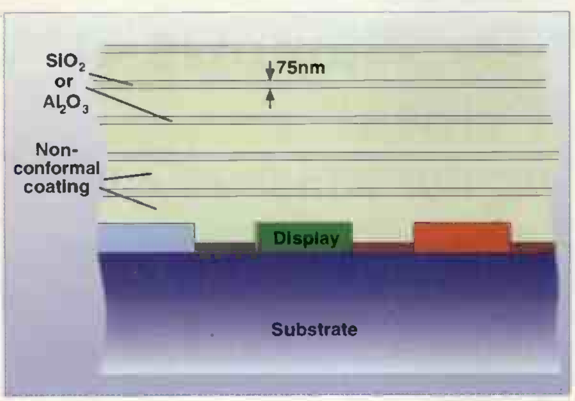
Organic LEDs on flexible plastic substrates: possible, but not easy

The Society for Information Display (SID) 2000 UK conference was held at the Electronic Information Displays show held at London's Docklands in November 2000.

Mike Weaver of US-based Universal Display Corporation spoke on the state-of-the-art for flexible organic LED displays (OLEDS).

In a down-to-earth presentation, he described the solutions the company has found to some of the problems of producing workable OLEDs on plastic films.

Five layers of high density dielectric could be the key to long life for organic LED displays.



Display lifetime is a big issue with OLED material, not only from the inherent emissive lifetime of the materials, but from poisoning.

Unfortunately, the chief poisons are water and oxygen. Apparently any exposure makes short work of destroying the OLED.

Glass and metals effectively block water and oxygen, but plastic substrates, "look like a sieve to water", according to Weaver.

To get over water and oxygen ingress without compromising flexibility, the company has developed a coating called Barix.

This consists of a series of layers of a dense transparent dielectric – aluminium and silicon oxides were given as examples – separated by surface levelling layers of non-conformal coating.

To keep the dielectric layers free from leaky pin-holes and grain boundaries, thickness is restricted to less than 70nm per layer.

This means around five layers to keep out moisture and air for the 10 000 hour planned operating life of the displays – 5000 hours is the average life of a car apparently.

Existing measurement equipment is an order of magnitude too insensitive to measure the permeability of the Barix structure, which Weaver estimates to be under 10^{-6}g/m^2/day.

"The best way to measure permeability at the moment to to build the display," said Weaver. Plastic has a permeability of up to 1g/m^2/day.

Current life of the company's prototype flexible OLEDs, measured by accelerated life testing, is 1000 hours. This is limited, he claims, by the display material rather than ingress as far as anyone can tell. "By mid-2001 we may be up there, to 10 000 hours," he said.

The Holy Grail of OLED research is an active matrix display on plastic, and constructing the matrix is causing headaches. Basically the substrate is damaged at 70°C and polysilicon is processed at 400°C.

There are two approaches to a solution according to Weaver. The first is an alternative plastic. A material called PEE is being worked on by the plastics industry which could be processed at up to 500°C. Meanwhile, work by Lucent and Philips may produce organic rather than polysilicon transistors which will be processed at low temperature.

Whatever the substrate, conductor technology now seems available to get power to the OLEDs, which are low-voltage, high-current devices.

Steve Bush

'Baird TV' might catch on yet

Remarkably, a demonstration of this 70 year-old technology may show a way forward for mobile videophones. The Narrow Bandwidth Television Association, a historian society, was showing a 32-line vertically-scanned 3:2 'door' format head and shoulders movie of someone talking. The 48 x 32 pixel moving monochrome image was perfectly recognisable and only requires a 12.5kbit/s uncompressed data stream. The image is clear, apparently, because the human brain is exceptionally good at filling in the missing data when faces are transmitted. No other images were shown on this display, but its maker said that perceived quality suffers when the subject is not a face.

LCD backround enhancer

DuPont has developed a holographic reflector material that increases the visual impact of standard 'watch-type' black-on white reflective LCDs.

The film is self-adhesive and replaces the reflective metal diffuser normally employed.

Displays with this backing have a 'brighter' more eye-catching background than those with a metal diffuser and either a blue, gold, green or red sheen – depending on the film chosen.

It works by 'bunching' reflected ambient light at approximately 90° to the display.

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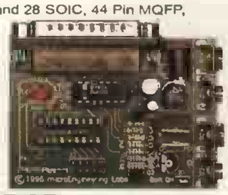

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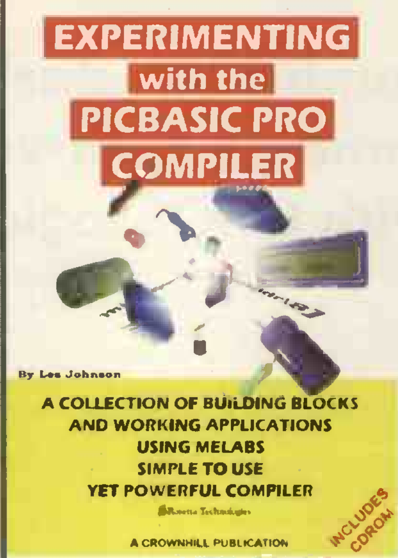

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CIRCLE NO. 107 ON REPLY CARD

Cambridge start-up finds a way to combine high-voltage switching circuits with CMOS logic

A spin-off from Cambridge University claims to have developed a technique for integrating high-voltage power-switching circuits alongside standard CMOS logic.

Called Cambridge Semiconductor, the start-up plans to begin prototype production of its invention. This invention is the result of ten years research in the University's Department of Engineering.

Combining logic and high-voltage switching circuits is one of the best routes to reducing power consumption of consumer goods.

"Power ICs will allow us to efficiently integrate power electronics into a standard CMOS process," said Professor Gehan Amaratunga, who heads up the new company.

The circuits can cope with voltages up to 600 or 700V, "which is what is required for mains connected power management", Amaratunga said. Able to handle currents of up to 1.5A, the devices are aimed at goods under 1kW.

These sorts of levels have only been achieved by one or perhaps two other companies, including Hitachi in Japan.

Typical circuits implemented by the chips would be mains AC-to-DC conversion, control circuitry, followed by efficient inversion back to AC to drive a motor.

"The whole thrust is to get away from the single phase AC motor," Amaratunga said.

The firm reckons any device supplied with mains electricity could benefit from the technology – including PCs and vacuum cleaners. Today's vacuum cleaners for example are around 50 per cent efficient, but they could be made more than 90 per cent efficient using power ICs.

Cambridge Semiconductor has received a £250 000 investment from the University's Challenge Fund.

***Richard Ball** Electronics Weekly*

Space-borne error detection lands in French design house

Single event upsets (SEUs) and soft errors have been a feature of space-borne electronics for many years. However, we are reaching the point where SEUs – which can cause catastrophic failure of a system – are appearing in earth-bound systems, due to decreasing feature sizes.

iRoC Technologies, a design house based in Grenoble, France, has developed a technique for adding fault tolerant circuitry which it has applied to a 32-bit Risc processor.

Most SEUs are caused by cosmic radiation – neutrons – flipping a bit inside a circuit by raising the local voltage above the threshold for a high. Other soft errors include temporary logic delays causing glitches and races.

iRoC reckons that three-quarters of SEUs are the result of reduced V_{dd} and noise margin – consequences of smaller geometries. At 0.13μm, says the firm, the soft error rate is one every few hours, while at 0.10μm it's as many as one every couple of minutes.

While memory can be checked through standard error-correction schemes, logic has no such fallback.

The aerospace solution of replicating logic several times over is just not suitable for modern electronics, especially in our price sensitive consumer world.

iRoC's technology is called a 'transient fault tolerant architecture', a collection of circuits distributed across an IC. It is implemented at the register transfer level of design.

"The extra circuitry on the logic detects the transient," explained Dr Michael Nicolaidis, chief technical officer at iRoC.

He claims the circuits are able to protect almost all logic and memory, with over 98 per cent coverage of the chip. Performance is unaffected and the whole process fits within a standard cell approach to design.

According to Nicolaidis, the extra circuitry adds around 17 per cent to the die size, a significantly better figure than duplicating or even triplicating logic.

To prove its point, iRoC has built the ROC S81, a 32-bit processor based on the Sparc architecture.

It is manufactured by STMicroelectronics using a 0.25μm process and runs at 100MHz. It is claimed to be immune to SEUs.

The firm also has a collection of EDA tools, including Shielder which enables insertion of the test circuits. It fits into a standard chip design flow.■

Richard Ball

750GHz logic uses niobium

Logic operating speeds of more than 750GHz have already been achieved experimentally in superconducting niobium, according to a report in the IEEE's Spectrum magazine. Rapid single flux quantum (RSFQ) logic, as it is called, works with the presence or absence of a magnetic flux quanta that represents information bits. The report predicts the feasibility of commercial 100GHz superconducting ICs within the next five years.

"We are creating a terrible disaster" claims STM's CEO

The earth cannot cope with the current pace of economic growth, Pasquale Pistorio, CEO of STMicroelectronics told an Electronica 2000 gathering.

"We have had this fantastic growth of the last fifty years but Planet Earth cannot sustain it," said Pistorio, "we are creating a terrible disaster. We are stealing from our children and grandchildren. Thousands of people are dying from pollution".

The problem cannot just be left to governments. "Companies and individuals should take the initiative," said Pistorio.

"Since 1994, ST has pursued a policy of sustainable development which gives good shareholders' returns, and is good for society and the environment".

"ST will have zero CO_2 equivalent emissions by 2010, by when we will be neutral to the environment," said Pistorio.

Biometrics and beyond

Biometrics used to be the territory of seventies James Bond films or sci-fi books populated with silver-suited types who could only get through doors by having their retinas scanned.

Now, the computing power needed for processing the biometric algorithms and the sensors for capturing the physical data are becoming cheap enough to be used by banks and health authorities and to be integrated into mobile phones and smart cards.

Beyond the basic technology, however, biometrics poses some difficult issues in terms of data protection, privacy and even human rights.

The commercial use of biometrics is mainly about data security as more services and data become Internet-accessible by a growing number of gadgets.

As the web enables stranger-to-stranger transactions, there's a need to authenticate who's on the other end of the line. "Access and convenience are the important drivers coupled with questions as to the long terms sustainability of the multiple PIN/password model," says Aline Coutts from the Nationwide Building Society.

The Nationwide has just completed a 6 month pilot study on speaker recognition as part of a wider European Community project called Picasso that has worthy aims to help people living in remote areas have access to the same information and services as people living in urban areas. (Other partners include Union Bank of Switzerland, University of Numegan, Perceptive, KPN Research and Vocalis.)

There is a growing range of hardware and software for implementing biometrics. There are several companies who make low cost silicon fingerprint sensors including Infineon, Polaroid, STMicroelectronics and Veridicom. These chips are based on capacitive or image sensing techniques and are being integrated into computer mice for logging onto corporate networks and smart cards.

Infineon promises to have flexible capacitive fingerprint sensors that are only 40µm thick by the end of next year specifically for smart card applications.

IriScan, the company with the sole patent for using for the iris as a biometric, has just formed a partnership with Panasonic for a PC access device based on iris recognition. IriScan is also talking to automotive manufacturers on immobilisation applications.

Paul Stanborough a consultant for IriScan reckons that iris recognition could be cost-effectively added to every mobile phone. "You just need a bit of software and the ability to grab an image. CMOS image sensors would be the way forward because they don't use much power." CMOS sensors could alternatively be used for verifying the user's facial features.

According to Graham Townsend, technical director of STMicroelectronics' CMOS image sensing division, the advantage of using optical sensing technology is that you are not limited by silicon size: "You have a lensing system that covers the sensor to enable an image to be taken that is much larger than the surface area of the chip.

"The state of the art is an optical area of 3 by 2.5mm. You can add a basic sensor to a system for under $10. Once you've got a CMOS image sensor integrated into your phone, it can be used for a host of camera applications including adding some sort of biometric."

Problems, problems

Human beings and the regulatory environment are the difficult variables in the biometric equation. If you live in the Yemen you probably won't bat an eyelid about the scary national ID scheme where everyone gets a swipe-card with a barcode that holds their fingerprint code and the details are kept on a central database.

In Europe or the USA, this would cause a national outcry about human rights and civil liberties.

It's all about personal biometric data (called the biometric template). Is it held by the person who owns it or on a central database? And if held centrally, how will it be safeguarded and what else might it be used for?

A recent story in the Sunday Independent about a UK scheme to combat rising credit card fraud sums up the issues nicely. The scheme requires shoppers to leave a real thumbprint on the back of their cheque or credit card slip so when there has been a fraudulent transaction, police can check it on a national fingerprint database.

"What it means is that it leaves a lot of personal information sitting around on cheques and there is a huge potential for abuse, said Deborah Clark, director public affairs for the civil rights organisation, Liberty. She fears that the scheme could lead to fingerprint scans replacing signatures as the means of security identification which in turn could lead to a national identity scheme via the back door.

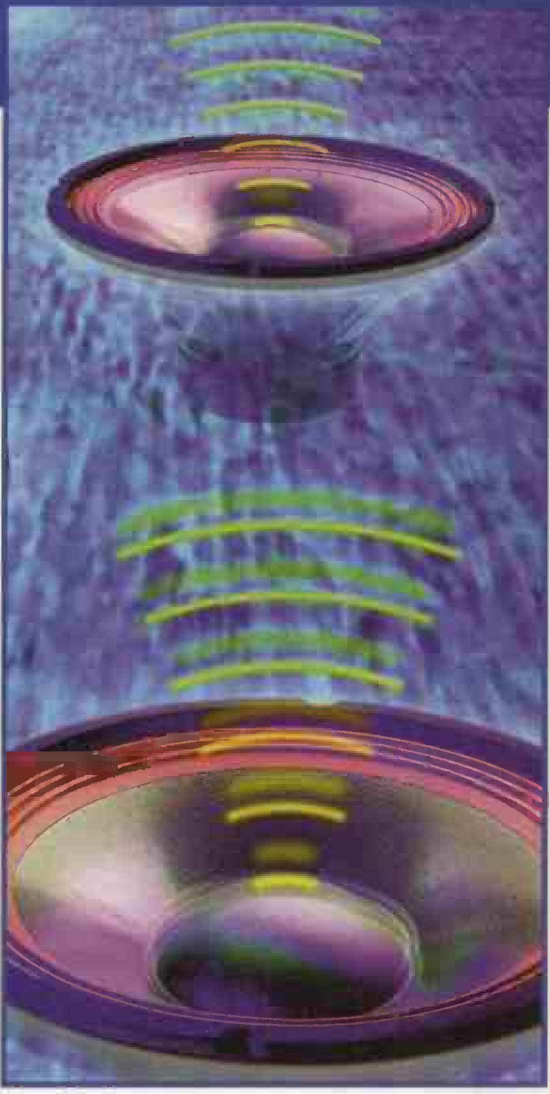
"It is the tip of the iceberg. If the police have access to these records there is a crime that has been committed, then that is one thing. But if the banks were holding databases of fingerprints, that is a completely different matter," says Clark.

Marek Rejman Green of BT suggests segmenting biometric templates as a way of addressing concerns about identity theft and privacy. In this way, the bits come together only at the time of a network transaction rather like a jigsaw puzzle.

No one holds the entire template and it circumvents problems of someone intercepting your biometric and you then requiring… er… a new finger or eye or face?

And biometrics change over time. The voice, for example, is affected by the psycho-physiological state of a person and alters with age and illness. So, if you've got a cold and are in bad mood, then presumably you might be denied access to your bank account. ■

Steve Bush *Electronics Weekly*



Phase-linear

There is a trend towards small, closed sub-woofer cabinets used in conjunction with electronic boost to compensate for the speaker's low-frequency roll off. But although such systems can potentially produce more realistic sound than their big-box counterparts, they fail to do so because of phase shifting in the equaliser, argues Graham Maynard. Here, Graham presents a general-purpose sub-bass booster with linear phase performance.

Hi-fi; stereo; semiconductors; digital; each of these milestone audio developments has contributed towards the accuracy, miniaturisation and power of affordable music reproduction systems.

However, and in spite of such technical developments, today's low-distortion tuners, CDs, and 20-20 flat amplifiers cannot ensure that we will enjoy 'fully realistic' sound reproduction. This is because their audio outputs are still used to drive permanent-magnet loudspeakers in much the same way as they were during the twilight years of the classic valve era.

I have no wish to denigrate any quality audio system, whether solid or hollow state. Most can generally be adjusted to perform well, but no matter how low their distortion, they cannot properly replicate lower bass sound waves with a flat 20Hz to 20kHz hi-fi response. This is because;

- loudspeaker transduction efficiency falls with reducing frequency
- socially acceptable playback levels make some low-frequency boost necessary to compensate for the natural fall-off in human hearing abilities.

Equaliser effects

You can use graphic equalisers and electronic crossovers feeding independently adjustable multi-channel amplifiers that drive separate bass, mid and treble loudspeakers to construct the most impressive of sound systems. Yet no matter how you adjust the controls to counter loudspeaker and hearing deficiencies, fully realistic reproduction remains frustratingly elusive.

This lack of fidelity is due to the frequency-selective-amplitude corrections for the natural roll-offs that occur within the audio spectrum. These corrections lead to our hearing becoming subjected to frequency-variable phase changes and time delays that affect waveform coherence and transient delivery.

Phase changes are why, for example, a little necessary bass boost from a Baxandall tone control or equaliser slider can sound okay. However, the greater amounts of amplitude correction that we might ultimately wish to correctly apply at lower listening levels, actually sound somewhat off-putting. A 'loudness' control is little better, and if one is used in addition to bass boost then the sound becomes atrociously soft and boomy.

Purists might use headphones, or stick to a flat amplifier response that they know does not alter signal phase or waveform coherence. Then, after selecting the best loudspeaker equipment that they can afford, they simply accept the natural low frequency roll-off that they can do nothing about.

Fortunately for experienced ears, flat amplification has little effect on the reproduction of choral works and the musicality of classical string and wind instruments. Yet, while it is the music and not the sound system that we listen to, it occasionally becomes obvious that some recordings have been made via studio equipment that had itself not been running fully flat!

Realism

For most listeners however, flat amplification and conventional loudspeakers will lead to a considerable loss of realism at playback. This is especially so when attempting to reproduce electronic and percussion instruments, full orchestra and open air performances, or the deep lows 'ambience' of larger auditoria where longer wavelengths have had the freedom to develop a 'spatial' sound that has been 'characterised' by the building itself.

You might not be able to hear 10Hz, but infra-sound beats and standing wave sensations are physically observed when we attend live classical and pop music performances. These become essential components of our learned experience.

When playback is carrying the subliminal harmonic and mixed signatures of such deep lows but their body-vibrating fundamentals are missing, then we cognitively realise that reproduction lacks the full set of characteristics that are essential to match our known experience, and thus to create the illusion of 'being there.'

sub-bass filter

Good recording equipment does transfer infra-sound frequencies to masters. When these are not filtered out prior to the disc printing run, then good CD players and systems can be made to reproduce them. This might be a tiny part of the sensory spectrum, and it is not always recorded, but it does remain very important: If you feel that the bass currently reproduced by conventional audio systems is not seriously compromised, then take a listen to known good music test tracks when played back through quality headphones.

Existing bass equalisers

Many active bass-equaliser circuits have already appeared in electronics publications. Most are additional in-line stages, though some use the power amplifier's negative feedback loop.

Those that start to roll off around 20Hz, or that incorporate a tuned-feedback loop to achieve their boost, make me cringe as I imagine the nature of the bass sound that would be reproduced.

With any frequency-selective amplitude adjustment or turnover, whether it be passive, electronic or loudspeaker in origin, there is always a phase change about and beyond its turnover frequency. If signal phase changes occur in the 15 to 100Hz range then low bass sound waves that initially had a coincidental starting point can recombine differently at playback and make an instrument sound as if it has been physically altered.

Occasionally, combinations of loudspeaker and equalisation characteristics so badly affect the sound that fundamental tones can run into quadrature with their own range of reproduced harmonics. Alternatively, they can become out of phase at frequencies that are not too far above and below what becomes an effective resonant 'Q' like turnover well within the audio spectrum.

When an amplifier or a loudspeaker amplitude response starts to roll off at, say, 20Hz, it is generally capable of good reproduction above 40Hz.

However, even though normal musical sounds seldom have fundamentals below 30Hz, reproduction can still be shown to be phase shifted and thus harmonically distorted. This is because a small deficit in the amplitude of a low-frequency fundamental introduces a larger level of distortion with respect to its entire family of harmonic overtones than would be the case if only a few of the upper orders had been similarly altered.

It is true that crossover networks also cause phase changes. Usually though, the aim here is to counter the reproduced shift in one driven channel with an equal but opposite effect in its continuing partner.

Why is phase important?

At the low-frequency limit of human hearing, where no other driver takes over, such phase distortions generate what we commonly term as a loudspeaker or system 'footprint'.

It has been said that we are less capable of noticing distortion at low frequencies, but I wonder what the reference has been. I also wonder whether this conclusion has been drawn while listening via conventional loudspeakers and amplifiers that start to roll off at 20Hz.

When phase changes occur, a range of frequency-related time delays develop. Previously accompanying wave components become separated at playback; i.e. they become displaced in time.

The resultant sound quite literally loses its coherence. Occasionally, the harmonic 'voice' of a bass instrument can be perceived before the peak of the first half cycle of its fundamental note. It is as if the bass loudspeaker cone is operating from farther and farther away with increasingly lower frequency.

There is thus a most definite need to maintain an overall flat phase response at all audible frequencies. Also, the associated amplitude linearity needs to extend to as far below hearing as is stably possible, without risking equipment reliability.

Filters that alter amplitude introduce phase related effects that extend out to two or three times, also to a half or a third of their nominal turnover frequency. Because of this, I personally consider that an amplitude roll-off should not be allowed above 5 to 7Hz for amplifiers and loudspeakers that are expected to perform linearly down to the lower limit of human hearing.

From a constructional viewpoint, minimising bass-driver and cabinet derived phase shifts below 100Hz then preclude the use of baffles. They also preclude large, ported or labyrinth cabinets and low-frequency auxiliary bass radiators that look like heavily coned loudspeakers without magnets.

These preclusions apply even when any of these options is driven via a specially-designed active equaliser element that maintains an overall flat audio response via that particular system. This is because air-column and mechanical resonances become independent of signal input once they have been exited. Resonant displacement can be controlled only via physical damping and/or motional feedback.

Yet the solution is simple.

The solution to the phase problem

Take a rigidly-coned long-throw driver motor, mounted within or against a small airtight enclosure. When crossed over and used resistively only at those frequencies that are below the new relatively-high resonant frequency, it will be more phase linear, more reliable and more compact than the conventional cabinets that we have come to accept.

However, a power transient induced resonance within such a compact cabinet that is fed by a poorly damping *LC* crossover can be unpleasantly distinct when compared to larger and conventional cabinets. As a result, an electronic crossover with a directly damping amplifier to voice coil connection becomes essential.

Also, the use of any resistor-capacitor type signal equalisation then merely introduces other separate

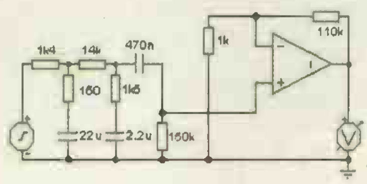


Fig. 1. Using a small cabinet, incorporating a rigid-coned speaker and mounted against a wall, offers many benefits over traditional bass enclosures. The drawback is that such a speaker needs considerably more power as frequency falls. This filter provides boost rising to 20dB at 15Hz.

but additional phase changes and/or a rumbly single note overhang effect. An electronically resonant circuit that generates necessary sub-bass boost has the same effect.

These distortions are as if the driver itself had been made very large, been mounted in a much larger enclosure, or had itself become under-damped. Thus the mechanical phase problems due to a loudspeaker being mounted in a conventional cabinet could erroneously be replaced by newly introduced electronic ones that sound equally undesirable.

You simply cannot reproduce a recording of clean and tight sounding bass on any system that is itself not clean and tightly damped in every regard; that means electronically, electro-dynamically, and physically.

Equalisation

Of course the above suggested compact loudspeaker solution is also seriously inefficient at generating sub-bass frequency sound-waves.

A considerable boost to signal amplitude becomes necessary at 15Hz with respect to 100Hz – approximately 20dB. At 15Hz, this 20dB boost then places serious demands upon amplifiers and drivers. A handling requirement of 100W at 15Hz becomes necessary to complement 1W at 1kHz.

Figure 1 shows a passive multi-stage resistor-capacitor filter network capable of providing the necessary boost. Its simulated amplitude response is the brown trace in the upper half of **Fig. 2**.

Also note however, that the phase response examination in the lower half of Fig. 2 reveals a considerable turnover lag at all audible bass frequencies. You will observe this same phase lag on any normal combination of active or passive equaliser circuits that act with respect to ground to provide the same slope of sub-bass increase.

This degree of phase shift runs bass instrument fundamentals into quadrature with the rest of their characterising harmonics, and renders boosted sub-bass reproduction woolly beyond tolerance.

During the summer of '98, I devised and test auditioned many different circuit architectures that were each individually capable of generating the same 20dB of amplitude boost at 15Hz. All were measured to check that their equalisation mirrored, and therefore compensated for, the natural roll-off of a dynamic loudspeaker working against a compact airtight enclosure. This meant a response that would hold 12dB per octave down to 10Hz.

In spite of their identical audio amplitude responses, each circuit sounded different. Some were grossly unacceptable due to phase distortion.

Then I realised that my dual-trace oscilloscope could be used to display the output waveform against signal generator input. This would allow me to observe any phase shifts that accompanied gain directly.

Eventually, I achieved a correct low-frequency amplitude boost characteristic. It always delivered output in the correct sense per sinewave half cycle. This means that it delivered zero crossing ac voltage input and output transitions that were in phase at all audible frequencies.

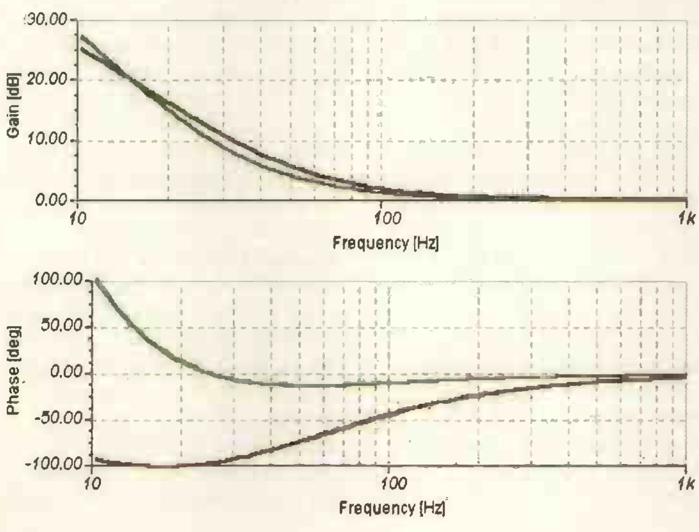


Fig. 2. Sub-bass amplitude and phase characteristics for the filter in Fig. 1, and the 'e-bass' circuit. The brown traces result from Fig. 1 while the green traces are the 'e-bass' circuit output characteristics.

I had heard, measured and disliked the more usual bass-frequency lag that results from filters similar to the one shown in Fig. 1. After much experimentation and repeated trials, my desire to counter this distortion resulted in the development of my final circuit, **Fig. 3**.

Both green traces in Fig. 2 show the much flatter but still boosted 'e-bass' response that I achieved and measured with conventional test equipment.

Circuit details

In Fig. 3, Tr_1 forms a low-Q 5Hz input filter. This is necessary to reduce infra-sound playback fluctuations. These can be caused by digital-audio bias and gain level settlement data, that should not be – but nevertheless occasionally is – imprinted upon some CDs.

This filter has a negligible effect on amplitude at audio frequencies, though it does introduce the unavoidably increasing phase lead that comes in below 15Hz. See traces 'a' in **Figs 4**.

All mid-high audio frequencies are twice phase reversed – first at Tr_1 collector, traces 'b', and then again at Tr_3 collector, traces 'd'. Transistor Tr_3's output, buffered by Tr_4, is thus shifted back to being fully in phase with Tr_1's base input, yet without delay.

Bass frequencies are separately taken from the emitter of Tr_1, traces 'a', with respect to the 180° phase shifted signal at the collector of Tr_1, traces 'b'. Sub-bass frequencies at Tr_2's emitter, traces 'c', thus develop an intermediate and frequency-dependent phase lead with respect to the higher frequencies at Tr_1's collector. The intermediate lead counters the typical lag of the brown trace in Fig. 2. This lag is caused by the passive resistor-capacitor high cut and thus sub-bass boosting networks acting on both Tr_1 and Tr_3 collectors with respect to ground.

The outcome at the emitter of Tr_4, is that all audio frequencies now remain in phase with the input. Bass frequencies have received a 12dB/octave characteristic boost to +20dB at 15Hz, with respect to 100Hz and above.

Audio output at Tr_4 emitter is fully in phase with the signal at input. As a result, a simple potentiometer arrangement between it and the first stage, Vr_3, is all that is necessary to tap off any desired level of phase-linear sub-bass boost between flat and +20dB at 15Hz with respect to 100Hz.

The greater the level of boost, then

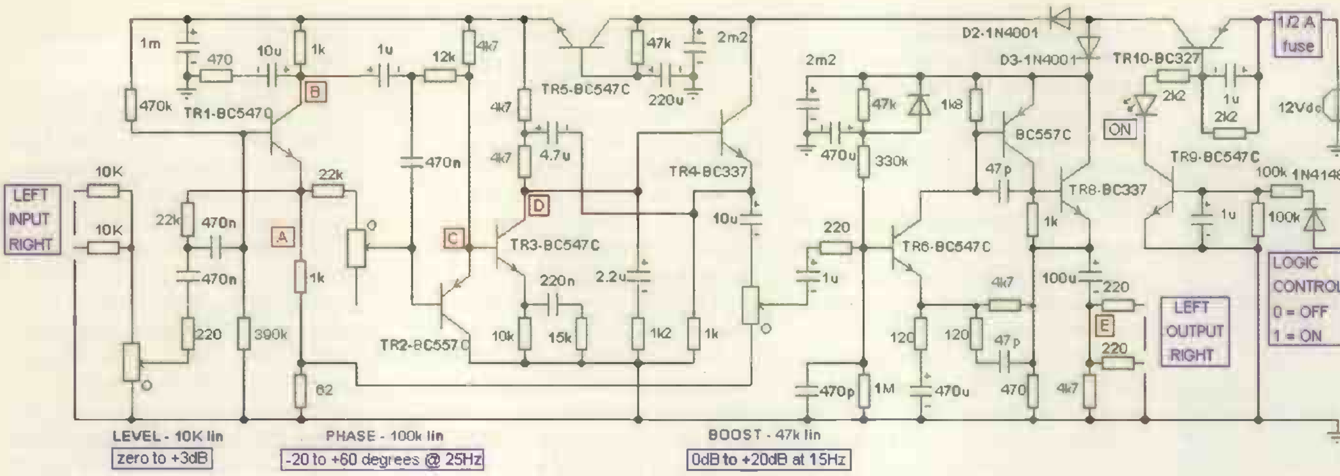


Fig. 3. Although more complex than the filter of Fig. 1, this 'e-bass' compensation circuit provides a much more faithful phase response, resulting in much-enhanced fidelity at sub-100Hz frequencies. A 5Hz input filter removes sub-sonic effects of CD gain and level settlement data that shouldn't occur – but sometimes does.

obviously the more that potentiometer Vr_1 might need to be backed off, to prevent sub-bass power-amplifier overdriving.

During listening tests, it became obvious to me that some CDs had been produced using equipment that had not maintained a flat phase response at sub-bass frequencies. Because of this, I introduced a third potentiometer.

Additionally, some unused equaliser phase shifting ability remained available between Tr_1 and Tr_2. So this potentiometer could prove useful in compensating for other undesirable loudspeaker and crossover shifts

I chose a component value that would allow a fully variable, but independent, sub-bass shift of between +60 and –20° at 25Hz. Thus phase was adjustable relative to unequalised audio that was being reproduced separately by the main bass-mid-treble component drivers. The result was a uniquely useful and user-controllable range of phase variation!

Initially, up to –60° had been available, but just as with ordinary bass controls, this amount of lag never sounded acceptable. Zero degrees is dead centre on the 100kΩ potentiometer. This also turns out to be the most common position set by ear.

Transistor Tr_4 bootstraps Tr_3 stably down to 7Hz. This optimises stage gain without any need for low impedance components at Tr_3 emitter.

Components $Tr_{6,7,8}$ form a simple two-stage, self biasing 2V rms output amplifier. This amplifier counters equalisation circuit losses when running flat.

Power supply

Transistor Tr_5 isolates the equaliser filter from supply-rail ripples. Thus the entire circuit eliminates loudspeaker thumping at power-up. This is something that op-amp circuitry cannot easily do.

An electronic on/off switch if formed by Tr_9 and Tr_{10}. This enables remote dc control from a pre-amp or

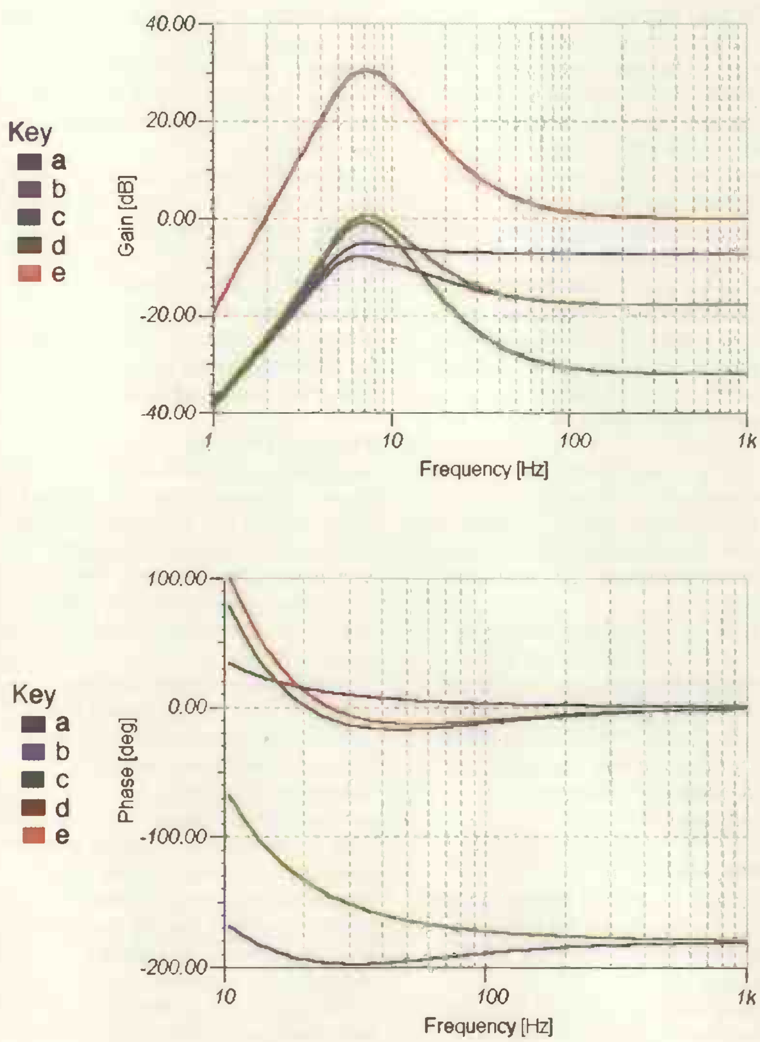


Fig. 4. Signal amplitude and phase characteristics at points within the 'e-bass' equaliser circuit.

head unit. These controlling transistors and their associated components can be omitted. If they are, diodes D_2 and D_3 will need to be powered directly from a main amplifier supply or via a separately switched rail.

This class-A constant-current circuit requires a single, positive, 11V to 16V dc supply at about 25mA. Fed via D_2 and D_3, it is tolerant of momentary psu interruptions, ripples and power-audio induced supply rail voltage dips when running from a commonned supply or car battery.

Diodes D_2 and D_3 could be connected to a raw a.c. supply. Where the equaliser must be installed at the end of long cable runs, a separately powered dc-to-dc converter might be considered. This would be inserted between the diodes plus the signal earth line, and Tr_9 plus the power earth line.

Adding a converter in this way would isolate audio circuitry from DC rails. I have not experienced any supply rail or common earth problems with prototypes, mainly due to the sharp roll off below 5Hz.

Remote control

During the design phase, I connected $Vr_{1,2,3}$ so that they could be controlled remotely from the listening position. I used a six metre screened multicore cable for this. No instability, interference or induced hum pickup problems were encountered.

My thanks to Andy Collinson at www.mitedu.freeserve.co.uk for the use of his TINA simulator program to generate the above amplitude and phase graphs directly from my final circuit. Andy's simulations confirm exactly my own longhand measurements made on 1998 prototypes.

In Fig. 4, amplitude trace 'e' shows a sharp turnover between the 12dB/oct boost and the 20dB/oct cut below 5Hz; this sharpness is achieved over several stages and thus is not resonant.

Also note that phase trace 'e' in Fig. 4 shows the overall phase deviation between input and output not exceeding ±13° down to 20Hz – figures that hold for the zero boost trace 'a' phase line: Can any conventionally cabinetted loudspeaker do better?

Realisation

The most significant test bed for this equaliser has been my son's 50Hz electronically crossed over sub-bass channel.

This crossover feeds a US-made Rockford Fosgate 100+100W 4Ω amplifier that maintains output down to 5Hz. It drives a Rockford Fosgate 15in diameter double 4Ω voice coil bass unit. The woofer is mounted in a sealed enclosure whose volume without the driver mounted is 2.3ft^3.

With one amplifier channel per voice coil, I observed a much superior cone control. The combination is capable of displacing up to one litre of air per half cycle; that is two litres peak to peak!

As this amplifier is capable of handling down to 1Ω per channel, I tried it in a bridged configuration with the voice coils in parallel to form a 2Ω load. The bridged set-up functioned exactly as expected; unnecessarily powerful with audible degradation in cone driving accuracy and damping.

An equalised power of 200W of cannot set any sound-pressure records when driving a compact sub-bass loudspeaker, but it is notably more realistic. The low frequencies of beating bowed cellos properly resound, and kick drums cleanly and tonefully kick air rather than thud-boomphing as if not a drum at all.

Also, reproductions from CD render more life-like stage and theatre hall acoustic ambiences. This aspect is hard to describe, it creates the illusion that your hearing is extended far beyond loudspeakers and the listening room.

Open-air recordings actually become more 'open'. Distant sounds clearly sound distant. This is because they are not modified by those rumbly bass loudspeaker colourations that we have become so used to, and which we can hear being differently exited on different systems as a result of air disturbance on a source microphone.

Positive clipping

Another positive attribute of the separately-powered sub-bass channel is that when a sub-bass amplifier runs into clipping, the rest of the audio system remains unaffected. Thus momentary sub-bass overdriving does not distort other bass-mid-treble output.

With a clean clipping sub-bass amplifier it can be determined that an insufficiency has occurred without it having had any mind catchingly deleterious effect upon listening pleasure. This is simply not the case with an inadvertently clipping main or whole bass amplifier, the effects of which can totally capture your attention and ruin our enjoyment for a long time after the momentary distortion has occurred.

This equaliser has proven effective with long-throw bass drivers as small as 8 inches in diameter. However, I trust that anyone who seriously attempts to produce realistic results will use a minimum single 12in, or twin 10in drivers. In the case of the twin drivers, one will face inwards, the other outwards and they will be wired oppositely.

Such sub-bass speakers should possess a genuinely continuous power rating of at least 100W. Also, the driving amplifier should clip cleanly and itself be capable of a continuous minimum 100W rms output. Its amplitude response should hold flat down to at least 5Hz so that its phase response will remain flat through the audible spectrum.

7Hz turn-over

To ensure that reproduction phase changes do not occur at normally audible frequencies, this equaliser's low-frequency boost has been set to turn over at 7Hz. This is similar to a normally specified half-power point. It is also the frequency at which the overall loudspeaker drive is allowed to fall away from a 12dB/oct boost characteristic. In other words, 7Hz is the frequency at which the equaliser and amplifier are no longer expected to provide increasing output with falling frequency.

Fortunately, there are few genuinely musical signals that require power below 30Hz. As a result, a rising power output is seldom necessary down to 7Hz. The 100W at 15Hz to match 1W at 1kHz really should be more than sufficient for loud sub-bass reproduction – even with modern synthesised instruments.

Nor should we forget that all of these turnovers, whether from flat to boost or cut, or from boost or cut to flat, alter the phase response for a factor of two to three beyond any nominal turnover frequency. Thus the 7Hz roll-off introduces a phase advance that adds to the input filter below 17Hz. Had the gain characteristic merely been allowed to satisfy the accepted 20Hz 'hi-fi' specification limit, then all audible sub-bass frequencies could have become phase distorted.

I have tested this circuit for several months with a 2.5Hz-input filter turnover, and 5Hz for the boost roll-off. The phase integrity was within ±20° down to 15Hz before the high-pass leading phase characteristics came in. But the resulting higher gain at infra-sound frequencies then over emphasised previously unnoticed CD flutter errors. Although inaudible, these errors did pressurise the listening room and introduce an unnecessary risk of additional non-linearity product generation.

Turnovers points of 5 and 7Hz were eventually deemed the sensible compromise between what was possible with good CDs, and what was realistically acceptable and practical on an everyday basis .

With 100 to 200W of continuous amplifier drive and an equivalently rated low-impedance sub-bass motor that is properly sealed against a compact and adequately rigid airtight enclosure, the cone simply cannot over travel. Air pressure in the cabinet prevents it from doing so.

Here, I must emphasise that cone stiffness and cabinet rigidity are essential. Normal driver cones or cabinets having sides that flex when you stand on them can leave a sub-bass system not necessarily vulnerable to, but capable of, occasionally producing a most unpleasant and whip cracking like release of sound energy.

Do not use a woofer for sub-bass if its entire cone does not move as a piston when it is pushed at a single point just inside the surrounding edge suspension; the cone must not be flexible.

With high supply rail voltages and low-frequency, high-power outputs, the SOAR limitations of the amplifier's output the devices become challenged. To overcome this problem, speakers with dual voice coils as low as 1.4Ω have already been manufactured in the US.

Drivers and enclosures

In recent years, loudspeaker manufacturers have reduced bass driver coil impedances considerably. This enables longer-wavelength air displacements to be generated with less risk of failure of the power amplifier's output devices.

To reveal the qualities and weaknesses of your own system, I recommend the highly competent Sheffield Labs Audiophile Reference test tracks that are supplied with recording plans in booklet form. These are compiled on the International Auto Sound Challenge Association competition CDs, also from the US. Check out the specialised links to manufacturers at www.iasca.com

For sub-bass loudspeaker construction, a simple starting point is to mount the chosen woofer externally against – not inside – an already sealed, thick walled cube. This cube should be lined and half-loosely filled with natural fibre wadding. It should have internal side dimensions that equal the chassis' outside diameter. Of course, one internal side dimension could be doubled if the depth is halved and the wadding is placed in line with the driver cone.

There are many 'per-chassis' Theile-Small parameter calculations that could be made. I would not wish to dissuade you from satisfying yourself about a cabinet's performance. However, calculated results could lead to one making an enclosure that is too large and hence not resistive enough for the 12dB/octave characteristic to be safely and smoothly applied.

When used singly, loudspeakers that are intended for sub-bass reproduction actually run more cleanly if they are mounted on to an already constructed cabinet facing inwards. This is because air flow through a tapered magnet vent is better able to cool the voice coil. As a bonus, harmonics that develop at the centre of the cone are less able to radiate directly. It is as if the crossover has an additional pole at no extra cost.

Component choices

I did not use any especially close tolerance components when making and testing prototypes. Resistors were sub-miniature 2% types. All miniature capacitors up to 10μF were selected within 10% of their rated value.

For neatness I used multiples of miniature 470μF radials to make up the larger electrolytic values. Initial dc testing was completed by checking that approximately half of the supply voltage developed at the collectors of Tr_4 and Tr_8 after dc bias levels had been allowed to settle.

All audio systems are different. As a result, the sub-bass crossover frequency will need to be determined by ear while simultaneously adjusting the equaliser potentiometers and any existing tone controls to level out an overall bass sound. This is not as difficult as might sound. Good results can be quickly achieved, but don't wrongly use maximum or impressive levels of boost, and don't use too high a crossover frequency.

Being realistic

This circuit maintains a tolerably flat phase response at all audio frequencies. However, it is no better at generating a low-frequency amplitude boost at the instant it occurs in real time than are any of the other frequency selective amplitude adjusting networks already in use. Here I am referring to the ineffectiveness of capacitors below a turnover frequency, and thus to the leading phase response that normally comes in below 30Hz.

Thankfully our brains appear to process waveforms on an averaging basis. When audible amplitude and phase responses are correct, we are much less distracted by the relatively few transient induced errors that here are now below the range of human hearing.

I am being entirely open here because most listeners are amazed by this equaliser's improvement to sub-bass sound when the significant phase shifts occur below 15Hz. Either we put up with larger loudspeakers and their audible weaknesses, or we accept equalised audio with occasional infra-sound phase errors that are much less noticeable.

An overall flat amplitude with flat phase sub-bass response actually sounds different from what is presently accepted as being 'normal' on conventional systems! Of course $Vr_{1\text{-}3}$ can be used to adjust the equaliser output towards a more direct and flat response if some source material already seems to be bass heavy or phase shifted. Medium boost with a leading phase can be applied if hardcore 'dance' playback is required.

Some discs sound disappointing. This led me to ask what types of loudspeakers are used in some of the production suites used monitor our prospective purchases.

I wonder if producers watch out for flutter errors that become permanently imprinted on their CD stamping masters after all dynamic and digital compression procedures have been completed, but before they go into production. These low frequency flutterings are to CDs what recorded rumbles were to some old vinyl discs.

Occasionally, cone movement can occur due to level compression or badly biased circuitry reacting to voice and solo instrument tones when no bass is being reproduced. The vast majority of CDs do play satisfactorily though.

Finally

As far as I am aware, this is the first 'phase linear' sub-bass equaliser circuit to be published. It was born of my unwillingness to accept bass phase distortion on everyday systems

The system has undergone repeated long term listening tests. These tests have been carried out with a determined empiricism and an enthusiastic disregard for any time that was necessary for development or evaluation.

I have not written about the many

previous designs that were completed, discarded, or superseded. Nor have I mentioned those that produced the same results but needed 1% capacitors, or which were less stable when driven with transients, etc.

A number of in car entertainment enthusiasts have listened to my system. They have been amazed at the much greater level of low-frequency output that can be correctly reproduced using a loudspeaker enclosure similar in size to the ones they are already using.

Like other *Electronics World* readers through the decades, I have had many opportunities to read about amplitude and phase relationships in audio articles. Soon after embarking upon this project, I was encouraged by the appearance of John Watkinson's 'Speakers' Corner' series.

John's writing sustained my determination to produce a circuit that would not generate the phase shifts that are normally introduced by amplitude equalisation when driving a compact and resistively loaded low frequency loudspeaker, which is otherwise much more phase linear in its own right.

I regard John's articles as being essential reading for anyone who has an interest in this form of bass reproduction; especially the October 1998 notes.

It is also worth mentioning that directly-driven resistive loudspeaker enclosures that operate below their resonant frequency do not hit amplifier output stages with the same momentary – and transistor popping – dynamic impedance dips that conventional loudspeakers systems do.

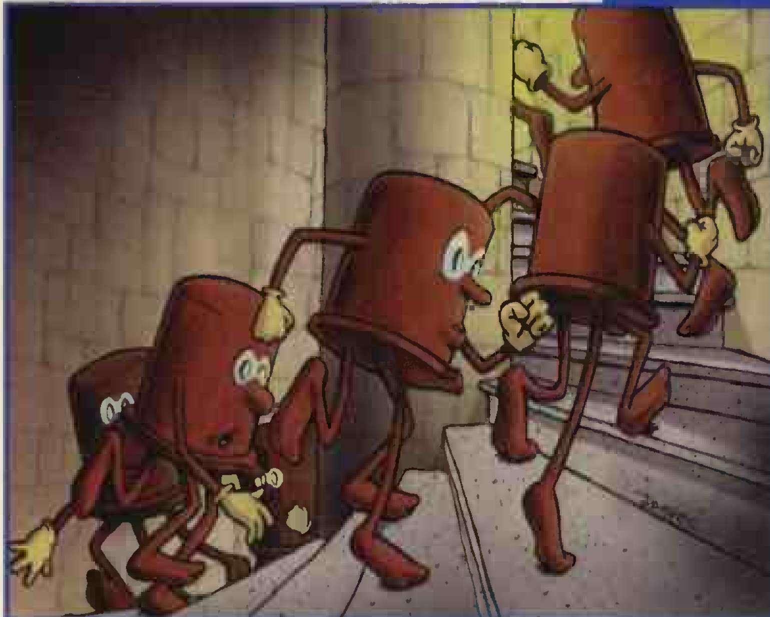
I am not claiming that this equaliser is the best or only way forward, but it does genuinely promote our striving for ever more realistic sound reproduction. I look forward to seeing other phase correction related circuits or letters appearing in future *EW* pages.

The circuit is straightforward and inexpensive. It is easy to insert it between an electronic crossover and a sub-bass amplifier. It may even be used as the basis for an additional and stand alone central sub-bass channel for use with already existing stereo systems. In this way, the original system might correctly be run flat.

Good sounding cinema, disco and professional loudspeakers need no longer be so large. Manufacturers could easily integrate an equaliser with an electronic crossover and sub-bass driving amplifier, to make 20Hz from 1ft^3 the domestic norm. Solid-state sub-bass can also be an excellent partner for tube amplified hi-fi. Many possibilities exist.

Phase linearly equalised sub-bass channels inconspicuously add depth to the sound stage in a surprising way that cannot be appreciated until experienced. All 'normal' systems then seem lacking, even when good and expensive reproducing equipment is already in use.

We cannot miss what we have not heard from our own CDs, but that is not a reason for doing nothing. Modern developments in amplifier and loudspeaker technologies already make the 'e-bass' approach a practical possibility. So come on – get your soldering iron out. ■



With the aim of helping you produce more effective analogue designs, Bryan Hart explains how it is possible to develop a set of 'ideal' transistor models that suit particular 'families' of applications. This second article deals with dynamic models, leading off with the charge-control model. The first covered static models.

An ideal transistor

Pioneered by Sparkes,[1,2] the charge-control model is applicable to dynamic switching conditions. In **Fig. 1**, its essential elements for V_{BC}<0 are shown superimposed on a schematic cross-section of the device in order to indicate the regions that they refer to.

The lettering denotes instantaneous values. Parameters defining the model may be unfamiliar to some of you, so I will outline them here.

Network element S_B represents a storage cell for a base charge, of magnitude q_B. This charge controls the transport current i_D via the relationship,

$$i_D = \frac{q_B}{\tau_T} \qquad (1)$$

In this equation, τ_T is the mean transit time of minority carriers crossing the base from the emitter to the collector.

By way of analogy, think of the relationship between the flow rate, quantity and transit time of a fluid in a pipe between two planes perpendicular to the flow. There is no potential difference across S_B. This is because the injection into the base, via the base contact, of a majority-carrier charge (q_B) causes an equal and opposite charge of minority carriers to be drawn in from the emitter region to maintain base charge neutrality. In this respect S_B resembles an infinite capacitance, and the capacitive association is reflected in its symbolic representation as a capacitor, albeit with curved plates. Thus, S_B requires a current,

$$\frac{dq_B}{dt}$$

only when the charge is changing.

Recombination of minority carriers entering the base from the emitter, and *vice-versa*, causes discharge of S_B. This is taken into account by the parallel current generator,

$$\frac{q_B}{\tau_B}$$

Here, τ_B is the overall minority carrier lifetime.

Under dynamic conditions,

$$i_B = \frac{dq_B}{dt} + \frac{q_B}{\tau_B} \qquad (2)$$

This equation resembles that relating the rate at which the quantity of water, q_B, is changing in a bath when it is being filled from a tap, i_B, and the water is leaking away at a rate,

$$\frac{q_B}{\tau_B}$$

through an ill-fitting plug.

Base-emitter voltage drop V_{JE} is associated with the injection of charge q_B into the base and is modelled by the diode D_E for which,

$$V_{JE} \approx V_T \log_e \frac{i_D}{I_S} = V_T \log_e \frac{q_B}{I_S \tau_T} \qquad (3)$$

As mentioned in the case of DC operation, parameter I_S depends only on the properties of the base region. Parasitic elements C_{JE}, C_{JC} and r_X are included in the model of **Fig. 2**. Junction-transition capacitances C_{JE} and C_{JC} are averaged

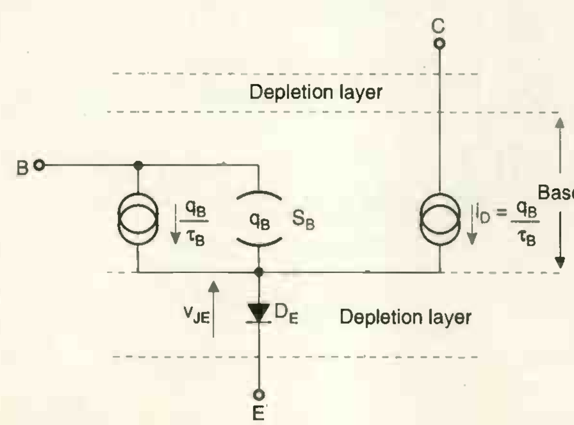


Fig. 1. Essential elements in the charge-control BJT model for a base-collector voltage of less than zero.

over their operating voltage ranges.

The base bulk resistance r_X is usually ignored in DC calculations. This is because the potential difference across it is negligible compared with V_{JE}. It is included in this model since it can affect the rate at which q_B can be injected, as in the example described below.

Under DC conditions a base current I_B is only needed to maintain the base charge constant at a value $Q_B=I_C\tau_T$, and equation 2 gives $Q_B=I_B\tau_B$.

From the two expressions for Q_B, the common-emitter current gain β is given by,

$$\beta=\frac{I_C}{I_B}=\frac{\tau_B}{\tau_T} \qquad (4)$$

As an example of the use of the model in Fig. 2, consider the circuit in **Fig. 3**. This shows the basic form of a scheme that I developed for producing two short-duration time-coincident rectangular voltage pulses of opposite polarity following the application of a voltage step $2V_G$, which is 1V in this case. This is balanced about earth and supplied from a pulse generator having an output resistance R_G, which is 50Ω.

The circuit has a long-tailed pair switch comprising $Q_{1,2}$ and tail current generator I_O. It also has a pulse-forming cable, with one-way delay time t_d and a characteristic resistance R_G of 50Ω, that is correctly terminated at each end.

Initially, Q_2 passes the full tail current but this is switched to Q_1 when the pulse edge arrives. A voltage change of $I_O(R_O/2)$ results at each collector. This voltage change ceases after the cable delay time t_d. When R_O is 50Ω, t_d in nanoseconds is around three times the cable length in metres.

Crucial to the design is an estimate of the switchover time, t_t, of the long tailed pair. This determines the minimum length of cable that can be used if the output pulses at the collectors of Q_1, Q_2 are to have flat tops.

In the model of **Fig. 4**, only those parts necessary to find t_t are shown. Recombination is ignored since it is anticipated that t_t is much greater than τ_B – an assumption that is subsequently justified.

During the switchover process,

$$i_B(R_G+2r_X)=V_G-(v_{JE1}-v_{JE2}) \qquad (5)$$

Using equation 3 shows that i_{D1}/i_{D2} is less than 0.01 when $v_{JE1}-v_{JE2}$ is $-5V_T$. This condition is taken as the start of the transition, which ends when $v_{JE1}-v_{JE2}$ is $+5V_T$, corresponding to $i_{D1}\div i_{D2}>0.99$.

The mean value of the right hand side of equation 5 is V_G so the mean base current, $i_{B(mean)}$, during the switchover process is,

$$i_{B(mean)}=\frac{V_G}{R_G+2r_X} \qquad (6)$$

This serves to remove a charge Q_T from the base of Q_2 whilst establishing a charge of equal magnitude in the base of Q_1.

$$Q_T=Q_B+Q_{JE}+Q_{JC} \qquad (7)$$

where Q_B is $I_O\tau_T$; Q_{JE} is C_{JE} (ΔV_{JE}), which is C_{JE} ($5V_T$);

$$Q_{JC}=C_{JC}(\Delta V_{JC})=C_{JC}\left[5V_T+I_O\frac{R_O}{2}\right]$$ Hence,

$$t_t=\frac{Q_T}{i_{B(mean)}} \qquad (8)$$

Suppose Q_1, and Q_2 each have the following representative values:

r_X=50Ω
$C_{JE}=C_{JC}$=2pF
τ_T=0.2ns
β=100.

Substituting this data in equations 6 to 8 shows that,

$$t_t=0.6\text{ns} \qquad (9)$$

Since t_t is much less than τ_B ($=\beta\tau_T\approx$20ns), neglecting the recombination is justified.

Each collector has a time constant, $C_L(R_O\div2)$, associated with it. Here, C_L is the stray capacitance. Assuming that C_L is 5pF, this means a 10% to 90% transition time of 0.275ns. Combining this with t_t in an rms sense, appropriate to cascaded systems, gives overall voltage transition times less than 0.7ns

The long-tailed pair offers the fastest known method of controlled BJT switching. This is because the control signal swing can be at its minimum value and the circuit can operate out of saturation.

Saturating circuits require a BJT model[2] that is a modified form of that in Fig. 2 but its use is not always necessary. Excess carrier-storage effects can be largely eliminated using a Schottky clamping diode in parallel with the base collector junction.

The hybrid-π small signal model

For small-signal wide-band and tuned amplifier design, an appropriate BJT model is the hybrid-π, introduced by Giacoletto[3].

From the charge-control model, it can be deduced by considering incremental changes about fixed bias levels, I_B, I_C, etc. Thus,

$q_B=Q_B+q_b$
$i_C=I_C+i_c$
$i_B=I_B+i_b$

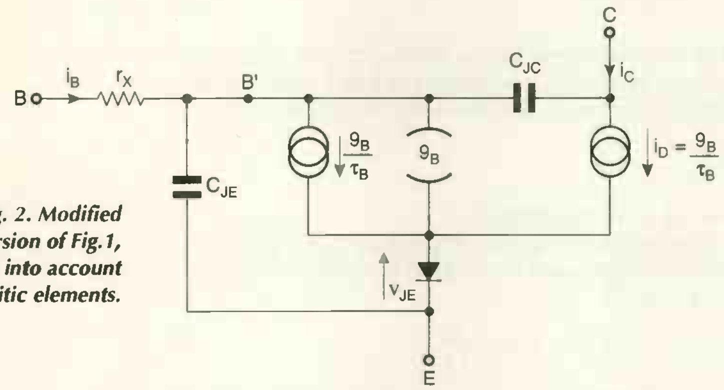


Fig. 2. Modified version of Fig.1, taking into account parasitic elements.

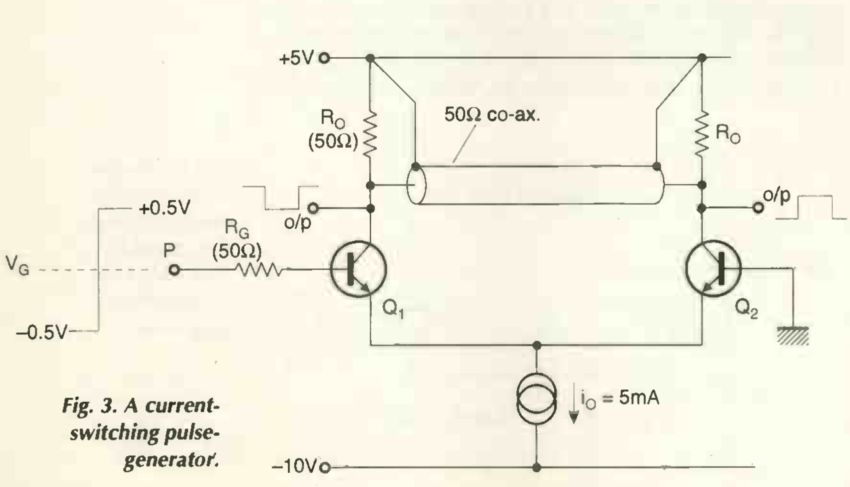


Fig. 3. A current-switching pulse-generator.

$v_{JE}=V_{JE}+v$

In these relationships, q_b is δQ_B, and hence is is much less than Q_B, etc.

Equations 1 and 2 now become,

$$i_c=\frac{q_B}{\tau_T} \tag{10}$$

and,

$$i_b=\frac{dq_b}{dt}+\frac{q_b}{\tau_B} \tag{11}$$

Differentiating equation 3, and rearranging, gives,

$$q_b=\frac{Q_B}{V_T}v=I_C\frac{\tau_T}{V_T}v=C_d v \tag{12}$$

Here C_d, which is,

$$I_C\frac{\tau_T}{V_T}$$

is defined as the base emitter diffusion capacitance.

From Equations 10 and 12,

$$i_c=\frac{q_b}{\tau_T}=g_m v \tag{13}$$

where,

$$g_m=\frac{I_C}{V_T}$$

is defined as the mutual conductance, or as 'transconductance'.

Also, the use of equations 12 and 4 gives,

$$\frac{q_b}{\tau_B}=\left(\frac{Q_B}{\beta\tau_T}\right)v=\left(\frac{I_C}{\beta V_T}\right)v=\frac{v}{r_\pi} \tag{14}$$

where,

$$r_\pi=\left(\beta\frac{V_T}{I_C}\right)=\frac{\beta}{g_m}$$

Equation 11 can now be rewritten as,

$$i_b=C_d\frac{dv}{dt}+\frac{v}{r_\pi} \tag{15}$$

The hybrid-π model in **Fig. 5** is based on a graphical interpretation of equations 13 and 15 together with the addition of the elements r_X, C_{JE} and C_{JC}. By definition, $C_\pi=C_d+C_{JE}$ and $C_{JC}=C_\mu$. The model parameters are either calculable from the bias and temperature conditions, or obtainable from terminal measurements. Thus, for a given collector current, $C_\pi+C_\mu$ can be found from the frequency dependence of the short circuit common-emitter small-signal current gain. By itself, capacitance C_μ, can be determined using a capacitance meter.

Element r_O, not so far discussed, is included to account explicitly for the implicit dependence of I_S, and hence g_m, on the voltage factor,

$$1+\frac{V_{CE}}{V_A}$$

mentioned previously. Normally V_A is much greater than V_{CE}, hence,

$$r_O\approx\frac{V_A}{I_C}$$

Note that r_π is independent of V_{CE} because β depends on the same voltage factor as g_m.

Among its widespread applications, the hybrid-π model has been used successfully in the design of amplifiers for high-speed optical transmission systems.

Conclusions

I have shown that the hybrid-π model is deducible from the charge-control model by considering incremental changes: in other words, by replacing q_B by q_b, etc. Similarly, DC models are obtained from the charge model by assuming q_B to be constant.

Relevant inter-relationships are summarised in **Fig. 6**. Apparently, the charge control model can be regarded as the 'big daddy' of a family of ideal models. ■

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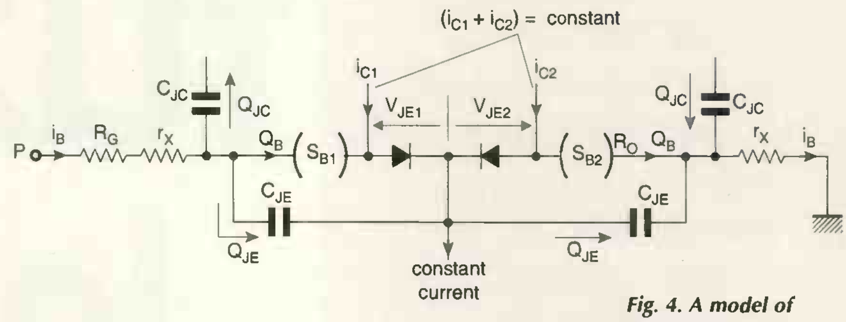


Fig. 4. A model of the circuit in Fig. 3 for calculating long-tailed pair switchover time.

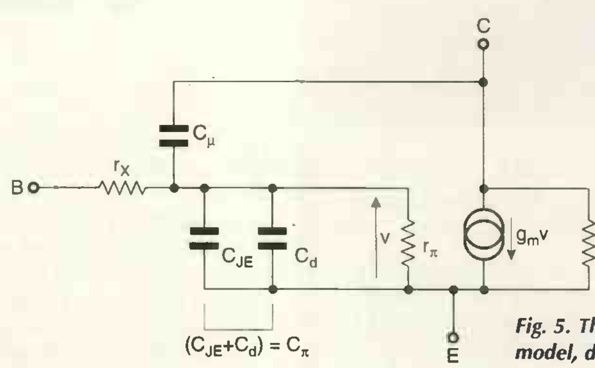


Fig. 5. The hybrid-π BJT model, derived from Fig. 2.

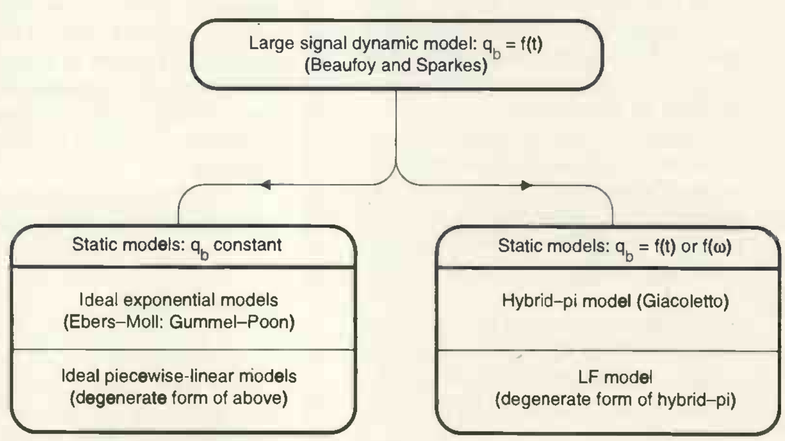


Fig. 6 A family-tree of ideal models.

Using fuzzy logic

Using conventional software methods to design fuzzy logic circuits leads to unnecessary complexity.
Douglas Clarkson looks at alternative ways of implementing fuzzy logic.

What is fuzzy logic?

According to Rudolf Graf's 'Modern Dictionary of Electronics':

*"**Fuzzy logic** – 1. A branch of logic that uses degrees of membership in sets rather than a strict true/false membership.*
2. A kind of logic using graded or qualified statements rather than ones that are strictly true or false. The results of fuzzy reasoning are not as definite as those derived by strict logic, but they cover a larger field of discourse.
3. A method of handling imprecision or uncertainty that attaches various measures of credibility to propositions. A form of logic used in some expert systems and other artificial intelligence applications.

__Fuzzy sets__ – Sets that do not have a crisply defined membership, but rather allow objects to have grades of membership from 0 to 1."

Fuzzy logic has come a long way since it was discovered in 1965 by Lotfi Zodeh in the USA. After being initially taken up by the Japanese, it was rediscovered in America, perhaps because the Japanese had had so much success with it.

In this first article, I outline the concept of fuzzy logic. In a second article I will be describing how to implement a basic fuzzy-logic unit.

This pair of articles will show you how to design a system with various analogue inputs, and how to process these inputs using fuzzy rules to create an analogue output. You will see that if conventional software methods are used to do this, then the resulting electronic circuit can become unnecessarily complex.

As I will show though, using a slightly different model can make it much easier to implement fuzzy logic.

Fuzzy-logic basics

Perhaps a better name for fuzzy logic would be 'continuous' logic, as **Fig. 1** indicates.

Take, for example, a pneumatic monitoring system that provides the pressure indications 'low', 'ok' and 'high'. These are essentially Boolean in structure. Whatever the pressure, only one function can be true. If the real pressure is at the threshold between the 'ok' and 'high' indications for example, there is ambiguity.

Fuzzy logic on the other hand embraces the concept of cross-over behaviour. In fuzzy logic, you can consider separate fuzzy logic functions. In this example, the functions are triangular in shape, rising to maximum value of 1, as in Fig. **1d**). These functions are called membership functions.

The whole point of a fuzzy controller is to have a series of such 'crisp' inputs, such as temperature, pressure, current, speed, etc., and derive appropriate 'crisp' outputs to control the system using fuzzy logic.

A balancing act

In **Fig. 2**, the position of a ball bearing is being maintained as centrally as possible on a see-saw guide. A geared-down DC motor can tilt the see-saw either way, depending on the value of drive voltage *V*.

A potentiometer indicates the shaft's position, producing an output voltage V_p. As the ball bearing moves along the slide, it changes the resistance of a second sensing circuit so that the system can detect the bearing's position using a bridge circuit.

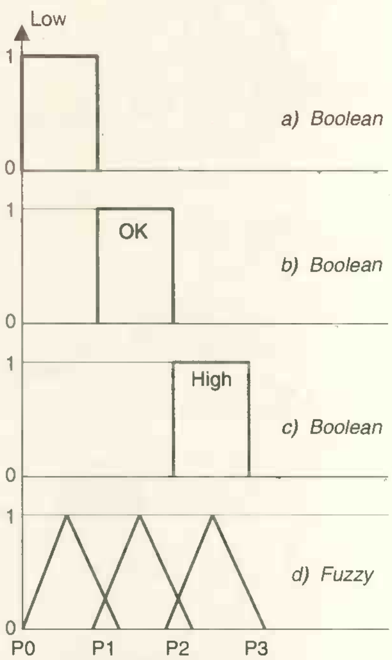


Fig. 1. Expression of states of a system in Boolean and fuzzy logic. 'Continuous logic' might be a better term than fuzzy logic.

Between the see-saw guide's end stops, the ball bearing can move freely. The problem is to develop fuzzy logic that will try to keep the unit horizontal, ideally with the ball bearing close to the centre position.

Figure 3 defines a series of conventional membership functions. The angle of the system is described by functions AR, AZ and AL while its position is described by PL, PZ and PR. The value of drive motor voltage is described by functions VNEG, NEG, ZERO, POS and VPOS.

While it is possible to identify up to nine rules that fully define the system, the system can be run with five rules as indicated in **Fig. 4**.

Table 1 indicates how the separate components of the membership functions behave for specific values of an input parameter. Such inputs are termed 'crisp' inputs.

For given values of ANGLE and POSITION crisp inputs, it is necessary to determine which – if any – rules apply. If you select values for which rules 3, 4 and 5 apply, then the process of calculation is indicated in **Fig. 5**.

In rule 5, the membership value of AR based on the input value of 12° is 0.35.

Similarly the membership value of function PR with input value of 3 is 0.5. What is required at this stage is to perform some function based on this pair of values that will give a measure of how much significance should be carried to the output function.

The most widely used approach is to take the minimum of these two values and to select the set of values in the hashed area of function VNEG. In a similar way, a value of 0.25 for the function ZERO based on rule 3 can be selected, based on intersection of AR/PZ of 0.4/0.25. Now this value for function ZERO can be applied to the output-function ZERO value.

Defuzzification – a process

If you combine, for example, two zones of the output func-

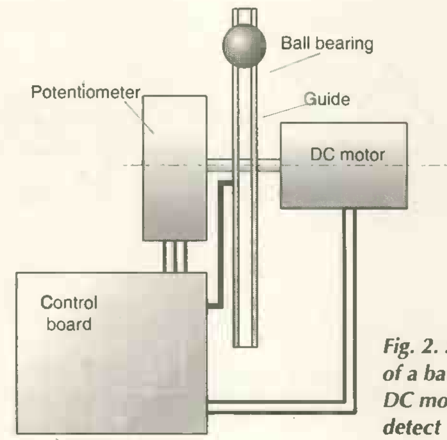


Fig. 2. Simple system of 'balancing act' of a ball bearing being balanced by a DC motor and a feedback system to detect the position of the ball.

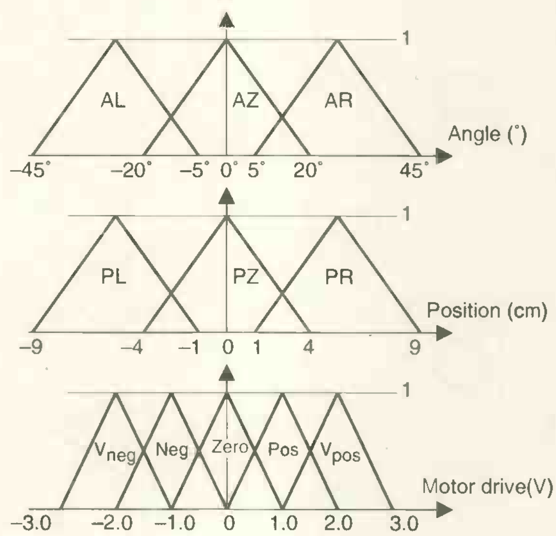


Fig. 3. Conventional membership functions of 'balancing act' example.

Table 1. Values of membership functions for 'crisp' input values of ANGLE AL, AZ and AR.

Crisp ANGLE	Membership value		
value	AL	AZ	AR
43	0	0	0.1
18	0	0.1	0.65
8	0	0.6	0.15
−2	0	0.9	0
−11	0.3	0.45	0
−39	0.3	0	0

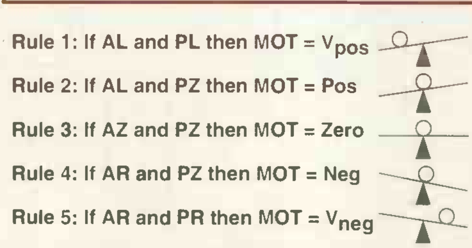


Fig. 4. Rules for 'balancing act' together with indication of positions of ball bearing and angle of motor, etc.

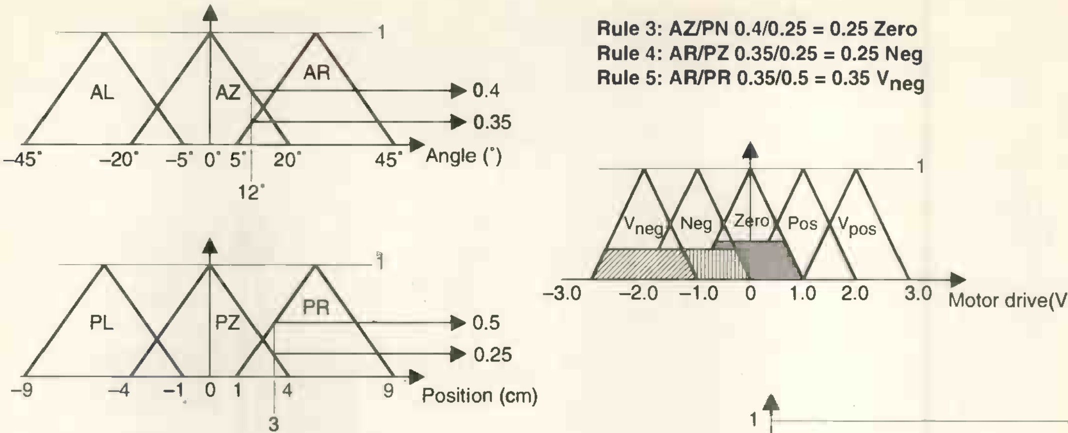


Fig. 5. Details of derivation of conventional intersection values of active rules for 'balancing act'.

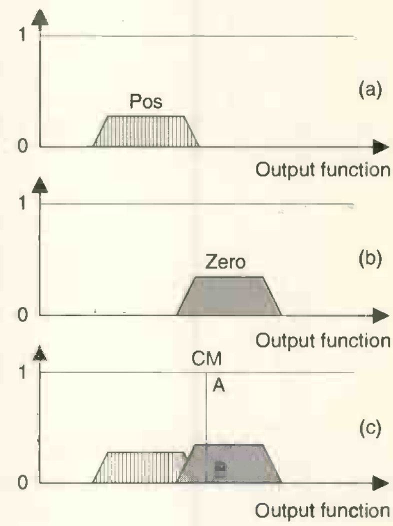


Fig. 6. General aspects of defuzzification

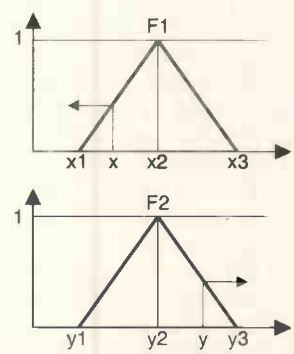
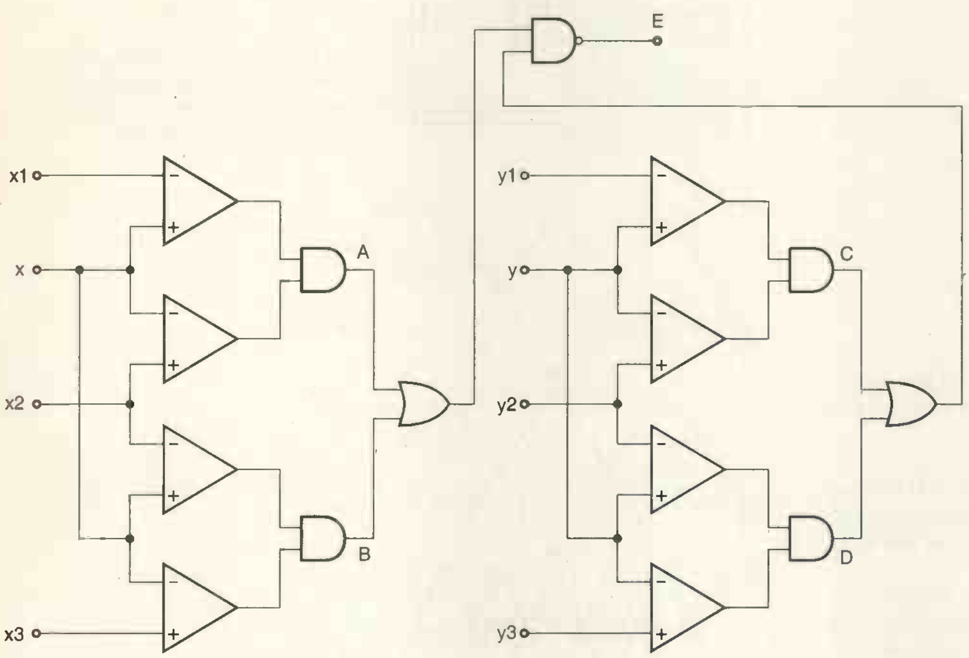


Fig. 7. Typical membership functions.



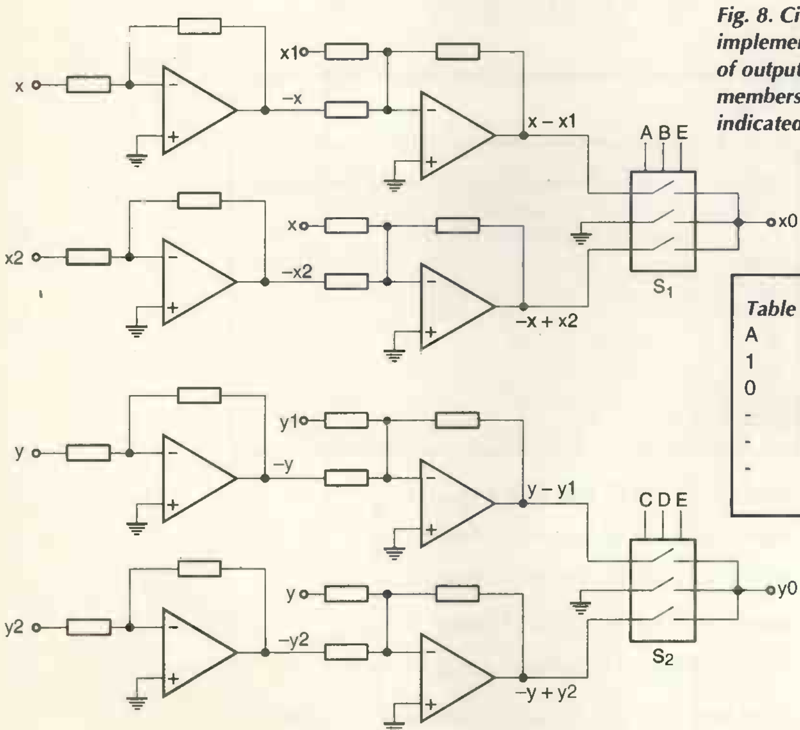


Fig. 8. Circuit implementation of selection of output value from a membership functions indicated in Fig. 7.

Table 2. Function selection for circuit implementation of Fig. 8.

A	B	C	D	E	xo	yo
1	0	-	-	0	x–x1	-
0	1	-	-	0	–x+x2	-
-	-	1	0	0	-	y–y1
-	-	0	1	0	-	–y+y2
-	-	-	-	1	0	0

tion NEG and ZERO, you can more closely identify the area of the combined output.

The problem now is to derive a single 'crisp' value of the output function MOTOR – a process known as defuzzification. While many methods are available, the most common one is known as the centroid, or centre of mass, approach.

In a simplistic approach, **Fig. 6**, imagine that line AB is the centroid. If you now imagine the hashed areas being cut out of cardboard, then line AB is the point at which the weight of cardboard on either side of it would be equal.

Many other methods can be used. If the output is considered to be a set of discrete points at $U(u_x)$ and value u_x, then the centroid value can be considered to be,

$$CM = \frac{U(u_1)u_1 + U(u_2)u_2 + U(u_3)u_3 ...}{u_1 + u_2 + u_3 ...}$$

This centre of mass calculation takes quite a bit of digital computing power to solve though, which can slow things down. An analogue solution, however, may even be faster. Provided that you do not have a zero on the denominator, an analogue division can be provided by an electronic circuit solution using analogue multipliers.

Implementing membership functions

In the example in **Fig. 7**, function F_1 has discrete values between x1, x2 and x3 and with variable input x. Similarly, the function F_2 operates between y1, y2 and y3 and with input y. Simplified membership functions required are given in **Table 2**.

Figure 8 outlines how this function table can be implemented in an electronic circuit to select the output value of each membership function. The logic of *A, B, C, D* and *E* is used to select the output using quad SPST CMOS analogue switches. Where any rule fails, *E* is logic 1 and a zero is selected for both xo and yo.

The next stage of the rule processing requires the intersection of the two values xo and yo derived for F_1 and F_2.

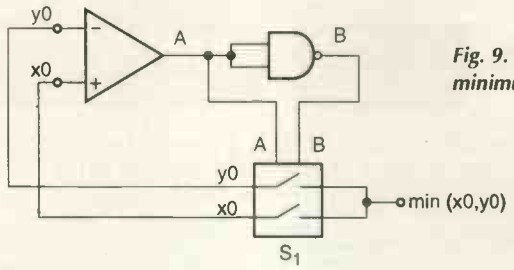


Fig. 9. Circuit for intersection: minimum of XO and YO.

This is generally taken as the minimum of the two values. So you need to design a circuit element that will select the minimum of the two voltage values.

Figure 9 outlines how this minimum function can be implemented.

At this stage, with the circuit of Fig. 9 processing the output from Fig. 8, the output value of a single specific rule has been determined.

The next stage requires this value to be transformed into an output value function, taking into consideration all the rules that may be active. At this point, things are getting complex, yet we are only at the stage of resolving only part of the 'output contribution' of a single rule. ■

In a second article, Douglas looks further at the complexity issue and describes practical circuitry.

Further reading

Fuzzy Controllers: Leonid Reznik, Newnes, 1997.

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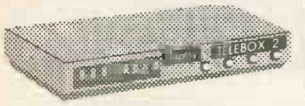

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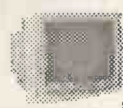

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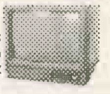

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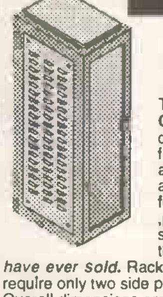

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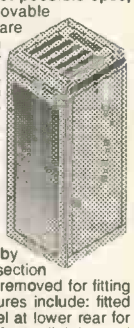

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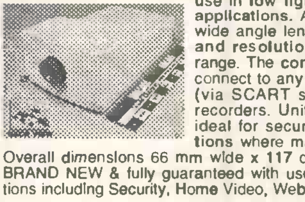

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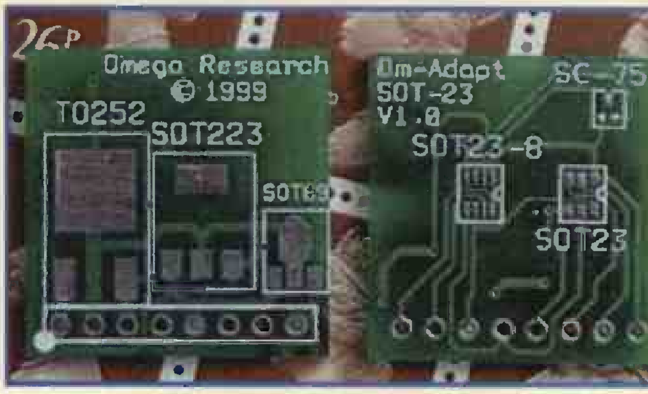

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Measuring true power

In his previous internet investigations, Cyril Bateman searched for conventional power measurement techniques. Here he reports on some more novel approaches, including PIC circuits and new chips that provide an alternative to the electro-mechanical system in your domestic power meter.

I discussed several circuits for measuring circuit power in my last internet article. Some involved measuring RMS voltage across a known resistor. Others computed the power consumed according to the product of circuit voltage and current.

Multipliers were used in the circuits involving computation. But other methods can be equally valid. The following circuit is perhaps the most novel of those I found in my Internet searches.

National's LM3812 integrated circuit is intended to monitor current levels and produce a duty-cycle output proportional to the current being monitored.[1] This circuit can be used to produce a novel, low-cost power meter.

Using the duty cycle output to gate the input voltage effectively provides a multiplication function but without using a dedicated multiplier circuit. A low-pass filter then provides a voltage output proportional to the power level.

While this approach sounds convoluted, it is easy to implement – and it doesn't cost much either.

In the circuit shown, the output voltage represents the input power, divided by ten to keep the output in range.[2] This circuit claims better than 3% accuracy with currents to ±7A and voltages from 2 to 5.25V, **Fig. 1**.

Another simple low-cost power meter uses the non-linear behaviour of a quad optisolator to act rather like a Gilbert cell and perform the needed multiplication. It also provides mains voltage isolation. This concept was originated in a much earlier 1994 design by W. Woodward, who also provides this updated circuit, **Fig. 2**.[3]

The optoisolator multiplier output is a current, drawn from C_1, that is proportional to the true instantaneous power delivered to the load. This current is then converted to a frequency in A_1 and A_2 to produce an output at A_2 pin 7 from 0 to 1200Hz. With the components shown, 1Hz represents 1W.

Using a copper-wire sense resistor with its temperature

Fig. 1. An alternative method for multiplying voltage and current involves the LM3812 current sensing IC. This easy-to-build circuit avoids the cost and complications found using conventional multipliers. It was designed by J. Kotowski and A. Johnston of National Semiconductors.

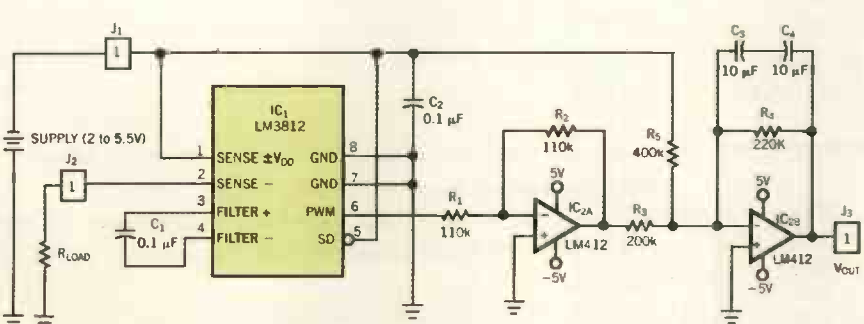
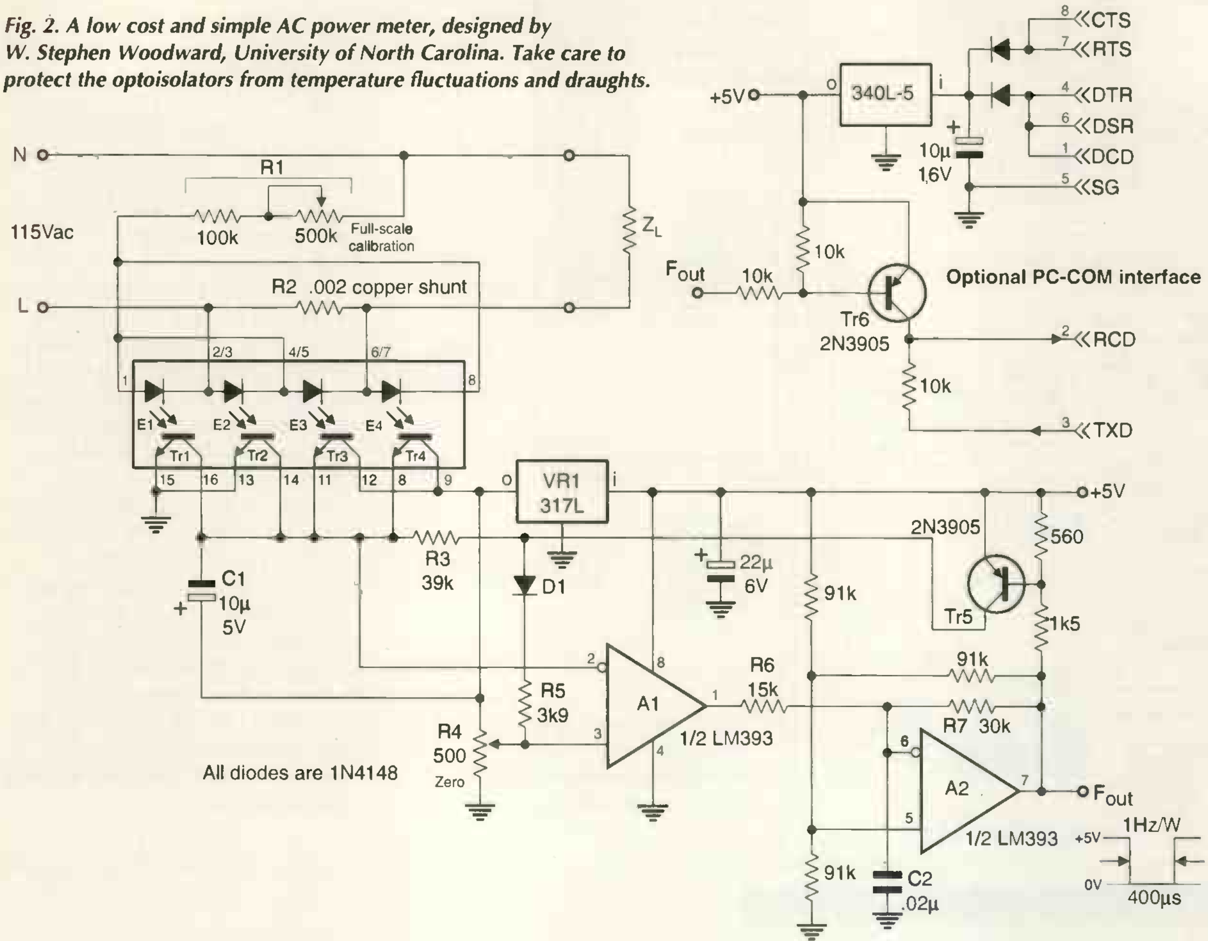


Fig. 2. A low cost and simple AC power meter, designed by W. Stephen Woodward, University of North Carolina. Take care to protect the optoisolators from temperature fluctuations and draughts.



coefficient of 3900ppm/°C serves to compensate for much of the LED and phototransistor scale-factor/temperature effects. However to preserve zero stability of the meter, the quad optoisolator and its components must be shielded from draughts.

This use of a quad optoisolator intrigued me, so I explored further. I was unable to find the original Woodward article on internet though. Searching for it however led me directly to my next circuit, which uses the optoisolator sensor together with a PIC to provide a digital readout.

A PIC power meter

A number of AC mains power meter designs have been based on low-cost PIC microcontrollers. Most use conventional methods to extract the voltage and current information.

This award winning circuit on the other hand, designed by Rick May, uses a quad optoisolator. Based on Woodward's circuit, the isolator generates an analogue voltage proportional to the power consumed by a load.

While not perhaps the most accurate power measurement method, the non-linear behaviour of the optoisolators permits an extremely simple, low cost and hand-held design. Instantaneous and average power up to 1200W are displayed, as well as kilowatt hours of consumption.

The essential power-to-voltage conversion using the quad optoisolators can be seen in the upper left third of the diagram. The remaining circuits convert this voltage into the desired output using the ADC0831 a-to-d converter with a PIC16C61, **Fig. 3**.

A five-page discussion of this circuit, together with the source code for the software routines used,

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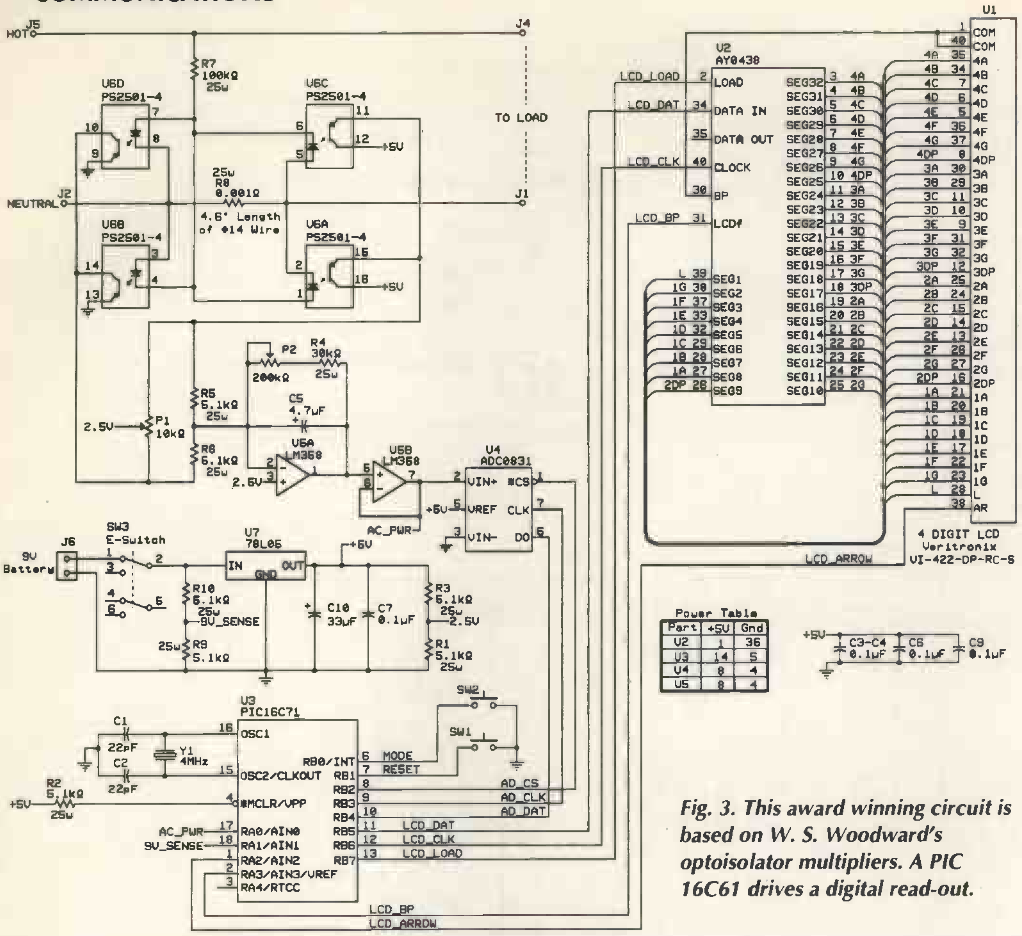


Fig. 3. This award winning circuit is based on W. S. Woodward's optoisolator multipliers. A PIC 16C61 drives a digital read-out.

can be downloaded from the Circuit Cellar website.[4]

In direct contrast to the simplicity of Rick May's circuit is a detailed reference design for a watt-hour meter using a PIC16C924. You can find this on Microchip Technology's site.[5]

Its schematic circuit spreads over five pages. To save space I have chosen to show only the AC mains to analogue-and-digital interface schematic. As you can see, this uses a transformer as the current input but a resistive divider for the voltage input.

To correct the input phase differences, a *CR* circuit of a 1kΩ resistor and three parallel capacitors totalling 0.378μF is used, **Fig. 4**.

The resistive voltage divider incorporates carefully-specified high-voltage 1MΩ resistors from Philips. In the interests of safety, these should not be substituted by lower rated components, which could be prone to flashover between element spirals.

Bugs or cookies?

Whenever the internet is accessed, each user leaves behind usage trails, recorded in log files on his or her computer and the various internet computers used.

More significant however are those attached to the many adverts commonly found on most web pages. These adverts not only slow down your web access. Many extract personal details from your computer while tracking your progress. This can happen even without 'clicking' on the advert: simply visiting the host page often suffices. Your details can be sold-on. At best this results in you receiving many unsolicited e-mails.

One way your private data can be extracted is by 'cookies'[A]. These were originally intended to travel only between the web host and your computer. A cookie was a temporary record of your access to a site so that, for example, if you were making an on-line purchase, it would retain your needs, allowing you to browse further before completing the sale.

Web pages often use components from a number of different servers, especially for adverts. Any server providing such content can install a cookie on your machine, to track your presence or extract information. Consequently without visiting the advert site, your data can be known.

The Windows 98 registration wizard inserts unique ID numbers that identify you, your computer and certain software installations, into cookies without asking your consent. Each time you visit the Microsoft site, this information can be accessed. According to *PC Magazine*, the ActiveX control has a bug that allows any Web site to retrieve your registration information.

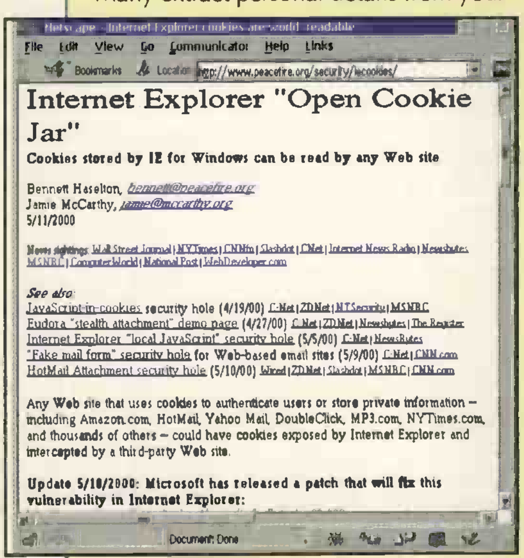
The easiest way to restrict cookie access is to completely disable cookie functions in your browser. But this will prevent you accessing a great many safe and desirable web pages. For some time I have set my machine to automatically accept only those cookies that will be returned only to the originating site. My browser then seeks my permission for all other cookies. This method of 'cookie' control works, but it can be a nuisance.

Today I examined the 'cookies' files of my four browsers on OS/2 and Windows98. I was concerned to find my name, e-mail address and other details held in various cookies as plain text – not encrypted in any way. How much larger and detailed might these files be had I not tried to control cookies?

A better approach may be to install cookie-management software. This can control cookie access and even prevent the downloading of adverts, thus accelerating web page access. I have now downloaded copies of 'Junkbuster'[B] and 'WebWasher'[C] and will watch their effect with interest.

Using Netscape, you will find your 'cookies' in a file called cookies.txt in the Netscape directory. For Internet Explorer, look in Windows\cookies.

Internet Explorer for Windows 95, 98, NT and 2000 can also be tricked into releasing your cookies to third-party sites. Microsoft has issued a patch to rectify this[D].



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Any Web site that uses cookies to authenticate users or store private information – including Amazon.com, HotMail, Yahoo Mail, DoubleClick, MP3.com, NYTimes.com, and thousands of others – could have cookies exposed by Internet Explorer and intercepted by a third-party Web site.

Update 5/18/2000: Microsoft has released a patch that will fix this vulnerability in Internet Explorer:

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Fig. A. If you access Internet using Internet Explorer with any version of Windows or NT operating systems, leaving your 'cookies' uncontrolled could allow your private data to become public.

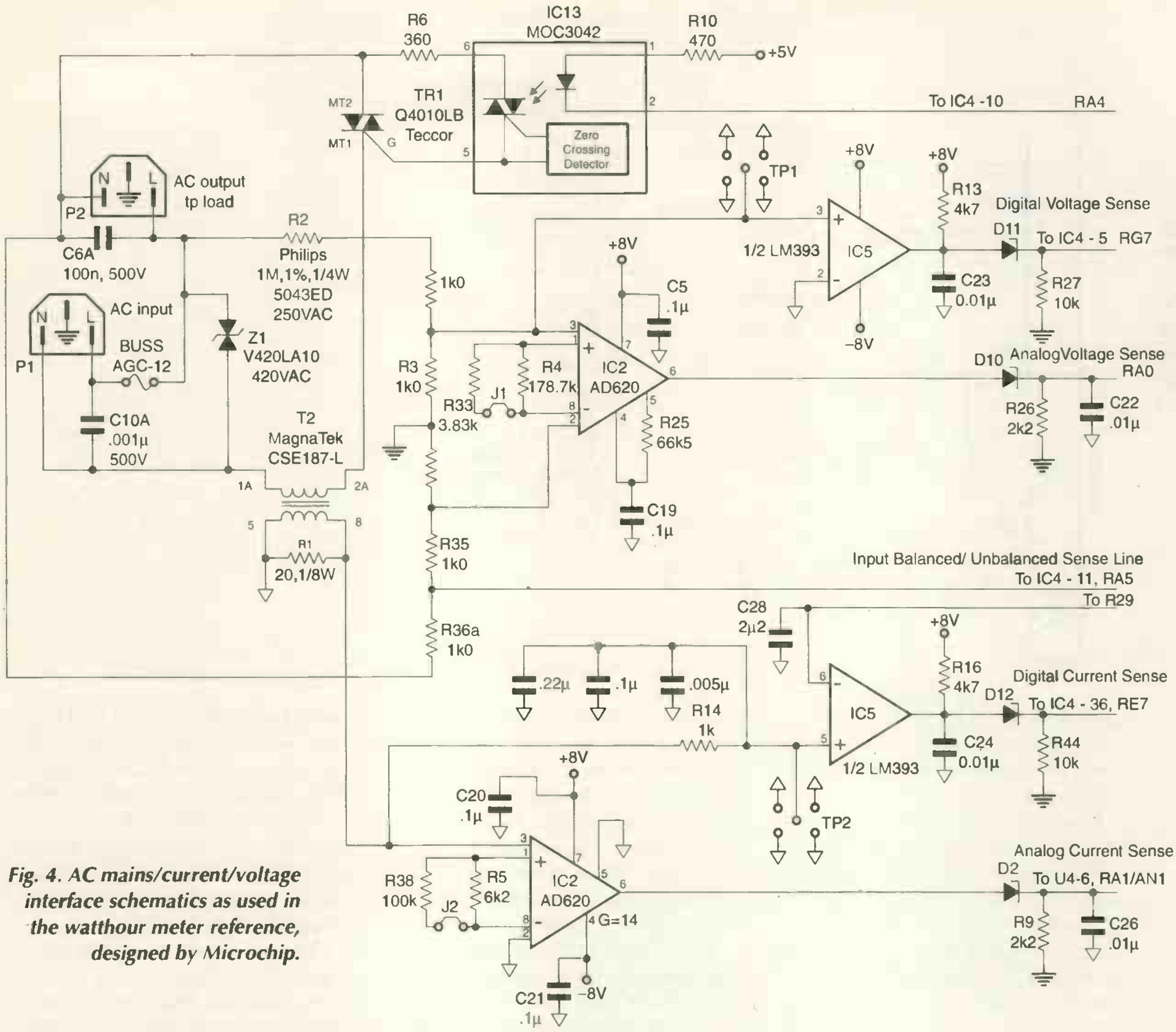

Fig. 4. AC mains/current/voltage interface schematics as used in the watthour meter reference, designed by Microchip.

Power factor

Complementary to the measurement of AC power is the power factor of your load. Knowing the power factor helps you to decide whether compensation capacitors are needed to attain a near unity power factor.

One of the simplest circuits I found uses only three ICs together with an ICL7106 DVM chip. It allows you to determine the power factor of an inductive load in the range 0.85 to 1 with 0.1% resolution and 1% accuracy.[6] To measure power factor, the meter has to solve the equation $V_m \cos\theta V_m$.

For safety, the meter is battery powered and uses current and voltage transformers to isolate and interface with the mains supply. The voltage waveform is rectified in a precision rectifier, then averaged in a long time-constant *CR* circuit to derive the mean voltage or V_m.

The load current waveform is amplified and converted to a balanced drive voltage. This is used to drive a synchronous rectifier which samples the voltage coinci-

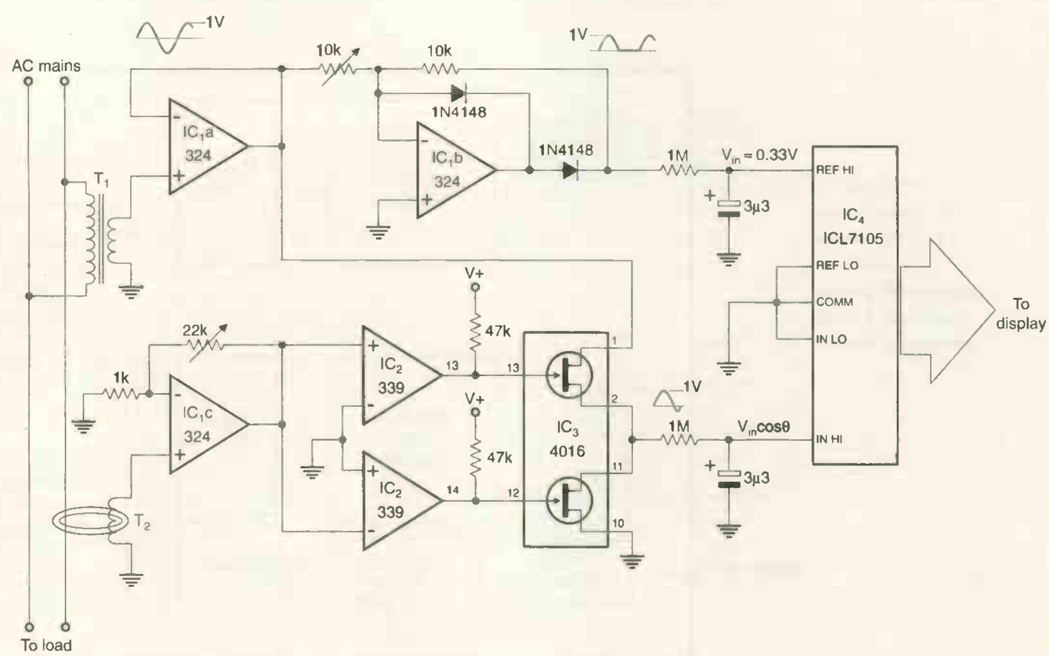

Fig. 5. Perhaps the simplest possible circuit devised for measuring the power factor of your mains-loading circuits.

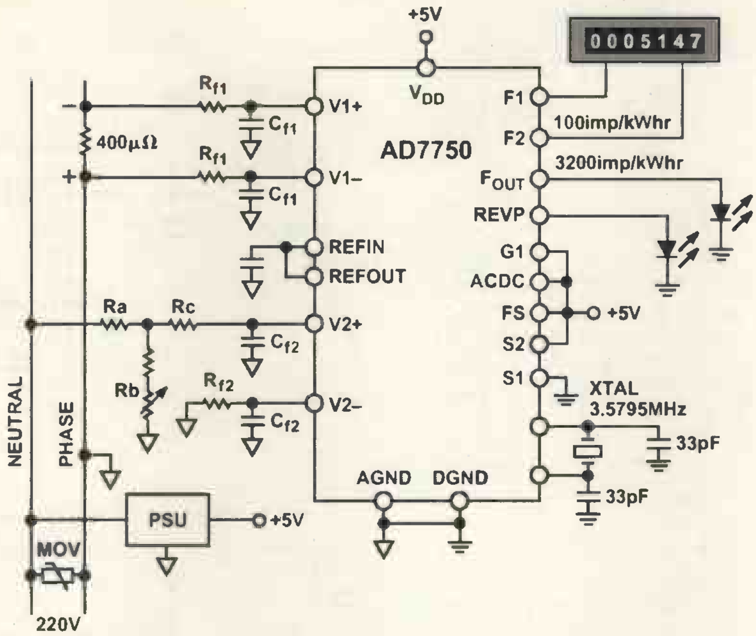

Fig. 6. Using a dedicated mains energy meter chip to measure the power consumption of your load.

dent with the phase of the load current. Output of this synchronous detector is also similarly averaged, deriving $V_m\cos\theta$.

Dividing these two voltages is easily performed using an ICL7106 DVM chip in ratio mode. With V_m input to the 'Ref Hi' input and $V_m\cos\theta$ to 'In Hi', the ICL7106 automatically divides $V_m\cos\theta$ by V_m to display the measured power factor, **Fig. 5**.

Measuring AC mains supply

The most common power meter is the billing meter attached to every home. To date, these have relied on electro-mechanical methods to measure your power consumption. Such meters are expensive to make and may lack accuracy.

Recent integrated-circuit designs have targeted this market, resulting in single chip solutions capable of driving either electronic or mechanical readouts, **Fig. 6**.

Both Cirrus Logic and Analog Devices offer dedicated ICs. These claim 0.1% and 0.2% reading accuracy respectively over a 1000:1 dynamic range. Other manufacturers may also have similar devices.

Not surprisingly, both makers demonstrate solutions that provide a variety of useful outputs besides the mass market's need for kW/h. These circuits can be used to measure RMS voltage, current and instantaneous power. Both makers' approaches provide outputs able to interface to a mechanical counter and an on-chip serial interface to a computer, **Fig. 7**.

Those of you interested in this topic will find a good introduction to the practical application of these circuits by downloading two application notes. These are CS5460RD.PDF from Cirrus logic[7] and AN-545.PDF from Analog Devices[8]. ■

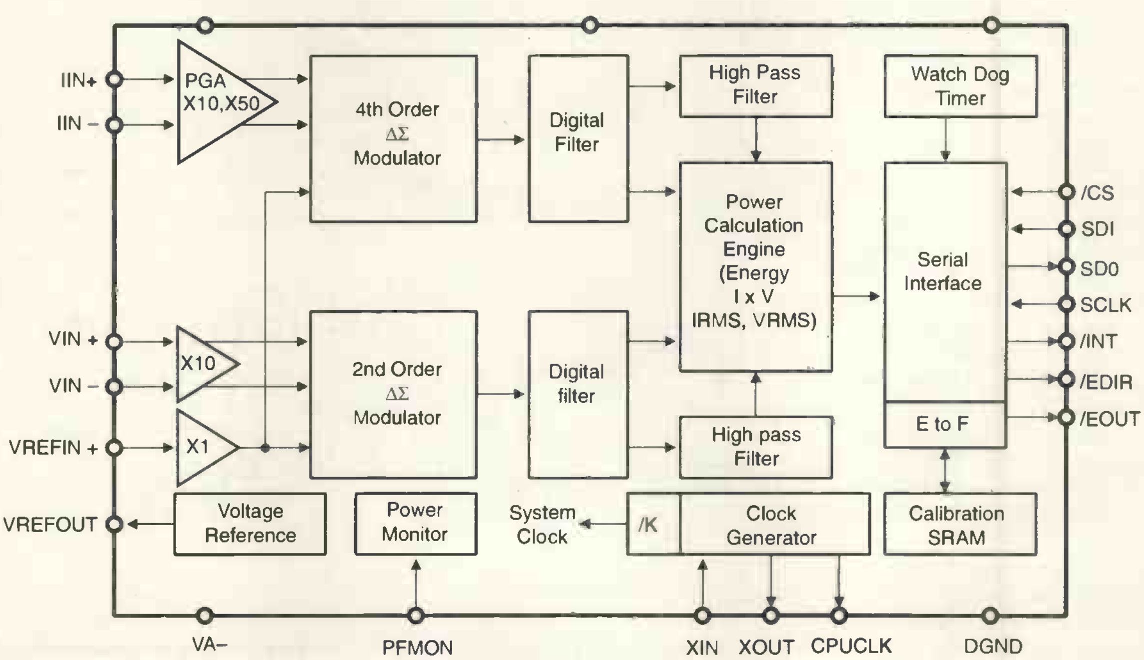

Fig. 7. Make-up of the Cirrus Logic CS5460 single phase bi-directional 'power meter on a chip' integrated circuit.

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Accurate milli-ohm meter

With a most sensitive range of 0 to 150mΩ, David Ponting's low-cost and self-contained milli-ohm meter measures low resistances without injecting a large current into the resistor under test. The meter can be used to locate solder bridges and PCB shorts as well as measuring resistances. Having a linear output, the circuit can drive a standard DVM display module, simplifying the design.

I was designing a circuit that needed an analogue ammeter capable of reading up to 2A DC. To save money, I decided that I would use a meter with a full scale deflection of 100mA, which I had to hand.

An accurate wire shunt was needed across the meter. It was not too difficult to calculate that with an internal resistance of 3kΩ, the voltage drop across the meter would be 0.3V for full-scale deflection. So the shunt needed to drop the same voltage at very nearly 2A. Hence its resistance had to be 0.15Ω.

If the resistance were 0.16Ω, the meter used would read 107 and if it were 0.14Ω the meter would read 93. This represents an unacceptable error of ±7%. Clearly, the accuracy of the shunt resistance is very important.

So I looked around to see if there were any off-the-peg resistance meters that would measure to 0.2Ω with some degree of accuracy. Most digital multi-meters have a lowest scale of 200.0Ω and have a resolution of 100mΩ. This is nowhere near good enough. In any case, leads and crocodile-clips or test points already provide errors of the order of 1Ω.

Beyond these multi-meters I found no low resistance measuring devices that could be said to be within the financial reach of the average electronics enthusiast.

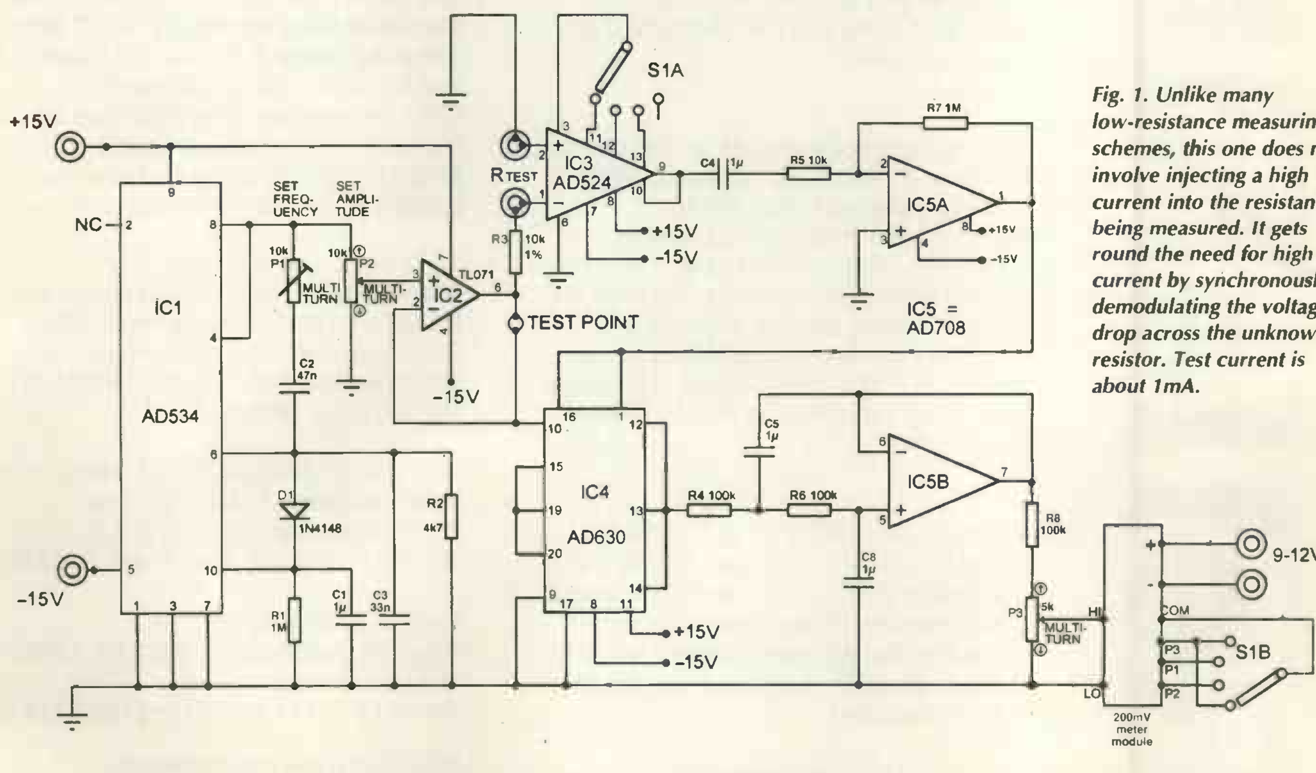


Fig. 1. Unlike many low-resistance measuring schemes, this one does not involve injecting a high current into the resistance being measured. It gets round the need for high current by synchronously demodulating the voltage drop across the unknown resistor. Test current is about 1mA.

Was it possible to make something that would do the trick? I searched around and found some circuits that got close, but all of those I found initially relied on injecting high currents into the unknown resistance in order to produce a large enough voltage-drop to be accurately measured.

High current would not matter in the resistance I was trying to determine in my meter-shunt, but making an accurate high-amperage power supply just to measure low resistance seemed excessive.

Then I found Analog Devices' Application Note number AN-306. It describes – albeit in scant outline – a circuit that will measure very small resistances without using high currents.

My circuit is an extension of this application note. It results in the practical realisation of an accurate milli-ohm meter.

Meter specifications

The meter has four ranges covering 150mΩ, 1.5Ω, 15Ω and 150Ω. All the scales are linear.

On its most sensitive range the meter will measure accurately and repeatably down to 1mΩ. It also gives a pretty good indication of the next decimal place, hundredths of micro-ohms.

Current through the resistance under test is fixed at 1mA, so most devices can be connected without damage. For example, the resistance across the closed contacts of a low-current switch or relay can safely be measured at the kind of current they will be expected to carry.

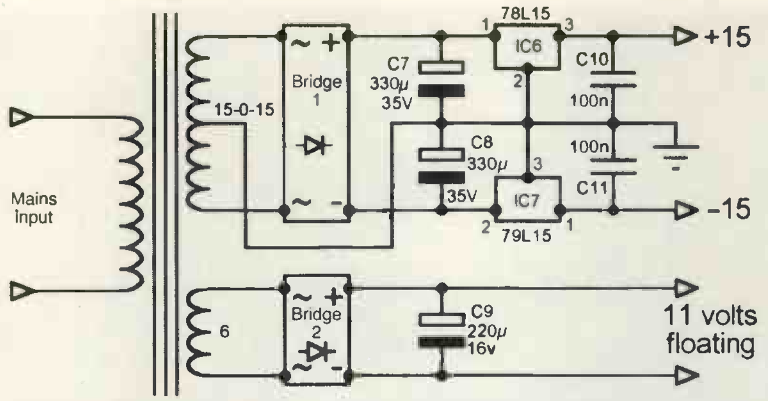
Figure 1 is the complete circuit diagram. Reading this from left to right, the first section of the circuit is built around AD534. This is an internally trimmed, precision-multiplier chip. Connected as shown, this IC functions as a sine-wave oscillator. Multi-turn potentiometer P_1 allows precise adjustment of the output frequency. Potentiometer P_2 sets the output amplitude. Op-amp IC_2, which is used here as a voltage-following buffer, can be almost any good-quality type.

With the proper adjustment of P_1 and P_2, a 10V peak, 1kHz signal is injected through both R_3 – the only precision resistor in the design – and R_t, the resistance under test.

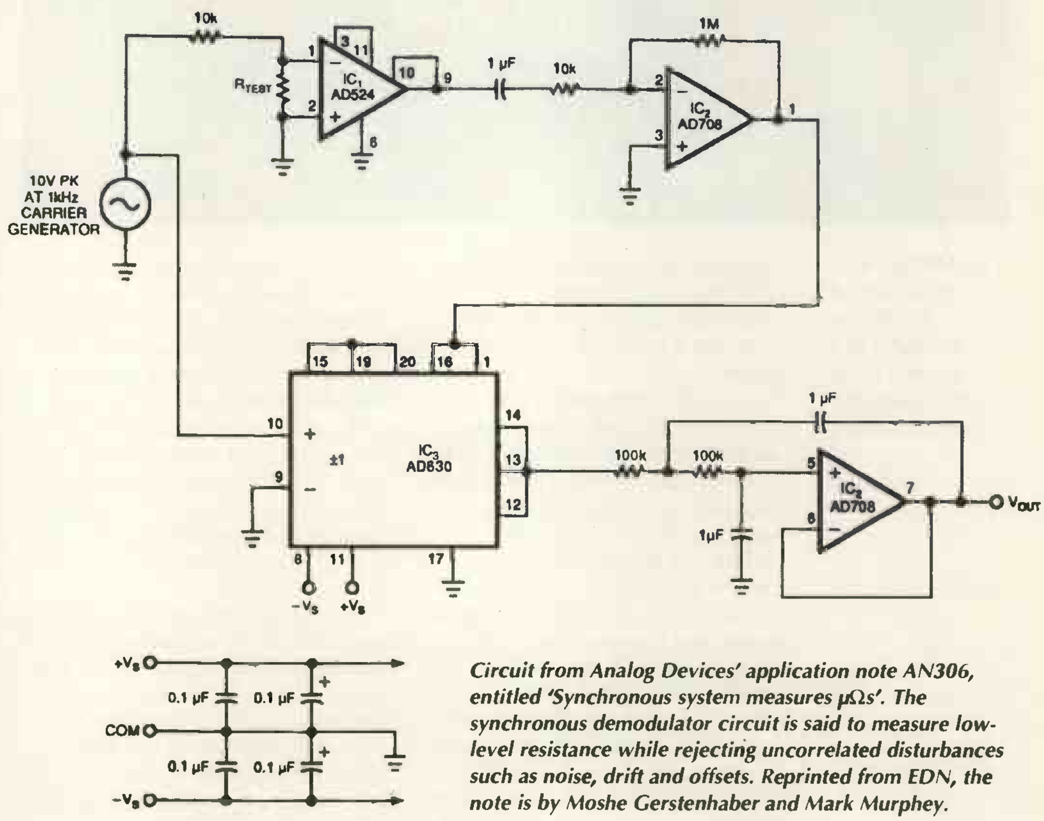
Voltage across R_t is detected and amplified by precision instrumentation amplifier, AD524. The output is further amplified by IC_{5A}, one of a pair of ultra-low-offset voltage op-amps in the AD708.

Now the signal feeds pins 1 and 16 of AD630, a balanced modulator/de-modulator chip. The 630 demodulates the signal and passes it to the second half of IC_5, which forms a low-pass filter. This filtering is necessary to attenuate those random drifts, noise and offsets not connected with the measurement of R_b.

Pin 7 of IC_5 outputs a voltage directly proportional to the resistance of R_t. This voltage is scaled by R_8 and P_3 and can be read directly as milli-ohms on either the 200mV scale of a digital multi-meter, or on the fundamental range of a dedicated voltmeter module.



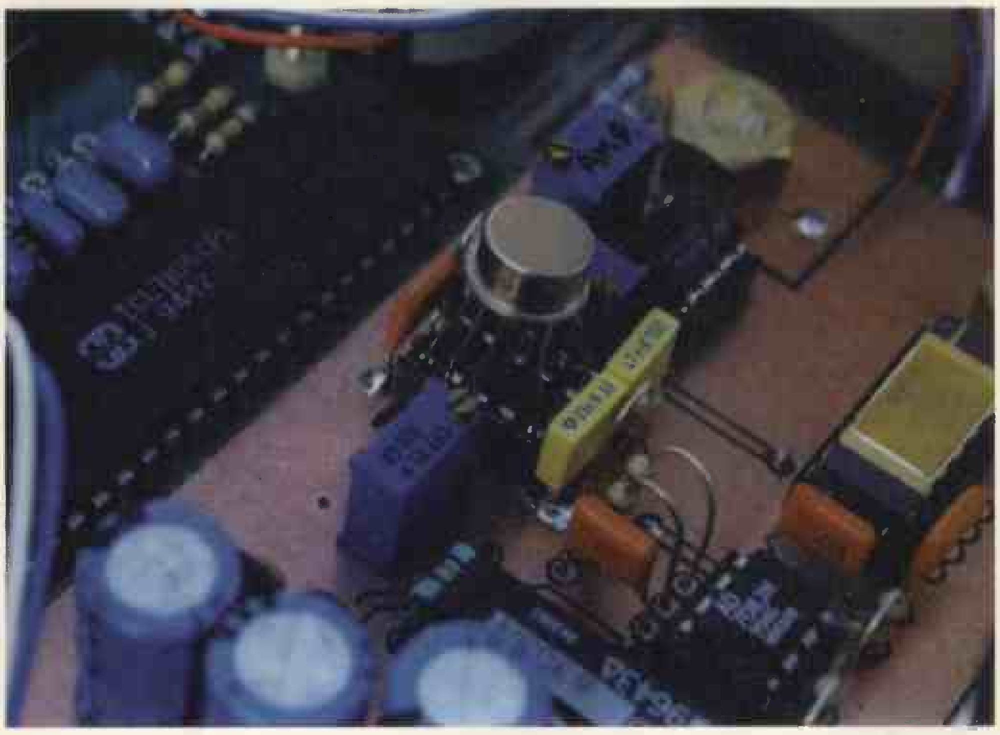
Power supply. Because the meter's current consumption is low, it is possible to use a 15-0-15V transformer to produce the two 15V rails. A separate floating supply is needed for the 200mV meter module.



Circuit from Analog Devices' application note AN306, entitled 'Synchronous system measures μΩs'. The synchronous demodulator circuit is said to measure low-level resistance while rejecting uncorrelated disturbances such as noise, drift and offsets. Reprinted from EDN, the note is by Moshe Gerstenhaber and Mark Murphey.

Why use an expensive oscillator IC?

You may think it a little extravagant in a simple design like this to use a somewhat expensive IC like the AD534 to do nothing more than



I couldn't get hold of a DIL version of the AD534 multiplier so I adapted a DIL socket to take the TO100 package.

regulators and rectifiers cause voltage drops.

I was fortunate in that the transformer I found for this circuit had an auxiliary 6V winding. When full-wave rectified and well smoothed, this winding did not need a regulator to provide the 9 to 12V floating supply needed to drive the voltmeter module.

It is possible to implement the circuit on a single-sided board. Double-sided board is better though as it allows lower grounding resistance. Where possible, the legs of components should be soldered to both sides of the board so that all the copper-fill becomes a single and substantial ground.

It is good practice to set each multi-turn potentiometer to its mid-point before soldering it. Doing this will ensure that the circuit – particularly the oscillator section – will at least function when the board is first powered up, even if the calibration is inaccurate.

produce a simple sine wave. The exact frequency of 1kHz is unimportant and the absolute purity of the sine wave is not really required.

In fact, the circuit shown around the AD534 does produce a pretty pure sine wave and does so at a very stable frequency. But its crucial contribution in the oscillator circuit is that its output amplitude is very stable.

Looking at the rest of the circuit diagram, the AD524 is switched so that pins 3 and 11 are joined. Without any external resistors, this conjunction produces a precise amplification factor of 1000. Similarly, a factor of 100 is achieved by joining pins 3 and 12, and a factor of 10 when connecting 3 and 13.

Further, if pin 3 is left unconnected, the amplification factor is unity, and hence the four positions of S_{1A} provide the separate ranges of the milli-ohm meter. If a dedicated 200mV meter module is used, pole B on S_1 is used to place the decimal point at the correct position in the display.

Four-terminal measurements

As with all attempts to measure low resistance, this design uses the 'four-wire' approach, which involves a pair of wires going to each end of the resistance being measured.

Ideally then, separate wires should go from IC_3 pin 2 and from ground, while a second pair should go from IC_3 pin 1 and from R_3. On my PCB, these four connections are paired off as close as possible to the terminals that take the component to be tested.

The meter measures any resistance beyond the point where the pairs meet. This means that the connecting terminals used to secure the ends of the component under test must make as good a contact as possible with the pads on the PCB.

Any connection terminals used must have as low a resistance as possible and must screw down firmly on the contacts of the component being measured. The ones I used were each made from a ¼in brass bolts with brass nuts. I even used large spring washers to ensure good contact with the PCB.

Any single wires used between the terminals and a component under test must be as short as possible and very thick. Crocodile clips used at their ends must be strong-spring-loaded and have sharp teeth to produce maximum electrical contact.

Power supply requirements

Figure 2 shows a suitable power supply. It is fairly standard. Because the circuit draws only a small current, it is possible to use a 15-0-15V, 100mA transformer to supply the ±15V rails, even though the

Setting up and calibrating

This will measure low resistance values to high precision. Between switch-on and warm up though, there is a small drift. Before you calibrate, it or use it for precise measurements, wait about 15 minutes. Warm-up drift is at most about 0.25%.

After warm-up, connect a counter or oscilloscope between the test point and ground. Adjust potentiometer P_1 until you measure 1kHz. Replace the counter with a digital multimeter set to read AC volts and adjust P_2 as carefully as possible for a reading of 6.37V. This odd figure is derived from $10(2/\pi)$, the mean value of a 10V peak, sinusoidal input.

Beg, borrow or steal a 1% or better resistor whose value is in the range 0.1 to 0.15Ω. It is worth some effort to get hold of a suitable reference resistor but close parallel soldering of ten 1Ω 1% resistors is good enough.

Set range switch S_1 to its most sensitive setting and screw down, as tightly as possible, your standard resistor to the meter's terminals. Then, in the case of a 0.1Ω standard, adjust P_3 so that the 200mV meter reads 100.0.

And that is it. All the other ranges are automatically adjusted.

You will soon notice that if the terminals are left open-circuit, the meter will read somewhere around 1900 on all ranges. In other words,

PCB files
We may be able to supply Proteus CAD files for the ohm-meter's PCB. If you are interested, please e-mail eworld.orders@rbi.co.uk using the words 'ohm meter' in the subject heading.

beyond the end of each useable scale, the output voltage is compressed.

It is difficult to generalise about an exact value at which the meter begins to become inaccurate. On the display module I used though, which reads up to 1999, I have found all readings up to 1500 are accurate. Above that, it is safer to switch to the next range.

Measurement tips

Getting good contact with the resistance under test is very important. Even with all the precautions that I took, a copper bar measuring 5 by 10 by 30mm screwed directly across the terminals still resulted in a measurement of 0.6mΩ. Consequently, I subtract something of the order of 0.6 from readings.

If you need to measure components that do not lend themselves to being attached via screw terminals, it would be better to lay out your PCB so that pins 1 and 2 of IC_3 float with respect to the terminals and ground planes. Then four separate wires can be brought out, two directly from pins 1 and 2, and two more directly from the R_3 terminal and the ground plane.

Each wire should terminate in a strong crocodile-clip. When measuring, the wires should be attached in pairs – pin 2/ground and pin $1/R_3$ – to the ends of the component.

After taking the measurement, all four connecting wires should be joined together and another reading recorded. This second reading provides the error factor that will need to be taken away from the reading of measured resistance.

Other uses

Having completed the final design, I find myself using it far more than I expected. For example, I found I could check the accuracy of the current reading on my bench power supply.

It uses a 0.1Ω nominal resistance in a power line to measure the voltage drop and hence the current flow. But I found the resistance to be significantly higher than 0.1Ω.

Shorting the resistor with an old 10Ω speaker potentiometer, I was able to adjust the combined resistance to a rather more precise 0.100Ω.

I have also measured various contact resistances of 'closed' switches and found how inaccurate it is to say that 'this point is directly connected to that point via the switch'!

And of course I measured the resistances of various odd bits of wire I found on my bench, just the ordinary kind of hook-up wire one uses to connect sections of an amplifier, for example. Suddenly the reason for hum via earth loops became very much more understandable.

There is one very important use for such a low resistance meter. You all know how easy it is to leave solder bridges across adjacent tracks on a printed circuit board during prototyping. And you also know how difficult it is to find them.

The resistance meter easily allows the tracing of such bridges. Using preferably the four-wire system, simply join one set of meter leads to one track and slide the joined second pair along the other. The measured resistance reaches a minimum at the position of the short.

The same effect can be achieved using the terminals and two short, thick leads but the sensitivity is not so high. ■

On the left is the milli-ohm meter with the copper shunt showing a reading of 0.6mΩ. Right is a 50mΩ, ±1% resistor being measured. Subtract the 0.6mΩ residual and the reading becomes 49.9mΩ.

Components

Resistors	
$R_{1,7}$	1M
R_2	4k7
R_3	10k, 1%
$R_{4,6,8}$	100k
R_5	10k
$P_{1,2}$	10k, multi-turn potentiometers
P_3	5k, multi-turn potentiometer.
Capacitors	
$C_{1,4\text{-}6}$	1μ
C_2	47n
C_3	33n
$C_{7,8}$	330μ, 35V
C_9	220μ,16V
$C_{10,11}$	100n
Diodes	
D_1	1N4148
$Bridge_{1,2}$	1A, 100V (piv)
Integrated circuits	
IC_1	AD534(JH)
IC_2	TL071
IC_3	AD524AD
IC_4	AD630JN
IC_5	AD708JN
IC_6	78L15
IC_7	79L15

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118

Track

1 Washington Post March, Band, 1909
2 Good Old Summertime, The American Quartet 1904
3 Marriage Bells, Bells & xylophone duet, Burckhardt & Daab with orchestra, 1913
4. The Volunteer Organist, Peter Dawson, 1913
5. Dialogue For Three, Flute, Oboe and Clarinet, 1913
6. The Toymaker's Dream, Foxtrot, vocal, B.A. Rolfe and his orchestra, 1929
7 As I Sat Upon My Dear Old Mother's Knee, Will Oakland, 1913
8 Light As A Feather, Bells solo, Charles Daab with orchestra, 1912
9 On Her Pic-Pic-Piccolo, Billy Williams, 1913
10 Polka Des English's, Artist unknown, 1900
11 Somebody's Coming To My House, Walter Van Brunt, 1913
12 Bonny Scotland Medley, Xylophone solo, Charles Daab with orchestra, 1914
13 Doin' the Raccoon, Billy Murray, 1929
14 Luce Mia! Francesco Daddi, 1913
15 The Olio Minstrel, 2nd part, 1913
16 Peg 0' My Heart, Walter Van Brunt, 1913
17 Auf Dem Mississippi, Johann Strauss orchestra, 1913
18 I'm Looking For A Sweetheart And I Think You'll Do, Ada Jones & Billy Murray, 1913
19 Intermezzo, Violin solo, Stroud Haxton, 1910
20 A Juanita, Abrego and Picazo, 1913
21 All Alone, Ada Jones, 1911

Total playing time 72.09

21 tracks – 72 minutes of music.
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NEW PRODUCTS

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Magnetics offer power feed in LANs

Pulse has launched a series of 10/100-base-TX local area networking (LAN) magnetic modules which it claims will eliminate the AC power requirements for IP (Internet Protocol) telephones and other low-power Ethernet devices. Capable of feeding up to 350mA on to standard LAN cabling, the H2005A and H2006A dual-port modules allow for enough power to drive many different applications over LAN cabling. These include IP phones, remote sensors, building thermostats, and wireless LAN access points. There are two part numbers in the series. The H2005A, designed for switch applications, provides maximum EMI suppression while facilitating power feed and return on LAN cabling. The H2006A is a more compact module for space-restricted applications such as IP phones, said the company. Each IEEE 802.3-compliant part is housed in an IC grade, transfer-moulded package, suitable for convection and infrared reflow soldering at temperatures up to 235°C. The new LAN magnetics are currently available in tubes or tape and reel packaging and are compatible with most high-speed pick-and-place assembly equipment. Details are listed in data sheet H327.

Pulse
Tel: 01483 401 700
Web: www.pulseeng.com

Enhanced BGA rework system

Metcal has enhanced its ball grid array (BGA) rework system. This system is available in two models – the CSP-3500 designed for smaller chip scale rework, and the

BGA-3590 with high-powered under-board heating for thermally demanding PCBs. The main system enhancement is the addition of an improved imaging camera that provides better definition when working on fine pitch components. A prism feature allows users to simultaneously look at the topside of the PCB and a superimposed image of the component underside. The images can then be accurately aligned in the X, Y and Theta axis. In addition, the improved vision system includes facilities for either solder paste or flux application without the need to remove the PCB from the machine. Other enhancements include simplified system controls, which are now accessible from the front of the machine, and improved board handling. In fact the under-board support has been completely redesigned to ensure easy adjustment and firm support without the risk of damage to the PCB.

Metcal
Tel: 023 8048 9100
Web: www.metcal.com

LDO consumes 650nA independent of load

A low-dropout (LDO) regulator featuring a low 650nA ground current has been designed by Micrel Semiconductor for use in battery-powered devices. Capable of driving a 10mA load, the MIC5231 is available in the company's proprietary IttyBitty SOT-23 package. Aimed at power sensitive applications, this micropower LDO uses the firm's FCap

Current-sensing motor drive IC claims to halve shutdown time

International Rectifier has introduced the IR2172 linear current sensing IC with built-in over-current shutdown. Suitable for motor drive applications, the device is claimed to eliminate external current sensing protection circuits. It is rated at 600V, for 230V three-phase AC inverter or brushless DC industrial motor drives used in conveyor belts and similar applications. The device has a 1.5μs over-current shutdown feature, with a wired-OR connection capability direct to the microprocessor or DSP. Conventional linear current-sensing opto- or Hall effect-based systems have typical shutdown times of three to four microseconds. Motor drive circuits using opto or Hall effect sensors require two external comparators and a level-shift operational amplifier to achieve the same functions, says the supplier. The motor driver comes in two industry-standard, compact package outlines: SO-8 and 8-pin DIP. The monolithic high-voltage IC construction is designed to eliminate optical or mechanical links between the current sensors and motor drive output circuits, and increases reliability while shrinking circuit size as much as 75 per cent, compared to circuits using Hall effect or opto-based couplers.

International Rectifier
Tel: 0208 8645 8001
Web: www.irf.com

Please quote *Electronics World* when seeking further information

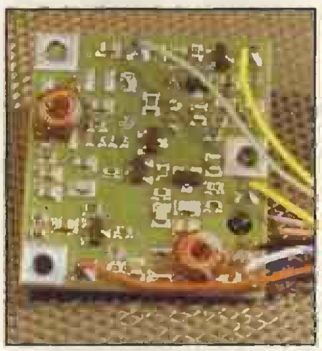
design, which is optimised for stability with low-cost ceramic or tantalum capacitors, or with no capacitor at all. Features includes two per cent initial accuracy, no need for an output capacitor, enable input, fixed output voltages: 2.75, 3.0, 3.3 and 5.0V, an input voltage range of 3.5 to 12V and a temperature range from –40 to 85°C.
Micrel Semiconductor
Tel: 01635 524455
Web: www.micrel.co.uk

Switching IC for wide varying loads

Ultimate Renaissance has made available the STR-G6551 integrated off-line switching IC. The device operates in fixed off-time mode with a maximum frequency of 60kHz. The STR-G6551 features a 3.9$R_{ds(on)}$ avalanche rated 650V, 158mJ FET and includes over voltage protection, under voltage lockout and thermal shutdown circuitry. It is available in a 5-pin isolated TO-220 package.
Ultimate Renaissance
Tel: 01793 439310
Web: www.ur-home.com

Reference clock with stress compensation

General Hybrid has developed a primary reference clock, the Alpha 1, for Quartzlock's Cosmo double oven-controlled crystal oscillators. The circuit, which incorporates a stress-compensated quartz crystal, is the main element in the



rubidium frequency standards, GPS and quartz oscillators made by Quartzlock. The circuit is based on an aluminium nitride substrate that has a thermal expansion similar to silicon. Applications include telecoms, metrology, radio transmitter referencing, calibration and navigation.
General Hybrid
Tel: 01295 256363
Web: www.ukgeneralhybrid.com

Pipes shed light on infra-red devices

Hero Electronics is introducing Signal Construct light pipes for use with infra-red devices. Conventional and surface

mount types are available. For arrays of devices such as IR LEDs, photodiodes or phototransistors, vertical and horizontal pipe arrangements are available with between three and ten light guides in a unit. For single devices, such as remote control receivers, straight bulkhead mounting pipes are the main option, but flexible pipes are available. Applications include set-top boxes, measurement systems and remote control transmitters.
Hero Electronics
Tel: 01525 405015
Web: www.heroelec.co.uk

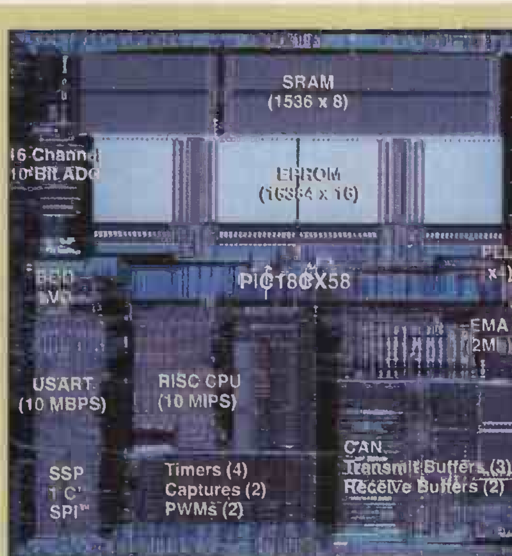
SBC for medical and POS applications

Advantech has introduced an EBX single board computer. The PCM-9550F/M's Pentium MMX 266MHz processor, EBX form factor, digital I/O and multimedia features make it a suitable embedded platform for medical and POS applications.

Power consumption is typically 3A at 5V and it operates without a fan in temperatures up to 60°C. The EBX form factor provides an embedded layout and PCI bus for expansion modules such as 100Mbit/s Ethernet or fax-modem. Eight digital inputs (TTL level signals) and eight digital outputs are provided. There are four com ports with +5 or +12V controlled by jumper select. There is also cash drawer support.
Advantech
Tel: 01908 304800
Web: www.advantech.com

3.3V VCXO comes with $\pm100\times10^{-6}$ pull range

Epson has introduced the VG-4231CA voltage controlled crystal oscillator with a pull range of $\pm100\times10^{-6}$. The VCXO is for use in networking and ADSL equipment and digital set-top boxes. There is an optional extended range from



PIC micros combine 10 Mips CPU with 32kbyte program memory

Microchip's PIC18C658 and PIC18C858 microcontrollers combine a 10Mips CPU core with 32kbyte OTP program memory, 1536byte user RAM and a controller area network 2.0B active peripheral interface. The intelligent CAN interface lets complex control algorithms and network interfaces be executed on the same microcontroller making them suitable for automotive and industrial control systems. The CAN interface contains a double-buffered receiver with two priority levels, six full acceptance filters and two acceptance masks. Three transmit buffers are available for application-specific prioritisation.
Microchip
Tel: 0118 921 5858
Web: www.microchip.com

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HP 4191A-002 1GHz Impedance Analyser	6500
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HP 16530A DSO Timebase Card	1000
HP 16531A 400MS/S DSO Card	1200
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HP 16555A Timing Analysis Module	1900
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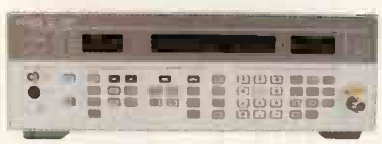
HP 436A RF Power Meter with option 022	750
HP 437B RF Power Meter	1250
HP 438A Dual Channel RF Power Meter	1950
HP 70100A 100KHz to 50GHz Power Meter Module	1000
Various HP 848x Power Sensors (from)	395
Various Marconi 69xx RF Power Sensors (from)	350
Marconi 6960A Power Meter	750
Wandel & Goltermann OLP-2 Optical Power Meter	900

POWER SUPPLIES

HP 6282A / 005 / 028 10V 10A DC Power Supply	150
HP 6284A 005/028 20V/3A DC Power Supply	150
HP E3615A 20V/3A DC Power Supply	195
HP E3631A 25V 5A DC PSU	650
Hunting.Hivolt Series 250 50kV, 5mA Power Supply	975



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HP 8082A Pulse Generator	850

RF SWEEP GENERATORS

Wavetek 2001 1.4GHz Sweep Generator	1300

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Anritsu MS612A 50Hz to 5.5GHz Spectrum Analyser	6500
HP 3561A 100KHz Dynamic Signal Analyser	3250
HP 3562A 2-Channel 100KHz Dynamic Signal Analyser	4500
HP 35660A 102.5KHz Dynamic Signal Analyser	2950
HP 35665A 102.4KHz Dynamic Signal Analyser	9750
HP 3585A 40MHz Spectrum Analyser	4500
HP 70000 22GHz Spectrum Analyser System	14500
HP 70000 2.9GHz Spectrum Analyser c/w Tracking Gen	13500
HP 8560A 2.9GHz Spectrum Analyser	10500
HP 8561A 1KHz-6.5GHz Spectrum Analyser	10850
HP 8562B 22GHz Spectrum Analyser	17500
HP 8591A / 021 1.8GHz Spectrum Analyser	4650
HP 8592A /021 Spectrum Analyser	9500
HP 8593A 22GHz Spectrum Analyser	12500
HP 8596E 12.8GHz Spectrum Analyser	19500
HP 8901A 1.3GHz Modulation Analyser	1250
HP 8903B 20Hz To 100KHz Audio Analyser	2750
Tektronix 492-02 21GHz Spectrum Analyser	4500
Lindos LA100 Audio Analyser (inc. LA101 & LA102)	2750

SIGNAL GENERATORS

Advantest R4262 4.5GHz Signal Generator	16500
HP 8642A /001 1GHz High Performance Signal Gen	2500
HP 8648B 2GHz Signal Generator	3650
HP 8656B /001 1GHz Synthesised Signal Gen	1150
HP 8657B 2GHz Signal Generator	4250
HP 8673C 0.05-18.6GHz Synthesised Signal Generator	19800
HP 8673G 26GHz Synthesised CW Signal Generator	5400
Marconi 2017 1GHz Low Noise Signal Generator	1000
Marconi 2019A 1GHz Signal Generator	1000
Marconi 2022 1GHz Signal Generator	650
Marconi 2031 2.7GHz Signal Generator	6950
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Anritsu MD6420A Data Transmission Analyser	3500
Anritsu MP1520B PDH Analyser	4300
Anritsu MS371A PCM Frame Analyser	4500
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HP 37732A / 005 8MBPS Telecoms/Datacoms Analyser	7950
HP 3788A/001 2MBPS Error Performance Analyser	2950
HP 4934A / 001 Tims Test Set With Battery Pack	3200
Marconi Triton Signalling Test Set	8500
Marconi 2840A Handheld 2MBPS BERT Tester	1950
Trend Aurora Duet Handheld ISDN Tester	3950
TTC Fireberd 6000 c/w Jitter (interfaces available)	4950
W&G PFA-35 2MBPS Communications Tester	4950
Wandel & Goltermann DST-1 E&M Signalling Tester	1250
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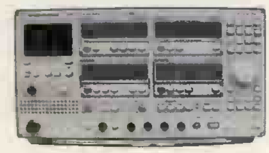

WIRELESS

HP 11757A / 70/140MHz Multipath Fading Simulator	4500
HP 83220A /022 PCS/DCS1800 Test Set	3950
HP 83220E PCS/DCS1800 MS Test Set	3500
HP 8920A 1GHz Radio Comms Test Set (Many configs)	3950
HP 8920A /2/3/4/5 1GHz Radio Comms Test Set	4950
HP 8920B Radio Comms Test Set (Many configs)	12500
HP 8922M GSM Test Set	9500
Marconi 2955 Radio Comms Test Set	1950
Marconi 2955B Radio Comms Test Set	3500
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Racal 6103 /001/002 GSM/DCS Test Set c/w SMS/Fax	9750
R&S CMTxx Radio Test Set - Various Models (from)	2250
R&S CTS55 Digital Radio Tester	4250
Schlumberger 4015 1GHz Radio Comms Test Set	4950
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–40 to +85°C; standard range is –20 to +70°C.
It operates at 3.3V and uses a CMOS IC inside with an output enable function. Besides the ADSL-frequency of 35.328MHz, it is available from 16 to 41MHz and comes in a 7 by 5mm ceramic package.
Epson Electronics
Tel: 00 49 89 14005 277
Web: www.epson-electronics.de

Micro interfaces on 0.635mm pitch

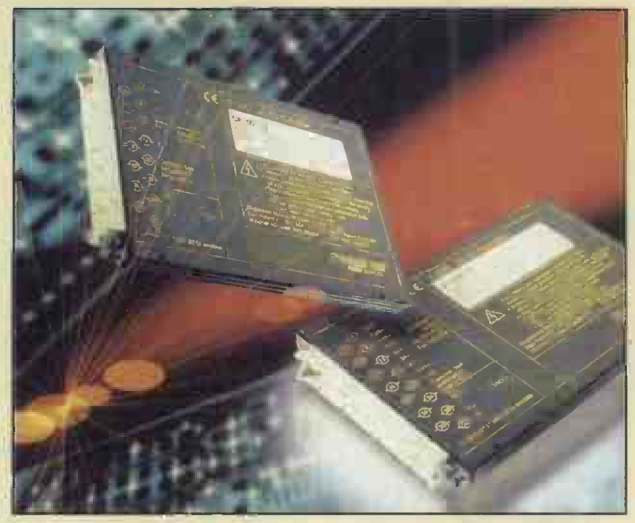
Samtec has introduced matched impedance micro interfaces in the MIT and MIS series on 0.635mm pitch with 38, 76, 114, 152, 190, 228 and 266 I/Os and a discrete ground plane with through-hole leads on 2.54mm pitch. They are tested for 50 and 75Ω systems for impedance, VSWR, attenuation, crosstalk, propagation delay and rise time at frequencies from 10MHz to 1GHz. Spacings between boards can be 5, 8 or 11mm. There is a choice of full surface mount, micro pitch interfaces for quiet and high speed applications, including 0.4mm pitch with 80 I/Os (QTF and QSF series), 0.5mm pitch with up to 300 I/Os (QTH and QSH), 0.635mm pitch with up to 250 I/Os (QTS and QSS) and 0.8mm pitch with up to 200 I/Os (QTE and QSE).
Samtec
Tel: 01236 739292
Web: www.samtec.com

Pushbutton operates passenger door panel

EAO has introduced a passenger door operating panel for use in refurbishing rolling stock. The door operating pushbutton panel has switches flush mounted or fitted with bezels. Bezels and actuators are available in various colours, and the pressel area can be supplied with raised symbols or engraved to requirements.
EAO Group
Tel: 01444 236000
Web: www.eao-group.com

Debugging support for H8/2664F Micro

Hitachi has released the E10T on-chip debugging tool that provides debugging support for the H8/3664F 16-bit flash microcontroller. The device has a five-pin interface that provides access to the on-chip debug system, letting the user set breakpoints on-chip and access an on-chip software trace facility. It is available in PCI and PCMCIA versions, and can be used in development and the final production version. The five pins comprise three I/O pins with NMI and reset functions. The tool provides the user with a hardware breakpoint on address and data, three-level branch trace, single stepping at C and assembly level, and the ability to download applications and program on-chip flash. The HDI source level debugger gives the user high level debugging in C with C level, mixed or assembly level display options. It can also run user



180W converter suits 19in racks

The 180W P-series DC/DC converter from Power-One measures 4TE wide and is suitable for 19in. rack systems. It provides single, dual, triple or quadruple outputs with voltages from 3.3 to 24V. Input voltage range is 16 to 150V DC across five models and they are protected against surges and transients occurring on the source line. The fully enclosed, black coated aluminium case includes an H15 or H1552 type connector and acts as heatsink and RFI shield. The units incorporate input and output filtering, over-temperature protection and input over and undervoltage lock-out. Each can be used as a modular power supply or as a part of a distributed power supply system.
Power One
Tel: 00 49 7666 931 962
Web: www.power-one.com

SMALL SELECTION ONLY LISTED - EXPORT TRADE AND QUANTITY DISCOUNTS - RING US FOR YOUR REQUIREMENTS WHICH MAY BE IN STOCK

HP New Colour Spectrum Analysers LAST FEW ONLY
HP141T+ 8552B IF + 8553B RF -1KHZ -110Mc/s - £500.
HP141T+ 8552B IF + 8554B RF -100KHz -1250M - £600.
HP141T+ 8552B IF + 8556A RF - 20Hz-300KHz - £400.
HP141T+ 8552B IF + 8555A 10 MC/S-18GHzS - £1000.
HP8443A Tracking Gen Counter 100KHz-110Mc/s - £200
HP8445B Tracking Preselector DC to 18GHz - £250.
HP8444A Tracking Generator • 5-1300Mc/s - £450.
HP8444A OPT 059 Tracking Gen • 5-1500Mc/s - £650.
HP35601A Spectrum Anz Interface - £300.
HP8970A Noise Figure Meter + 346B Noise Head - £3k.
HP8755A+B+C Scalar Network Anz PI - £250 + MF 180C - Heads 11664 Extra - £150 each.
HP3709B Constellation ANZ £1,000.
HP11715A AM-FM Test Source - £350.
FARNELL TVS70MKII PU 0-70V 10 amps - £150.
MARCONI 6500 Network Scaler Anz - £500. Heads available to 40GHz many types in stock.
Mixers are available forANZs to 60GHz.
Marconi TF2374 Zero Loss Probe - £200.
Racal/Dana 2101 Microwave Counter - 10Hz-20GHz - with book as new £2k.
Racal/Dana 9303 True RMS Levelmeter + Head - £450.
TEKA6902A also A6902B Isolator - £300-£400.
TEK CT-5 High Current Transformer Probe - £250.
HP Frequency comb generator type 8406 - £400.
HP Sweep Oscillators type 8690 A+B + plug-[illegible] from [illegible] to 18GHz also 18-40GHz.
HP Network Analyser type 8407A + 8412A + 8601A [illegible]0Kc/s - 110Mc/s - £500 - £1000.
HP 8410-A-B-C Network Ana[illegible] [illegible] GHz or 18 GHz - plus most other units and displays used in this set-up - 8411a-8412-8413-8414-8418-8740-8741-8742-8743-8[illegible] 8650. From £1k.
Racal/Dana 9301A-9302 RF millivoltmeter - 1[illegible]z - qty stock £250-£400
Racal/Dana Modulation Meter Type 9009-9008 - [illegible] 1.5GHz - £150/£250 - 9009A £350.
Marconi RCL Bridge type TF2700 - £150.
Marconi Microwave 6600A 1 sweep osc., mainframe with 6650PI - 18-26.5 GHz or 6651 PI - 26.5-40GHz £750 or PI only £600. MF only £250.
Gould J3B test oscillator + manual - £150.
Barr & Stroud Variable filter EF3 0.1Hz-100Kc/s + high pass + low pass - £150, other makes in stock.
Racal/Dana 9300 RMS voltmeter - £250.
HP 8750A storage normalizer - £400 with lead + S.A. or N/A
Marconi mod meters type TF2304 - £250 - TF2305 - £1,000.
Racal/Dana counters-99904-9905-9906-9915-9916-9917-9921 50Mc/s-3GHz - £100 - £400 - all fitted with FX standards.
HP180TR. HP181T, HP182T mainframes £300 - £500.
HP432A-435A or B-436A-power meters + powerheads to 60GHz - £150 - £1750 - spare heads available.
HP3586A or C selective level meter - £500.
HP86222A+B Sweep PI -01-2.4GHz + ATT £1000-£1250.
HP86290A+B Sweep PI-2 - 18GHz - £1000 - £1250.
HP8620C Mainframe - £250. IEEE £350.
HP8165A Programmable signal source - 1MHz - [illegible] - £1k.
HP3455/3456A Digital voltmeter - £400.
HP5370A Universal time interval counter - £1[illegible]
HP5335A Universal counter - 200Mc/s £1000
TEKTRONIX 577 Curve tracer + adaptors - £900.
TEKTRONIX 1502/1503 TDR cable test set - £400.
HP8699B Sweep PI YIG oscillator .01 - 4GHz - £300. 8690B MF-£250. Both £500.
Dummy Loads & Power att up to 2.5 kilowatts FX up to 18GHz - microwave parts new and ex equipt - relays - attenuators - switches - waveguides - Yigs - SMA - APC7 plugs - adaptors etc. qty. in stock.
B&K Items in stock - ask for list.
Power Supplies Heavy duty + bench in stock - Farnell - Weir - Thurlby - Racal etc. Ask for list. [illegible] in stock, all types to 400 amp - 100Kv.
HP8405A Vector voltmeter - late colour - £400.
HP8508A Vector voltmeter - £2500.

LIGHT AND OPTICAL EQUIPMENT

Anritsu ML93A & Optical Lead Power Meter - £250.
Anritsu ML93B & Optical Lead Power Meter - £350.
Power Sensors for above MA96A - MA98A - MA913A - Battery Pack MZ95A.
Anritsu MW97A Pulse Echo Tester.
PI available - MH914C 1.3 - MH915B 1.3 - MH913B 0.85 - MH925A 1.3 - MH929A 1.55 - MH925A 1.3GI - MH914C 1.3SM - £500 + one P.I.
Anritsu MW98A Time Domain Reflector.
PI available - MH914C 1.3 - MH915B 1.3 - MH913B 0.85 - MH925A 1.3 - MH929A 1.55 - MH925A 1.3GI - MH914C 1.3SM - £500 + one P.I.
Anritsu MZ100A E/O Converter.
+ MG912B (LD 1.35) Light Source + MG92B (LD 0.85) Light Source £350.
Anritsu MZ118A O/E Converter.
+MH922A 0.8 O/E unit + MH923 A1.3 O/E unit £350.
Anritsu ML96B Power Meter & Charger £450.
Anritsu MN95B Variable Att. 1300 £100.
Photo Dyne 1950 XR Continuous Att. 1300 - 1500 £100.
Photo Dyne 1800 FA. Att £100.
Cossor-Raytheon 108L Optical Cable Fault Locator 0-1000M 0-10kM £200.
TEK P6701 Optical Converter 700 MC/S-850 £250.

TEK OF150 Fibre Optic TDR – £750.
HP81512A Head 150MC/S 950-1700 £250.
HP84801A Fibre Power Sensor 600-1200 £250.
HP8158B ATT OPT 002+011 1300-1550 £300.
HP81519A RX DC-400MC/S 550-950 £250.
STC OFR10 Reflectometer – £250.
STC OFSK15 Machine jointing + eye magnifier – £250.

MISCELLANEOUS ITEMS

HP 4261 LCR meter – £650.
HP 4274 FX LCR meter – £1,500.
HP 3488 Switch Control Unit + PI Boards – £500.
HP 75000 VXI Bus Controllers + E1326B [illegible] quantity.
HP 83220A GSM DCS/PCS 1805-1990MC [illegible]ertor for use with 8922A – £2,000.
HP 1630-1631-1650 Logic ANZ's in stock.
HP 8754A Network ANZ 4-1300MC/S + 8502A + cables – £1,500.
HP 8754A Network ANZ H26 4-2600MC/S + 8502A + Cables – £2,000.
HP 8350A Sweeper MF + 83540A PI 2[illegible]GHZ + 83545A PI 5.9-12.4GHZ all – £3,500.
HP MICROWAVE TWT AMPLIFIER 489A [illegible]GHZ-30DB – £400.
HP PREAMPLIFIER 8447A 0.1-400MC/S – £200. Dual – £300.
HP PREAMPLIFIER 8447D 0.01-1.3GHZ – £400.
HP POWER AMPLIFIER 8447E 0.01-1.3GHZ – £400.
HP PRE + POWER AMPLIFIER 8447F 0.01-1.3GHZ – £600.
HP [illegible]74 Gain Phase Meter 1HZ-13MC/S OPT 001 Dual – £400.
MARCONI 2305 Modulation Meter-50KHZ-2.3 GHZ – £1,000.
MARCONI 2610 True RMS Meter – £450.
MARCONI 893B AF Power Meter (opt Sinad filter) – £250-£350.
MARCONI 6950-6960B Power Meters + Heads – £400-£900.
MARCONI SIGNAL SOURCE-6055-6056-6057-6058-6059 – FX range 4-18GHZ– £250-£400.
RACAL 1792 COMMUNICATION RX – £600 early – £1,000 – late model with back lighting and byte test.
RACAL 1772 COMMUNICATION RX – £400-£500.
PLESSEY PR2250 A-G-H COMMUNICATION RX – £500-£[illegible]
TEK MODULE MAINFRAMES – TM501-502-503-504-506-TM5003-5006.
TEK PI 5010-M1 – Prog Multi Interface – £250 – [illegible] Prog [illegible]MC/S Function Gen – £400 – S1 Prog Scanner – £250 – DM Prog DMM – £400.
TEK 7000 OSCILLOSCOPE MAINFRAMES – 7603-7623-7633-7834-7854-7904-7904A-7104 – £150-£1,000.
TEK 7000 PI's – 7A11-7A12-7A13-7A18-7A19-7A22-7A24-7A26-7A29-7A42-7B10-7B15-7B53A-7B80-7B85-7B92A-7D15-7D20.
TEK 7000 – 7S11-7S12-7S14-7M11-S1-S2-S3A-S4-S5-S6-S51-S53-S54.
HP POWER SUPPLIES – 6621A-6623A-6624A-6632A-6652[illegible] Qtys available [illegible] 100 types £POA.

SPECIAL OFFERS

MARCONI 2019A SYNTHESIZED SIGNAL GENERATORS – 80KC/S-1040MC/S – AM-FM – £400 inc instruction book – tested.
MARCONI 2022E SYNTHESIZED SIGNAL GENERATOR – 10KC/S-1.01GHZ AM-FM – £500 inc. instruction book – tested.
R&S APN 62 LF Sig Gen 0.1Hz – 260KHz [illegible] book – £250.
HP 8922 RADIO COMMUNICATION Test Sets [illegible]H-M. Options various. £2,000-£[illegible]000 each.
HP 83220 A-E GMS UNITS for above £1,000-£1,500.
WAVE TECK SCLUMBERGER 4931 Radio Communication Test Set internal spectrum an[illegible] £1,800-£2,000.
ANRITSU MS555A2 RADIO Comm anz. TO1000MC/S. No CR Tube in this model £450.
TEK 2445A-4CH-150OMCS SCOPE + new X1-X10 probe instruction book. £500 each.
HP4193A VECTOR IMPEDANCE METER + Probe Kit – 400KHZ to 110 MC/S. 3.5k.

WE KEEP IN STOCK HP and other makes of RF Frequency doublers which when fitted to the RF output socket of a S/Generator doubles the output frequency EG.50-1300MC/S to 50-2600MC/S price from £250 – £450 each.

SPECTRUM ANALYZERS

HP 3580A 5HZ-50KHZ – £750.
HP 3582A Dual 0.2HZ-25.5KHZ – £1,500.
HP 3585A 20HZ-40MC/S – £3,500.
HP 3588A 10HZ-150MC/S – £7,500.
HP 8568A 100HZ-1.5GHZ – £3,500.
HP 8568B 100HZ-1.5GHZ – £4,500.
HP 8590B 9KC/S-1.8GHZ – £4,500.
HP 8569B 10MC/S (0.01-22GHZ) – £3,500.
HP 3581A Signal Analyzer 15HZ-50KHZ – £400.
TEK492 50KHZ-21GHZ OPT 2 – £2,500.
TEK492P 50KHZ-21GHZ OPT 1-2-3 – £3,500.
TEK492AP 50KHZ-21GHZ OPT 1-2-3 – £4,000.
TEK492BP 50KHZ-21GHZ – £3,000-£4,000.
TEK495 100KHZ-1.8GHZ – £2,000.
HP 8557A 0.01MC/S-350MC/S – £500 + MF180T or 180C – £150 – 182T – £500.
HP 8558B 0.01-1500MC/S – £750 – MF180T or 180C – £150 – 182T – £500.

HP 8559A 0.01-21GHZ – £1,000 – MF180T or 180C – £150 – 182T – £500.
HP 8901A AM FM Modulation ANZ Meter – £800.
HP 8901B AM FM Modulation ANZ Meter – £1,750.
HP 8903A Audio Analyzer – £1,000.
HP 8903B Audio Analyzer – £1,500.
MARCONI 2370 SPECTRUM ANALYZERS – HIGH QUALITY – DIGITAL STORAGE – 30HZ-110MC/S Large qty to clear as received from Gov – all sold as is from pile complete or add £[illegible] for basic testing and adjustment – callers preferred – pick your own from over sixty units – discount on qtys of five.
A EARLY MODEL GREY – horizontal alloy cooling fins – £200.
B LATE MODEL GREY – vertical alloy cooling fins – £300.
C LATE MODEL BROWN – as above (few only) – £500.

OSCILLOSCOPES

TEK 465-465B 100MC/S + 2 probes – £250-£300.
TEK 466 100MC/S storage + 2 probes – £200.
TEK 475-475A 200MC/S-250MC/S + 2 probes – £300-£350.
TEK 2213-2213A-2215-2215A-2224-2225-2235-2236-2245-60-100MC/S – £250-£400
TEK 2445 4ch 150MC/S + 2 probes – £450.
TEK 2445A 4ch 150MC/S + 2 probes – £600.
TEK 2445B 4ch 150MC/S + 2 probes – £750.
TEK 468 D.S.O. 100MC/S + 2 probes – £500.
TEK 485 350MC/S + 2 probes – £550.
TEK 2465 4ch-300MC/S – £1,150.
TEK 2465A 4ch-350MC/S – £1,550.
TEK 2465ACT 4ch-350MC/S – £1,750.
TEK D.S.O. 2230 -100MC/S + 2 probes – £1,000.
TEK D.S.O. 2430 -150MC/S + 2 probes – £1,250.
TEK D.S.O. 2430A -150MC/S + 2 probes – £1,750.
TEK D.S.O. 2440 -300MC/S + 2 probes – £2,000.
TEK TAS 475-485 - 100MC/S-200MC/S 4 ch + 2 probes – £900-£1.1K.
HP1740A – 100MC/S + 2 probes – £250.
HP1741A – 100MC/S storage + 2 probes – £200.
HP1720A – 1722A - 1725A - 275MC/S + 2 probes – £300-£400.
HP1744A – 100MC/S storage - large screen – £250.
HP1745A – 1746A - 100MC/S - large screen – £350.
HP54100A – 1GHz digitizing – £500.
HP54200A – 50MC/S digitizing – £500.
HP54501A – 100MC/S digitizing – £500.
HP54100D – 1GHZ digitizing – £1,000.

MICROWAVE COUNTERS – ALL LED READOUT

EIP 351D Autohet 20Hz-18GHz – £750.
EIP 371 Micro Source Locking – 20Hz-18GHz – £850.
EIP 451 Micro Pulse Counter – 300MC/S-18GHz – £700.
EIP 545 Microwave Frequency Counter – 10Hz-18GHz – £1K.
EIP 548A Microwave Frequency Counter – 10HZ-26.5GHz – £1.5K.
EIP 575 Microwave Source Locking – 10Hz-18GHz – £1.2K.
EIP 588 Microwave Pulse Counter – 300MC/S-26.5GHz – £1.4K.
SD 6054B Micro Counter 20HZ-24GHZ – SMA Socket – £800.
SD 6054B Micro Counter 20HZ-18GHZ – N Socket – £700.
SD 6054D Micro Counter 800MC/S-18GHz – £600.
SD 6246A Micro Counter 20Hz-26GHz – £1.2K.
SD 6244A Micro Counter 20Hz-4.5GHz – £400.
HP5352B Micro Counter OPT 010-005-46GHz – new in box – £5k.
HP5340A Micro Counter 10HZ-18GHz – Nixey – £500.
HP5342A Micro Counter 10HZ-18-24GHz – £800-£1K – OPTS 001-002-003-005-011 available.
HP5342A + 5344S Source Synchronizer – £1.5K.
HP5345A 500MC/S 11 Digit LED Readout – £400.
HP5345A + 5354A Plugin – 4GHz – £700.
HP5345A + 5355A Plugin with 5356A 18GHz Head – £1K.
HP5385A 1GHz 5386A-5386A 3GHz Counter – £1K-£2K.
Racal/Dana Counter 1991-160MC/S – £200.
Racal/Dana Counter 1992-1.3GHz – £600.
Racal/Dana Counter 9921-3GHz – £350.

SIGNAL GENERATORS

HP8640A – AM-FM 0.5-512-1024MC/S – £200-£400.
HP8640B – Phase locked – AM-FM-0.5-512-1024MC/S – £500-£1.2K. Opts 1-2-3 available.
HP8654A – B AM-FM 10MC/S-520MC/S – £300.
HP8656A SYN AM-FM 0.1-990MC/S – £900.
HP8656B SYN AM-FM 0.1-990MC/S – £1.5K.
HP8657A SYN AM-FM 0.1-1040MC/S – £2K.
HP8657B SYN AM-FM 0.1-2060MC/S – £3K.
HP8660C SYN AM-FM-PM-0.01-1300MC/S-2600MC/S – £2K.
HP8660D SYN AM-FM-PM-0.01-1300MC/S-2600MC/S – £3K.
HP8673D SYN AM-FM-PM-0.01-26.5 GHZ – £12K.
HP3312A Function Generator AM-FM 13MC/S-Dual – £300.
HP3314A Function Generator AM-FM-VCO-20MC/S – £600.
HP3325A SYN Function Generator 21MC/S – £800.
HP3325B SYN Function Generator 21MC/S – £2K.
HP8673-B SYN AM-FM-PH 2-26.5 GHz – £6.5K.
HP3326A SYN 2CH Function Generator 13MC/S-IEEE – £1.4K.
HP3336A-B-C SYN Func/Level Gen 21MC/S – £400-£300-£500.
Racal/Dana 9081 SYN S/G AM-FM-PH-5-520MC/S – £300.
Racal/Dana 9082 SYN S/G AM-FM-PH-1.5-520MC/S – £400.
Racal/Dana 9084 SYN S/G AM-FM-PH-.001-104MC/S – £300.
Racal/Dana 9087 SYN S/G AM-FM-PH-.001-1300MC/S – £1K.
Anritsu MG3601A SYN AM-FM 0.1-1040MC/S – £1.2K.

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Johns Radio, Whitehall Works, 84 Whitehall Road East, Birkenshaw, Bradford BD11 2ER. Tel: (01274) 684007. Fax: 651160

CIRCLE NO. 115 ON REPLY CARD

Please quote *Electronics World* when seeking further information

code in real-time and single step, with step-over, step into and step out of functions. HDI provides a register display, three level branch trace, C level watch-points and a memory window.

Hitachi
Tel: 01628 585163
Web: www.Hitachi.co.jp

Half-pitch DIP switch has 1.5mm profile

Omron's A6H half-pitch surface mount DIP switch has a 1.5mm profile, making it suitable for space sensitive applications where occasional or installation-specific setting changes are required, such as in security and vending equipment. It can be reflow soldered and is available in four and eight pole versions, with a six pole version planned.

Measuring 4.5mm wide, and 6.31, 8.85 and 11.39 mm long for four, six and eight pole models respectively, it will switch 25mA at 24V DC and provides 100MΩ insulation resistance. Available with tape seal for immerse cleaning and on embossed tape reels for automatic pick-and-place processes, the switch can operate at –20 to +70°C with zero icing or condensation.

Omron
Tel: 020 8450 4646
Web: www.omron.co.uk

Sun SPARC cluster system launch

I-Bus/Phoenix has launched the G2077 NEBS tested Sun Sparc cluster system, which is a CompactPCI platform equipped with two independent system boards, Sun Microsystems CP1500 SBCs, configured as a cluster server.

The system comes fully integrated with dual hot-swap, Intraserver ITI-8241C-S quad Ethernet, dual Ultra-2 SCSI and 4MB video board, 10 user expansion slots, as well as Sun Cluster 2.2 or Veritas HA Foundation Suite installed.

I-Bus Phoenix
Tel: 023 9242 4800
Web: www.sales@ibus .co.uk

Memory products with dual-bank flash

Silicon Storage Technology has introduced four ComboMemory products that incorporate a dual-bank flash memory architecture and allow for concurrent operations between the two internal flash memory banks and the SRAM. They contain 16Mbit flash and 2 or 4Mbit SRAM. Housed in an 8

by 10mm BGA package with stacked-die technology, the devices are for mobile wireless communications applications such as Bluetooth and WAP-enabled mobile phones and Internet-based PDAs. The data can be read from either bank while an erase or program operation is in progress in the opposite bank. The flash memory is partitioned into 4 and 12Mbit banks with top or bottom sector protection options for storing boot codes and configuration data.

Silicon Storage Technology
Tel: 01628 585163
Web: www.ssti.com

Line interface hybrids support ISDN S chips

Advanced Power Components has released line interface hybrids to support Infineon ISDN S interface chips. The hybrids contain the interface components required between the chip and the telephone connector port, including line isolation and secondary protection. The Infineon transceivers operate on a 3.3V supply. There are three package styles for the hybrids. The standard sized APC14131E provides 1500V supplementary isolation. The APC8305 mini hybrid and PCMCIA-height APC8403 provide basic isolation. They are verified against the requirements of layer one testing.

APC
Tel: 01634 290588
Web: www.apc-plc.com

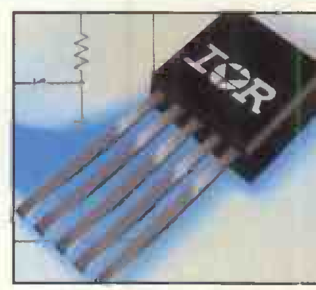
Data recovery up to 2.5Gbit/s

Micrel Semiconductor has developed a clock and data recovery IC that combines onboard clock synthesis with operation from 28Mbit/s to 2.5Gbit/s. The AnyRate SY87702L can recover data from serial streams running at any rate up to OC-48. The device can handle commonly used communications speeds and protocols including OC-3 STM-1, OC-12 STM-4, OC-48 STM-16, Fast Ethernet, Gigabit Ethernet, FDDI, Escon, fibre channel, 2x fibre channel, Infiniband and digital video rates outlined in SMPTE 259 and 292. It also supports proprietary rates.

Micrel Semiconductor
Tel: 0207 823 3224
Web: www.micrel.com

200V integrated SMPS in TO-262 packaging

International Rectifier has introduced a 200V integrated switched mode power supply IC in a TO-262 package. The IR4007 is a DC-to-DC converter control IC with an avalanche-characterised Mosfet built-in. The device can handle momentary transient voltages above its rated breakdown voltage as long as they do not exceed the specified avalanche energy limit of 100mJ. It has



proprietary input and output feedback circuits. Topologies include 48V in, 3.3V output flyback converters up to 5A. Two operating modes are built-in – quasi-resonant mode and pulse ratio control mode. In quasi-resonant mode the transformer primary inductance and an additional capacitor provide a resonant signal to trigger the Mosfet when the drain-source voltage is at a minimum, reducing power dissipation.

International Rectifier
Tel: 0208 645 8001
Web: www.irf.com

40 MHz comms chip

Cypress has introduced a 40MHz intelligent control communications processor for Lonworks control networks. Each device contains three 8-bit

CPUs, onboard memory, 11 general purpose I/O pins and a complete, interoperable implementation of the Ansi EIA709.1-A-1999 control network protocol standard. The CY7C53120E4 is for embedded Internet systems that remotely monitor and control electrical devices through online connections. It integrates 4kbyte of onboard E^{2}PROM and is made using a 0.35μm silicon oxide, nitride oxide silicon process.

Cypress Semiconductor
Tel: 01707 378700
Web: www.cypress.com

Microcontroller peaks at 50Mips

Dallas Semiconductor has announced the DS89C420 8051-compatible microcontroller with peak processing speeds of one machine instruction per clock cycle, or 50Mips, at a maximum clock speed of 50MHz. It has 16kbyte of flash memory that allows programming in any of three modes – in-system, in-application or standard parallel programming. The device is pin and instruction-set compatible with existing 8051-based systems.

Dallas Semiconductor
Tel: 00 1 972 371 3832
Web: www.dalsemi.com

Flexible connectors pitch at 2.54mm

Robinson Nugent has announced the FPC 5 series of flexible printed circuit connectors on straight (vertical entry) or right angle (side entry) mounting. With a contact pitch of 2.54mm and pin counts from three to 21, they are tin plated phosphor bronze and rated at

1A. The contacts are housed in a UL rated PBT body and have an operating range of –40 to +105°C. The flex circuit interface wipes on the contacts removing debris and oxidation on the contact surface. Also available is a 1.25mm pitch side entry version.

Robinson Nugent
Tel: 01227 794495
Web: www.robinsonnugent.com

SBC supports AMD's 133MHz Elan 520 CPU

The Microspace PC/104 MSM586SV single board computer from Digital-Logic supports an AMD 133MHz Elan 520 CPU, 10/100BaseT Ethernet network interface, SXGA 69000 graphics controller and PCI video data path. Interfaces include

keyboard and mouse, four serial RS232Cs and a floppy disk port. The board also has an IDE hard disk connection and a printer interface. Features include disk-on-chip 2000 slot, watchdog and E^2PROM and IrDA support.

Digital-Logic
Tel: 00 41 32 681 58 00
Web: www.digitallogic.ch

Prescaler performs divide-by-two function

SiGe Microsystems has introduced the D602 prescaler performing a divide-by-two function for frequency synthesisers up to 6GHz. Consuming 16mA from a 3V supply, it has a residual phase noise performance of –145dBc/Hz at 1kHz offset. Power-down mode, reduces supply current to less than 100nA and input sensitivity is typically –27dBm. Applications include local multipoint distribution systems, satellites and 5GHz WLANs.

SiGe Microsystems
Tel: 01223 597806
Web: www.sige.com

WAN switch doubles the bandwidth

A wide area network (WAN) access switching chip announced by Mitel Semiconductor, the MT90866 switches up to 4096 by 2432 channels, twice the bandwidth of alternative designs, claims the company. The device manages voice, data and video traffic in large-scale carrier-class equipment that bridges information on packet-based data networks to the circuit-switched public telephone network. When different types of packet-based and circuit-switched traffic converge onto the network infrastructure, the result can be delay, jitter, echo and data loss – conditions that degrade quality of service. According to Mitel, its switch incorporates a phase locked loop (PLL), a block of circuitry that minimises these effects by optimising system timing and synchronisation. It is a specialised 3V time division multiplex (TDM) timeslot interface that supports voice, data and video traffic. The device incorporates three switching functions into a single product: 4096 by 2432 channels between backplane and the local streams, 2432 by 2432 channels among local streams, and 2048 by 2048 channels among backplane streams.

Mitel Semiconductor
Tel: 01793 518528
`Web: www.mitel.com

Wirewound inductors from 3.3nH to 4.7µH

NIC has introduced wire-wound SMT inductors for handheld devices, telecoms, high-frequency and wireless designs. Available in D (0805) and C (1008) sizes, the NIN-H units have inductance values from 3.3nH to 4.7µH with current ratings up to 1.0A. Inductance tolerances of 20 and 10 per cent are available, while minimum Q factor is up to 65 and minimum self-resonant frequency up to 6GHz. Maximum DC resistance can be down to 0.08Ω and maximum DC current up to 600mA. Operating range is –40 to +125°C and they come in embossed plastic tape packages for automatic pick-and-place. ■

NIC Eurotech
Tel: 01280 813737
Web: www.niccomp.com

Self on Audio

Douglas Self

The cream of 20 years of Electronics World articles (focusing on recent material)

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Scientific electronics based on empirical data

Douglas Self has been writing for Electronics World and Wireless World over the past 20 years, offering cutting-edge insights into scientific methods of electronics design.

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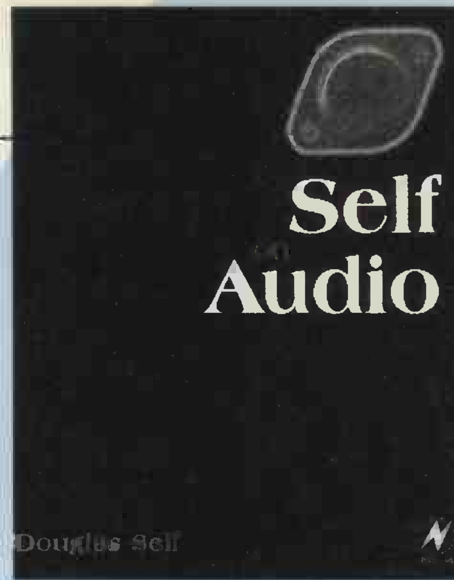

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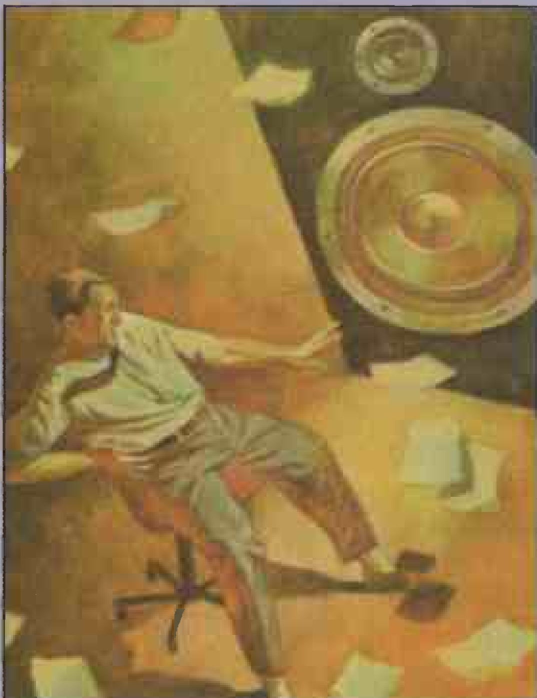


SPEAKERS' CORNER

Dome myths exploded

"All of the problems associated with designing a dome loudspeaker can be solved by relegating it to the function of a dust cap in the middle of a cone", argues John Watkinson.

A reader from Sweden e-mailed to ask what I meant by the term 'dome sound'. And by the time I had finished thinking what to say, I had quite a lot of material.

There is more to building loudspeakers than assembling a collection of parts. In order to obtain high quality, the parts have to work together as a coherent system.

The full frequency range high-quality moving-coil drive unit is an impossibility. That leaves us with the need for at least two drive units in a practical speaker.

I have argued at length that one of the critical aspects of loudspeaker design is to obtain an inaudible crossover so that the existence of two drive units is not revealed in the sound. Inaudible crossover is a performance related term. This means that the crossover network must separate the original wide-band waveform into two sub-bands as a mirror of the way the sound from the two drive units will add back together acoustically.

When designing a real speaker, the designer has to balance the strengths and weaknesses of available components in such a way that the strengths are revealed and the weaknesses are concealed or eliminated. The design process is an art form. It requires a degree of compromise so that the resulting speaker has a balanced performance in various domains with no obvious flaws.

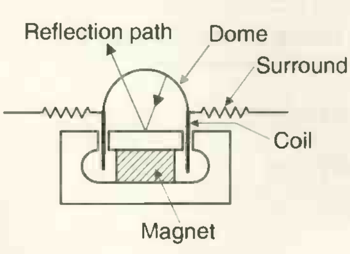


Fig. 1. Basic construction of a dome tweeter. Such speakers invariably have a solid centre piece that reflects sound from the back of the dome.

The dome's inherent limitations

The problem with the dome tweeter is that it severely limits the scope of the speaker designer to make design trade-offs. These limits are not due to any lack of attention to quality, but are inherent in the dome design.

Figure 1 shows a typical dome tweeter. The coil is attached to the perimeter of the dome, the whole of which is suspended by a flexible surround.

Figure 2 reveals one of the myths of the dome. In a) is an ideal sound radiator. It is a sphere that can somehow change its radius as a function of the audio waveform. I would dearly love to know how to do this, but for the moment no-one knows how.

Now the dome looks superficially similar, because it is a section of a sphere, but there the similarity ends. Figure **2b)** shows that the entire spherical section simply moves from side to side along a single axis. Acoustically this is totally different from **2a)**. The dome is not a spherical radiator

Looks can be deceiving. It is physics and performance that matter though – not appearance. For example, few people realise that the old air-cooled Volkswagen microbus had a better drag coefficient than the E-type Jaguar.

How the dome works

In order to determine how a dome works, it is necessary to treat the surface as an infinite number of point radiators and then to integrate the contributions from each at the listening point. For the time being, I will assume that the dome is rigid.

Figure 3a) shows that the

difference in path length between the centre and the perimeter causes an aperture effect. When the height of the dome is half a wavelength, there will be significant cancellation. Figure 3b) shows that if the dome is concave, exactly the same response is obtained.

Practical domes are made small enough and shallow enough so that the first null is outside the audio band. Within these limits the actual shape of the dome is irrelevant as it could be convex, concave or flat without affecting the frequency response. In fact the dome shape is used for rigidity, and to fool the gullible.

Considering off-axis sound, the rigid dome also suffers an aperture effect. **Figure 4** shows that there will be a response drop when the path length difference from the two sides of the dome approaches half a wavelength. This is classical beaming from a radiator of finite area.

The only solution to obtain radiation over a reasonable angle is to keep the dome small in diameter. A typical figure is 19mm. A radiator as small as 19mm will have a high fundamental resonance, and the small diaphragm area will limit the volume velocity that can be obtained.

All of these factors together mean that the dome tweeter is fundamentally a narrow-band device. If the dome is made larger, it will have more volume velocity and a lower resonance, allowing it to be used at lower frequency. But the greater size will lower the frequency of the first null and make the beaming worse.

The designer of a two-way speaker having a dome tweeter is basically stuck with an impossible choice of crossover frequency in the 1.5 to 2kHz region. The woofer will be asked to operate at frequencies where its diaphragm is far too large and it will beam.

Figure 5a) shows the result. At low frequencies the woofer is omnidirectional. As frequency rises the woofer gets more and more directional until at the crossover the wide directivity of the tweeter takes over. Off-axis, the frequency response is like a dog's hind leg and the reverberant field will be coloured.

Figure 5 also shows the frequency and power responses of almost any small two-way speaker with a dome tweeter. The frequency response usually shows a small dip at the crossover where the designer is trying to make the crossover frequency as low as possible. The power response shows a huge dip at the crossover.

Effectively a huge chunk of the spectrum is missing in the reverberant field. This gives the listener the impression that the audio has literally been cut in half. So much for the inaudible crossover!

In practice, domes can't be rigid because vibrations can travel right across the dome and bounce off the other side. The palliative here is the soft dome which is made from a lossy material. The resultant problem is that at high frequencies the centre of the dome decouples as in Fig. 4b), leaving an annular radiator which has appalling directivity.

Finally, most domes have solid centre-pole pieces. Sound radiated from the back of the dome bounces off the pole piece and passes through the soft dome as shown in Fig. 1. This causes time smear and comb filtering in the frequency response.

All of the problems of a dome can be solved by relegating it to the function of a dust cap in the middle of a cone, but then it wouldn't look right to someone who is judging by appearances. Perhaps the solution is to colour the cone black and leave the dome a bright colour. ■

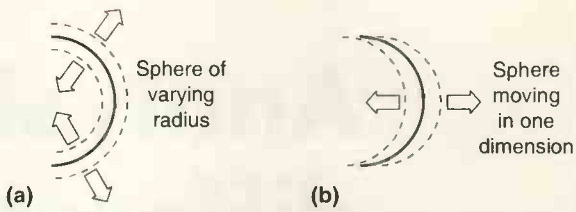


Fig. 2. The myth of the dome. The ideal motion of a), where the diaphragm expands and contracts, is simply not met. You can see from diagram b) that the diaphragm moves in one dimension only.

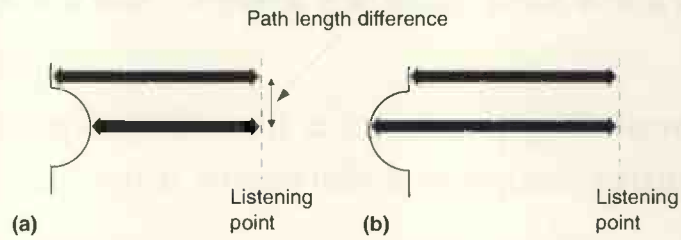


Fig. 3. Whether the dome is convex, as in a), or concave as in b) makes no difference to its performance.

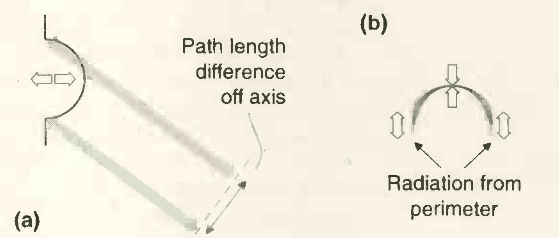
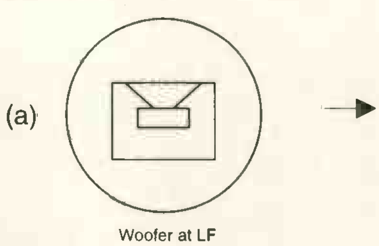
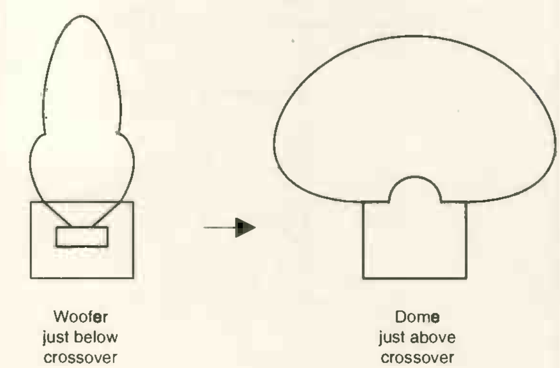
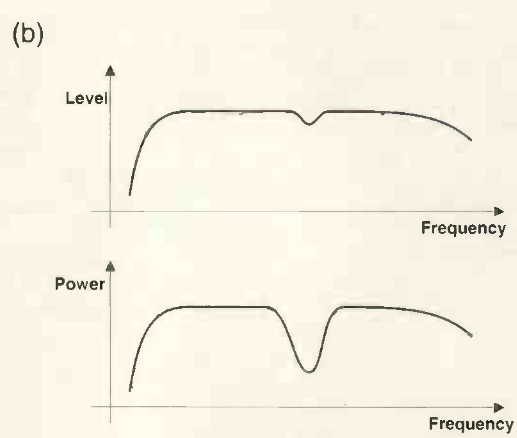


Fig. 4. From a), you can see that the dome must be as small as possible in order to minimise directivity problems. Diagram b) shows how the dome centre decouples at high frequencies.

Fig. 5. At a), the excessive beaming of the woofer at a high crossover frequency causes a large dip in the power response, b). This is audible.







Analysing the differential amplifier

Lionel Gay presents a theoretically exact analysis of the emitter-coupled differential amplifier.

The emitter-coupled differential amplifier evolved from the cathode-coupled differential amplifier and the behaviour of the circuit is well known. See, for example, Slaughter's paper, published in 1956[1].

No approximations are made in the analysis that follows. As a result, it is applicable to any circuit that incorporates a pair of identical active two-port devices that are defined by their identical hybrid parameters. In addition, the analysis can be readily modified to take account of a pair of non-identical active devices.

Consider **Fig. 1**, where V_i, V_{01}, and V_{02} denote signal potentials. Assuming that Tr_1 and Tr_2 have identical h parameters, the task is to set up expressions for V_{01}/V_i, V_{02}/V_i, the input resistance R_i, and the two output resistances R_{01} and R_{02}.

In order to calculate V_{01}/V_i, it is helpful to draw the equivalent circuit of **Fig.** 2. Next, set up expressions for R_J, R_K, and hence V_{01}/V_i, R_N, and R_i. In order to do so, you can make use of the equations listed in Appendix X of reference 2.

Assuming that the signal frequency is low so that the h parameters are real, it follows that,

$$R_j = \frac{h_{ib} + D_{hb}\left(R_5 + R_7\right) + R_7\left(1 + h_{fb} - h_{rb} + h_{ob}R_5\right)}{1 + h_{ob}\left(R_5 + R_7\right)}$$

where,

$$D_{hb} = h_{ib}h_{ob} - h_{fb}h_{rb}$$

Given the common-emitter h parameters, the other two sets of h parameters can be calculated. See, for example, Table A8.1 on page 480 of reference 2.

$$R_K = R_3 + \frac{R_4\left(R_6 + R_j\right)}{R_4 + R_6 + R_j}$$

and,

$$\frac{V_{01}}{V_i} = \frac{-R_2\left(h_{fe} - h_{oe}R_K\right)}{h_{ie} + D_{he}\left(R_2 + R_K\right) + R_K\left(1 + h_{fe} - h_{re} + h_{oe}R_2\right)}$$

Here, R_i is the parallel combination of R_1 and R_N. So,

$$R_i = \frac{R_1 R_N}{R_1 + R_N}$$

and,

$$R_N = \frac{h_{ie} + D_{he}\left(R_2 + R_K\right) + R_K\left(1 + h_{fe} - h_{re} + h_{oe}R_2\right)}{1 + h_{oe}\left(R_2 + R_K\right)}$$

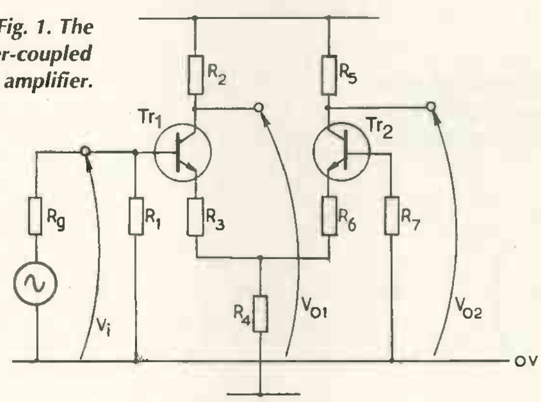
In order to set up an expression for $V_{02} \div V_i$, draw the equivalent circuit shown in **Fig. 3**, where I_1, I_2, I_3, and I_4 denote signal currents.

$$\frac{V_{02}}{V_i} = \frac{-I_4R_5}{V_i} = \frac{-I_4R_5}{I_1R_N} = -\frac{R_5}{R_N}\frac{I_4}{I_3}\frac{I_3}{I_2}\frac{I_2}{I_1}$$

$$\frac{I_4}{I_3} = \frac{h_{fb} - h_{ob}R_7}{1 + h_{ob}\left(R_5 + R_7\right)}$$

$$\frac{I_3}{I_2} = \frac{-R_4}{R_4 + R_6 + R_j}$$

Fig. 1. The emitter-coupled differential amplifier.



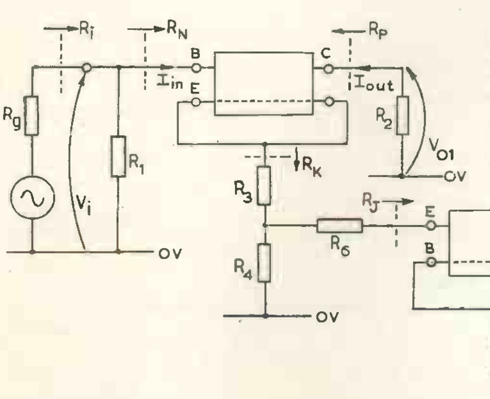


Fig. 2. Equivalent circuit of the emitter-coupled differential amplifier.

and,

$$\frac{I_2}{I_1} = \frac{h_{fc} - h_{oc}R_2}{1 + h_{oc}(R_K + R_2)}$$

So,

$$\frac{V_{02}}{V_i} = \frac{R_5 R_4 (h_{fb} - h_{ob}R_7)(h_{fc} - h_{oc}R_2)}{Y_1}$$

where,

$$Y_1 = R_N(R_4 + R_6 + R_j)(1 + h_{ob}(R_5 + R_7))(1 + h_{oc}(R_K + R_2))$$

R_{01} is the parallel combination of R_P and R_2, see Fig. 2, so that,

$$R_{01} = \frac{R_p R_2}{(R_p + R_2)}$$

where,

$$R_p = \frac{\left(h_{ie} + \frac{R_1 R_g}{(R_1 + R_g)}\right) \times (1 + h_{oe}R_K) + R_K(1 - h_{re})(1 + h_{fe})}{Y_2}$$

and where,

$$Y_2 = D_{he} + h_{oe}\left(\frac{R_1 R_g}{R_1 + R_g} + R_K\right)$$

Resistance R_{02} is the parallel combination of R_T and R_5, **Fig. 3**, so that,

$$R_{02} = \frac{R_5 R_T}{R_5 + R_T}$$

where,

$$R_T = \frac{(h_{ib} + R_A)(1 + h_{ob}R_7) + R_7(1 - h_{rb})(1 + h_{fb})}{D_{hb} + h_{ob}(R_A + R_7)}$$

$$R_A = R_6 + \frac{R_4(R_3 + R_Q)}{R_4 + R_3 + R_Q}$$

$$R_Q = \frac{\left(h_{ic} + \frac{R_1 R_g}{R_1 + R_g}\right)(1 + h_{oc}R_2) + R_2(1 - h_{rc})(1 + h_{fc})}{Y_3}$$

Where,

$$Y_3 = D_{hc} + h_{oc}\left(\frac{R_1 R_g}{R_1 + R_g} + R_2\right)$$

With reference to Fig. 2 , where I_{in} and I_{out} denote signal currents, the mutual conductance of the circuit is, by definition, given by $g_{m1}=I_{out} \div V_i$.

Since,

$$\frac{V_{01}}{V_i} = -I_{out}\frac{R_2}{V_i},$$

$$\frac{V_{01}}{V_i} = -g_{m1}R_2$$

$$g_{m1} = \frac{I_{out}}{I_{in}R_N} = \frac{A_I}{R_N}$$

Where, A_I denotes current gain,

$$A_I = \frac{h_{fe} - h_{oe}R_K}{1 + h_{oe}(R_2 + R_K)}$$

With reference to Fig. 3,

$$\frac{V_{02}}{V_i} = -g_{m2}R_5$$

where,

$$g_{m2} = \frac{I_4}{I_3 R_N}$$

■

References

1. Slaughter, D W, 'The Emitter-Coupled Differential Amplifier,' *IRE Transactions – Circuit Theory*; March, 1956, pp. 51 to 53.
2. 'Physical Principles and Applications of Junction Transistors' by J. H. Simpson and R. S. Richards, pub. Clarendon Press, Oxford, 1962.

Computed results

Here is the input data for a BBC-Basic program that the author used to make some calculations. Also shown is the resulting output data.

Input data:
R_g/ohm=75
R_1/ohm=1.2E4
R_2/ohm=R_4/ohm=R_5/ohm=1.0E3
R_3/ohm=R_6/ohm=R_7/ohm=4.7
h_{ie}/ohm=1.05E3
h_{re}=240E–6
h_{fe}=200
h_{oe}.ohm=110E–6.
Note that '.' is used to signify either a decimal point, as in '1.0E3', etc., or a full-stop, or the multiply operator, as in 'h_{oe}.ohm'.

Output data:
V_{01}/V_i=–48.66
V_{02}/V_i=48.16
R_{01}/ohm=981.6
R_{02}/ohm=983.4
g_{m1}.ohm=4.866E–2.
This data excludes h_{ib}, h_{ic}, etc.

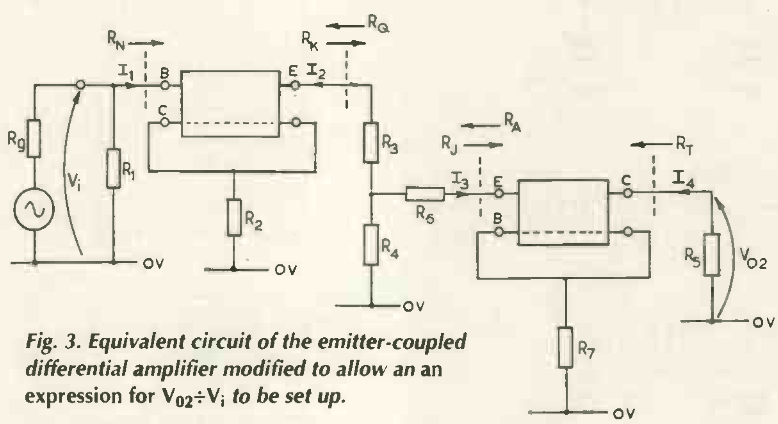


Fig. 3. Equivalent circuit of the emitter-coupled differential amplifier modified to allow an an expression for $V_{02} \div V_i$ ***to be set up.***

CIRCUIT IDEAS

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

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

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

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

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

Digital metal detector provides visual indication

This BFO-based metal detector uses a 1MHz crystal reference in conjunction with a search coil oscillator, operating at the same nominal frequency.

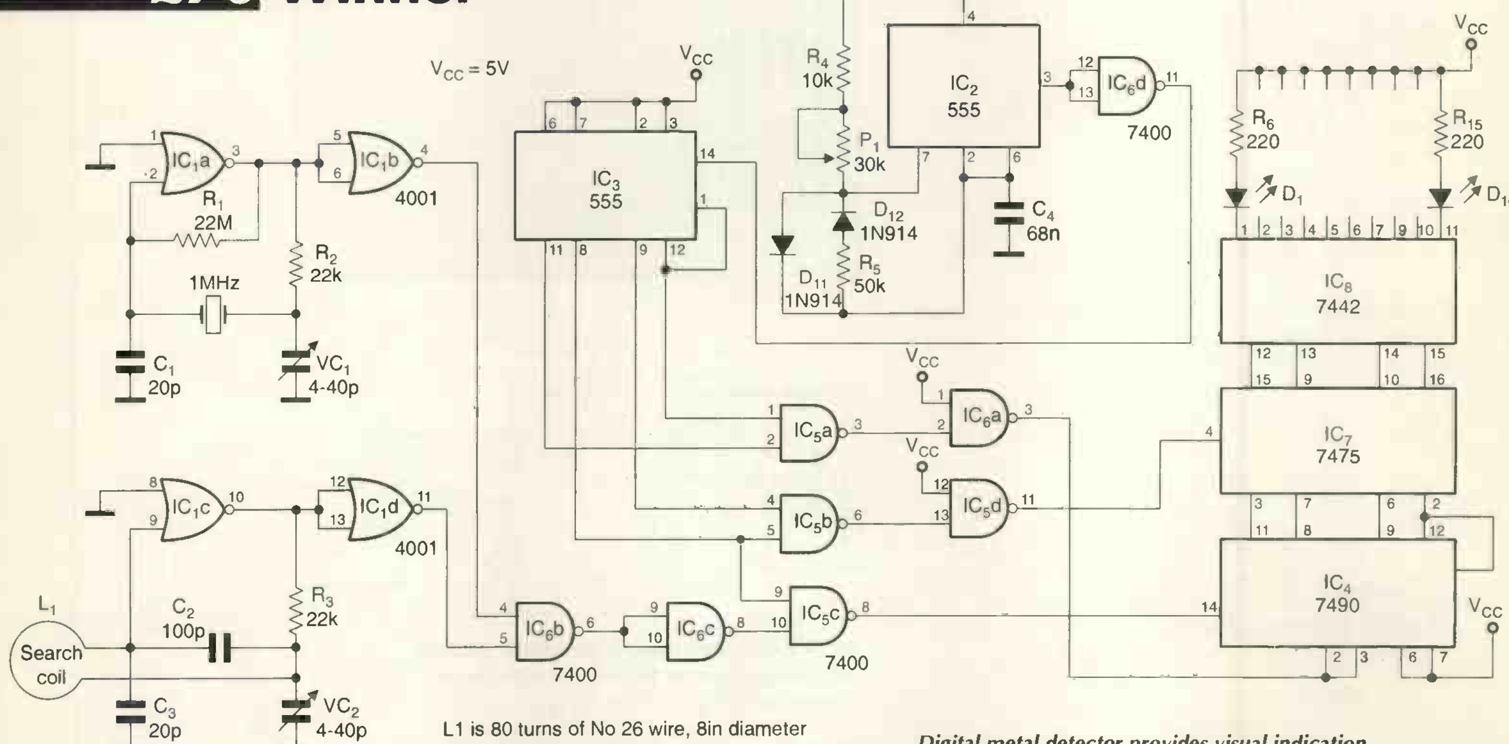
The squarewave reference and search signals are mixed by IC_6b to produce the detected beat frequency. IC_2 generates a frequency of about 500Hz for display driving. The detected and display signals are combined through IC_3 and NAND gates, and fed to U_4 and IC_7 to drive LEDs D_1 through D_{10}.

With VC_1 and VC_2 suitably adjusted, it will be possible in the no metal situation to set P_1 so that all LEDs are OFF, even in the proximity of ground. If the search head is moved over a metal object, e.g. ferromagnetic, the LEDs will light sequentially.

If the metal is non-ferromagnetic. The LEDs will light sequentially in the opposite direction.

Saheed Mansoory Algalandis
Tabriz
Iran
D91

£70 Winner



Digital metal detector provides visual indication and distinguishes between ferrous and non-ferrous metals.

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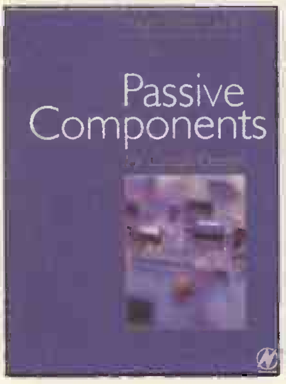

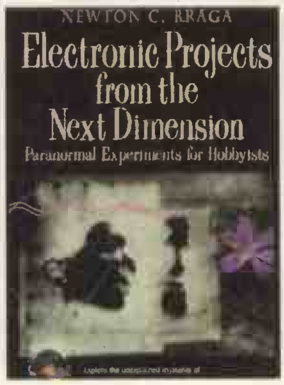

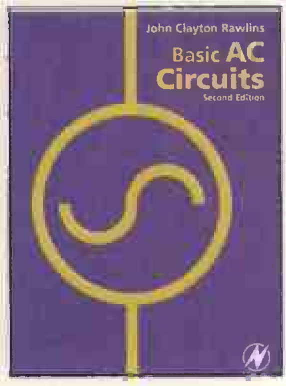

PASSIVE COMPONENTS FOR CIRCUIT DESIGN

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Contents: Preface; Fundamentals; Fixed resistors; Variable resistors, potentiometers and diodes; Capacitors; Inductors and inductive components; Inductive devices; Transducing components; SMT; Hardware; Index

Readership: Technician engineers, circuit designers, advanced hobbyists

Pages: 304pp

Price: UK £22.00
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ELECTRONIC PROJECTS FROM THE NEXT DIMENSION

For years paranormal scientists have explored the detection and documentation of spirits, auras, ESP, hypnosis, and many more phenomena through electronics. Electronic Projects from the Next Dimension provides useful information on building practical circuits and projects, and applying the knowledge to unique experiments in the paranormal field. The author writes about dozens of inexpensive projects to help electronics hobbyists search for and document their own answers about instrumental transcommunication (ITC), the electronic voice phenomenon (EVP), and paranormal experiments involving ESP, auras, and Kirlian photography.

Although paranormal studies are considered esoteric, Electronic Projects from the Next Dimension teaches the technical skills needed to make devices that can be used in many different kinds of experiments. Each section indicates how the circuit can be used in paranormal experiments with suggestions about procedures and how to analyze the results.

Contents: White noise generators for use in instrumental transcommunication (ITC) and electronic voice phenomenon (EVP) experiments; Practical circuits for image experimentation, such as a wireless sparkling image generator, horizontal bar generator, brontophic sound, magnetic field generator, high-voltage generators (Kirlian Machine's I & II); Paranormal skills experiments with temperature change, polygraph, electro-shock, random number generation, UFO detection, and ghost-finding.

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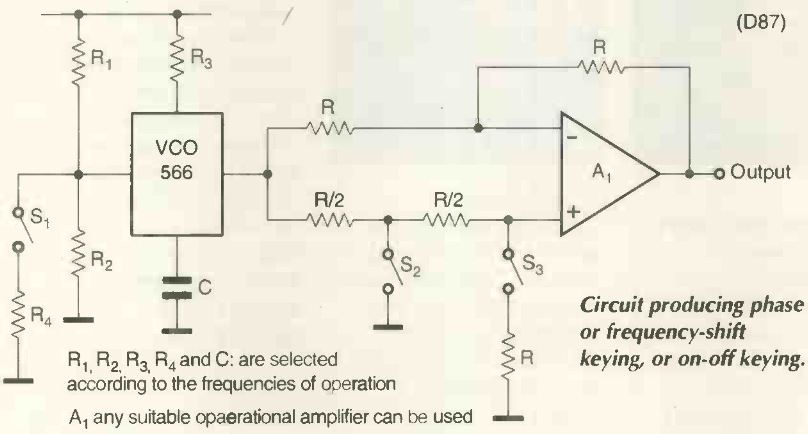
Generating amplitude, frequency and phase-shift keying signals

Amplitude-shift keying, frequency-shift keying and phase-shift keying signals can all be produced by this circuit. It operates as follows:

1. With switches S_2 and S_3 opened, the circuit produces frequency-shift keying signal under control of switch S_1.
2. With switches S_1 and S_3 opened, the circuit produces phase-shift keying signal under control of switch S_2.
3. With switches S_1 and S_2 opened, the circuit produces on-off keying signal under control of switch S_3.



Circuit producing phase or frequency-shift keying, or on-off keying.

On-off keying is a special case of amplitude-shift keying where the signal is keyed entirely off or on. If switches S_1, S_2 and S_3 are monolithic analogue switches, the circuit is digitally controllable.

Muhammad Taher Abuelma'atti
Dharan
Saudi Arabia
D87

Versatile test oscillator

A feature of this instrument, a variant of the Franklin oscillator, is that the tank circuit inductor is untapped. It can thus be used with a wide range of inductors resonating at frequencies up to 100MHz.

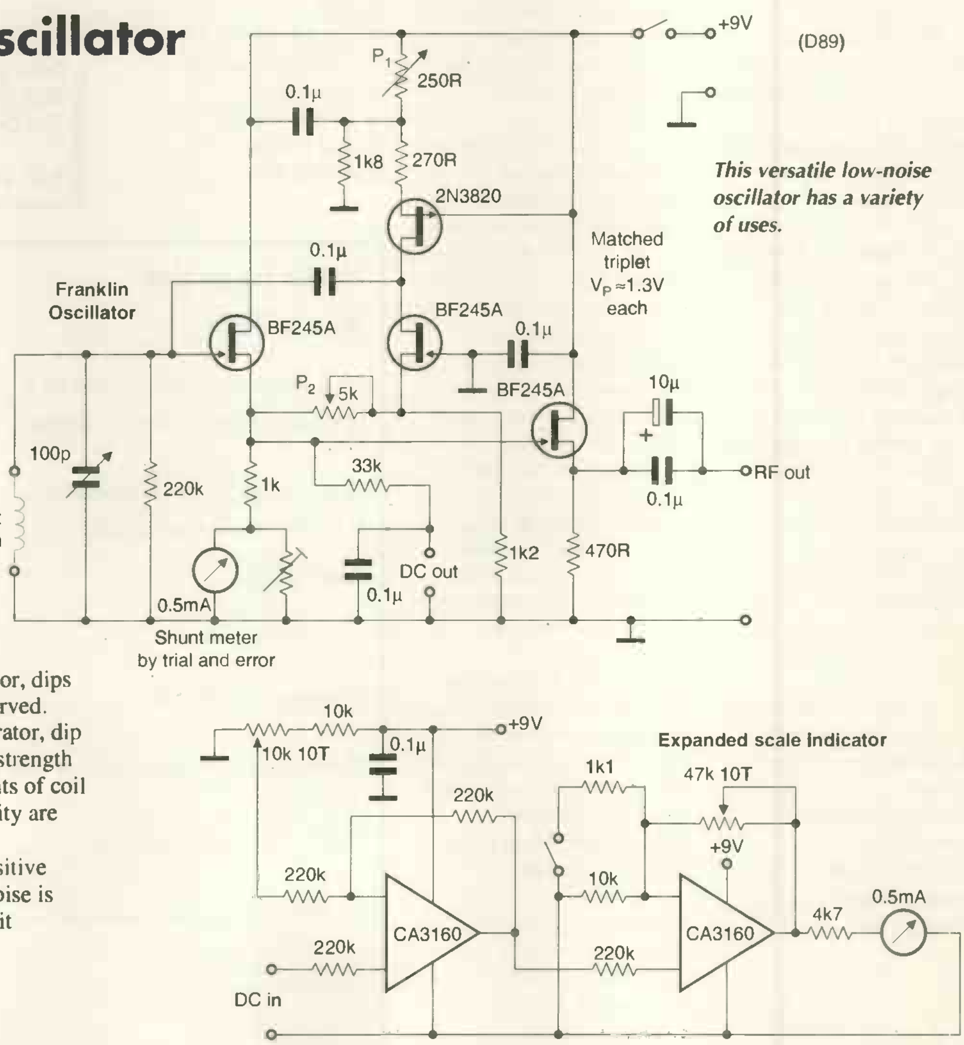
The active load provided by PFET 2N3820, with its typical pinch-off voltage of 1.5V, provides the maintaining amplifier with a high dynamic output impedance. This minimises loading on the tuned circuit.

Potentiometer P_1 is set for maximum amplitude of oscillation, while P_2 adjusts the amount of regeneration.

With the regeneration set to just provide oscillation, as monitored by the expanded scale indicator, the circuit forms a sensitive absorption wavemeter. As a dip oscillator, dips up to the eleventh overtone can be observed.

The unit can be used as a signal generator, dip oscillator, absorption wavemeter, field strength indicator etc. Comparative measurements of coil and capacitor Q factor and crystal activity are possible.

An RF output for connection to a sensitive frequency counter is provided. Phase noise is generally so good that the one hertz digit indication will not vary.



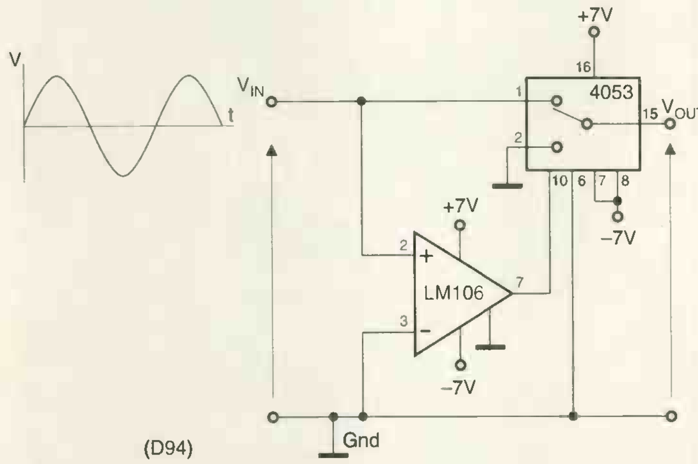
This versatile low-noise oscillator has a variety of uses.

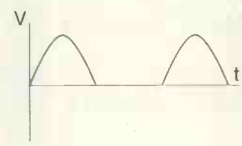
Wim de Ruyter
Oudkarspel
The Netherlands

Ideal rectifier works at video speed

Rectification by diodes has its advantages such as size and cost. There are applications though that require very precise splitting and control of the waveform. This is achieved by the circuit shown.

High-speed comparator LM106 is used to detect the point at which the input waveform goes below 0V. At that instant, the output of the comparator goes low, forcing the 4053 to switch to the grounded input. Voltage V_{out} then follows this input; thus clipping the negative half of the input waveform.





Precision half wave rectifier works at video speeds.

Complex waveforms may be generated if the voltage at the grounded input of the 4053 is derived from a potentiometer. To achieve a good response at high frequencies, decouple both ICs and the power supply.

A single ground point and short tracks are essential.

K P Cummings
Nottingham
D94

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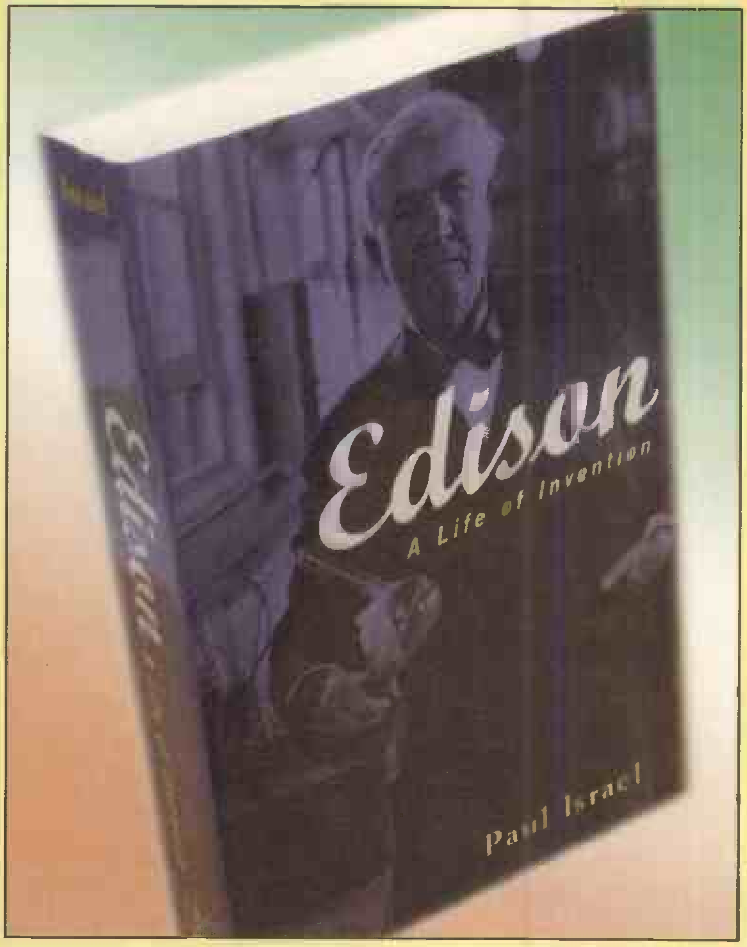

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Audio level indication by tri-colour led

A single tri-colour LED is used here to convey an idea of audio signal level. It is relatively simple and inexpensive, and is particularly attractive in cases where front panel space is at a premium. It is much less cumbersome than a full blown bargraph, and far more informative than a simple 'overload' LED.

At low signal levels, the LED glows green, with varying intensity according to the amplitude. As the signal level increases, the colour changes through amber, then finally to red.

The circuit is based on quad op-amp IC_1. I used the LF347, but other devices would no doubt work just as well.

Op-amp IC_{1b} is configured as precision half wave rectifier, the output of which charges C_4. Attack and decay rates are determined largely by R_5 and R_4 respectively. Input sensitivity is determined by R_3. These parameters are characterised for speech signals at a nominal 0dB level.

Op-amp IC_{1a} simply buffers the capacitor from the following stages. These require a low-impedance source.

Output of IC_{1c} is a voltage proportional to the green segment current. At low levels, D_4 does not conduct, and the output simply follows the input. Above D_4's 'knee' point, the non-inverting input is clamped, inverting action takes over, and the output starts to fall again. This is the 'amber' phase of the circuit's operation.

The output of IC_{1d} is a voltage proportional to the red segment current. This remains at zero until the signal level increases sufficiently to allow D_3 to conduct. Beyond this point, the output rises in line with the increase in signal. A point is reached where the green segment current has diminished to zero, and only red segment current flows.

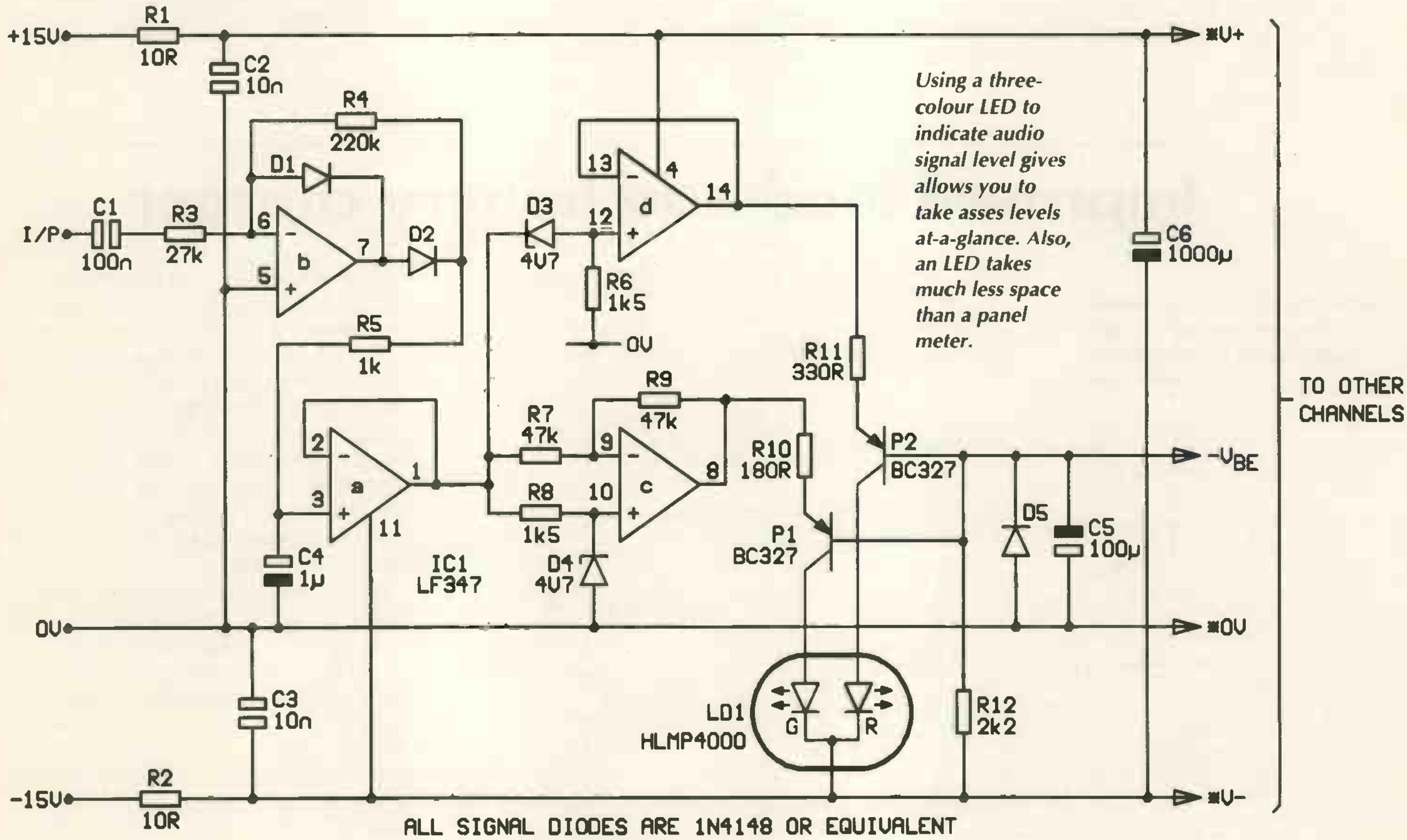
Both transistors are in a common-base configuration. The base bias voltage is set at approximately $-V_{be}$ by R_{12}/D_5 so that the emitters form 'virtual-earth' inputs. Voltage outputs of $IC_{1c,d}$ appear across resistors $R_{10,11}$. The resulting currents flow via the transistors to the common-cathode tri-colour LED. This arrangement eliminates any need to compensate for the LED segment forward voltages. Furthermore, the relatively heavy LED currents flow 'rail-to-rail', and not via the ground return.

The LED segment currents are modulated smoothly, rather than being switched as is the case in a bargraph. Demands on the power supply are therefore reduced, and simple decoupling suffices – as provided by R_1, R_2 and C_6. Multiple indicator circuit blocks can share this supply, and also the $-V_{be}$ bias voltage.

Local HF decoupling for the IC is essential. This is provided by ceramic capacitors $C_{2,3}$.

With no signal applied, a small residual current flows, resulting in a very faint glow from the LED. This should not be a problem in practice.

Graham Booth
Auditel Ltd
Amersham
Bucks
E51



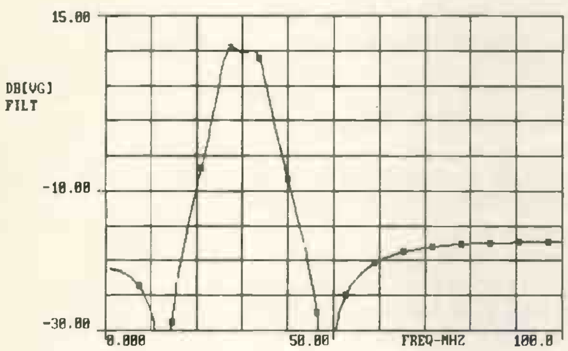
Using a three-colour LED to indicate audio signal level gives allows you to take asses levels at-a-glance. Also, an LED takes much less space than a panel meter.

£50 Winner

CMOS tripler needs no adjustment

Having a low output impedance and high input impedance, advanced-CMOS logic is useful for non-linear signal processing.

In this example a low-impedance square-wave is generated from a 10MHz source using one advanced-CMOS logic gate. A bandpass fitter then extracts the third harmonic while providing voltage gain by impedance transformation.



A second-order 'coil-saving' elliptic design provides adequate selectivity without adjustment, and gives good rejection at the fundamental and fifth harmonic. A second gate then squares up and buffers the output.

In a square wave, the third harmonic is only 7.44dB lower in amplitude than the input. The filter gives about 10dB of voltage gain, so even with circuit losses there is no shortage of signal to operate the second gate. High signal levels combined with *LC* filtering should ensure that the output has low noise.

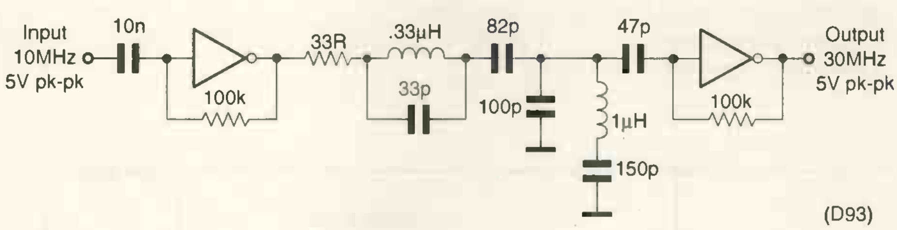
The plot shows a simulation of the filter response allowing 20Ω of source resistance and 4.5pF of load capacitance, typical of an AC gate. The idea can be adapted to work at different frequencies.

At low frequencies though, Schmitt trigger gates should be used to avoid self-oscillation during signal transitions. At higher frequencies, more attention may need to be paid to the effect of the receiving gate capacitance.

The circuit was built with two sections of a 74AC00 quad 2-input NAND gate each with one input wired high. It worked without further adjustment.

J N Wells
St Albans
D93

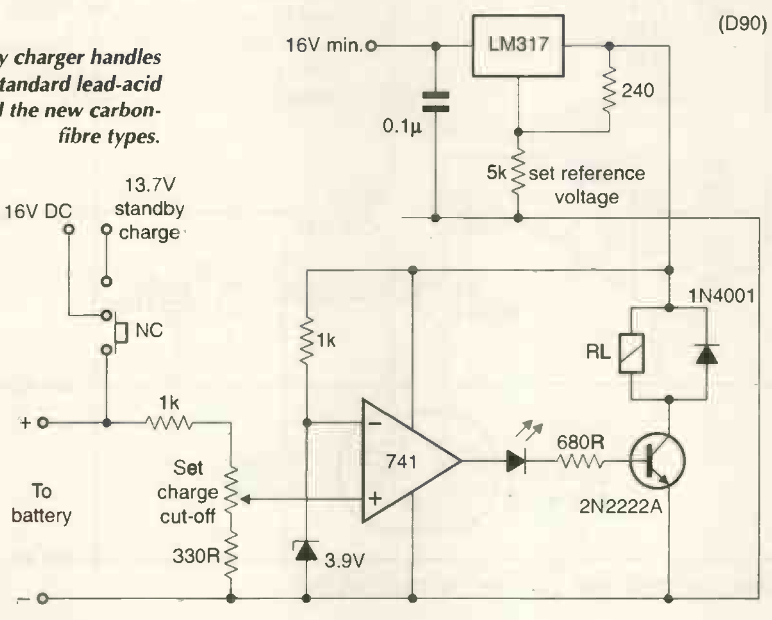
Apologies, but we are unable to announce the winner of the National Instruments Labview kit until next month. Making a decision is proving more difficult than we expected! Ed.



Frequency tripler takes advantage of non-linear characteristics of advanced-CMOS logic.

Improved lead-acid battery charger

Improved battery charger handles sealed and standard lead-acid batteries and the new carbon-fibre types.



The lead-acid battery charger shown on page 830 of the October 1999 issue will not give a 100% full charge to a flat battery, if set up as described, using a fully charged battery (12.6V or more) as the reference voltage. This is because on charge, a final voltage of 15.6V is required to guarantee 100% capacity.

My modified circuit shown addresses this problem. It also changes the battery over to a trickle charge regime on reaching full charge.

An adjustable regulator output is also shown, permitting the charging of standard lead-acid batteries (15.6V), sealed types (15V), the new Elecsol carbon-fibre battery (14.5V) or even 6V batteries.

Andrew Bird
Burntwood
Staffordshire
D90

8051-type controllers include ISP

New 8051-type controllers are appearing that lend themselves to in-circuit programming and reprogramming – drastically reducing development time for those of you without a full-blown emulator. The one described here has a separate loader ROM so none of the 64K application ROM is lost. Both ROM areas are reprogrammable flash memories.

Winbond's W78E516 microcontroller is an 8051 derivative that can be programmed *in situ*. This is also the first 8051-based device to provide two separate on-board flash ROMs. One is a 64k block of application ROM, referred to as APROM, while the other is a loader ROM of 4k, called LDROM.

Either ROM can program the other without external programming voltages or hardware. And, because both ROMs are blank when the devices are manufactured, software developers are free to implement their own loader program rather than being tied to a specific serial type interface.

Normally, these devices are supplied either blank or pre-programmed with a boot loader that allows the APROM to be programmed by a serial connection to a PC. Because either ROM can program the other, developers can program the device with application code, then program the LDROM with a bespoke boot loader program.

Source code for the LDROM is freely available from Flint Distribution. This code may be

Registers used for in-system programming.

Register name: SFRCN

Bit	Name	Function
7		Reserved
6	WFWIN	Selects the ROM for In-circuit programming – 0=APROM, 1=LDROM
5	OEN	ROM Output enable
4	CEN	ROM Chip enable
3,2,1,0	CTRL[3:0]	Flash control signals (See below)

Mode	WFWIN	CTRL[3:0]	OEN	CEN	SFRAH, SFRAL	SFRFD
Erase APROM	0	0010	1	0	-	X
Program APROM	0	0001	1	0	Address in	Data in
Read APROM	0	0000	0	0	Address in	Data Out
Erase LDROM	1	0010	1	0	-	X
Program LDROM	1	0001	1	0	Address in	Data in
Read LDROM	1	0000	0	0	Address in	Data out

Register name: CHPCON

Bit	Name	Function
7	SWRESET	Reading as logic 1 indicates that the device is in FO4KBOOT mode. MCU will perform a software reset if this and FBOOTSL and FPROGEN are set
6	–	Reserved
5	–	Reserved
4	ENAUXRAM	1: Enable on-chip 256 byte auxiliary RAM 0: Disable on-chip 256 byte auxiliary RAM
3	0	Must be zero
2	0	Must be zero
1	FBOOTSL	1: Loader program is in LDROM, APROM can be reprogrammed 0: Loader program is in APROM, LDROM can be reprogrammed
0	FPROGEN	1: ISP function enabled 0: ISP function disabled

*F04KMODE

modified and reprogrammed into the LDROM either via the APROM or using a standard device programmer.

In-system programming

ISP functions available on the W78E516 are erase, read and write, and 'act on the alternative ROM to that currently selected'. Access is through four registers. Three are address pointers, namely SFRAH and SFRAL, SFRFD. These contains data read from or written to flash. The fourth is the control register SFRCN, which selects ISP mode.

Erase is accomplished by a single command. All functions are carried out independently of the CPU, while the device is in 'idle' mode.

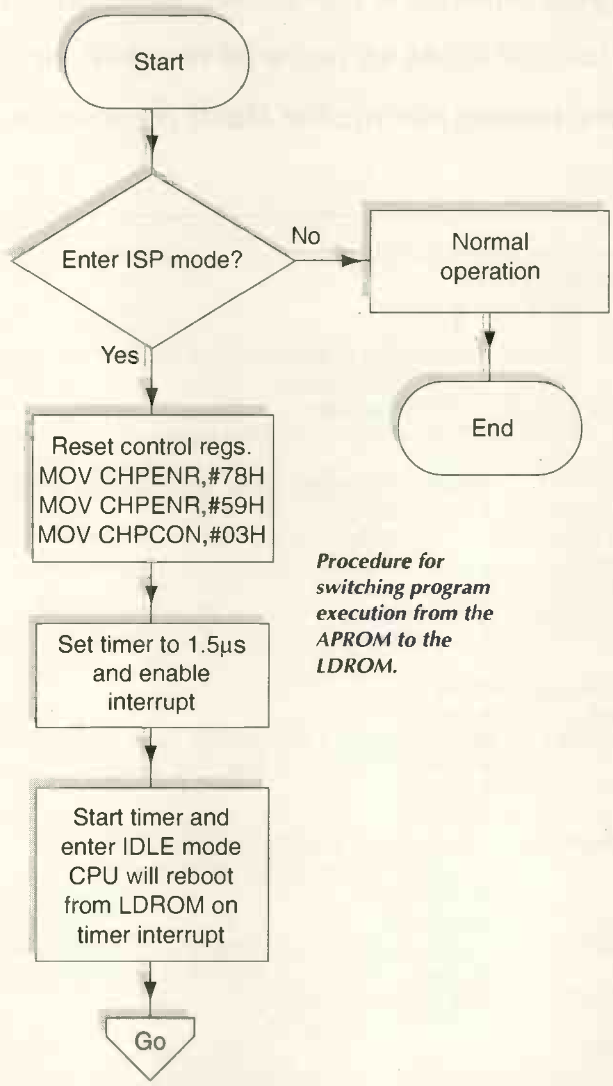
Once the controlling registers are programmed, the device is put into idle mode. While in idle mode, the CPU clock is disabled but peripherals and interrupt logic remain active.

Usually a timer generates an interrupt after a predefined time, based on how long the ISP function takes to execute. This interrupt returns the CPU to active mode so that the device can continue ISP services or normal operation.

During normal operation, the W78E516 boots from APROM, jumping to the LDROM under software control when required. It is also possible to force the device to boot from the 4k LDROM via hardware. This mode has an associated flag known as F04KBOOT.

An F04KBOOT flag allows the system to distinguish between normal or forced execution of LDROM code. Forced execution might occur for example if a reset is instigated while there is no set condition, or while waiting for a hardware reset if set.

The F04KBOOT flag can also trigger a hardware test routine within an application.



Procedure for switching program execution from the APROM to the LDROM.

Development environment

There's support for the W78E516's in-system programmability – and other 8051 type devices – in the form of a development platform available from Flint Distribution. This platform is based on hardware that connects to the standard 40-pin DIL 8051 socket. The hardware allows code to be downloaded to the device and executed *in situ*.

At the heart of the system is a W78E516 programmed with a boot loader. It runs from an 11.0592MHz crystal and connects to the PC via a serial interface. A PC application allows users to select a '.bin' file and download it to the device while in F04KBOOT mode.

The program first erases the APROM, and then programs and verifies the device with the selected file. In this way the APROM can be fully utilised by the programmer. There is no overhead of a monitor program or operating system.

File size checking is also provided, to ensure that the file is no greater than 64k and data are verified during file transfer.

Hardware

The system was designed for use with bread-boarded prototypes. Because of this, a standard 80C5X 40-pin DIL pin-out is provided. This provides compatibility with standard devices and with third-party adapters for conversion to PLCC or QFP footprint.

A key factor in the hardware design is that as many I/O lines as possible are kept free for use. Because the PLCC version of the device has an extra four-bit port, this is used to drive an LED. This LED indicates that the boot loader is running and therefore the device is waiting for code to be downloaded to the APROM.

The port is also used to invoke F04KBOOT mode. The only other two I/O lines used for the ISP function are RXD and TXD. Connecting these ports to DIP switches allows users to choose between connection to the board's RS232 driver during program download; or to the pins of the 40-pin DIL header when programs are running and require serial communications.

The system further allows control of the /EA input, used to switch between external and internal data access. All bus control signals are also available, giving maximum flexibility in expanded memory applications.

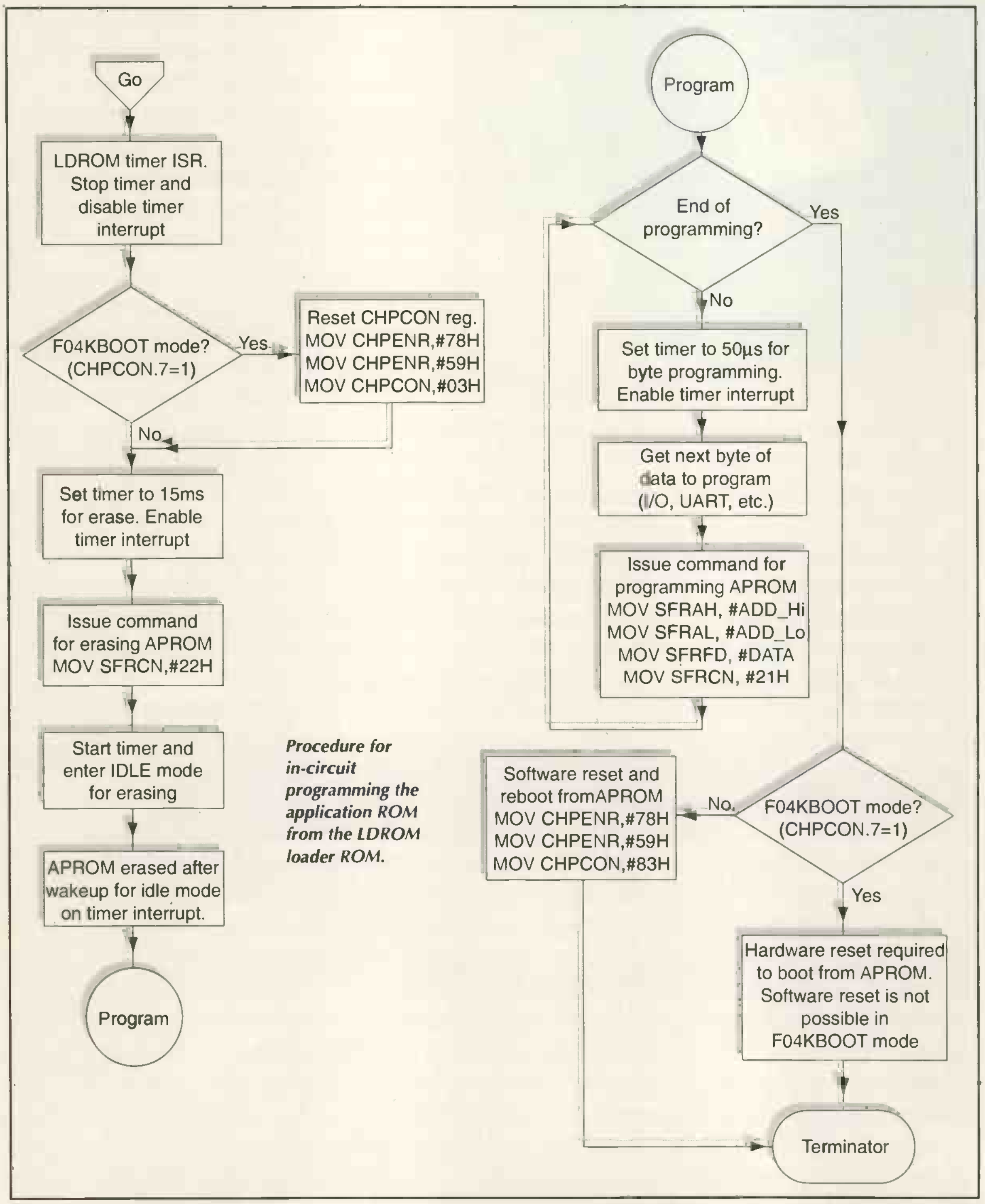
The only pins that are not available to the developer are the reset and oscillator pins. Most of the device resources remain available to the user.

A number of new ISP devices are becoming available from Winbond, such as a low-voltage version of the W78E516 and the W78LE58, again a low voltage device which features a 32k APROM and 4k LDROM. Either of these devices can be used with the ISP kit, and may be ordered separately.

When coupled with a public-domain 8051 assembler or C compiler, the result is an extremely low cost 8051 development system, where maximum MCU resources are accessible to the developer, and software overheads are kept to a minimum.

For more information on the ISP kit and to download the associated programs, please visit the Flint website, www.flint.co.uk. ■

Offer over page



Procedure for in-circuit programming the application ROM from the LDROM loader ROM.

Reader offer

Buy your in-circuit programming development kit for just £60 – save £25

Until 31 March 2001, Flint Distribution is offering the new W78E516 in-circuit programming development kit to *Electronics World* readers for just £60 including VAT and postage – a saving of £25 over the normal selling price *without* VAT. Simply mention that you are an *Electronics World* reader when you order.

At the heart of this development platform is a W78E516 controller running from an 11.0592MHz crystal and pre-programmed with a boot loader.

The loader, contained in a separate flash memory called LDROM allows the controller's APROM application memory area to be programmed via a serial connection to a PC. No monitor program or operating system is needed in the controller so the APROM can be fully utilised by the developers.

The target device plugs into a standard 805C5X 40-pin DIL socket, which is also compatible with third-party adapters for conversion to PLCC footprint. Devices can also be removed and programmed with a range of alternative programmers including systems from Dataman, Needhams, BP Microsystems and MPQ.

A PC application is included to allow binary code to be downloaded to the device by forcing it to boot from LDROM rather than APROM. File size is automatically checked to ensure that APROM space is not exceeded, and data verification is incorporated to ensure reliable downloading.

Because either ROM can program the other, developers are free to program the device with application code, which can in turn program the LDROM with a bespoke boot loader programme. LDROM source code is freely available from Flint, allowing any amount of development or modification.

The kit includes:

- in-circuit programmable evaluation board with W78E516 microcontroller
- serial cable for interfacing the board with your PC
- DOS software for downloading object code to the target board
- Windows software for downloading code to the target board

Developers can use the ISP kit with new 8051 microcontrollers in the Winbond family, such as low-voltage versions and the W78LE58, which provides 32k APROM and 4k LDROM, also with low-voltage capability.

For further details, please contact Flint Distribution at Walker Road, Bardon Hill, Coalville, Leicestershire, LE67 1TU. Tel 01530 510333 or see the website www.flint.co.uk.

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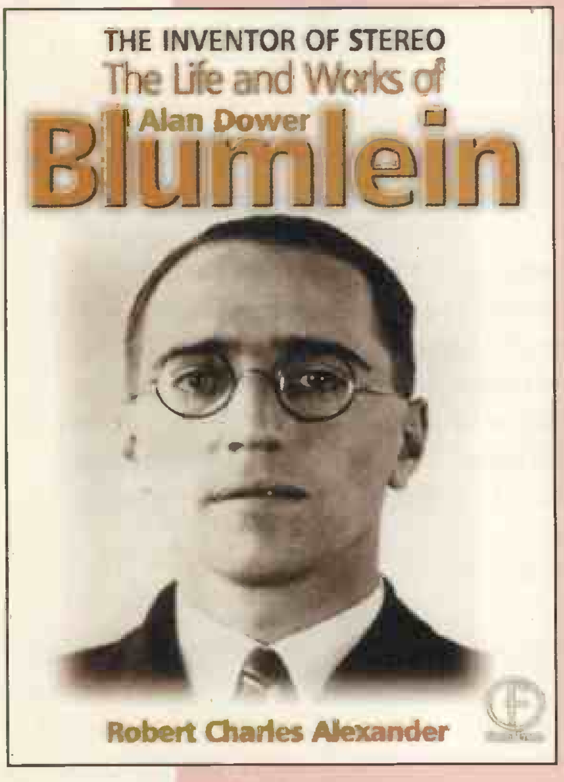

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This book is the definitive study of the life and works of one of Britain's most important inventors who, due to a cruel set of circumstances, has all but been overlooked by history.

Alan Dower Blumlein led an extraordinary life in which his inventive output rate easily surpassed that of Edison, but whose early death during the darkest days of World War Two led to a shroud of secrecy which has covered his life and achievements ever since.

His 1931 Patent for a Binaural Recording System was so revolutionary that most of his contemporaries regarded it as more than 20 years ahead of its time. Even years after his death, the full magnitude of its detail had not been fully utilized. Among his 128 patents are the principal electronic circuits critical to the development of the world's first electronic television system. During his short working life, Blumlein produced patent after patent breaking entirely new ground in electronic and audio engineering.

During the Second World War, Alan Blumlein was deeply engaged in the very secret work of radar development and contributed enormously to the system eventually to become 'H2S' - blind-bombing radar. Tragically, during an experimental H2S flight in June 1942, the Halifax bomber in which Blumlein and several colleagues were flying, crashed and all aboard were killed. He was just days short of his thirty-ninth birthday.

For many years there have been rumours about a biography of Alan Blumlein, yet none has been forthcoming. This is the world's first study of a man whose achievements should rank among those of the greatest Britain has produced. This book provides detailed knowledge of every one of his patents and the process behind them, while giving an in-depth study of the life and times of this quite extraordinary man.

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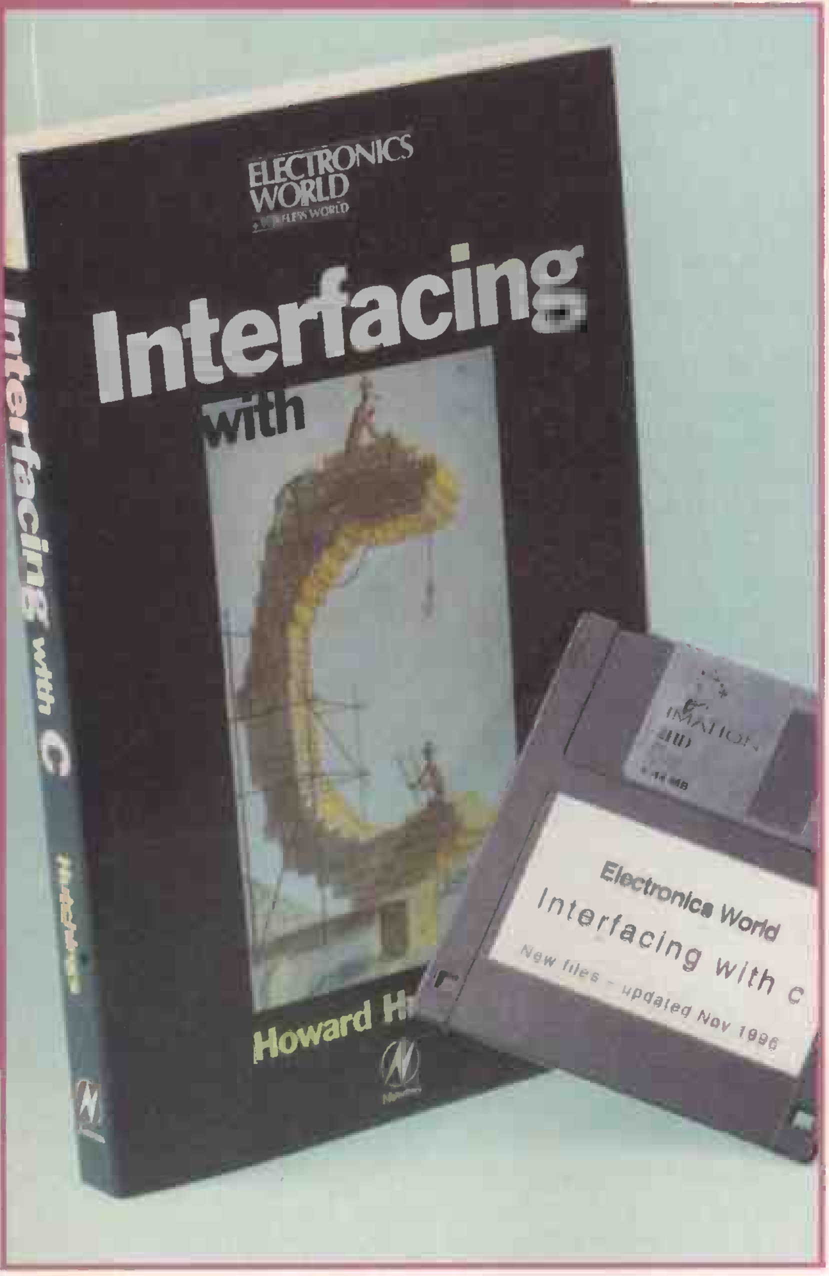

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Introduction to current conveyors

Giuseppe Ferri *et al* present a guide to current conveyors in three parts, the first of which gives you a brief history of this frequently overlooked circuit configuration.

Giuseppe Ferri and Pierpaolo De Laurentiis, are with the Università di L'Aquila, L'Aquila, Italy and Giovanni Stochino is with Ericsson Lab Italy SpA, in Rome.

In the autumn of 1966, A.S.Sedra, working on his Master's thesis[1], had to design a voltage-controlled waveform generator. He implemented a current-controlled oscillator, so he had the additional task of converting the voltage to current.

At that time, the best known solution to convert a voltage to a current was to take a p-n-p bipolar junction transistor with its base grounded and connect its emitter via a resistor to the control voltage. But that solution presented several drawbacks. The finite base-emitter voltage caused a large offset in the voltage-to-frequency conversion. Moreover it depended on temperature and on the flowing current, worsening the stability and linearity of the entire circuit.

To improve the current-to-voltage conversion, Sedra came up with the circuit of **Fig. 1**. This was the first type current conveyor, or convertor,[1-3] and it was given the abbreviation CCI, where the 'I' is the Roman numeral '1'.

Both X and Y are output nodes. If you look closely at Fig. 1, you will see that the current in Tr_3, which is approximately i_X, is 'copied' by the dual-output current mirror formed by $Tr_{3\text{-}5}$.

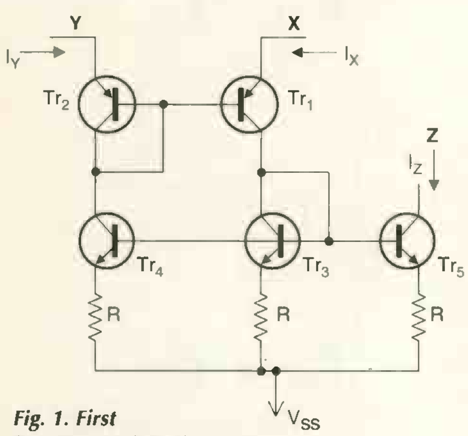


Fig. 1. First implementation of current conveyor, CCI. Here, $i_X = i_Y = i_Z$.

Fig. 2. Symbol of a first-generation current conveyor, or CCI.

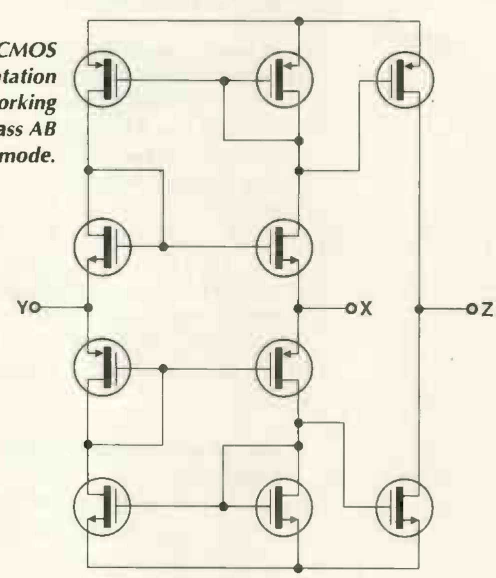


Fig. 3. CMOS implementation of a CCI working in Class AB mode.

So, $i_X=i_Y=i_Z$. Moreover, as the same current is flowing in Tr_1 and Tr_2, their base-emitter voltages will be same too. This means that $v_Y=v_X$, regardless of the values of v_Y and i_Y.

In mathematical terms, the CCI, whose symbol is given in **Fig. 2**, is an active device described by the following matrix relationship,

$$\begin{bmatrix} i_Y \\ v_X \\ i_Z \end{bmatrix} = \begin{bmatrix} 0 & 1 & 0 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix} \begin{bmatrix} v_Y \\ i_X \\ v_Z \end{bmatrix}$$

This type of current conveyor is particularly useful whenever currents at very low impedances have to be measured. This is why the current is duplicated with unity gain on high-impedance terminal Z.

Historically, the CCI has formed a useful alternative to oscilloscope current probes based on Hall effect devices. They have allowed measurements over a range of 0 to 100MHz to be made at input impedances of less than 1Ω.

Note that the performance of the circuit of Fig. 1 can be improved. For example, more elaborate current mirrors can be added, as can operating capabilities. The complementary version of the circuit could also be considered. Here, each transistor would be replaced by its opposite polarity counterpart and the negative supply replaced with a positive one.

It is also possible to connect two complementary current conveyors to come up with a current conveyor capable of AB operation. When the CCI works in Class AB, it can be biased from low dc current values. It can also process current signals having magnitude larger than the bias current I_o, with reduced harmonic distortion.

Fig. 3. Shows a CMOS implementation of the Class-AB CCI. A typical supply would be 3.3V.

Second-generation current conveyors

In 1970 two years later than the CCI invention, Sedra and Smith proposed an extension of their original circuit[2]. They called it the

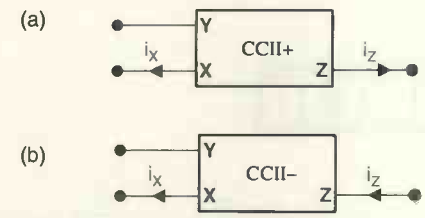


Fig. 4. In a) is the CCII+ while b) shows the complementary CCII–.

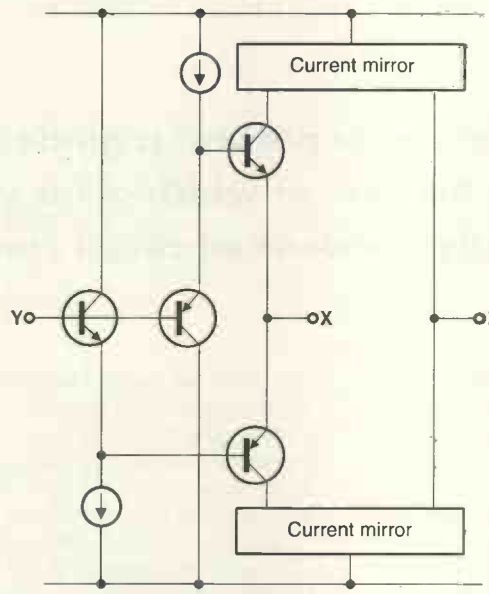


Fig. 5. Internal block structure of the CCII01.

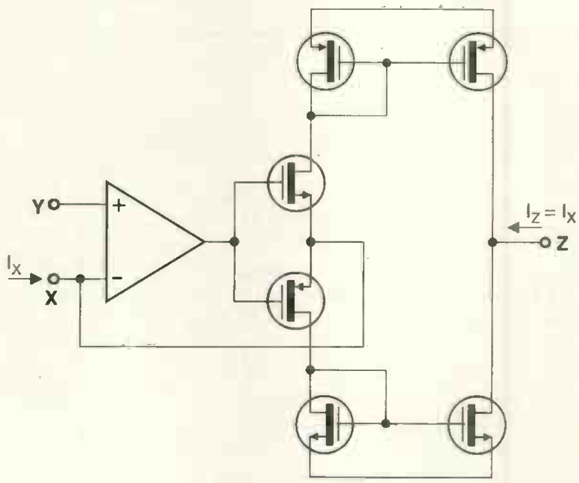


Fig. 6. This positive current-conveyor, abbreviated to CCII+, uses complementary MOS devices to provide better voltage-follower performance.

second-generation current conveyor, or CCII. This novel circuit was a more versatile device. We will be concentrating mostly on this second-generation in this article and its two successors.

The idea of Sedra and Smith was to make the input impedance on the terminal Y infinite, i.e. we have the following ideal behaviour,

$$\begin{bmatrix} i_Y \\ v_X \\ i_Z \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 1 & 0 & 0 \\ 0 & \pm 1 & 0 \end{bmatrix} \begin{bmatrix} v_Y \\ i_X \\ v_Z \end{bmatrix}$$

Thus we have $i_Y=0$, $v_X=v_Y$, that is to make the impedance on terminal X zero, $i_Z=\pm i_Y$. This makes the output impedance on terminal Z infinite.

The '±' sign means that the current supplied to X can be conveyed to Z with either positive (CCII+) or negative (CCII–) polarity. **Figure 4** shows the circuit symbols used for the CCII+ and CCII– respectively.

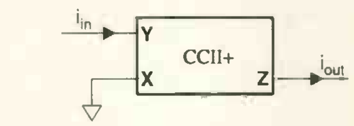
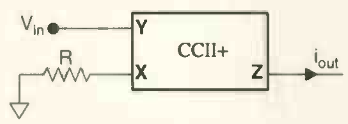
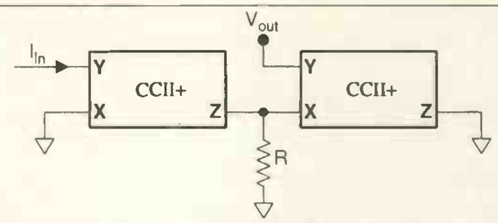
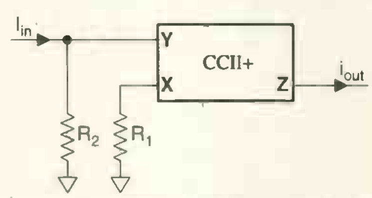
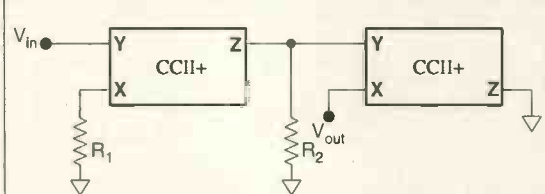
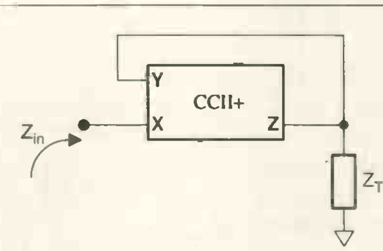
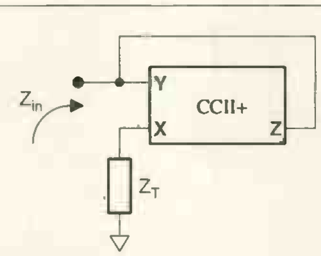
Generally speaking, the current conveyor can be considered a 'double follower' in that it can form a voltages or current follower.

The main advantages of the current conveyor approach is that the traditional closed-loop gain band-width conflict of negative feedback voltage op-amp is avoided. The benefits of negative feedback – noise reduction, improvements of input and output impedances, etc. – are not available. However, a wider bandwidth at high gain levels is inherent.

When the current conveyor was introduced, the semiconductor industry was absorbed with the development of monolithic op-amps. However, a lot of papers were published on the conveyor's theory and applications, showing that the traditional op-amp has no clear benefits over a current-conveyor.

Because of their combined voltage and current buffering action, current conveyors may be used to synthesise a great number of analogue circuit functions. They can also reduce the number of components needed to implement the circuits[3-7], as you will see from the **Table** on the right.

Table 1. Basic topologies using current conveyors, demonstrating that many analogue functions can be implemented with fewer components relative to traditional op-amp solutions.

Functional element	Schematic	Mathematical expression
Current buffer		$i_{out} = -i_{in}$
Voltage-controlled current source		$i_{out} = V_{in}/R$
Curent-controlled voltage source		$V_{out} = RI_{in}$
Current amplifier		$i_{out} = I_{in}R_2/R_1$
Voltage amplifier		$V_{out} = V_{in}R_2/R_1$
Negative-impedance converter (current controlled)		$Z_{in} = Z_T$
Negative-impedance converter (voltage controlled)		$Z_{in} = Z_T$

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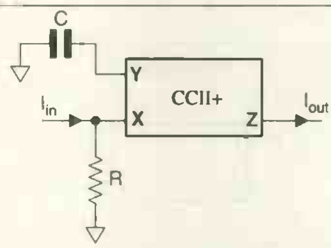
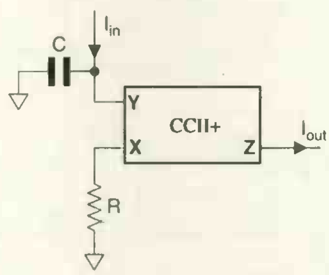
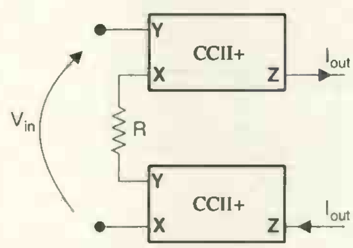
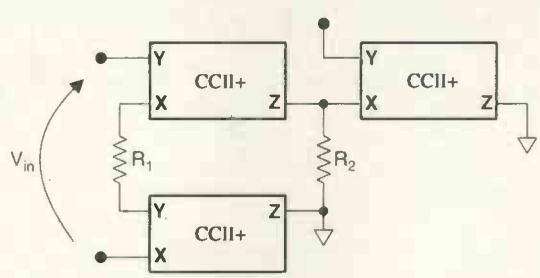
Functional element	Schematic	Mathematical expression
Current differentiator		$I_{out} = RC\frac{dI_{in}}{dt}$
Current integrator		$I_{out} = \frac{1}{RC}\int I_{in}(t)\, dt$
Differential voltage-to-current converter		$I_{out} = \frac{V_{in}}{R}$
Instrumentation amplifier		$V_{out} = V_{in} \cdot \frac{R_2}{R_1}$

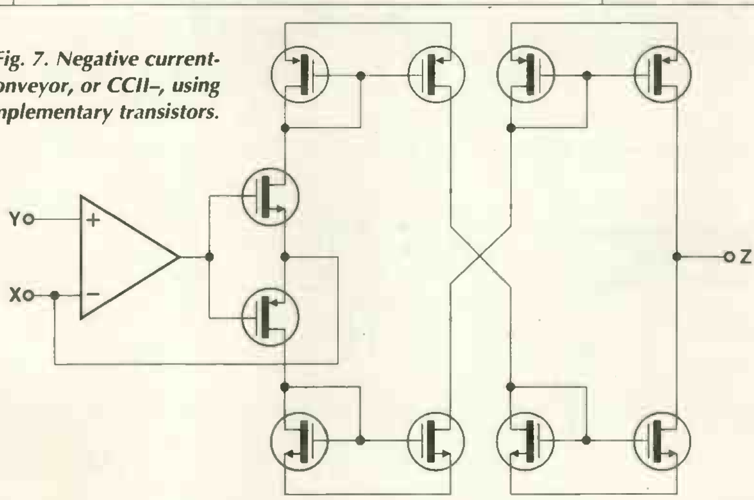


Fig. 7. Negative current-conveyor, or CCII–, using complementary transistors.

First commercial conveyor chips

The first monolithic commercial current-conveyor was the CCII01, developed by LTP Electronics. **Figure 5** shows its simplified schematic. This device, capable of working from ±5V up to ±15V supplies, it features an equivalent slew-rate of 2000V/μs and a 100MHz bandwidth[3].

A commercial CCII is the PA630A 8925, produced by Phototronics.

In order to perform a better following action from a voltage view-point, it is possible to take two complementary transistors in the negative feedback loop of an op-amp. This technique has been demonstrated to work well with CMOS technology.

An example of CCII+ is given in **Fig.** 7. Here, the op-amp can be realised as a standard two-stage amplifier. In order to implement a CCII–, only two current mirrors need to be added, as in Fig. 7. ■

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2. A. Sedra, K. Smith, 'A Second-Generation Current Conveyor and Its Applications', IEEE Trans. on Circuit Theory, CT-17, pp. 132-133, February 1970.
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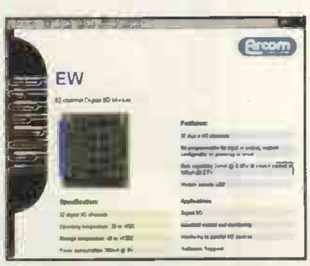

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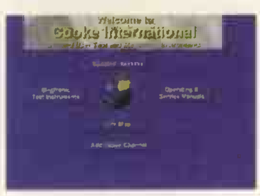

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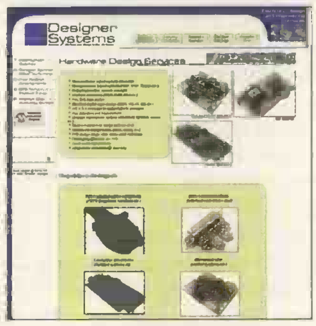

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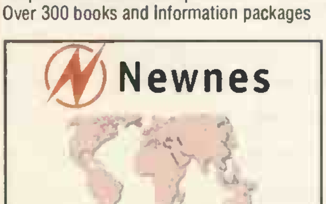

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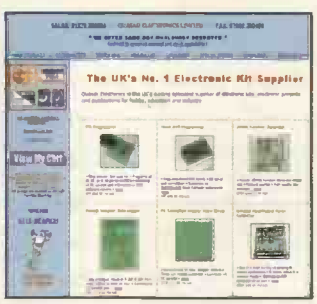

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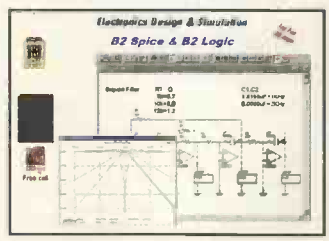

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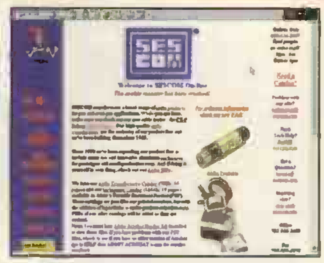

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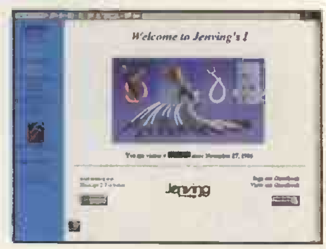

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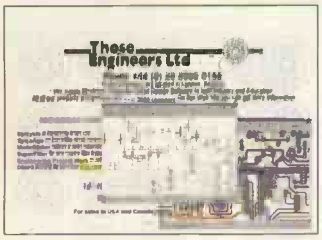

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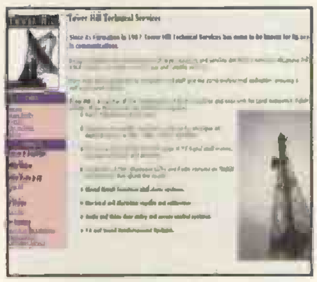

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Following on from the newsgroup discussion last month there is a UK Email group for TV technicians where you can send an Email to everyone in the group. There's just over 30 people in the group at present. For more details and how to register look at the egroup home page. Just a general comment though - you do have to be careful who you give your Email address to so that you can avoid "spamming" - that is getting lots of unwanted Email about dubious Russian site (amongst others).

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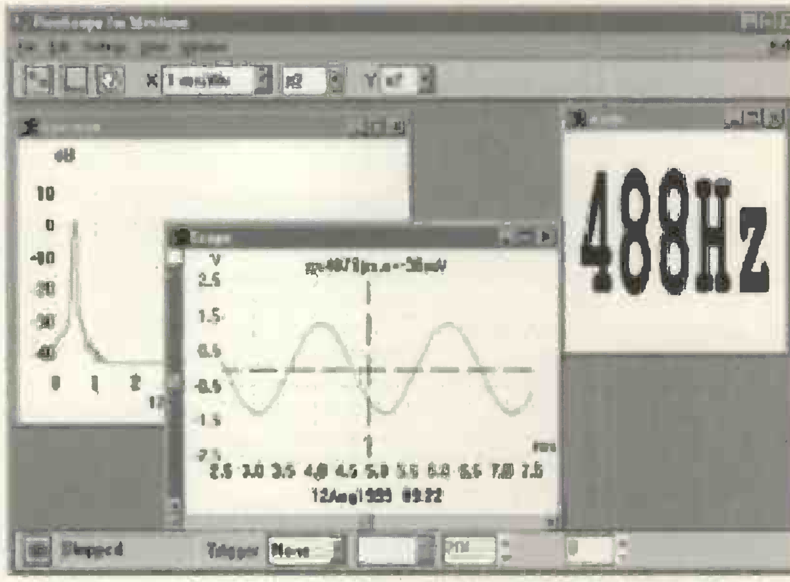

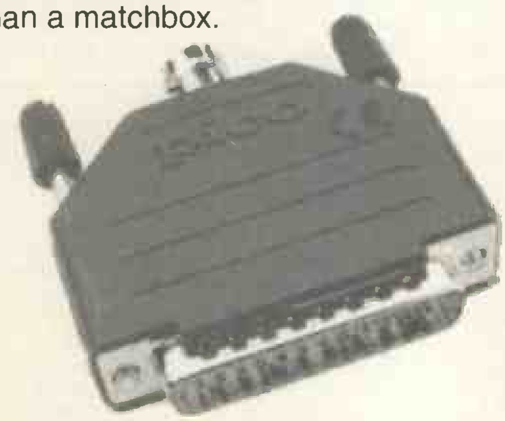

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Minister for electronics?

I think we can accept that coal was the fuel that powered the 19th century industrial revolution.

I believe it is also true to say that oil, for transport and chemicals, was the dynamic fuel that powered the 20th century.

What then is the fuel of the 21st century? Is that fuel not a material but a technology? Electronics?

Strange then at a recent meeting of IMAPS UK (The International Microelectronics and Packaging Society) a society dedicated to the advancement of microelectronics and packaging in the UK, not one member could identify or name the Government Minister responsible for this key sector of industry. It was agreed that a phone call to Lord Sainsbury at the Department for Trade and Industry would probably be the best bet – but nobody appeared to know.

This is a sad reflection on what I believe is the Government's lack of support for the electronics industry. Despite this, electronics in the UK does exist but in the rest of the developed world one does see strong Government support.

If we don't have a 'Minister for Electronics' is it perhaps time to appoint one?

David Lowrie
Chairman IMAPS UK
Baldock
Hertfordshire

X-ray specs

The interesting article on X-rays in the October 2000 issue perpetuates a popular misconception. In describing existing radiographic techniques it states, "Traditionally the detector is ordinary photosensitive emulsion, but in digital radiography it is replaced with either a semiconductor sheet or a phosphor screen."

In fact, in medical radiography the primary detector of X-rays is a phosphor screen, sometimes referred to as a 'salt screen'. The phosphor screens used have been developed over the years to give high-resolution images at ever-lower doses of radiation.

It is the light emitted from the phosphor screen that is recorded by the photosensitive emulsion. The emulsion is optically sensitised so that its peak spectral sensitivity matches the peak emission wavelength of the phosphor.

Film for medical radiography is normally coated on both sides with a silver-halide emulsion. A phosphor screen is held in contact with each side of the film during exposure. The amount of X-radiation recorded directly by the silver halide emulsion is minimal compared with the emitted radiation recorded from the phosphor screen.

In many of the semiconductor radiographic techniques described in the article, the key element is the same as with the technique listing photographic film; a phosphor screen.

Silver-halide emulsions are used directly as X-ray detectors in some dosage badges and for radiographic inspection of welds. In these cases a lead-foil screen is used to increase sensitivity by through electron emission caused by absorbed X-rays.

Guy Selby-Lowndes
Billingshurst
Kent

Curve tracing in Sweden

Having a long-standing interest in curve tracers, I was very interested to see Ian Hickman's description in the August 2000 issue.

I built my first tracer in the late fifties – actually a valve curve tracer, later adapted for transistors – and my second in the early seventies.

I recently decided to build a third, more up to date tracer. But Mr Hickman's reasoning relating to suppressing the bright spot resulting from the inactive part of the cycle made me wonder why no one – including myself earlier – has used double rectification of the collector voltage? That would eliminate the inactive bright spot altogether and double the refresh rate. This is desirable when a set of 8 curves is to be displayed on an ordinary oscilloscope. Of course, it requires a transformer with a centre tap – or several if many voltages are wanted – and twice as many diodes, but diodes are cheap.

If it is not going to be visible in

Phase shifter for headphones

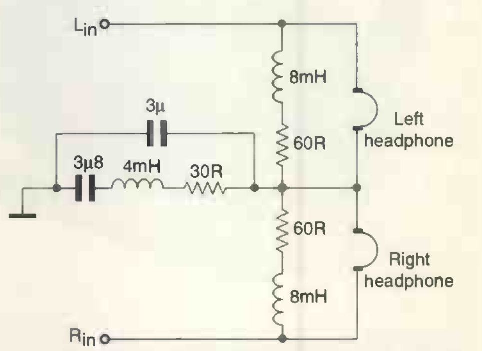
In response to A T Granger's letter on page 747 of the September 2000 issue, here is a circuit in passive and active forms that will solve his head phone problems.

Phase shifts caused by the circuit put you at the front of the stage even though you are listening through headphones. The passive form was used by Dr Bauer here in the USA. I use it with a Walkman.

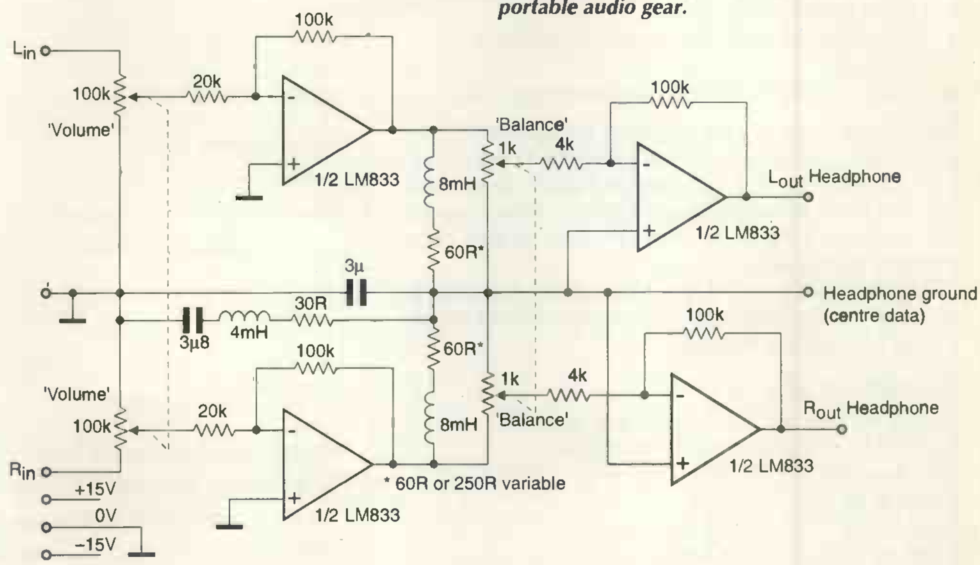
The active alternative works on all outputs. I use it on tape, CD and tuner outputs. It requires +12V and –12V at 100mA.

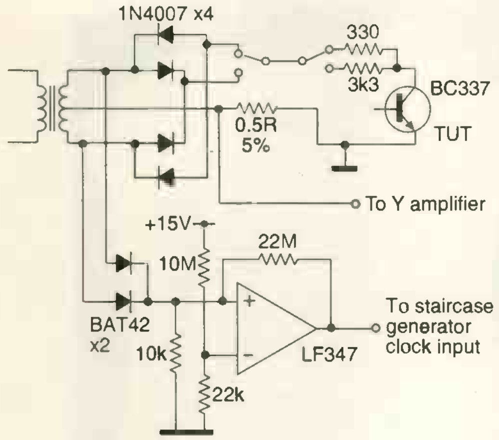
Do not use any polarised capacitor in the circuit.

James Watson
Orangevale
California
USA

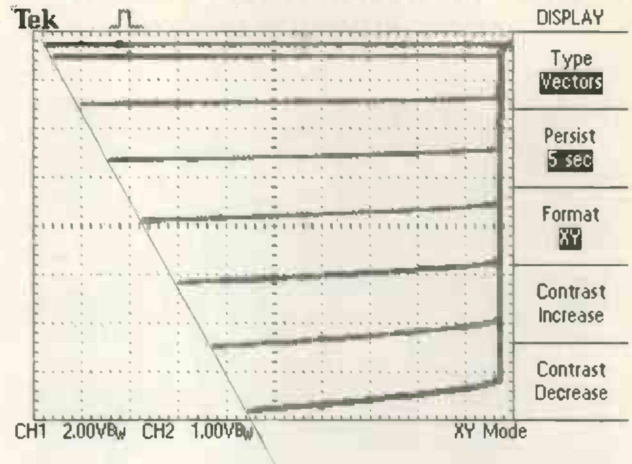


When you are listening through headphones, these audio phase-shift circuits create the illusion that you are standing in front of what you are listening to. The passive form is useful for portable audio gear.





Adding diodes in this way eliminates the oscilloscope bright spot problem. The right-hand switch can have many positions for selecting any resistance between 5Ω and 33kΩ.



the displayed curve set, the staircase generator needs to be clocked very close to the zero crossing. I have tested the simplified circuit shown and found it to perform well from about 2V up to several tens of volts.

The BAT42 Schottky diodes conduct at a lower voltage than the 1N4002 diodes, so the clocking takes place essentially before any collector voltage is applied to the transistor under test. Because the clock pulse is very narrow, C_5 in the original circuit, was reduced to 33pF and R_{47}, R_{48} were replaced by 1N4148 clamp diodes.

Another matter, which Mr Hickman touched on briefly, is the possibility of including a series resistor in the collector supply. I made this resistor switchable in 12 positions, as shown simplified in the schematic. This adds the possibility to have a selectable projected load line in the curve set which may be interesting for graphical design purposes. The curve set shown with the circuit diagram was produced using a Tektronix TDS210 DSO.

For the rest of the circuit, I used Mr Hickman's original design. I found it well designed and up to date. I look forward to seeing more of that kind! ■

Evert Olsson
Sweden

Better buffers rebuffed

I would like to comment on Dave Kimber's article on page 858 of the November 2000 issue, entitled, 'Better Buffers III'. In it, he claims that, "crossover distortion is impossible to eliminate".

I don't know whether Mr Kimber has been living in a cave all of his life or whether he is just plain ignorant and uninformed. To be sure, his assertion is just plain gobbledygook and pure nonsense.

We over here in the US had this problem of crossover distortion – sometimes also referred to as switching distortion – solved some thirty-five years ago.

There are fundamentally three ways to eliminate this effect. The first is, of course, pure class A operation but, that is obvious. The second method was a stroke of genius and was conceived by Sidney S. Smith, of Marantz fame. He is probably the most brilliant audio engineer that ever lived. His idea was to use power diodes in the output stage for the compound pair rather than a resistor.

Using power diodes allowed the transfer function to be *always* totally exponential throughout virtually the entire signal cycle, therefore allowing the power devices to remain on virtually through the signal cycle. This was done in the Model 15, that even by today's standards is still a magnificent product.

The second method is one that I created while director of engineering at SAE in the early seventies. What I did was to shunt the output stage emitter resistors with power Schottky diodes. This allowed the forward transconductance to remain much more linear throughout the entire signal cycle.

The proof of the pudding is in the performance and indeed, these amplifiers would produce THD figures of less than 0.01% at 20kHz at full power with no crossover notches.

Great pains were also taken to do a complete spectral analysis in order to make sure that these distortion residuals were indeed harmonic products and not higher order switching products.

Along with my output stage design was a totally new dual-differential input stage wherein the entire amplifier was also fully complementary.

In order for the output stage to work properly, I found it to be necessary to dramatically lower the open loop gain whereby the open loop frequency response was significantly raised, obviously making the amplifier much faster. This also aids in eliminating crossover distortion.

As a final note, it seems that the engineering world in Europe has not yet caught up with the times regarding full complementary topologies.

Even in Douglas Self's book, he doesn't even raise the issue. This topology, which I created, has been standard fare both here in America as well as in Japan for almost 30 years.

I think Mr Kimber and others need to join the modern world and get caught up to date.

James Bongiorno
Lompoc
California

David replies:

Class A does, of course, eliminate crossover distortion but there may still be odd order distortion because the transfer functions of the two halves of the output stage are not fully complementary.

Using diodes in the base circuit of the output transistors may help reduce crossover distortion, as I mentioned. I am glad that some people have had success with this method.

I cannot see how putting diodes across the emitter resistors will help with crossover distortion though. Either at the output transistor emitters proper, or at the CCEF 'emitter', I would expect diodes to help counteract the effects of high current beta droop; this could reduce large signal distortion.

In order to eliminate crossover distortion in Class B amplifiers, the transfer functions of the two halves must exactly fit in the crossover region just like two jigsaw pieces. I don't think anybody has yet achieved this, but some get closer than others. The remaining distortion is then reduced by carefully applied negative feedback, which requires adequate open-loop gain.

I am fascinated by the idea of reducing crossover distortion by reducing open loop gain, as Mr Bongiorno appears to claim. Perhaps he would like to give more details? I fear he may have fallen into the common trap of regarding high open-loop bandwidth as being intrinsically a good thing – see Self's comments on this.

Dave Kimber G8HQP
Davant Technology Limited

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