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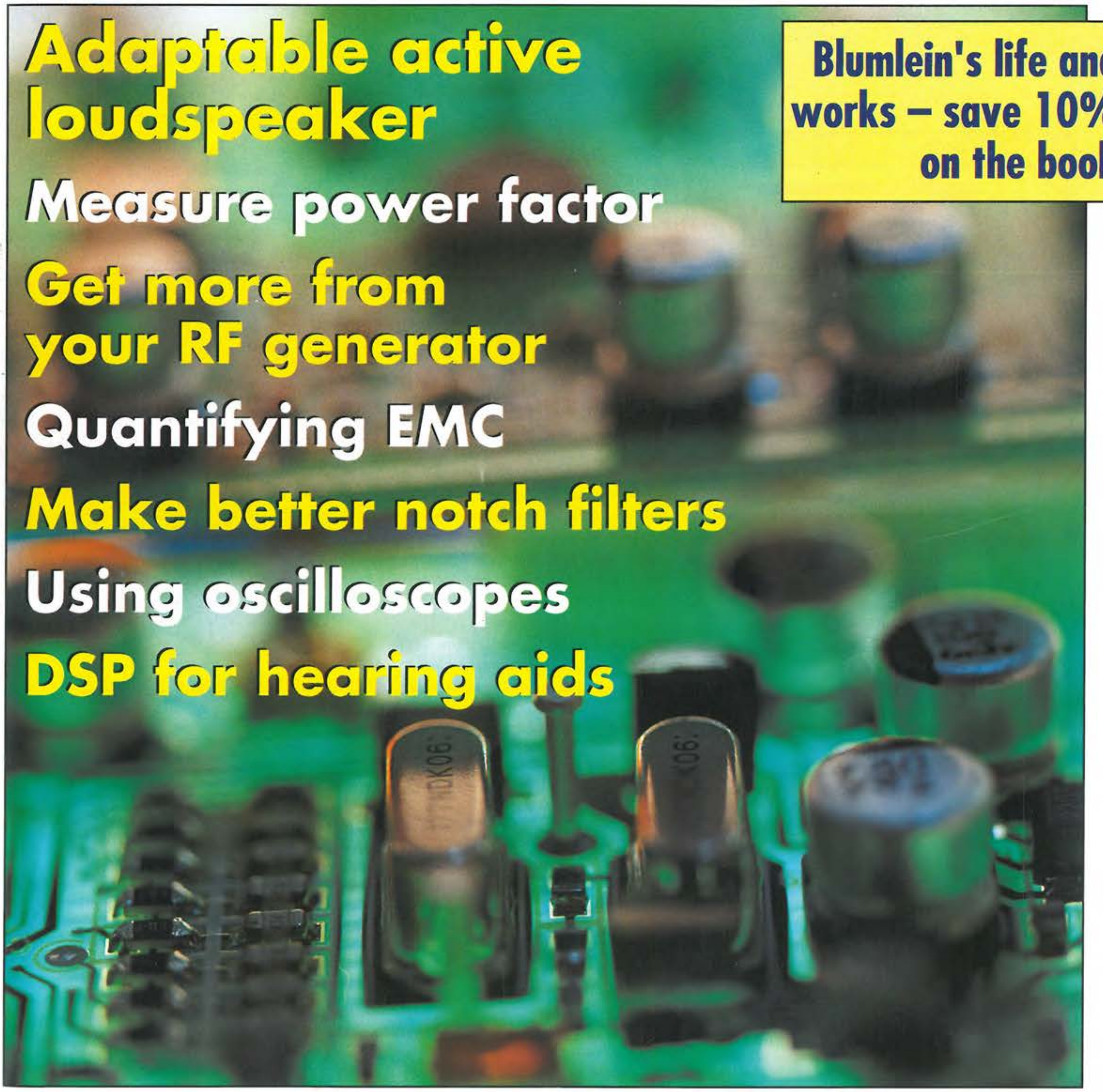

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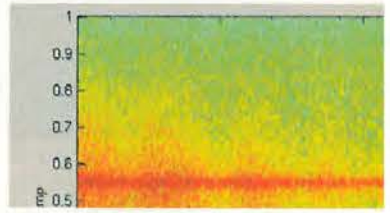
Using parallel processing DSP, this behind-the-ear hearing aid manages around 130Mips at just 10MHz clock frequency. Read about it on page 100.

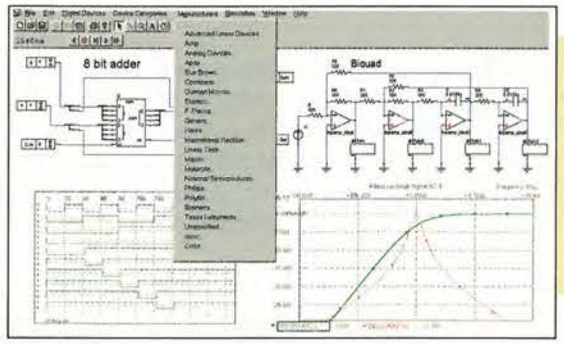
Features of the adaptable loudspeaker design on page 105 are a squarewave-in, squarewave-out active crossover, electronic bass roll-off compensation and spherical enclosures for minimal diffraction.



DSP is being used to detect and correct sputtering in spray paint guns. Find out how on page 98.

New book – see page 165.

March issue on sale 3 February



Want to save 25% on the list price of RD Research's brand new Spice based simulator? Electronics World readers can. Details on page 126.

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REED BUSINESS INFORMATION

A murmur of breaking glass?

Linux – a free, stable and powerful alternative operating system. But if Linux is to start breaking Windows, it needs more diverse applications support.

Back in July 1998, with Windows 98 about to be launched, *Electronics World* carried an editorial about the Microsoft operating system monopoly. With Windows 2000 almost here, it is an appropriate time to review the situation.

The 1998 editorial was concerned with the abuses that often follow in monopolistic situations. As the subsequent letters pages in *Electronics World* adroitly pointed out, the bias was clearly against Windows; yes indeed, for what reasonable person could support a greedy giant selling faulted software at such vast profit?

However, not even I imagined the extent of the bullying that Judge T P Jackson revealed.

Moreover, the bullying has not stopped, despite the court ruling. Microsoft is now telling systems engineers that they will lose their Microsoft certification unless they take, and pass, an exam for Windows 2000.

It seems that Microsoft is determined to stuff Windows 2000 down the throats of NT4 and 98 users whether they like it or not. Naturally, those on the receiving end of this outrageous threat are unhappy, but then they are only small fry and unlikely to have sufficient muscle to resist Microsoft.

Looking back a few years, it is astonishing how we all hurried to the store, money in hand and only too eager to hand the stuff over, to acquire such a poorly-designed, over-priced and unstable operating system.

With the benefit of hindsight, the stimulus was not that Windows was good – although the hype certainly was – but that the alternatives were so poor. Everyone was only too pleased to get something marginally better. This situation caused the start of the rot.

True, the enforced popularity of Windows, fuelled by hype not just from Microsoft but by some sectors of the PC press, led to a standardisation of a sort with Windows as the common operating system. But even today this 'standardisation' is fragmented between NT, 98 and 95. Some are even sticking to 16-bit Windows 3.1 or 3.11, no doubt hoping for something better than 95, 98 or 2000 to appear.

However, judging from the number of published complaints, standardisation on a system with faults is worse than no standardisation. There have been improvements to the various Windows versions, but not the clear-cut root-and-branch improvements that were expected – particularly with regard to stability. Bill Gates' very own public blue-screen experience showed everyone that operating system instability remains an issue.

I believe that users were hoping that 95 would be more stable, and they hoped the same for 98. But it has been pointed out that Windows is, like a house built on sand, still reliant on DOS despite all the camouflage.

Houses built on shaky foundations can be propped up by a process called under-pinning. This is a laborious and not very satisfactory practice, but it does stop the house falling down – and it is cheaper than rebuilding. Is there an analogy here with Windows?

Thank heavens then for the meteoric rise of Linux. It is now supported by an impressive number of big names, many of who have signed up in the last few months. Its success seems assured.

The contrast to Microsoft's product could not be starker. Linux source code is free to anyone who wants to use it, Windows is not. Linux was not designed with money as the main objective.

Linux has an established reputation for built-in stability. It has done so to the extent that it is currently being bandied about by the *cognoscente* that if you manage to crash Windows once a month, then you can expect to crash Linux in a year; if you crash Windows once a year, you can expect Linux to crash once every 12 years...

Despite Linux having such widespread support and a reputation for stability, the field of electronics engineering is not well served. I know of no program for circuit simulation, pcb design or autorouteing written for Linux.

Surely here is an opportunity? The first CAD company in this field with a Linux program is bound to attract the interest of every designer disaffected with Windows – and this will definitely place the operating system ahead of the competition. ■

Simon Wright

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UPDATE

UK engineers denied special advisory role in government

The UK's government has snubbed its country's engineers by rejecting a cross-party call to appoint a special engineering adviser in the Cabinet Office.

Lichfield Tory Michael Fabricant, who staged a special half-hour debate on the subject at Westminster recently, said it was wrong that issues to do with engineering were merely part of the brief of Chief Scientific Adviser Sir Robert May.

"The government's continued refusal to accept that engineering is wealth creating, and different from science, helps to perpetuate the myth that engineers are of a lower status," said Fabricant. "That is a dangerous and ignorant over-simplification of reality."

Fabricant's concerns over the implications for the status of engineers were backed by fellow Chartered Engineer and Labour MP Claire Curtis-Thomas.

"The government's persistence in not employing an engineering adviser to the Cabinet sends an unfortunate message to the two million or so engineers in the UK, which is that their status and their contribution to the quality of our lives are not acknowledged or understood at the highest level of government," said Curtis-Thomas.

Fabricant said he had no criticism of Sir Robert, who despite having training in chemical engineering was essentially a natural scientist, but added: "Engineering forms part of his brief because the government sees engineering as no more than a branch of science, a sub-set on a Venn diagram."

And he continued: "By every measure, engineers enjoy – if that is the word – a lower status in the UK than in France, Germany, the USA or Japan,"

Trade and Industry Minister Patricia Hewitt rejected the plea, saying: "We are very happy with our Chief Scientific Adviser. My understanding is that, contrary to what has been said, most other countries have a Chief Scientific Advisers who embraces engineering."

Hewitt insisted the government had "a deep understanding and appreciation of the extraordinary contribution that engineers make today to the quality of our lives and the strength of our economy."

Invention of the century... the PC?

The PC has ousted the TV as the most important home technology invention of the 20th century, according to US market research.

Thirty per cent of those surveyed by Technocopia named the PC as the century's single most important home technology development, with 22 per cent choosing the television for second place.

According to Technocopia founder Hillary Rettig: "The PC has reconfigured our lexicon and daily habits in a manner which few innovations can match."

The also-rans in the survey included the refrigerator which received eight per cent of the votes, followed by the air conditioner with five per cent, AC electricity and the electric washing machine tied at four per cent, and the radio and the remote control device tied at two per cent.

A further 24 other devices were mentioned including: electric can openers, electric blankets and pinball machines.

SiGe technology boosted by copper interconnect promises 40GHz chips

Silicon germanium (SiGe) transistors and copper interconnects have been combined in a single IC process by IBM Microelectronics.

Adding the SiGe devices to a 0.18μm CMOS process has "little effect on the CMOS device properties and design-rules", said IBM in a paper presented at the IEDM conference in the US.

Including three layers of copper wiring improves speed – important if the process is to be useful in RF applications.

Transition frequency of the process is 90GHz, which would enable application frequencies above 40GHz, the firm claims.

One of the downside of the process is a reduced breakdown voltage of the SiGe heterojunction bipolar transistors (HBTs). Collector-emitter breakdown is 2.3V, while collector-base breakdown is 5.9V. Higher breakdowns are possible, but with f_T reduced to 25GHz.

Several companies are using IBM's SiGe process, such as Intersil for its Prism II wireless LAN chipset. The addition of copper would improve speed significantly.

TV on the Radio... A combined television and mobile phone has been developed by Samsung Electronics. The company mounted a 1.8in LCD on the mobile hand-set and built in a miniature TV receiver. Talk time of the phone is claimed to be 170 minutes, while in TV mode the batteries last for 200 minutes, the firm said. The TV/phone will be available in Korea early next year.

100kbit/s mobile phones in years, not months

Suggestions that GPRS mobile phones with data speeds of 100kbits/s will be available in the middle of this year have been branded as "wildly optimistic" by mobile operator Orange.

In its report 'The Future of Mobile Data Technologies' the company said there has been much conflicting and frequently misleading information about the availability and potential of next generation technologies.

Orange believes deployment of GPRS in 2000 is "extremely unlikely" as terminals will not be available until 2001. It says the data rates offered will initially be 14.4kbit/s rising to 57.6kbit/s by the end of 2001. Orange is itself pushing HSCSD technology as a soon-to-be-available alternative rather than waiting possibly two years for GPRS.

Rival company Vodafone believes GPRS will be available in 2000 and intends to start trials in the first quarter. It sees the technology being introduced in the last quarter of the year. But a spokeswoman said; "We really see it as a technology that will become widely available in 2001."

Orange's report also shows EDGE, the last evolution before 3G mobiles, will happen in 2002 at the earliest.

GSM EXTENSIONS AT A GLANCE

- HSCSD is the final evolution of circuit switched technology. It enables data rates up to 64kbit/s.
- GPRS (General Packet Radio Service) is the first GSM packet switch implementation. Data rates up to 115kbit/s are expected.
- EDGE (Enhanced Data for GSM Evolution) represents the final evolution of GSM. Data rates of 384kbit/s will be possible.

School design & technology opt out attacked

Controversy... The design and technology opt out.

The government's decision to make it easier for school children to drop Design and Technology (D&T) at age 13 has been heavily criticised.

A group including the Engineering Council, the Engineering Employers Federation and the Royal Academy of Engineering, is urging the government to think again due to the importance of technology now and in the future.

The decision to allow D&T to'be one of the few National Curriculum subjects that can be dropped has already been taken.

The subject can be dropped now, although only if the pupil cannot keep up or can show work-based experience. The government is seeking to make this process easier. "There is nothing we can do to stop this happening," said Ruth Wright, education officer at the Engineering Council. "We can only affect the process and try to make sure it's as stitched down as possible."

As a former teacher, one of Wright's fears is that career guidance at school will form the basis of the decision to drop D&T. The government has admitted this needs re-organisation. One result could be a loss of female talent. "If students skew off in the arts direction very strongly at 13, they're not going to be able to pull back if they change their mind later," said Wright.

- A Y2K recruitment campaign has been launched to encourage more British companies to join a scheme aiming to encourage more students into industry. The Year in Industry scheme has been running for 14 years, providing pre-university students with a year's experience in industry. It has achieved an 82% return rate for students entering industry so far.

New gate insulator promises yet higher performance silicon

Motorola says it has developed a new class of semiconductor material that will enable the production of smaller and faster transistors, potentially boosting the performance of current chip designs.

According to Motorola, the technology presents a way around a silicon performance roadblock as the company grapples with the limitations of silicon dioxide.

The material is called perovskite. It has never been used in chip making processes before, but Motorola says that it allows for the production of transistors that are as much as four times thinner than using current materials. Although the research is still at an early stage, the company has managed to produce working chips using the material.

"As devices continue to shrink in size, the gate oxide of the transistor also needs to become thinner. However, we are quickly reaching the limit where we can no longer thin the silicon dioxide which has been used as a gate insulator for the last 30 years," said Jim Prendergast, v-p and general manager of Motorola's Physical Sciences Research Lab. "The solution is to use a new family of materials that appear electrically to be much smaller than their actual physical thickness."

This is achieved by using materials with a higher dielectric constant than the standard silicon dioxide, which limits current leakage. By growing a strontium titanate crystalline material on silicon substrates, Motorola has demonstrated electrical properties over ten times better than silicon dioxide.

Tory ghost haunts UK's chip capability

Restrictions imposed by the last Tory government could jeopardise the country's capability in high-level chip design by barring UK participation in European R&D programmes, writes David Manners.

Next generation European microelectronics R&D will focus on system-on-chip technology, said Dr Jurgen Knorr, chairman of MEDEA, the pan-European chip R&D organisation, in Paris late last year.

However, UK participation in European programmes is restricted to companies with under 250 employees and to universities with small company links. This debars both UK universities and small companies because university R&D tends to get funded by larger companies, while small companies have found it too expensive and time-consuming to get into European programmes.

"We had a meeting with the DTI but they do not expect any change in the rules," said MEDEA director Gerard Matheron.

At the Department of Trade and Industry, Dr Tim Scragg, said: "We're taking a strong interest. I believe we need to have the whole infrastructure from the supply side to the design side. The rules on participation are always being reviewed".

"There is a danger that the UK microelectronics design community could lose their edge if they are excluded from the systems knowledge gained from collaborative R&D", warned industry analyst Malcolm Penn.

According to Ian Burnett, chairman of JEMI, which represents companies supplying the chip industry, many small UK firms which qualify under DTI rules are excluded from participating because they do not have the time or resources to apply for European projects. "Exclusion means our members only get European-generated IP when it is generally available – not when the participants get it. In areas of design, knowledge is power and months really count," said Burnett.

NTL and Alcatel trial 6Mbit/s ADSL phone links in UK

Telecoms operator NTL has teamed up with Alcatel for a trial of broadband ADSL technology applications for businesses in Surrey.

The trial, one of two or three in the UK, involves a "small number of customers" in Woking and Guildford and is more about configuration of the network rather than the ADSL (asymmetric digital subscriber line) technology itself, which NTL already agrees does works. The intention is to roll it out towards the end of this year.

Part of the trial will involve direct access for home workers via a cable modem to their ADSL linked company networks. It is believed to be the first time a cable modem to ADSL modem link has been optimised in the UK.

"We're aiming for seamless integration across a number of transmission technologies," said Stephen Rowles, NTL's group MD of business solutions.

The company sees a role for copper cable-based ADSL technology even in its largely optical fibre network. "There is a place for ADSL among our fibre," said Rowles, "and as the local loop gets unbundled we are well placed to roll it out."

The trials will demonstrate applications at speeds up to 6Mbit/s. Pricing of the service has not yet been determined but NTL is "watching others very keenly."

In a separate move Internet service provider Cerbernet is inviting up to 50 companies to take part in the second phase of BT's ADSL broadband communications trial.

A full commercial service, using BT's ADSL network technology, will be available in March 2000. The first phase of the trial began in early 1999.

"Cerbernet has worked with BT in the development and trials of ADSL technology since day one, and Cerbernet has more experience with the business applications of ADSL than any other UK service provider," said technical director of Cerbernet, Justin Keery.

BT is upgrading 400 local exchanges with ADSL technology, which supports up to 8Mbit/s down existing telephone lines to the user.

Smart records... A smartcard for medical records, claimed to be the first of its kind, has been introduced by Gloucestershire firm EMR Medicard. Using technology from ORGA Card Systems, the card is aimed toward "at risk" people, including those with diabetes, epilepsy and heart related problems. Information stored in the card could be of particular benefit to accident and emergency crews, the firm said. The card stores details such as name, ID number and medical data. Several west of England A&E departments plan to start using the cards in January.

Retina chip implant may let Stevie see for the first time

Soul star Stevie Wonder, who went blind shortly after birth, apparently wants to undergo an experimental medical procedure to regain limited vision through the use of a chip implanted on his retina.

Wonder lost his sight as a premature1y-born infant after being given too much oxygen while in an incubator.

The procedure, which involves placing a chip on the retina and stimulating cells within the eye, and the visual centre of the brain, would in theory allow the patient to regain sight for up to 30 minutes at a time, according to published reports.

Such a device would have to run on very low power levels so as to not injure the eye by generating heat, says Gerald Chader, the chief scientific officer of the Foundation Fighting

Blindness. He described the area where the chip is implanted as having the consistency of wet tissue paper.

Such a chip would not confer full sight on its users, but instead allow them to see varying shades of light and shapes Chader said. ■

Spray without sputter

Ever tried spray painting? If you have, you will appreciate the benefits of an electronic system that can hear when your spray gun's about to clog. Such technology is not yet within the reach of domestic users, but as Roy Rubenstein explains, the prize-winning technique that can hear clogged jets does work. And it could well have implications in other areas.

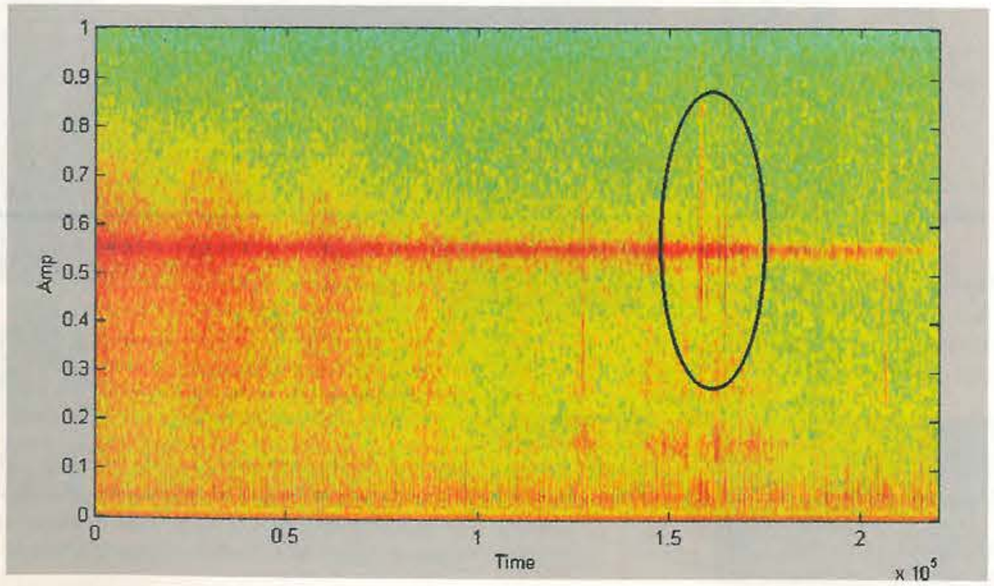
If you want to summarise the essence of DSP, it's about transforming signals or extracting information from them. The nature of the information being looked for can be extraordinarily subtle.

Take a car production line – and in particular the spray guns used to paint them. At present there is no way to give the user feedback regarding the state of the spray gun. Instead preventative cleaning is required if problems such as the gun clogging are to be avoided. Inevitably though, taking time to clean the guns affects productivity.

Joerg Kuechen of German firm Gavitec has looked at using the wavelet transform to monitor the spray process. Wavelet theory offers an alternative approach to the traditional fast Fourier transform (FFT) when analysing signals. In particular it is suited to tackling short duration signals, an area where the FFT falls short.

Kuechen has investigated applying the wavelet transform to the audio signal given off by the gun, to quickly detect any abnormalities. For example, air in the lacquer can cause 'spitting', resulting in 'bulbs' being deposited on the car's surface.

When the paintwork counts... Abnormalities such as spitting and bulbs can cause havoc with a car's paintwork.



Blobs... Shown is the occurrence of spitting - when air in the lacquer causes bulbs to be deposited onto the sprayed surface. The 2D plot was calculated using wavelets where the spitting – shown within the ellipse – may only last for 10ms durations.

His entry was the competition winner of Hunt Engineering's innovation competition. The Somerset-based DSP systems specialist has used the competition to gain application ideas for its products.

To monitor the spray gun, a small microphone is placed near the nozzle and its output signal is sampled at 44kHz to a 16-bit resolution. Kuechen's design uses two Texas Instruments TMS320C6201 very-long-instruction-word DSPs to process the audio signal.

The first DSP performs the wavelet transform, and passes the results to a second device for analysis. Up to 24 nozzles can be monitored using the two DSPs.

When the gun begins to clog, or the viscosity of the paint changes, the pressure and the air mixture can be changed to ensure the quality of the spraying. Productivity also improves by reducing the downtime needed for the gun's cleaning.

Hunt Engineering has supplied Kuechen with an eight channel a-to-d card and two C6201 DSPs. He has six months to implement the process control system.

According to Kuechen, the technique can be applied to a variety of industrial processes. One application already being considered is the control of a laser welding process. Here impurities in the material being welded, the focus of the laser and other difficulties can require that the laser power be adjusted. These difficulties can be tackled by applying the same technique to the sound of the welding process. ■

See www.hunteng.co.uk for more...

LISTEN with DSP

Fancy processors are all well and good, but what real benefit does a microprocessor powered toaster bring?

Digital signal processing (DSP) on the other hand is bringing tangible benefits in the form of digital hearing aids.

Traditionally analogue systems, the digital revolution has not passed hearing instruments by, and DSP is allowing manufacturers to use a whole new range of algorithms.

Swiss precision... Phonak's behind-the-ear hearing aids use its VLIW, parallel processing DSP architecture. The device manages around 130Mips at just 10MHz clock frequency. By the way, they're not normally transparent.

"Since the first hearing instruments appeared on the market the complexity has grown by about 100 times," says Gos Leenen, head of DSP and IC development at the newly merged Beltone/Philips Hearing Technologies. "The most complex devices on the market have more than one million transistors."

Companies shifting to the brave new digital world include the recently merged Beltone and Philips, GNReSound, Phonak and Siemens Hearing Instruments. DSPs are also coming from Mitel Semiconductor, Motorola, Infineon Technologies and Texas Instruments.

"We are in one of the few fields where chip area is a concern," Leenen at Philips points out. Just to make a chip and package designed to fit snugly in the ear canal is a major task. To then make it last a whole week before changing the battery is another incredibly difficult challenge.

Most companies use a two or three chip configuration. Philips has a complex Asic for all signal processing and data conversion, while the second chip is an EEPROM.

Philips' Asic holds everything except the non-volatile memory, including a-to-d converter, DSP, microcontroller and their associated RAM and ROM, oscillators and a special class D output stage. "Then we have quite extensive power supply circuitry," says Leenen.

The DSP is designed specifically for the hearing algorithms required. "We use very long instruction words (VLIW) to do up to ten things in parallel," says Keenen. "We want to do things in parallel to keep the clock speed low."

Therefore voltage can be reduced and hence power is lower. More parallelism leads to larger area, but the trade off in terms of overall power consumption is better this way.

Current drawn by the device depends on the algorithm loaded at the time, which can be changed by the user. The least complex algorithms consume about 0.6mA, the most complex around 0.85mA. The latter gives about five days battery life.

The custom DSP approach has also been used by Swiss firm Phonak for its latest digital hearing instruments.

Hearing must be one of the most important of the senses, and DSPs are coming to the aid of the deaf. Richard Ball reports on how.

"We have a very special tailored architecture," says Christian Berg, R&D manager at Phonak. Parallel execution units and a VLIW architecture allow for low clock speed and low voltage.

"The highest clock frequency we have is 10MHz, but we still have hundreds of Mips," Berg points out.

Like Philips, Phonak insists that a custom DSP is needed in such a power sensitive application. Voltage and hence clock frequency must be kept low, which means using a parallel processor.

A general-purpose DSP would have an order of magnitude greater power consumption, Berg claims: "They would have to drain 10mA."

Phonak uses several further techniques to reduce power. "We try to optimise the system for minimum memory access," Berg says. The DSP uses gated clocks and the multiple execution units.

The very long instruction words to the DSP can be up to 650 bits long.

The DSP and pulse width modulation controller for speakers are fabricated in a 0.25μm digital chip, while three a-to-d converters, FSK remote control receiver and power management are made in a 0.35μm mixed signal process. Along with a 64K EEPROM, the chips are stacked together in a single package.

"At 1V it used to be exotic – but not any more," says Berg.

In order to reduce problems as much as possible, Phonak always tries to use well understood semiconductor processes. "We have a 0.35μm mixed signal and a 0.25μm digital process and the technology is absolutely superb," Berg says.

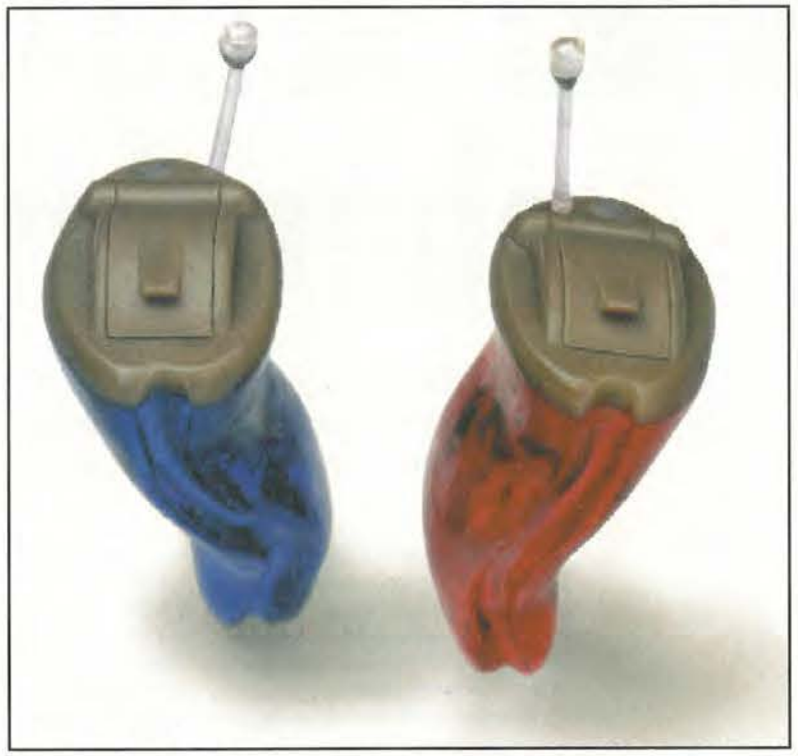
Using unknown, exotic technology is not on the cards: "You're bound to run into trouble."

Texas Instruments (TI) has a roadmap for its C5000 family of DSPs that leads to 0.9V devices next year. These, the firm says will be targeted at, among other things, hearing instruments.

"We see the hearing aid market as very demanding in terms of power," says Gweltaz Toquet, TI's manager for C5000 cores. "Next year we will be able to provide 30Mips at 0.9V."

However, the use of Mips as a metric can be misleading, as these types of processor often have special instructions replacing several simple instructions.

"The C5000 has some very specific instructions such as FIR that can reduce the number of instructions and therefore the power consumption versus general purpose DSPs," Toquet points out.



Small is beautiful... Hearing instruments that fit completely in the ear canal are the smallest available. This poses some considerable problems for system designers – not least of which is getting the required DSP algorithms at the right power consumption.

The evolution of hearing instruments from analogue systems through hardwired digital to the latest programmable DSP powered models has brought software to the party.

There is an increasing importance placed upon software, Toquet says. Hearing instruments are being designed that allow a choice of algorithm, depending on the environment and these can be tailored to an individual's hearing loss.

Toquet says that although its lowest power C5000 devices will not be ready until next year, designers are using existing chips to test out algorithms and system designs. ■

Infineon's Carmel DSP and hearing applications

Infineon Technologies' Carmel DSP is also being touted for use in hearing applications. Like the Philips and Phonak devices, it uses a VLIW approach with multiple execution units. Carmel can perform up to 15 basic operations in parallel.

At normal voltages of 2.5V, power consumption is claimed to be 180mW at 120MHz. Dropping the supply to 1V (reduces power by 80 per cent) and dropping the clock down to a few megahertz would bring Carmel into the realm of the hearing aid DSP.

This idea of using VLIW and multiple execution units is gaining popularity in the DSP and microprocessor world. Trading off a larger area for a reduced clock speed results in lower power consumption, a trade off many designers are happy to make.

The IMEC phenomenon

It may surprise you that the largest independent research centre in Europe is in Belgium, and that last year it made 46 patent applications and was granted 14 patents. Richard Ball looks at IMEC.

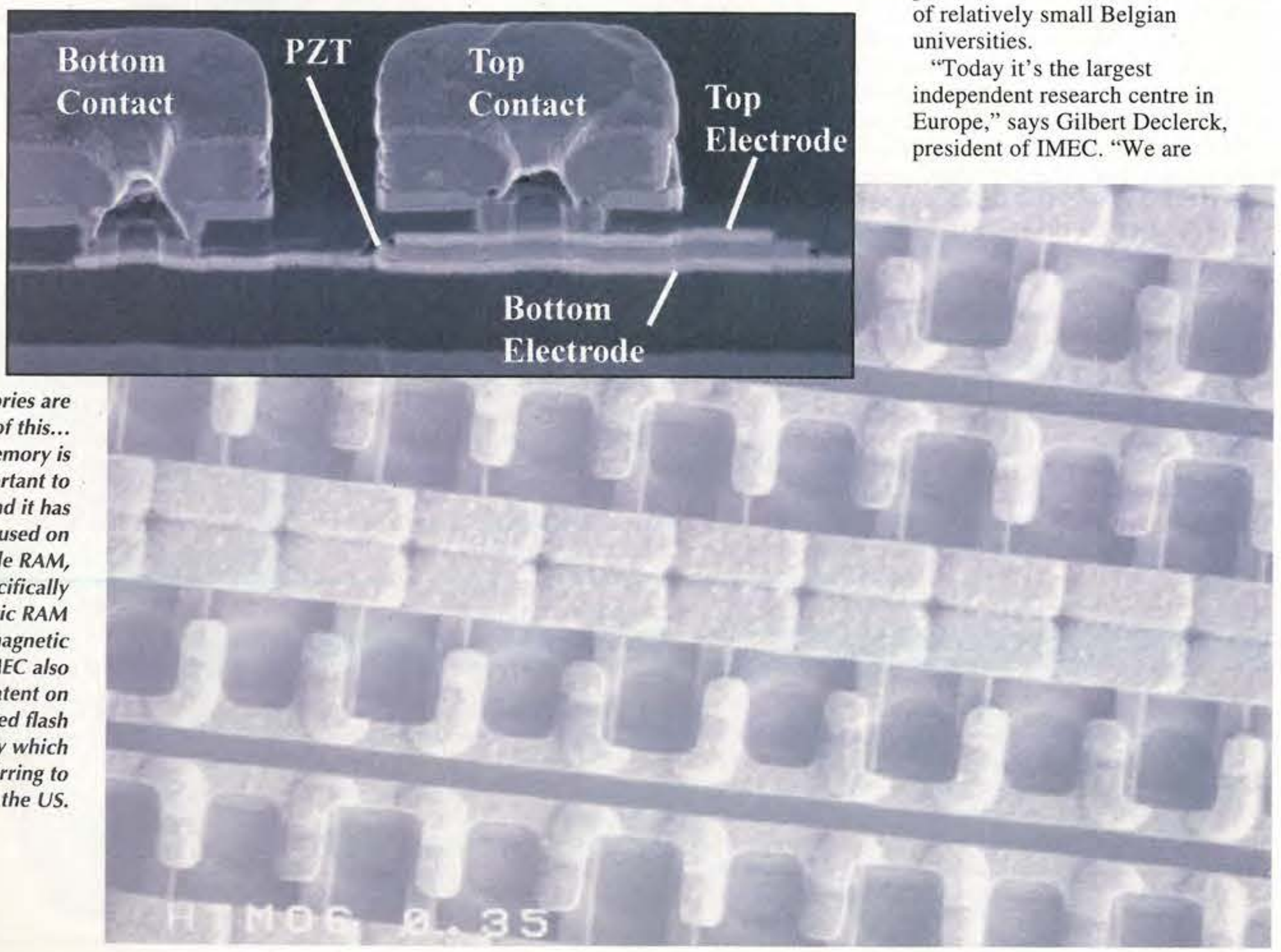
Tucked away in a small corner of Belgium is Europe's largest independent microelectronics research centre.

From the small university town of Leuven, IMEC has produced some of the most important semiconductor process technologies of the nineties.

From humble beginnings in 1984, IMEC currently has a budget of \$78m and employs over 850 people. The centre's research programmes are typically about five years ahead of commercial industrial needs. As such it works with almost every field of semiconductor technology from lithography and dielectrics to ferroelectric materials and high-level design tools.

Last year, the centre published 183 papers, delivered 372 conference papers, made 45 patent applications and was granted 14 patents – not bad for a collaboration of relatively small Belgian universities.

"Today it's the largest independent research centre in Europe," says Gilbert Declerck, president of IMEC. "We are working on the design of integrated information and communication systems, semiconductor process technology, silicon technology and device integration, material components, packaging and training."



Memories are made of this... Memory is important to IMEC, and it has focused on non-volatile RAM, specifically ferroelectric RAM and magnetic RAM. IMEC also has a patent on embedded flash technology which it is transferring to AMI in the US.

This year's contract research is worth \$45m. While just under a third comes from Flemish industry, the biggest source – over 40 per cent – comes from industrial partners, split 50/50 between US and European firms.

With EU funded research dropping away, much more work is being done with semiconductor firms from the US and Europe. "In order to cope with the complexity we are working with all the major industrial companies," says Ludo Deferm, IMEC's v-p of business development.

No single company could hope to develop all the critical technology themselves. Cooperation is the key, and for big companies such as AMD, Intel and Motorola to do so means IMEC is both trusted to be independent and up to the task.

At the highest level of IC development, IMEC is working on design methods, including a C++ development environment. VHDL is not good enough anymore, says Deferm, therefore IMEC is using C++ for an object oriented approach.

IMEC's programme is called OCAPI, which is testing at various companies.

In process technology, IMEC's major research area, it is figuring out how to continue scaling down the size of CMOS devices.

"We need higher frequency operation, so we need to scale down the CMOS process as fast as possible," points out Deferm. "The 193nm programme has been successfully started and is running at high speed."

193nm lithography is needed for 0.13, 0.10 and 0.07μm transistors.

"But this programme is not ready and will not be for two years, therefore it is behind what the industry needs," Deferm says.

Semiconductor firms have begun using 248nm lithography for 0.13μm chips, but are using exotic techniques such as optical proximity correction and phase shift masks in order to do so.

"Another programme which has just started, and is another bottleneck, is high-k gate materials," says Deferm.

Materials with high-k, a high dielectric constant, are needed in order to get a thinner gate insulating material without electrons tunnelling through.

"There are materials developed for DRAM, but these are not suitable," says Deferm.

Materials such as tantalum pentoxide and strontium titanium oxide don't interface directly to silicon, and can't be used in a standard high-temperature CMOS process.

IMEC recently agreed to work together with ASM International to use atomic layer CVD (chemical vapour deposition). This can put down layers of molecules at a time, producing well ordered structures with relatively few defects.

IMEC hopes this will help fix the gate oxide problem at the 100nm level.

At the back end of the chip process is the metallisation. IMEC is researching the use of low-k dielectric material and copper for metal.

Low-k means a value of less than 2.0 – half that of the silicon dioxide being used today. Reducing the capacitance of the material surrounding metal lines reduces crosstalk and improves propagation delay through long lines.

While researching these new materials will cause some headaches for semiconductor companies, the problems are not insurmountable, believes Deferm: "I don't see any limitation for CMOS in the next ten years."

Memory is also important, but IMEC's size limits its research capability, so it has focussed on non-volatile RAM, specifically ferroelectric RAM and magnetic RAM.

"We focused on non-volatile memory technology. We have a patent on embedded flash technology and we are transferring this to AMI in the US," Deferm says. "Also we started about five or six years ago a ferroelectric memory technology."

This is now being transferred to STMicroelectronics. ■

Magnetic memory

New materials research has led to a possible magnetic RAM technology, quite different to Motorola's attempts to develop MRAM. "We worked with tunnelling cells to see if we could use this as a memory," says Deferm. Like FRAM, but unlike other forms of non-volatile memory, MRAM can use a low voltage. "The problem with MRAM is power. To make a magnetic field you need a large amount of power," Deferm points out. The requirement is to develop materials that change resistance with a very small magnetic field, which would thereby reduce current and power. Magnetic materials also bring other problems, including a processing temperature limit of around 400°C. Therefore oxide layers need to be sputtered and not deposited in the normal way.

Packaging

Packaging technology is also on IMEC's agenda. It is advocating the use of 'system on package', or SOP. A consumer communications product such as a mobile phone could always use further integration. But putting digital baseband, RF and analogue filters on a single chip is horrifically expensive – if not impossible. IMEC's alternative is multi-chip modules, with active substrates to form passives such as inductors and capacitors for filters not easily realised on a chip.

Adaptable active speaker system

Unhappy with existing loudspeaker configurations, Christof Heinzerling set to work designing a three-way active system. It involves a squarewave-in, squarewave-out crossover network, electronic bass roll-off compensation and spherical enclosures for minimal diffraction.

I have enjoyed numerous live concerts, but I have yet to hear a sound system that could faithfully reproduce live sound in my living room. So I decided to build the ultimate loudspeaker system.

I investigated many technical papers before embarking on the design. There seems to be a lot of detail on bass reproduction, but there is little information on closed-box all-round concepts of the type that could suit my needs. But, after a lot of searching, trial and error, I believe I have found the solution.

Design goals

My wife appreciates good sound quality, but prefers it to come from an enclosure that is ideally invisible. Despite the enclosure size restriction, I decided that the bass should extend

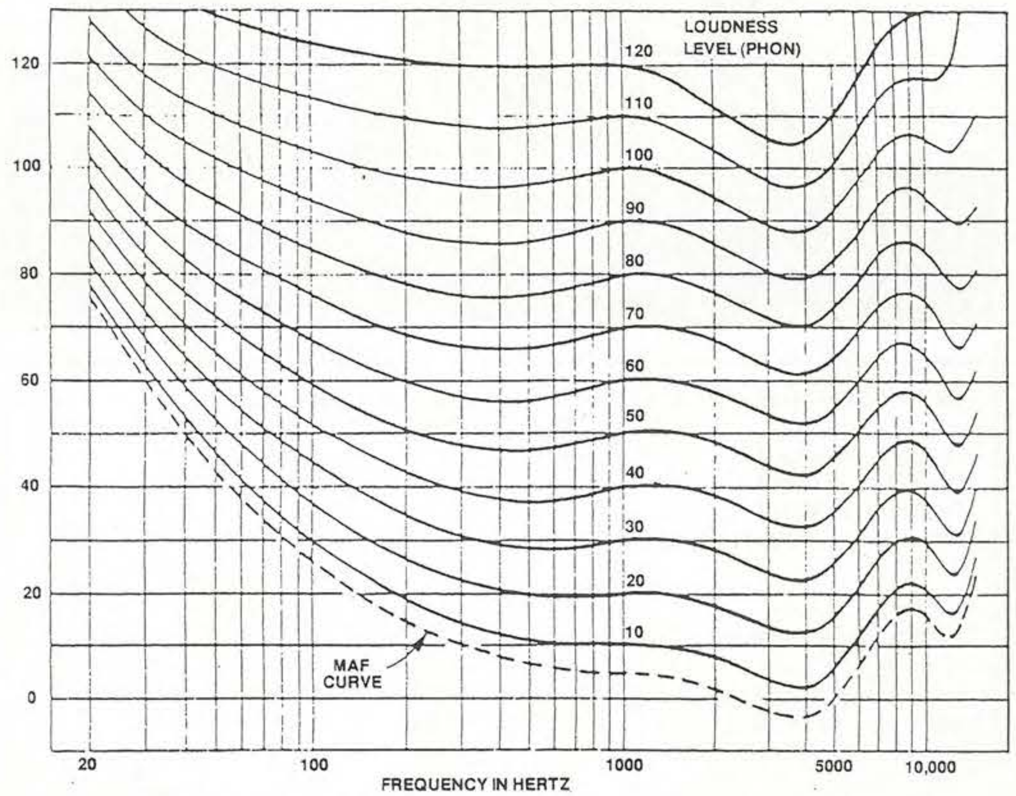
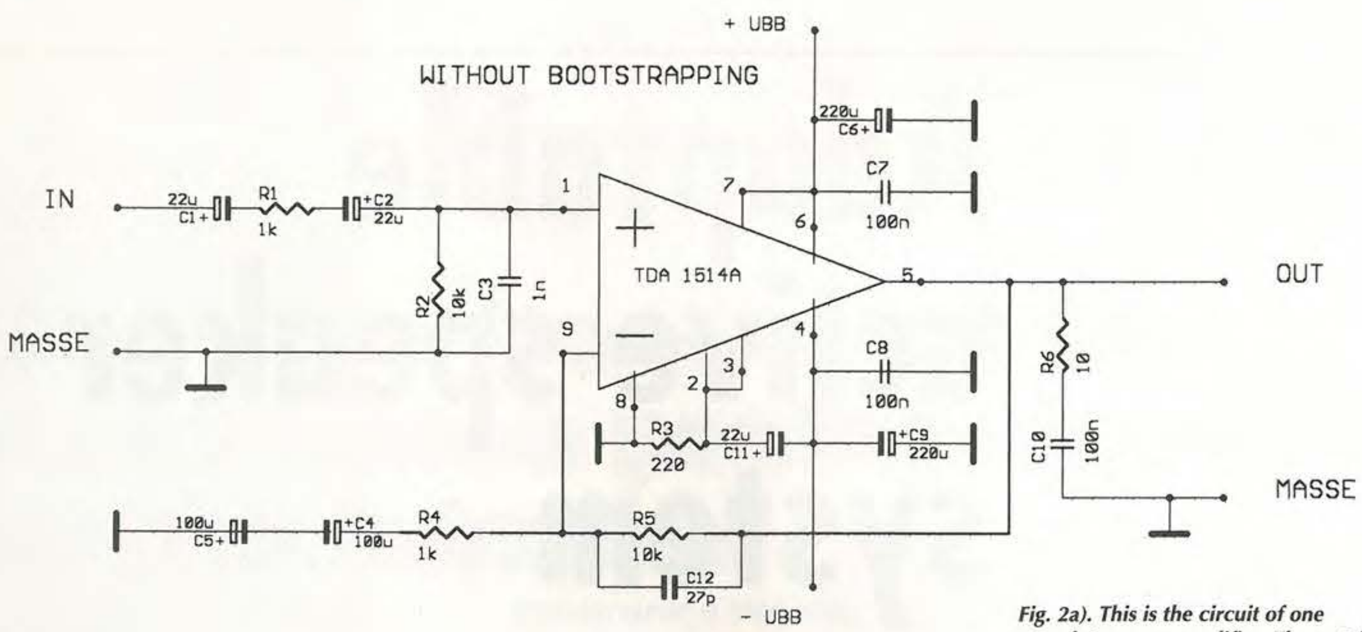


Fig. 1. If the bass unit's response is to be lowered from 40Hz to 20Hz, a sound-pressure increase of nearly 20dB is necessary to achieve the same perceived level, as these loudness contours show[2].



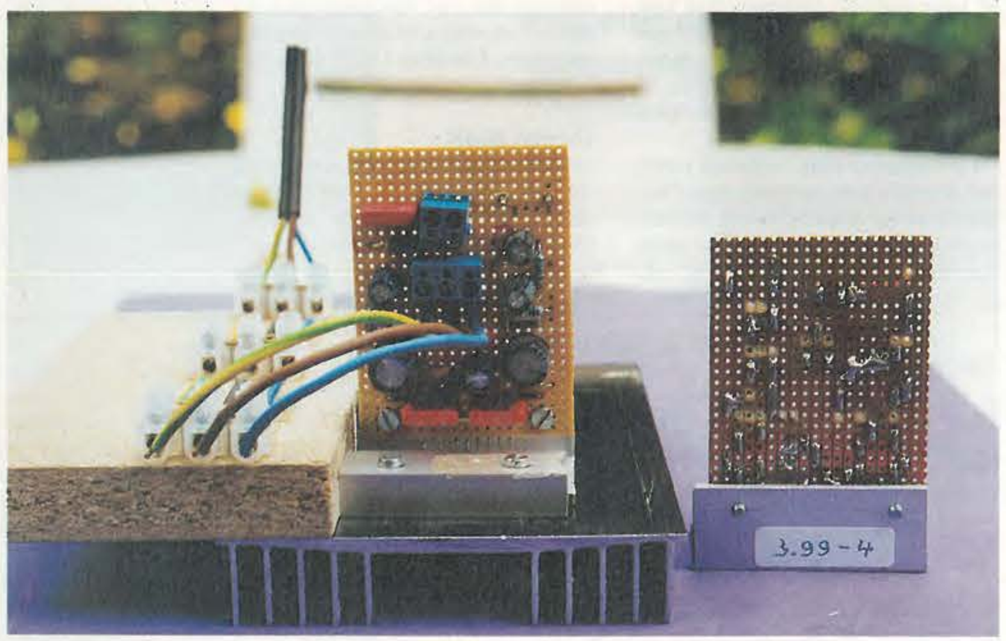


Table 1. Specifications of the TDA 1514A 50W high-performance audio amplifier.

	min	typ	max
Supply voltage range	±10V		±30V
Peak output current	6A		
Total quiescent current			90mA
Power out at ±23V, THD –60dB,			
R_L=8Ω		28W	
R_L=4Ω		48W	
Total harmonic distortion at 32W		–90dB	
Slew rate		14V/μs	
Closed loop voltage gain	20dB	30dB	60dB
Signal-to-noise ratio at 50mW	80dB	83dB	
Output offset voltage		7mV	200mV
Input bias current		0.1μA	1μA
Supply voltage ripple rejection	58dB	64dB	

Fig. 2a). This is the circuit of one complete power amplifier. The metal area on the back side of the chip connects the negative supply voltage $-U_{BB}$. The reservoir capacitors are soldered directly to pins 4/6. Corner frequencies are 3.2Hz and >100kHz.

Fig. 2b). The strip-line board measures 5cm by 5.5cm. Each is mounted on a 2mm-thick aluminium angle bracket measuring 20mm by 30mm and 50mm long. This bracket is large enough to allow testing. In use, the main heatsink is needed to remove additional heat cause by continuous operation.

down to 20Hz, and that it should be possible to build an enclosure within a book shelf.

My solution had to be affordable and easy to build. This meant using electronic modules that required little setting up, but nevertheless did not compromise the sound reproduction.

John Linsley-Hood found that 2-3W output is a sufficient sound level for normal listening requirements.[1] I believe that there is no need for very high-powered amplifiers in a domestic environment. If you have heard a live rendering of Mussorgsky's 'The Great Gate of Kiev', you will understand that it is impracticable to reproduce such crescendos involved in your living room.

Although the sound system described here is designed for quality rather than quantity, it can still produce high-quality loud music in a domestic environment.

The concept

In **Fig. 1**, taken from reference 2, you can see that at 20Hz, around 80dB is necessary to produce the same sound pressure in contrast to 10dB at 1kHz. Similarly, 100dB at 20Hz is necessary to produce the same pressure as 60dB at 1kHz. A large bass membrane area is indicated by these figures. To keep the enclosure small, it is possible to mount the woofer in the side of the cabinet, but this degrades the stereo image, so multiple smaller speakers are a good compromise.

In spite of the very low 20Hz requirement, I chose closed box, as suggested once in Speakers' Corner[3]. As the sound pressure rolls off at exactly 12dB/octave it is possible to compensate for the roll-off with well known filter circuits.[4,5] So a blameless physical – and also mathematical solu-

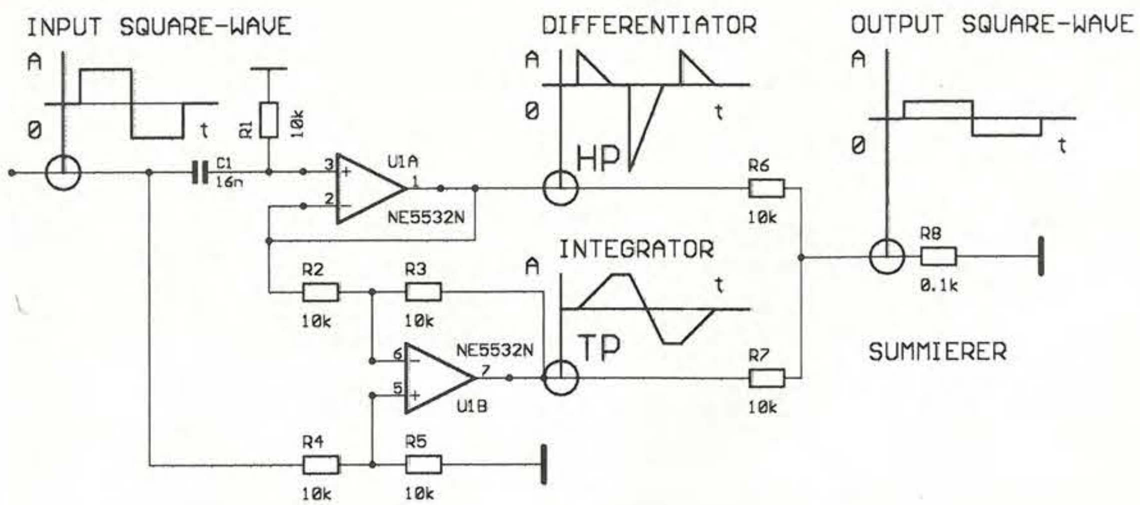


Fig. 3. Principle of the applied crossover-filter in its simplest form. It shows that an input square-wave is exactly reproduced as an output square-wave. Input steps will not be distorted, resulting in a 'correct-step difference filter'.

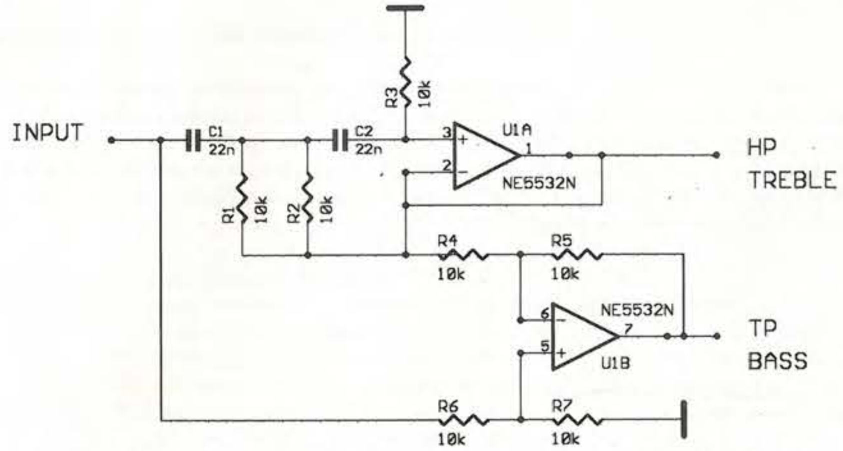


Fig. 4a). Simple high-pass filter is replaced by a second-order Butterworth high-pass filter ($R_1 // R_2 = \frac{1}{2} R_3$) with a corner-frequency of around 1kHz.

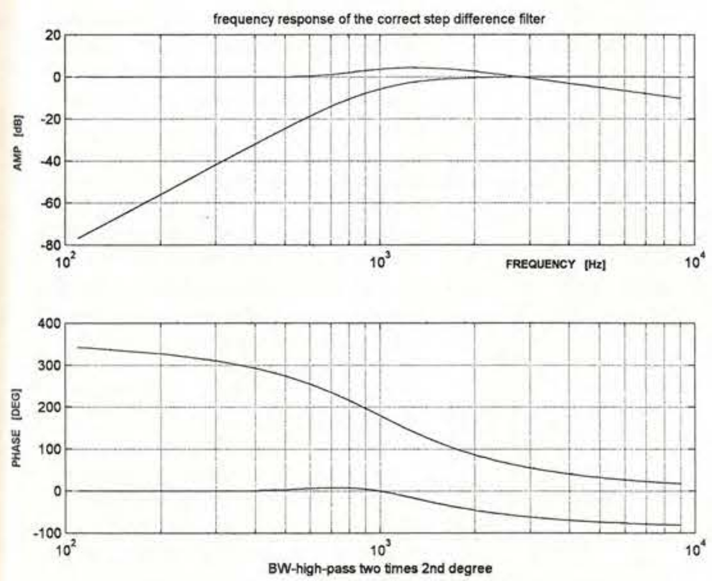


Fig. 4b). Amplitude and phase response of the high-pass section are standard. The response of the low-pass channel is for illustration purposes only. Response roll-off is poor, at only 6dB/oct independent of the order and character of the used filter.

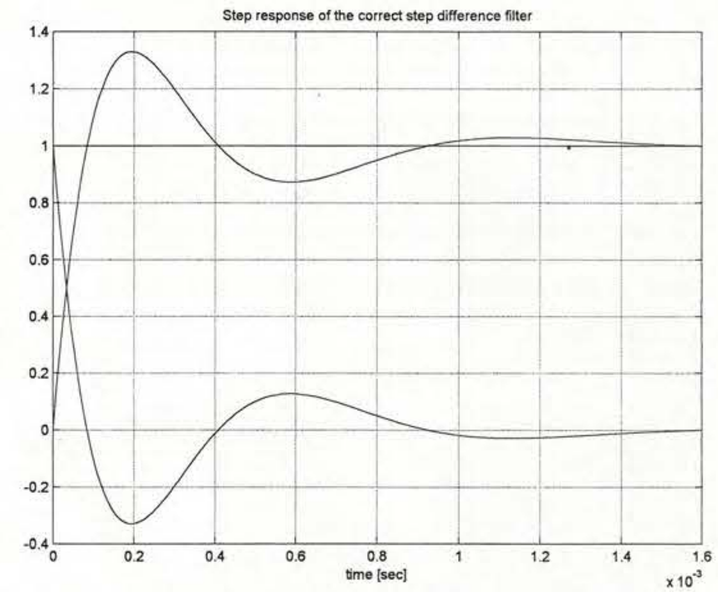


Fig. 4c). The step response of the high-pass filter shows overshoot but the difference channel equalises it. That means that the overshoot of a Butterworth filter adds no distortion in the time domain.

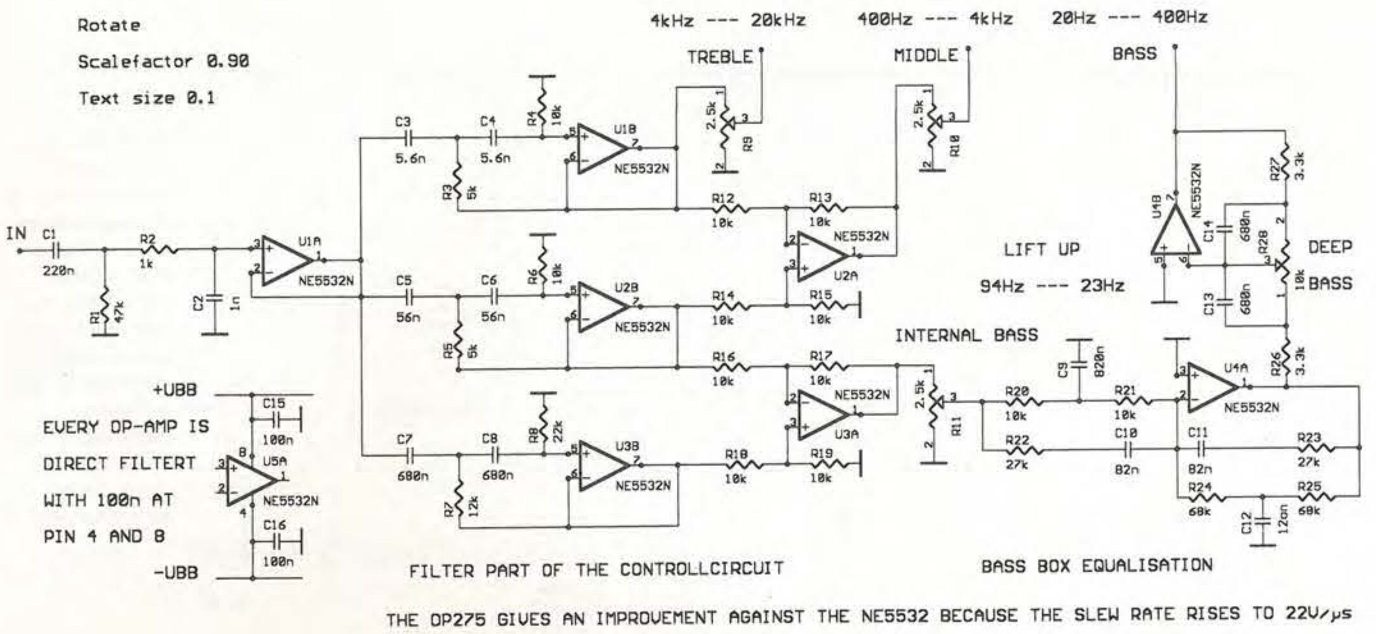


Fig. 5. This is the whole control circuit. Op-amp U_{A1} supplies the high passes of the three frequency bands. The upper frequency band is then subtracted from the next lower band in a difference amplifier. In this way, the corner-frequencies are reproduced exactly. The allowable load is 600Ω so linear potentiometers can be connected without buffering. These feed the inputs of the power-amplifiers or, in the case of the woofer, the bass filter. The bass filter provides box equalisation and bass-lift to compensate for attenuation in the sub-bass area. The frequency values represent a convenient distribution so adapting the sound level to room acoustics is easy. If you replace the NE5532 with an OP-275 you will have to pay double the price, but the difference is clearly audible.

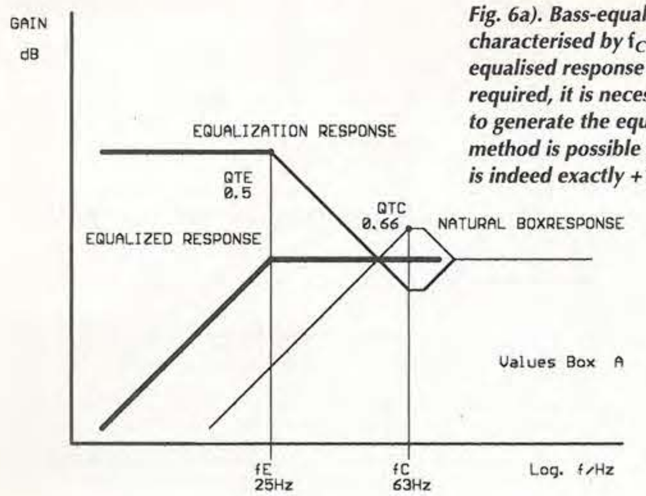
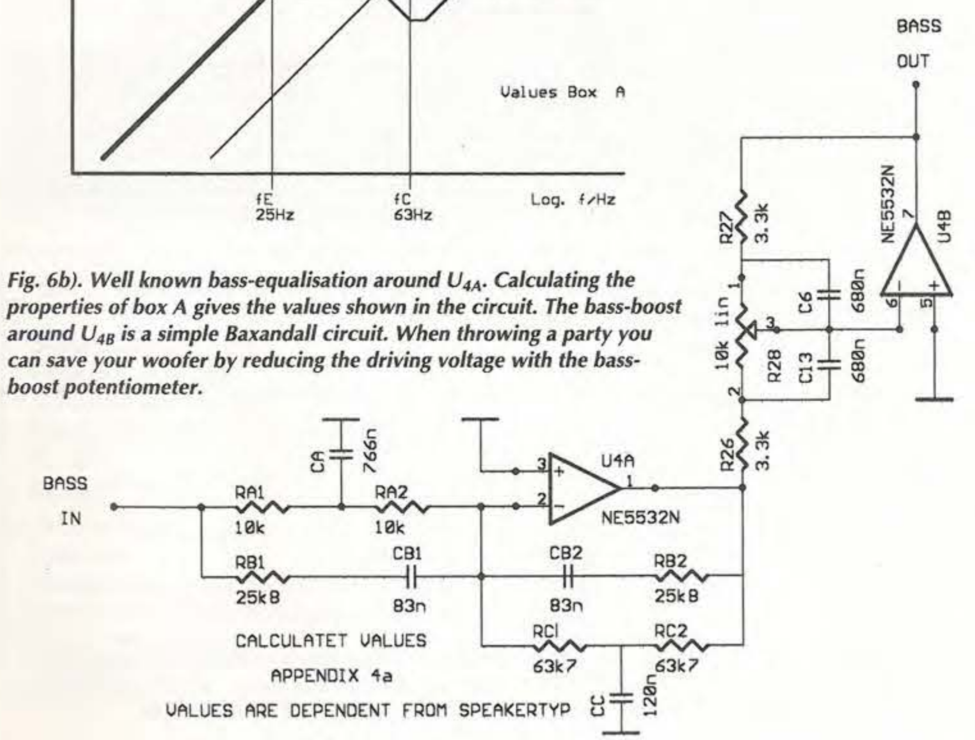


Fig. 6a). Bass-equalisation. The natural box response is characterised by f_C=63Hz and Q_{TC}=0.66. As an equalised response of f_E=25Hz and Q_{TE}=0.5 is required, it is necessary to implement the circuit in 6b) to generate the equalisation response. This simple method is possible because the slope of the closed box is indeed exactly +12dB/octave.

Fig. 6b). Well known bass-equalisation around U_{4A}. Calculating the properties of box A gives the values shown in the circuit. The bass-boost around U_{4B} is a simple Baxandall circuit. When throwing a party you can save your woofer by reducing the driving voltage with the bass-boost potentiometer.



tion – can be arrived at, resulting in clear bass reproduction.

Using two bass speakers in a front-rear arrangement results in excellent clarity[6]. The action of the front speaker is compensated by the reaction of the rear speaker. Even with high sound pressures, there is no movement of the loudspeaker cabinet and therefore there is no unwanted cabinet radiation.

Amplifier choice

My next task was to find a suitable amplifier. In reference 7, Ben Duncan explains that the TDA1514A from Philips Semiconductors is an overlooked jewel, **Table 1**. Its distortion spectrum shows that the third and higher harmonics fade away into the noise floor.

The chip costs around £2. Surrounding it with the necessary passive components mounted on a stripline board together with an aluminium angle heat sink results in an inexpensive high-quality 40W amplifier for about £5. Every speaker has its own power amplifier, receiving full power from an impedance of practically 0Ω.

To separate the frequency domains, I used a well-proven yet simple filter.[8,9] Because of its transfer character, I like to call it a 'correct-step difference filter'.

Since the power amplifier I chose is so easy to implement, I decided to design a three-way speaker arrangement. Each speaker handles a narrow

Is loudness control necessary?

Take a look at the phon values of the loudness contours at 1kHz and 20Hz in Fig. 1. You will notice a difference of 40phon if you look at the 1kHz-60phon contour. At the 1kHz-20phon contour you will see a difference of 60phon.

That means that if the sound image at normal sound pressure is balanced, there must be a considerable deficit if the sound pressure is reduced to a low level. And this deficit gets worse as the frequency gets lower. The same effect arises at the high-frequency end, worsening with the age of the listener.[13]

Because of these effects, correction at both ends of the frequency spectrum is needed to ensure that all parts of the spectrum are properly represented as the volume control is raised and lowered.

sound band, in spite of the fact that the slope of the filter difference output is only –6dB/oct. Since the sound level of each frequency domain is variable, it is easy to adapt the system to the room acoustics.

To prevent diffraction in the middle and high-frequency sections, I chose spherical enclosures.[10] This results in a stereo image of such clarity and width that I can say that, together with the extended bass reproduction, this is the sound of my dreams.

As all elements of the system are easy to calculate, this speaker concept is adaptable to different speaker chassis and room criteria.

The electronics

One complete power amplifier using the TDA1514A is shown in **Fig. 2a)**. Input components $C_{1,2}$ and $R_{1,2}$ form a high-pass filter with a corner frequency of 1.3Hz, a voltage division of 10/11 and a low pass of greater than 100kHz.

The non-inverting operation yields amplification of 11, nearly the lower allowable value proposed by Philips. Together with the divider the resultant voltage gain is exactly 10 at 1kHz.

The lower corner within the feedback circuit is formed by R_4, C_4 and C_5 and yields a frequency of 3.2Hz.[11] Components C_{11} and R_3 are proposed by the manufacturer and set the muting time.

Essential for stability is the R_6, C_{10} combination across the output pins and the direct mounting of the 220µF capacitors C_6 and C_9 on the strip-line board. Bootstrapping between pins 5 and 7 (220µ-150Ω-82Ω) enhances the output power by about 10%, but it was omitted to keep costs down.

Figure 2b shows the mechanical construction with a 0.1in-pitch strip-line board. It is mounted on an aluminium angle bracket to facilitate testing and assembling. The assembly needs to be mounted on a main heat sink with a thermal resistance of about 1.5K/W measuring 25cm x 10cm x 2cm. Note that the metal area of the chip back-side connects to the negative supply voltage $-U_{BB}$.

As the negative supply voltage is only 26V, it is not strictly necessary to isolate the chip from the sink. It is safer though to isolate the chip and to earth the main heat sink.

An essential part of this design is the 'correct-step difference filter'. As the music signals are extensively characterised by transients, I found that this filter is a good choice. An input step is exactly reproduced as an output step when the two output signals are summed.

In **Fig. 3** you can see that the input squarewave feeds an upper high-pass channel C_1, R_1 and a differentiated signal appears at its output, point HP. In the difference amplifier around U_{1B} this signal is subtracted from the original input signal, resulting in the wave form of an integrated signal at point TP. Summing these two outputs at R_8 reconstitutes the original square-wave form, but with reduced amplitude due to the voltage divider.

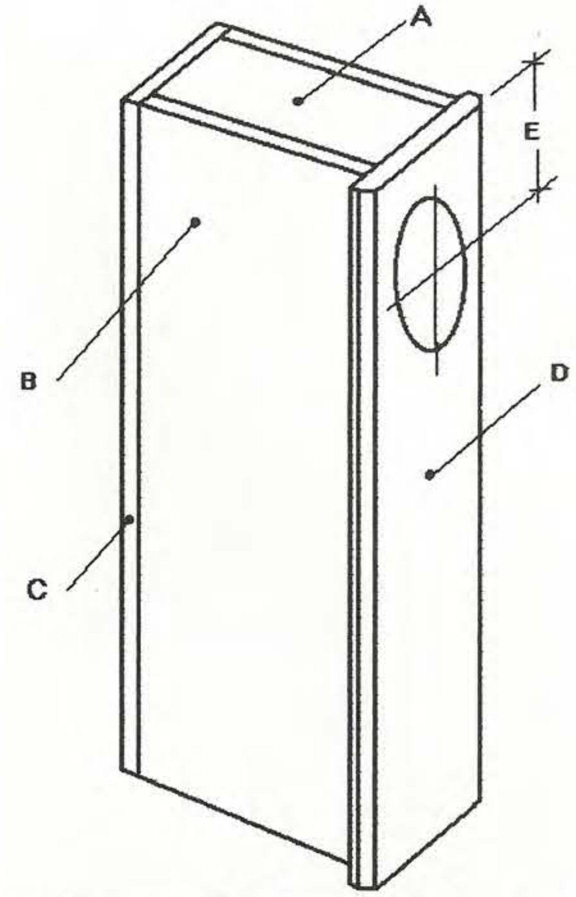
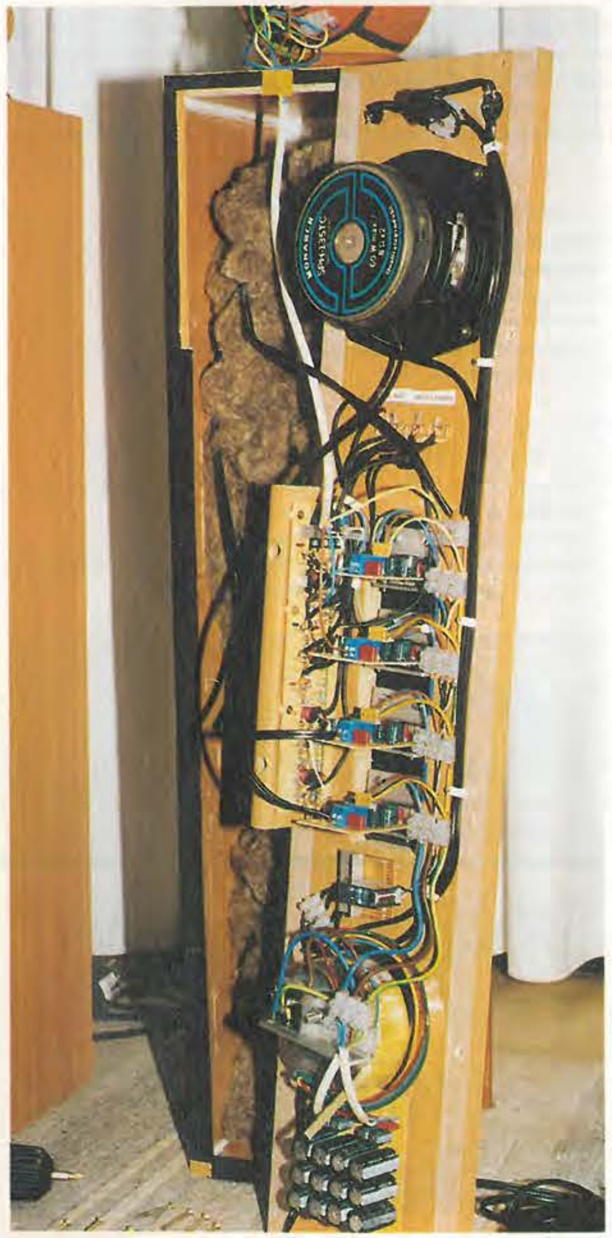


Fig. 7. Construction of the bass enclosure is simple. As the front panel extends beyond the periphery of the box, the construction seems smaller. The minimum speaker diameter determines board A. The rest is a compromise between box volume, box proportions and aesthetics. Some proposals are made in Table 2.

Fig. 8. This is the inexpensive material-mix to build the spherical boxes. The styropor spheres are an optimal acoustical solution as the styropor damps well and the sphere is the optimal form for radiating sound. The soft surface can be stabilised with ordinary papier mâché. If you coat all this with mineral material you get an attractive enclosure.

Fig. 9. Both faces of the cabinets are smart. The 5.5in 60W mini-woofer sets the width at 19cm (by 82cm by 23cm) of the bass enclosure. The spheres have a diameter of 15cm and 7.5cm.

Fig. 10. The rear board carries all the circuits. The electronics side of the design is easy to test. Note the damping-wool.



Filter details

Figure 4a) shows a second-order Butterworth filter in the high-pass section. It has a corner frequency of 1kHz. The values shown result in an amplitude of –3dB, a phase of +90° and an ascending slope of 12dB/octave at the HP output. All this is well known. If you subtract this HP signal from the input signal you get the frequency responses in **Fig. 4b)**.

The high-pass section is conventional. The low-pass section is only for illustration purposes. From it, you can ascertain that there is a slope of only –6dB/oct. This slope is independent of the degree and character of the high-pass filter. Its amplitude is about +1.9dB and it has a phase angle of –34° at the corner frequency of 1kHz.

The slope of only –6dB/octave is a disadvantage and seems to call for a speaker chassis with an extended frequency range. But if you look at the double corner frequency of 2kHz you will see that the high-pass stage has reached its final amplitude and a phase of 43° is measured. The difference channel is now unnecessary. There seems to be no need for an extended frequency behaviour of the bass unit above double the corner frequency.

In the time domain, the properties of the filter are excellent. A Butterworth filter shows overshoot, which results in undesirable distortions in transient musical events. But if you look at the step response of filter **Fig. 4c)** you will see that this over swing is compensated in the difference output, making it insignificant. In other words, transient events will be exactly reproduced independent of the applied filter characteristic, resulting in clear sound.

On carrying out blind tests listening to music reproduced with various filter types, the 'correct-step difference filter' came out best every time, producing a natural sound.

Control circuitry

Now I'll explain the control circuit, **Fig. 5**. In reference 12, Ben Duncan tested the distortion behaviour of audio op-amps and of those tested, the NE5532 came out best. But as this amplifier is compensated to provide stability at unity gain, its slew rate is only 9V/μs.

If you replace this amplifier with an OP-275, the slew rate rises to 22V/μs and a clearer sound results. The voltage noise density is slightly higher, but this is of little consequence. Its price though is double that of the NE5532, at about £1.

Distortion is well below 0.001% if the maximum signal amplitude stays below 3V. In the interests of low noise and minimal errors, 10kΩ resistors were used throughout.

The front filter is straightforward. Determined by R_2/C_2, the upper corner is high enough, at over 100kHz. Op-amp U_{1A} drives the three high-pass filters for the treble, the mid-range and bass.

The treble filter has a low corner frequency of 4kHz. Its output, at op-amp U_{1B}, feeds the treble potentiometer whose inner resistance doesn't exceed around 600Ω. This treble signal also connects to the inverting branch of the upper difference amplifier U_{2A} to limit the mid-frequency domain to 4kHz.

With a corner frequency of 400Hz, the mid-range filter works in the same way. This frequency is relative high. It allows the use of small mid-range speakers with low membrane mass, while the woofer only has to work up to 1kHz.

The second-order high-pass in the bass branch, comprising $R_7/R_8/C_7/C_8$, works together with the high-pass of the input circuit C_1/R_1. This yields a third-order Bessel high-pass filter with 20Hz lower corner frequency at the output of op-amp U_{3B}. At the output of op-amp U_{3A} is the bass channel, which drives the internal bass potentiometer

As mentioned above, the closed-box equalisation is easy to realise. In **Fig. 6a)** the fine line shows the natural response of the built in speakers with f_C and Q_{TC}. Correction values f_E and Q_{TE} are chosen to produce the bold line, i.e. the equalised response. Equalisation is provided by the well known network shown in **Fig. 6b)**.

Calculations for determining the resistor and capacitor values are presented in a separate panel. The values here are for the small box 'A'.

Finally, there follows the deep-bass booster. It is an ordinary Baxandall bass control with a range of ±12dB and an upper corner frequency of 94Hz.

Setting up

A quick check of the signal voltages within Fig. 5 can be made to ensure proper working. Assuming an output power of about 40W at the bass speaker with a 4Ω load, an effective output voltage of 12.7V is needed, which requires 1.27V at the input of the power amplifier.

As there is bass lift, Fig. 6a), about 0.2V needs to be delivered to the internal bass potentiometer and also to the input. This is because the filter circuits here have a voltage gain of unity, which represents a good compromise between distortion and noise.

The control potentiometers are linear. As the power amplifiers have a capacitor input, no precautions are needed to suppress offset voltages at the output of the control circuit. As a result, the circuitry is simple and is safe.

Conventionally, the power supply is built with a toroidal transformer of 2x18V/120W and 16x2200μF/35V for box 'A'. Supply voltage of the control circuit is derived from the supply of the power amplifiers and stabilised by 7815 and 7915 regulators. Every op-amp's pin 4 and 8 are direct connected to earth via an 100n capacitor.

All the circuits are low impedance so there should be little hum and noise. With an input resistance of 680Ω there was –81dBV hum and noise at the woofer and –88dBV at the tweeter connections measured in JIS A.

Designing the enclosures

There are three cases to design for the different speakers. Firstly the bass speakers.

My experiments with vented constructions did not lead to a satisfying bass reproduction. In addition the electronic filters, mechanical constructions and the damping associated with vented enclosures were relatively complicated. So I chose the infinite baffle solution, which is very easy to damp effectively with sheep's wool. This agrees with reference 3, as above earlier.

In **Fig. 7**, front board D is 19mm beech plywood. Building the side and rear boards A, B and C of 16mm chipboard with beech imitation veneer is a cheap solution. The front and rear panel involve the baffle cuts.

To avoid precise wood working, the front panel protrudes on all sides around 1cm. The woofers drive in opposite directions, so there is no need for heavy wood construction. Dimensions for diverse layouts are shown in **Table 2**.

For the mid and high-range speakers, I stumbled upon an ideal, low-cost enclosure solution while playing with my granddaughter and her styropor sphere kit.

I used wool-filled styropor spheres with diameters of 7.5 and 10cm for the small 4cm tweeters. As the mid-range speakers need a small volume of air, hollow spheres of 15, 20 and 25cm diameter can be used depending on the enclosure configuration you choose.

In **Fig. 8** you can see the spheres with some speakers at the right side and other materials I used. At 2cm, the thickness of the styropor wall provides

Table 2. Details of the bass enclosure, Fig. 7, all dimensions in cm.

	Box A	Box B	Box C
Minimal speaker diameter	13.8	16,6	21
Board A, 19mm chipboard (x2)	14 by 20	17 by 25	21.5 by 25
Board B, 16mm chipboard	80 by 20	98 by 25	110 by 25
Board C, 16mm chipboard	80 by 17.2	98 by 20.2	110 by 24.7
Board D, 19mm plywood	82 by 19,2	100 by 22,2	112 by 26,7
Volume	21.5 litre	40.3 litre	57.4 litre
Distance E	approximately equal to speaker-diameter		

Fig. 11a). The boxes stand before the windows at a distance of 3.3m. They radiate into an acoustically-damped living-room. With a total height of 100cm and a width of 19.5cm they are well suited to the living-room, as you can see.

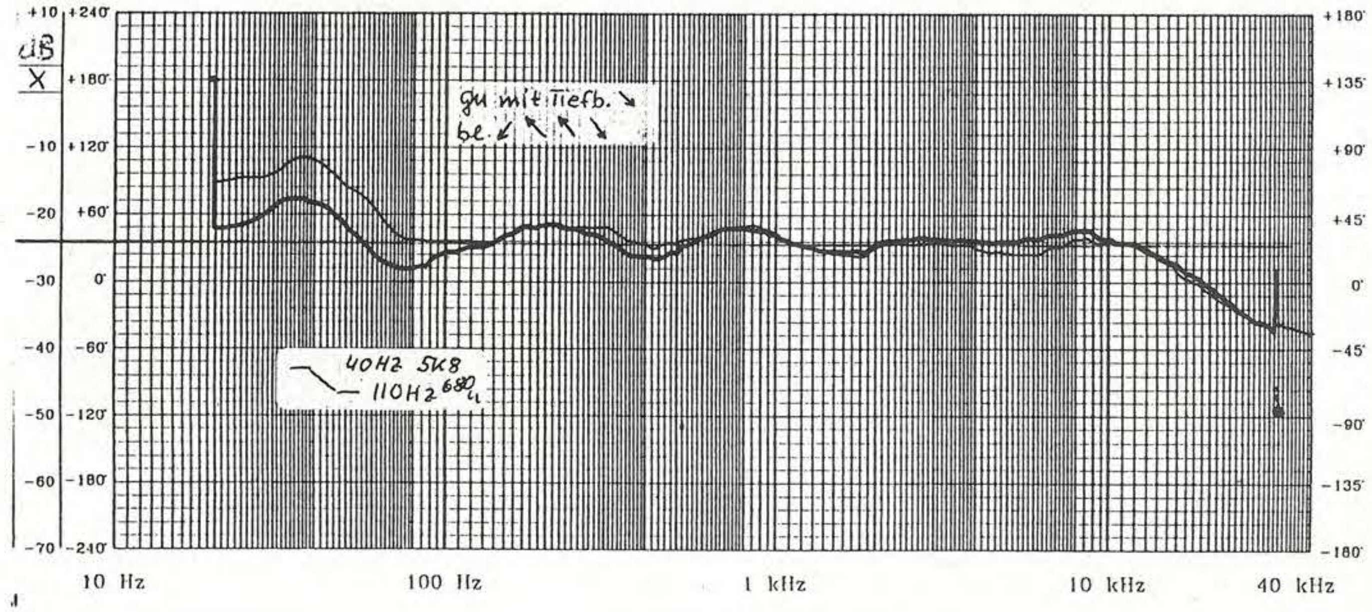


Fig. 11b). Measured relative sound level at a distance to both speakers of 3.3m, Box A. The curves are measured with averaging of $^1/_2$ octave. As the frequency spreads down to 20Hz, a cello sounds very good.

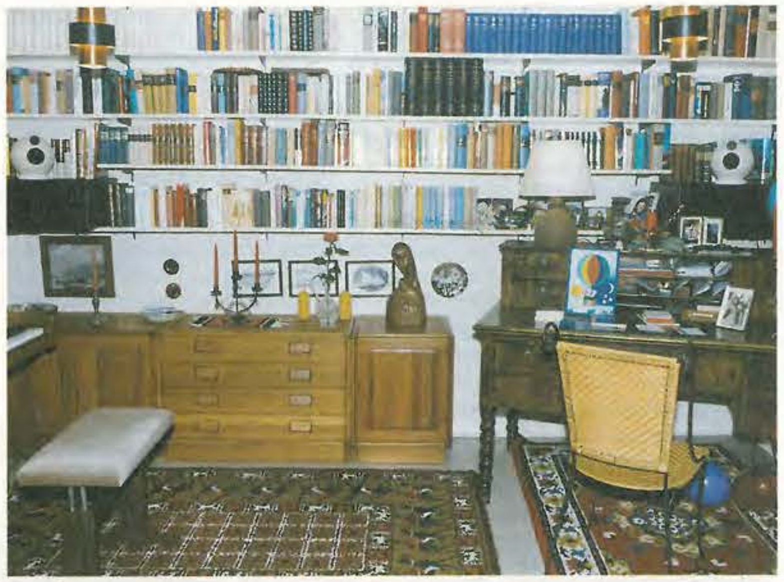


Fig. 12a). Here, the boxes are integrated within book shelves. The bass part radiates vertically in a push-pull manner. The distance between the walls is 4m while the distance of the mid-range sphere to the wall is about 40cm.

Fig. 12b). Construction of one arrangement shows the bass section with dimensions of 64cm by 44cm by 24cm and two 6.5in speakers operating in push-pull. The high-frequency spheres are 20cm and 7.5cm in diameter.

very good damping so no sound from the rear of the speaker radiates into the environment. The sphere surface cause no diffractions, and this yields a precise stereo image; the instruments are fixed in the room and the stereo area is enlarged.

The closed boxes are damped with sheep's wool. The speakers in the mid-range sections exhibit Q_{TC} values of about 0.6-0.7, which I can live with.

Because the surface of the styropor parts is very soft, they have to be stabilised. Here begins the artistic phase of the design, because I recommend coating the surfaces with papier mâché. The stability increases enormously, but each sphere requires about an hour of art work.

With the speaker mounted, the sphere will roll due to its weight so a counterbalance is needed.

A papier mâché surface doesn't look very nice. I applied a second coat of a cement-based material, which resulted in a very high quality surface. This coating increased the weight too, giving me a perfect enclosure for a total of about £2.50.

The tweeter spheres are so set up that their voice coils are vertically in line with those of the mid-range drivers, whose voice coils are in turn lined up with those of the bass units. In this way, all the speakers radiate from the same plane, resulting in phase-linear radiation. **Figure 9** shows the results of my efforts.

Mounting details

Mounting the speakers into the styropor is a little difficult. A 2cm deep hole

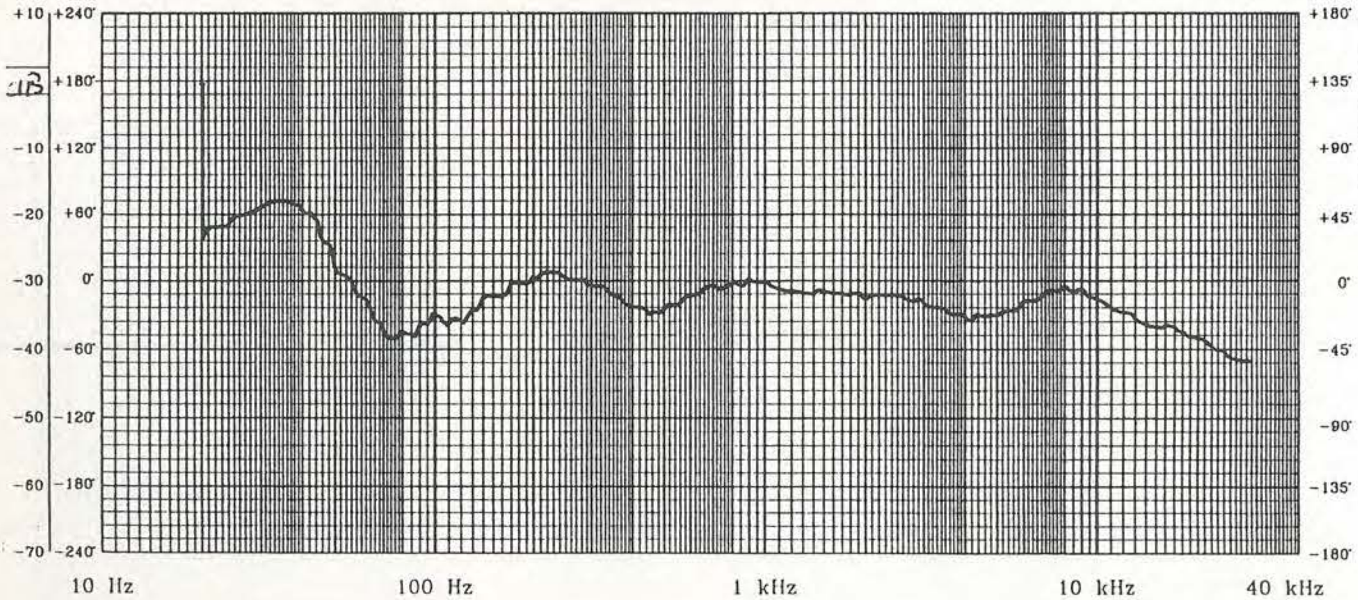


Fig. 12c). Relative sound-pressure level of the arrangement standing within the left edge, measured with an averaging of 1/3 octave. The measuring position is about 4m from the speakers built into the bookshelves. With no averaging, two heavy deviations appeared at 35Hz and 75Hz due to room resonances. As the Q values are very high, it is difficult to equalise this room influence with simple circuits.

Table 3. Determining component values for the low-bass equaliser.

	Box A	Box B	Box C
Possible speaker type	**SPH135TC**	**W170/4**	**SPH200TC**
Baffle cut-out	123mm	150mm	185mm
Free-air resonance	40Hz	37Hz	28Hz
Total Q factor	0.42	0.42	0.24
Complete equivalent vol.	2 x 16litre	2 x 40litre	2 x 65litre
Impedance	2x8Ω	4Ω	2x8Ω

From this speaker data and the volume information in Table 2, the following values can be arrived at, as detailed in the equations in the separate panel on page 164.

R_A	10k	10k	10k
R_B	25.8k	30.65k	70.45k
R_C	63.7k	102.2k	63.98k
C_A	765.6n	915.1n	1.234μ
C_B	83.1n	67.7n	80.21n
C_C	120.2n	89.56n	192.8n

Table 4. Speaker specifications versus the different boxes.

		Box A	Box B	Box C
Woofer		SPH-135 TC	W170/4	SPH-200TC
	f_s/Hz	40	37	28
	Q_{ts}	0.42	0.42	0.24
	V_{as}/litre	16	40	65
	Diam./cm	13.8	18.8	21.0
Mid-range		TPC80 RW-4	MSH-116/4	SPH-100 KEP
	f_s/Hz	93	95	55
	Q_{ts}	0.33	0.4	0.32
	V_{as}/Litre	1.9	2.7	6.5
	Diam./cm	8.0	12.4	12.5
Tweeter		HK 10 UF-8	HK 10 UF-8	KTN 25 F
	f_s/Hz	2800	2800	1800
	Diam./cm	4.8	4.8	4.8
Speaker price, approx.		£75	£82	£170

The SPH, MSH and SPH prefixed speakers are from Monacor, the W and KTN types are from SEAS and the TPC and HK types are from Gradient.

whose diameter suits the mounting requirements of the chosen speaker needs to be sawn. Next, stabilise the styropor surface with PVA glue.

Smaller speakers can now be fixed with silicone adhesive. With larger mid-range units, it may be advisable to glue a wooden ring on to the face of the sphere and screw the speaker to the ring, ensuring air-tightness,

I used screw terminals on the bottoms of the mid-range spheres to make it easier to connect the cables. To set up the spheres on the bass enclosure, a wooden board of appropriate thickness and with an appropriate hole is the simplest solution. Cutting a groove into the lower side of it allows you to make the cables invisible.

In all the suggested layouts, the rear panel is long enough to carry all the modules, **Fig. 10**. Mounting the reservoir capacitors at the lowest point is advisable since it is the coolest. Just above them should be the mains transformer. An inrush limiter comprising 2x100Ω resistors follows and in the middle of the panel there is the main heat sink with the four power amplifiers.

At the left-hand side are the control circuit and the cables connecting the level-setting potentiometers. The rear speaker comes next and at the upper end there are the power switch, the fuse and the LEDs indicating 'mains voltage present' and 'power on'.

As all sections and one speaker are mounted together, the whole arrangement is easy to test. In Fig. 10a), on the left within the enclosure, you can see the damping material. To avoid creating earth loops, do not connect cable screening at the power amplifier end.

In my environment, the speakers are set up before the windows, **Fig. 11a)**, so that they radiate into the damped living room. This results in an equalised sound pressure indicated as Fig. **11b)**.

There is no need to sit exactly between the two loudspeakers. Sometimes it seems as though the speakers are not working, yet the room is filled with music.

An alternative solution, where speakers built following the same design procedure are mounted in a book shelf, is shown in **Fig. 12a)**. The black boxes you see are bass units measuring 64cm by 44cm by 24cm.

Looking at Fig. **12b)**, you can see the push-pull arrangement of the bass speakers radiating in vertical direction and the associated spheres. The influence of the corner position is evident in Fig. **12c)**. The extreme values at the lower frequency end are caused by room influences.

This alternative uses very inexpensive speakers. In every case the sound pressure achievable is astonishing, proving Mr Linsley-Hood's statement.

In summary

Listening to music from these boxes comes close to my dream. They sound good, but they also look good – so good in fact that my wife finds them acceptable!

Speaker sources

Seas and Monacor loudspeakers are available from Wilmslow Audio, tel. 01465 286603, www.wilmslow-audio.co.uk.

Gradient products can be found at www.speakerland.com.

Continued on page 164...

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Power factor and how to measure it

Michael Slifkin and Yosef Chernobrodov explain the importance of power factor and present a means of measuring it.

Michael and Yosef are with Jerusalem College of Technology in Israel

Power factor is one of those parameters that engineers learn about at the beginning of their careers but often have only a sketchy idea of what it means. Yet, an understanding of power factor and what it implies is important to anyone working with electrical and electronic equipment.

Four parameters describe the quality of the mains supply and all are important in their own way. These are the frequency, the voltage, the harmonic content and the power factor.

At one time, but less so now with the advent of cheap quartz clocks, the mains frequency was used to drive clocks and synchronous motors. There was hardly a lounge or parlour in the land that did not have an electric clock on the mantel driven from the mains.

Except during times of high loading and voltage shedding, the mains frequency is maintained at 50Hz to a high degree of accuracy. So accurate were the clocks maintained that the CEGB would, in the event of power shedding and a drop in frequency, make up the cycles lost during the night so that next morning the clock would be accurate again.

At one time, it was not uncommon to run choppers from synchronous motors driven from the mains via a gearbox to give a range of accurate rotational frequencies.

In the UK, a voltage drop is very rare but in less developed countries it is not uncommon to have nominal voltage fall to as low as 180V from 230V during peak usage. An instrument not seen in the UK, but common in such countries is a small plug-in voltmeter fitting in a domestic socket to monitor the mains voltage.

Harmonic content

Harmonic content is a big problem in all situations where there are appliances that use a lot of electricity, such as inductance motors. These can create havoc with electronic instruments. Large spikes caused by switching heavy loads in the near vicinity can cause severe overloading and the destruction of ICs on a large scale.

The power factor is a less obvious phenomenon, but not one that should be ignored. There seems to be a lot of misunderstanding about the power factor. During the course of preparing this article we surfed the Web to see what others had written and found at least two Web sites which stated that the power factor could take values from zero to infinity. One of the Web sites actually belongs to a firm manufacturing power factor correction equipment! These have since been corrected.

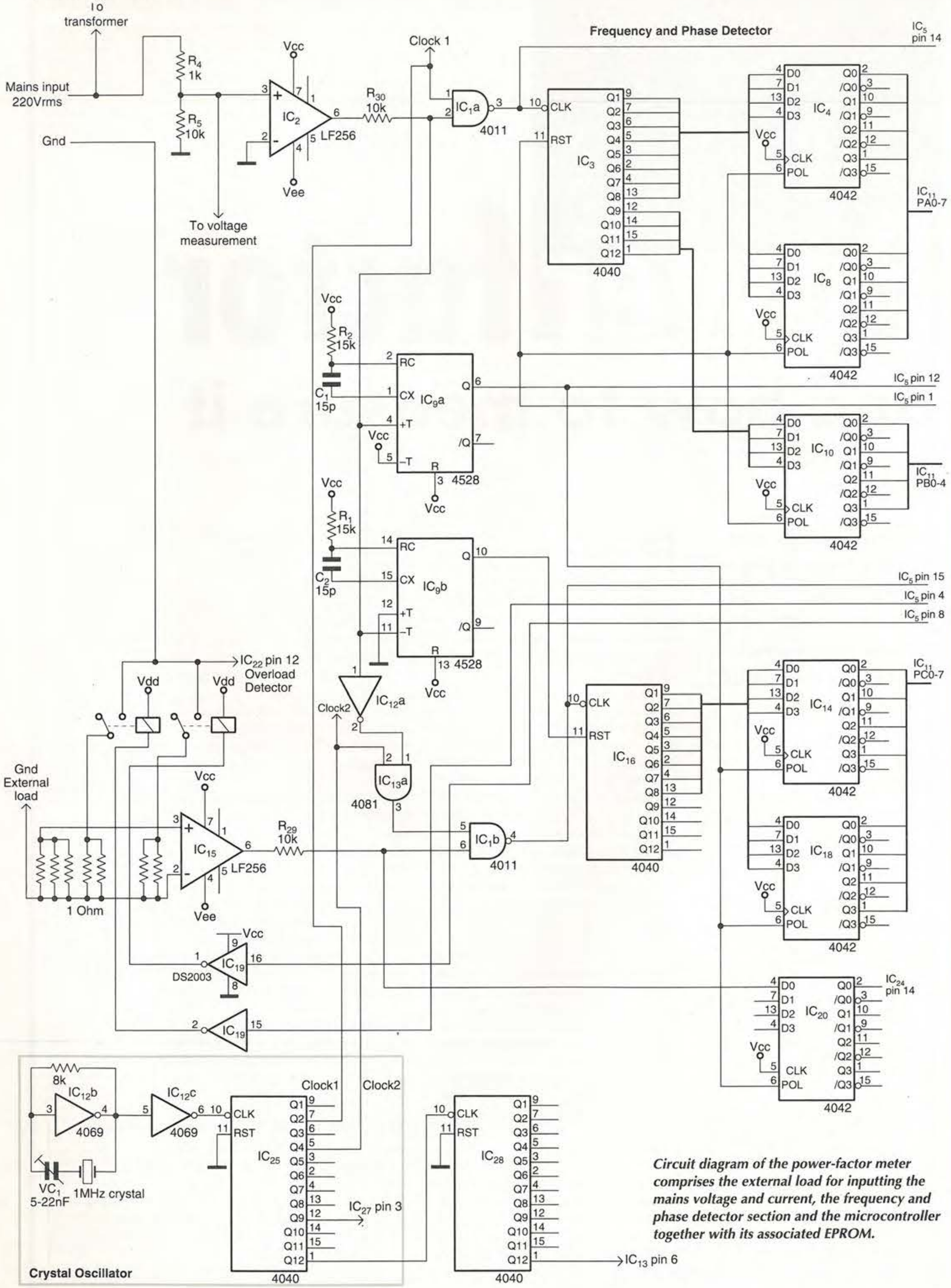
So what is power factor?

In an AC circuit, power is expended when current flows through a resistive element such as an electric fire. The

Specifications of the power factor meter

Input voltage	60 to 240V rms
Current input	500mA-17.5A
Frequency range	40 to 70Hz

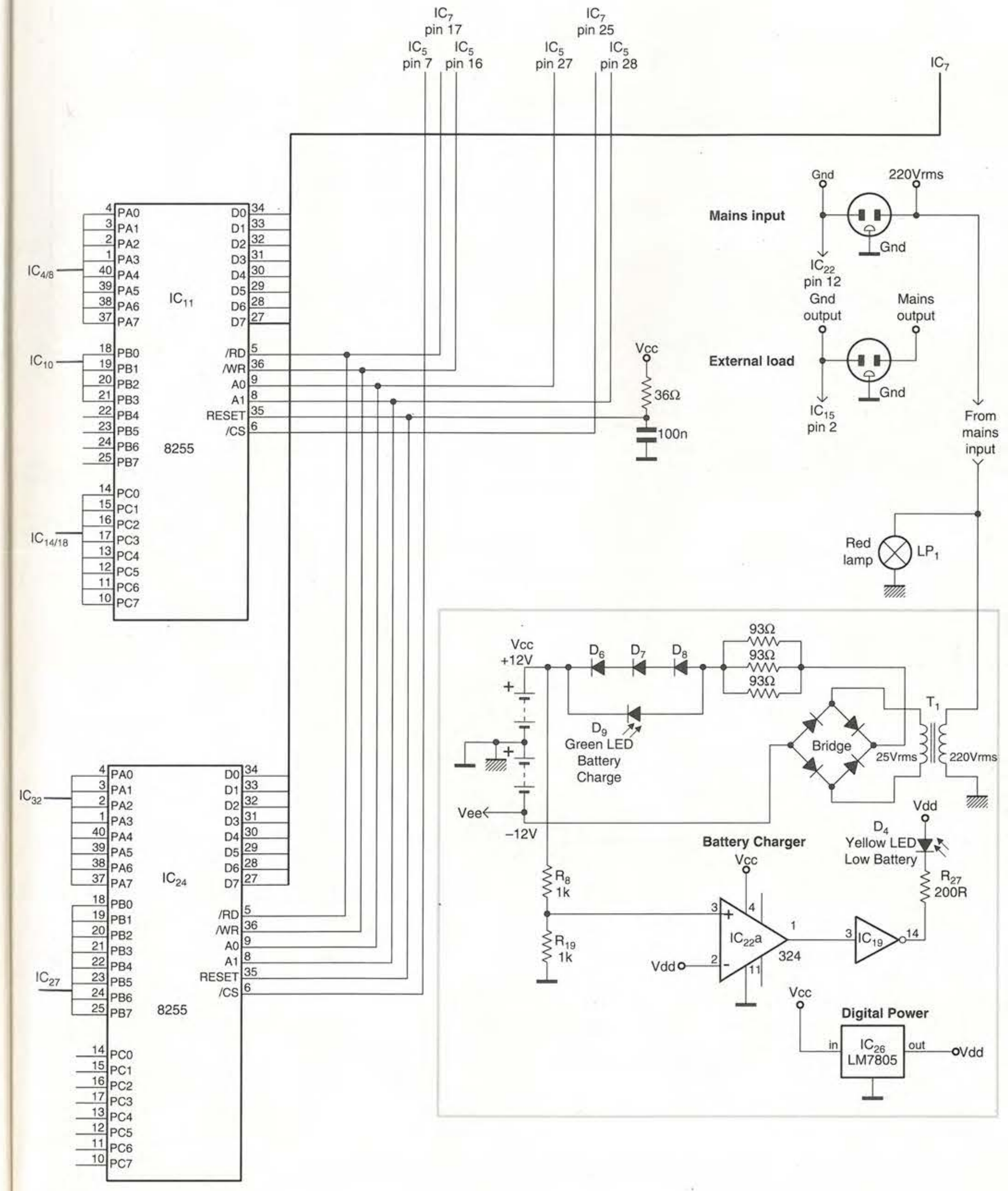
Power factor and cosine to one digit in second place after the decimal point. Frequency is accurate to 0.1Hz, the phase angle to 1° and the voltage to 1V.



Circuit diagram of the power-factor meter comprises the external load for inputting the mains voltage and current, the frequency and phase detector section and the microcontroller together with its associated EPROM.

power used is recorded on the power meter supplied by the electricity company.

For peak current *I* and peak voltage *V*, the power is simply *V*×*I*/2. If there are reactive elements in use such as capacitance or inductance often associated with fluorescent lights and electric motors, then although current flows through these, there is no consumption of energy even though extra current is being drawn from the electricity supply. This is because the current is exactly 90° out of phase with the voltage. Unfortunately for the electricity companies, this extra current is not recorded on the power meter.

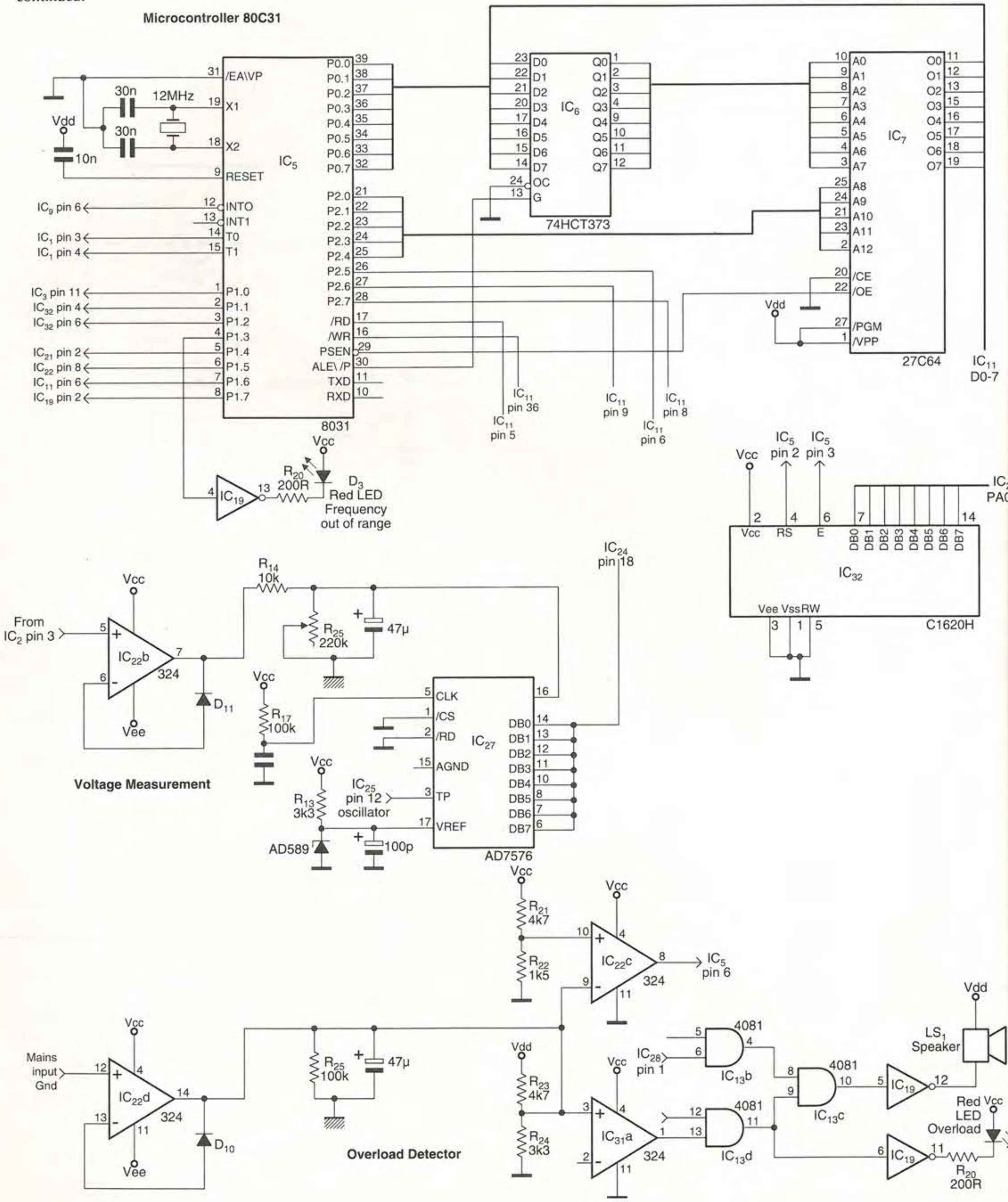


In capacitance, the current leads the voltage whereas in inductance, the voltage leads the current. In most situations, the power factor is likely to be influenced by inductance. Most heavy consumers of electricity use machines such as motors that are inductive.

This situation is termed 'lagging power factor' as the voltage lags behind the current. Hence, most power correction systems normally consist of banks of capacitors, which are switched automatically to adjust the power factor upwards.

Continued on page 147...

Circuit diagram continued.



Top notches

Holes and poles – active filters come in all shapes and sizes, but one important class is the notch filter, and its inverse, the 'notch pass', or narrow band selective filter. The notch knocks a narrow 'hole' in the frequency response of signal path, while its inverse, the narrow band-pass produces a frequency response like a tent-pole sticking up from zero response. Here, Ian Hickman looks at various implementations of these useful circuits, some familiar and some less so – a feast of useful circuits.

Regardless of the frequency band in which they are to be used, filters all fall into one or other of a limited number of categories. The main categories are low-pass, high-pass, band-pass, band-stop and all-pass; I have produced a CAD program to assist in the design of yet another category – the all-stop filter. I hope to see it published in the April issue.

This article is concerned with one class of active filters, the notch, and its cousin the notch-pass. I lump these together as one class, as their realisation is closely connected, both types commonly using variants on one or other of a few basic designs. These are the Wien bridge, the twin-tee and the bridged-tee circuits, plus sundry assorted others. So in what follows, the basic circuits are considered in that order. I show how a notch or a narrow selective pass-band – or even an oscillator – can be realised with each.

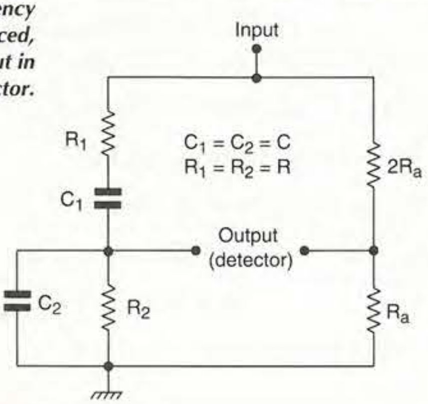


Fig. 1. The Wien bridge. At the frequency 1/(2πCR)Hz, attenuation through the frequency selective arm is a factor of three, the same as the aperiodic arm. At this frequency the bridge is balanced, producing no output in the detector.

Wien-bridge-based options

Figure 1 shows the basic Wien bridge. At the frequency $f_o = 1/(2\pi CR)$Hz, the attenuation through the frequency-selective arm is a factor of three, the same as the aperiodic arm. At this frequency, the bridge is balanced, producing no output in the detector, which might be, for example, an earpiece.

This was the original purpose of the Wien bridge, used in the days before frequency counters to measure the frequency of a signal – assuming it was vaguely sinusoidal. Either the two *C*s or the two *R*s must be varied as a pair to tune to the null. Then, knowing the value of *C* and *R*, the frequency is deduced from the above formula.

The frequency-selective arm can be analysed purely by mental arithmetic, or is it mental algebra? At the frequency where the reactance of *C* equals *R*, both the series and the parallel *CR* arms have a phase angle of 45° leading. Thus the voltage drop across the upper arm will be in phase with that across the lower, and so the voltage at their junction will be in phase with the applied voltage.

Further, the reactance of the upper arm will be $\sqrt{2}R$, while that of the lower is $R/\sqrt{2}$, or half that of the upper. Hence the attenuation at f_o must be given by $v_o/v_i = {}^1/_3$, with the output in phase with the input.

Clearly, if an instrumentation amplifier replaces the detector, you have a simple notch filter. But the shape or selectivity of the notch is fixed by this arrangement, regardless of the operating frequency. For example, if *R* is 100kΩ and *C* is 1nF in Fig. 1, the response at 1.59kHz – down to –∞dB in principle – is already showing appreciable attenuation well away from the notch. In fact, it will be 3dB down relative to the 'flat' response at dc and high frequency, at 480Hz and 5.3kHz, i.e. about 3.2 times higher and lower than the notch frequency, **Fig. 2**.

It is possible to arrange feedback around the circuit to sharpen up the notch. An example of the benefits of this appears in a total harmonic distortion meter design published earlier[1].

A selective amplifier... The Wien bridge also lends itself to a very convenient notch-pass or selective band-pass arrangement, **Fig. 3**.

Imagine that the input is short circuited, and that there is a signal at,

$$f_o = \frac{1}{2\pi CR}\text{Hz}$$

at the output. This is using the Wien network back to front, so the voltage at the op-amp's inverting input will be 2/3 of that at its output, and forms negative feedback.

Provided that the value of R_3 is less than $2{\times}R_4$, the positive feedback will be less than the negative, and the circuit will be stable. If R_3 is greater than $2R_4$, the positive feedback outweighs the negative, and the circuit will oscillate. The more nearly R_3 approaches $2R_4$ from below, the greater the gain at f_o. At dc, the gain is zero, due to the series *C*, and also at infinity, due to the shunt *C*.

Figure 4 shows the response for the case where R_3 equals 20k and R_4 equals 10.2kΩ. The peak gain is +37.5dB, and is 0db at 3.2 times higher or lower than f_o in Fig. 7. A 1% increase in R_4, or a 1% decrease in R_3, results in a 3.5dB decrease in gain. This makes close-tolerance high-stability resistors a must.

As always, when obtaining a high-Q response from a basically low-Q circuit by means of positive feedback, the result is very sensitive to component tolerances and stability.

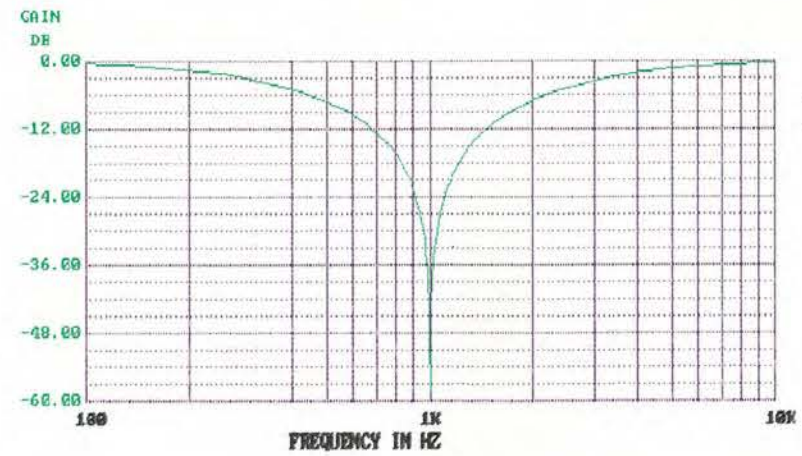


Fig. 2. Frequency response of the Wien bridge.

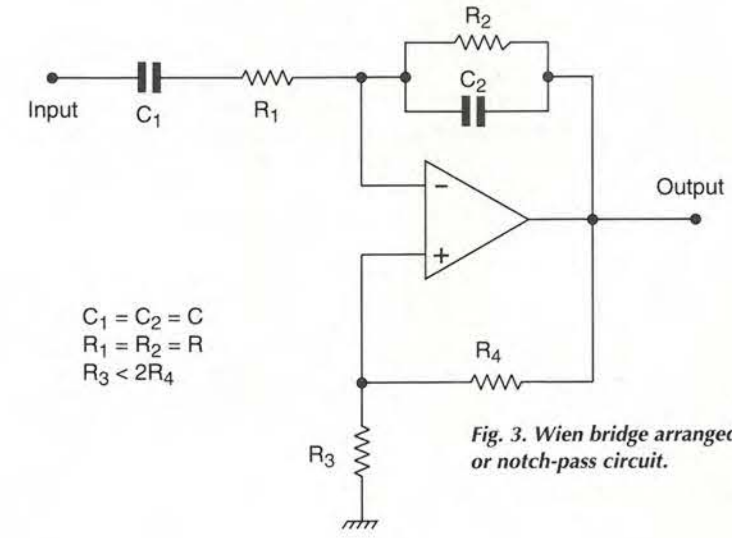


Fig. 3. Wien bridge arranged as a selective amplifier or notch-pass circuit.

...and an oscillator. If the series and shunt *CR* arms in Fig. 3 are interchanged, and the input short circuited, you have the basic conventional Wien bridge oscillator. But the Fig. 3 arrangement would also make a usable sine-wave oscillator, provided R_3 includes a resistor with a negative temperature coefficient, such as a thermistor. Alternatively, R_4 could have a positive temperature coefficient component, provided for example by a very low wattage filament lamp.

The Wien network has also been used in an ingenious tone control circuit[2].

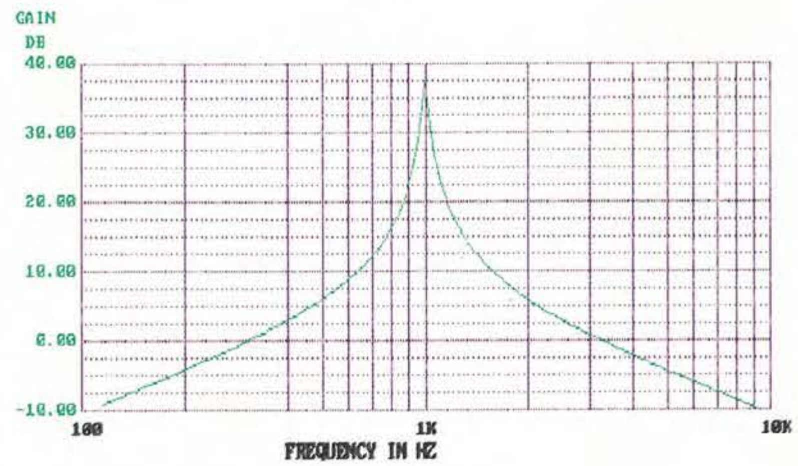


Fig. 4. Frequency response of the Wien bridge amplifier or notch-pass circuit, Fig. 3.

Twin-tee based circuits

The twin tee is not the most convenient of circuits to work with, on account of using three resistors and three capacitors to achieve its null. However, both its input and output ports are unbalanced.

The Wien bridge on the other hand has an unbalanced input, but needs a differential input amplifier at its output. This makes the twin tee handy in fixed-frequency applications. But the notch turns out to be even 'lazier' than that of the Wien bridge, being 3dB down on the zero and infinite frequency response at 4.25 times higher and lower than f_o, rather than 3.2 times.

Figure 5 shows the basic twin-tee circuit, and **Fig. 6** its circle diagram. A circle diagram shows the response at various points in a circuit, as the frequency varies from zero to infinity. Figure 6 shows the locus of the tips of the vectors representing the voltages at X and Y.

Superimposed on the circle diagram is the voltage vector diagram for the particular frequency f_o at which the notch occurs. The vectors showing the voltages across C_2 and R_2 at f_o have been offset a tad for clarity, as they overlie others.

As you can see, the response at B passes through zero (ground G) at f_o. The basic twin-tee circuit uses the component values shown, where again,

$$f_o = \frac{1}{2\pi CR}\text{Hz}$$

As with the Wien bridge, the circuit can be analysed mentally.

Imagine that unity gain buffer amplifiers are inserted at the points marked X and Y in Fig. 5, and that all the capacitors and resistors are equal. As the frequency varies from zero to infinity, the voltage at point X will rise from zero with phase angle 90° leading, to equal the input v_i at A, and in phase with it.

Likewise, the voltage at point Y will fall from v_i to zero, 90° lagging. At f_o, the voltage at point X will be 3dB down on the input, leading by 45°, and at Y will also be 3dB down, but lagging by 45°. The voltage between X and Y is applied by the buffers to the series C and R, and forms the base of another (semi)circle diagram. Thus at f_o, the output voltage is zero.

Without the buffers, assuming that the input is driven by a zero impedance source – effectively ground – the impedance seen looking back into point Y at f_o is not R but $R/2$. This is because at the notch frequency, the output is also effectively ground.

So C_3 must be $2\times C$ to give the required 45° phase at Y. Likewise looking in at point X, there are two capacitors of value C in parallel, so the resistor from there to ground must be $R/2$. The voltage vector diagram of Fig. 6 is then the same as for the case of equal Cs and Rs with buffers; only the current vector diagram, which is not shown, differs.

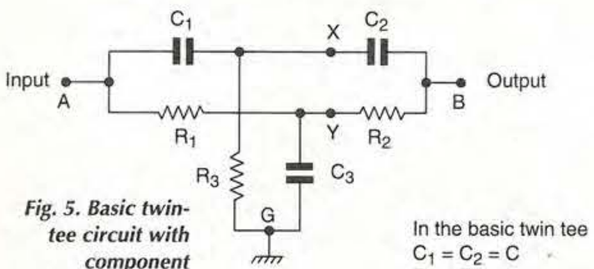


Fig. 5. Basic twin-tee circuit with component values. The text also considers the case where unity gain buffer amplifiers are inserted at X and Y, with all equal component values.

A sharper notch... Loading of the output CR on the lead and lag input sections is responsible for the lazy shape of the notch. With the equal components plus buffers case mentioned above, the –3dB point is only 2.4 times above and below the notch frequency, as against 3.2 times for the Wien bridge.

However, the basic twin-tee circuit less buffers can be sharpened up even more, by the judicious application of positive feedback, as in **Fig. 7**.

To see how this works, consider the input frequency rising from 0Hz, up towards the notch frequency. If the $R/2$ and $2C$ arm were grounded, the response would fall as with the circuit of Fig. 5. But in fact, it is bootstrapped very nearly up to the same voltage as that at B in Fig. 5. Consequently, the attenuation due to the twin-tee network is greatly reduced at frequencies approaching f_o.

But at exactly f_o, this argument must break down, so a zero response will still in fact be observed. With the values shown, the notch is sharpened up considerably, **Fig. 8**. A less extreme notch results from tapping the feedback point further down the op-amp's output. In this case the feedback voltage may need buffering, to keep its impedance low.

The extra selectivity is bought by the positive feedback, and at a price. As always with positive feedback, distortion, noise and variation of gain with component tolerances and ageing, are all increased.

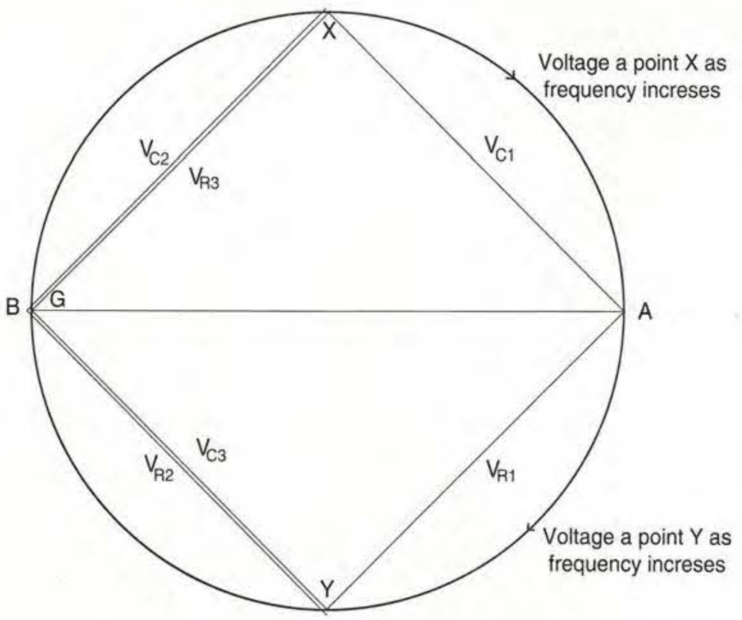


Fig. 6. Circle diagram for the twin-tee network, with superimposed, the voltage vector diagram for the frequency where the notch occurs.

Chebishev response

The twin-tee circuit is capable of some other useful tricks[3].

For example, if capacitor C_1 is connected not to the input, but tapped down a potentiometer chain at a fraction k, a notch is still observed, but while the output is still equal to the input v_i at 0Hz, it is only $k\times v_i$ at infinite frequency. Thus the circuit provides a second-order Chebishev (or Tchebyshev) response. As k approaches zero, the notch moves out to infinite frequency, leaving a low-pass response.

Further, let the fraction of the output fed back to the shunt $R/2$, $2C$ arm, as in Fig. 7, be m. Then with the 22Ω resistor changed to 30Ω (m=0.97), and k=0.5, the response becomes as in **Fig. 9** – a second-order elliptic response.

Such sections can be used to build up an elliptic filter of any desired order. A search through my files failed to unearth a copy of reference 3, but I assume it gives the relevant design equations.

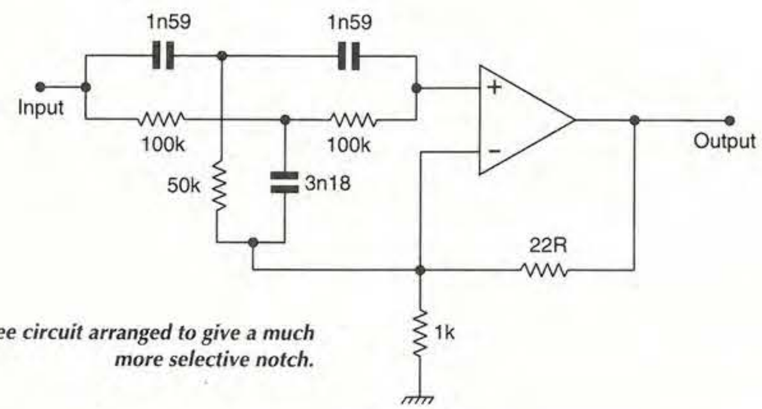


Fig. 7. Twin-tee circuit arranged to give a much more selective notch.

...and an oscillator. The notch provided by the twin-tee filter only extends down to minus infinity decibels if the component values are suitable. But this does not mean that they must be exactly R, R, $R/2$, C, C, and $2C$.

If these component values are all roughly right, then tweaking the value of any two, obviously easier for Rs than Cs, will trim it up to perfection. Otherwise, the locus of point B may fail to pass through 0V.

If it misses, there's an equal chance of it being to the left or to the right of the 0V origin G in Fig. 6. This means that there may be a small output in antiphase with the input at the notional notch frequency. This was exploited in reference 4 to make a twin-tee-based oscillator where the maintaining amplifier was an emitter follower.

So how can that work? **Figure 10a)** shows a twin tee where the phase lead at X exceeds 45° at the same time as the lag at Y. The vectors shown are for the nominal f_o case, and show a small antiphase output. The dotted circle shows the locus of the tip of the output vector as the frequency varies from zero hertz to infinity.

Note that if the same current passes through a resistor and capacitor in series, the voltages across the two components must be in quadrature. Hence angle YBX in Fig. 10a) is 90°, as also are angles GXA and GYA, since any angle inscribed in a semicircle is a right angle.

If now such a modified twin tee is connected as in Figure **10b)**, an oscillator results. It looks odd, having the twin-tee's input connected to ground, but look at it this way. Relative to the *emitter*, assume a large antiphase input at A. Then, relative to the emitter again, there will be a small in-phase signal at B.

So relative to ground, the input at the base is marginally larger than the output at the emitter, by a few percent. Provided that the gain of the emitter follower is within a few percent of unity, oscillation results.

According to reference 4, if the modification a to the R and C values in the

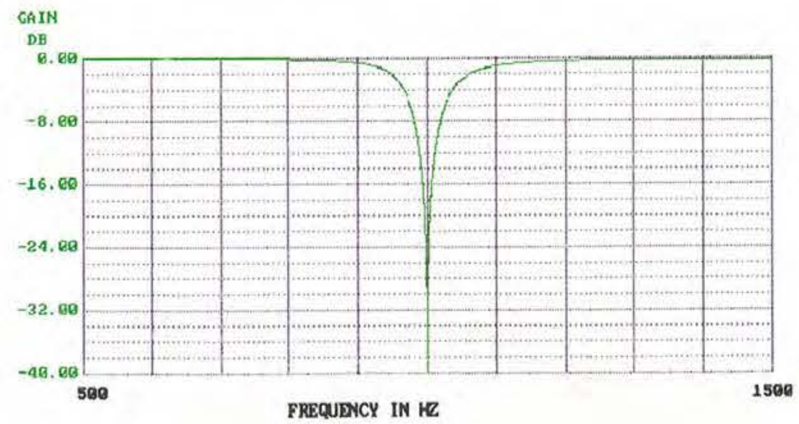


Fig. 8. Frequency response of the twin-tee selective notch, Fig. 7.

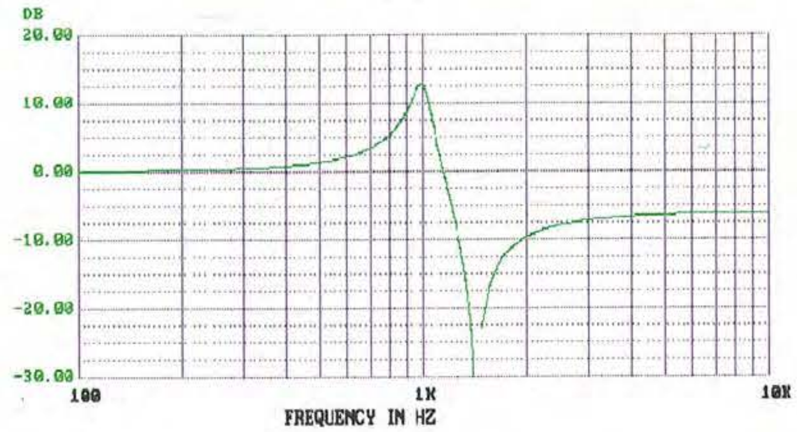


Fig. 9. Second-order elliptic response of the selective notch circuit in Fig. 7 but with values modified as described in the text.

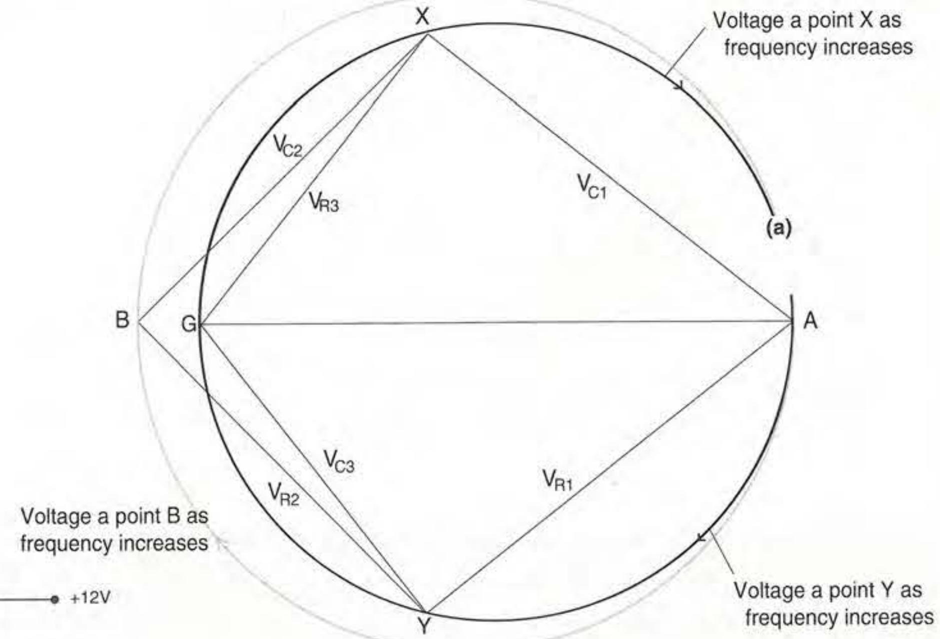
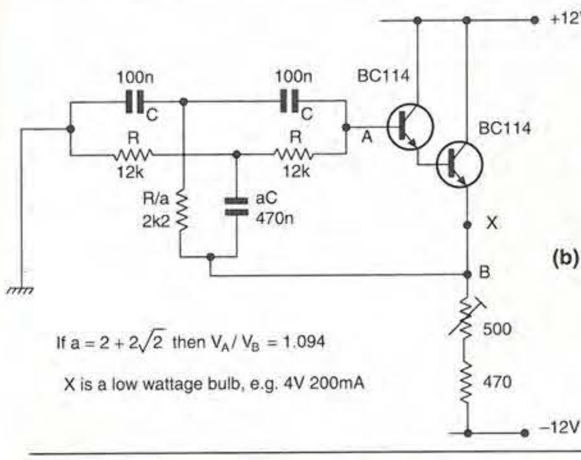


Fig. 10a). Showing how a twin-tee circuit can provide a small phase-inverted output.



(b) *b) Emitter follower oscillator using a twin tee.*

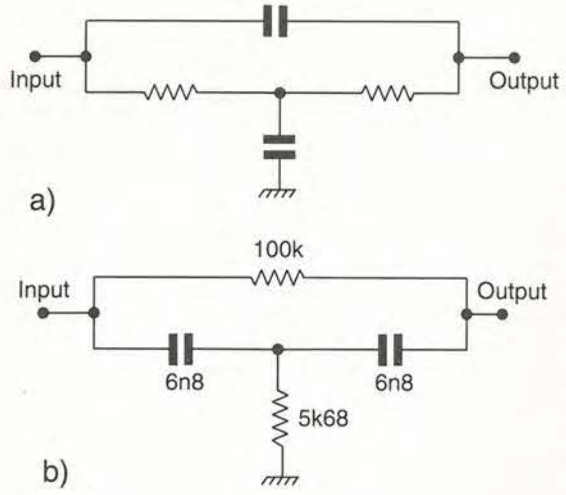


Fig. 11. Bridged tee circuits may have either a C or R shunt leg. a) capacitive shunt leg. b) resistive shunt leg.

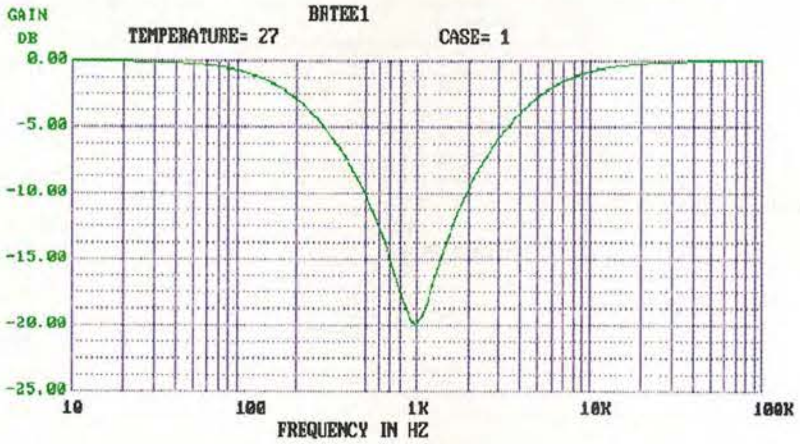


Fig. 12. Response of the bridged tee circuit with resistive shunt leg, Fig. 11b).

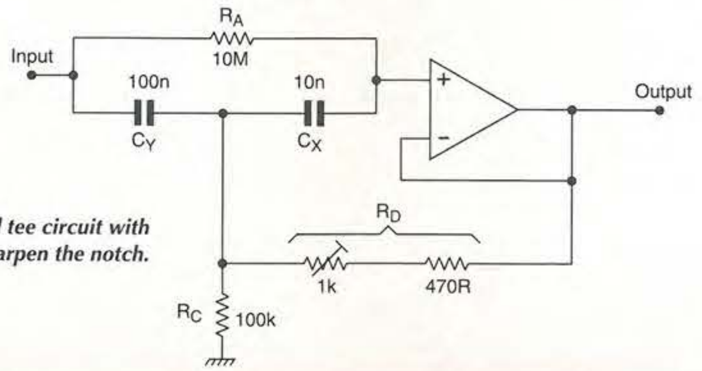


Fig. 13. A bridged tee circuit with feedback to sharpen the notch.

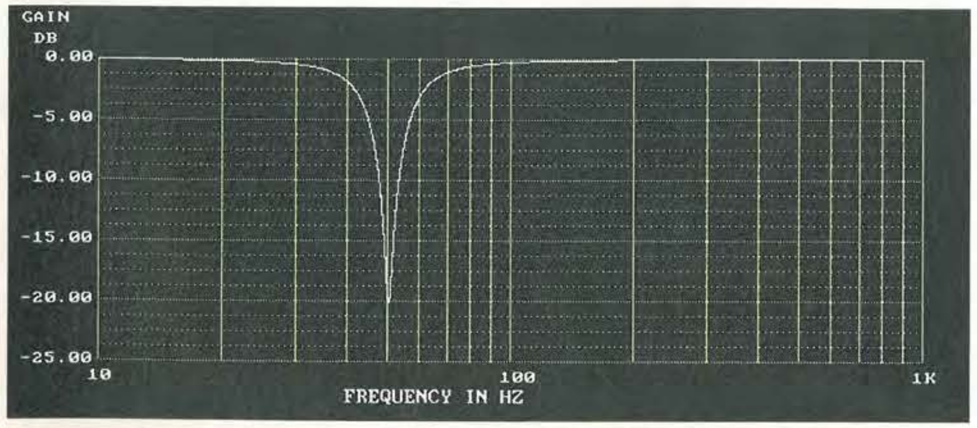


Fig. 14. Circuit of Fig. 13 gives a 20dB notch at 50Hz, but is flat within 3dB from 60Hz up.

shunt leg is $2+2\sqrt{2}$, then the voltage at B exceeds that at G by a factor 1.094. A small low-wattage bulb inserted in the emitter lead at Z then stabilises the amplitude of oscillation, giving a low distortion sine-wave output.

Note that the lamp carries the dc emitter current, adjustment of which 'preloads' the bulb with dissipation. As a result, the small additional loading due to the twin tee stabilises the amplitude.

The gain of a non-inverting op-amp buffer with 100% negative feedback is very close to unity indeed, so such a circuit would only need a value of a marginally in excess of 2.

Bridged-tee based circuits

Whereas the twin tee uses three capacitors and three resistors, the bridged tee uses only two of each. The bridge across the tee may be either a capacitor or a resistor, as shown in **Fig. 11**.

The economy in components of the bridged tee comes at a price; the notch depth can never be infinite. However, where something less is acceptable, the circuit is definitely useful.

For example, with the values shown in Figure **11b)**, a 20dB notch depth is achieved. Although the capacitor values are equal, a useful notch depth requires considerable asymmetry in the resistor values, as shown. The resultant frequency response is as in **Fig. 12**.

A handy hum rejection filter... As with the twin tee, feedback can be used to sharpen up the notch. A typical circuit, taken from reference 5, is shown in **Fig. 13**.

Note that in this design, both the resistor and capacitor values are asymmetrical. This, in conjunction with the feedback, results in a 20dB notch at 50Hz, flat to within 3dB at 60Hz and upwards, **Fig. 14**.

Usefully, the notch frequency may be trimmed by varying R_D, without affecting either the notch depth or bandwidth. Reference 5 gives the full design equations for this circuit, but obviously you can change the frequency simply by scaling the capacitors.

...and easily tuned selective amplifier. Here is a handy circuit that was published years ago[6] by analogue design guru Bob Pease while he was still Robert Pease of Teledyne Philbrick. Not that it was invented by him, but as he said, it is too good not to pass on.

Shown in **Fig. 15**, the circuit has several nice features. The tuned frequency is adjusted by means of R_2. Furthermore, this does not affect either the centre frequency gain or the bandwidth.

Centre frequency gain is given by,

$$\frac{R_3}{2R_1}$$

and bandwidth by,

$$\frac{1}{\pi C_o R_3}$$

The centre frequency is,

$$f_o = \frac{1}{2\pi C_o \sqrt{R_1 \| R_2 \times R_3}}$$

Other notch circuits

Many other notch and notch-pass circuits exist; one could fill pages with them. But for the present, I will show just a couple more.

The first needs a low, ideally zero, source impedance and must feed into a high, ideally open-circuit load. Further, the notch it provides is not sharpened up by feedback. Despite this, it is of interest because it is canonical, unlike the twin-tee notch.

A notch circuit is necessarily a two-pole or second-order arrangement, and canonical in this context means using only one CR time constant per pole. Depending on how they are arranged, the two CRs can provide either a low-pass or high-pass response with a 12dB per octave roll-off in the stop-band, a band-pass response with 6dB per octave roll-off either side, or a notch.

The notch means that there is a zero on the $j\omega$ axis, and the response either side can be sharpened up by bringing the associated poles, at the same value of ω closer to the $j\omega$ axis.

The canonical notch in question appeared in reference 7 and the circuit is shown in **Fig. 16**. Its response is 4.5dB down at one octave above and below the notch and still a decibel down at 200Hz.

The easiest way to understand how the circuit works is by drawing a circle diagram. If you care to try, start off by drawing the semicircle representing the leading voltage at the op-amp's non-inverting input. The locus of the tip of the vector representing the op-amp's output can then be drawn in, remembering that it must be such that the voltage at the inverting input must always equal that at the non-inverting.

Finally, the voltage applied to the R_1 C_1 arm is the difference between the input voltage and that at the op-amp's output terminal. You will find that the voltage at the junction of R_1 and C_1 does indeed pass through zero.

My final offering is another notch, from reference 8. This uses a simulated lossy inductor, realised with an op-amp, capacitor C_2 and a potentiometer, as in **Fig. 17**. The potentiometer is adjusted so that the inductor resonates with C_1 at the desired notch frequency f_o.

If the residual resistance at resonance is R, the inputs of the other op-amp are connected to a balanced bridge, so its common-mode rejection of A_1 means that there is no output. The notch frequency is,

$$f_o = \frac{1}{2\pi\sqrt{R_1 R_2 C_1 C_2}}$$

and the bandwidth,

$$bw = \frac{\sqrt{R_1 R_2 \frac{C_2}{C_1}}}{R_t}$$

where the resistance of the potentiometer $R_t = R_1 + R_2$.

For a ratio of C_2/C_1 of 250, a 50Hz notch will be flat to within –3dB below 45Hz and above 56Hz. Be aware that with a high Q, the lower op-amp may saturate on large signals, due to the internal magnification around this circuit. For more information, see reference 9. ■

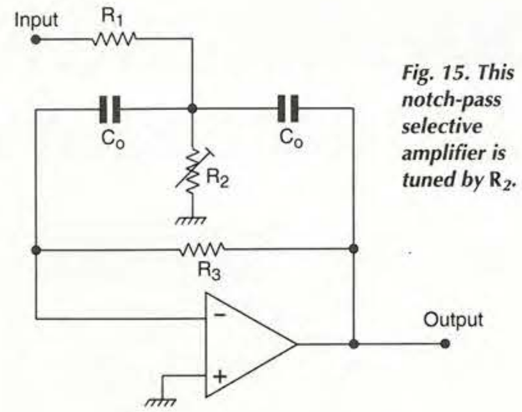


Fig. 15. This notch-pass selective amplifier is tuned by R_2.

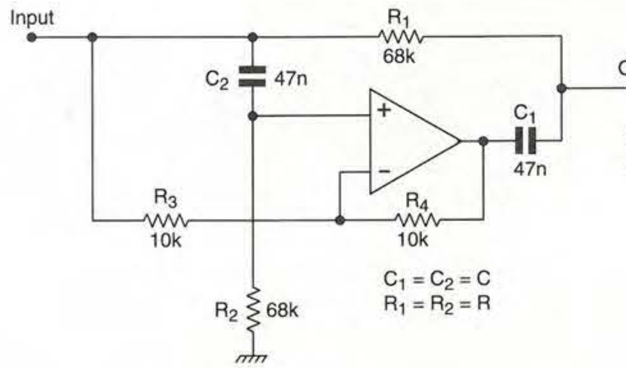


Fig. 16. A canonical notch circuit. Notch frequency $f_o = 1/(2\pi CR)$.

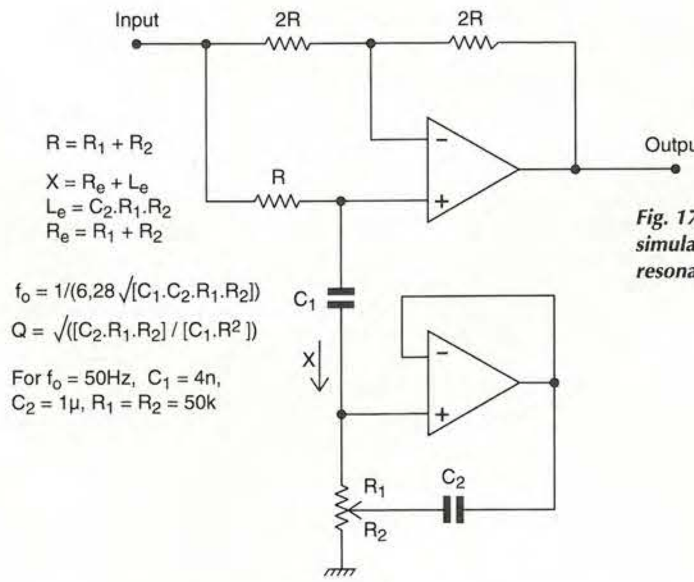


Fig. 17. Notch circuit using a simulated inductor, which resonates with C_1.

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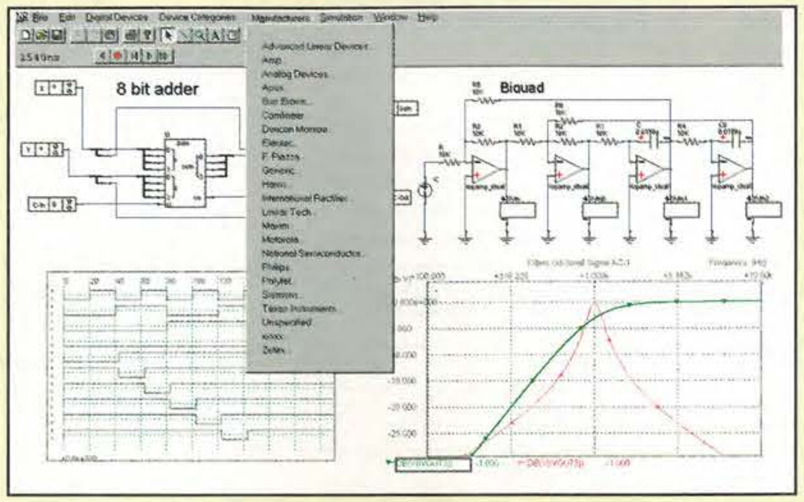

On 1 January 2000, following two years of development, RD Research is launching a significant new CAD package. Called B2Spice 2000, this new SPICE simulation software is being offered to *Electronics World*'s UK readers with a unique 25% introductory discount until 1 February 2000. Overseas readers can take advantage of this offer until 1 March 2000.

B² Spice 2000 combines the benefits of two widely used circuit simulation programs – B² Spice and B² Logic – to provide true mixed-mode simulation in one new and enhanced package.

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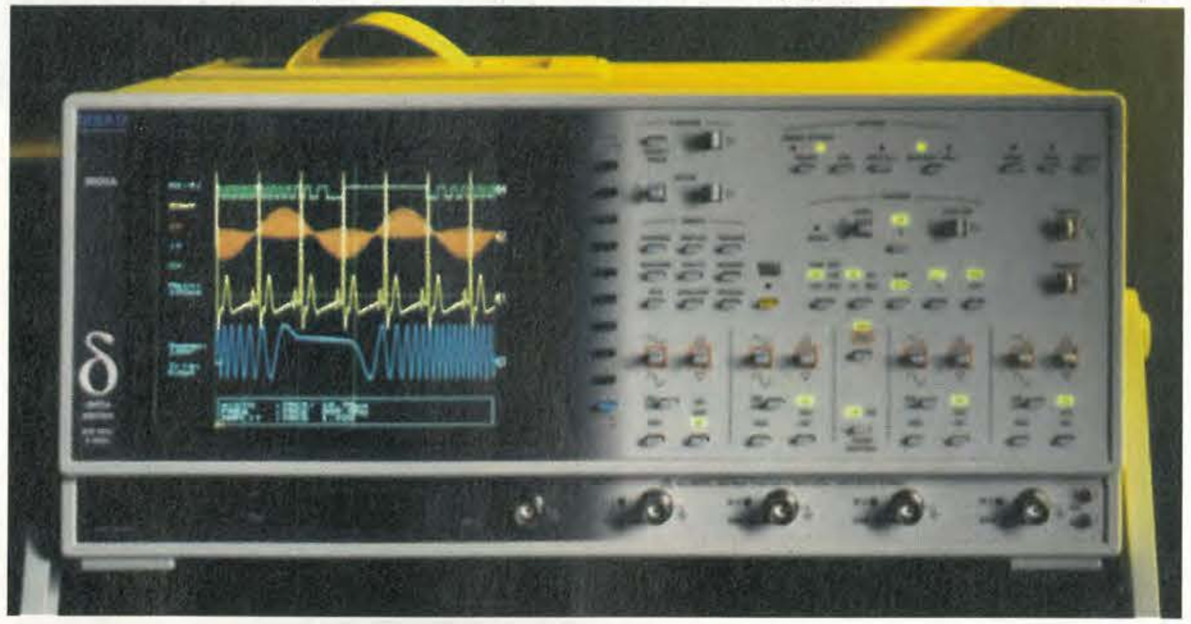


Get the most from your scope

In this set of articles, Les Green explains what types of oscilloscope are available, and looks at how to apply them for best results. His first article covers oscilloscope basics, benefits and limitations.

It is not possible to see electricity so we have to rely on instrumentation. This makes it necessary to know about the limitations and accuracy of every piece of test gear we use, in order to get answers that can be believed.

This article will help you pick the right type of oscilloscope for your application, and to get the best measurements possible using a few tricks of the trade.

What's in an alias?

Technically, an alias occurs in a sampling system when there is a significant amount of signal at a frequency of more than twice the sampling rate.

Viewing a sine wave with a totally inadequate sampling rate can create a signal that looks quite normal, except that it appears to be at a frequency much lower than its actual value. Instead of an assumed name it has an assumed frequency. This only really causes a problem when the situation goes unnoticed.

Note that the use of stroboscopic lights to slow down or freeze motion is a deliberate application of this aliasing phenomenon.

Where a pulse's width is narrower than the sample period, the sample may or may not occur during the pulse. If the samples are not synchronised with the pulse then the pulse will only be visible for some of the time; this is another example of aliasing.

Types of oscilloscope

There are three basic types of oscilloscope; the real-time scope, the storage scope and the sampling scope.

By far the cheapest and most common is the real-time scope, sometime called an analogue scope. They have also been referred to as real-time oscilloscopes, or RTOs. I doubt that there are any engineers around who do not know roughly how a real-time oscilloscope works. Similarly, the concept of a storage scope is not hard to grasp. However, I would imagine that only a very few engineers have even heard of a sampling scope.

While real-time scopes get very difficult to make above the 400-500MHz region, sampling scopes above 20GHz are common. Unfortunately they only work on repetitive signals, preferably with repetition rates well above 1MHz, and they have 50Ω inputs – which are destroyed by voltages much greater than 5V rms.

Typically a diode sampling bridge is opened for a very short time once every 10μs. The input waveform is then reconstructed from the sampled waveform – which is a deliberate alias of the input waveform – and can be displayed on a conventional scope tube without any form of storage.

Older models are typically difficult to use as the trigger control needs to be set very carefully to get any sort of picture at all. It is all too easy to get a screen full of disconnected dots – which is not very useful. However, if you want to look at signals well above 1GHz or rise-times less than 300ps you do not have much choice.

Storage scopes come in two basic types; tube-storage, i.e. analogue, and digital storage. A tube-storage scope uses a special type of display tube that stores the trace on an inner surface structure. They are difficult to use and will not store a trace for a long time. This technology is long dead but I have no doubt that some such scopes are still in use because they are not actually broken.

Size matters

You can write this in bold letters, "Equipment that is difficult to use, or too heavy to move about, will not be used as much as it should be". It is often the case that engineers are pushed for time; tests that are difficult to do, on equipment that is wretched to use, will tend to get reduced to a minimum or skipped entirely.

Such test equipment is a liability, not an asset, and would be better off in the bin. If you love your old tube-storage scope I apologise unreservedly.

In addition to the basic types mentioned, there are at least two other hybrid types available. Sampling scopes are now available in a digital storage format. They are easier to use than the older type but should not be confused with digital storage oscilloscopes, or DSOs. You have to read the fine print to see that their single-shot sampling speed is low – typically less than 1MS/s – so they are not useful as a general purpose DSO replacement.

The other hybrid type is the combined real-time and digital storage scope. These tend to be neither good real-time scopes nor good DSOs but they do have a place in general purpose applications.

So we now have sampling scopes for use above 1GHz, with digital versions being necessary for repetition rates below about 10kHz. This is the last time that I will be mentioning them. For the rest of the measurements I will only compare real-time scopes and DSOs.

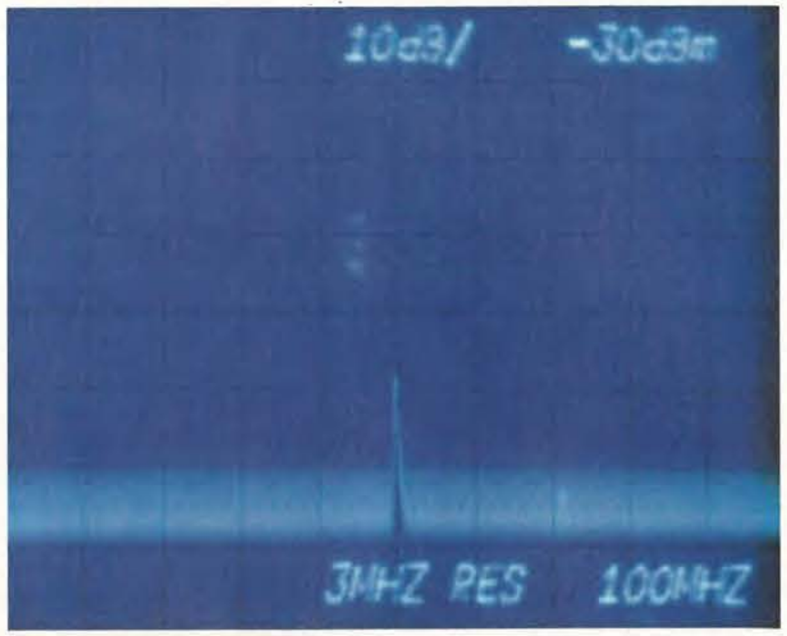


Fig. 1. An output stage oscillation on a faulty DC-300MHz amplifier, measured using a conventional 10:1 probe on a Tektronix 7603 frame with a 7L12 spectrum analyser plug-in. Top of the screen is –30dBm and the frequency of the oscillation is 1170MHz. This oscillation could not be seen on a 400MHz Tektronix 2465B real-time scope.

Types of measurement

You must use the right piece of equipment to measure your circuit's performance. An AVO 8 is a perfectly good piece of equipment. For any alien life forms reading this article, an AVO 8 is a moving coil multimeter characterised by being in a solid, heavy, black case that can not be pulled off the work bench by 15A test leads and having a working life at least equal to that of the average engineer.

But if you try to measure a 100MHz voltage with an AVO, you will not get a sensible answer. If you then blame the meter for giving you the wrong answer, you are only demonstrating your complete and utter ignorance of the subject.

Just because you have a scope probe in your hand though, it does not mean that you should measure everything with it. Scopes are good at looking at changing voltages; if you use one to measure steady direct voltages, then the answers you get will not necessarily be very accurate.

Scope accuracies are generally around the 1% to 3% range. If you then add another 1% for the 10:1 probe you are using, and another 1% for the input resistance tolerance of your scope, you get a fairly inaccurate answer. If that is all the accuracy you need then there is no problem. However, even the most modest of bench DMMs will give you a far more accurate answer.

Probe adjustment

While we are on the subject of 10:1 probe accuracy there is another point that is sometimes overlooked – probe adjustment. For rectangular waveforms above 1kHz, the pulse response is totally dependent on the adjustment of the probe. The 'probe calibrator' output on the front of an oscilloscope facilitates this.

Each probe has to be set up for each oscilloscope *and* for each channel. There is no guarantee that the inputs of any particular scope are matched to any degree at all. If you swap a probe from one channel to another it is vital that you re-tweak the probe to get the correct pulse response.

Calibration is generally done at 1kHz. Beware of probe calibrators that change frequency with the timebase. Although they have a definite advantage for experienced users in that the VHF performance of the probes can be optimised (there are often hidden trimming points within 10:1 probes) they are a definite liability for the general user.

The problem is that if the timebase is set too high the probe will give a square edge response regardless of the trimmer position. Thus the unwary user can think that the probe is trimmed correctly when it is not.

It is common practice to probe a circuit with an oscilloscope, checking for oscillations and noise, before trying to get an accurate reading on a DMM. Note that a DMM set to read DC volts will measure a 5V DC level with 1V of 1MHz sinusoidal oscillation on it and tell you the answer is 5V, maybe.

It is also possible for a DMM to give an incorrect reading due to internal asymmetrical slew-rate limiting. The point is that the result is no longer defined and it's not the DMM's fault, but yours!

I should also mention that putting a DMM on some circuits causes them to oscillate. Some DMMs have large input capacitances – say 100pF – which can cause problems.

Test leads can also capacitively couple to/from sensitive nodes causing oscillation. To fix this, all you have to do is to put a resistor in series with one or both test leads; 100Ω to 1kΩ is usually enough, but it needs to be at the probe tip, or as close to it as possible, for maximum effect. Alternatively, keep your scope connected to the circuit when you probe with your DMM; this at least warns you of any circuit malfunction that may occur.

Without wishing to stray from scopes too far, I must point out that it is not just DMMs that cause oscillations. Even a 10:1 scope probe with 10MΩ input resistance and 15pF input capacitance can cause or stop an oscillation. Unfortunately these oscillations can be well out of band for the equipment being used. Low-level gigahertz-rate oscillations just will not show up on an oscilloscope.

A faulty DC-300MHz preamplifier board caused the oscillation shown in **Fig. 1**. When connected to a 400MHz Tektronix 2465B real-time scope – arguably the best real-time scope in the world – this oscillation was not visible. In fact gigahertz oscillations often cause mysterious DC offsets that just don't make any sense and which change as you probe around the circuit.

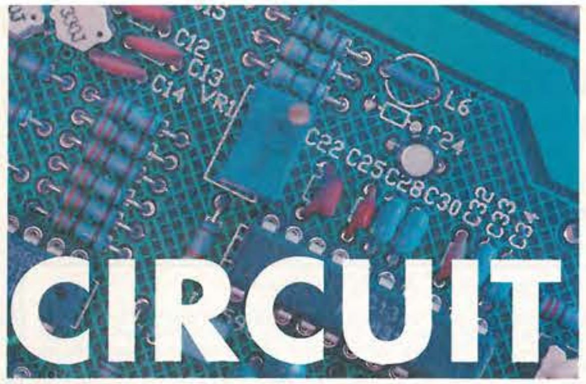
When a spectrum analyser is best

For anyone working with transistors having an f_t of 2GHz or more, a final check on a piece of equipment would have to include a quick scan with a spectrum analyser.

Even the fastest scopes will not detect the peaking of a noise band that is indicative of a circuit being on the edge of stability. One could easily make a case for insisting that such testing be done on prototypes, as it does not cost much to hire a simple spectrum analyser for a few days.

It is easy to use a spectrum analyser for such a task. All that is needed is a conventional 10:1 probe and probably a BNC to type-N adapter. The spectrum analyser will almost certainly have a 50Ω type-N input but this does not matter; hf signals will easily pass through the probe and register on the spectrum analyser. The presence of an oscillation, or latent oscillation, is more important than its actual amplitude. ■

In his next article, Les looks at the problems involved in making specific types of measurement with an oscilloscope.



CIRCUIT IDEAS

Winner

Richard's spectrum analyser extender wins the first National Instruments digital multimeter worth over £500.

10MHz spectrum analyser monitors 550MHz signal

Needing to monitor the output of a 500-600MHz amplifier and only having a 10MHz spectrum analyser brought to mind an article on the use of gates as oscillators.* The design works reliably, but it is not without its limitations and could easily be improved.

A *74HC04* inverter forms the oscillator, whose frequency is determined by the supply voltage, varying between 80MHz and 250MHz for a supply variation of 2-5.5V. Third harmonic is 750MHz, the fundamental at 16dBm being filtered out.

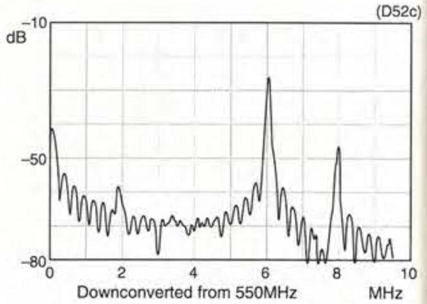
The mixer is based on the National *LMX2216*, which contains a low-noise amplifier, not used here. Output from the mixer goes through a simple band-pass filter to reduce aliasing on the spectrum analyser. Harmonics and sum-and-difference frequencies from the oscillator mean that one must take care in measurement to observe the correct signal.

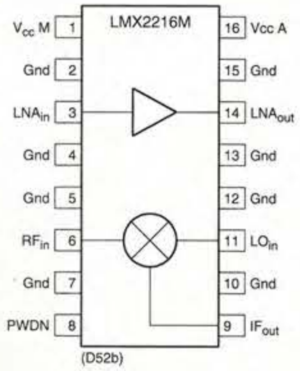
Mixer and oscillator should be mounted on copper-clad board with all ground connections soldered to the copper. Any tracks should be kept few and short (a 1in wire represents 25nH). Active devices should be decoupled on supply pins. Small capacitors are s-m COG types and rf connections should be made via 50Ω connectors.

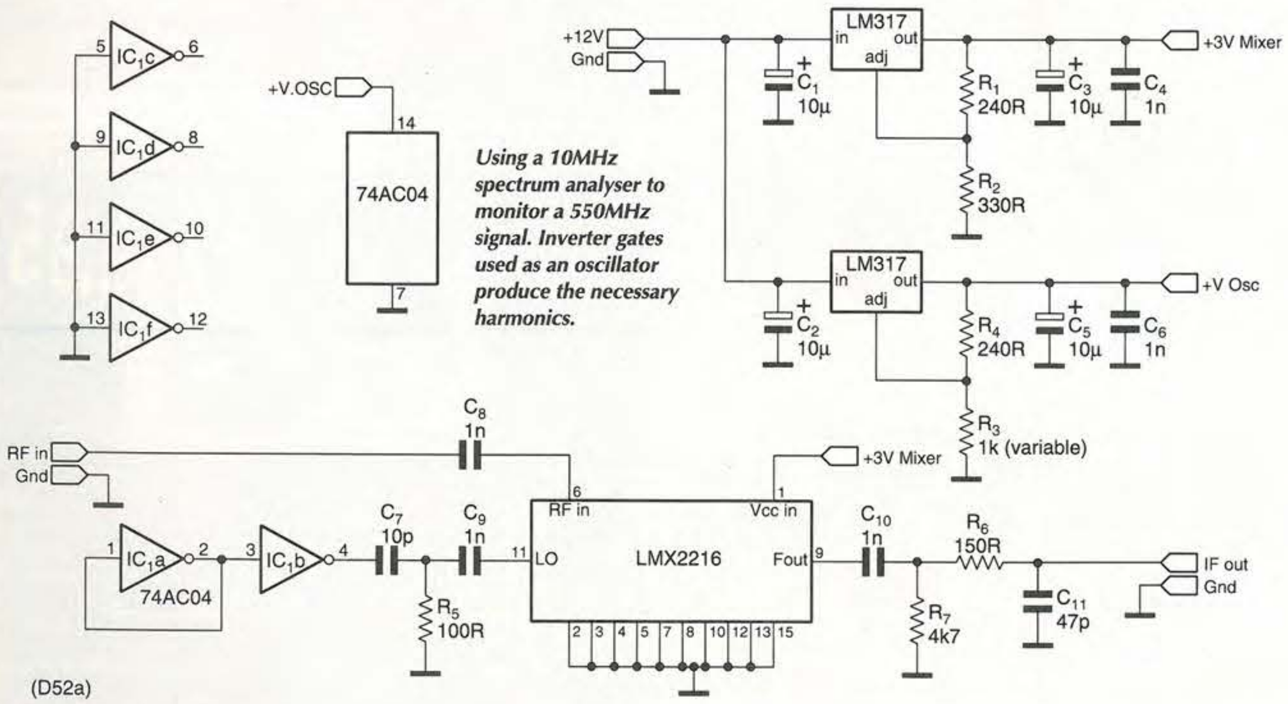
Richard Jacklin
Midhurst
West Sussex
D52

*Forster, I, 'When is a gate not a gate?', *Electronics World*, December 1996, p.956.

V_{supply}	Frequency	V_{supply}	Frequency
2	80	4	209
2.5	121	4.5	225
3	157	5	239
3.5	186	5.5	5.5







Using a 10MHz spectrum analyser to monitor a 550MHz signal. Inverter gates used as an oscillator produce the necessary harmonics.

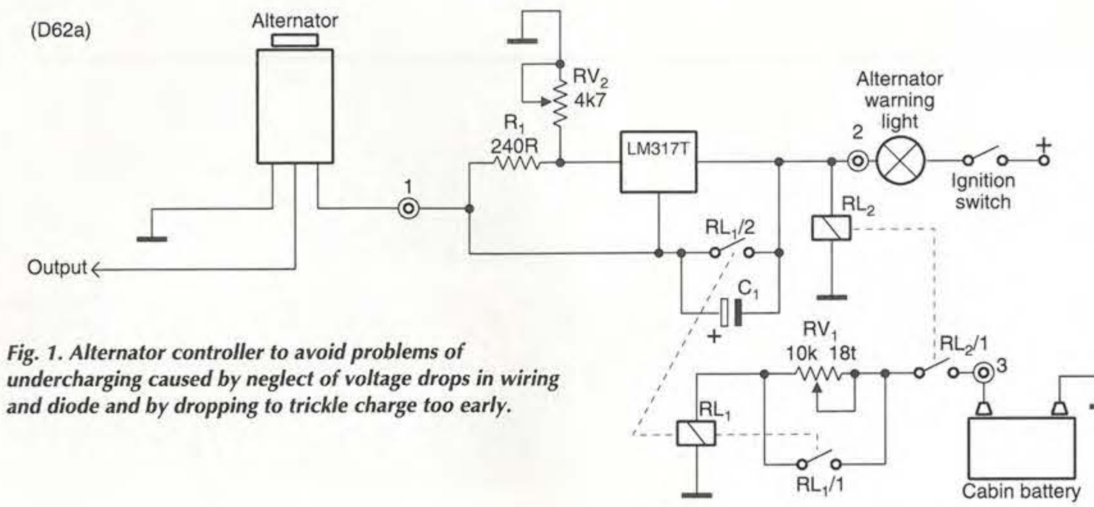


Fig. 1. Alternator controller to avoid problems of undercharging caused by neglect of voltage drops in wiring and diode and by dropping to trickle charge too early.

Involved with boats or battery charging?

Alternator controllers supplied with the alternator are inefficient at charging discharged batteries deeply because, since battery voltage is measured at the alternator, the measurement does not take account of wiring or blocking-diode voltage drops.

Further, the regulator normally starts to trickle charge too soon; at about 13.5V or 14V instead of the correct 16V (the voltage across a fully charged lead-acid battery). Add-on controllers to avoid these problems are expensive and often charge at too high a current.

To cater for the undercharging, the scheme shown in **Fig. 1** is effective, in that the regulator is "tricked" into detecting a battery voltage 2V lower than it is by the use of an LM317T regulator, set up so that the output is always lower than the input. The LM317T is switched by a relay to prevent overcharging, the potentiometer being set to make the relay RL_1 operate at 16V, switching out the LM317T with one contact and latching RL_1 on. At switch off, the relays drop out and are ready for normal operation at the next switch-on.

Figures 2 and **4** illustrate the method of use, Fig. 2 showing charging with a built-in regulator and Fig.4 indicating the use of a separate one. **Figure 3** shows the connections. On the alternator, remove the wire connected to terminal F and connect the green wire from the controller, connecting blue from the controller to the wire just removed. (A separate regulator has red and black.)

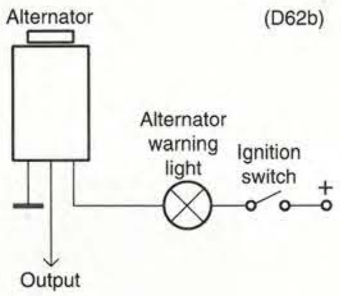
Connect green from the controller to alternator positive, removing the original wire and connecting blue to it instead.

At switch-on, the alternator light comes on and the start-up light goes out, while the battery voltage rises slowly to 16V and then decreases to 13.8. Sealed batteries need more care, so set the maximum to 15.5V.

The *LM317T* needs a heat sink.

Andrew Bird
Burntwood
D62

Fig. 2. Charging system with built-in regulator.



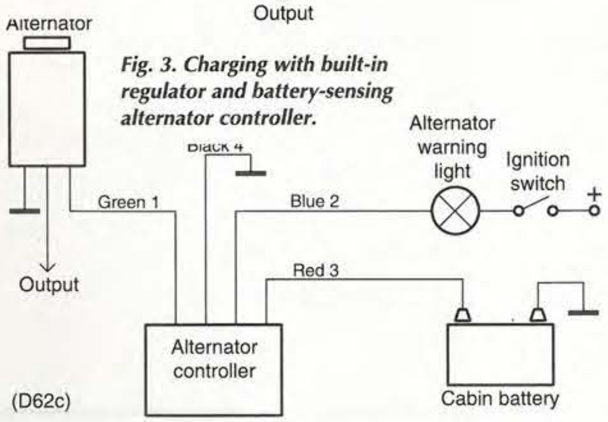
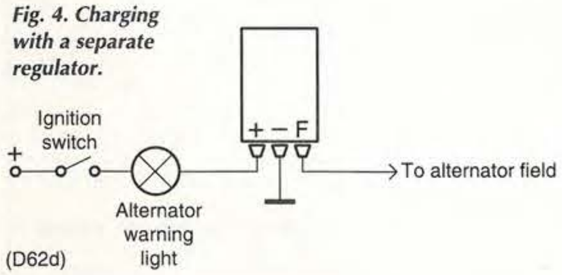


Fig. 3. Charging with built-in regulator and battery-sensing alternator controller.

Fig. 4. Charging with a separate regulator.



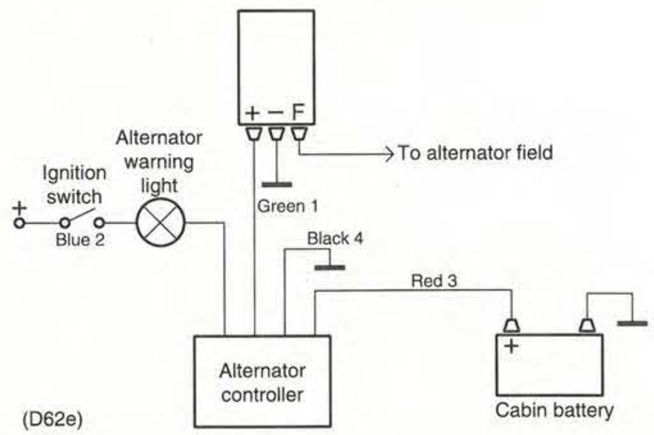


Fig. 5. Separate regulator with battery-sensing controller.

Self-powered amplifier/squarer

Sinusoidal inputs to this circuit are amplified and squared without the use of an additional power source.

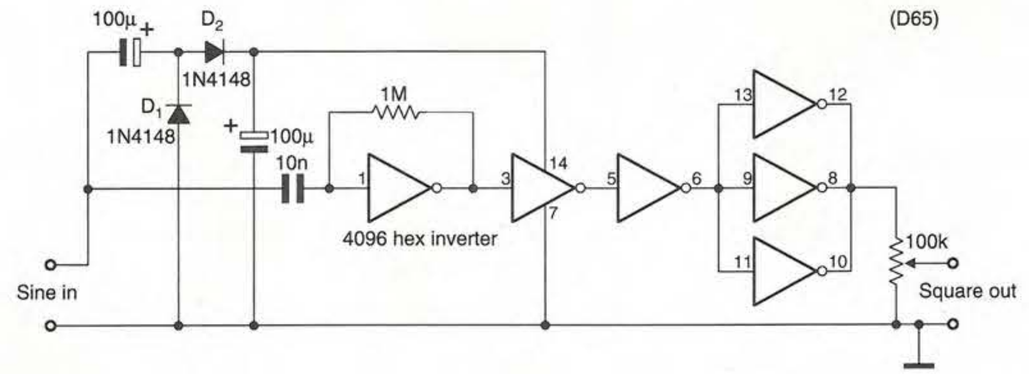
Inputs in the range 20Hz-20kHz drive a voltage doubler rectifier to supply the power and the first stage of the *4069* hex inverter functions as a linear amplifier, succeeding gates taking care of the squaring. Output with an input of 1-2.5Vrms produces a 50:50 m:s ratio.

Minimum input voltage for reliable working is 750mV, although using germanium diodes lowers the minimum.

Flavio Dellepiane
Genova
Italy
D65

£50 winner

This amplifier and squarer needs no external power rail, taking its power from the input signal.



Two identical crystals – what's the difference?

This produces a rectangular wave output at the difference frequency of two crystal oscillators, needing no counter/dividers and using only two integrated circuits.

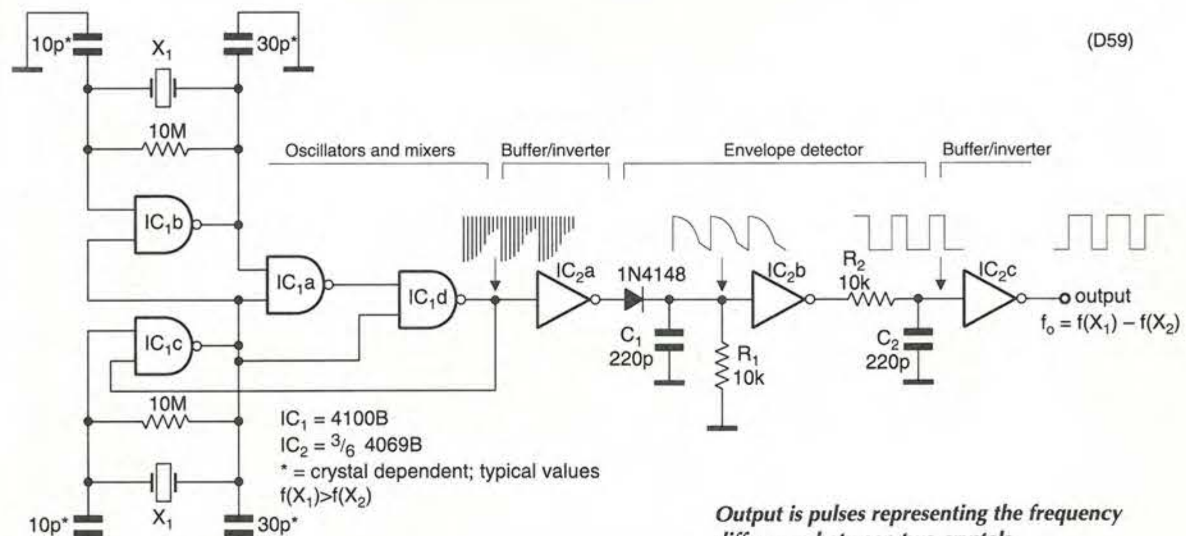
The nand gate, IC_1 forms both two crystal oscillators and a mixer, the output of the mixer being at the frequency f_1-f_2, the difference between the two oscillators.

After buffering, this signal goes to the envelope detector, which suppresses all hf components to give the output pulse at the difference frequency.

With components shown, the detector bandwidth is up to 100kHz with a crystal frequency of 3MHz. ■

Pekka Vähäkangas
Lakiala
Finland
D59

£50 winner



Output is pulses representing the frequency difference between two crystals.

NEW PRODUCTS

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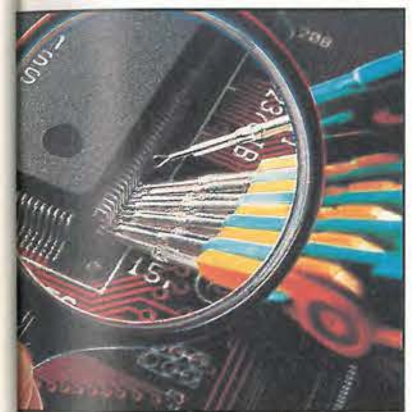
Test clips

Micro SMD grabber test clips for fine-pitch IC packages are available from Warwick Test Supplies. Made by Pomona, the clips are for prototype design and debugging. Three grabber styles are available – short and long tip for 0.8 to 0.5mm lead pitches and one for lead pitches down to 0.3mm. They can be used with QFP, PQFP, SSOP, TSOP and TSSOP packages.



The body design allows for side-by-side stacking for adjacent lead probing, the pincer tips being Teflon insulated to guard against shorting. All styles can be used at up to 100MHz and are supplied with flying leads for logic analyser attachment.
Warwick Test Supplies
Tel: 01926 851007
Enquiry No 502

P-channel MOSFET

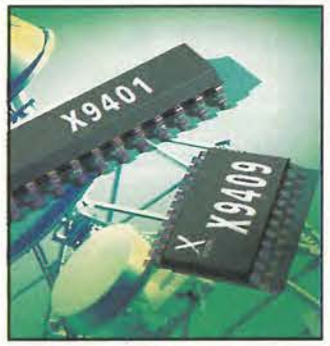
Fairchild now produces a 30V p-channel MOSFET – the FDS6675 – using the firm's Powertrench process. On-state resistance is 0.014Ω at a V_{GS} of 10V. It is for use as a power-management tool switching loads such as HDD, backlight or docking power switches. It can also be used as a battery switch to control the charging current. Gate charge is 30nC at 5V typical.
Fairchild Semiconductor
Tel: 01793 856819
Enquiry No 501

Leaded resistor range

VTM has introduced a leaded resistor, the BW, that can be supplied on reels suitable for auto-insertion machines. Virtually non-inductive and moulded in a cylindrical axial case, the resistor has a range from R010 to R062 (E12) in one per cent tolerance rated at 1.5W, 70°C. Applications include pulse circuits, switch controllers, frequency converters, current sensing in voltage regulators and power supplies.
VTM
Tel: 01494 738600
Enquiry No 504

Digital pots

Xicor now makes two quad 64-tap digital potentiometers. The X9401 and X9409 operate from one power supply and are for use in digital communications subsystems, cellular basestations and fibre optic transceivers. The non-volatile pots



operate from 2.7 to 5.5V and are for applications with typical standby current less than 0.2μA from – 40 to +85°C. End-to-end resistance options are 10 and 100kΩ. Both have four 6-bit EEPROM registers per potentiometer for storing tap settings and system parameters. This lets the designer automatically save current resistance values during power down and preset resistor values.
Xicor
Tel: 01933 700544
Enquiry No 505

Dual band noise amp targets GSM900 and 1800 front-ends

Infineon Technologies has introduced the PMB2362 dual band low noise amplifier IC for GSM900 and 1800 front-ends. It forms a GSM dual-band mobile phone chip-set with the Smarti PMB6250 transceiver and E-Gold+ baseband chip. The amp uses the firm's B6HF technology to give a transition frequency of 25GHz. One on-chip amp is for inputs between 0.9 and 1.0GHz and the other for 1.8 to 1.9GHz. Noise is 1.5dB at 0.95GHz and 2.0dB at 1.85GHz. Gain for the first amp is 17dB and for the second tuneable between 16 and 21dB. The device includes band switch, power down and low gain functions. The low gain mode reduces the gain by a fixed step of 20dB.
Infineon Technologies
Tel: 01344 396315
Enquiry No 503

Voice evaluation

DT Electronics has introduced a kit containing microphone, speaker and evaluation board with processor, preamplifiers, speaker driver and an RS232 interface to help engineers familiarise themselves with the Oki MSM6679A voice recognition processor. For use with a PC, the board includes the circuitry needed to demonstrate the main functions of the processor and the software lets designers use the speaker-independent and speaker-dependent capabilities of the chip. Software source code is included.
DT Electronics
Tel: 01203 466500
Enquiry No 506

Fibre-Channel adapter

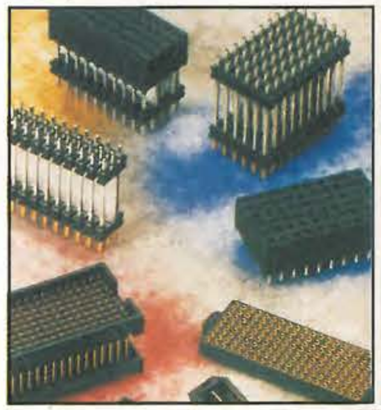
New from AMC is the 6526 single port Fibre-Channel adapter for CompactPCI systems. It uses standard Eurocard dimensions and gas-tight pin and socket connectors. Powered by the Hewlett-Packard Tachyon Fibre Channel protocol engine, the adapter supports applications from server storage clusters to digital video and image display systems. The architecture uses the Interphase IchipTPI Asic and an embedded 20-bit physical interface. The Fiberview GUI management tool for installation and configuration of Interphase Fibre Channel adapter cards is available separately.
Advanced Modular Computers
Tel: 01753 580660
Enquiry No 507

Electrical safety tester

Seaward has launched a portable appliance tester – the Europa Pac – that lets electrical contractors or facilities managers see if an electrical unit is safe and print results directly using an accessory thermal label printer via the RS interface. The LCD display shows test time, test icons and measured value. It can also be used to test IT equipment. The device can do earth continuity, AC and DC earth screen, insulation, load and leakage and European tests such as touch and substitute leakage test.
Seaward Electronic
Tel: 0191 261 8666
Enquiry No 508

Matched connectors

Genalog is producing matched impedance connectors on 0.5, 0.8 and 2mm pitch. Tested for impedance, VSWR, attenuation, crosstalk, propagation delay and rise time at frequencies from 10MHz to 1GHz, they provide board spacing from 5 to 32mm depending on the pitch selected. Also available are interfaces that let the number of rows, pins per row and board spacing be specified with standard connectors. These are available with 1.27 or 2mm pitch with up to 500 and 300 I/Os respectively. Surface-mount interfaces can be processed with standard BGA techniques, and design features permit one-pass infra-red soldering of high-density through-hole types.
Genalog
Tel: 01580 753754
Enquiry No 510



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PC audio accelerator

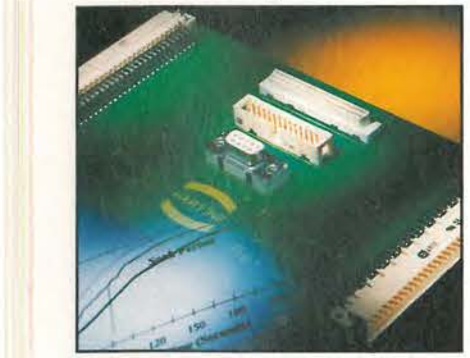
Philips Semiconductors has announced the Thunderbird Avenger PCI audio accelerator that enables 5.1 channel playback of games, music and movies. It converts all types of media into 5.1 channel output and combines hardware acceleration, two-speaker 3D virtualisation and multi-channel speaker support. The programmable DSP controller provides 5.1 speaker playback of music CDs, videos, MP3 files and Midi files. Movies designed for 5.1 playback can be improved using QSound Labs' algorithms for the accelerator. The device provides a similar theatre experience on four channel speaker systems and can provide two speaker and headphone 3D virtualisation.
Philips Semiconductor
Tel: 00 31 40 272 2091
Enquiry No 509

DSP core

LSI Logic is making available a DSP core based on the ZSP DSP architecture. The ZSP400 is for system-on-a-chip designs using Coreware for Asic implementations. Performance is more than 400 million multiply-accumulates per second for one core.
LSI Logic
Tel: 01344 413209
Enquiry No 512

SM connectors

Surface mount compatible connectors that can handle pin-in-hole intrusive reflow soldering have been launched



by Harting. They are compatible with automatic pick-and-place assembly machines. As a result, they can be reflow soldered simultaneously with other SM components on single and double-sided plated-through boards.
Harting
Tel: 01604 766686
Enquiry No 513

SiGe VCOs

Maxim has introduced the Max 2622 and 2623 SiGe voltage-controlled oscillators for ISM-band radios and Dect cordless phones. The 2622 has an 855 to 881MHz tuning range, making it suitable for systems operating in the European 866 to 868MHz ISM band, which uses an IF of around 11MHz. The 2623 has an 885 to 950MHz tuning range, making it suitable for Dect systems in receive and transmit mode. The 2623 can also be used in US ISM-band systems with IF frequencies around 11MHz. The soon-to-be-released 2624, for the US ISM band, can be used with 45 and 70MHz IF. The 2622 and 2623 have outputs that can be matched to 50Ω using two capacitors. They are packaged in eight-pin μmax packages and work from –40 to +85°C.
Maxim
Tel: 0118 930 3388
Enquiry No 514

Reference for 0.04% accuracy

Linear Technology is manufacturing the new LT1461 2.5V, 3ppm/°C, low dropout series reference with an initial accuracy of ±0.04 per cent at 1mV. It draws 50μA maximum supply

current. Operating temperature can be 0 to +70°C, –40 to +85°C or –40 to +125°C. Input voltage range is 2.8 to 20V. Applications include battery powered equipment, industrial controls and measurement equipment. Dropout voltage is 300mV maximum at an output current of 1mA.
Linear Technology
Tel: 01276 677676
Enquiry No 515

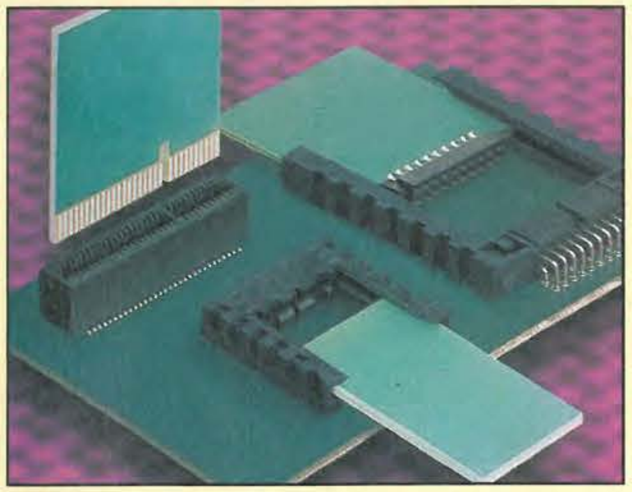
DC/DC in SM

The SM packaging of Datel's UNR 3.6 to 10W non-isolated DC-to-DC converters is compatible with pick-and-place and solder reflow processes. The converters measure 2.54 by 2.54 by 1.19cm and weigh 17g. They can be installed using vacuum-pickup, automatic pick-and-place equipment.
Datel
Tel: 01256 880444
Enquiry No 516

PC card bus connectors

Hirose has introduced a single-slot PC card bus connector, the IC11S. There's

three ejector button options for either side of the connector – pop up, fold away or ridged. They are for use with types I, II or III cards and have eight gold plated contacts. Options include standard or reverse PCB orientation.
Hirose
Tel: 01908 260616
Enquiry No 517



Mini-edge card connectors

Samtec's MEC1 mini-edge card connectors have a double row of beryllium copper contacts on 1mm pitch for micro card interfaces. These sockets are available in seven sizes from five to 68 contacts per row for a total of 136 I/Os in less than 645mm² of PCB area. The sockets are normally polarised to provide error free mating, however, smaller sizes with up to 30 contacts per side may be specified without polarisation to increase I/O density. The firm also has MB1 micro bay interfaces for mating with 0.8 and 1.6mm mini-cards moulded with 20, 30, 40 or 50 contacts on 1mm pitch with 25.4mm long card guides to align daughtercard traces to the contacts.
Samtec
Tel: 01236 739292
Enquiry No 511

Cable to PCB connector

The P50LE from Robinson Nugent is a SM cable to board connector using a contact design developed for laptop, blind mating plug-in devices. It has a

cable connection option with a mated height less than 12mm, a one-touch locking and ejector system, and self alignment when mating. Contact counts are 40, 50, 68, 80 and 100 positions. The 100 position model is 80mm long. Current rating is 0.5A per contact and the insulator material is UL 94V-0 rated.
Robinson Nugent
Tel: 01227 794495
Enquiry No 518

Embedded PCI

Quicklogic has announced three embedded standard product (ESP) QuickPCI devices – the QL5030, QL5130 and QL5232 – bringing the family up to five devices. They combine embedded PCI controllers with programmable logic. PCI bus performance is up to 600Mbyte/s with zero wait states and independent back-end clock speeds up to 160MHz. Reference development kits with boards, devices and software drivers are available. They are for 32 and 64-bit PCI buses at speeds of 33, 66 and 75MHz.
Quicklogic
Tel: 00 1 408 990 4000
Enquiry No 519

Resettable fuse

Raychem has launched the Polyswitch VTP240 resettable fuse for battery protection. Hold current is 2.4A enabling overcurrent and overtemperature protection for

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handheld wireless products. It is a polymeric positive temperature coefficient fuse that functions as a low-resistance series element between battery cells and pack circuits. When the terminals of a pack are shorted, causing an overcurrent, the resistance of the device rapidly increases, reducing the discharge of the battery to a safe and protected level.
Raychem
Tel: 00 32 70 233 041
Enquiry No 520

PC audio accelerator

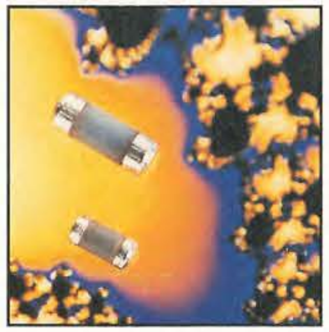
Aluminum electrolytic capacitors with a 150°C temperature rating for automotive, aerospace and industrial applications have been released by Vishay. The axial-style devices are for smoothing, filtering, coupling and decoupling circuits. The EBH and EGH have capacitance options from 220 to 22,000µF with tolerances of ±20 per cent. Voltage rating is 10, 16, 25 or 40V with maximum current ratings up to 3.9A.
Vishay Intertechnology
Tel: 00 49 07661 37 253
Enquiry No 523

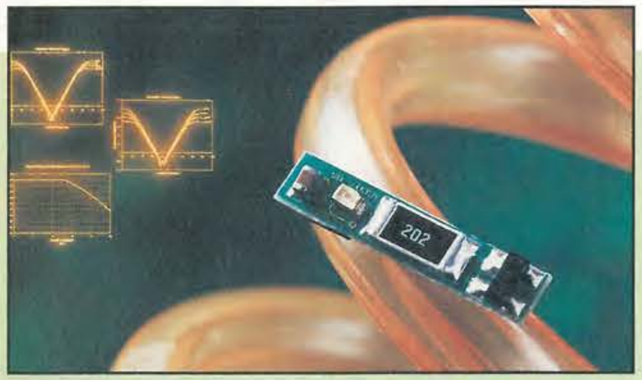
PCB calculation software

The CITS 25 field-solving controlled impedance calculator from Polar has been upgraded to include impedance calculation equations for coated, embedded, offset and other coplanar structures. Windows-based, it can be used on its own or with CAD tools, and can calculate design and yield for manufacturers involved in fabricating controlled impedance boards.
Polar Instruments
Tel: 01471 253081
Enquiry No 524

Cylindrical inductors

BC Components has launched cylindrical thin-film inductors that can work up to 2GHz. They can be used by pick-and-place machines and their behaviour is independent of orientation. The components are available in Mini-Melf (IMA 0204) and Micro-Melf (IMU 0102) formats. Inductance values are from 10 to 100nH with tolerances of ±10 per cent





Magnetic sensor comes with reverse battery protection

For position sensing in pneumatic cylinders, the AG007-07 magnetic sensor from Rhopoint is based on giant magneto-resistive (GMR) technology. It has a digital switch with controlled magnetic operate and release points. Features include reverse battery protection. It can also detect speed and direction of motion. The GMR effect changes the electrical resistance when stacked layers of ferromagnetic and non-magnetic materials are exposed to a magnetic field.
Rhopoint Components
Tel: 01883 717983
Enquiry No 521

Coaxial adapters pass test

PSP Electronics has released a range of coaxial adapters for electronic test applications. Made by Pomona Electronics, they come in various configurations to enable conversion between connector types including BNC, N, SMA, TNC, UHF, banana plugs and binding posts.
PSP Electronics
Tel: 0208 903 9061
Enquiry No 522

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and self resonant frequencies from 1500 to 4000MHz.
BC Components
Tel: 00 31 40 259 0724
Enquiry No 526

Configurable processor

Ambar has introduced a configurable processor based E5 system-on-chip developed by Triscend. It combines an 8032 microcontroller, programmable logic, RAM, high-speed bus and other system functions. The firm's Fastchip development software is available to customise the microcontroller. The microcontroller has 40MHz operation up to 10Mips, up to 40 000 gates of programmable logic and up to 64kbyte of on-chip dedicated SRAM. Features include debug capability, standalone operation from an external memory, power-down and power-management modes, embedded support functions, SRAM-based configurable system logic matrix and a dedicated system bus.
Amber Components
Tel: 01296 397396
Enquiry No 525

500V Electrolytics

For power supplies and industrial drives, Panasonic's 500V TS-UP aluminium electrolytic capacitors are available with capacitance values between 47 and 470µF and are rated for surge voltages up to 550V DC. Case diameters are from 40 to 22mm with case profiles down to 25mm.
Panasonic
Tel: 01344 853349
Enquiry No 527

Modular connector

A Molex modular connector combines power and signal functions. The Compodre consists of a shroud with internal pods that accept any combination of Mini-Fit and Micro-Fit connectors. It is for applications requiring varying amperages or a combination of power and signal contacts. These include power supplies, computers, copiers, generators, vending and gaming equipment, HVAC equipment and industrial machinery. The modular design requires no tools for assembly. Polarised pods help prevent improper assembly and polarised shrouds help prevent mismating. An active latch on the shroud produces an audible click to indicate proper connection.
Molex
Tel: 01420 488488
Enquiry No 528

Flat panel controller

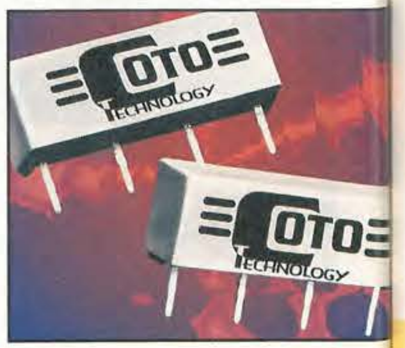
Silicon Image has announced the Panellink Sil 851 flat-panel monitor controller. There are plans to incorporate it into 38.4 and 46cm LCD monitors. The monitors interface with computers and other video sources via the DVI specification. The controller combines a digital receiver, image processing and display adaption layer. It is for displays up to SXGA resolutions (1280 x 1024 pixels). Features include automatic scaling up to XGA or SXGA resolutions, built-in dithering, programmable on-screen-display controller to support localised character sets, gamma, brightness and contrast control via three programmable look up tables, and software development tools.
Silicon Image
Tel: 00 1 408 873 3111
Enquiry No 529

Graphics card

Imagine Graphics has released the Jeronimo 2000 dual monitor graphics card that lets users operate two displays from one AGP or PCI slot at resolutions up to 1920 x 1080 x 32bpp at 85Hz, supporting 2D and 3D acceleration, OpenGL and DirectX. For CAE, CAD, DTP, digital content creation and financial trading applications, it is compatible with Windows 95, 98, NT 4.0 and 2000. Built with two Permedia Three chips from 3Dlabs, the card delivers images and scenes with DVD motion compensation, colour space conversion and 125 million multi-texture pixels per second fill rate. There is an option for users to receive real-time video feeds from a baseband source such as a VCR, camcorder or standard composite Pal and NTSC source. The optional external Appian TV tuner also lets users view multiple channels of live television broadcasts. It is bundled with Hydravision display management software so users can position windows and dialogue boxes, direct and manipulate information across multiple monitors, and set independent resolution and refresh rates for each monitor.
Image Graphics
Tel: 01727 844744
Enquiry No 530

Reed relay

Coto Technology has added a reed relay to its line of Spartan SIPs. The 5V 9007-05-40 has 1kΩ coil resistance and includes an external magnetic shield. The SIP is



available in Form A with an optional diode. It is 0.635cm wide and 2.06cm long with 0.5cm spacing between the four leads. Contact rating is 10W and it provides up to 200V or 0.5A switching. Coto Technology is for security, industrial and telecoms applications.
Coto Technology
Tel: 001 401 943 2686
Enquiry No 531

16-bit transceiver

Vitesse Semiconductor has introduced a 16-bit transceiver for telecoms backplane applications that do not require tight Sonet jitter specifications. The VSC7164 incorporates four functions – multiplexer, demultiplexer, clock recovery unit (CRU) and clock multiplication unit (CMU). It serialises a 16-bit data bus at 155.52Mbit/s onto

redundant PECL outputs at 2.488Gbit/s. The integrated CMU generates the 2.488GHz clock from either a 77.76 or 155.52MHz reference clock input. Serial 2.488Gbit/s data on the selected redundant input buffer is recovered in the digital CRU. The output of the CRU is deserialised onto a 16-bit data bus at 155.52MHz. It has a 3.3V supply and is packaged in a 100-pin, 14mm PQFP.
Vitesse Semiconductor
Tel: 01634 863494
Enquiry No 532

16-bit transceiver

Clear Logic has introduced its CL10KA family of no-NRE Asics that are functionally identical to Altera's user-programmable Flex 10KA FPGAs including pin-out and I/O characteristics. As a result, the Altera devices can be used to prototype and debug the CL devices. They use vertical link-configured Asic technology. Any IP core for the Altera architecture will function identically in the CL architecture. Once the designs have been prototyped in hardware using the FPGA, designers can send their bit streams to Clear Logic through the internet and receive Asic samples within two weeks. The CL devices consist of logic units containing a four-input look-up-table with a register. The logic units are arranged in blocks of eight with local connections between them. For every row of LBBs there is a 2kbit block of embedded SRAM.
Clear Logic
Tel: 001 408 361 2600
Enquiry No 533

Protocol controller IC

Infineon has introduced a multi-channel network interface controller, the Munich 128x PEB20324, for datacomms and telecoms. It can handle up to 128 full duplex serial PCM channels. It combines four independent 24132-channel HDLC controllers each with a dedicated 64-channel DMA controller and a serial PCM interface controller. Processed data is transferred to host memory via the PCI 2.1 interface or demultiplexed bus interface. The IC performs layer-two HDLC formatting and deformatting at data rates from 8kbit/s to 2048Mbit/s or V.11 OIX.30 protocols up to data rates of 38.4kbit/s.
Infineon Technologies
Tel: 0990 550 500
Enquiry No 534

16-bit transceiver

The new Panasonic EVQPP rectangular touch switches from Flint have snap action and push-on SPST, suitable for reflow soldering. Profile is 6.8mm maximum and footprint 3.7 by 2.5mm with a rectangular knob. Operating force is either 1.6 or 2.4N. They can be used as signal input

switches on portable CD and MD players. Options include straight or J-bent terminals, with or without ground terminals. The self-cleaning contact is rated at 50mA, 12V DC. n
Flint Distribution
Tel: 01530 510333
Enquiry No 535

Power factor

Continued from page 118

The combination of resistive power and reactive power gives rise to total power as shown in the phasor diagram, **Fig. 1**. It is assumed that the voltage is horizontal so that the phasor diagram shows the direction of the current.

Only the resistive power is available to the consumer. In the phasor diagram resistive, i.e. usable, power is denoted as kilowatts, or kW. The total power is normally denoted as kVA. The reactive power is denoted as kVAr.

Phase angles

The angle between the total power and the resistive power is usually represented by the Greek letter psi, ψ. The ratio of resistive to total power is called the power factor and can take any value from zero representing very bad, to one representing very good. For those who still remember their high school trigonometry, this ratio is also the cosine of ψ, usually written $\cos\varphi$. Although we have drawn the diagram with current leading, we could equally have drawn it with the current lagging. This makes no difference to the definition of power factor that can clearly not exceed 1.

There are disadvantages in having too low a power factor. Considerably more current has to be supplied than is actually used. Thus, much more demands are made on the supply equipment and on the wiring of the consumer. Too low a power factor with inadequately rated wiring can give rise to overheating and even fires.

There can be a voltage drop at the consumer's outlet due to excessive current in the supply. The electricity supply companies do not like this as they too have to use much higher rated wiring then they would need and are supplying electric current that they do not get paid for. Supply companies lay down minimum power factors for their customers and usually extract hefty payments from those not conforming.

There used to be myths floating around, certainly in SE Lancashire 50 years ago, that it was possible to cheat the supply company by putting a large capacitance or inductance in series with one's supply. This was usually fostered by stories in the tabloid press about strangers offering devices in the local pub that would enable one to reduce ones electricity bill.

However, as you can see from the phasor diagram, one can certainly take more electricity from the supply, but there is no way one can reduce the bill using the same apparatus. On the contrary. Any added capacitance or inductance will have some resistance associated with it. While of no use to the consumer, it will in fact increase his or her electricity bill, and the chance of overloading the wiring.

So how do I measure it?

The measurement of power factor is important not only for the supplier, but also for the consumer. We have designed and built a cpu controlled digital meter to measure both power factor, voltage and frequency of the supply.

What the meter does not measure is harmonic content. We should emphasise that the measurement of the power factor is complicated where there is considerable harmonic content. This device simply measures the difference in phase between the current and the voltage.

The instrument works by measuring the interval between where the current crosses zero and where the voltage crosses zero, by counting pulses of known frequency 32.5kHz between

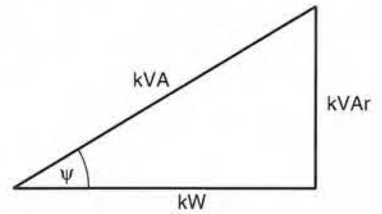


Fig. 1. A combination of resistive and reactive power gives rises to total power.

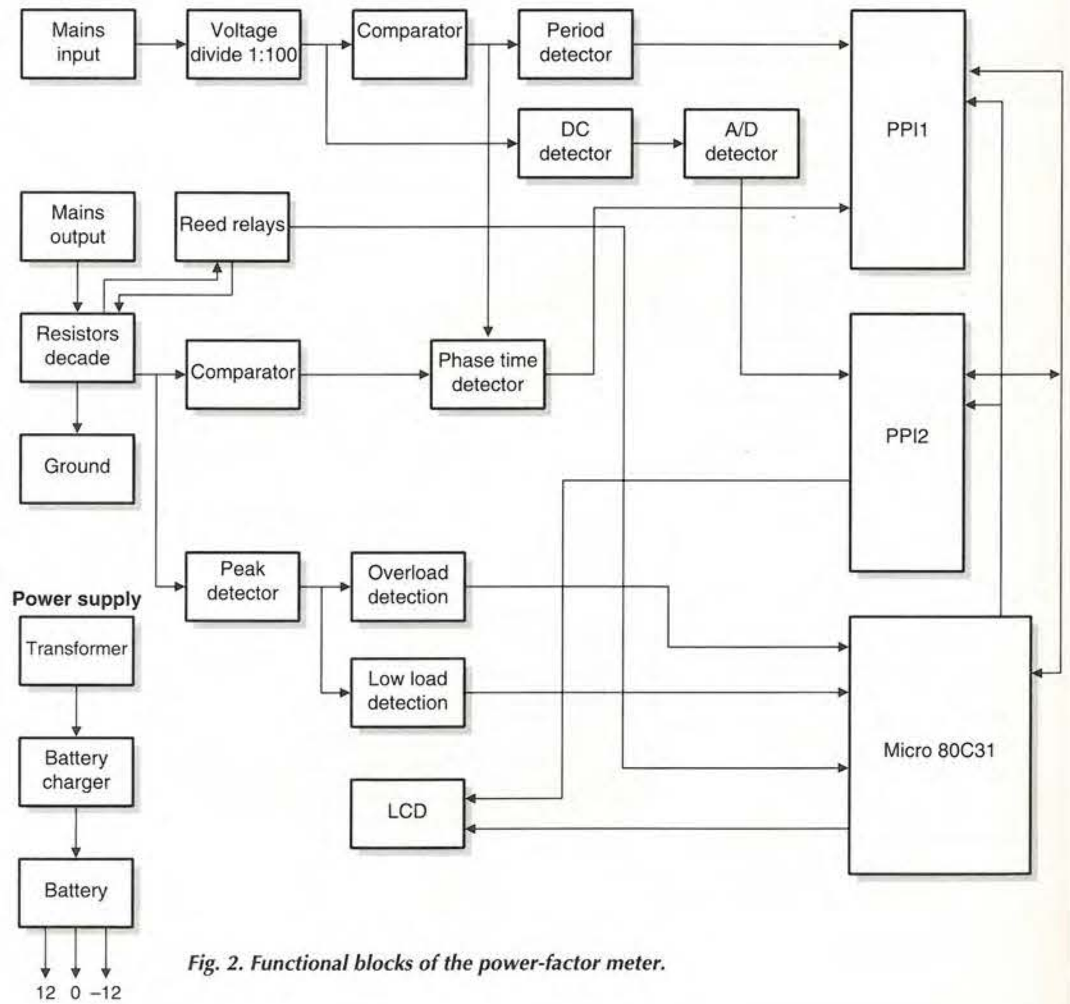


Fig. 2. Functional blocks of the power-factor meter.

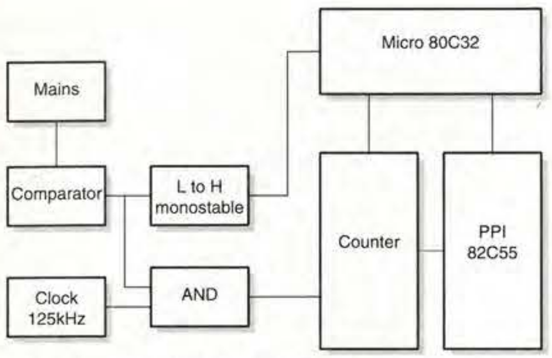


Fig. 3. Elements of the circuit involved in frequency measurement.

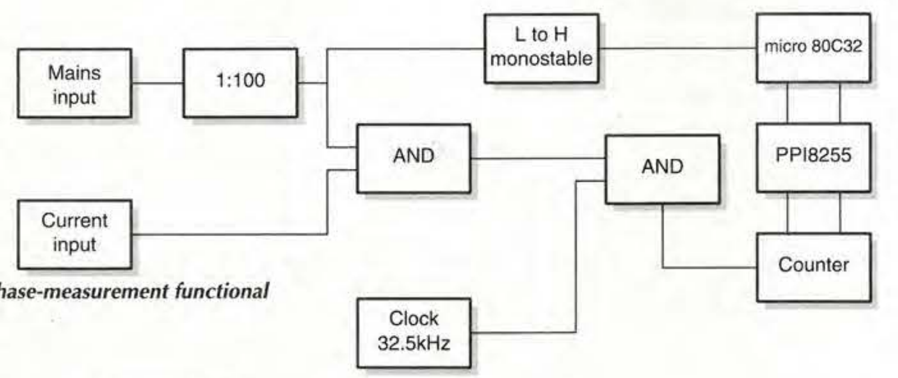


Fig. 4. Phase-measurement functional blocks.

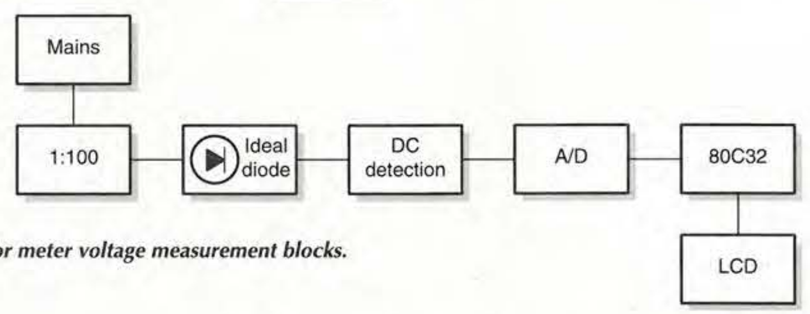


Fig. 5. Power-factor meter voltage measurement blocks.

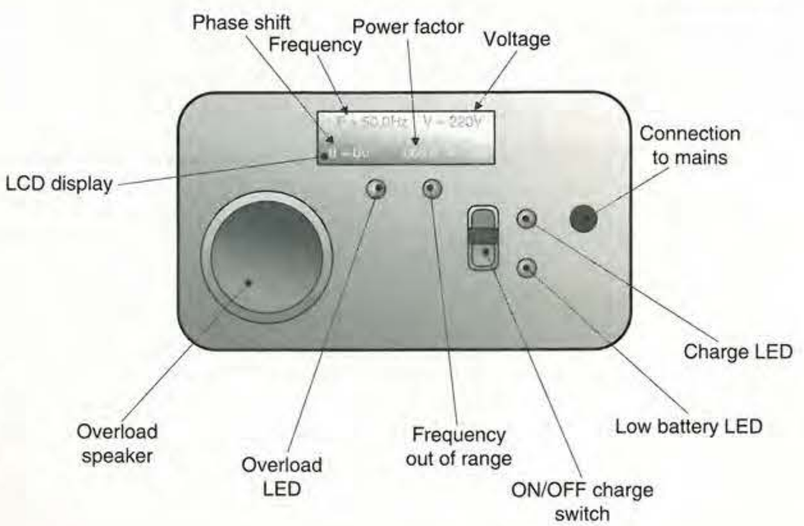


Fig. 6. Front panel of our prototype power-factor meter.

these two states. Knowing the frequency of the pulses and the frequency of the supply, which is also measured, this interval is converted into a phase angle or power factor

As the main interest in this device would presumably be in countries with rather erratic supply, we have also included a rechargeable battery. It is possible that at the extremes of low voltage and erratic frequency, the device would not work well without it.

The circuit diagram consists of several sub-modules. These are the external load for inputting the mains voltage and current, the frequency and phase detector section, the microcontroller together with its associated EPROM which contains the program to operate the instrument.

Auxiliary circuits

The crystal oscillator gives the pulses used to measure the frequency and phase differences. Other auxiliary circuits are the overload detector, the power supply and the battery charger, which can of course be omitted.

The alphanumeric LCD display made by Citizen is a sub-unit consisting of the display itself together with two onboard driver chips, HD44780A00 and HD44100H . This unit now appears to be obsolete but similar units are available from Seiko and Sanyo, with the same device number and programming. A wide range of characters is available on the display, which has a 16x2 matrix. You will need to get the data sheet to learn what codes give what alphanumeric characters.

The digital and analogue sections have their own separate earths. The instrument is shown in schematic form in **Fig. 2**. The units marked PPI are programmable peripheral interfaces.

Figure 3 shows in schematic form how the frequency is measured while **Fig. 4** illustrates the measurement of phase. **Figure 5** shows how the voltage is measured and **Fig. 6** a drawing of the front panel and the LCD.

To monitor the mains it is necessary to use a number of resistors in parallel and series to allow for a large variation in voltage. We have provided three ranges,

0.5 to 7.5A 0.33Ω
7.5 to 12.5A 0.2Ω
12.5 to 17.5A 0.14Ω

We have included a control circuit so that if the current is too much for the lower range resistor than the others resistors are switched in using the reed relays or switched out to the lower ranges if required. If however it is not possible to find a sufficiently low resistor, i.e. there is overload, then the overload indicator light comes on and more importantly the loudspeaker emits a very loud tone.

Although we have included this section in our instrument box, this is rather problematic, as due to the high currents and voltage there is a possibility of fire. It would almost certainly be safer to put these resistors and control in a separate grounded metal box.

Software

While the hardware of this project might seem rather complex, this is nothing in comparison to the programming. There are several hundred lines of code. It is not possible in an

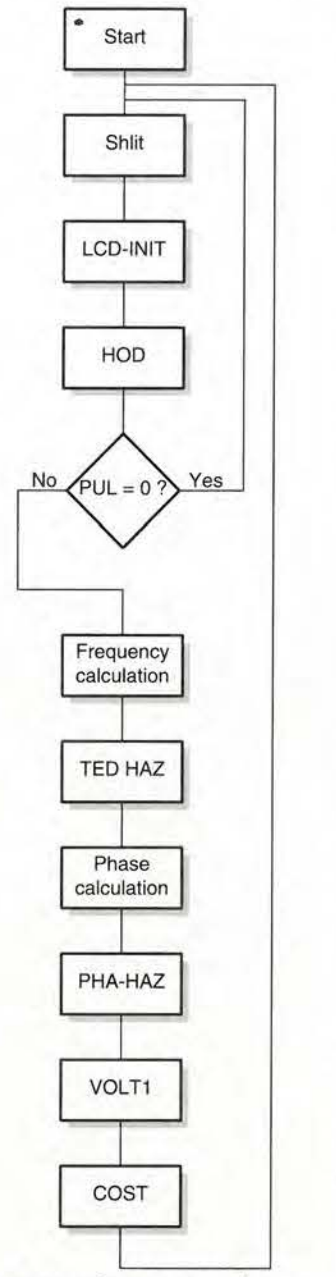


Fig. 7. Power-factor meter software flow outline. Labels in each function block represent subroutines in the source code and are explained in the text.

Hex dump of the power-factor meter's eprom.


```
000000 02 0e fd 02 0e 84 78 2f a6 8c 78 2e a6 8a 78 2d a6 8b 75 8b 00 90  ..˝...x/¶.x.¶.x-¶.u...
000016 10 00 e0 f5 1f 90 90 00 e0 78 32 f6 d2 90 c2 90 75 8a 00 75 8c 00  ..‡ı....‡x2ˆ".¬.u..u..
00002c 05 1e e5 1e d3 94 0a 40 03 75 1e 01 22 c2 92 e5 22 90 20 00 f0 d2  ..Â."..@.u.."¬íÂ". "
000042 92 74 0a 12 0e b5 c2 92 22 d2 92 74 96 12 0e b5 75 22 01 e5 22 d3  ít...µ¬í""ít...µu".Â""
000058 94 03 50 19 d2 92 c2 91 74 30 90 20 00 f0 74 32 12 0e b5 c2 92 78  ..P."í¬ët0. t2..µ¬íx
00006e 22 74 01 26 f6 50 e0 c2 91 c2 92 74 38 90 20 00 f0 d2 92 74 14 12  "t.&ˆP‡¬ë¬ít8. "ít..
000084 0e b5 c2 92 74 08 90 20 00 f0 d2 92 74 14 12 0e b5 c2 92 90 20 00  .µ¬ít.. "ít...µ¬í. .
00009a 74 01 f0 d2 92 74 14 12 0e b5 c2 92 74 07 90 20 00 f0 d2 92 74 14  t."ít...µ¬ít.. "ít.
0000b0 12 0e b5 c2 92 74 38 90 20 00 f0 d2 92 74 14 12 0e b5 c2 92 74 0f  ..µ¬ít8. "ít...µ¬ít.
0000c6 90 20 00 f0 d2 92 74 14 12 0e b5 c2 92 74 06 90 20 00 f0 d2 92 74  . "ít...µ¬ít.. "ít
0000dc 14 12 0e b5 c2 92 22 c2 91 75 22 02 12 00 39 d2 91 75 22 66 12 00  ...µ¬í"¬ëu"...9"ëu"f..
0000f2 39 75 22 3d 12 00 39 ae 23 af 24 74 0a 8e 04 8f 05 ff 12 0e c5 78  9u"=..9Æ#Ø$t.....˘..=x
000108 2c a6 05 ed 24 30 78 2c f6 78 2c 86 22 12 00 39 ae 23 af 24 74 0a  ,¶.Ì$0x,ˆx,."..9Æ#Ø$t.
00011e 8e 04 8f 05 ff 12 0e f1 78 2c a6 05 ed 24 30 78 2c f6 78 2c 86 22  .....˘..Òx,¶.Ì$0x,ˆx,."
000134 12 00 39 22 75 22 f2 12 00 39 75 22 3d 12 00 39 ae 23 af 24 74 0a  ..9"u"Ú..9u"=..9Æ#Ø$t.
00014a 8e 04 8f 05 ff 12 0e c5 78 2c a6 05 ed 24 30 78 2c f6 78 2c 86 22  ....˘..=x,¶.Ì$0x,ˆx,."
000160 12 00 39 ae 23 af 24 74 0a 8e 04 8f 05 ff 12 0e f1 78 2c a6 05 ed  ..9Æ#Ø$t.....˘..Òx,¶.Ì
000176 24 30 78 2c f6 78 2c 86 22 12 00 39 75 22 df 12 00 39 22 c2 91 75  $0x,ˆx,."..9u"ﬂ..9"¬ëu
00018c 22 02 12 00 39 d2 91 75 22 4e 12 00 39 75 22 4f 12 00 39 75 22 54  "...9"ëu"N..9u"O..9u"T
0001a2 12 00 39 75 22 20 12 00 39 75 22 43 12 00 39 75 22 4f 12 00 39 75  ..9u" ..9u"C..9u"O..9u
0001b8 22 4e 12 00 39 75 22 4e 12 00 39 75 22 45 12 00 39 75 22 43 12 00  "N..9u"N..9u"E..9u"C..
0001ce 39 75 22 54 12 00 39 75 22 45 12 00 39 75 22 44 12 00 39 75 22 2e  9u"T..9u"E..9u"D..9u".
0001e4 12 00 39 74 c8 12 0e b5 74 c8 12 0e b5 74 c8 12 0e b5 22 a2 95 b3  ..9t»..µt»..µt»..µ"¢.≥
0001fa 40 0d a2 94 b3 40 02 d2 97 a2 94 40 02 d2 94 a2 96 b3 40 0d a2 97  @.¢.≥@.".¢.@.".¢.≥@.¢.
000210 40 02 c2 94 a2 97 b3 40 02 c2 97 22 75 22 56 12 00 39 75 22 3d 12  @.¬.¢.≥@.¬."u"V..9u"=.
000226 00 39 90 60 00 e0 78 38 f6 78 38 e6 75 f0 f4 a4 ff ae f0 78 3a a6  .9.`.‡x8ˆx8ÊuÙ§˘Æx:¶
00023c 06 08 a6 07 74 ff 8e 04 8f 05 ff 12 0e c5 78 38 a6 05 ed 75 f0 65  ..¶.t˘....˘..=x8¶.Ìue
000252 a4 ff ae f0 78 3a a6 06 08 a6 07 74 64 8e 04 8f 05 ff 12 0e c5 ed  §˘Æx:¶..¶.td....˘..=Ì
000268 78 38 f6 78 38 e6 75 f0 05 84 ff 7e 00 ef 75 f0 05 a4 fc af f0 78  x8ˆx8Êu..˘~.Ôu.§,Øx
00027e 38 a6 04 ec 75 f0 05 84 ad f0 7e 00 78 39 a6 05 ed d3 94 02 40 08  8¶.Ïu..≠~.x9¶.Ì"..@.
000294 78 38 e6 24 05 78 38 f6 78 38 e6 75 f0 64 84 ff 7e 00 78 39 a6 07  x8Ê$.x8ˆx8Êud.˘~.x9¶.
0002aa ef 24 30 78 39 f6 78 39 86 22 12 00 39 78 38 e6 75 f0 64 84 af f0  Ô$0x9ˆx9."..9x8Êud.Ø
0002c0 7e 00 78 38 a6 07 ef 75 f0 0a 84 fd 7c 00 78 39 a6 05 ed 24 30 78  ~.x8¶.Ôu..˝|.x9¶.Ì$0x
0002d6 39 f6 78 39 86 22 12 00 39 78 38 e6 75 f0 0a 84 af f0 7e 00 78 39  9ˆx9."..9x8Êu..Ø~.x9
0002ec a6 07 ef 24 30 78 39 f6 78 39 86 22 12 00 39 75 22 20 12 00 39 75  ¶.Ô$0x9ˆx9."..9u" ..9u
000302 22 76 12 00 39 22 75 22 63 12 00 39 75 22 6f 12 00 39 75 22 73 12  "v..9"u"c..9u"o..9u"s.
000318 00 39 75 22 f2 12 00 39 75 22 3d 12 00 39 ae 23 af 24 74 06 be 00  .9u"Ú..9u"=..9Æ#Ø$t.æ.
00032e 08 d3 9f 40 04 78 3d 76 64 ae 23 af 24 74 06 6f 4e 70 04 78 3d 76  .".@.x=vdÆ#Ø$t.oNp.x=v
000344 63 ae 23 af 24 74 07 6f 4e 70 04 78 3d 76 63 ae 23 af 24 74 08 6f  cÆ#Ø$t.oNp.x=vcÆ#Ø$t.o
00035a 4e 70 04 78 3d 76 63 ae 23 af 24 74 09 6f 4e 70 04 78 3d 76 63 ae  Np.x=vcÆ#Ø$t.oNp.x=vcÆ
000370 23 af 24 74 0a 6f 4e 70 04 78 3d 76 62 ae 23 af 24 74 0b 6f 4e 70  #Ø$t.oNp.x=vbÆ#Ø$t.oNp
000386 04 78 3d 76 62 ae 23 af 24 74 0c 6f 4e 70 04 78 3d 76 62 ae 23 af  .x=vbÆ#Ø$t.oNp.x=vbÆ#Ø
00039c 24 74 0d 6f 4e 70 04 78 3d 76 61 ae 23 af 24 74 0e 6f 4e 70 04 78  $t.oNp.x=vaÆ#Ø$t.oNp.x
0003b2 3d 76 61 ae 23 af 24 74 0f 6f 4e 70 04 78 3d 76 61 ae 23 af 24 74  =vaÆ#Ø$t.oNp.x=vaÆ#Ø$t
0003c8 10 6f 4e 70 04 78 3d 76 60 ae 23 af 24 74 11 6f 4e 70 04 78 3d 76  .oNp.x=v`Æ#Ø$t.oNp.x=v
0003de 60 ae 23 af 24 74 12 6f 4e 70 04 78 3d 76 5f ae 23 af 24 74 13 6f  `Æ#Ø$t.oNp.x=v_Æ#Ø$t.o
0003f4 4e 70 04 78 3d 76 5f ae 23 af 24 74 14 6f 4e 70 04 78 3d 76 5e ae  Np.x=v_Æ#Ø$t.oNp.x=v^Æ
00040a 23 af 24 74 14 be 00 03 9f 50 09 74 29 be 00 04 d3 9f 50 03 02 05  #Ø$t.æ...P.t)æ..".P...
000420 39 ae 23 af 24 74 15 6f 4e 70 04 78 3d 76 5d ae 23 af 24 74 16 6f  9Æ#Ø$t.oNp.x=v]Æ#Ø$t.o
000436 4e 70 04 78 3d 76 5d ae 23 af 24 74 17 6f 4e 70 04 78 3d 76 5c ae  Np.x=v]Æ#Ø$t.oNp.x=v\Æ
00044c 23 af 24 74 18 6f 4e 70 04 78 3d 76 5b ae 23 af 24 74 19 6f 4e 70  #Ø$t.oNp.x=v[Æ#Ø$t.oNp
000462 04 78 3d 76 5b ae 23 af 24 74 1a 6f 4e 70 04 78 3d 76 5a ae 23 af  .x=v[Æ#Ø$t.oNp.x=vZÆ#Ø
000478 24 74 1b 6f 4e 70 04 78 3d 76 59 ae 23 af 24 74 1c 6f 4e 70 04 78  $t.oNp.x=vYÆ#Ø$t.oNp.x
00048e 3d 76 58 ae 23 af 24 74 1d 6f 4e 70 04 78 3d 76 57 ae 23 af 24 74  =vXÆ#Ø$t.oNp.x=vWÆ#Ø$t
0004a4 1e 6f 4e 70 04 78 3d 76 57 ae 23 af 24 74 1f 6f 4e 70 04 78 3d 76  .oNp.x=vWÆ#Ø$t.oNp.x=v
0004ba 56 ae 23 af 24 74 20 6f 4e 70 04 78 3d 76 55 ae 23 af 24 74 21 6f  VÆ#Ø$t oNp.x=vUÆ#Ø$t!o
0004d0 4e 70 04 78 3d 76 54 ae 23 af 24 74 22 6f 4e 70 04 78 3d 76 53 ae  Np.x=vTÆ#Ø$t"oNp.x=vSÆ
0004e6 23 af 24 74 23 6f 4e 70 04 78 3d 76 52 ae 23 af 24 74 24 6f 4e 70  #Ø$t#oNp.x=vRÆ#Ø$t$oNp
0004fc 04 78 3d 76 51 ae 23 af 24 74 25 6f 4e 70 04 78 3d 76 50 ae 23 af  .x=vQÆ#Ø$t%oNp.x=vPÆ#Ø
000512 24 74 26 6f 4e 70 04 78 3d 76 4f ae 23 af 24 74 27 6f 4e 70 04 78  $t&oNp.x=vOÆ#Ø$t'oNp.x
000528 3d 76 4e ae 23 af 24 74 28 6f 4e 70 04 78 3d 76 4d ae 23 af 24 74  =vNÆ#Ø$t(oNp.x=vMÆ#Ø$t
00053e 28 be 00 03 9f 50 09 74 3d be 00 04 d3 9f 50 03 02 06 69 ae 23 af  (æ...P.t=æ..".P...iÆ#Ø
000554 24 74 29 6f 4e 70 04 78 3d 76 4b ae 23 af 24 74 2a 6f 4e 70 04 78  $t)oNp.x=vKÆ#Ø$t*oNp.x
00056a 3d 76 4a ae 23 af 24 74 2b 6f 4e 70 04 78 3d 76 49 ae 23 af 24 74  =vJÆ#Ø$t+oNp.x=vIÆ#Ø$t
000580 2c 6f 4e 70 04 78 3d 76 48 ae 23 af 24 74 2d 6f 4e 70 04 78 3d 76  ,oNp.x=vHÆ#Ø$t-oNp.x=v
000596 47 ae 23 af 24 74 2e 6f 4e 70 04 78 3d 76 46 ae 23 af 24 74 2f 6f  GÆ#Ø$t.oNp.x=vFÆ#Ø$t/o
0005ac 4e 70 04 78 3d 76 44 ae 23 af 24 74 30 6f 4e 70 04 78 3d 76 43 ae  Np.x=vDÆ#Ø$t0oNp.x=vCÆ
0005c2 23 af 24 74 31 6f 4e 70 04 78 3d 76 42 ae 23 af 24 74 32 6f 4e 70  #Ø$t1oNp.x=vBÆ#Ø$t2oNp
0005d8 04 78 3d 76 40 ae 23 af 24 74 33 6f 4e 70 04 78 3d 76 3f ae 23 af  .x=v@Æ#Ø$t3oNp.x=v?Æ#Ø
0005ee 24 74 34 6f 4e 70 04 78 3d 76 3e ae 23 af 24 74 35 6f 4e 70 04 78  $t4oNp.x=v>Æ#Ø$t5oNp.x
000604 3d 76 3c ae 23 af 24 74 36 6f 4e 70 04 78 3d 76 3b ae 23 af 24 74  =v<Æ#Ø$t6oNp.x=v;Æ#Ø$t
00061a 37 6f 4e 70 04 78 3d 76 39 ae 23 af 24 74 38 6f 4e 70 04 78 3d 76  7oNp.x=v9Æ#Ø$t8oNp.x=v
000630 38 ae 23 af 24 74 39 6f 4e 70 04 78 3d 76 37 ae 23 af 24 74 3a 6f  8Æ#Ø$t9oNp.x=v7Æ#Ø$t:o
000646 4e 70 04 78 3d 76 35 ae 23 af 24 74 3b 6f 4e 70 04 78 3d 76 34 ae  Np.x=v5Æ#Ø$t;oNp.x=v4Æ
00065c 23 af 24 74 3c 6f 4e 70 04 78 3d 76 32 ae 23 af 24 74 3c be 00 03  #Ø$t<oNp.x=v2Æ#Ø$t<æ..
000672 9f 50 09 74 5b be 00 04 d3 9f 50 03 02 08 25 ae 23 af 24 74 3d 6f  .P.t[æ..".P...%Æ#Ø$t=o
000688 4e 70 04 78 3d 76 30 ae 23 af 24 74 3e 6f 4e 70 04 78 3d 76 2f ae  Np.x=v0Æ#Ø$t>oNp.x=v/Æ
00069e 23 af 24 74 3f 6f 4e 70 04 78 3d 76 2d ae 23 af 24 74 40 6f 4e 70  #Ø$t?oNp.x=v-Æ#Ø$t@oNp
0006b4 04 78 3d 76 2c ae 23 af 24 74 41 6f 4e 70 04 78 3d 76 2b ae 23 af  .x=v,Æ#Ø$tAoNp.x=v+Æ#Ø
0006ca 24 74 42 6f 4e 70 04 78 3d 76 29 ae 23 af 24 74 43 6f 4e 70 04 78  $tBoNp.x=v)Æ#Ø$tCoNp.x
0006e0 3d 76 27 ae 23 af 24 74 44 6f 4e 70 04 78 3d 76 26 ae 23 af 24 74  =v'Æ#Ø$tDoNp.x=v&Æ#Ø$t
0006f6 45 6f 4e 70 04 78 3d 76 24 ae 23 af 24 74 46 6f 4e 70 04 78 3d 76  EoNp.x=v$Æ#Ø$tFoNp.x=v
00070c 22 ae 23 af 24 74 47 6f 4e 70 04 78 3d 76 21 ae 23 af 24 74 48 6f  "Æ#Ø$tGoNp.x=v!Æ#Ø$tHo
000722 4e 70 04 78 3d 76 1f ae 23 af 24 74 49 6f 4e 70 04 78 3d 76 1d ae  Np.x=v.Æ#Ø$tIoNp.x=v.Æ
000738 23 af 24 74 4a 6f 4e 70 04 78 3d 76 1c ae 23 af 24 74 4b 6f 4e 70  #Ø$tJoNp.x=v.Æ#Ø$tKoNp
00074e 04 78 3d 76 1a ae 23 af 24 74 4c 6f 4e 70 04 78 3d 76 18 ae 23 af  .x=v.Æ#Ø$tLoNp.x=v.Æ#Ø
000764 24 74 4d 6f 4e 70 04 78 3d 76 16 ae 23 af 24 74 4e 6f 4e 70 04 78  $tMoNp.x=v.Æ#Ø$tNoNp.x
00077a 3d 76 15 ae 23 af 24 74 4f 6f 4e 70 04 78 3d 76 13 ae 23 af 24 74  =v.Æ#Ø$tOoNp.x=v.Æ#Ø$t
000790 50 6f 4e 70 04 78 3d 76 11 ae 23 af 24 74 51 6f 4e 70 04 78 3d 76  PoNp.x=v.Æ#Ø$tQoNp.x=v
0007a6 10 ae 23 af 24 74 52 6f 4e 70 04 78 3d 76 0e ae 23 af 24 74 53 6f  .Æ#Ø$tRoNp.x=v.Æ#Ø$tSo
0007bc 4e 70 04 78 3d 76 0c ae 23 af 24 74 54 6f 4e 70 04 78 3d 76 0b ae  Np.x=v.Æ#Ø$tToNp.x=v.Æ
0007d2 23 af 24 74 55 6f 4e 70 04 78 3d 76 09 ae 23 af 24 74 56 6f 4e 70  #Ø$tUoNp.x=v.Æ#Ø$tVoNp
0007e8 04 78 3d 76 07 ae 23 af 24 74 57 6f 4e 70 04 78 3d 76 05 ae 23 af  .x=v.Æ#Ø$tWoNp.x=v.Æ#Ø
0007fe 24 74 58 6f 4e 70 04 78 3d 76 03 ae 23 af 24 74 59 6f 4e 70 04 78  $tXoNp.x=v.Æ#Ø$tYoNp.x
000814 3d 76 02 ae 23 af 24 74 5a 6f 4e 70 04 78 3d 76 00 78 3d e6 75 f0  =v.Æ#Ø$tZoNp.x=v.x=Êu
```


Continued over page

article like this to give the source code. However the object code is shown. The source code program with annotations is obtainable from *Electronics World*.

To give some idea of the complexity of the program we present a flow chart of the program in **Fig. 7**. Subroutine SHLIT checks whether the internal resistors are of the right values and if not changes them. Resetting and initialisation of the LCD is carried out by LCD-INIT and HOD writes the alphabetic messages to the screen.

Code headed TED-HAZ sends the frequency to the LCD display while PHA-HAZ sends the phase to the LCD display. The phase angle is calculated by the routing COST, which then sends it to the LCD display. Finally, VOLT1 calculates the voltage and then sends it to the LCD display.

There is a reset switch when one needs to take new measurements. In view of the frequencies involved the lay out is not important but as there is a considerable amount of components to be included, it will need to be built on stacked circuit boards. There should be adequate spacing between them for good circulation of air and good heat shielding needs to be used for transistors and similar components carrying heavy current. Finally one should make sure that the cable used to sample the mains current is of adequate thickness.

It was difficult to test the instrument because of the high quality of our electricity supply. We artificially altered the power factor by using large values of inductance and capacitance. However these were done only briefly in order not to annoy the electricity company.

This instrument should be useful in any situation where voltage, frequency and power factor needs to be monitored. It could also be adapted to control power factor by switching in banks of capacitance to move the power factor closer to 1. ■


```
00082a 64 84 ff 7e 00 78 3e a6 07 ef 24 30 78 3e f6 78 3e 86 22 12 00 39
000840 75 22 2e 12 00 39 78 3d e6 75 f0 64 84 af f0 7e 00 78 3d a6 07 ef
000856 75 f0 0a 84 fd 7c 00 78 3e a6 05 ed 24 30 78 3e f6 78 3e 86 22 12
00086c 00 39 78 3d e6 75 f0 0a 84 af f0 7e 00 78 3e a6 07 ef 24 30 78 3e
000882 f6 78 3e 86 22 12 00 39 22 75 89 55 d2 8c d2 8e 74 9b 90 d0 00 f0
000898 74 8b 90 e0 00 f0 12 00 4b 12 01 89 75 a8 01 c2 94 12 01 f7 e5 1f
0008ae 75 f0 10 a4 ff ae f0 78 32 e6 75 f0 10 84 fd 7c 00 ef 2d fd ee 3c
0008c4 fc 74 01 8c 06 8d 07 12 0e d5 78 25 a6 06 08 a6 07 7e 01 7f 00 78
0008da 2f e6 12 0e e3 78 2e e6 2f ff 50 01 0e 74 01 12 0e d5 78 34 a6 06
0008f0 08 a6 07 78 25 86 04 08 86 05 ed 2f ff ec 3e fe 74 01 12 0e d5 78
000906 25 a6 06 08 a6 07 78 34 86 04 08 86 05 74 7c d3 9d 74 03 9c 40 02
00091c d2 93 78 34 86 06 08 86 07 74 7c d3 9f 74 03 9e 40 03 02 08 a9 78
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000948 86 07 ef d3 94 1a ee 94 06 40 03 02 08 a9 c2 93 e5 1e b4 01 07 78
00095e 25 86 0a 08 86 0b e5 1e b4 02 07 78 25 86 0c 08 86 0d e5 1e b4 03
000974 07 78 25 86 0e 08 86 0f e5 1e b4 04 07 78 25 86 10 08 86 11 e5 1e
00098a b4 05 07 78 25 86 12 08 86 13 e5 1e b4 06 07 78 25 86 14 08 86 15
0009a0 e5 1e b4 07 07 78 25 86 16 08 86 17 e5 1e b4 08 07 78 25 86 18 08
0009b6 86 19 e5 1e b4 09 07 78 25 86 1a 08 86 1b e5 1e b4 0a 07 78 25 86
0009cc 1c 08 86 1d ae 0a af 0b ac 0c ad 0d ef 2d fd ee 3c fc aa 0e ab 0f
0009e2 ed 2b fb ec 3a fa ac 10 ad 11 eb 2d fd ea 3c fc aa 12 ab 13 ed 2b
0009f8 fb ec 3a fa ac 14 ad 15 eb 2d fd ea 3c fc aa 16 ab 17 ed 2b fb ec
000a0e 3a fa ac 18 ad 19 eb 2d fd ea 3c fc aa 1a ab 1b ed 2b fb ec 3a fa
000a24 ac 1c ad 1d eb 2d fd ea 3c fc 78 25 a6 04 08 a6 05 74 0f 2d fd 50
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000a66 25 86 06 08 86 07 12 0d a3 78 29 a6 06 08 a6 07 78 29 86 20 08 86
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000a92 05 12 0f 06 78 27 a6 06 08 a6 07 74 0a 12 0e e3 78 27 a6 06 08 a6
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000ad4 03 78 27 86 06 08 86 07 8b 04 8a 05 12 0f 06 78 29 a6 06 08 a6 07
000aea 74 0a 12 0e e3 78 29 a6 06 08 a6 07 8f 04 8e 05 78 25 86 06 08 86
000b00 07 cd cc cd 12 0d a3 78 29 a6 06 08 a6 07 74 06 be 00 04 d3 9f 50
000b16 03 78 2b 06 78 29 a6 20 08 a6 21 78 2b e6 d3 94 09 40 09 78 29 12
000b2c 0f 0e 78 2b 76 00 78 2b e6 24 30 78 2b f6 78 29 86 23 08 86 24 12
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000c8c fe 78 27 86 04 08 86 05 ef 2d fd ee 3c fc 8c 20 8d 21 74 64 ff 12
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000dd6 ef 25 e0 ff ee 33 fe 0a c5 f0 23 c5 f0 30 e7 f0 ea 54 07 f9 09 6a
000dec 7a 00 7b 00 60 01 08 c3 ec 9e 40 16 70 04 ed 9f 40 10 ed 9f fd ec
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000e44 8f f0 84 fd ee c4 fe 54 f0 2d ff ee 54 0f 3c fe ad f0 7c 00 22 79
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000ede cf d5 e0 f6 22 f8 8f f0 a4 ff e5 f0 ce 88 f0 a4 2e fe 22 ee c0 e0
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000f62 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00
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Software availability

If you want the object code in electronic form, e-mail jackie.lowe@rbi.co.uk and it will be forwarded to you as text embedded in an e-mail. It will not be sent as an attachment to avoid distributing viruses. If you want the object code on disk, again free of charge, send us a PC formatted 3.5in floppy with a protective self-addressed envelope and enough stamps to allow us to return the disk to you. Post it to the Quadrant House address below.

For those of you wanting to exploit the design further, the source code is available on disk, or as an e-mail attachment, for £25 fully inclusive. E-mail the above address with your credit card number, expiry date and card-holder address or fax 0181 652 8555 with the details. Alternatively, send your order to Power factor soft., Electronics World Editorial, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.

Joe Carr presents six tips to help you get more out of your RF signal generator.

Get more from your RF generator

Signal generators come in a number of different flavours, capabilities and quality levels. Some are used for troubleshooting equipment and circuits, while others are used for making more precise measurements. What they all have in common is that they produce some sort of controlled output signal.

Unfortunately, the output signal is not always clean. Although the purity of the output signal is one of the things that differentiates lower quality and higher quality generators, they all produce signals other than the one desired.

Figure 1 shows a typical spectrum output. This display is what might be seen on a spectrum analyser. The main signal is a continuous sine wave, so ideally you would expect only one single spike, with a height proportional to the output level. But there's a lot of other signals in there.

First, note that the main signal is spread out by phase noise. This noise is random variation around the main frequency. When integrated over a specified bandwidth, e.g. 300 to 3000Hz, the phase noise is called residual FM.

Second, there are harmonics present. If the main signal has a frequency of F, the harmonics have frequencies of nF, where n is an integer. For example, the second harmonic is $2F$, and the third harmonic is $3F$. In many cases, the third harmonic is stronger than the second, but generally the higher harmonics are weaker than lower harmonics.

There are also sometimes sub-harmonics. These are integer quotients of the main signal. Again, if the F is the main signal frequency, $nF/2$ represents the sub-harmonics. Typically, unless something is interfering with the output signal, sub-harmonics are not as prominent. One thing that does make sub-harmonics prominent, however, is the use of frequency multiplier or divider stages – which is the case in many modern generators.

Finally, there are miscellaneous spurious signals, or

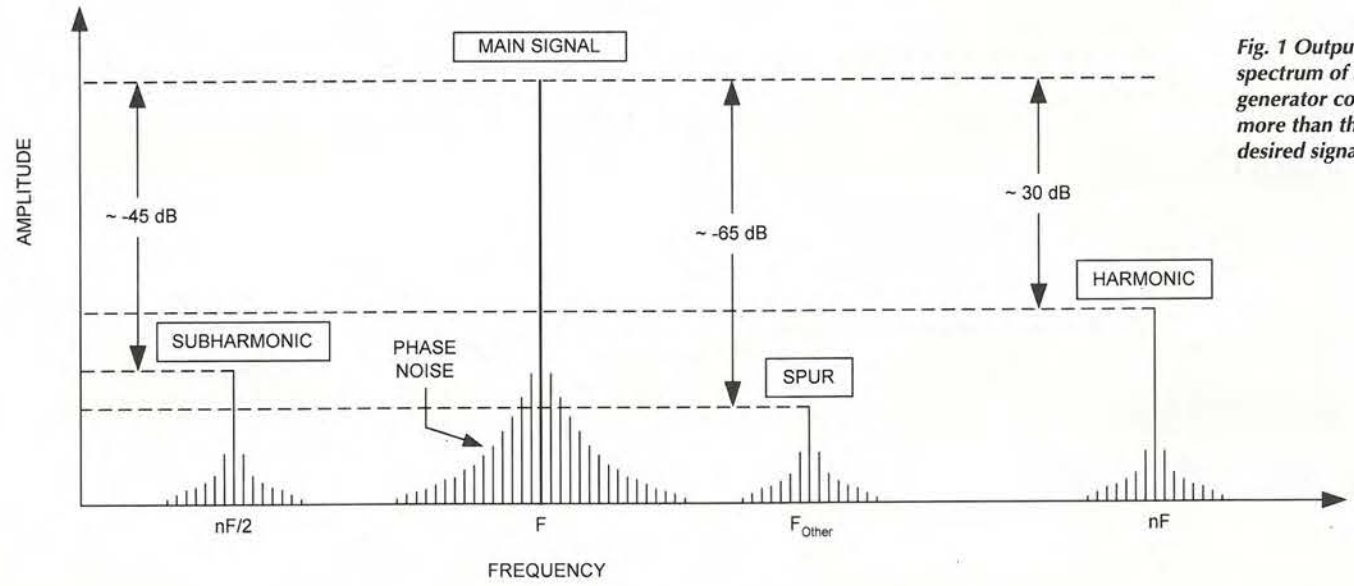


Fig. 1 Output spectrum of a signal generator contains more than the desired signal.

'spurs', found on some generators. These might be due to power supply ripple modulating the output signal, parasitic oscillations, digital noise from counter or phase-locked-loop circuits, and other sources.

Method 1: Improving spectral purity

Certain high quality measurements are very sensitive to extraneous signals coming out of a signal generator. Many signal generators put out harmonics that are –30dB down from the main signal, i.e. the carrier, while other signals may be either higher or lower than this level. The way to get rid of these extraneous signals is to place a frequency-selective filter between the output of the signal generator and the device under test.

Figure 2 shows the use of a low-pass filter to eliminate the harmonics and any spurs that are above the main signal. Select a filter with a –3dB point somewhere between the main signal and the first extra signal, and an attenuation slope enough to reduce the 'bad' signals as much as possible.

If there are any sub-harmonics – or spurs lower than the main signal – then either use a band-pass filter or add a high-pass filter with a –3dB cut-off between the main signal and the sub-harmonic.

There is a cautionary note, however. Real filters do not have the nice flat response seen in some textbooks. They will have pass-band ripple, and some odd responses out of band. Also, *LC* filters are notoriously unpredictable when you terminate them in an impedance other than the design impedance.

Understand the pass-band response and the insertion loss of the filter before using it.

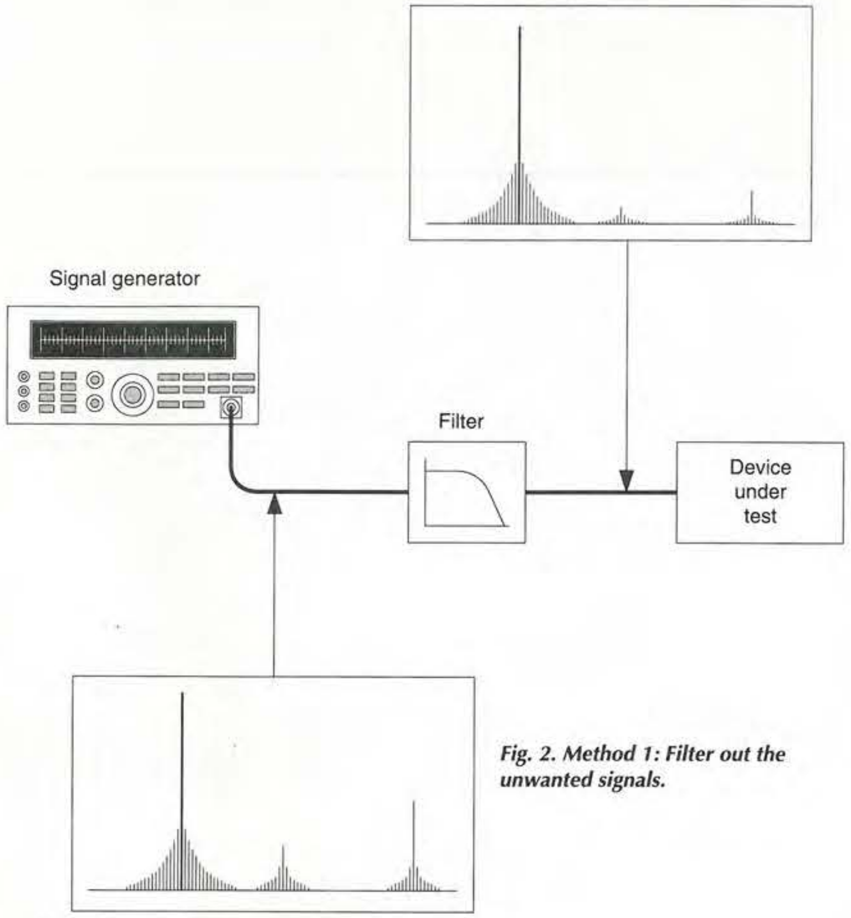


Fig. 2. Method 1: Filter out the unwanted signals.

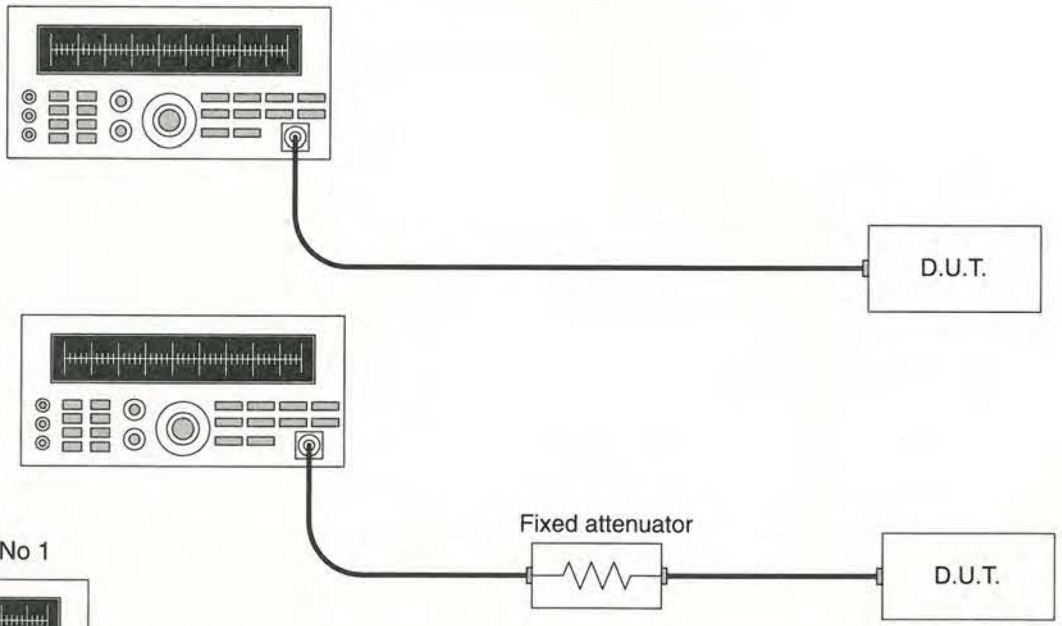


Fig. 3 . Fixing mismatch loss and error can be done by the simple expedient of adding an attenuator in the line.

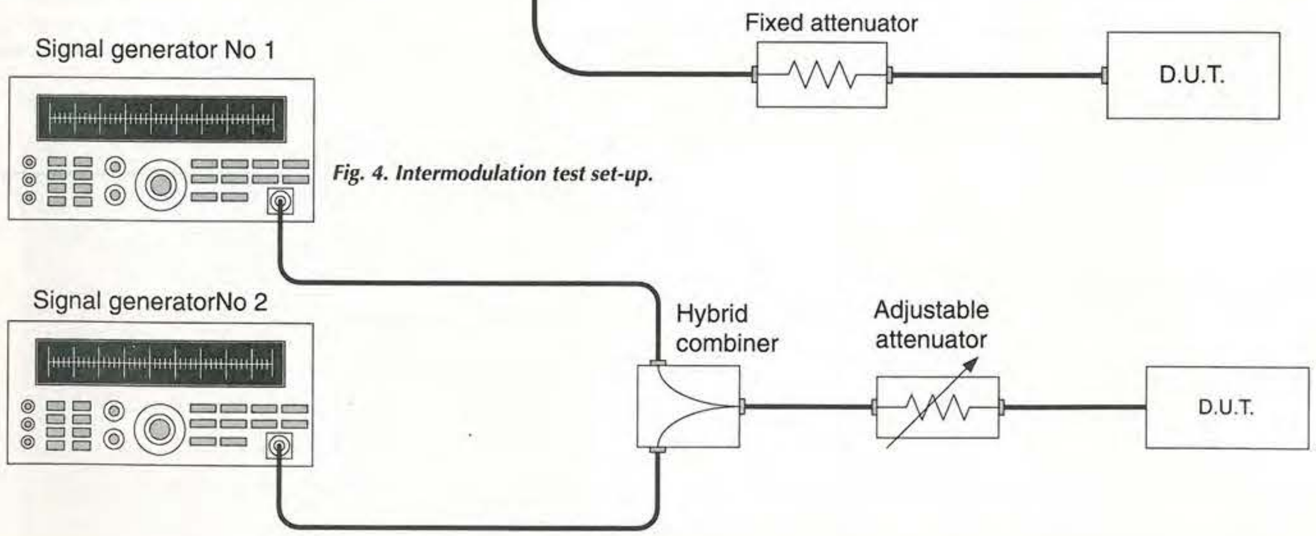


Fig. 4. Intermodulation test set-up.

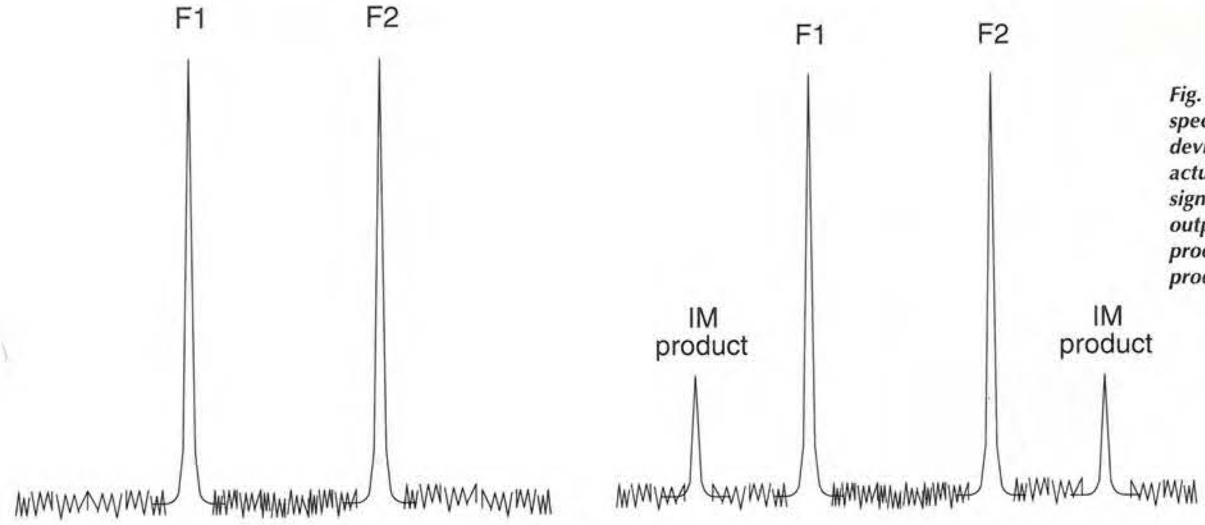


Fig. 5a). Desired spectrum applied to device being tested; b) actual spectrum if the signal generator outputs interact to produce a third-order product.

Method 2: Improving mismatch loss

Mismatch error occurs because the load and the signal generator are not impedance matched. In any electronic circuit, the maximum power transfer occurs when the impedances are matched. There may be an inherent mismatch problem in either the signal generator or the load, and almost certainly in the cables or other devices connected in line with the signal generator.

For example, assume a signal generator with a VSWR of 1.9:1, and a device under test with a VSWR of 1.6:1 connected in the normal way, **Fig. 3a**. The mismatch loss can be found once we know the reflection coefficients: source,

$$\rho_s = \frac{SWR-1}{SWR+1} = \frac{1.9-1}{1.9+1} = \frac{0.9}{2.9} = 0.31$$

device under test,

$$\rho_s = \frac{SWR-1}{SWR+1} = \frac{1.6-1}{1.6+1} = \frac{0.6}{2.6} = 0.23$$

and mismatch loss,

$$\begin{aligned} Loss_{mismatch} &= 20\log(1+\rho_S\rho_D) \\ &= 20\log[1+(0.31\times0.23)] \\ &= 20\log 1.07 \\ &= 20\times0.03 = 0.6\text{dB} \end{aligned}$$

Figure **3b)** shows how to deal with this problem. Insert a 10dB fixed attenuator in line with the line between the signal generator output and the device under test. You will have to adjust the signal generator output level control 10dB higher than normal to compensate for the extra attenuation.

The reason this works is that fixed resistive attenuators tend to be designed with very low reflection coefficients. Suppose we have the same components in Fig. 3b) as in Fig. 3a), but add an attenuator with ρ_A=0.31. The mismatch loss becomes,

$$\begin{aligned} Loss_{mismatch} &= 20\log[1+\{\rho_S\rho_D(\rho_A^2)\}] \\ &= 20\log[1+(0.31\times0.23\times0.31^2)] \\ &= 20\log(1+0.0069) \\ &= 20\times0.003 = 0.06\text{dB} \end{aligned}$$

Method 3: Improving third-order intercept performance

One of the most important specifications for an amplifier or radio receiver is the *third-order intercept point*, also referred to as TOIP or IP3. This specification tells you something about the device's dynamic performance – especially in the presence of multiple input signals.

If you listen to any shortwave receiver, AM or FM BCB receiver, or any scanner receiver you will realise that most

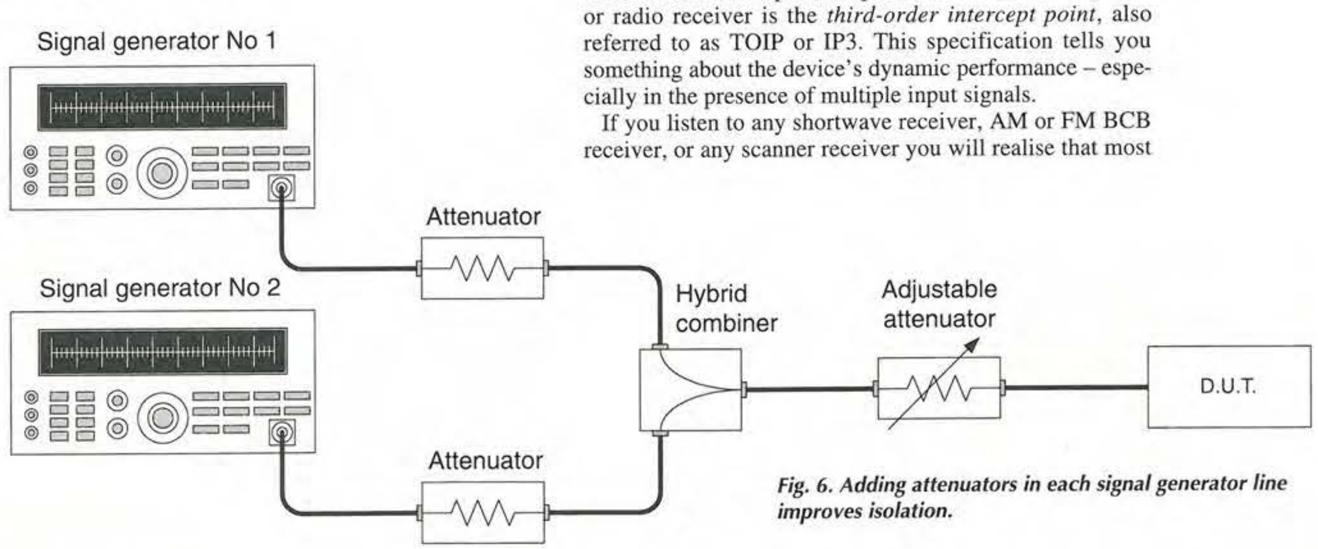


Fig. 6. Adding attenuators in each signal generator line improves isolation.

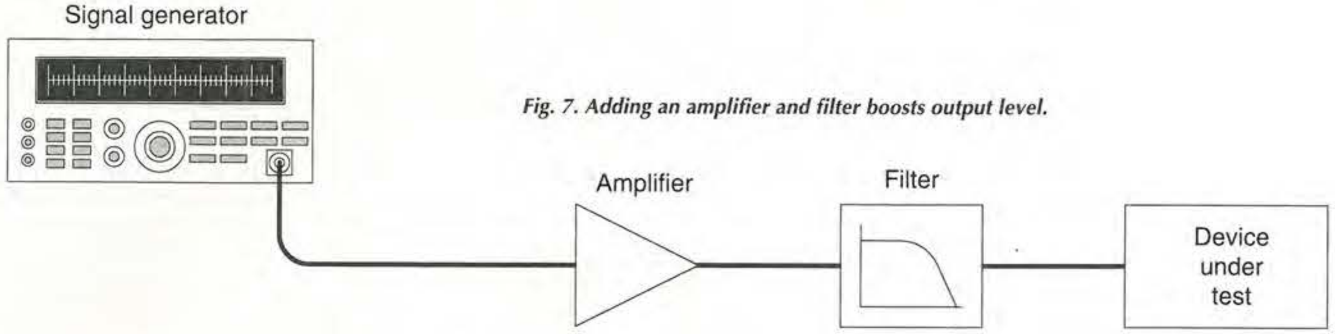


Fig. 7. Adding an amplifier and filter boosts output level.

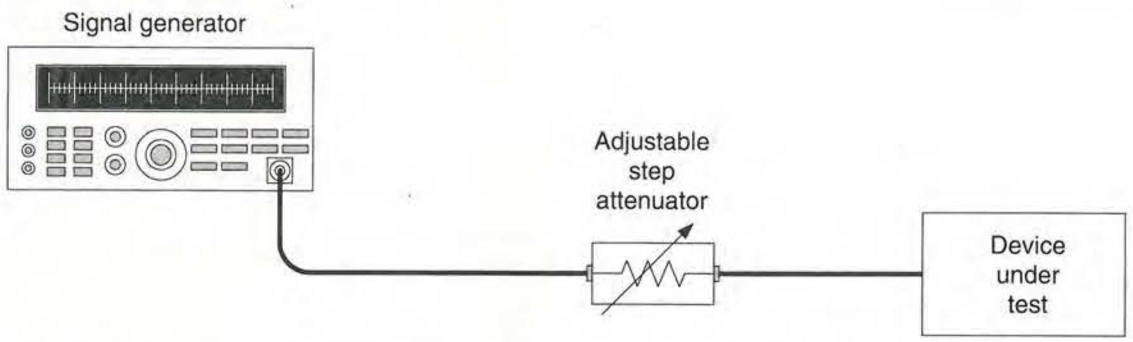


Fig. 8. Adding a calibrated attenuator allows lower signal levels to be accommodated.

areas of the country are polluted with too many radio signals. When multiple strong signals are received at the same time, receiver or amplifier non-linearity occurs, and heterodyne products are created.

If F_1 and F_2 are two input signals – one of which might be the desired signal – these 'intermodulation' products will have frequencies equal to $mF_1 \pm nF_2$, where m and n are integers. The third-order harmonics are those in which m=2 and n=1, or m=1 and n=2, i.e. m+n=3, and these are the most difficult to handle.

The worst case is usually the $2F_1-F_2$ and $2F_2-F_1$ third-order products. This is because they will fall close to F_1 and F_2 and may be within the device pass-band. A problem with these products is that they increase at a rate three times the number of decibels as the fundamental signal. If F_1 or F_2 goes up 1dB, then the third-order products go up 3dB.

Figure 4 shows the basic set-up for measuring the third-order intercept point, and also certain other parameters. The two signal generators produce frequencies F_1 and F_2. They are set to identical output levels, usually quite high such as −10dBm or −20dBm.

Initially, the receiver is tuned to one frequency, F_1 for example, and a reference level established equal to the minimum discernible signal, or in some procedures an S1 signal level. The receiver is then tuned to the third-order product frequencies, and the attenuator decreased, raising signal level until the same reference level is produced. The IP3 can be calculated from these data points.

But look what happens if the signal generator is not working quite the way we hoped, **Fig. 5**. The spectrum of Fig. **5a)** is what we hope to see. Frequencies F_1 and F_2 standing up smartly above the noise level – which is hopefully quite low.

Figure **5b)** shows the same spectrum with the third-order intermodulation products present. When you see this at the output of a amplifier or receiver being tested, then you might assume that the intermodulation products are generated in the device under test. But not always. Sometimes, the signal from one generator gets into the output stages of the other generator, and causes an IM response that is due solely to the test set-up!

A couple things can be done to prevent the problem. First, if the combiner used to merge the signals into one line is a resistive star type circuit, then there is only 6dB of isolation between the ports. Using a hybrid combiner with a larger amount of port-to-port isolation helps tremendously because it reduces the signal reaching the other generator's output stages.

Another fix is to insert 10dB or 20dB fixed attenuators in each signal line, **Fig. 6**. These attenuators provide an additional amount of isolation between the signal generators. Of course, you will have to adjust the output levels of the signal generators to overcome the extra loss.

A cautionary note is in order: be certain that the signal generator output can be cranked up to a higher level without producing spurious output signals, harmonics and other extraneous signals. One of my own signal generators works well from 0 to 90 percent of full output, but at output levels greater than 90 percent the spectrum blossoms with unwanted signals. Ouch!

Method 4: Extending upper output range

Signal generator output controls are calibrated in terms of output voltage, usually microvolts or millivolts, or the power level, for example dBm, i.e. decibels relative to 1mW into 50Ω.

A typical generator produces output levels up to some value from about 0dBm or perhaps +20dBm, or some value in between. But what do you do if the signal generator maximum is, say, +10dBm, and you need a signal level of +30dBm (1W)? Or, how about the case where you have a signal generator like mine that is wonderful at lower levels, but falls apart at higher levels?

The solution is simple and obvious: amplify the output. But there are some cautions. **Figure 7** shows the use of an external amplifier to boost the output level of the signal generator. Because all amplifiers can become nonlinear, and produce a bit of harmonic distortion in their own right, a low-pass filter is inserted in the path between the amplifier output and the device under test.

You must also make sure that the selected amplifier can do the job. Make sure the IP3 specification of the amplifier is sufficiently high that the signal generator cannot overdrive it. The maximum input drive level – usually specified in dBm – and the output power level – also expressed in dBm or possibly watts – must be sufficient to handle the job. Otherwise, adding the amplifier might add problems.

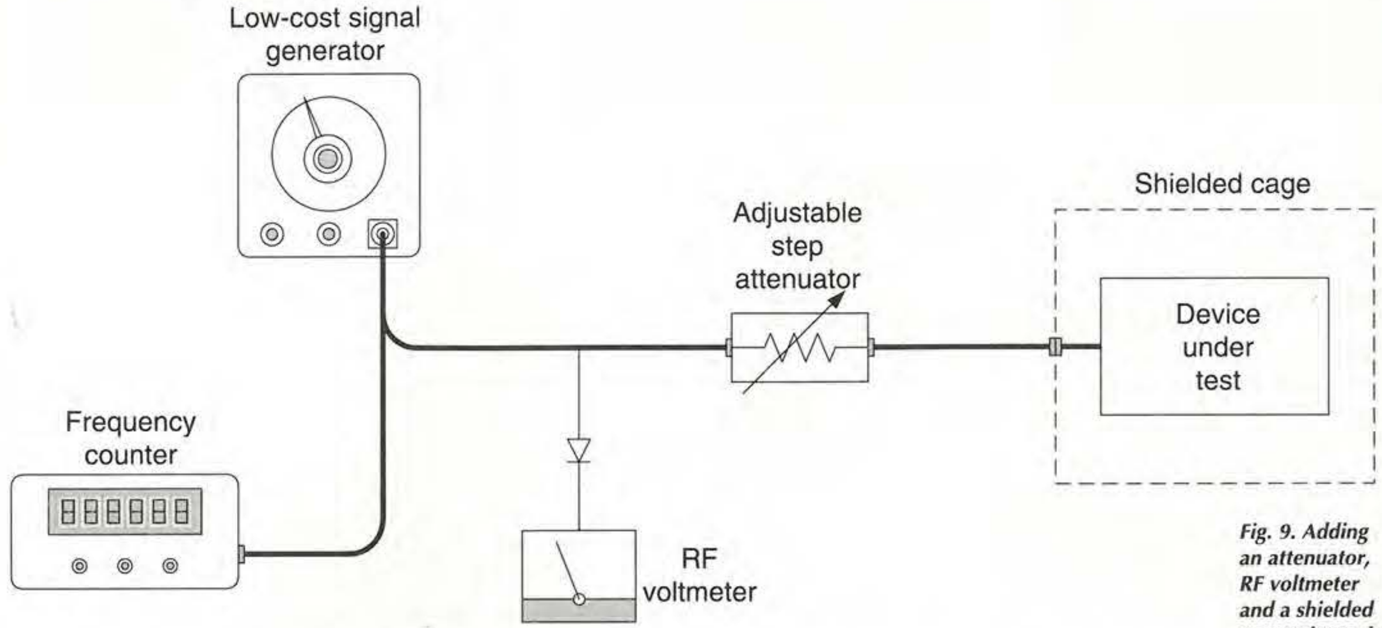


Fig. 9. Adding an attenuator, RF voltmeter and a shielded cage solves a lot of problems with cheapie generators.

Method 5: Reducing the output level

There are times when you might want to reduce the signal generator output level. One reason for doing this is that you need a very small signal at the device under test, but need a higher signal to act as a reference or be fed to a frequency counter.

I sometimes do this when using one of my elderly analogue signal generators. It has an inaccurate analogue frequency dial, but the output attenuator is well calibrated. To get a higher level for the counter, while providing a low level signal to the amplifier or receiver being tested, I use a set-up like **Fig. 8**. In other cases, you will simply need a lower signal level than the generator can provide. Figure 8 works for that purpose as well.

The attenuator should be a calibrated type. You can obtain continuously variable calibrated attenuator. These are costly, but some tend to come on to the surplus market.

A lower cost alternative is to use a precision step attenuator. These devices have switch selectable attenuation levels in various steps. The total attenuation is the sum of all the individual attenuations. You can build step attenuators, but for precision work you are well advised to buy one. The resistors for precision attenuation levels are really odd values, although they can be approximated.

Method 6: Using a cheapie signal generator

Many of the tests and measurements done with signal generators can only be done with rather expensive instruments. Unfortunately, a large number of people are forced to use low-cost 'service shop' grade signal generators. And these can be a problem.

Of course, if you are lucky enough to find surplus 'laboratory-grade' or military signal generators, then you are way ahead of the game. Good quality signal generators are easily found on the surplus market. I've seen high quality units go wanting for a buyer at hamfests, even though the asking price was quite reasonable. Indeed, a friend of mine use to make quite a living buying hamfest specials, reconditioning them, and re-selling them.

But what do you do if the only signal generator you can use is a service-grade instrument? Indeed, what's wrong with service grade instruments?

In answer to the second question, there are a couple of problems. First, the frequency is not well calibrated, although that can be overcome with a frequency counter. Second, the output level is not well calibrated. Indeed, there might be a single knob output control that has no markings at all, or 'relative' 0-9 level markings.

Finally – and here is the biggee – the output level control might be irrelevant for low-level measurements because more signal will escape around the cabinet flanges than goes through the output connector! Making a cabinet RF-tight is a pretty hard trick, and cheapie signal generators simply don't do it very well. Of course, you could break the generator open and try shielding everything in sight.

Figure 9 shows a method for overcoming these problems. Like the previous method, this one uses a calibrated step-attenuator to set the output level. In order to make the output level mean something an RF voltmeter – or AF voltmeter if an audio generator is being used – must be connected to allow the output level to be set to a specified point.

In order to make the device under test able to handle the low signal level without being swamped by signal that escapes around the signal generator cabinet flanges a metal shield cage is used to hold the device under test.

The cage can be made of metal window screen, or perforated aluminum sheet metal available from hardware stores. As long as the box is 'RF tight,' i.e. it works as a reasonably decent Faraday cage, the device under test will only see signal coming through the coaxial cable, and not radiated through the air.

In summary

I hope that the methods discussed in this article will make it possible for you to do a lot more with your signal generator than is presently the case. Even if your signal generator is a low-cost service grade type, these methods will help. ■

Bench testing for EMC

Ian Darney attempts to bring EMC down to Earth using a simple model structure that is easy to verify on the bench.

Confusion and frustration are the normal by-products of any attempt to glean a simple set of circuit guidelines from the literature on electromagnetic compatibility. On the practical side, a plethora of design tips and stratagems can be found, many of which conflict with other design considerations.

Delving into the subject of electromagnetic theory reveals a morass of ever more complex mathematics. A simple method of applying the design process to interference problems is nowhere to be found.

The purpose of this article is to introduce such a method. The approach is based on the idea of creating an accurate circuit model of the signal coupling between cable and structure, or between conductors routed over the structure. The model can then be used to analyse the interference signals, whatever form they may take.

Conclusions of such an analysis can only be assessed in the light of practical results, and bench testing is the simplest way of obtaining such data. So, the design and use of such equipment are described too.

Confidence in the validity of the approach is established by building a test rig, measuring the coupling characteristics between a twin conductor and structure, and comparing analytical results with test data.

The key features of my approach are the utilisation of general-purpose software to analyse the results, and the simplicity and ease of use of the test equipment. This reduces EMC from a black art to a design problem.

Background

One of the basic assumptions of circuit theory is that the voltage across any two-terminal device is a function only of the current in that device. This is in direct contradiction to the concepts of electromagnetic theory, which lead to the conclusion that the voltage of any conductor is a function of the current in all conductors.

In fact, this simplifying assumption is one of the reasons why circuit theory is the more useful. By avoiding inconsequential detail, it allows the bigger picture to become visible. Circuit diagrams enable us to visualise what is happening to the various signals in a system.

Since circuit components and electromagnetic parameters are entirely different entities, any combination can only result in confusion. Maintaining strict separation can be achieved by defining components of circuit models as being of type 'circuit', just as some numbers can be defined as being of type 'integer'.

Equations can be derived to relate the electromagnetic field round a set of conductors to the currents and voltages in those conductors. Since these are the simplest relationships possible, the parameters can be described as 'primitives'; a term borrowed from the ideas of operational systems. Primitive equations describe the behaviour of the assembly as an antenna.

In electronic systems, signals are routed from one element of the system to another on interconnecting cables that act as multi-conductor transmission lines. The most convenient way of analysing the coupling between such signals is to define a set of loops, each loop linking a pair of conductors. Parameters involved here can be described as 'loop' impedances.

To distinguish between 'primitive', 'loop', and 'circuit'[1], parameters, an extra letter is added to the appropriate symbol, e.g. L_p, L_l, and L_c.

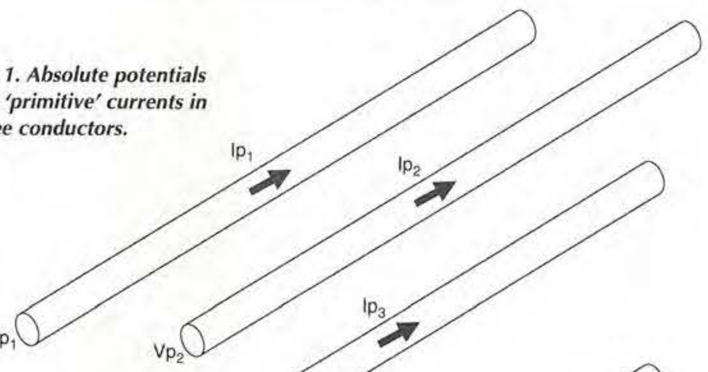


Fig. 1. Absolute potentials and 'primitive' currents in three conductors.

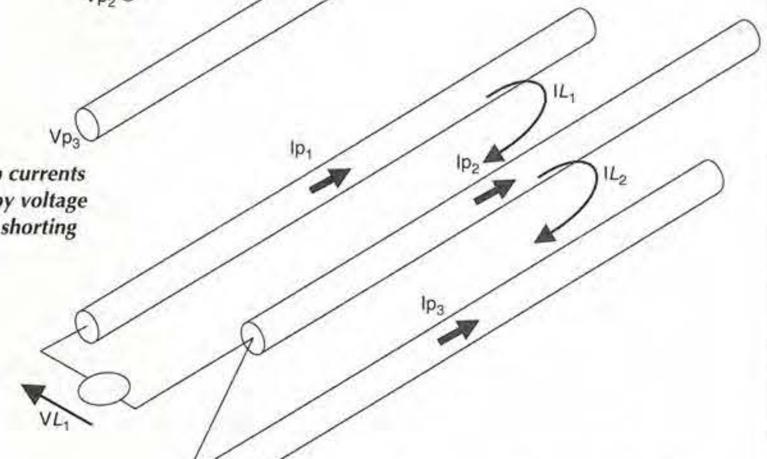


Fig. 2. Loop currents generated by voltage source and shorting link.

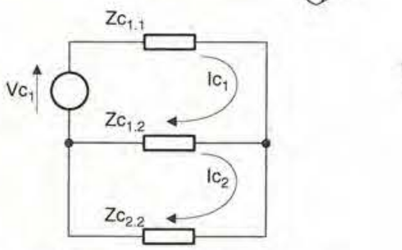


Fig. 3. This circuit gives a set of equations similar to the loop equations.

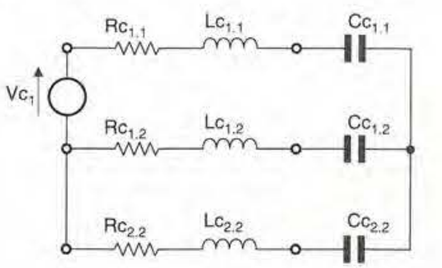


Fig. 4. Initial circuit model, identifying individual components.

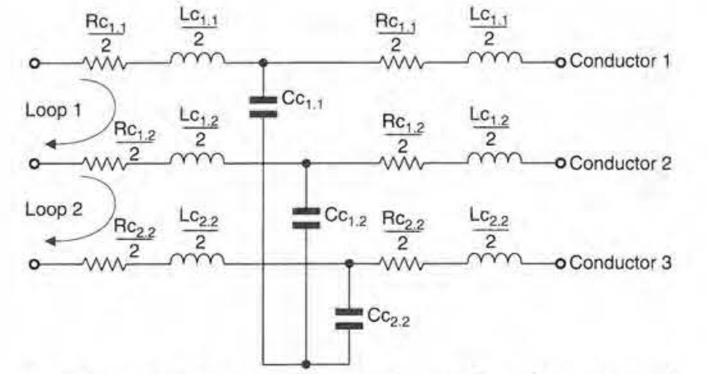


Fig. 5. The model is improved by converting each branch to a T-network.

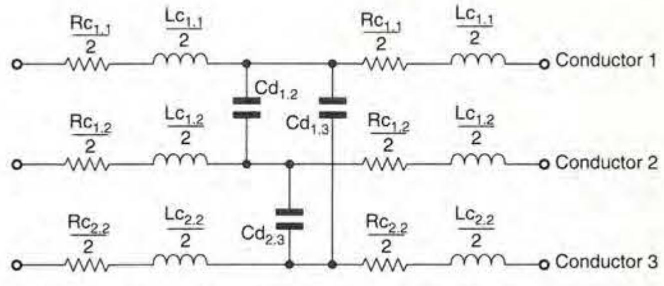
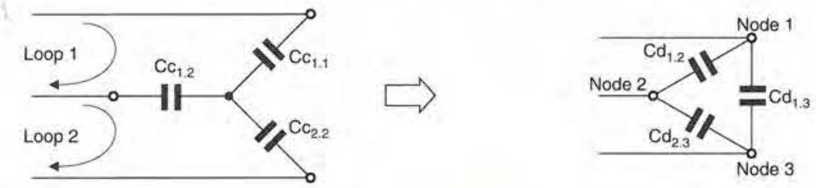


Fig. 7. General circuit model of three conductor lines.



$$Cd_{1,2} = \frac{Cc_{1,1} \cdot Cc_{1,2}}{Cc_{1,1} + Cc_{1,2} + Cc_{2,2}}$$

$$Cd_{1,3} = \frac{Cc_{1,1} \cdot Cc_{2,2}}{Cc_{1,1} + Cc_{1,2} + Cc_{2,2}}$$

$$Cd_{2,3} = \frac{Cc_{1,2} \cdot Cc_{2,2}}{Cc_{1,1} + Cc_{1,2} + Cc_{2,2}}$$

Fig. 6. Star-to-delta transformation helps visualise how a capacitor connects two conductors.

Creating a circuit model

By using such identifiers in a systematic way, formulae can be derived for the inductors, capacitors and resistors associated with a three-conductor assembly. The procedure is set out in the panel entitled 'Formulation' and summarised in **Figs 1 to 4**. Note that the process does not involve anything more difficult than addition, subtraction, and substitution.

However, the response of the circuit of Fig. 4 deviates from the actual response as the frequency approaches resonance. A much better simulation can be achieved by representing each impedance as a T network, as shown in **Fig. 5**.

This model is perfectly adequate for analysing the performance of the assembly. However, it is more conventional to visualise capacitance as a component that couples two conductors. A star-to-delta transformation as shown on **Fig. 6** takes care of this. For delta components, the component symbol is followed by the letter 'd', and the identification numbers refer to nodes.

This leads to the circuit model of **Fig. 7**. If the physical dimensions of the line are known, equations (8), (10), and (11) can be used to determine component values.

Circuit analysis software

Even though the model is crude in terms of transmission-line theory, it presents a daunting challenge to anyone attempting to analyse its response, armed with only a hand calculator. Fortunately, there is no need to attempt such an exercise. Circuit analysis software[1] is readily available, and the processing power of desktop computers is more than adequate to the task.

With such a facility, all that is necessary is to draw the circuit on the screen and define component values. A selection is made of whether frequency analysis or transient analysis is desired, the test limits are defined, and the signals to be examined are then selected. At the touch of a key, the results are computed and presented on the screen as a smooth curve.

Test gear

At this point, all the analytical tools have been introduced. The second half of the exercise is to design and build some test gear to allow interference coupling parameters to be measured. Such an exercise is described below.

The first step was to build a test rig. This consisted of a length of copper pipe to simulate a structure, and a wire pair to represent the cable under test. A strip of wood laid along the pipe was used to achieve a fixed separation between cable and structure.

The test equipment was about as simple as it could be; an oscilloscope, a signal generator, and a few locally purchased components.

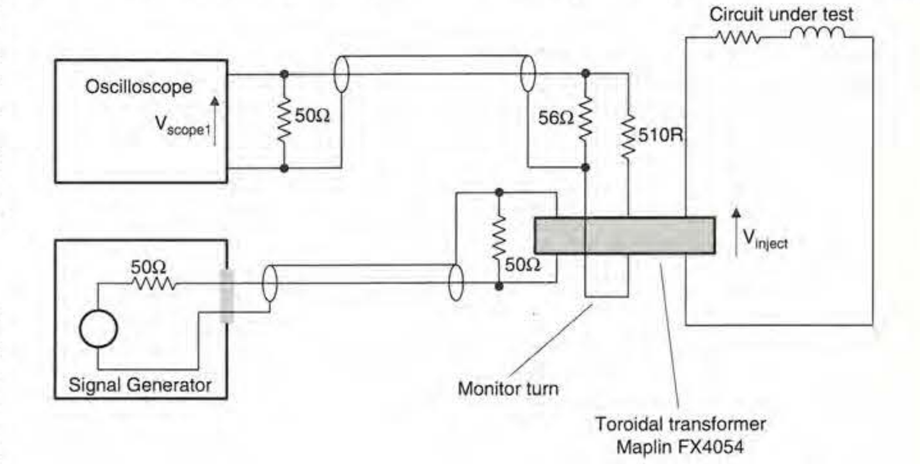
Injection transformer

A small injection transformer was constructed, using ten turns of enamelled copper wire on a toroidal core as the primary winding. This winding was connected to the output of the signal generator via a co-axial cable, **Fig. 8**. A 50Ω resistor was connected in parallel with the primary winding to minimise reflections at the transformer end of the cable.

To monitor the injected voltage, a tightly wound single turn of wire was added round the core, and the loop closed by 510Ω in series with 56Ω. A co-axial cable connected the output across the 56Ω to one input of the oscilloscope. The resistors act as a potential divider, and as characteristic impedance to terminate the cable. Figure 8 also defines the relationship between transformer output voltage and the oscilloscope input.

Unlike the transformers used in EMC test facilities, this transformer lacks the ability to clamp over the cable under test. From the point of view of bench testing, this is no real disadvantage. The cable can be threaded through the transformer core during assembly, and removed afterwards; a few minutes work.

Being small, the transformer is a low power device. Operating at high power levels when investigating interference is an unsociable activity, so low power levels are desirable.



$$V_{inject} = \frac{510 + R_{in}}{R_{in}} \cdot V_{scope\,1} \quad \text{where} \quad R_{in} = \frac{56.50}{50 + 56}$$

Fig. 8. Injection transformer circuitry – part of the kit needed to evaluate the model.

It is also a low-cost assembly, easy to modify to suit the application.

Current transformer

To monitor current in the cables, an identical toroid was used to construct a current transformer. This time the primary was the intended loop-under-test and the secondary was ten turns of enamelled copper wire. A 30Ω resistor was connected across the secondary, allowing any voltage across the resistor to be monitored by the oscilloscope. The 22Ω resistor was used to match the transformer termination to the co-axial line.

Figure 9 illustrates the set up, and defines the relationship between current in the cable and voltage at the oscilloscope.

Loop admittance test

The test equipment was connected to the test rig, as shown on **Fig. 10**. One conductor of the wire pair was connected to structure at both ends to form a loop, and simulate a 'return' conductor with both ends grounded.

The 'signal' conductor was terminated at one end in a short circuit to structure, and at the other end by a load consisting of 510Ω and 56Ω resistors in series. This load's purpose was to allow a subsequent test to measure the common mode rejection of the set-up.

The injection transformer was used to inject a sinusoidal voltage of about a volt peak to peak into the cable/structure loop, to simulate a signal induced by an external source. Such an external source could be electromagnetic radiation causing spurious currents in the structure, or an adjacent cable carrying other signals in the system.

The current transformer was used to monitor the resultant current. Measurements were carried out at a number of spot frequencies, and a record kept of frequency, input voltage, and output current. When resonance was detected, several measurements were taken at and around the peak or the trough in the response.

Analysing the results

Physical data, as recorded in Fig. 10, was used to calculate component values for the circuit model of the set-up, using equations (8), (10) and (11). Then the general circuit model of Fig. 7 was converted to the specific circuit model of **Fig. 11**.

The model was completed by shorting the terminations at the right-hand side, adding the monitor resistors to the left-hand side, including a 1V source to represent the output of the transformer, and adding a 0.2Ω resistor to simulate the load presented by the current transformer.

Using Geseca and Spiceage, as mentioned earlier under 'Circuit analysis software', I calculated the frequency response of the model. In this case, the output selected was the current in the 0.2Ω resistor. This resulted in the solid curve of **Fig. 12**.

Test data was then processed to give admittance values, and the results added to the curve as a set of test points.

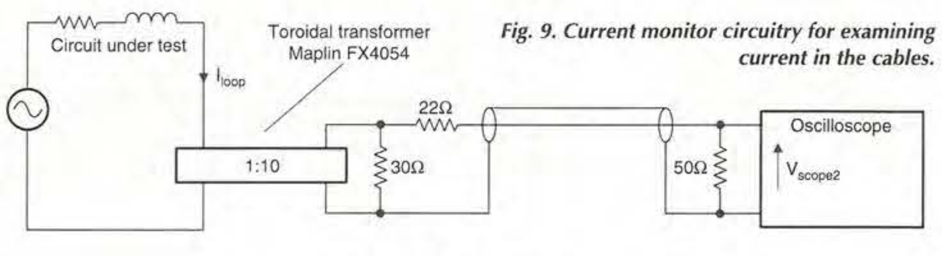


Fig. 9. Current monitor circuitry for examining current in the cables.

$$I_{loop} = 10.\left[\frac{50+22}{30}+1\right].1/_{50}.V_{scope\,2}$$

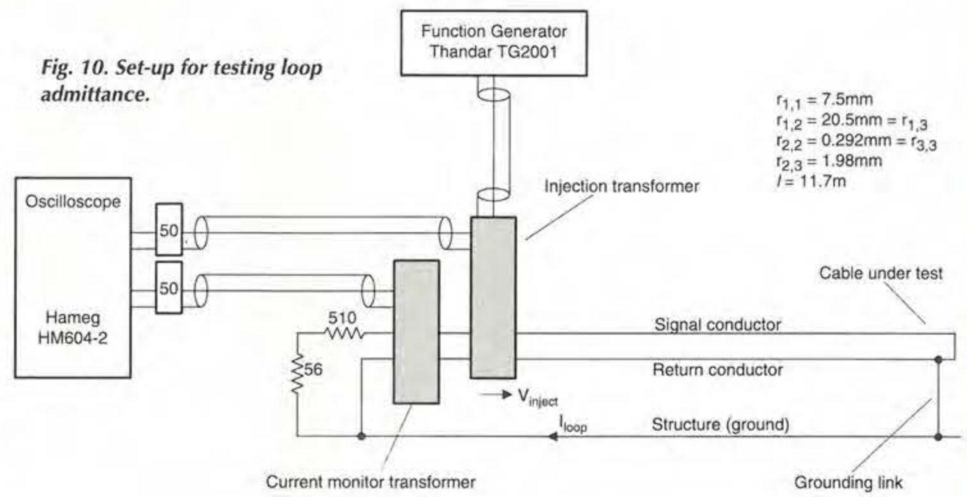


Fig. 10. Set-up for testing loop admittance.

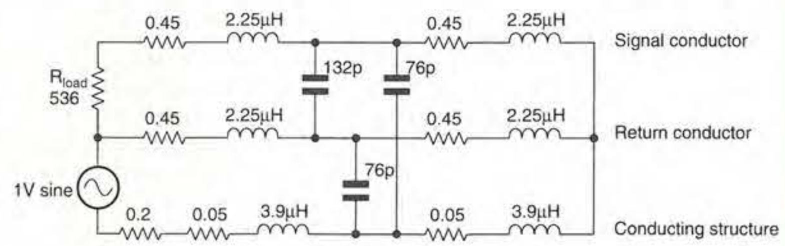


Fig. 11. Circuit model – twin cable – both ends grounded.

Assessing the results

It is immediately obvious that there is a close correlation between test results and those derived from theory. This confirms that the concepts used in the creation of the model, the physical measurements, the functioning of the test equipment, the method of measurement, and the processing of the data, were essentially correct.

The model demonstrates that the cable/structure loop behaves as a resistor, inductor, and transmission line as the frequency increases; i.e. it is applicable over a wide bandwidth.

Since all currents and voltages in the assembly are simulated, any characteristic can be investigated. This includes common-mode rejection and the effect of 'floating' either pair of cable terminals.

Using software to simulate electronic components at either end of the cable, the response to interference can also be predicted. Both frequency and transient analysis are possible.

Using the test equipment, actual response can be compared to the predictions, and the model refined. Both emission and susceptibility characteristics can be determined.

The model can be constructed from either electrical measurements or from physical data; it is not restricted to the simple configuration described here.

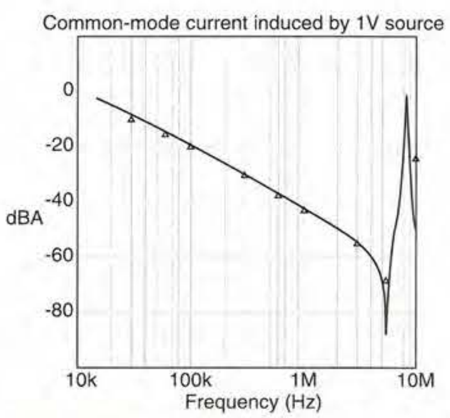


Fig. 12. Loop admittance characteristic – twin cable – both ends grounded. Solid line is theoretical, triangles represent test data.

In summary

I have outlined a method of creating circuit models of electromagnetic coupling in cabling, and explained how to design simple test equipment to measure that coupling.

Combining testing with modelling can solve a wide variety of EMC problems. ■

Reference

1. Geseca for Windows': 'Spiceage for Windows'. (Those Engineers Ltd, Mill Hill, London NW7 4BP).

Working out

Assume that the three isolated conductors of Fig. 1 represent a section of a wiring assembly. If there is current in the conductors, then there is also an electromagnetic field; and *vice versa*. The absolute potential of each conductor is related to the current in all three conductors by the primitive equations,

$$\begin{aligned} V_{p(1)} &= Z_{p(1,1)} \times I_{p(1)} + Z_{p(1,2)} \times I_{p(2)} + Z_{p(1,3)} \times I_{p(3)} \\ V_{p(2)} &= Z_{p(2,1)} \times I_{p(1)} + Z_{p(2,2)} \times I_{p(2)} + Z_{(p2,3)} \times I_{p(3)} \\ V_{p(3)} &= Z_{p(3,1)} \times I_{p(1)} + Z_{p(3,2)} \times I_{p(2)} + Z_{p(3,3)} \times I_{p(3)} \end{aligned} \qquad (1)$$

Here, the parenthesised numbers in the subscripts define the conductors. One feature of primitive impedances is that they are symmetrical; if i and j are used to identify conductors, then,

$$Z_{p(i,j)} = Z_{p(j,i)}$$

If the current is generated by a voltage source between the end terminals of conductors 1 and 2, with a short between conductors 2 and 3, as in Fig. 2, then the loop currents of Fig. 2 can be related to the primitive currents of Fig. 1.

$$\begin{aligned} I_{p(1)} &= I_{l(1)} \\ I_{p(2)} &= I_{l(2)} - I_{l(1)} \\ I_{p(3)} &= -I_{l(2)} \end{aligned} \qquad (2)$$

Relationships between current and voltage of equation 1 remain unaltered, so the primitive currents can be replaced by the loop currents. Loop voltages are the difference in potential between pairs of conductors, so a set of loop equations can be derived,

$$\begin{aligned} V_{l(1)} = V_{p(1)} - V_{p(2)} &= Z_{l(1,1)} \times I_{l1} - Z_{l(1,2)} \times I_{l2} \\ 0 = V_{p(2)} - V_{p(3)} &= Z_{l(2,1)} \times I_{l1} - Z_{l(2,2)} \times I_{l2} \end{aligned} \qquad (3)$$

where,

$$\begin{aligned} Z_{l(1,1)} &= Z_{p(1,1)} - Z_{p(1,2)} - Z_{p(2,1)} + Z_{p(2,2)} \\ Z_{l(1,2)} &= Z_{p(1,2)} - Z_{p(1,3)} - Z_{p(2,2)} + Z_{p(2,3)} = Z_{l(2,1)} \\ Z_{l(2,2)} &= Z_{p(2,2)} - Z_{p(2,3)} - Z_{p(3,2)} + Z_{p(3,3)} \end{aligned} \qquad (4)$$

Because primitive impedances are symmetrical, so are loop impedances. For loop parameters, the numbers in the subscripts refer to loops.

The next action is to create a circuit diagram that contains two loops, with one impedance common to both loops, Fig. 3. This is an exercise in lateral thinking, and is the most important step in the whole procedure.

Circuit equations for this model are,

$$\begin{aligned} V_{c(1)} &= \left(Z_{c(1,1)} + Z_{c(1,2)}\right) \times I_{c(1)} - Z_{c(1,2)} \times I_{c2} \\ 0 &= -Z_{c(1,2)} \times I_{c(1)} + \left(Z_{c(1,2)} + Z_{c(2,2)}\right) \times I_{c2} \end{aligned} \qquad (5)$$

For the model to replicate the action of the twin loop assembly then mesh currents and voltages of equation set 5 must be identical to loop currents and voltages of the loop equations 3. Circuit impedances can be related to loop impedances, and hence to the primitives,

$$\begin{aligned} Z_{c(1,1)} &= Z_{l(1,1)} + Z_{l(1,2)} = Z_{p(1,1)} - Z_{p(1,2)} - Z_{p(1,3)} + Z_{p(2,3)} \\ Z_{c(1,2)} &= -Z_{l(1,2)} = Z_{p(2,2)} - Z_{p(2,3)} - Z_{p(1,2)} + Z_{p(1,3)} \\ Z_{c(2,2)} &= Z_{l(1,2)} + Z_{l(2,2)} = Z_{p(3,3)} - Z_{p(2,3)} - Z_{p(1,3)} + Z_{p(1,2)} \end{aligned} \qquad (6)$$

The inductance of each circuit element can be calculated by substituting, in equation set 6,

$$Z_{p(i,j)} = L_{p(i,j)} = \frac{\mu \times l}{2\pi}\left[\ln\left(\frac{2l}{r_{i,j}}\right) - 1\right] \qquad (7)$$

where,
l is the length of the section,
$r_{i,j}$ is the separation between conductors i and j,
$r_{i,i}$ is the radius of conductor i and
μ is the permeability. This gives,

$$\begin{aligned} L_{c(1,1)} &= \frac{\mu \times l}{2\pi} \ln \frac{r_{1,2} \times r_{1,3}}{r_{1,1} \times r_{2,3}} \\ L_{c(1,2)} &= \frac{\mu \times l}{2\pi} \ln \frac{r_{1,2} \times r_{2,3}}{r_{2,2} \times r_{1,3}} \\ L_{c(2,2)} &= \frac{\mu \times l}{2\pi} \ln \frac{r_{1,3} \times r_{2,3}}{r_{3,3} \times r_{1,2}} \end{aligned} \qquad (8)$$

Capacitors can be calculated by substituting, in equation set 6,

$$Z_{p(i,j)} = \frac{1}{C_{p(i,j)}} = \frac{1}{2\pi \varepsilon l} \times \ln \frac{1}{r_{i,j}} \qquad (9)$$

Here, ε is the permittivity. This gives,

$$\begin{aligned} C_{c(1,1)} &= \frac{2\pi \times \varepsilon \times l}{\ln \frac{r_{1,2} \times r_{1,3}}{r_{1,1} \times r_{2,3}}} \\ C_{c(1,2)} &= \frac{2\pi \times \varepsilon \times l}{\ln \frac{r_{1,2} \times r_{2,3}}{r_{2,2} \times r_{1,3}}} \\ C_{c(2,2)} &= \frac{2\pi \times \varepsilon \times l}{\ln \frac{r_{1,3} \times r_{2,3}}{r_{3,3} \times r_{1,2}}} \end{aligned} \qquad (10)$$

Resistors can be obtained by inspecting equation set 6 and then calculating end-to-end resistance,

$$\begin{aligned} R_{c(1,1)} &= R_{p(1,1)} = \frac{\rho \times l}{\pi\left(r_{1,1}\right)^2} \\ R_{c(1,2)} &= R_{p(2,2)} = \frac{\rho \times l}{\pi\left(r_{2,2}\right)^2} \\ R_{c(2,2)} &= R_{p(3,3)} = \frac{\rho \times l}{\pi\left(r_{3,3}\right)^2} \end{aligned} \qquad (11)$$

Here, ρ is the resistivity of the conductors. If each impedance of Fig. 3 is represented by R, L and C in series, then Fig. 4 is the result.

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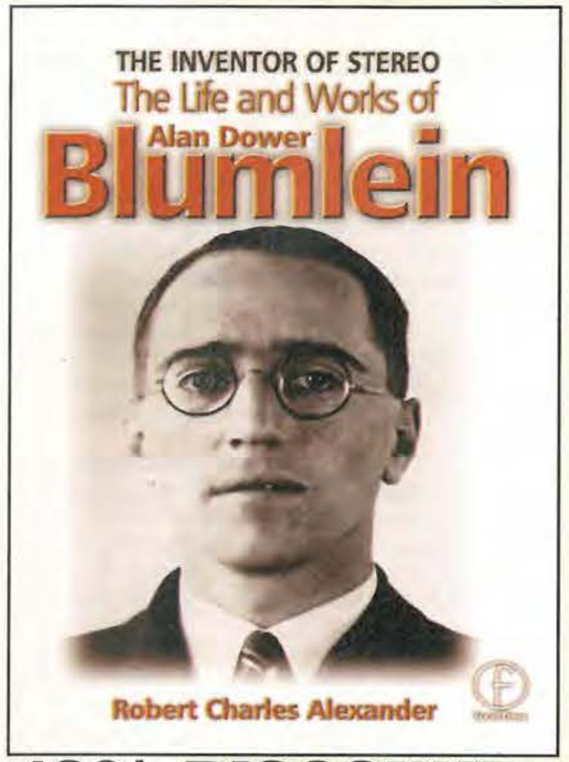

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This book is the definitive study of the life and works of one of Britain's most important inventors who, due to a cruel set of circumstances, has all but been overlooked by history.

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.

Contents

Design details

Calculating equalisation network, box 'A'
The following equations are from reference 7. First choose R_A for the desired input-impedance:

$$R_B = 2kR_A$$

$$R_C = \left(\frac{\omega_C}{\omega_E}\right)^2 R_A$$

$$C_A = \frac{2Q_{TC}(1+k)}{R_A\omega_C}$$

$$C_A = \frac{1}{2\omega_C Q_{TC} R_A (1+k)}$$

$$C_C = \left(\frac{\omega_E}{\omega_C}\right)^2 C_A$$

$$k = \frac{\frac{\omega_C}{\omega_E} - \frac{Q_{TC}}{Q_{TE}}}{\frac{Q_{TC}}{Q_{TE}} - \frac{\omega_E}{\omega_C}}$$

In practice, k must be greater than 0, which means,

$$\frac{\omega_E}{\omega_C} < \frac{Q_{TC}}{Q_{TE}} < \frac{\omega_C}{\omega_E}$$

As an example, I will show how the smallest box 'A' is calculated.
To find box-volume V_B,

Width×depth×height=14×20×(80–3.2)cm
=21.5 litre

Due to the damping material, effective box volume expands by about 20%. As the volumes of the speaker's components amount to about $0.2V_B$, the full value of 21.5 litre is maintained.
To find total Q for the two bass speakers,[15]

$$f_C = f_S\sqrt{\frac{2V_{AS}}{V_B}+1}$$

$$= 40\text{Hz}\times\sqrt{\frac{2\times 16l}{21.5l}+1}$$

$$= 40\text{Hz}\times 1.577 = 63.1\text{Hz}$$

$$Q_{TC} = Q_{TS}\sqrt{\frac{2V_{AS}}{V_B}+1}$$

$$= 0.42\times 1.57 = 0.662$$

The necessary equalised values are,

f_E=25Hz, Q_{TE}=0.5

Calculated values	Actual values
k=1.291	–
R_A=10kΩ	R_A=10kΩ
R_B=25.8kΩ	R_B=27kΩ
R_C=63.7kΩ	R_C=68kΩ
C_A=766nF	C_A=820nF
C_B=83nF	C_B=82nF
C_C=120nF	C_C=120nF

Next, analysis is needed to check the deviation,

$$\omega_C = \frac{1}{R_A\sqrt{C_A C_B}}$$

$$\omega_E = \frac{1}{R_C\sqrt{C_B C_C}}$$

$$Q_{TC} = \frac{R_A}{2R_A + R_B}\sqrt{\frac{C_A}{C_B}}$$

$$Q_{TE} = \frac{R_C}{2R_C + R_B}\sqrt{\frac{C_C}{C_B}}$$

Calculated values	Measured values
f_C=63.1Hz	f_C=61.4Hz
f_E=25Hz	f_E=23.6Hz
Q_{TC}=0.662	Q_{TC}=0.673
Q_{TE}=0.5	Q_{TE}=0.505

I believe these deviations to be acceptable. You will find these values at the righthand side of Fig. 5. You should now be able to design the box with a pocket-calculator.

Data for mid-range box 'A'
The internal volume of the 15cm diameter mid-range sphere is,

$$V_B = \frac{4}{3}\pi r^3 \approx 4.189r^3$$

$$r = \frac{1}{2}(15\text{cm} - 2\times 2\text{cm}) = 5.5\text{cm}$$

$$V_B = 0.7\text{litre}$$

From Table 4, you can find that f_s=93Hz, Q_{ts}=0.33 and V_{AS}=1.9litre. And using the f_C equation given earlier, you get:

$$f_C = 93\text{Hz}\times\sqrt{\frac{1.9l}{0.7l}+1}$$

$$= 93\text{Hz}\times 1.927 = 179\text{Hz}$$

and,

Q_{TC}=0.33×1.927=0.64

With an f_C of 179Hz, the 400Hz corner of the mid-range speaker is a good choice. The value of Q_{TC} is slightly over the damping of a Bessel behaviour, but I can live with that.

The Butterworth high-pass filter
Select the corner frequency f_{HP} and the value of R_3 in Fig. 4a).

$$C_1 = C_2 = \frac{1}{\pi f_{HP} R_3\sqrt{2}}$$

and,

$$R_1 \| R_2 = \frac{\sqrt{2}}{4\pi f_{HP} C_1}$$

So with R_3 at 10kΩ and f_{HP} at 4kHz, C_1=C_2=5.63nF and R1||R2=5kΩ (two 10kΩ resistors in parallel).

The Baxandall bass control
To obtain maximum bass voltage, contact point 3 with point 2 of potentiometer R_{28} in Fig. 6b). As inversion is involved, we get:

$$A_{V\,IB} = -\frac{Z_2}{Z_1}$$

$$Z_1 = R_{26}$$

$$Z_2 = R_{28}\|XC_{14} + R_{27}$$

$$A_{V\,IB} = -\frac{\frac{R_{28}\times 1/j\omega C_{14}}{R_{28}+1/j\omega C_{14}} + R_{27}}{R_{26}}$$

$$= -\frac{\frac{R_{28}}{1+j\omega C_{14}\times R28} + R_{27}}{R_{26}}$$

$$= -\frac{R_{27}+R_{28}+j\omega C_{14}\times R_{27}\times R_{28}}{R_{26}\times(1+j\omega C_{14}\times R_{28})}$$

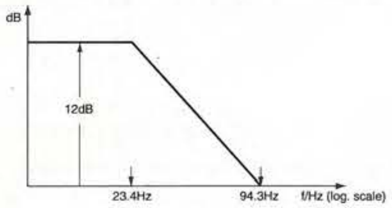
$$A_{V1B} = -\frac{R_{27}+R_{28}}{R_{26}}\times\frac{1+j\omega C_{14}(R_{27}\|R_{28})}{1+j\omega C_{14}R_{28}}$$

$$A_{V(0\text{Hz})} = \frac{R_{27}+R_{28}}{R_{26}} = 4.0 \Rightarrow 20\log 4.0$$

$$f_{\text{Denominator}} = \frac{1}{\omega C_{14}R_{28}} = 23.4\text{Hz}$$

$$f_{\text{Nominator}} = \frac{1}{\omega C_{14}(R_{27}\|R_{28})} = 94.3\text{Hz}$$

Now it is possible to draw Bode diagram:



Why has the low-pass filter 6dB/octave roll-off?
To answer this, first look at a simple low-pass filter:

$$H_{HP} = \frac{sT}{1+sT}$$

$$H_{LP} = 1 - \frac{sT}{1+sT} = \frac{1+sT-sT}{1+sT} = \frac{1}{1+sT}$$

This is a first-order low-pass filter. Now take two high-pass filters in series.

$$H_{2\times HP} = \frac{(sT)^2}{1+2sT+(sT)^2}$$

$$H_{2\times LP} = \frac{1+2sT+(sT)^2-(sT)^2}{1+2sT+(sT)^2}$$

$$= \frac{1+2sT}{1+2sT+(sT)^2}$$

If we compare this result with a Gaussian low-pass filter,

$$H_{GaussLP} = \frac{1}{1+2sT+(sT)^2}$$

you will see that there is no 2sT term as with the simple subtraction from unity. In every case only the highest degree disappears, but the rest remain. So there is no hope of a steeper slope – sorry. In every case the correct-step behaviour is preserved. ■

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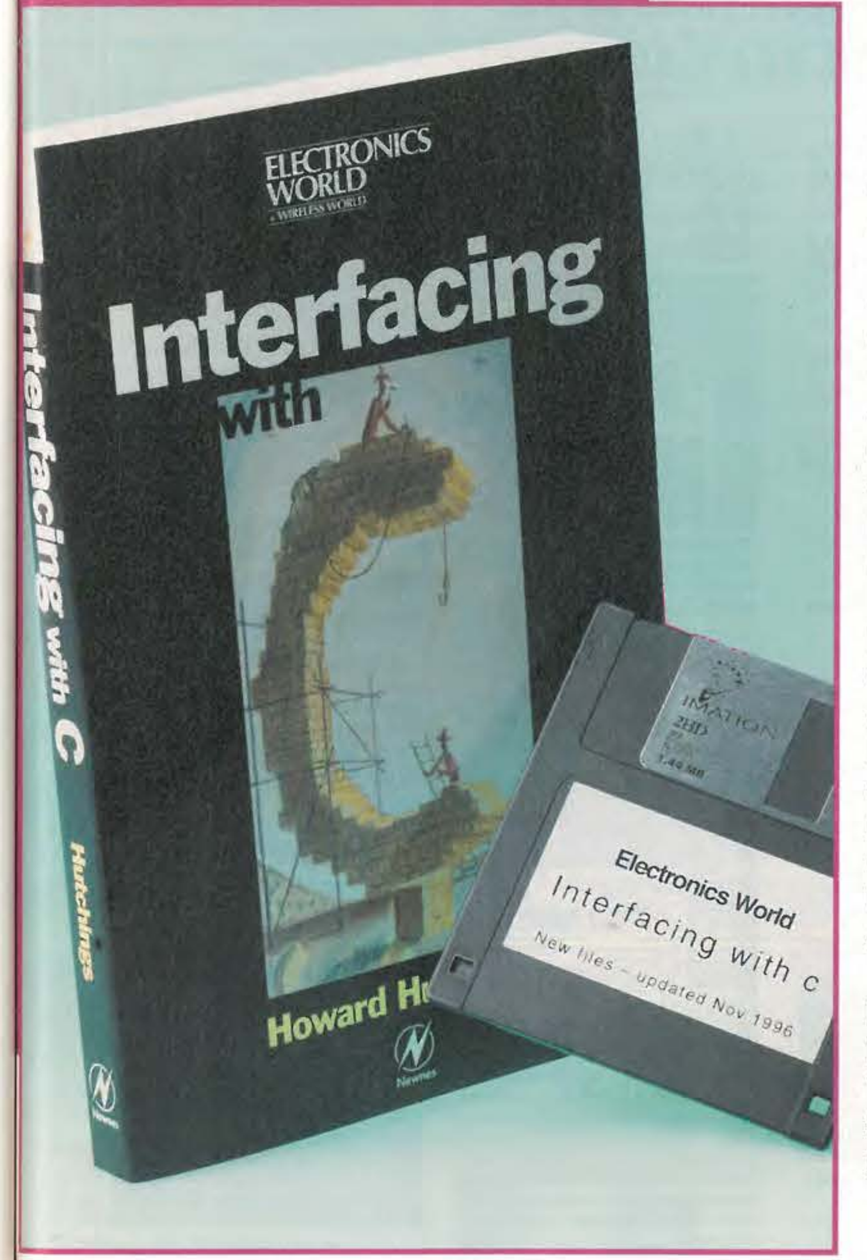

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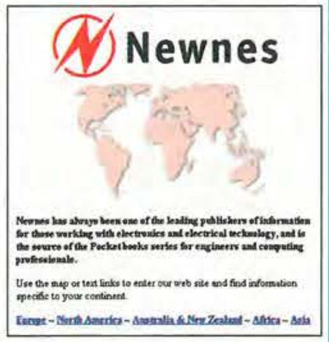

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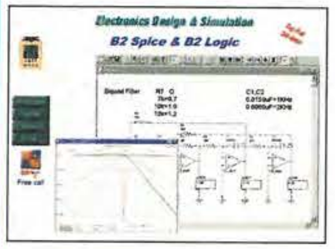

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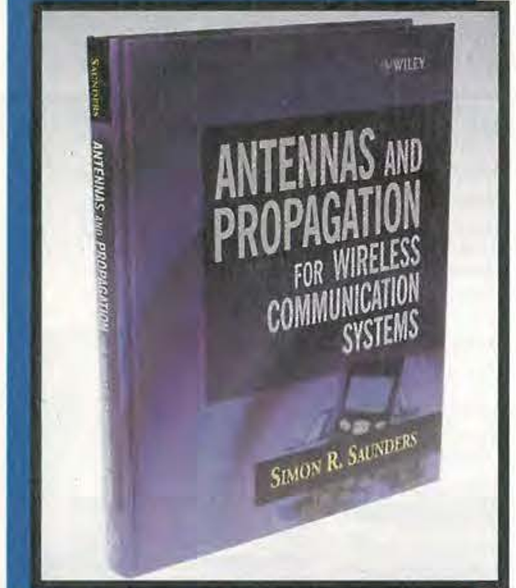

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Digital TV – a botched start to the Millennium?

While digital TV offers many advantages over analogue tv in terms of picture and sound quality and programme choice, I do not consider that enough has been done to integrate the three modes of reception – namely conventional terrestrial, cable and satellite – in one receiver.

Nor has enough been done to ensure digital tv's economical servicing by the trade. I ask myself how much consideration has been given to the implications of the fact that digital tv receivers will break down, and will have to be serviced?

By the time analogue TV is discontinued, there may be as many as 40 million digital TV sets in the UK. This represents a huge volume of production, with huge potential for economies of scale if all receivers were made identical in the reception and signal processing areas. In these areas, individual manufacturers can do little to affect the outcome in terms of picture quality and appeal anyhow.

I am sure that the savings here would more than pay for the inclusion of facilities for reception of all three modes, and a diagnostic tool allowing easy servicing via a PC.

There would be no need for set-top boxes, and manufacturers would be free to chose their power supplies, tube circuitry, controls, speakers and cabinets to give their sets consumer appeal. There could be firms similar to M.C.E.S. that could repair the modules quickly and cheaply, so that those in the trade who cannot cope with SM components or digital fault diagnosis could still get their customers' sets mended economically. A new module might even be affordable.

Digital satellite TV is chaotic, with many different encrypting systems from rival broadcasters. Most have Common Interfaces (CI), but products complying with this standard are expensive – at about £80 each.

BSkyB has given the public set-top boxes that make it difficult to access programmes from alternative sources, and that lack CI . Serious satellite viewers will need a further set-top box, and LNB and output switching – neither of which is handy or cheap.

If someone ever cracks the Videogard code used by BSkyB, users will be left with a lot of expensive junk, and providers will have to give the public more 'free' set top boxes. It would serve them right.

Robert E Littlewood
Gainsborough
Lincolnshire

RMS power indeed

I enjoyed the two articles by Joe Carr on the measurement of RF power (Nov/Dec '99) – a topic not well understood In general. Joe displayed some considerable understanding of the topic, and yet I would venture to say has not appreciated his own mis-use of the term 'RMS power'.

RMS power is itself a valid mathematical expression, but not used in practice. RMS power is the square root of the average of the squares of a number of readings of power. It is *not* the same as the product of RMS voltage and RMS current, since the product of two RMS values is not automatically RMS itself.

The reading of power that *is* useful, since it indicates the

Two gangs and the 2kΩ pot famine

Ian Hickman's distortion meter in the August 1999 issue incorporated a wire-wound 2kΩ ten-turn potentiometer. It appears that these are only available in ten-off quantities and cost over £20 each. If we receive ten or more orders for the pots, *Electronics World* will buy them in and redistribute them. Send an envelope marked 'Twin gang' with your name and address on it to Electronics World Editorial, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. Assuming we get enough orders we will advise each requester of the final price before ordering, provided that a daytime telephone number is given on the envelope.

Who invented the synchrodyne?

The term 'synchrodyne' became known in Britain when D G Tucker applied it in 1947 to his novel synchronised-oscillator direct-conversion receiver. However, I have discovered that this was not the first use of the term synchrodyne.

Browsing through some old magazines in a French second-hand shop, I came across a rather grandiose full-page advertisement for a radio receiver named the 'Synchrodyne'. The magazine was L'Illustration for January 28 1928.

In the advertisement, the Synchrodyne is described as a superhet, designed by "the celebrated scientist Lucien Levy, rendered easy of operation by 'automatic regulation'". What this meant is not clear. There is a reference to single-knob tuning so perhaps it was merely ganged tuning.

The principle of first use would seem to give the term 'synchrodyne' to Levy, since he pre-empts Tucker's later – clearly independent – coinage of the same word. Fortunately Tucker himself suggested an alternative[2]. This is 'homodyne' – a term that first appeared in the pages of *Electronics World* when it was still *Wireless World*..

In 1924 F M Colebrook used the word homodyne to describe a synchronous receiver comprising a greatly oscillating reacting detector, whose frequency is synchronised by a received carrier.[3] Tucker proposed using 'homodyne' for the class of receivers in which demodulation is assisted by enhancing the carrier in some way.

George Short
Brighton
Sussex

References

1 Tucker, D G, "The synchrodyne" *Electronic Engineering*, vol 19, p 75-76 March 1947

2 Tucker, D G , "The history of the homodyne and synchrodyne" *J Brit IRE* April 1954 p 143-154

3 Colebrook, F M "Homodyne" *Wireless World & Radio Review* 1924, vol 13 pp 645-648

This advertisement appeared in 1928 – almost two decades before Tucker applied the word 'Synchrodyne'.

heating power available, is proportional to the square of RMS voltage (bolometric power principle), or the product of RMS voltage and RMS current. It is normally termed average power.

S. Pepper
Hertford

Thanks to all of you who wrote in on this point. I did of course let the term slip through on purpose so that readers who weren't aware of the inappropriateness of RMS power could learn a valuable lesson – honest. Joe's reply to Mr Pepper was, "Good point, $I_{rms} \times V_{rms} = P_{rms}$ only in restive circuits." – Ed.

How long's a centimetre in ounces?

About 8% of males are colour blind. Nearly all clock-radios have a red LED digital display. Nearly all video recorders have a green fluorescent display.

Do colour-blind males find it easier to read the time on a video recorder?

R N Soar
Doncaster
South Yorkshire

The perfect transistor is a current conveyor

Cyril Bateman describes, 'The perfect transistor', December 1999, pp. 1049-53, which is a relatively new device from Burr Brown, the OPA660.

In fact the concept of the same analogue building-block, which exhibits precise unity voltage and current gain, was first proposed by Sedra and Smith in 1970[1] when semiconductor ICs were in their infancy. They called it a current-conveyor.

At that time it was merely a circuit concept without practical implementation. The second generation current-conveyor, or CCII, offers as much, if not more versatility than the operational amplifier. One particular advantage of the CCII is its current output capability, making it ideal for transconductance and current amplifier applications.

However, it was not until the development of mass-producible fully-complementary bipolar technology almost 20 years later that the capability of creating a high performance single-chip CCII became practical.

In 1990 working in collaboration with Elantec Inc, LTP Electronics designed and created the first monolithic dual current-conveyor, part number CCII01, which we then marketed on behalf of Elantec. Interestingly, the Burr Brown OPA660 has virtually the same topology as our device.

Unfortunately when we launched the CCII01 early in the 1990s[2], the electronics industry looked somewhat sceptically at this new analogue building-block. Despite many passionate enthusiasts around the world who realised the value of the CCII01, we were forced to discontinue sales as our partner manufacturer, Elantec, stopped manufacturing. This was because the volume of sales at that time was simply too low.

It is now 30 years since Smith and Sedra proposed the current conveyor, and though it is not being marketed as a current-conveyor, the OPA660 is just that. I am very pleased to see its arrival in the Burr Brown stable. I am sure that when applications engineers get some experience with it, they will realise its true potential and it will become as useful a part in the analogue designer's tool-kit as the ubiquitous op-amp.

John Lidgey
Professor of Electronics
Oxford Brookes University
Director of LTP Electronics Ltd,
Oxon

References

1. A S Sedra and K C Smith and, 'A second generation current conveyor and its applications', IEEE Transactions on Circuit Theory, Vol CT-17, pp 132-134, February 1970
2. C Toumazou, F J Lidgey and M A Vere Hunt, 'Advanced BJT technology makes current-conveyors a practical reality', New Electronics, March 1993, pp24 & 25

We hope to present a full article on the history, development and application of the 'perfect transistor' – the current-conveyor – in the next issue – Ed.

Wien and Christmas

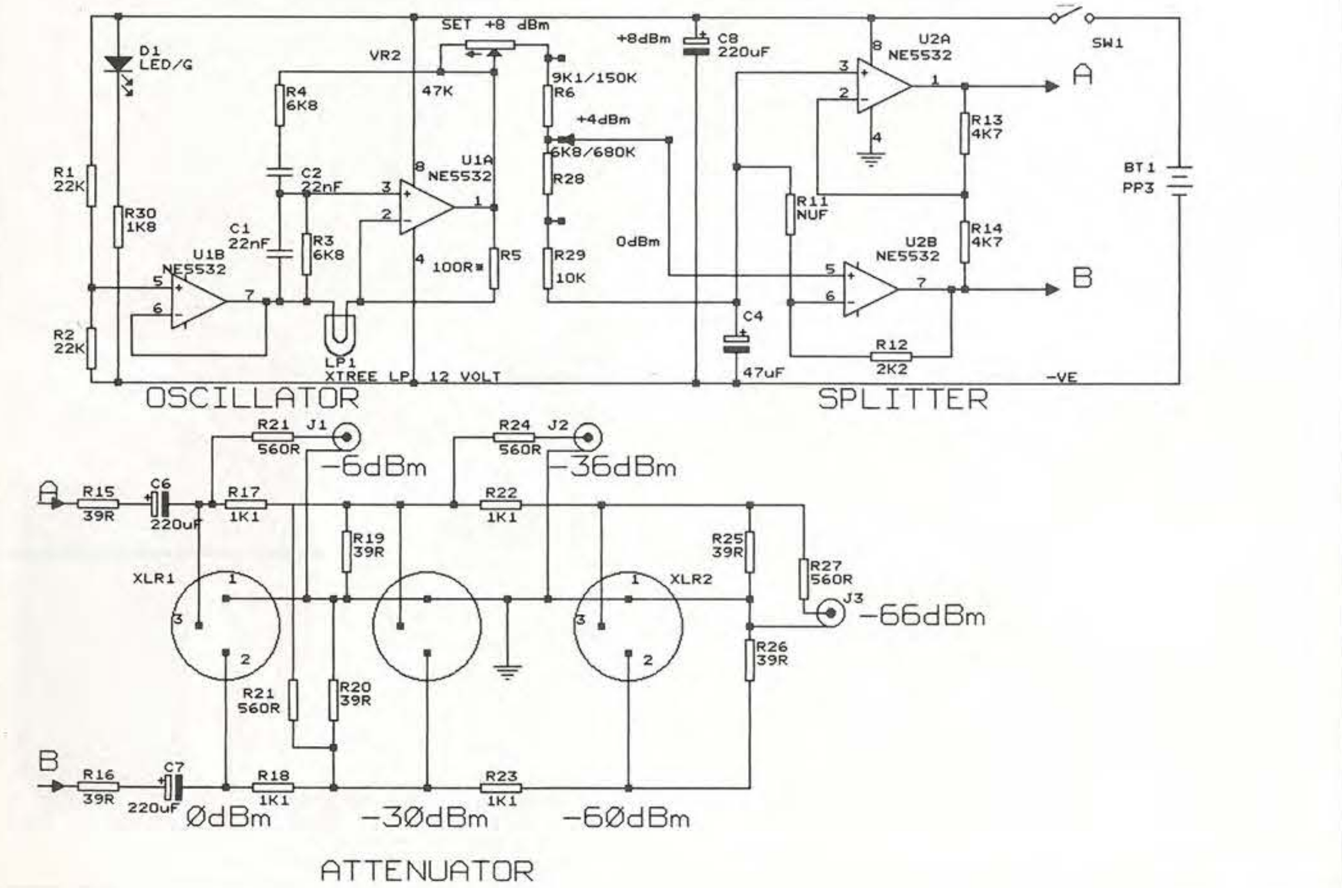
I was Interested in Ian Hickman's article about Wien-bridge oscillators in the January issue. But for my needs, the R52 thermistor mentioned was rather extravagant.

Wanting a simple Wien oscillator as a 1kHz tone source, I found that a cheap alternative is a 12-volt Christmas-tree lamp. Since it has a positive temperature coefficient rather than the negative one of the R53, it has to be placed in the ground leg of the feedback network, and a fixed resistor of around 100Ω is needed in the upper leg.

I have made two of these oscillators and they have been very satisfactory for their purpose. Woolworths is useful source of spare lamps.

The circuit below shows the simple battery powered tone source with its attenuators.

Michael Tong
Twickenham
Middlesex



Phase-linear misconception

I think Mr de Boer's letter in the November issue misinterpreted my aim in designing the phase-linear crossover on page 779 of the September issue. The rate of attenuation of the high-pass response was not of prime concern as the design was primarily intended for an electrostatic high-frequency unit with a wide frequency response.

The aim was to produce a crossover with the lowest possible distortion. I would gladly trade off a 'mellow' attenuation for phase linearity, flat time delay, good damping and impulse response (virtually no ringing, undershoot, overshoot) fast rise and settling times.

I built the design mentioned in National Semiconductor's Audio Design handbook in 1981, but its poor sonic performance led me to design the phase-linear circuit. The old design suffered from what can only be described as 'smearing' of the signal, which made it uncomfortable to listen to over long periods.

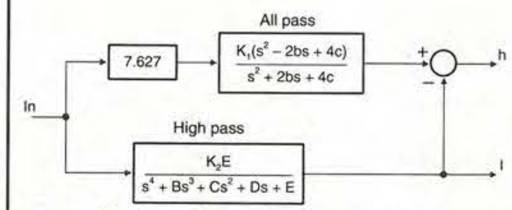
Secondly, the transfer function shown by Mr de Boer is not correct as he opted to ignore the all-pass filter's effect from his version of what the high-pass output response should be – input less low-pass.

Normalised transfer functions (ω_0=1) of the phase-linear crossover are shown in the diagram. Terms b and c are the 2nd order Bessel coefficients and B, C, D, and E, are the 4th order Bessel coefficients. K is the sections gain. Ignoring the gain of 1.906, the correct high out normalised transfer function will be:

$$H_h(s) = \frac{s^2 - 6s + 12}{s^2 + 6s + 12} - \frac{105}{s^4 + 10s^3 + 45s^2 + 105s + 105}$$

Solving this equation shows that there are zeros at the origin, which indicates that this function will act as a high-pass filter. It also has the same behaviour as the low-pass filter. ■

Peter Latsky
Fordingbridge



Normalised transfer functions (ω_0=1) of the phase-linear crossover.

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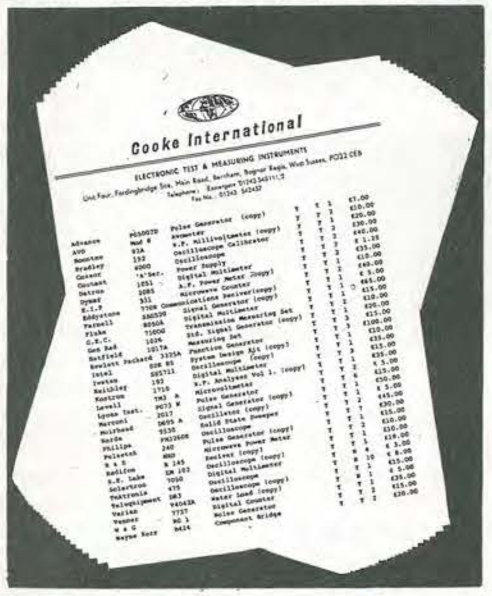

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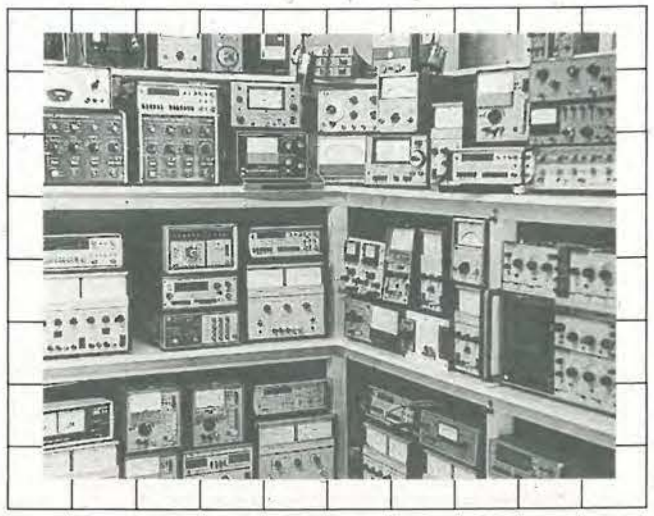

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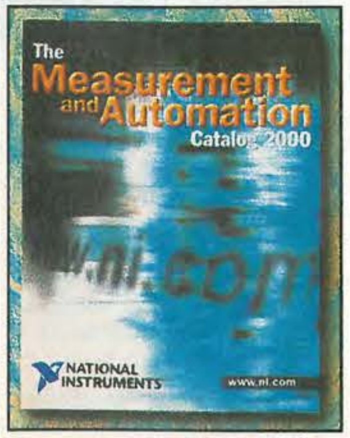

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