Recruitment special: starts page 560

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Self explains muting relays

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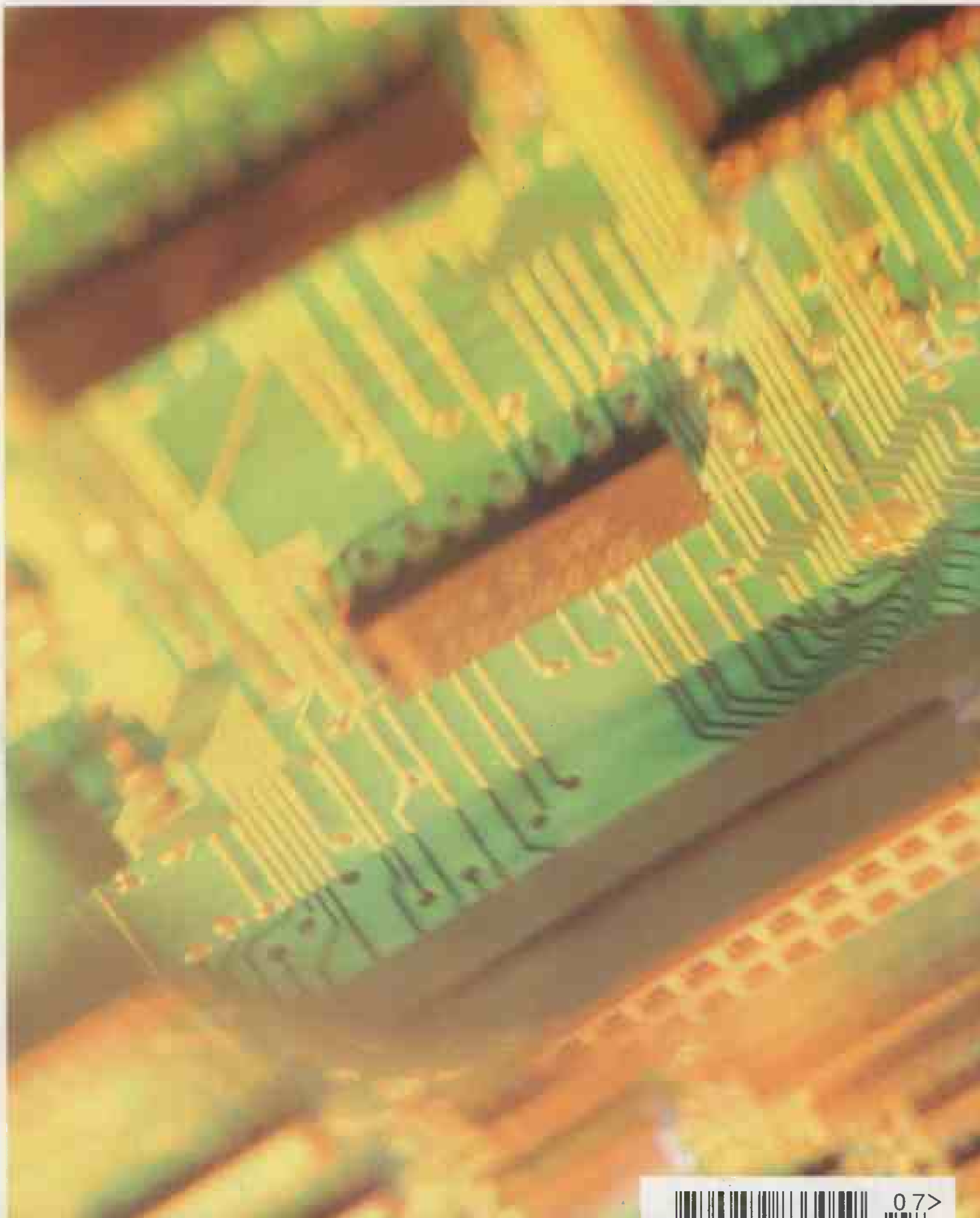

The route to simulation: new review series starts with Workbench

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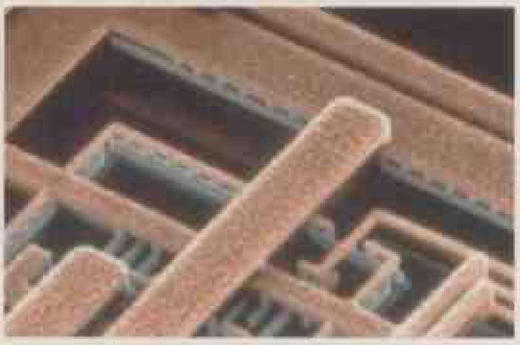

Copper interconnects offer significant benefits over aluminium, but can they be mass produced? Find out on page 532.

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585 HOTTER SPICE 2
In this second article presenting a breakthrough in power mosfet modelling for audio design, Ian Hegglun adds compensation for electrothermal effects and answers the question of whether sub-threshold conduction is needed for simulating crossover distortion.

591 EARLY APPLIED
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611 PROGRAMMING THE EEPROM HC11
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This month's cover mount is an advanced thermally-conductive material that can fill air voids and surface irregularities without the need for silicone grease. See page 607 for more.

In 1949, Maurice Wilkes built the first computer designed to store programs in 1.6m long memories. Read more on page 580.

August issue on sale 24 June

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Tuning step size	100 Hz (5 Hz BFO)	100 Hz (1 Hz for SSB and CW)	100 Hz (1 Hz for SSB and CW)
IF bandwidths	6 kHz (AM/SSB), 17 kHz (FM-N), 230 kHz (W)	2.5 kHz(SSB/CW), 9 kHz (AM) 17 kHz (FM-N), 230 kHz (W)	2.5 kHz(SSB/CW), 9 kHz (AM) 17 kHz (FM-N), 230 kHz (W)
Receiver type	PLL-based triple-conv. superhet		
Scanning speed	10 ch/sec (AM), 50 ch/sec (FM)		
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REED BUSINESS INFORMATION

Digital terrestrial: a launch too soon?

Launching a brand-new entertainment product at a time when consumers are scarcely flush with spending money is a brave stratagem, yet digital television has successfully managed to allure around a quarter of a million viewers since its launch.

With an either/or choice between satellite (BSkyB) and terrestrial (British Digital Broadcasting, that's ONdigital to you and me), the pundits were expecting digital consumers to opt overwhelmingly for Sky with its established brand image and earlier launch date.

In point of fact ONdigital's 'through the air' sales pitch seems to have struck a chord. In its first four and a half months of operation, Ondigital succeeded in signing up 110 000 customers – only 10 000 fewer than BskyB achieved.

Digital terrestrial television, DTT, will receive a further boost this summer when the first TV receivers with built-in digital decoders appear in the shops, at a subsidised price of £500 or so; outboard set-top boxes are an unattractive solution.

So far so good: the road map for the digital replacement of analogue terrestrial transmissions is coming into clear focus and DTT has been proven as viable technology. Or has it?

Not if this viewer's experience is anything to go by. DTT has been oversold. In exactly the same way as BSB made extravagant claims for its miniature Squarial, the marketeers at ONdigital have convinced themselves that DTT can charm birds out of trees and turn base metal into gold.

Their technical advisers must be squirming...

To begin with, ONdigital's advertising claims of 'crystal-clear pictures' and 'the sharpest sound with the immediacy of CDs' are no more than claims. Although noise-free, the pictures are noticeably 'soft', while the data rate used for audio fails to reach the sampling levels necessary to match CDs and is in fact no better than the also digital NICAM used on analogue transmissions. Good analogue reception is every bit the equal of good DTT reception and to infer otherwise is deceit.

The promotional literature is also less than honest with its promises of improved pictures. While the DTT set-top box can theoretically produce results on-screen that beat most analogue signals, these results will only be achieved on sets having a direct RGB input. Nowhere is this stated in the literature, and anyone stuck with the more commonplace arrangements of UHF or composite video input will enjoy pictures that are no more than average and probably no better than analogue.

Whereas analogue reception degrades gradually, receiving DTT is an either/or affair. It may be adequate in winter, when there are no leaves on the trees, yet disappear in summer.

The suggestion that viewers with indifferent analogue reception will see better pictures on digital is not borne out by experience. In fact the whole success of DTT is predicated on optimised antenna installations – a phenomenon that's extremely rare in reality.

Most viewers have corroded, misaligned rooftop aerials and a coaxial down-lead that is more attenuator than feeder; those that don't rely on loft antennas or set-top devices that is. All these may be adequate for indifferent analogue reception but for the more demanding DTT process they just don't cut the mustard.

Articles in the trade press also indicate that many antenna fitters are inexperienced in optimising installations for digital and are not equipped with the special meters and equipment necessary. Fortunately ONdigital is committed to replacing aerial installations – with some exceptions – but the company may end up having to change many more of these than anticipated.

Not everyone has the choice of ONdigital either. Coverage from the 81 digital transmitter sites currently being built is stated by the Digital TV Group to be between 73 and 90 per cent of population as opposed to the 99.4 per cent of analogue, meaning than many viewers will be disappointed. Significant areas of the country have no reception as yet and it will be a long time before all relay transmitters are digitised – if ever.

Even where viewers receive rock-solid analogue pictures, digital reception may suffer constant break-up on account of local impulsive interference. Causes include domestic light switches and thermostats, DECT cordless phones and passing road traffic, and for all the robustness that digital transmissions may show in the face of multipath reception, they are pitiably vulnerable to the natural hazards of the electromagnetic spectrum.

Whereas these effects produce minor flecks and buzzing on analogue television, with digital they result in picture loss for two to three seconds and extremely unpleasant audible pops and cracks way above normal listening levels.

In short digital terrestrial television is flawed – significantly flawed. This is not to deny its advantages; the future is inevitably digital and eventually improved receiver design and increased transmitter power will eliminate the current bugbears.

Nor would I decry the commercial achievements of British Digital Broadcasting and the attractive assortment of programming offered. Time will tell if their business is a success – and we'll all be in trouble if it is not. It is nonetheless unfortunate that early adopters must pay the price of field-trialling at their own expense a system that is demonstrably inadequate and that may also prejudice further take-up. ■

Andrew Emmerson

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UK high-tech firms turn recession tide

The UK is emerging from the risk of recession on the back of a reshaped high-tech industry, according to several reports out this week.

A study by Deloitte & Touche says the high-tech industry is benefiting from unprecedented growth, with the UK's top 50 fastest growing technology companies recording an average increase in turnover of 1857 per cent from 1995 to 1997. At the same time, the government released figures showing a 0.1 per cent increase in gross domestic product, which analysts say means the country has avoided a recession.

"The main message from this survey is one of great optimism. The UK has a technology business that is driven by highly focused entrepreneurs," said William Touche, a partner with Deloitte & Touche.

Scotland is doing particularly well with one quarter of the top 50 firms located in the area, higher than in London and the south east. Other evidence of the UK's burgeoning high-tech industry came from the Invest in Britain Bureau (IBB) which reported an increase in overseas investment in UK high-tech firms.

"We looked back over the past three months to test our drive in attracting high-tech investments rather than large scale manufacturing jobs. Britain is doing exceptionally well," said an IBB spokesman. The IBB reported a 39 per cent rise in overseas high-tech investments compared to the first quarter of 1997.

"Hopefully this news adds to the government's White Paper. It makes clear where we are going," added the spokesman. The White Paper was introduced to encourage start-ups and investments in the high-tech industry.

However, a note of caution was offered by the Federation of Electronics Industry (FEI). "It depends what products you are selling, what industry you are serving. The fastest growing is the mobile telecoms market but if you are involved in the infrastructure – base stations – side of the market, it is not growing as fast as other areas," said Richard Hinds, the FEI's components director.

Alex Mayhew-Smith
Electronics Weekly

Copper chip interconnects "ready to roll"

IBM Microelectronics has a copper interconnect on silicon on insulator (SOI) process available for prototyping that can draw gate lengths on silicon down to 0.11μm.

"For the past couple of years our process technology has been very showy," said Dr Lisa Su, project manager for CMOS Logic Technologies at IBM Microelectronics, pointing out that the 0.11μm features are drawn using a 248nm ultraviolet wavelength.

At the moment, the process is available to customers only for prototyping but, "it is ready to roll," said Su.

Before volume production is available to OEMs, an ASIC library for the process has to be developed. According to IBM, this is currently underway.

The company has used the process internally to make PowerPCs and ASICS. It expects customers to use it for IBM BlueLogic system-level chips and for ASICS used in high-end servers.

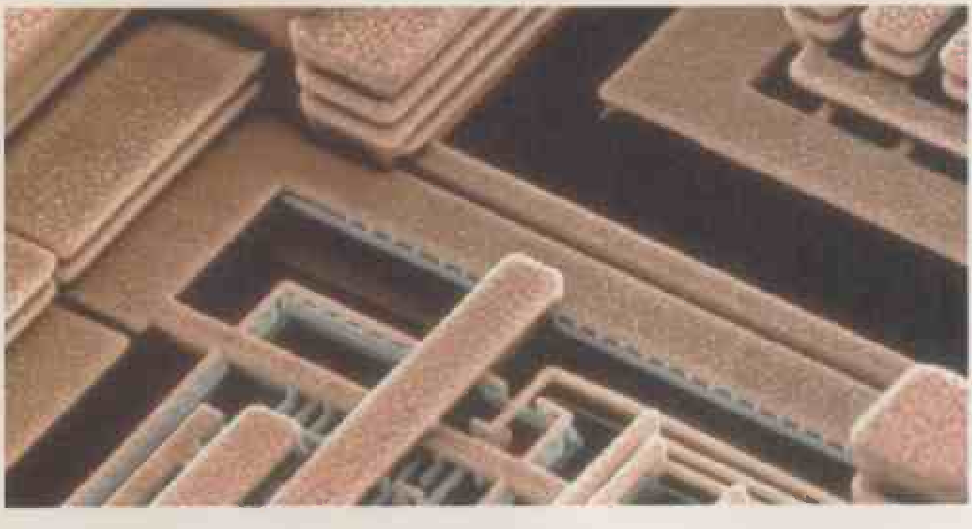
Using copper interconnects speeds up chips by 10 to 20 per cent compared to aluminium, says IBM, while using SOI wafers improves transistor performance by up to 35 per cent and offers up to a threefold reduction in power.

Although SOI wafers cost three times the price of silicon wafers, according to Andre´ Auberton-Herve´, president of SOI wafer suppliers SOITEC, the extra cost for finished silicon is only ten per cent, claims Su.

IBM calls its copper/SOI process SA27E. "The reason SOI is so attractive is that it is an ideal transistor," said Su. "The difficulties with it have been in getting good enough materials without defects. What has kept people away from SOI is that scaling has been going so fast that SOI couldn't catch up. It could only be useful if SOI and silicon scaling were equivalent. Now SOI has caught up."

David Manners *Electronics Weekly*

Cross section of the structure of IBM's copper interconnect process



UK leads Europe in IC design capability

The UK has taken the lead in Europe in high-level IC design. Local firms now account for well over half of all the revenues generated by Europe's independent design houses, according to a report by analysts Future Horizons.

"The level of intellect that's there is staggering," said Malcolm Penn, chairman of Future Horizons.

The report says that out of 140 design houses in Europe, 60 are in the UK and that, out of total 1998 design revenues of $1.06bn generated in Europe, $604m are generated by UK design houses.

The reason why the UK generates more than half Europe's revenues with less than half Europe's number of companies is, "because of the UK background in telecomms and mixed signal where you have to have the real brainpower," said Penn.

"Where you're not just linking millions of gates together but you're hand-crafting the transistors to get performance – these are fabulously intellectual skills and a very scarce resource," he continued.

"It's back to the days when you had linear designers who would sit for months in the desert drinking Tequila and then come back with a brilliant design," said Penn.

Driving the growth of the design house is the move to system level chips. Because OEMs and semiconductor houses do not need to employ high-level designers permanently, it is cheaper to out-source design work.

"Typically these companies are five to 15 engineers with sophisticated software but inexpensive workstations," said Penn. "The software cost is horrendous but after that it's all brainpower."

Typically the UK's design houses are located in clusters: in Bristol where they have spun-off from Inmos, in Swindon where they have spun off from GEC-Plessey Semiconductors (now Mitel Semiconductor), and around the Universities of Edinburgh and Cambridge.

European high-level IC design revenues will grow at 34 per cent compound annual growth rate over the next five years, said the report.

David Manners

1000MHz PC has fridge technology

Advanced Micro Devices (AMD) has demonstrated a PC running at 1GHz using an innovative cooling system.

The prototype PC, called the Super-G, uses AMD's forthcoming K7 microprocessor cooled by a system from KryoTech. Cooling the microprocessor allows engineers to increase the clock speed without worrying about overheating the chip.

"The Super-G is derived from two years of technical co-operation between Kryo-Tech and AMD," said Al Quick, chairman and CEO of KryoTech.

The Super-G cooling system uses a vapour phase refrigeration technology similar to that used in refrigerators.

KryoTech said it will ship Super-G PCs later this year and will release pricing information at a future date.

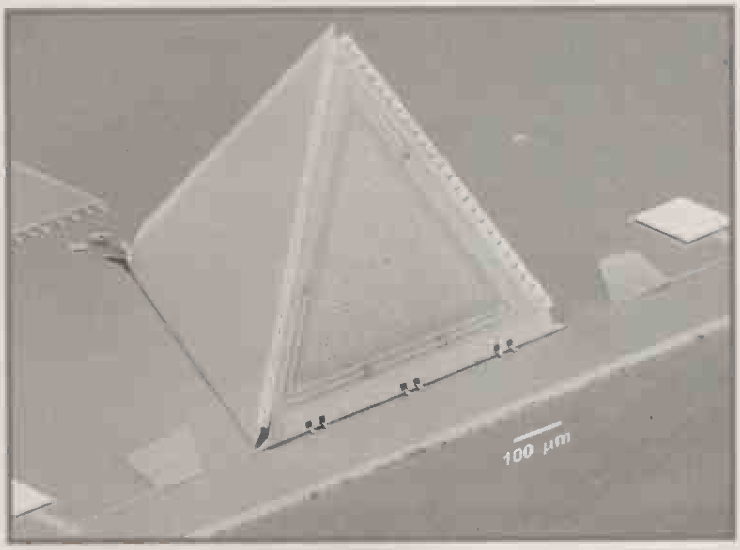
AMD competitor Intel has also demonstrated a 1GHz PC but it is believed to have used liquid nitrogen rather than a commercial system to cool the processor.

No hot heads at BT

BT has denied that it has instructed its staff to use headsets with their mobile phones in response to a former employee's plan to sue the telecom giant.

The company confirmed that hands-free sets are available to those employees requesting them but said this has been the case for some time. "Our position hasn't changed on this issue at all," said a BT spokesman. "The fact we give employees the chance to have hands-free sets if they want doesn't change our view that there's no evidence that mobile phones are harmful."

Stephen Corney, a former BT engineer, is suing the company over permanent brain damage allegedly caused by using mobile phones for several hours a day as part of his job.



Midget mike... Bell Labs researchers have built a tiny microphone on a silicon chip in their goal towards building a low-power, single-chip radio. Peter Gammel and colleagues at Bell Labs, the research arm of Lucent Technologies, built the microphone using micromachining.

Tempo rivals Dixons with free Internet

In the latest battle to win subscribers to free Internet service, Dixons' rival Tempo has set up a service offering free calls.

The high street chain is offering a free service, the same as that offered by Dixons' Freeserve, but with the addition of free connection calls during off-peak hours.

Tempo's screaming.net service will offer 118 off-peak hours a week Internet connection at no charge. The service will be available at first from Tempo stores through free CDs containing the necessary software.

Users will have to sign up with telecoms service provider Localtel to use the service, which will make a return on normal voice calls. This, the firm says, will be cheaper than BT rates.

Initially a limited number of CDs will be available so as not to swamp the service. Later on these will be on offer at the rate of 100000 to 200000 a month.

Digital radio waves hello to Web access

The BBC is working with UK start-up RadioScape to enable Web access through digital radio transmissions.

The initiative will mean that Web sites containing headline news and share prices can be transmitted to a PC without connection to the Internet. The information will then be continually updated using digital radio, or digital audio broadcasting (DAB) as it is called.

"This is where DAB is going to become compelling," said Peter Florence, managing director of London-based digital radio specialist RadioScape. "This is the year it is really happening."

RadioScape is working on the infrastructure of the system and is receiving funding from the Department of Trade and Industry. The results of its work will be made available to those involved in the digital radio community. "We'll be defining and developing software and then making it available to help people build up data services and broadcast data services in DAB," said Florence.

Work on the project has already begun and is anticipated to take six months. Broadcasts could start by the year end.

To receive the radio transmissions, software and a DAB PC card will need to be installed in the PC. The system will also allow the normal data associated with DAB transmissions to be read by the PC. This includes data appended to radio programmes offering such information as the title of the current music and the band playing it.

Florence stresses the value of DAB's ability to add data available from digital radio, especially when it could attract "some people who don't want to get into the whole Internet thing".

Melanie Reynolds

LCD adapts to bright and dim

Biggin Hill-based liquid crystal display (LCD) manufacturer Densitron has developed a display technology that improves visibility in different ambient lighting conditions.

"It's a major breakthrough for us," said Nick How, product manager at Densitron.

The display shows a negative image and uses a backlight in low light conditions. When light levels are high, the display is positive, and reflects light.

Called Chameleon, the technology has applications where there is both high and low ambient light, such as marine and automotive environments and mobile phones.

"The technology automatically senses the ambient light and adjusts the image on the display from a positive to a negative image," said How. "As ambient light stops getting reflected, transmitted light takes over."

The technique can be applied to any size of display, said How, with minimal cost increase to the glass. Because less backlighting is needed, overall system cost could be reduced, he claimed.

"And a by-product is that reflected light is double that of a normal LCD technology," said How. Up to 70 per cent of incident light is reflected.

Backlights can use LED or electro-luminescent techniques.

Densitron has no plans to license Chameleon to other LCD manufacturers, preferring to keep it for its own displays.

Richard Ball

Positive or negative? Densitron's Chameleon display technology changes from a positive to negative image as ambient light levels change. In bright light, the positive image is reflective, saving battery power otherwise required for the backlight.

Sub 100nm geometry from optical litho?

Flash memory with a minimum feature size of just 0.08µm has been manufactured using conventional optical lithography by Bell Labs.

Researchers claim this is the smallest working device ever manufactured using optical lithography. Stepper manufacturers and semiconductor engineers believed optical techniques would reach a limit at 0.12µm, or 120nm.

The achievement could result in big savings for semiconductor makers, by not having to switch to new manufacturing technology.

To make the device, Bell Labs used a 193nm laser in combination with phase shift masks. Photo-resists were also developed to work with the shorter wavelength light.

The resulting flash storage cell has a floating gate measuring 80x160nm.

Beyond the 193nm used by Bell Labs, semiconductor manufacturers are looking towards extreme ultra-violet (EUV) lasers, ion-beam and electron-beam.

New chip solves music industry's web copyright fears

Lucent Technologies, Texas Instruments (TI) and e.Digital are developing a device that can download from the Internet copyright protected music.

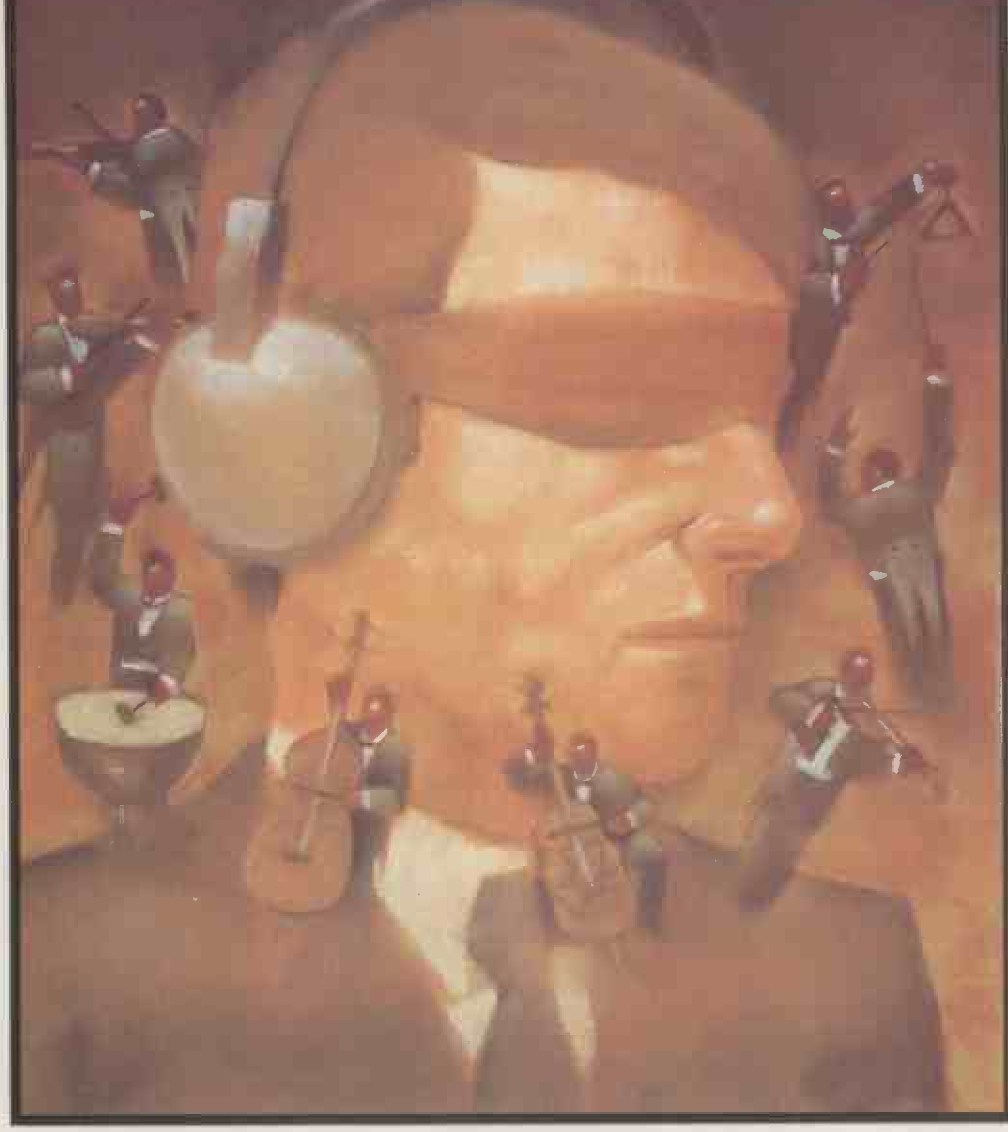
The device will compete with other portable digital music products that play audio encoded in the popular MP3 format.

e.Digital will make the portable devices based on a TI DSP running Lucent's enhanced perceptual audio coder (EPAC) technology.

"The quality of the sound that we've heard with EPAC on our hardware platform is exceptional," said Fred Falk, CEO of e.Digital.

Lucent and its partners hope that their EPAC-based devices will find favour among main music publishers who are shunning MP3 because it lacks copyright protection.

EPAC uses an 11-to-1 compression ratio to provide CD-quality sound and was developed at Lucent's Bell Labs. ■



When considering a stereo system, most people look at distortion, frequency response and power output. Purists also look at things like damping factor, noise and slew rate. But what about the system's stereo image? Does it matter? If so, how can it be quantified? John Watkinson provides the answers.

Stereo from all angles

This set of articles looks at every aspect of stereo and in particular those areas where quality can be lost. Starting with the human ear, it considers the direction sensing mechanism, the way that the illusion of intensity stereo is created, the way stereo microphones should work and possible pitfalls, and then the way in which loudspeakers ought to work in stereo but frequently don't.

In addition to explaining the theory of stereo, I will present practical advice on what to do to make a tangible difference.

Where is it coming from?

Without doubt, the ability to convey some impression of the spatial nature of sound adds considerably to the realism of an otherwise accurate reproduction of the frequency and time domain information.

It is axiomatic that the more accurate a reproduction system is, the more realistic it will sound. Various measurement techniques have evolved for audio equipment in which an objective measurement is made and compared with some criterion based on subjective tests of human hearing.

Measurements of frequency response and linear and non-linear distortion are routinely made to assess the quality of audio equipment and all have internationally agreed measurement units.

The mechanism by which humans determine sound direction, and the accuracy to which it can be done are well known. Accordingly, this ought to be the criterion for the spatial accuracy of a stereo reproduction system – just as the ear's sensitivity

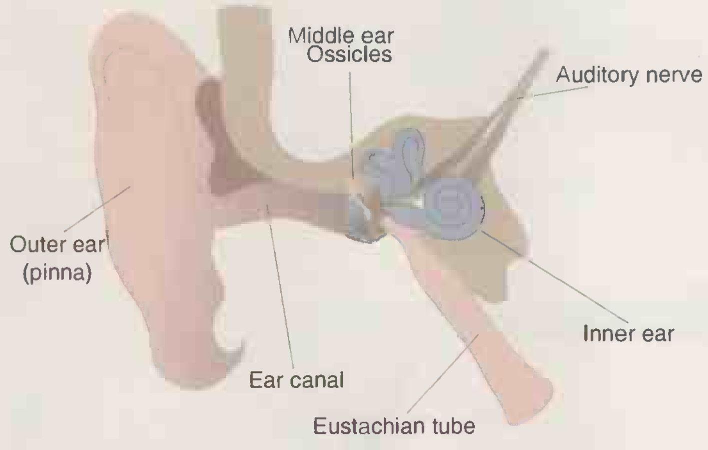


Fig. 1. Structure of the human ear. The outer ear works at low impedance, the inner ear works at high impedance. The middle ear is an impedance matching device.

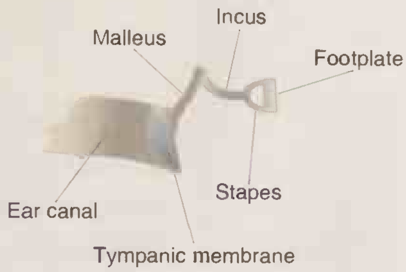


Fig. 2. The malleus tensions the tympanic membrane into a conical shape. The ossicles provide an impedance-transforming lever system between the tympanic membrane and the oval window.

to distortion is the basis for audio linearity criteria.

I find it bizarre that there are no internationally agreed techniques for measuring the spatial accuracy of stereo reproduction systems. Nor are there even any units of measurement, although other industries purveying spatially-disposed images have evolved such means.

The consequences of this lack are all around us. It is well known that loudspeakers displaying identical conventional measurements sound different from each other and from the original. This confirms that existing measurements give an incomplete picture of performance.

In the absence of a psychoacoustically based unit of spatial accuracy, the comparison of speakers becomes a minefield of subjectivism and technical progress is hampered.

I have been carrying out fundamental research into stereo imaging accuracy and I have consistently found that whenever a traditional approach is replaced by an approach dictated by imaging theory, the results are simply more realistic.

As this series evolves, it will become clear that there is considerable scope for improvement in the realism of stereo reproduction and that an additional metric is required to help bring that about.

Enter the ear...

The ultimate criterion for sound reproduction is that the human ear is fooled into thinking it has heard the real thing. It follows that good understanding of human hearing has to form the basis for all of the value judgements which will be necessary as system design proceeds.

Figure 1 shows that the structure of the ear is traditionally divided into the outer, middle and inner ears. The outer ear works at low impedance, the inner ear works at high impedance, and the middle ear is an impedance matching device.

The visible part of the outer ear is called the pinna which plays a subtle role in determining the direction of arrival of sound at high frequencies. It is too small to have any effect at low frequencies. Incident sound enters the auditory canal or meatus.

The pipe-like meatus causes a small resonance at around 4kHz. Sound vibrates the eardrum or tympanic membrane which is stretched between the outer ear and the middle ear.

The inner ear, or cochlea, works by sound travelling though a fluid. Sound enters the cochlea via a membrane called the oval window. If airborne sound were to be incident on the oval window directly, the serious impedance mismatch would cause most of the sound to be reflected. The middle ear remedies that mismatch by providing a mechanical advantage.

The tympanic membrane is linked to the oval window by three bones known as ossicles. These act as a lever system such that a large displacement of the tympanic membrane results in a smaller displacement of the oval window, but with greater force.

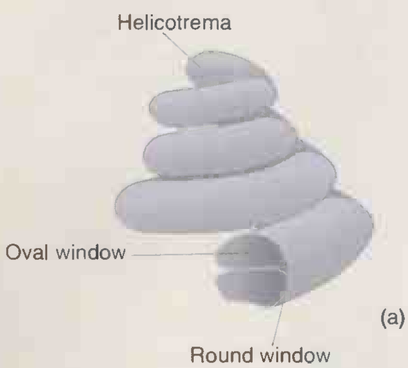
Figure 2 shows that the malleus applies tension to the tympanic membrane rendering it conical. The malleus and the incus are firmly joined together to form a lever. The incus acts upon the stapes through a spherical joint. The area of the tympanic membrane is greater than that of the oval window, creating a further mechanical advantage. Small pressures over the large area of the tympanic membrane are converted to high pressures over the small area of the oval window.

The middle ear is normally sealed, but ambient pressure changes will cause static pressure on the tympanic membrane that is painful. The pressure is relieved by the eustachian tube, which opens involuntarily while swallowing.

The cochlea, shown in **Fig. 3a)**, is a tapering spiral cavity within the bony walls which is filled with fluid. The widest part, near the oval window, is called the base and the distant end is the apex.

Figure **3b)** shows that the cochlea is divided lengthwise into three volumes by Reissner's membrane and the basilar membrane. The scala vestibuli and the scala tympani are connected by a small aperture at the apex of the cochlea known as the helicotrema.

Vibrations from the stapes are transferred to the oval window and become fluid pressure variations



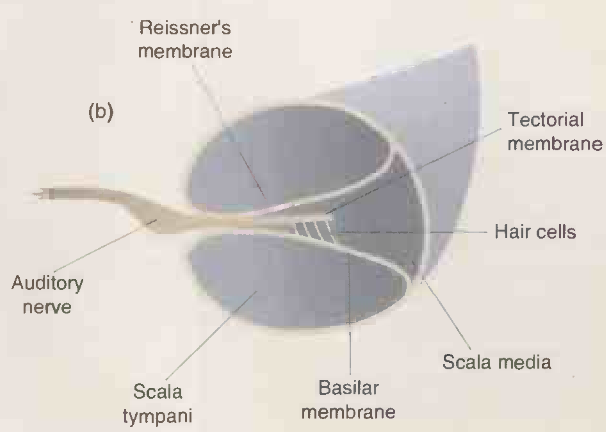


Fig. 3. At a) is the cochlea and b) is a section through it. Diagram c) shows the cochlea unrolled for clarity.

which are relieved by the flexing of the round window.

Effectively the basilar membrane is in series with the fluid motion and is driven by it except at very low frequencies where the fluid flows through the helicotrema, bypassing the basilar membrane. Figure 3c) shows that the basilar membrane tapers in width and varies in thickness in the opposite sense to the taper of the cochlea.

How various frequencies affect the mechanism

The part of the basilar membrane that resonates as a result of an applied sound is a function of the frequency. High frequencies cause resonance near to the oval window, and low frequencies cause resonances further away.

The distance from the apex where the maximum resonance occurs is a logarithmic function of the frequency so that tones spaced apart in octave steps will excite evenly spaced resonances in the basilar membrane.

The existence of resonance at a location on the membrane which is a function of frequency is predicted by place theory. Essentially the basilar membrane is a mechanical frequency analyser.

A knowledge of the way it operates is essential to an understanding of musical phenomena such as pitch discrimination, timbre, consonance and dissonance and to auditory phenomena such as critical bands, masking and the precedence effect.

The vibration of the basilar membrane is sensed by the organ of corti, which runs along the centre of the cochlea and contains elements which can generate vibration as well as sense it. These are connected in a regenerative fashion so that the Q factor, or frequency selectivity, of the ear is higher than it would otherwise be.

The deflection of hair cells in the organ of corti triggers nerve firings and these signals are conducted to the brain by the auditory nerve.

Nerve firings are not a perfect analogue of the basilar membrane motion. A nerve firing appears to occur at a constant phase relationship to the basilar vibration; a phenomenon called phase locking, but firings do not necessarily occur on every cycle. At higher frequencies firings are intermittent, yet each has the same phase relationship.

Ear resonances

The resonant behaviour of the basilar membrane is not observed at the lowest audible frequencies below 50Hz. The pattern of vibration does not appear to change with frequency and it is possible that the frequency is low enough to be measured directly from the rate of nerve firings.

At its best, the ear can detect a sound pressure variation of only 20 micropascals RMS and so this figure is used as the reference against which sound pressure level is measured. The sensation of loudness is a logarithmic function of sound pressure level and consequently a logarithmic unit – the decibel – is used in audio measurement.

The dynamic range of the ear exceeds 130dB, but at the extremes of this range, the ear is either straining to hear or is in pain. The frequency response of the ear is not at all uniform. It changes with sound pressure level.

The subjective response to level is called loudness and is measured in phons. The phon scale and the sound pressure level scale coincide at 1kHz, but at other frequencies the phon scale deviates because it displays the actual sound pressure levels judged by a human subject to be equally loud as a given level at 1kHz.

Figure 4 shows the so-called equal loudness contours which were originally measured by Fletcher and Munson and subsequently by Robinson and Dadson. Note the irregularities caused by resonances in the meatus at about 4kHz and 13kHz.

Usually, people's ears are at their most sensitive between about 2kHz and 5kHz. The generally accepted frequency range for high quality audio is 20Hz to 20 000Hz.

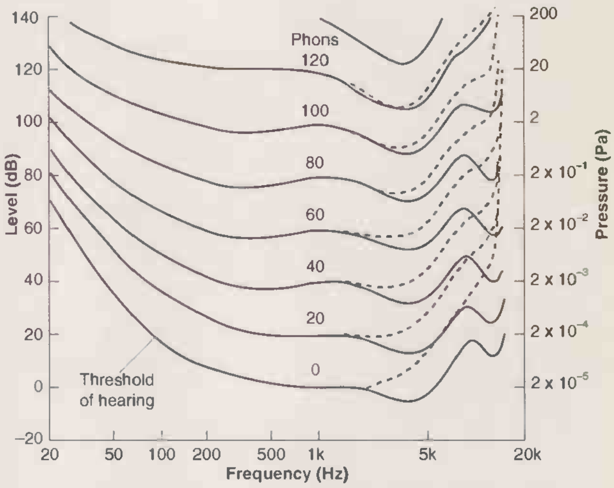


Fig. 4. Contours of equal loudness showing that the frequency response of the ear is highly level dependent. The solid line is typical of someone aged 20 and the dashed line a 60-year old.

Where did the bass go?

The most dramatic effect of the curves of Fig. 4 is that the bass content of reproduced sound is disproportionately reduced as the level is turned down.

If an adequately powerful yet high quality reproduction system is available the correct tonal balance when playing a good recording can be obtained simply by setting the volume control to a level that sounds natural. Many musical instruments and the human voice change timbre with level and there is only one level that sounds correct for the timbre.

A further consequence of level-dependent hearing response is that recordings mixed at an excessively high level will appear bass-light when played back at a normal level. Such recordings are more a product of self-indulgence than professionalism.

Figure 5 shows an uncoiled basilar

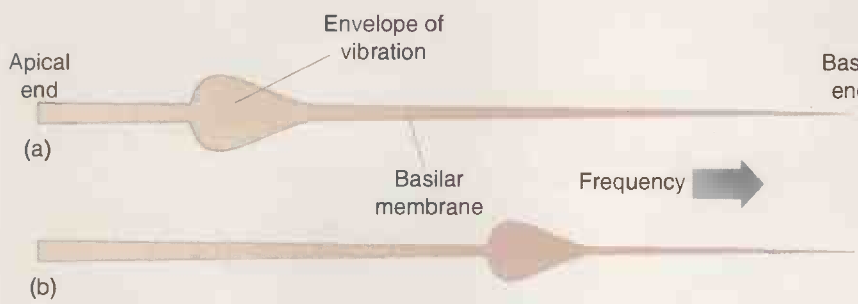


Fig. 5. Basilar membrane symbolically uncoiled. This shows that when the single frequency at a) is changed, as in b), the peak of the envelope moves.

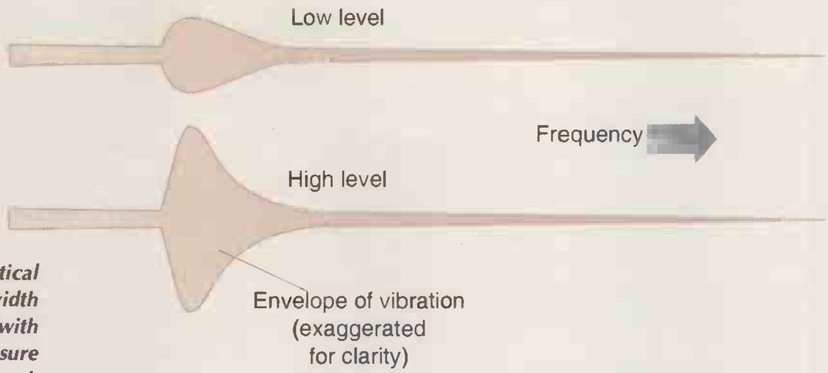


Fig. 6. Critical bandwidth changes with sound pressure level.

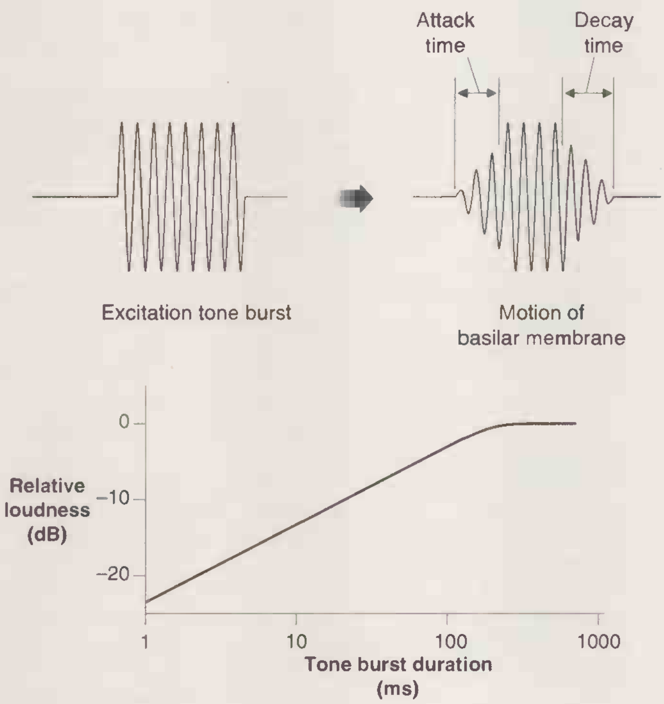
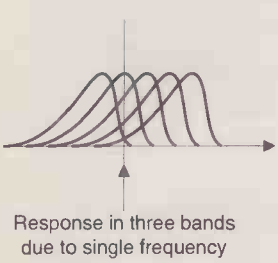
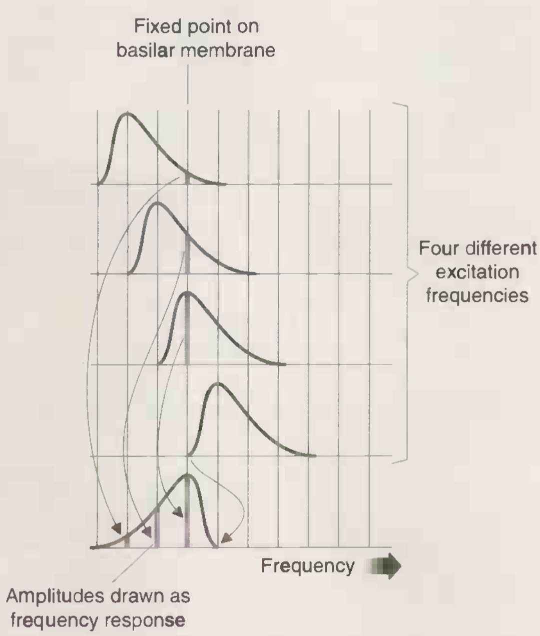


Fig. 7. The resonant nature of the ear prevents accurate response to a tone burst, a). In b), response is seen to increase during the tone burst as resonance builds up.

Fig. 8. If the ear behaved like a fixed filter bank, the filter response could be derived as shown in a). This theory does not hold though, because a single tone would cause response in several bands, b).





membrane with the apex on the left so that the usual logarithmic frequency scale can be applied. The envelope of displacement of the basilar membrane is shown for a single frequency at **a)**. The vibration of the membrane in sympathy with a single frequency cannot be localised to an infinitely small area. Nearby areas are forced to vibrate at the same frequency with an amplitude that decreases with distance.

Note that the envelope is asymmetrical because the membrane is tapering and because of frequency dependent losses in the propagation of vibrational energy down the cochlea.

If the frequency is changed, as in **b)**, the position of maximum displacement will also change. As the basilar membrane is continuous, the position of maximum displacement is infinitely variable allowing extremely good pitch discrimination of about one twelfth of a semitone which is determined by the spacing of hair cells.

In the presence of a complex spectrum, the finite width of the vibration envelope means that the ear fails to register energy in some bands when there is more energy in a nearby band. Within those areas, other frequencies are mechanically excluded because their amplitude is insufficient to dominate the local vibration of the membrane.

The Q factor of the membrane is responsible for the degree of auditory masking, defined as the decreased audibility of one sound in the presence of another.

The term critical bandwidth is used in psychoacoustics to describe the finite width of the vibration envelope. The envelope of basilar vibration is a complicated function. It is clear from the mechanism that the area of the membrane involved will increase as the sound level rises. **Figure 6** shows the bandwidth as a function of level.

Transform theory teaches that the higher the frequency resolution of a transform, the worse the time accuracy. As the basilar membrane has finite frequency resolution measured in the width of a critical band, it follows that it must have finite time resolution. This also follows from the fact that the membrane is resonant, taking time to start and stop vibrating in response to a stimulus.

There are many examples of this. **Figure 7a)** shows the impulse response and Fig. 7**b)** shows the perceived loudness of a tone burst increases with duration up to about 200ms due to the finite response time.

The ear has evolved to offer intelligibility in reverberant environments. It does this by averaging all received energy over a period of about 30ms. Reflected sound that arrives within this time is integrated to produce a louder sensation, whereas reflected sound arriving after that time can be temporally discriminated and is perceived as an echo.

A further example of the finite time discrimination of the ear is the fact that short interruptions to a continuous tone are difficult to detect. Finite time resolution means that masking can take place even when the masking tone begins after and ceases before the masked sound. This is referred to as forward and backward masking.

Some treatments of human hearing liken the basilar membrane to a bank of fixed filters each of which is the width of a critical band. The frequency response of such a filter can be deduced from the envelope of basilar displacement as has been done in **Fig. 8**.

The fact that no agreement has been reached on the number of such filters should alert the suspicions of the reader. The fact that a third-octave filter-bank model cannot explain pitch discrimination some thirty times better is another cause for doubt.

The response of the basilar membrane is centred upon the input frequency. No fixed filter can do this.

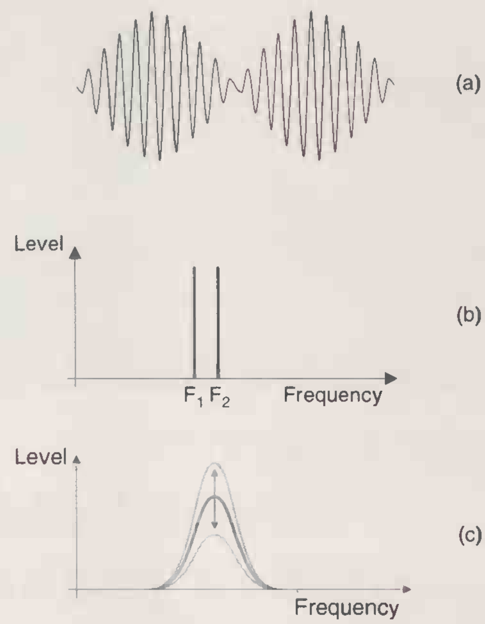
However, the most worrying aspect of the fixed filter mode, is that according to Fig. **8b)**, a single tone would cause a response in several bands which would be interpreted as several tones. This is at variance with reality. Far from masking higher frequencies, we appear to be creating them!

Figure 9 shows an electrical signal **a)** in which two equal sine waves of nearly the same frequency have been linearly added together. Note that the envelope of the signal varies as the two waves move in and out of phrase.

Clearly the frequency transforms calculated to infinite accuracy is that shown at **b)**. The two amplitudes are constant and there is no evidence of the envelope modulation. However, such a measurement requires an infinite time. When a shorter time is available, the frequency discrimination of the transform falls and the bands in which energy is detected become broader.

When the frequency discrimination is too wide to distinguish the two tones as in **c)**, the result is that they are registered as a single tone. The amplitude of the single tone will change from one measurement to the next because the envelope is being measured.

The rate at which the envelope amplitude changes is called a beat frequency. This beat frequency is not present in the input signal. Beats are an artefact of finite frequency resolution transforms. The fact that human hearing produces beats from pairs of tones proves that it has finite resolution.



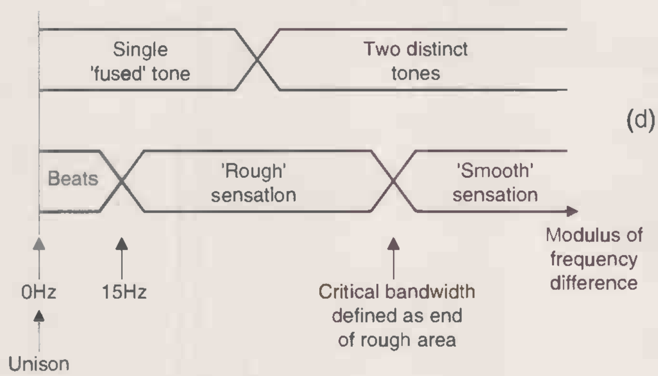


Fig. 9. Section a) shows two sine waves of similar frequency added and b) shows the waveform's spectrum to infinite accuracy. Assuming finite accuracy, only a single frequency is distinguished, c), whose amplitude changes with the envelope of a), giving rise to beats. Perception of a two-tone signal as frequency difference changes is shown in d).

Measurement of when beats occur allows measurement of critical bandwidth. **Figure 9d)** shows the results of human perception of a two-tone signal as the frequency difference changes. When it is zero, described musically as unison, only a single note is heard.

As the difference increases, beats are heard, yet only a single note is perceived. The limited frequency resolution of the basilar membrane has fused the two tones together. As it increases further, the sensation of beats ceases at 12-15Hz and is replaced by a sensation of roughness or dissonance.

The roughness is due to parts of the basilar membrane being unable to decide the frequency at which to vibrate. The regenerative effect may well become confused under such conditions. The roughness persists until the frequency difference has reached the critical bandwidth. Beyond this bandwidth, two separate tones will be heard because there are now two discrete basilar resonances. In fact this is the definition of critical bandwidth.

My next article on this topic continues with how the human hearing mechanism senses direction.■

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Input capacitance	12pF if oscilloscope i/p is 20pF
Compensation range	10-60pF
Working voltage	600V DC or pk-pk AC

Switch position 'Ref'
Probe tip grounded via 9MΩ, scope i/p grounded

543

Douglas Self tackles the vital subject of relay muting and protection in audio amplifiers. He reveals some little known secrets of relay operation and shows how to make them switch faster than is apparently possible. As a bonus, there's an upgrade to his precision preamp.

Muting relays

Most power amplifiers incorporate an output relay that not only provides muting to prevent transients reaching the loudspeakers, but also protection against destructive DC faults.

Loudspeakers are expensive, and no amplifier should ever be connected to one without proper DC-offset protection. This applies with particular force to experimental amplifiers.

Sensible preamplifiers – ie those with AC-coupled outputs – do not require DC protection, but the muting of thumps is no less important. Electronic switching at preamp outputs is feasible, but still presents technical challenges if high standards of linearity are to be combined with a reasonably low output impedance.

Electronic output switching is impracticable at power amplifier signal levels; however, if the amplifier is powered by a switch-mode supply, then turning it off is an option if positive and negative rails can be relied upon to collapse quickly and symmetrically

Protection circuit operation

Basic functions of a power-on thump elimination and DC protection circuit are as follows:

- Delay relay pull-in until amplifier turn-on transients are over.
- Drop out relay as fast as possible when AC power is removed.
- Drop out relay as fast as possible when DC fault occurs.
- Drop out relay on excess temperature, etc. Speed non-critical.

Figure 1 is a block-diagram of a system to perform these functions. Since this is in part a protection system, simplicity and bullet-proof reliability are essential.

The main dynamic parameters of a relay are the pull-in and drop-out time. For this kind of application, the pull-in time is more or less irrelevant, as it is milliseconds compared with the seconds of the turn-on delay.

Relay contacts bounce when they close, but the duration of pull-in contact bounce is not important for this application.

All the relays I examined showed clean contact-breaking on drop-out, and this is essential for fast muting. **Table 1** gives details of three power-amp relays and the Fujitsu relay used in the Precision Preamp '96 article.[1]

The specifications for the P&B relay are very conservative. The example measured pulled-in at 72% of the must-operate voltage, and dropped out at 350% of the must-drop-out voltage. Likewise the real operating times are much less than those specified.

The critical parameter for audio muting is the drop-out time, for this puts a limit on the speed with which turn-off transients can be suppressed. It seems at first that the drop-out time must be solely a function of the relay design, depending on the force in the bent contact spring and the inertia of the moving parts. This is partly true, as mechanical factors set a minimum time, but that time is greatly extended by the normal relay-driving circuits.

Relay-on timing

The delay required at amplifier turn-on depends on the amplifier characteristics.

If there are long time-constants, and voltages that take a while to settle, then the muting period will have to be extended to prevent clicks and thumps. Five seconds is probably the upper limit before the delay gets irritating; one second is long enough for a silent start-up with most conventional amplifiers.

This delay function can be performed in many ways, but there are a few points to consider. The tolerance on the length of the turn-on delay is not critical, and an *RC* time-constant is quite adequate to define it.

It is convenient – and significantly cheaper – to run the relay control circuitry directly from the main HT rails rather than creating regulated sub-rails or extra windings on the mains transformer. The emphasis is therefore on discrete transistor circuitry.

Figure 2 shows the relay control system I used in the Precision Preamplifier '96. Note that there was an error in the original diagram that is corrected here.

Capacitor C_{224} charges through R_{211} until D_{207} is forward-biased and Tr_{205} turns on. This turns on Tr_{206} and energises the relay; the extra current-gain of Tr_{206} enables the timing circuitry to run at low power. The on-timing delay here is 2 seconds.

A series dropper resistor for the relay is usually required; here it is R_{218}. The highest voltage relay-coil available is usually 48V, though 24V is more common, and power amplifier rails are often much higher than this.

The reverse diodes across the relay coils prevent Tr_{206} being damaged by the inductive spike created when the coil is suddenly de-energised. For relays of the size used in power amplifiers, signal diodes cannot cope

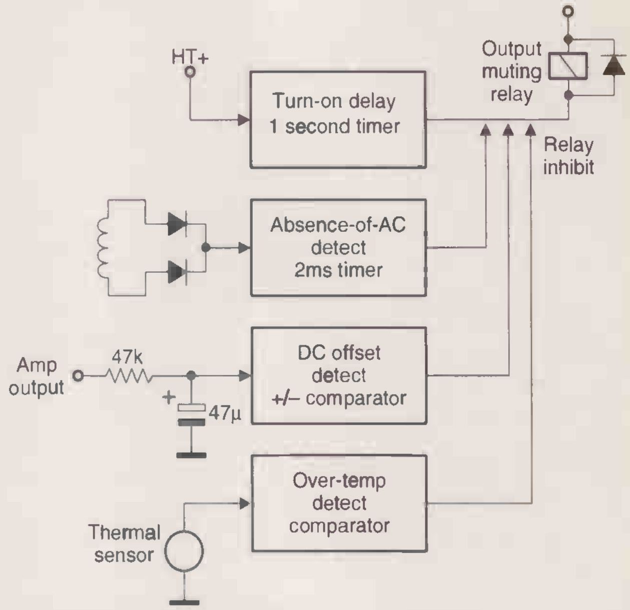

Fig. 1. Block diagram of a relay control system.

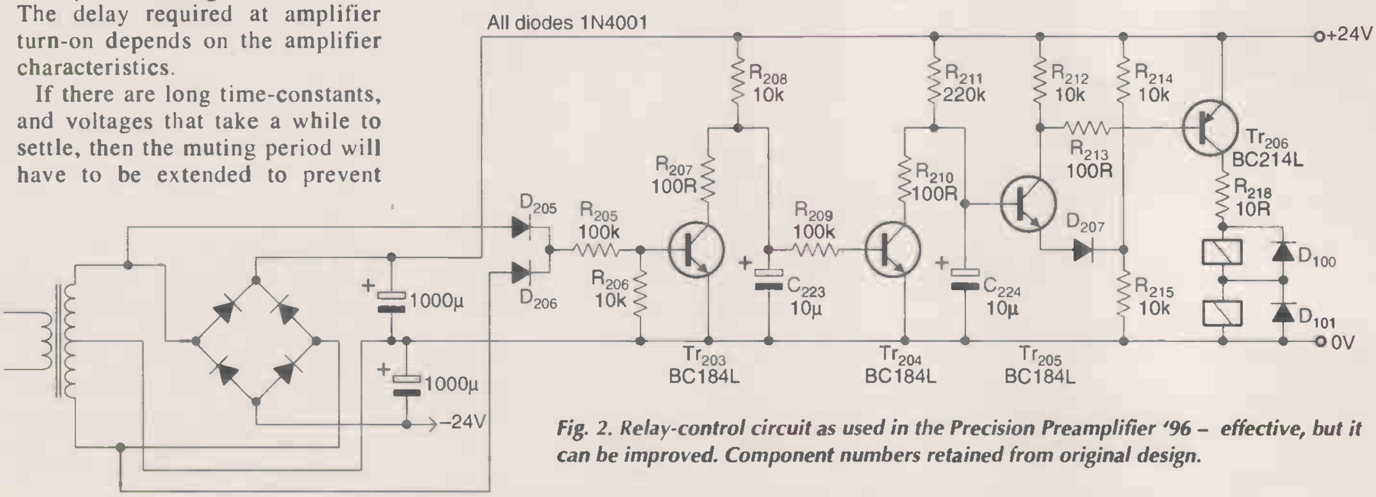

Fig. 2. Relay-control circuit as used in the Precision Preamplifier '96 – effective, but it can be improved. Component numbers retained from original design.

Table 1a. Relay specifications as presented by their manufacturers.

	P&B	Oko	Schrack	Fujitsu
Nominal voltage	24V	18V	12V	12V
Must-operate voltage	18V			8.4V
Drop-out voltage	2.4V			1.2V
Coil resistance	660Ω			320Ω
Coil inductance	0.55H			
Pull-in time maximum	15ms			5ms
Pull-in time typical	9ms			
Drop-out time maximum	10ms			3ms
Drop-out time typical	7ms			

Table 1b. Measured relay specifications.

	P&B	Oko	Schrack	Fujitsu
Operate voltage	13V	13V	7V	6V
Drop-out voltage	8.5V	6.5V	2.5V	2V
Pull-in time	14ms	10ms	10ms	2.7ms
Drop-out time	1.0ms	1.3ms	2.4ms	1.2ms
Diode drop-out time	5.4ms	6.9ms	11ms	4.2ms
27V-clamp drop-out time	1.8ms	2.4ms	2.7ms	1.3ms

*P&B is Potter and Brumfield

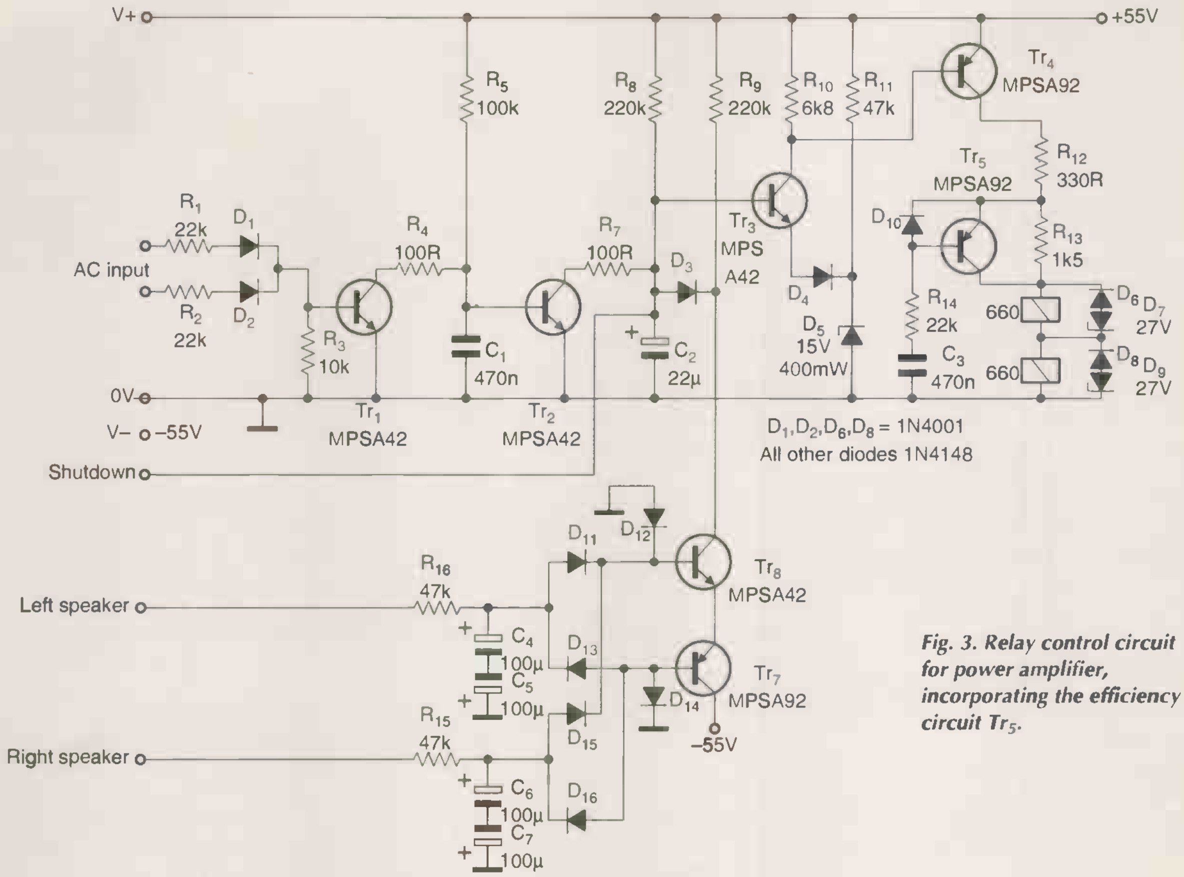

Fig. 3. Relay control circuit for power amplifier, incorporating the efficiency circuit Tr_5.

with the stored energy and the 1N4001 type should be used.

Off timing criteria

The relay drop-out must be as fast as possible. If a relay is powered directly from the supply rails, then it will drop out eventually as the rails collapse, but this will be far too slow to catch turn-off noises.

The drop-out voltage may well be less than a third of the pull-in voltage, and this slows things down even more. A specific fast off circuit is required, and there are several ways to achieve this.

Mechanical detection. This is a mains switch that closes or opens a control circuit before the mains power contacts are opened. It could give perfect relay operation, but I am not aware that any such mains switch has ever been produced.

Detecting the loss of DC supply. This technique involves a subsidiary supply rail with a small reservoir capacitor. When the mains is switched off, the capacitor discharges quickly and either removes the relay power directly, or resets the turn-on delay timer. The latter is usually easier to implement.

This method is inherently slow, because the relay-off threshold must be below the ripple troughs. Therefore in the worst case, an entire half-cycle of mains must pass before the capacitor becomes fully discharged, so the delay may be 10ms.

In practice there are component tolerances to be allowed for, and the threshold must be set low enough to prevent spurious operation if the mains voltage is below normal. It is usually prudent to ensure circuitry works with mains down to at least –20%. This extends the minimum delay to about 16ms. The reservoir capacitor will have a large ripple voltage across it, and its ripple-current rating must be carefully observed.

Detection of loss of AC. Detecting the loss of AC supply, as opposed to the rectified DC, is potentially quicker as there is no reservoir capacitor to discharge before the circuit operates. An AC waveform is effectively appearing and disappearing every half-cycle, so the circuit must distinguish between the zero-crossings that occur every 10ms, and genuine loss of power.

AC loss detection

The most straightforward method of AC-loss detection exploits the fact that properly-defined zero-crossings are very brief; all that is required is a timer that will not complete and drop out the relay until a period greater than the width of the zero-crossing has expired. This delay can readily be reduced to less than 1ms.

Referring to Fig. 2, Tr_{203} is normally held firmly on by the incoming AC, via D_{205} on positive half-cycles and D_{206} on negative ones, and thus keeps C_{223} fully discharged.

At the zero-crossings, Tr_{203} has no base drive and turns off, allowing C_{223} to begin charging through R_{208}. If the absence of base drive persists beyond the preset period, which means the AC has been interrupted, then C_{223} charges until Tr_{204} turns on and rapidly discharges the main timing capacitor C_{210} through R_{207}, dropping out the relay.

When a relay is driven by a transistor, it is standard to put a reversed diode across the coil. Without it, abrupt turn-off of current causes the coil voltage to reverse, driving the collector more negative. For the relays here, the worst spike measured was –120V, which is enough to destructively exceed the V_{ceo} of most transistors.

This apparently innocent, and indeed laudable practice of diode protection conceals a lurking snag; drop-out time is hugely increased by the reversed diode. It is roughly five times longer, which is very unwelcome in this particular application. This is because the diode gives a path for current to circulate while the magnetic field decays.

This is a good point to stop and consider exactly what we are trying to do: the aim is not to totally suppress the back-EMF but rather to protect the transistor.

If the back-EMF is clamped to about –27V by a suitable Zener diode in series with the reverse diode, the circulating current stops much sooner, and the drop-out is almost as fast as for the non-suppressed relay.

In general, drop-out is speeded up by a factor of about four on moving from conventional protection to Zener clamping. For the relays examined here, a 500mW Zener appeared to be adequate.

Preamp enhancement

The preamp relay controller can be improved upon; it works well under most circumstances, but it could be faster. Testing showed that the delay between loss of AC and the relay power being removed could be as long as 17ms, depending slightly on the phase of the mains when it was cut. The relay drop-out time was 5ms giving a total of 22ms before the preamp output is muted.

The following circuit improvements were made to speed up relay drop out.

The on-timing reference divider $R_{214,215}$ is replaced with a 15V Zener diode. This sharpens up the relay pull-in, making a more 'precise' click. It also prevents the voltage on C_{224} rising beyond that required to turn on Tr_{205}; discharging it when the time comes is therefore quicker.

Base drive to Tr_{203} is increased by reducing R_{205} to 22kΩ. This defines the zero-crossing as twice as narrow, allowing the time-constant R_{208}-C_{223} which bridges this period to be made shorter. Capacitor C_{224} therefore starts discharging sooner after AC is lost.

Impedance of the zero-crossing time-constant R_{208}-C_{223} is increased by changing the values from 10kΩ-10µF to 100kΩ-470nF. This simultaneously reduces the time-constant mentioned in the previous paragraph. It is now possible to use a non-electrolytic timing capacitor, which reduces tolerances and makes the circuit more designable.

Base drive to Tr_{204} is increased to speed up the discharge of C_{224} by reducing R_{205} from 100kΩ to 100Ω.

Finally, a 27V Zener clamp is applied to each relay, as described above.

After these improvements, the electronic delay was reduced from 17 to 5.4ms; the total delay including contacts opening now was 9.5ms worst-case.

After adding Zener clamping to the relays this fell to 6.3ms worst-case, the average being 4.5ms; the improved circuit is four times faster.

These component changes can be simply retro-fitted to existing Preamp '96 circuit boards using **Table 2**.

Other relay functions

The extra protective functions of a power amp relay require OR-ing together several error signals for DC offset, temperature shutdown, etc.

If a DC fault occurs in a power amplifier, this typically means that the output slams hard to one of the rails and stays there. Assuming the loudspeaker does not suffer instantaneous mechanical damage, it will overheat after a relatively short period as the DC flows through it.

DC offset protection cannot prevent a loudspeaker hitting its mechanical limits, but it will stop it catching fire if the relay opens promptly. Once more, time is of the essence.

Usually, DC offset is detected by passing the amplifier output through an *RC* time-constant long enough to remove all audio, followed by a DC-detect circuit that responds to offsets of either polarity.

To allow a safety margin against false triggering on bass signals, I decided that the *RC* filter must accept full output at 2Hz without the detector acting. For example, if it triggers at ±2V, then for supply rails of ±55V there must be 29dB of attenuation at 2Hz; with a single pole this means a –3 db frequency about 0.07Hz.

This sort of low-pass filtering inevitably introduces a time delay; if the output leaves 0V and moves promptly to one of the rails, this will be 50ms with the circuit of **Fig. 3**.

Detecting offsets of either polarity requires a little thought. **Figure 4** shows a common circuit; a positive voltage turns on Tr_8 by forward biasing its base, while a negative voltage turns on Tr_7 by pulling down the emitter. The presence of DC is indicated by the collector voltage falling.

This solution is simple but highly asymmetrical, requiring either +1.05V or –5.5V to pull the collectors down to 0V. For positive voltages the stage is common-emitter with high voltage gain, but for negative ones it works in common-base with a lower voltage gain, set by the ratio of R_{16} and R_9. It is difficult to make this ratio large without R_{16} becoming too small and hence C_4 inconveniently big.

If you're unlucky – and chances are you will be – the offset will have

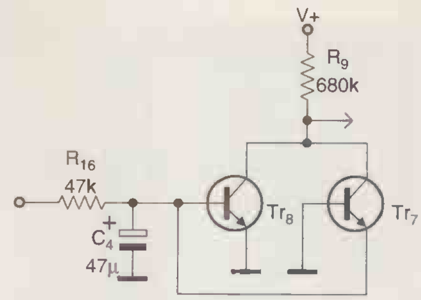

Fig. 4. Simple DC-detect circuit with asymmetrical thresholds at +1.05V and –5.5V.

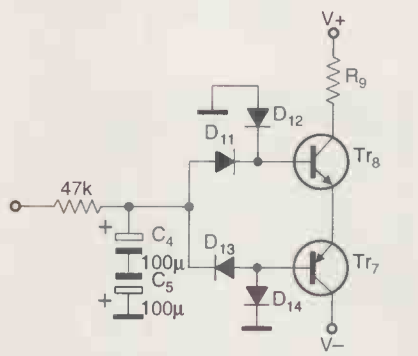

Fig. 5. Improved DC-detect circuit; fully symmetrical thresholds at ±2.4V. RC filter can cope with either polarity.

Table 2. Component revisions for the preamplifier.

	Old value	New value
R_{205}	100kΩ	22kΩ
R_{208}	10kΩ	100kΩ
C_{223}	10µF	470nF
R_{209}	100kΩ	100Ω
R_{215}	10kΩ	15V 400mW Zener
Relay	1N4001	1N4001 + Suppression 27V, 500mW Zener

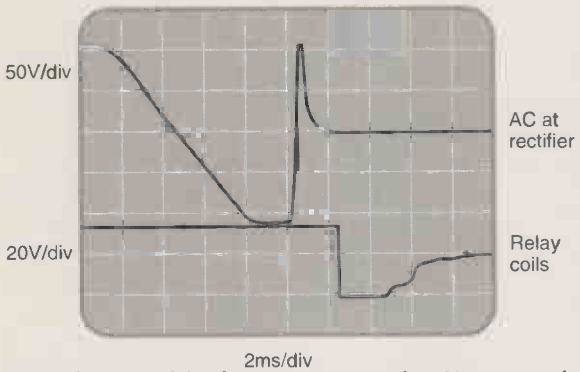

Fig. 6. Electronic delay for power-amp version. Upper trace is transformer secondary with the usual flat-topped mains waveform, lower is voltage across both relay coils.

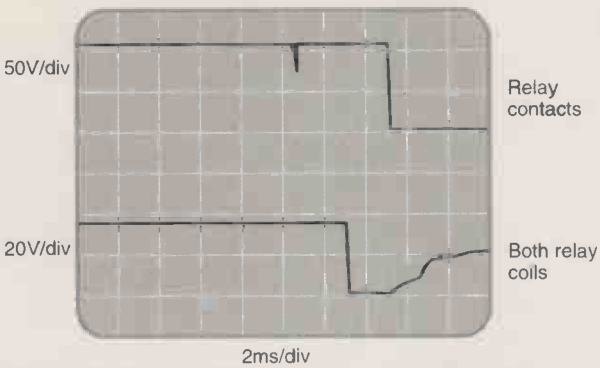

Fig. 7. Relay drop-out delay for power-amp version. Upper trace is contact timing, lower is voltage across both relay coils.

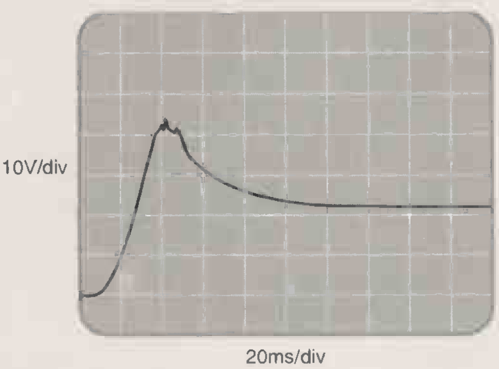

Fig. 8. Pull-in voltage across both relay coils with efficiency circuit added.

the wrong polarity for C_4, which will degrade if left reverse-biased for long periods. Two ordinary electrolytics back-to-back is the cheapest solution.

The improved DC detector in **Fig. 5** is fully symmetrical. Positive voltages turn on D_{11} and Tr_8; Tr_7 also conducts as its emitter is pulled up by Tr_8, while its base is held low by D_{14}. Negative voltages turn on D_{13} and Tr_7; Tr_8 conducts with its base fed by D_{12}. The threshold is now ±2.4V, as for each polarity there are two diodes and two base-emitter voltages in series. The higher threshold is not a problem as the typical amplifier fault snaps the output hard to one of the rails.

One exception to this statement is HF instability. If an amplifier bursts joyfully into HF oscillation, it will almost certainly show slew-limiting as well. This is unlikely to be very symmetrical so there will be a DC shift at the output.

The magnitude of this is not very predictable, but a 2.4V threshold will detect most cases. This should save your tweeters, though it may not save the amplifier from internal heating due to conduction overlap in the relatively slow output devices.

Figure 5 can be adapted for stereo simply by adding two more diodes, as in Fig. 3. Note that a positive offset on one channel and a negative one on the other – admittedly highly unlikely – do not cancel out; a fault is still signalled.

Power amplifier relay control

Figure 3 shows a power amp relay controller, designed for ±55V rails and 24V relays such as the P&B *T90* type outlined in Table 1. The main differences are the inclusion of DC offset detection and an efficiency circuit to minimise dissipation in the relays, which are now larger than in the preamp, and require more power.

The DC-detect circuit rapidly discharges on-timing capacitor C_2 through D_3 when Tr_8 collector goes low. An extra OR input for thermal shutdown acts via a series diode in the shutdown line.

The circuit now uses *MPSA42/MPSA92* transistors to withstand the higher supply voltages; as usual higher V_{ceo} means lower current-gain, which must be allowed for in the detailed design.

The electronic delay until coil switch-off averages 2ms, the timing being shown in **Fig. 6**. The AC was interrupted at centre screen, and a large positive-going off-transient can be seen just to the right. This is due to the leakage inductance of a large transformer.

The loss of AC cannot be detected until this transient decays to zero, so the delay is slightly extended. This was not a problem with the preamp version as it uses a small toroid with much less leakage inductance.

Figure 7 shows the relay coil voltage. At switch off it goes straight down through zero until clamped by the two Zeners at around –50V. This puts 105V on Tr_4 collector, which is no problem as it is rated at 300V V_{ceo}. The relay contacts open just as clamping ceases and the coil voltage returns slowly to zero. Drop-out time is 1.8ms, giving a total delay of 3.8ms.

Efficiency circuit

All relays have a pull-in voltage that is greatly in excess of that required to keep them closed. It is therefore possible to save considerable power by applying full voltage only briefly, and then reducing it to a level which is still safely above the maximum drop-out voltage.

From Table 1 there is plenty of scope for this. By comparing the specified and measured performance, you will see that the P&B relay can be trusted to pull in at 18V and not drop out above 8.5V.

The initial pulse is provided by Tr_5 and R_{13}. At switch-on, Tr_4 is off and Tr_5 does nothing. After the on-timing delay Tr_4 conducts, Tr_5's emitter is pulled up, and its base receives a pulse of current via R_{14} and C_3. Resistor R_{13} is shorted by Tr_5 and the relays get a voltage reduced only by R_{12}; see **Fig. 8**.

After 40ms, C_3 is fully charged and Tr_5 turns off; this is at least four times longer than the minimum pulse to pull-in the T90 relay, but may be adjusted to suit other types by altering C_3. Diode D_{10} protects Tr_5 at switch-off.

In Fig. 3, the initial voltage is 22V per relay and the holding voltage 12V, giving an initial power consumption of 1.85W, falling to 960mW long-term. The total power saving is just under a watt.

Running relays at a reduced holding voltage not only avoids the inelegance of consuming power for no good reason, but also speeds drop-out time by reducing the magnetic energy stored. It could be argued that such a power saving is negligible. In a big Class-A amplifier it might be, but it makes sense in modest Class-B amplifiers idling for much of the time – which is of course almost all of them. ■

Reference

1. Self, D. 'Precision Preamplifier '96,' *Electronics World*, Jul/Aug,Sept 1996.

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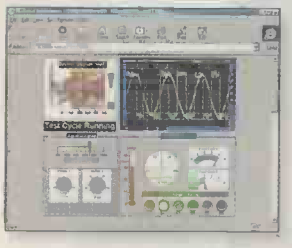

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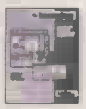

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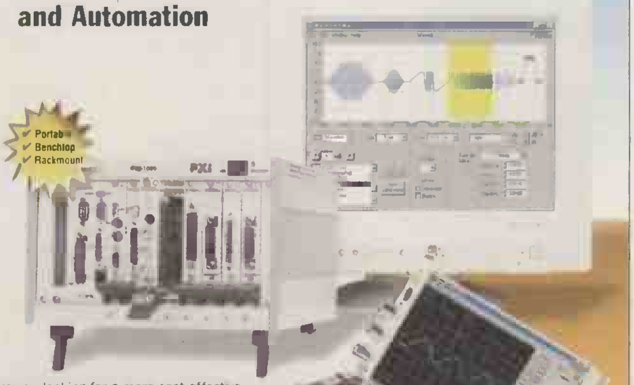

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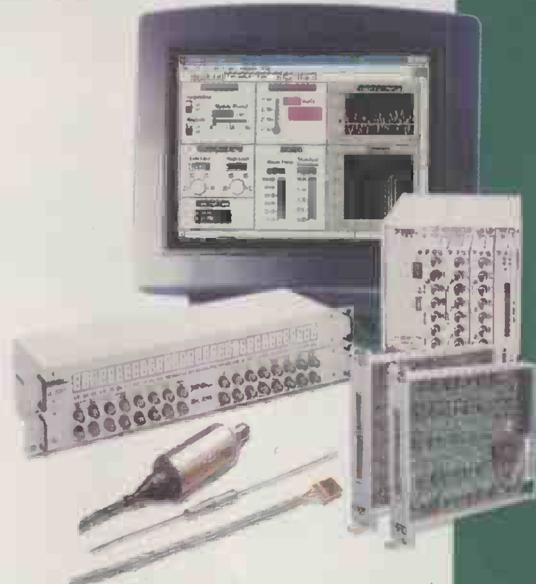

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Like its modern counterparts, this early attempt at simulation was difficult and expensive to implement – and it didn't always do what its creator wanted it to.

The route to simulation

Choosing a circuit simulation package is complex and making a mistake is costly. Starting with a review of Workbench and an outline of what the problems are, Rod Cooper's guide is essential reading for all first time buyers.

Review subjects

This first review covers ***Electronic Workbench*** version 5.12, whose maker is IIT Ltd of Canada. Workbench's UK supplier is Adept Scientific plc, tel 01462 480055.
Electronic Workbench's price is £199.

Subsequent reviews cover ***CircuitMaker Pro*** from Microcode Engineering, ***Tina Pro*** from Designsoft, Labcenter's ***Lisa***, which is part of Proteus IV, and ***Pulsar and Analyser*** from Number One Systems, which are modules from the Easy PC package.

My recent series 'The route to pcb-cad' looked at CAD programs specifically written for designing printed circuits using a PC. The series comprised a set of reviews and was intended to help first-time buyers of such a package make the right choice.

At the end of this series, I pointed out that if you went to the trouble of generating a net list for autorouteing a pcb, it was a natural progression to use the same net list for running a simulation of the circuit.

This set of reviews is intended primarily for those of you who, having bought your pcb-cad program, wish to take the almost irresistible step of adding a compatible budget-priced simulator.

For this reason, programs covered are either already adjuncts of the pcb-cad products reviewed previously, or they have an easy – or at least compatible – net-list interface. In addition though, these reviews will be of interest to those who just want to buy a stand-alone simulator.

Selecting what to review

People starting in the field of simulation require a program that is not too difficult to operate, so I have picked only simulation programs that I found relatively easy to use.

The second criterion is that they should not necessarily be equipped with every feature and every simulation under the sun. Rather they should have a number of basic simulations, with sufficient scope for broad, non-specialised use.

The third criterion is that, for the range of simulations offered, the programs should have a modest price tag. Programs that I felt did not meet all three criteria have not been included.

This review sets out to indicate what attributes the programs offer – or what is lacking, as the case may be. This gives the potential purchaser the chance to pick out a specific program that will suit the type of simulations he wants to run. It is then up to him to check through the demo version of the program to see if he likes it, and if it will do what he wants in his own field.

Spice, but you wouldn't know it

Spice in its raw form takes a lot of time and effort to learn properly. Most of the circuit simulators reviewed here are based on Spice. However, the GUI provided with some of them is so well developed that for most of the time you would not realise you were using Spice.

You do not need a full grasp of Spice in order to use the packages described here, but there are occasions when a basic grasp of Spice theory does help.

Spice has some peculiarities, which cause the first-time buyer some unnecessary dismay. Typical of these was a program returned under a 30-day trial period.

The buyer wanted to simulate transformers. He noticed the Spice requirement that 'all nodes in a circuit had to have a path to ground'. In the case of transformers, this seemed nonsensical, so he returned his software.

All he had to do to satisfy Spice though was to emulate real life and put in the very high leakage resistances that all transformers have. Even easier, if he had read a little about Spice, he would have spotted a device called RSHUNT, which did this for him automatically, for all circuit nodes.

Another typical example of difficulties with Spice is in simulating two or more series-connected capacitors. The connection between them will be isolated from ground if the capacitor model is ideal. Also, getting oscillators to work can be a problem.

However, there are ways and means of getting round these difficulties. Indeed, for many Spice-induced obstacles, there is often a known work-around, so it's usually just a question of reading the available literature.

Not all simulators use Spice. One of the advantages of using Spice-based simulators is that you can transfer Spice net lists relatively easily between programs.

Also, Spice data is available from device makers that can be added to the model library. These may be factors that influence your purchase.

Why bother to simulate?

The cynical would say that it's just so much less effort, and more convenient to turn on the PC and play around with a simulation program, rather than go down to the lab and set about plugging in 'scopes and signal generators.

But there are several advantages of simulation compared to bench testing. The big attraction is the potential to save money.

Manufacturers see simulation as an inexpensive way of detecting design errors before any money is spent making and bench-testing even the prototype hardware. Investment in test-gear can be reduced – although it may be unwise to go too far down this road.

Educationalists also see simulators as a way of saving money. For example, when teaching students it is possible to test circuits without destroying any expensive components – either deliberately, or in error. It is possible to fully demonstrate a design in the classroom, without incurring the cost of building it in an expensive laboratory.

There are also some good non-financial reasons for using simulation. The range of tests in a simulator is usually much greater than that available in the average workshop. For example, simulator signal generators can typically be set to run from 1Hz to 1GHz in one sweep, the output amplitude can be adjusted from 0 to plus or minus a gigavolt, and duty cycle from 1 to 99% in a choice of waveforms. It would be difficult or impossible – and certainly expensive – to achieve such versatility in a laboratory with real instruments.

Another reason for simulating is that some simulator operations just have no easy, inexpensive real-world equivalent, such as the word generator in digital analysis. Also, the purity of signal sources in simulation is not compromised by physical constraints.

Often a simulator can avoid doing tedious mathematical calculation. Another advantage is that simulators can produce test results quicker than by the conventional tests on real instruments. However, this is not always true.

Sometimes a circuit that refuses to run on a simulator can consume disproportionate amounts of time, not because it is inherently difficult, but due to some peculiarity of the simulation program. It is very easy to fall into the trap of spending more time getting the simulator to comply than it would take doing a real test on the bench.

Against all these advantages, there is the initial difficulty of learning how to operate the program correctly. A simulator has to be learnt just like any program and the degree of difficulty experienced is wide and varied.

Spice outline

Spice was invented by the University of California at Berkeley in the USA as a means of simulating the function of IC designs. The name Spice is an acronym for Simulation Program (with) Integrated Circuit Emphasis.

Like the decibel, Spice was far too useful a concept just to be left to itself. The concept has been expanded and adapted to cover many other devices – many not connected at all with ICs.

Spice originated in the early seventies, so could be considered an old program. But it has been kept up-to-date by extensive up-grading, becoming the *de facto* standard in the process. If you want to find out more about Spice, especially the benefits, restrictions and incompatibilities, or if you want to edit your own Spice models, there are several books that go into Spice in depth. I will mention them in a subsequent article.

Only a few years ago, learning Spice was a difficult process. But with the present generation of programs, any engineer with a knowledge of real instruments and who is familiar with Windows should be up and running in an hour or so. Because of this, most firms are finding that the need for a 'simulation specialist' is much reduced or non-existent.

Reliability issues

Simulation is not an easy way out of testing a new circuit. On the contrary you have to watch your step.

A view taken by many designers is that simulation does not replace bench testing, but complements it. Used with discretion, a simulator can save a lot of bench-testing time. But now and then it can give a misleading result.

As the art of modelling components for simulation has progressed, such odd results have decreased. Even in a budget-

Net-lists or schematic capture?

Some users still prefer to type out a net list in order to make a simulation, instead of using schematic capture, maintaining that this gives a better insight into what is actually taking place. This may be true, but proponents of schematic capture regard such insight as a lengthy diversion from the main task of getting the simulations done.

Unless you have good typing speed, the net list is a slow and cumbersome method. It suffers from being abstract and unintuitive. I read in one net-list-entry program manual that 'it takes only ten minutes to type out the circuit as a net list'. Using exactly the same circuit, it took just 90 seconds to schematic-capture it in CircuitMaker.

The net list does not have universal acceptability – show a group of engineers a net list and the chances are that some of them will not understand it.

Designers have to ask themselves if they really want an insight into the workings of their simulator via the net list. It may all be very interesting, but is it relevant to the job? For some people, such as teachers, perhaps it will be, but for commercial designers who simply use simulators as just another tool, it may be merely a distraction.

There are better things to learn in an age of information overload than the particular way your work-tools operate. Therefore, I make no apology that the reviewed simulators all use schematic entry.

However, a simulator that uses schematic capture will stand or fall on the quality of its schematic drawing system. Some are excellent, others serviceable, but I believe none in this review is difficult to use.

priced simulator of the present generation, the results in the majority of cases are usually acceptable.

Another example of a source of error is that, at this price level, there is usually no compensation for strays introduced by the PCB design. You can of course make a guess what they are, based on experience, and add them into the circuit.

Now for the first of the reviews.

Review 1:
Electronic Workbench V5.12

This is the latest version of a popular program which has sold 100 000 copies to date.

Workbench features a schematic drawing and capture program coupled with a mixed-mode simulator. The two sections are fully integrated, i.e. you don't have to transfer out of the schematic program to work the simulator. For practical purposes, it is one program.

Simulations can be displayed on the same screen as the schematic. There is no pin limitation. The simulator, based on Spice 3F5, uses both the virtual instrument concept and conventional analyses. It is claimed to be 5 to 10 times faster than version 4.

The software comes on CD and installation is easy. Security is by registration number. Besides running on WIN95 and 98 and NT, there is a Windows 3.1 version. As Workbench V.5 is 32-bit, the Windows 3.1 version installs the W32s extension.

Recommended hardware is a 486 or later with 8MB RAM, or 12MB with Windows NT. Disk space needed is 20MB.

Documentation

The main documentation comes in four soft-back books, and there is a quick-reference card and an installation sheet as well.

The first book is a user guide that runs through the basic system and contains several tutorials. This is complemented by a larger technical reference book, which goes into more depth. The review copy contained references to the higher EDA version of Workbench, but sorting out what was applicable to version 5 was no problem.

A third, slimmer volume lists all the ICs and models available in the library, and a final booklet explains how to import and export net lists.

All this documentation is very well presented and readable. There is plenty of explanation *en route*, but no glossary. A chapter in the technical reference book is devoted to a basic explanation of how Spice works, which is useful reading if you are not conversant with it.

A useful 'Help' section in the program complements the books.

Schematic capture

For an insight into the schematic-capture part of this program, refer to the review of Workbench version 4.0 already published in *EW* October 96. Briefly, this indicated that generally, the schematic drawing section was easy to use, flexible and versatile.

With its multiple parts bins, the schematic drafter is suitable for on-screen designing and experimentation. Although there have been several small improvements to the drawing section since V.4, the fundamentals have not altered much since this earlier review – with the big exception of the wiring-up system. Here there has been a significant change, as the auto-wirer has now been complemented with a manual system of wiring up the symbols. This removes one of the criticisms of earlier versions.

Like most autowirers, the one in Workbench is acceptable for simple circuit diagrams but tends towards complicated routes and untidiness on larger diagrams. For this reason, a manual method is be welcomed.

As before, some sleight of hand has to be developed during wiring up to avoid selecting a component for editing instead of connecting it. Both are done with the left mouse button and it is easy to mix the two up until some manual skill has been gained.

A few moments with the demonstration program will illustrate this. However, most existing users of the package will find schematic drawing in version 5.12 generally much easier.

The notional size of the working area of the drawing sheet is about 7 by 20in. The enlarged library holds more than 100 analogue and 200 digital symbols backed by 4000 Spice device models.

With this version of the program it is not possible to make up and add your own symbols. You can, however, add extra Spice models, but these must be assigned to a generic 'black box' symbol. You can also alter existing models if you are conversant with Spice.

As well as a Spice net list, Workbench

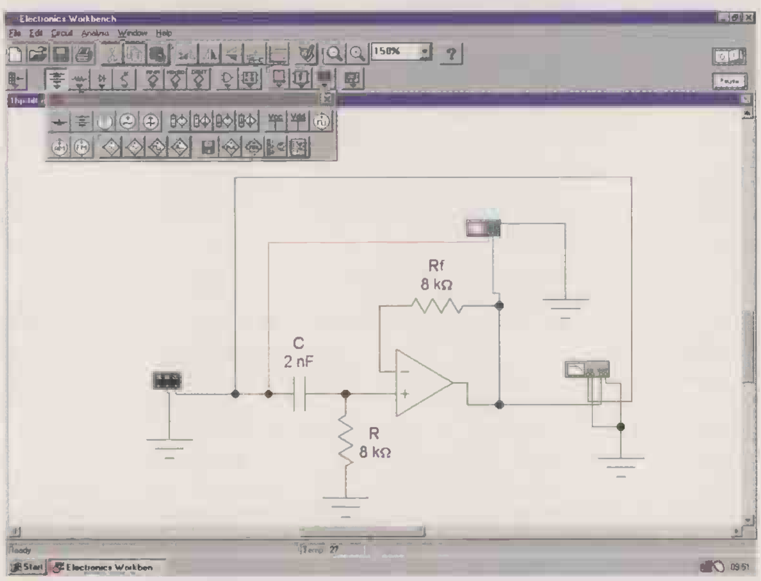

Fig. 1. A simple schematic, taken from Workbench's sample library, illustrating how the virtual instruments are wired into circuit and a typical open parts bin.

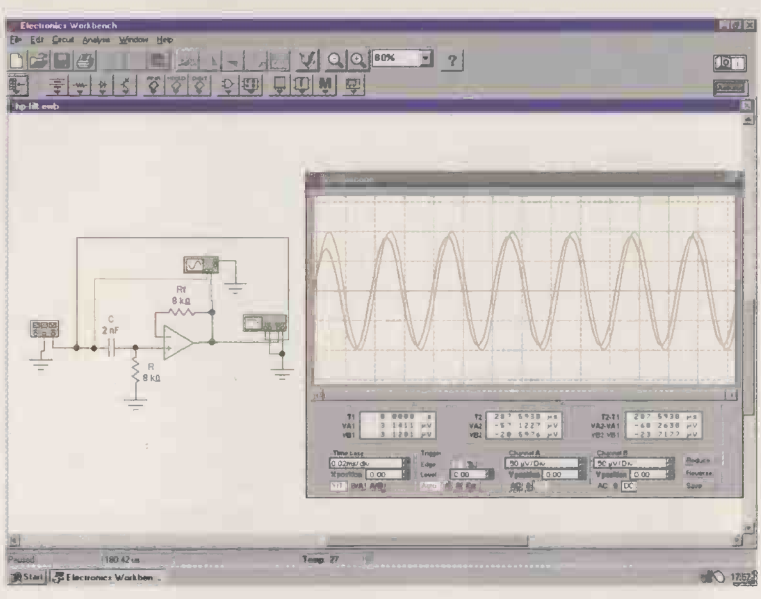

Fig. 2. The same circuit with the 'scope instrument expanded to show traces from the two points indicated on the circuit by red and blue wires.

schematic capture can export files to several PCB design programs, namely OrCad, Tango, Eagle, Protel, Layo1.

Simulation section

Simulation relies strongly on virtual instruments. These are made to resemble the operating panels of real instruments as far as it is possible on a PC.

The virtual instruments consist of function generator, twin-beam oscilloscope, multimeter, bode plotter, logic converter, word generator, and 16-channel logic analyser.

In addition to these main instruments, there are ammeters and voltmeters, and an impressive range of signal sources, such as AM and FM, and various voltage-controlled waveforms. They are all easy to insert into the schematic and to wire up, behaving just like regular component symbols, and are also pleasantly intuitive to set up and use.

The method of use for the main virtual instruments is to expand the icon to a more usable size when you want to run a simulation. It then appears in the form of an overlay on the schematic drawing, **Figs 1** and **2**. By dragging the instrument to one side of the diagram, to get a view of the diagram, it is possible to change component values, or the components themselves, while the simulator is running. This is a useful attribute.

Two points to note about the instruments. Firstly the 'scope has two levels of expansion for more detailed examination. Secondly, the bode plotter has no graduated plot. If you want to make an accurate measurement you have to run a conventional analysis instead.

In general, all these virtual instruments have a wide choice of parameters and plenty of adjustment range. The actual figures occupy too much space to list here but can be viewed in the product's promotional leaflet.

To complement the virtual instruments, there is a range of analyses, *viz.* noise, distortion, AC frequency sweep for gain and phase, Fourier, transient, and DC operating point. There is no plot of output or input impedance versus frequency – a notable omission.

A typical analysis graph for the circuit of Fig. 1 is shown in **Fig. 3**. The analyses are menu-driven in a clear, easily understood format and plain terminology.

To run an analysis of this type you specify the circuit node you want to sample, from such a menu. To do this, you modify the circuit diagram to show the nodes, again through a menu, then select a node. Fig. 3 shows just one node, the output, but you could add more if required. Unlike the virtual instruments, you can expand the analyses to full-screen if you wish.

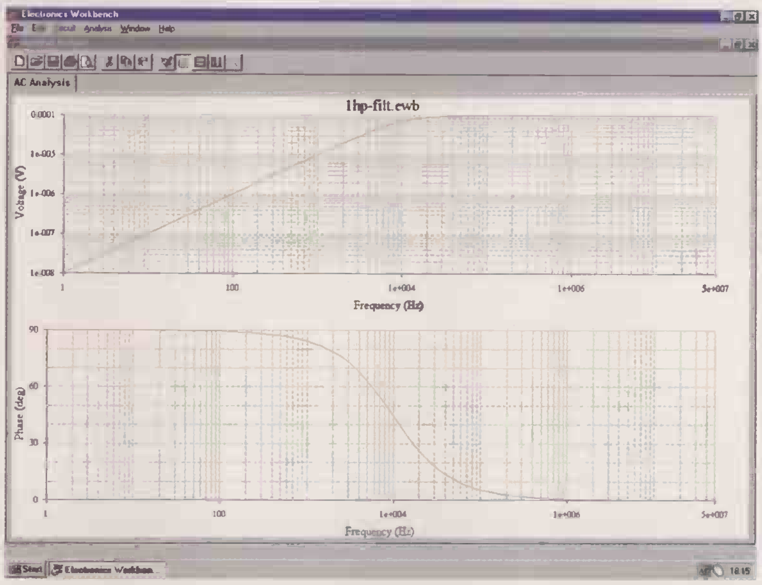

Fig. 3. An AC analysis of the circuit. Note particularly the method of calibrating the x axis.

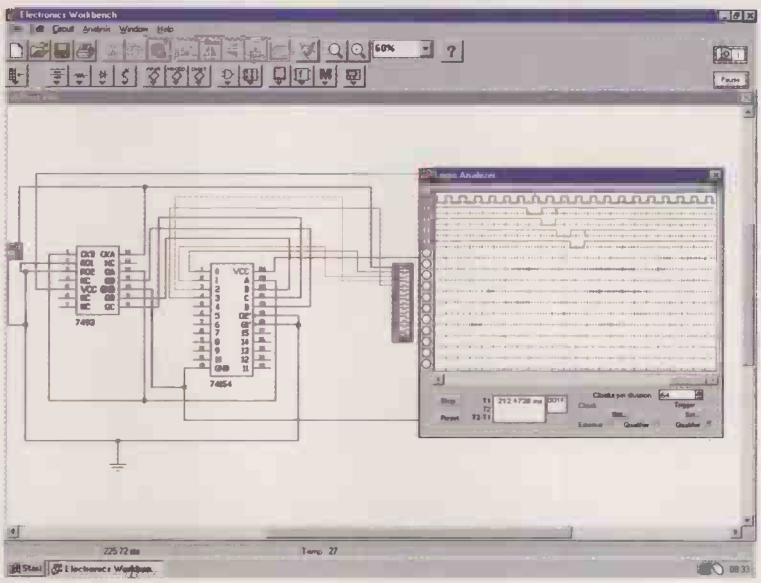

Fig. 4. The digital analyser in use showing how useful the coloured wire system is.

There is no probe tool as there is in Tina and CircuitMaker. The analyses are stored in a stack for ready reference so you can flip back to one if you wish to re-view it.

As with Workbench's predecessor, you can use a range of colours for individual wires on the more complex schematics to improve the readability of the diagram. This is a good, practical idea. Even better, you can transfer this colour scheme to the graphs. **Figure 4** shows this in practice on the digital analyser. Note that there is no glitch control – this is explained next month.

If you are interested in transformer simulations, for switch-mode supplies for example, then you will find Workbench's transformer modelling better than many others in this price bracket.

Version 5 retains its educational bias with features useful to teachers, such as fault injection. ■

Summary

Workbench's comfortable learning curve, assisted by the excellent combination of printed manuals and complementary help files, makes it an attractive purchase for a first-time buyer. With the few exceptions detailed above, the program is easy to run.

The package also has appeal as a quick check for the more experienced designer. The range and scope offered by the package is excellent for the price, the range of signal inputs being particularly good.

You should still check carefully that the parameters you want to simulate are included. For example there is no glitch control and no input/output impedance graphs. Also check that the results are presented to your liking and that the schematic symbols provided cover everything needed. There is no way of adding to them.

Considering the range of simulations, at £199 the program represents very good value for money.

Sensitivity demystified

Joe Carr explains not only how to decipher sensitivity figures for receivers, but also how to verify them for yourself.

Radio receiver specifications can be verified using test equipment and a few simple procedures. Such tests are made to evaluate receivers, to troubleshoot problems, and to verify performance. A number of receiver parameters are important, but perhaps the one that is most commonly discussed is sensitivity. Let's take a look at how these tests are done.

What is sensitivity?

Receiver sensitivity is a measure of how well a receiver will pick up very weak signals.

As with most engineering measurements, the notion of sensitivity is an operational definition. In other words, there are standard procedures that will yield coherent results by which different receivers – or the same receiver before and after repairs – can be compared.

Sensitivity is basically a game of SNR, i.e. signal-to-noise ratio. Or more properly, the signal-plus-noise-to-noise ratio, (S+N)/N. For every receiver or amplifier there is a basic noise level consisting of the noise produced external to the receiver and noise produced inside the receiver. Even a receiver with its antenna input terminated in a shielded matching resistor, rather than an antenna or signal generator, will show a certain amount of thermal noise.

One important consideration when making sensitivity measurements – or comparing receiver sensitivity specifications – is bandwidth. Thermal and other forms of noise are gaussian distributed over all possible bandwidths.

The value of the noise at any given instant depends on the bandwidth of the channel. For most receivers this means the IF selectivity bandwidth, although in some cases the audio bandwidth is less than the IF so that number would dominate.

Table 1 shows the thermal noise expected from a 50Ω resistor at various bandwidths. Always make sure that the bandwidths at which various sensitivity numbers are compared are the same.

Figure 1 depicts two different definitions of SNR. Basically, you can't hear signals down in the noise. The minimum discernible signal is operationally defined as the

R8500 communications receiver, photo courtesy ICOM.

Table 1. Thermal noise expected from a 50Ω resistor at various bandwidths.

BW (Hz)	Thermal noise (μV)
500	0.01
1000	0.014
1500	0.017
2000	0.02
2500	0.022
3000	0.025
3500	0.027
4000	0.028
4500	0.03
5000	0.032
5500	0.033
6000	0.035
6500	0.036
7000	0.037
7500	0.039
8000	0.04
8500	0.041
9000	0.042
9500	0.044
10000	0.045

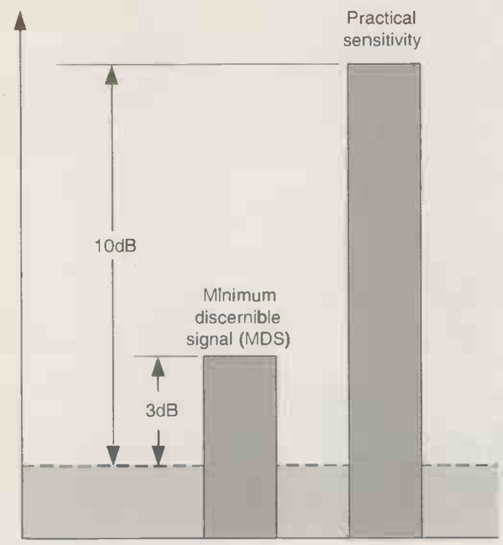

Fig. 1. When evaluating sensitivity, it's important to know how SNR is defined.

signal level that is the same as the noise floor, or the signal level that is 3dB above the receiver's noise floor. But that sensitivity is not all that useful for most applications.

I'm sure that there are people out there who can listen to a signal that is only 3dB above the noise floor. Most of us, however, require a higher SNR to be practical.

Although some definitions use 6dB, 12dB or 20dB, the standard for practical sensitivity is that it is the signal level that produces a 10dB SNR. This definition is found on most CW, AM and SSB receivers.

Signal Generator. The signal generator selected to make sensitivity measurements must have very high isolation figures. Most 'service grade' signal generators are useful for doing troubleshooting, but they are not satisfactory for making sensitivity measurements. The reason is that signal escapes around the cabinet flanges and control bushings. If you have a sensitive receiver or spectrum analyser then you can detect this unwanted signal.

Want to give it a try? Connect a shielded dummy load to the output of the signal generator, and turn the signal generator's output down to zero. Connect a whip or wire antenna to the receiver's antenna input, and then tune the receiver across the signal generator frequency with the RF gain cranked all the way up.

The signal generator should also have a calibrated output control. The correct calibrations are either dBm – i.,e. power decibels relative to one milliwatt in a 50Ω load – or microvolts.

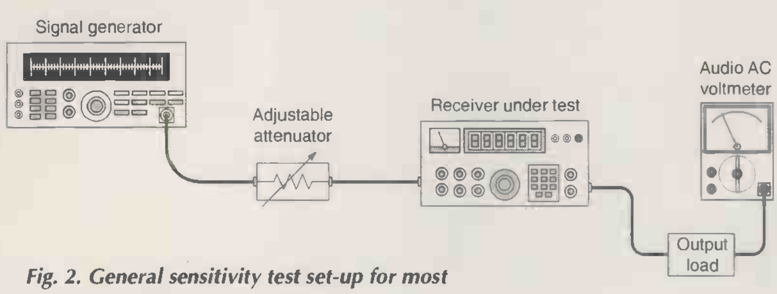

Fig. 2. General sensitivity test set-up for most receiver sensitivity measurements.

Some signal generators have an output meter that can set relative output, but can become 'calibrated' if a calibrated step-attenuator is connected between the output of the signal generator and the receiver under test. You can find the exact level if you can measure the high level output of the signal generator.

'Laboratory grade' signal generators may be beyond the means of many people, but there is a relatively vigorous market in used or surplus equipment. Sources of such equipment are listed on the world wide web. If you don't need the latest digitally synthesised signal generators, then you will be able to find good signal generators at low cost.

Test set-up

Figure 2 shows the test set-up for most receiver sensitivity measurements. The attenuator is optional, and may not be needed if the signal generator is adequately equipped with a good quality calibrated output attenuator. When measuring an AM receiver, set the signal generator modulation for 30 percent depth and 1000Hz.

An audio AC voltmeter is used to measure the receiver's output level. Ideally, the instrument should be calibrated in decibels as well as volts, and should have RMS reading capability.

The receiver must be correctly set up, or the measurement will be in error. In most test set-ups, the receiver's RF and AF gain controls are turned to maximum, and the squelch is turned off. Further, the automatic gain control, or AGC, must be either turned off, or in the case of some models, clamped with a DC level according to the manufacturer's directions.

Minimum discernible signal sensitivity. To make the minimum discernible signal measurement you need to find the signal level in dBm or μV that is 3dB above the receiver noise floor. To do this,

- Connect the equipment as in Fig. 2, and set the receiver and signal generator to the same frequency.
- Turn the signal generator output fully down to zero.
- Set the RF gain and AF controls to maximum. You may want to set the audio output control to a convenient level if you don't have a dummy speaker load.
- To make the measurement, you first measure the RMS value of the noise – i.e. 'hiss' – output on the AC voltmeter, and then increase the signal generator output level until the receiver output level increases by 3dB.

You can also determine the numerical value of the receiver noise floor by the same approach. Measure the output noise level, and then find the minimum discernible signal by the procedure above. The receiver noise floor level will be the same as the signal generator output level (less any attenuation in line).

Standard output conditions. A sensitivity specification used for consumer radio receivers uses a standard output approach. A typical receiver sensitivity specification might read, "xxμV for 400mW in an 8Ω load when modulated 30% by 1kHz."

The same equipment set up of Fig. 2 can be used for this measurement. A power of 400mW into an 8Ω load is the same as 1.789V RMS, which can be read on the AC voltmeter. I recommend using an 8Ω noninductive resistor for the load rather than the loudspeaker since the sound levels involved in the measurements are pretty annoying.

Adjust the signal generator output level for an RMS output voltage of 1.789V, and read the output level from the signal generator controls.

Full-power sensitivity. Some older receivers use the full-power sensitivity figure. This is the signal level that will produce the full rated audio output power.

Set the signal generator for 1kHz modulation with 30% depth. Tune the radio and signal generator to the same frequency, and crank up the output until the audio output power is at the full-rated power level – for example 400mW, 1W, or in the case of one radio I owned 7.5W. The signal level that produces this condition is the 'sensitivity' of the receiver.

10dB (S+N)/N test. The 10dB test method is the same as the 3dB minimum discernible signal method, except that the signal generator level is increased until the output is 10dB above the noise floor.

An alternative method is sometimes used on AM receivers:

- Set up the signal generator and receiver as discussed above.
- Set the output of the receiver to produce at least 0.5W audio output, or if the rated output power is lower than 1W set it for at least 50mW audio output power.
- Turn off the modulation. If the audio output drops at least 10dB then the signal generator setting is 10dB S+N/N level. If the level drops less than 10dB, then readjust the signal generator output level upward a small amount and try again.

On-site effective sensitivity test. This test is only done on-site where the receiver is installed. It is intended to get some idea of how well the receiver performs in its actual installed environment. **Figure 3** shows the test set-up.

- Measure the 10dB S+N/N sensitivity as discussed above (see Fig. 2 for set-up) and write down the figure.
- Connect the hybrid combiner, two-position coaxial switch, antenna and dummy load into the circuit.
- Set the switch to the dummy load and measure the sensitivity. It will be considerably worse than the 10dB sensitivity.
- Set the coaxial switch to the antenna, and again measure the 10dB sensitivity. It should be lower still.

The effective sensitivity is $SNR_{10dB}-(SNR_{LOAD}-SNR_{ANT})$. The figure $SNR_{LOAD}-SNR_{ANT}$ is the degradation factor. For example, suppose the 10dB SNR is –122dBm, the SNR when the load is connected is –77dBm and when the antenna is connected it is –70dBm. The effective on-site SNR is:

$$SNR_{EFF}=SNR_{10dB}-(SNR_{LOAD}-SNR_{ANT})$$
$$SNR_{EFF}=-122dBm-[(-77dBm)-(-70dBm)]$$
$$SNR_{EFF}=-122dBm-[-7dBm]=-115dBm$$

The effective sensitivity is only valid for the given site and conditions present when the test is performed. If the site is changed, or if the noise generators and other signals present change, then the test must be repeated.

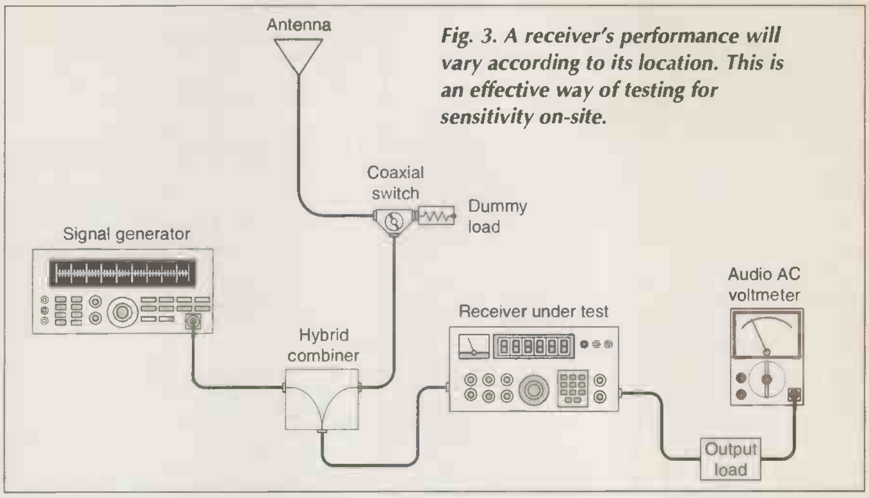

Fig. 3. A receiver's performance will vary according to its location. This is an effective way of testing for sensitivity on-site.

FM receiver sensitivity

There are two basic methods for measuring the sensitivity of FM receivers: 20dB quieting and 12dB SINAD. The 20dB quieting method is typically used on FM broadcast band receivers. It was once popular for communications receivers as well. More recently, the 12dB SINAD method is preferred.

20dB quieting method. This method relies on the fact that the FM detector will suppress noise once the limiting signal level is reached. The well-known capability of FM to eliminate noise relies on the fact that most noise amplitude modulates the carrier.

If the amplitude can be clamped below the level where the noise is effective, then the frequency variations can be detected to recover the audio. This effect is called 'quieting,' i.e. the reduction of noise as the signal level increases.

To measure the 20dB quieting sensitivity:

- Connect the receiver and signal generator as in Fig. 2. Keep the signal generator output at zero. The modulation – deviation – should be set to whatever is appropriate for the class of receiver being measured.
- Turn the RF gain all the way up. Set the audio output to produce a convenient reading in the high end of the AC voltmeter scale.
- Measure the output noise level and write it down.
- With modulation off, turn the signal generator

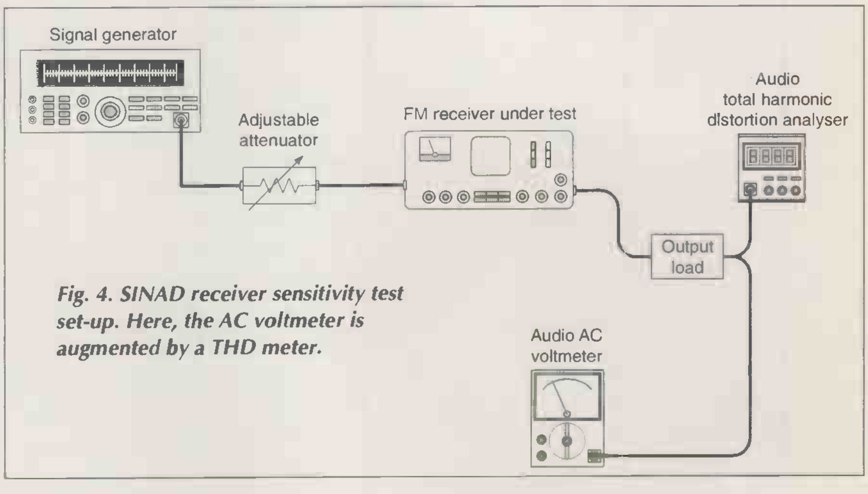

Fig. 4. SINAD receiver sensitivity test set-up. Here, the AC voltmeter is augmented by a THD meter.

output level up until the reading on the AC voltmeter drops 20dB. The signal generator output level that accomplishes this is the 20dB quieting sensitivity – typically less than 1μV.

SINAD sensitivity. The sensitivity of FM receivers is often expressed in terms of SINAD. This approach to signal-to-noise ratio recognises that the problem of detection depends on not simply signal and noise level, but also distortion. The SINAD (signal-noise-distortion) method is described by equation (1).

$$SINAD = \frac{signal + noise + distortion}{noise + distortion} \qquad (1)$$

In terms of decibels, the following equation is used:

$$SINAD(dB) = 20\log\frac{V_{signal} + V_{noise} + V_{distortion}}{V_{noise} + V_{distortion}} \qquad (2)$$

Here, *SINAD(dB)* is the SINAD sensitivity expressed in decibels, V_{signal} is the output voltage due to signal, V_{noise} is the output voltage due to noise and $V_{distortion}$ is the output voltage due to distortion.

The standard 12dB SINAD sensitivity corresponds to a 4:1 S/N ratio, in which the sum or noise and distortion is 25 percent of the signal voltage. As signal levels get higher, the SINAD and 10dB S/N values tend to converge.

Figure 4 shows a typical test set-up for the SINAD measurement. The output AC voltmeter is augmented by a total-harmonic-distortion analyser, both of which measure the output signal across the audio load (speaker, load resistor, etc.)

- Set the signal generator frequency and receiver frequency to the same value.
- Set standard conditions: modulating frequency 1kHz sine wave; deviation set to 60 percent of the peak deviation used for that service. For an FM broadcast band receiver for example, deviation is ±75kHz, so set the signal generator deviation to 0.6×±75kHz=±45kHz. For a communications receiver designed for ±5kHz deviation, set deviation to 0.6×±5kHz=±3kHz.
- Adjust the receiver audio output to approximately 50 percent of the receiver's rated audio output.
- Adjust the signal generator output until the input signal is high enough to produce 25 percent distortion. This is the 12dB SINAD sensitivity.

Special SINAD sensitivity meters are available that combine the total-harmonic-distortion analyser and audio voltmeter functions in one instrument.

In summary

Measuring a receiver's sensitivity is relatively easy if correct procedures and decent equipment are used. Now you should be able to tell whether the specifications claimed in advertisements are reasonably accurate – or not. ■

A rough guide to Europe

Working abroad in Europe could awaken your mind and boost your career, says Tony Atherton

For some people a time working abroad has been a natural part of expanding their life and career. As opportunities in former colonies have diminished so those in Europe have increased.

Yet, while we are all Europeans now – or so we are told – for most people the European dream of living and working on the continent is little more than someone else's funny idea. Even so, you will not be a pioneer. Soldiers, footballers, construction workers and politicians have all been there before you.

European Union (EU) rules guarantee freedom of movement to nationals of all member countries, allowing you to work or set up business anywhere in the EU. As a British citizen you do not need a work permit, although in most EU countries you will need a resident's permit if you intend to stay more than three months.

The changing rules make life easier but they do not guarantee success and happiness. Working in Europe is very different from holidaying there but at least a holiday can start the acclimatisation process.

While language will be the main stumbling block, electronics means you can run your life in English with cash and credit cards, e-mail, and telephone banking.

As for shopping, department store and supermarket staff rarely speak to you in Britain let alone abroad and there is always cyber shopping. All that is fine if you want to isolate yourself from the local community and so miss the point of being there in the first place.

The three basic recruitment questions apply just as much to working in Europe as in Britain.

- Can you do the job?
- Will you do the job?
- Will you fit in?

The third one takes on a new dimension when you move to another country. It is not simply a question of will you get along with your new colleagues, but will you fit into the community?

If you are heading for a large city with an international presence then you can lose yourself in the expatriate community, speak english all the time, read British newspapers, drink British beer and be a little Briton in a different climate. If that is all you want then you could stay at home and turn up the central heating.

Alternatively, your may want to experience the culture, work with the nationals, learn their way of life and speak their language. There are advantages. If employers can recruit locally they will not risk employing someone who may be pining for home within three months.

For this total immersion approach you must learn the language. A national training company, such as Linguarama, will charge around 1500 per week for intensive training but check out local trainers in the Yellow Pages. Also try schools, Sixth Form Colleges and friends to see if anyone will tutor privately.

Research is the key. Do you know anyone who has lived and worked in the country of your choice, or retired there? Does your existing employer offer any opportunities to work with subsidiaries or partners elsewhere in the EU? Do you know any nationals of your chosen country who work here?

Job Centres can put you in touch with Euroadvisers from the Overseas Placing Unit of the Employment Service. These Euroadvisers are part of the European Employment Services partnership (EURES) and are in contact with their opposite numbers in other EU countries. Try the Eures Web site as some jobs are advertised there. The current shortages include IT specialists. Ask at your Job Centre for a copy of the booklet Working in [Country Name]. These include details of resident permits, taxation, social security, education, Job Centres, and so on.

Visit your local reference library for a good careers encyclopedia (such as Cassells) and the Executive Grapevine Volume 1. The latter identifies the recruitment agencies that deal with both electronic engineering and EU countries. The CEPEC Guide is similar.

In electronics one immediately thinks of the Netherlands, Germany, France, Italy and Ireland. However, all EU countries have some electronics industry. The telecoms and IT industries are universal and generate opportunities everywhere, sunny or not. Finally do not forget the EU itself. Someone has to provide its IT and telecommunications services in Brussels and Strasbourg, for instance; likewise for the UN in Geneva.

Working abroad provides a huge range of experiences you will not get in the UK. As part of a career plan, especially in the early years, a spell in Europe will awaken your mind and possibly boost your career. You may even get a tan. ■

Tony Atherton: 01962 885534
tony.atherton@btinernet.com

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Senior position to work on future ASIC technology for set top box applications. Minimum of three years ASIC design experience with knowledge of VHDL/VERILOG, logic synthesis, test insertion, verification, etc. Ref: BC2200

Consultancy

M3 £20-40K

Junior and Senior position to work on analogue and mixed signal ASIC design. Good BSc with interest or experience in customer driven ASIC design. Ref: BC2201

Processors

Middx to £50K

A number of positions to work on the design and verification of RISC and DSP processors. HDL/VHDL experience necessary with team/project leadership skills for senior position. Ref: BC2202

Communications

M4 c. £45K

Team leader to run a group of ASIC Designers working on mobile comms, chip sets. A hands on role requiring substantial experience of embedded codes and customer devices. Ref: BC2203

Satellites

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You will work on state of the art VLSI systems for space-borne DSP environment. Wide experience of CAE tools and in depth understanding of digital/DSP design. Ref: BC2204

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South Coast to £35K

Good degree plus 2 years solid ASIC design skills to work on digital video systems. Your experience should include VHDL, H/S digital systems, RF (to 2GHz), FPGA's, DSP, etc. Ref: BC2205

Consultancy

Cambs to £40K

Minimum of 2 years experience in the mobile communications sector working on low cost high volume product design. VHDL/SYNOPSIS essential with the ability to specify and implement complex IC's. Ref: BC2206

ASIC and Clients

M3 £28K + Car

Solid experience of design flows with particular emphasis on the front end from RTL coding, synthesis, simulation and timing analysis. Lots of customer contact. Ref: BC2207

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A number of positions with an IC design consultancy. CMOS and BiCMOS experience an advantage. Work will involve RF and Analogue IC design with an excellent salary and benefits package. Ref: BC2208

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Development of digital hardware for communications equipment. In depth experience of FPGA's, ASIC's and gate arrays with VHDL and some PCB design. Ref: BC2209

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Surrey £30K

Creation of basic cell libraries, memory and complex IP models, model validation using simulation tools and technical support to field sales team. Ref: BC2211

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High calibre digital design engineers with 2 years experience including strong HDL and synthesis skills. Ref: BC2212

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CIRCUIT IDEAS

Non-intrusive continuity tester

Without injecting a direct current into the circuit being tested, or activating any semiconductor junctions by the millivolt-level ac, this tester provides an audio signal to indicate a low resistance between test points.

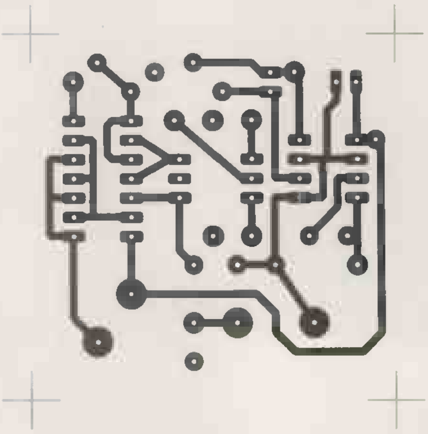
A switched-capacitor inverter provides both 6V for the hex Schmitt inverters and a square wave from its summing node. A reflected impedance of under 1Ω or 4Ω from the half or full transformer secondary provides a 4kHz tone from the piezoelectric transducer.

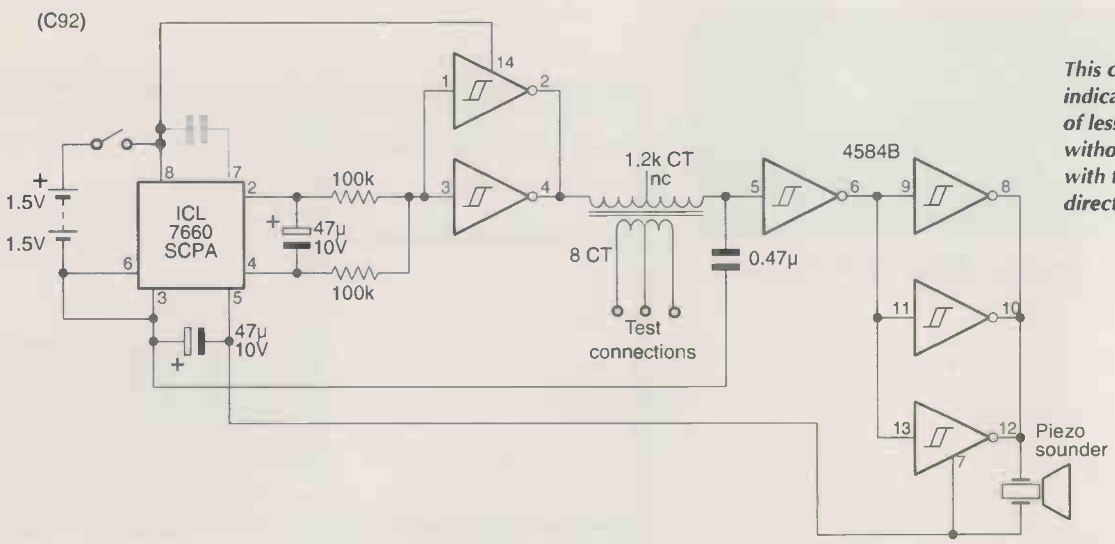
If required, a capacitor between pins 7 and 8 of the inverter lowers the frequency of the square wave to match the transducer resonant frequency.

John A Haase
Fort Collins
Colorado
USA
C92

Further reading

Einar Abill, 'Sense continuity below an ohm,' *Electronic design*, September 5, 1995.





This continuity tester indicates resistance of less than 1Ω without interfering with the circuit; no direct current is used.

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

Like life, *Electronics World* may seem surreal at times, but it is certainly not exclusive. Clearly, the best circuit ideas are ones that save time or money, or stimulate the thought process. This includes the odd solution looking for a problem – provided it has a degree of ingenuity.

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

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

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

Liquid-level controller

Low power consumption, resistance to interference from mains spikes and low cost are features of this water level controller.

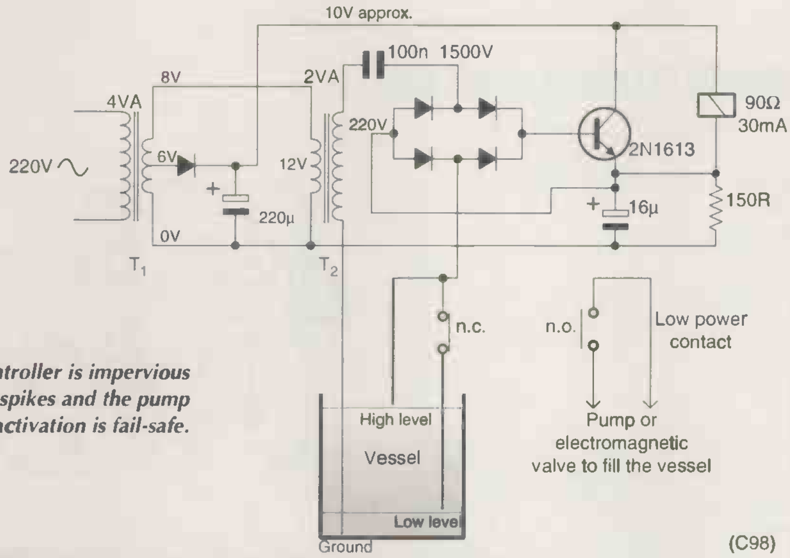
A step-down mains transformer provides voltage for the control circuit and a step-up transformer for the sensor measuring voltage of 100-200Vac to avoid electrolysis. Measuring current is 1-3mA, limited by the 100nF, 1.5kV capacitor.

If the vessel is empty, no current flows through the bridge rectifier, the transistor is off and the relay is activated. The normally closed contact is open and the low-level electrode is not in circuit, while the normally open contact is closed and the pump or valve starts to fill the vessel.

As the water rises, it first touches the low-level contact, but has no effect, since it is out of circuit; the level rises until the high-level contact is touched and the bridge takes current, the transistor turns on and the relay coil is short-circuited, so the low-level contact is in circuit and the pump stops. As the vessel empties and clears the low contact, the cycle starts again.

Mains spikes have no effect, since they reinforce the actions of the circuit. Distance between the vessel and circuit may as much as 10m, but the ground line must not be near the other two lines to avoid stray capacitance.

Roland Vanthomme
Sambreville
Belgium
C98



Liquid level controller is impervious to mains spikes and the pump activation is fail-safe.

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CIRCLE NO. 115 ON REPLY CARD

Mosfet hf mixer

Figure 1 shows a typical dual-gate hf mixer, which provides reasonable, low-cost intermodulation and noise performance.

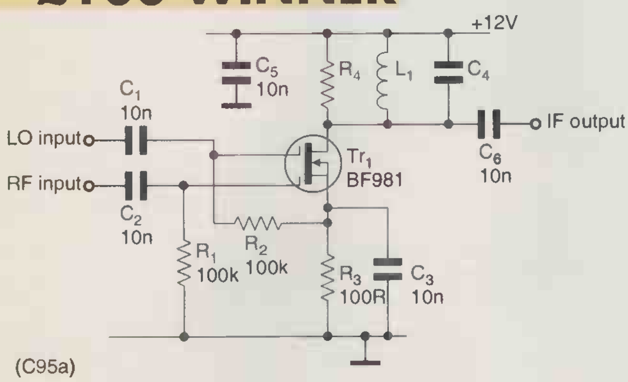
The improved circuit in **Fig. 2** employs the mosfet as an amplifier rather than a mixer, an arrangement that allows the mosfet to be biased for best noise figure.

Transistors $Tr_{2,3}$ provide the multiplication needed for mixing and act as switches. Inductor L_1 and C_4 resonate at the if and R_4 represents the if load.

Local oscillator input must be around 20mV pk-pk from a low impedance such as 50Ω and output inductor L_1 is bifilar wound on a ferrite core for best balance. These modifications to the original circuit provide higher gain and a lower noise figure and higher intermodulation intercept to give an improvement in third-order dynamic range.

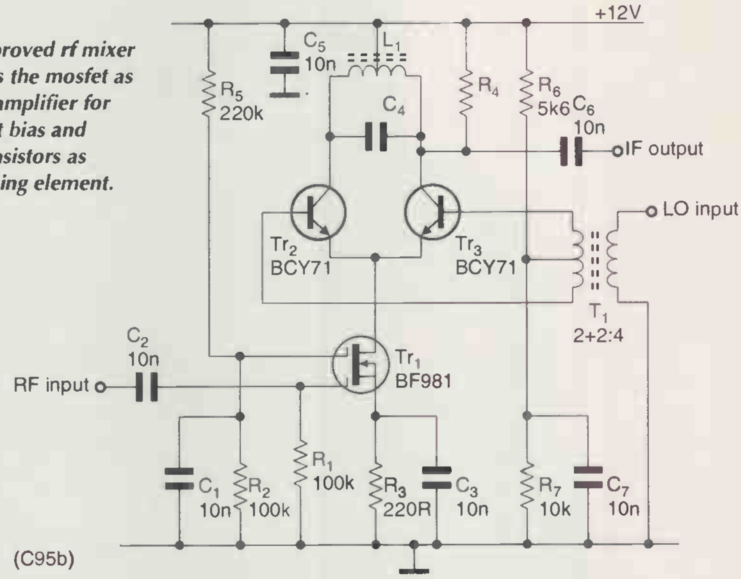
P Goodson
Bracknell
Berkshire
C95

£100 WINNER



(C95a)

Improved rf mixer uses the mosfet as an amplifier for best bias and transistors as mixing element.



(C95b)

Wide-band amplitude modulator

Gain of this amplitude modulator is constant for carrier frequencies between 10kHz and 50MHz. The circuit linearly modulates the carrier by up to 100% using a modulation signal between 10Hz and 200kHz.

Modulating input varies the current through Tr_3 and therefore the gain of the emitter-coupled amplifier $Tr_{1,2}$. Output of the amplifier is a composite of an am-modulated carrier and the modulating frequency, the latter normally being removed by a filter.

Since the modulating frequency appears at the two collectors in common-mode, the differential op-amp amplifier rejects it, while amplifying the two am signals normally. The avoidance of filtering confers a wide-band performance.

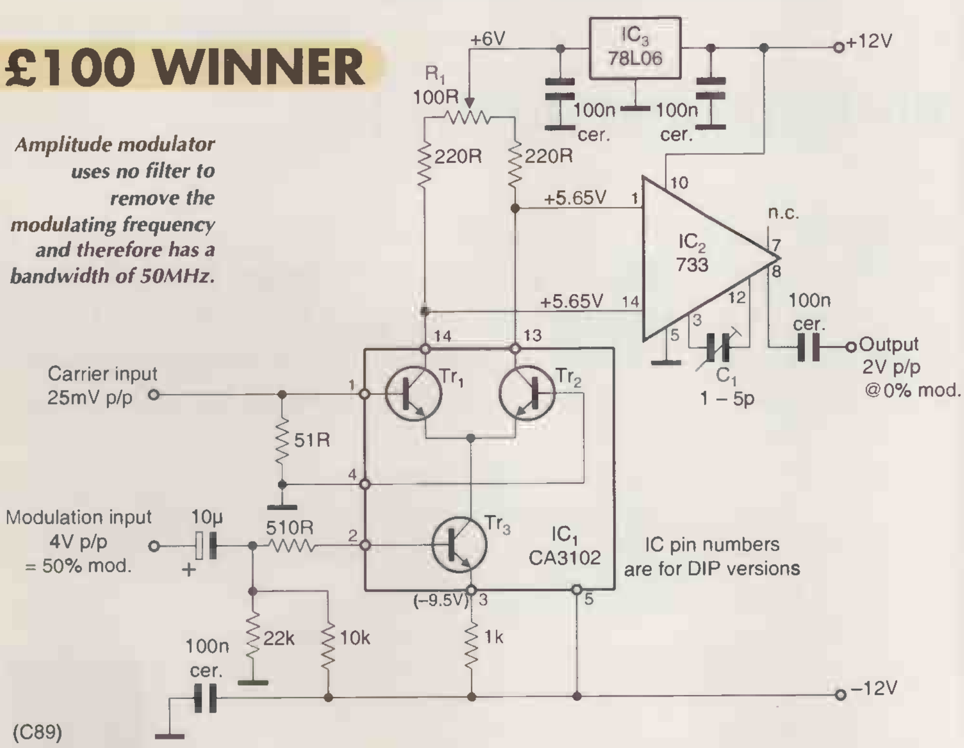
Trimmer C_1 increases the gain of the op-amp to counteract precisely the fall-off in gain of the emitter-coupled amplifier, while the potentiometer R_1 sets symmetry of the rf envelope when modulated to 50% by a 200kHz input, the carrier being anything over 2MHz.

The circuit is affected only minimally by temperature.

John Gibson
Berkeley
California
USA
C89

£100 WINNER

Amplitude modulator uses no filter to remove the modulating frequency and therefore has a bandwidth of 50MHz.



(C89)

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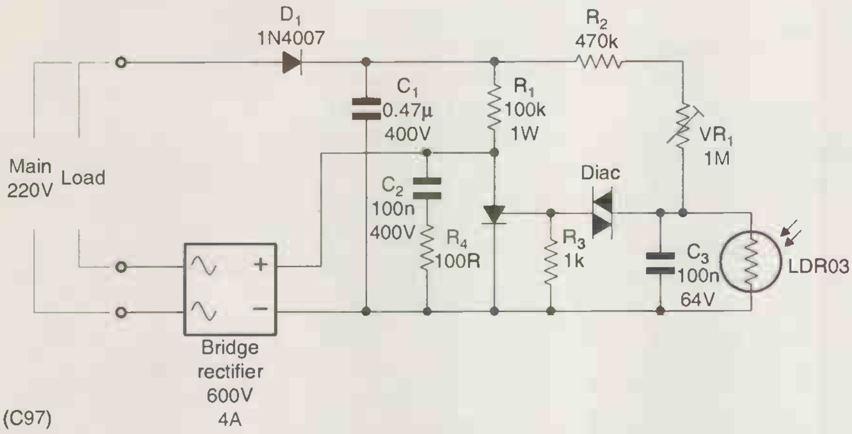
Twilight switch

This circuit switches power to a load when ambient light falls below a set level. In some circuits of this type, the light sensor will switch the load off again if light impinges on it. In this circuit, the thyristor is supplied with a sustaining current from the rectified supply, so that the load remains on in the presence of mains power.

Capacitor C_2 and R_4 prevent rapid power-on voltage rises to avoid false thyristor firing. The variable resistor adjusts sensitivity.

Jean-Marc Brassart
Saint-Laurent-du-Var
France
C97



(C97)

Once fired, the thyristor in this light switch stays on, regardless of light on the sensor.

Triode audio amplifier with bootstrapping

Positive feedback reduces the demands on low-mu triodes such as the 6080,which are easily obtained, avoids the need for inter-stage transformers and exotic triodes.

The use of triodes as the output stage in valve amplifiers has much to recommend it, but there is the problem of the large drive needed by some types of valve. It is found that a form of bootstrapping overcomes the problem, following similar design techniques used by Macintosh.

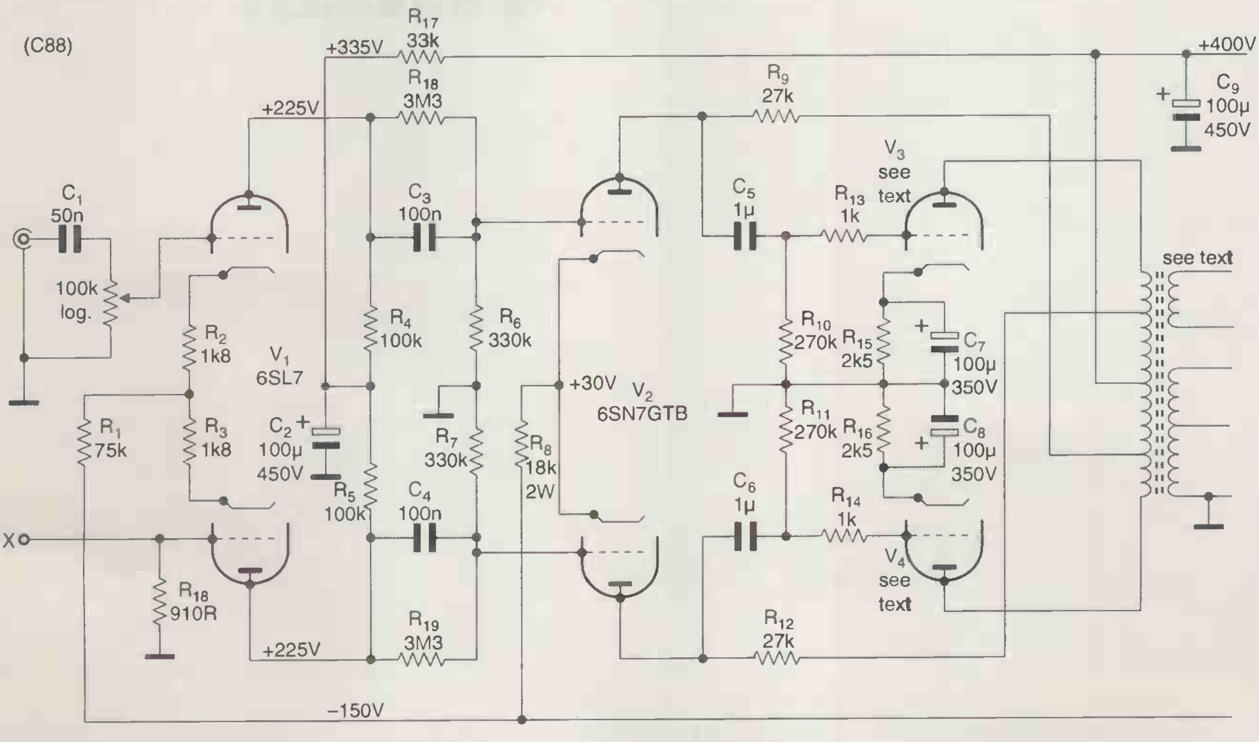
Positive feedback, applied to the anode load resistors of the driver stage, comes from taps on the ultra-linear output transformer, many of which are available. This removes the need for the drivers to supply the large voltage swings needed by low-μ triodes such as the 6080. An output transformer having 43% taps is suitable for use with the 6080 types or, with a μ of 4.5, 25% taps can be used.

A single 6080 with a 4300Ω impedance transformer, running from a 350V rail, produced 10W of clean audio, while another with two 6080s in push-pull and a transformer impedance of 2150Ω on a supply of 400V gave 30W. V_3 and V_4 are both sections of a 6080, each section having its own cathode resistor and bypass.

Feedback comes from the 8Ω tap of the transformer to point X, reversal of the output transformer secondary stopping any oscillation.

John L Stewart
King City
Ontario
Canada
C88

(C88)



£100 WINNER

New FIR speaker crossover

Finite impulse response filters produce a linear phase response and a constant group delay and may be designed using all-pass elements and a summing network, a linear phase response being obtained when all coefficients are symmetrical.

The two-way crossover filter shown here is a member of a new class of phase-linear FIR crossovers.

Varying the coefficients produces differing low-pass filters, as shown below. Taking the difference between the delayed input and the low-pass output produces a high-pass characteristic.

With four all-pass elements, you can obtain slopes up to 24dB/octave. Adding elements with an extended summation network produces higher-order crossover, but for use with standard high-quality speakers, second or fourth-order crossovers are adequate.

Corner frequency depends on R_0C_0 in each element and on values in the summation network.

For an average filter,
$R_{1\text{-}5}$ 20kΩ.

Least-squares filter:
R_1 8.66kΩ
$R_{2,4}$ 34kΩ
R_3 48.7kΩ.

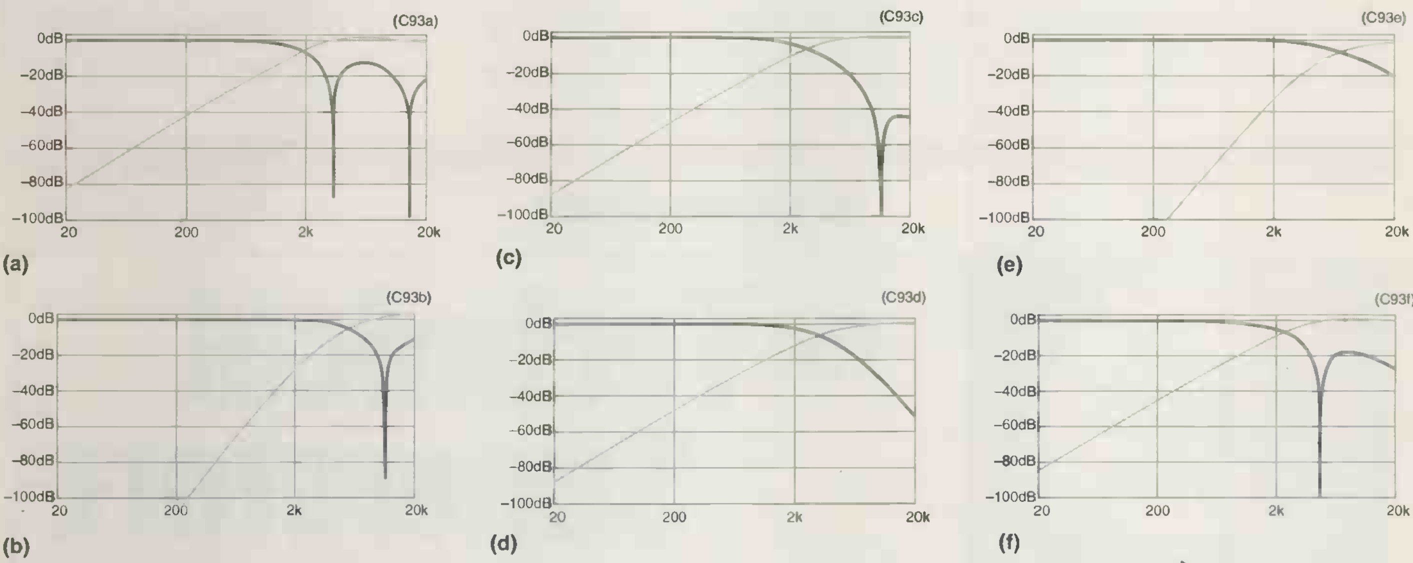
Optimised average filter:
R_1 7.32kΩ
$R_{2,4}$ 24.9kΩ
R_3 35.7kΩ.

Optimal low-pass filter:
R_1 6.19kΩ
$R_{2,4}$ 24.9kΩ
R_3 37.4kΩ.

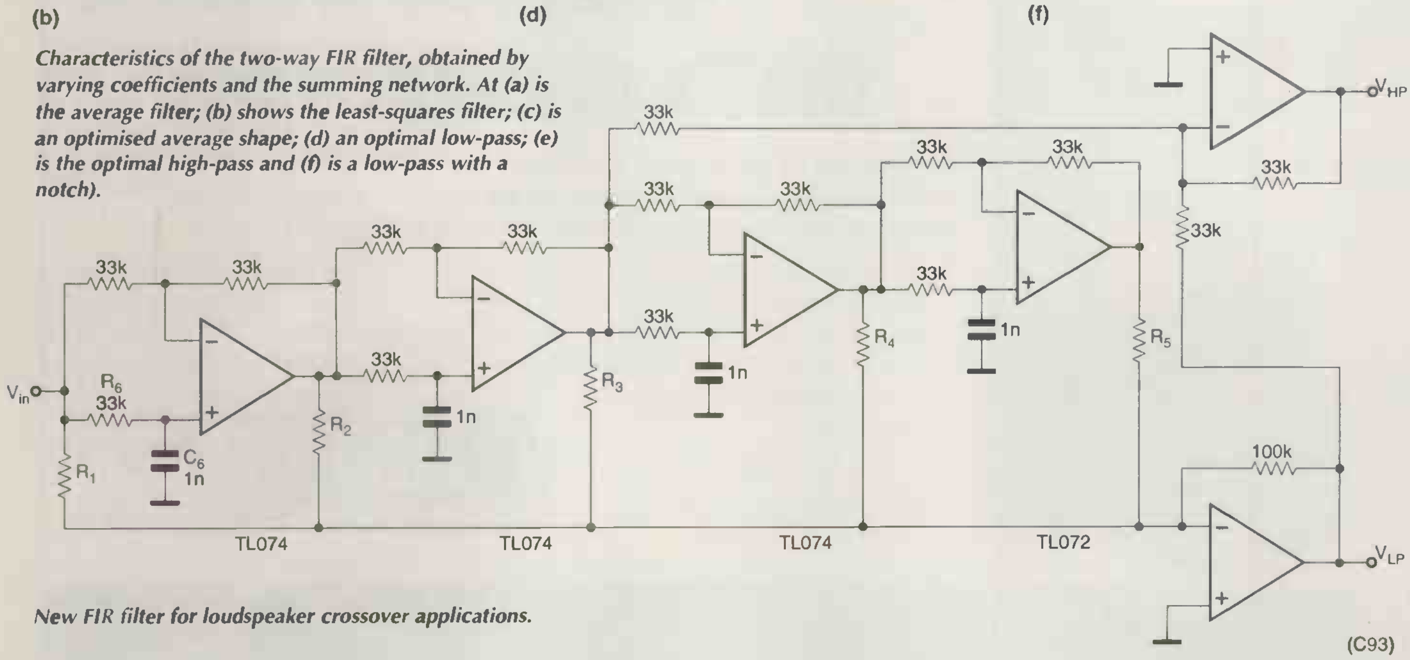
Optimal high-pass filter:
R_1 6.19kΩ
$R_{2,4}$ 24.9kΩ
R_3 61.9kΩ.

Low-pass filter with notch:
R_1 12.4kΩ
$R_{2\text{-}4}$ 24.9kΩ.

Gerd Schmidt
Frankfurt
Germany
C93



Characteristics of the two-way FIR filter, obtained by varying coefficients and the summing network. At (a) is the average filter; (b) shows the least-squares filter; (c) is an optimised average shape; (d) an optimal low-pass; (e) is the optimal high-pass and (f) is a low-pass with a notch).



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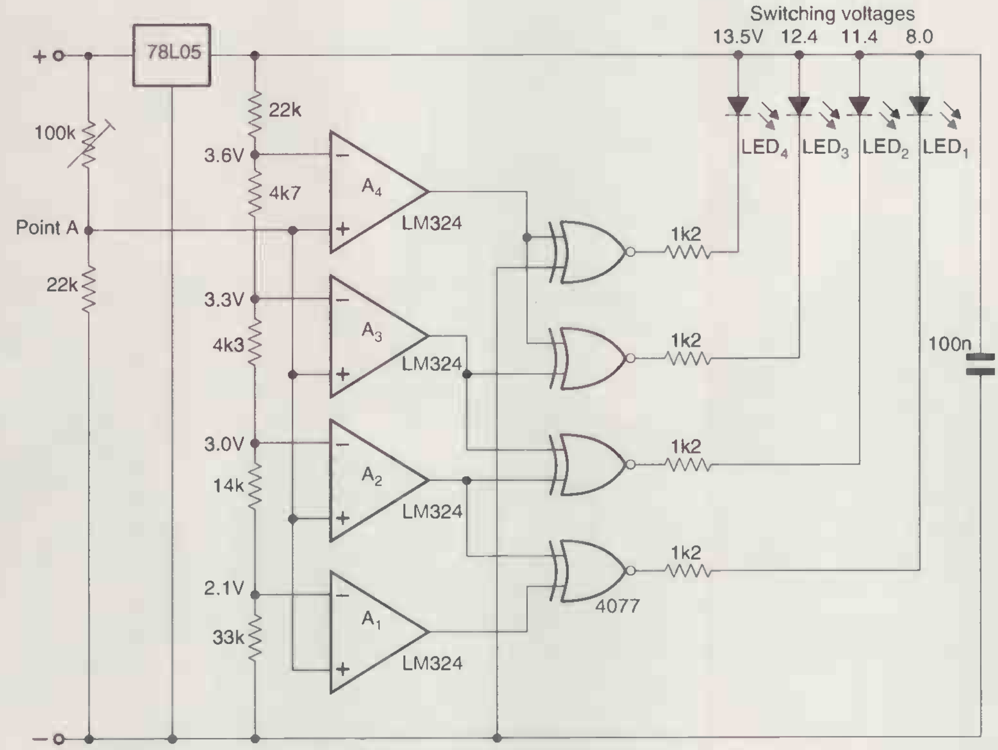
Low-current battery monitor

The use of exclusive-Nor gates in this four-led battery monitor, which originally had three of the four leds illuminated for much of the time, has allowed only one of the leds to be on at a time, so saving current.

As a reminder, an exNor gives a 1 output if one input is 0 and a 0 output if both inputs are 0. In the circuit shown, a *78L05* regulator supplies a tapped resistor chain with 5V to give four reference voltages, the four op-amps in an *LM324* being used as comparators, in which the voltage-divided input is compared with the four reference voltages. The use of the exNors means that as each led comes on, the one below it goes out, only the highest therefore being illuminated. Switching points with values shown are 8V, 11,4V, 12.4V and 13.5V. No led is on below 8V, since the circuit supplied by the battery is long gone at that voltage. ■

Neville Frewin
Fontainebleau
South Africa

Four-led battery monitor shows only one led at a time, instead of a maximum of four, due to the use of exclusive-Nor gates.



(D4)

Two-op-amp sine generator, MkII

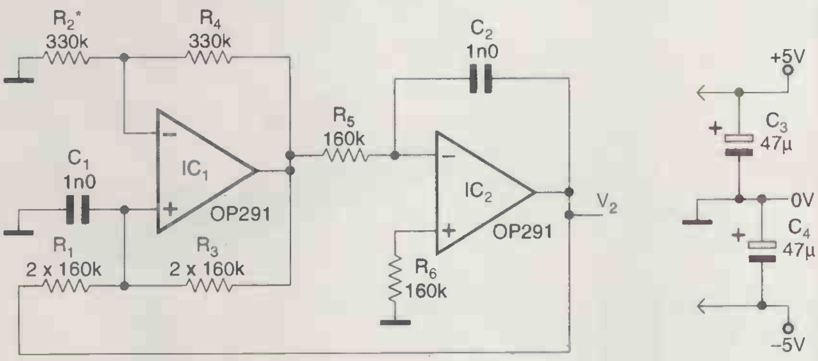
An earlier contribution[1] showed how two op-amps may be made to generate low-distortion sine waves, the circuit being based on a non-inverting integrator that needed two capacitors. This new circuit[2] needs only one capacitor, but two extra resistors, which are easy to match and readily available in low-temperature coefficient form at low cost.

Op-amp IC_1 is the non-inverter, IC_2 being the inverting integrator. To obtain oscillation, phase lag from each integrator must be 90° and loop gain just greater than unity; adjustment of the gain to make v_1 just clip at the ±5V rails is critical and is done here by reducing R_2 and inserting a 20kΩ trimmer in its ground leg. Lowest distortion is gained if C_2 is slightly greater than C_1 or R_5 slightly increased.

Common-mode signals into the inputs of IC_1 are nearly ±3V and it is necessary to use an *OP-291* or similar that has near rail-to-rail inputs as well as output. Frequency of the circuit shown is 1kHz.

CJD Catto
Cambridge
D3

Oscillator uses a non-inverting integrator described earlier, but modified to use one capacitor and two resistors.

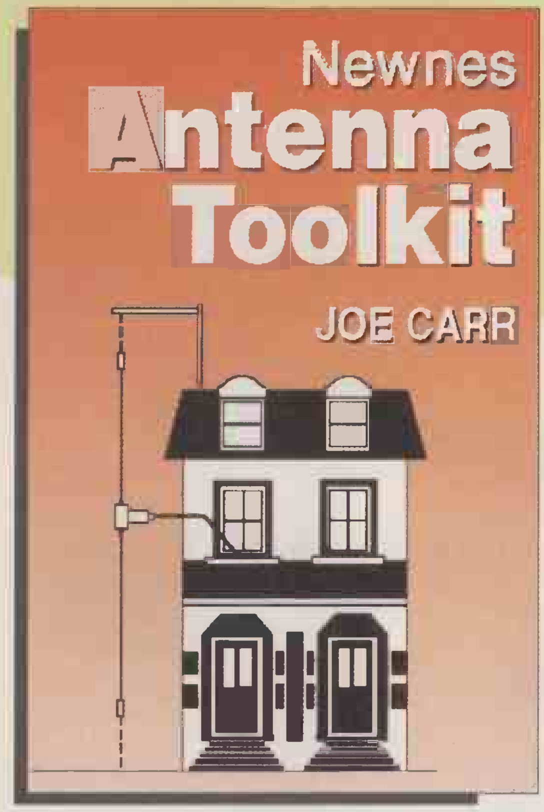


(D3)

References

1. Catto, CJD. Two op-amp sine generator. *Electronics World*, April, 1999, p.291.
2. Burnill, J. Integrator with no signal inversion. *Electronics World*, May, 1995, p.431.
3. Hickman, I. A perfect variable oscillator? *Electronics World*, June, 1998, p.485.

(Ref. 3 was inadvertently omitted from the article in Ref.1.)



Antenna Toolkit

by Joe Carr

Combined with antenna design software on CD-rom, Newnes' new book *Antenna Toolkit* provides a complete design solution. Prepared by antenna expert Joe Carr, this package is written for beginners and advanced users alike.

On the CD-rom is a suite of powerful software running on the pc. The software calculates the critical lengths and other parameters of the antennas in the book by having the user select the antenna type and set the frequency.

The main menu screen is in the form of tabs, one for each chapter of the book plus other topics.

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SPEAKERS' CORNER

Getting more copper inside a loudspeaker's coil gap by using square cross-sectioned wire instead of circular has got to improve performance, hasn't it? John Watkinson investigates.

As I have demonstrated in earlier articles, the efficiency of a moving coil speaker is pretty miserable. Anything that offers an improvement in efficiency is worth looking at.

The efficiency depends upon the *Bl* product and the overall moving mass. Making the magnet more powerful is an obvious approach, but this is expensive. An alternative is to look at ways of getting more *Bl* from an existing magnet.

This is where square and rectangular wire comes in to consideration. **Figure 1a)** shows a coil made with round wire. The packing efficiency is poor because circles don't fit together at all well. The result is that only part of the magnetic field that we have payed for is being used. Figure **1b)** shows that if square wire is used, the air voids are removed and the packing efficiency is better.

How much better?

The question is though, how much better? I have kept coming across the above argument, but the result was never quantified, so I decided to work it out. It turned out to be more complex than might at first be thought.

It is important to compare like with like, so that as much as possible everything should be kept the same except for the cross-sectional shape of the wire. So consider two coils, having the same overall length, the same DC resistance and the same mass. They differ only in their wire cross section.

This isn't easy because if the cross-sectional area of the wire is kept the same, the coil having the square wire will be shorter because of the better packing. This will reduce the linear travel of the cone and the overhung coil length and so we are not comparing like with like.

On the other hand, if the coils are made the same length, the coil having the square wire would have more turns and hence higher resistance.

I thought that one approach would be to compare coils of the same length and DC resistance. The cross-sectional area of the square coil would have to be greater so that the same resistance was obtained despite the larger number of turns. The result of this approach is that the coil wire itself would be heavier, but the effect of the changed moving mass could be estimated.

The panel entitled shows the calculations. The result is that the *Bl* product goes up by 8 percent because more turns are in the gap. The gap volume occupied by the wire can be reduced by 8 percent because of the improved packing. This means that the magnet could be smaller, or with the same magnet the field would be stronger.

Clearance and former losses

In practice, the magnet gap is wasted by clearance spaces and the coil former, so nothing like the 8 percent improvement is realised. The gain is more like 2 percent. The square wire has given about a ten percent improvement in *Bl*. Unfortunately, the panel also shows that the mass of the coil has gone up by 17 percent, so we lose!

Another approach to the problem is to allow rectangular wire. This would make it possible to match the mass, resistance and coil length with that of an equal circular wire coil.

Figure 2 shows that if the rectangular wire has the same cross-sectional area it will have the same mass and resistance per unit length as the round wire. Also, if the diameter is the same as the section height, the coil will have the same length too.

Note that the *Bl* product doesn't change. But the thickness of the coil falls to about three quarters of the value for round wire. This allows a smaller gap volume and consequently a smaller magnet for the same field strength.

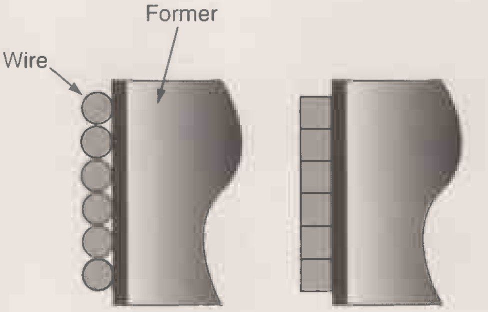
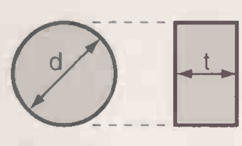


Fig. 1. Round wire at a) wastes magnetic field volume. Efficient packing of square wire at b) should be more efficient – but by how much?

However, the magnet won't be three quarters as big because as before some of the gap is used up with coil former and clearance space. Maybe a ten percent reduction in magnet size, or a ten percent improvement in *Bl* for the same magnet would be feasible.

Bearing in mind the phenomenal cost of square wire and the enormous difficulty in obtaining it, it is likely that the saving on the magnet would be eclipsed by the extra cost of the wire below a certain size of drive unit. It is only in large drive units where the magnet cost becomes an issue and small magnetic efficiency gains become worthwhile.

So the efficiency argument for



$$A=\frac{\pi}{4}d^2 \quad = \quad A=dt$$

$$\text{so, } t=\frac{\pi}{4}d^2=0.785d$$

Fig. 2. Rectangular wire can have the same weight and resistance as round wire, but needs only 3/4 the volume so the magnet can be smaller.

Efficiency versus weight

If wire length is L then for equal resistance,

$$\frac{L_S}{A_S}=\frac{L_R}{A_R} \qquad (1)$$

If coil circumference is C, number of turns is L/C. Length of coil, L_C, is number of turns multiplied by w or d.

$$L_C=\frac{L_S}{C}\times w=\frac{L_R}{C}\times d \therefore \frac{L_S}{L_R}=\frac{d}{w} \qquad (2)$$

From eqn 1,

$$\frac{4L_R}{\pi d^2}=\frac{L_S}{w^2} \therefore \frac{L_S}{L_R}=\frac{4w^2}{\pi d^2}=\frac{d}{w} \text{ (from eqn 2)}$$

$$\therefore \frac{d^3}{w^3}=\frac{4}{\pi} \therefore \frac{d}{w}=1.084 \rightarrow Bl \text{ is 8\% better}$$

If mass is M,

$$\frac{M_S}{M_R}=\frac{L_S\times A_S}{L_R\times A_R}=\frac{d}{w}\times\frac{4w^2}{\pi d^2}=\frac{4}{\pi}\times\frac{w}{d}$$

$$=1.17\rightarrow \text{ mass is 17\% higher.}$$

So assuming constant DC resistance and coil length, the coil using square wire is 8% more efficient, but 17% heavier.

$A_R=\frac{\pi}{4}d^2$ $\qquad$ $A_S=w^2$

square or rectangular wire is tenuous if used as a replacement for round wire in an existing design. It is certainly less effective than the improvement obtained by going from copper to aluminium. However, square or rectangular wire has the characteristic that successive turns can easily be bonded together so that the coil former can be dispensed with. This has a number of advantages.

Benefits of no former

The elimination of the coil former reduces the amount of wasted volume in the gap. The wasted volume is now only due to the inner and outer coil clearance spaces. A self supporting coil has better cooling because both sides are exposed and can radiate to both poles.

Combining the elimination of the coil former with the slightly better packing of the square wire gives a tangible reduction in gap volume. The more powerful the speaker, the greater the ratio of coil volume to clearance volume and the more relevant the approach becomes. Self supporting coils of this kind are found in all kinds of applications such as vibration table actuators and in the positioners of giant disk drives.

For small or low powered speakers, it doesn't make economic sense except for manufacturers who operate at very high volumes and can get specialist wires at reasonable cost through volume purchase. If square or rectangular wire is contemplated, the speaker has to be designed from the outset to use it. If the coil isn't self supporting, most of the advantage is lost. ■

Maurice Wilkes: pioneer of modern computing

Wilkes addressing IBM staff at Wembley Conference Centre in 1977.

Chris Hipwell has been talking to Maurice Wilkes, the innovator who, in 1949, built the first computer designed to store programs.

Fifty years ago, on 6 May 1949, a machine read a paper tape containing a program for computing a table of squares and printed the results. Edsac, the first computer designed to embody the concept of the stored program, was up and running. It would remain operational, providing a service for scientists and researchers at Cambridge University, until 1958.

Those bare facts hardly do justice to the achievement of Maurice Wilkes who had, with a team rarely numbering more than half a dozen people, transformed a concept proposed by American computing pioneers into a working system. It had taken two-and-a-half years but even so Wilkes' implementation of the stored program concept was completed ahead of its American counterpart, Edvac.

Remarkably Wilkes, now aged 85, still works as a consultant at the AT&T Research Laboratory in Cambridge. This institution, formerly funded by Olivetti and Oracle, is a think-tank that provides a link between academia and the commercial IT world.

In Wilkes' case it also provides a link with the present generation of researchers at the University Computing Laboratory, which he was director of for 34 years from 1946 to 1980.

It was at his office in the AT&T laboratory that we talked about his early days and the unusual conditions prevailing in Britain when he set out to design Edsac in 1946. Despite the austerity and power-cuts of the post-war years, for Wilkes it was the best of times.

Although Britain was on its knees in economic terms, "everybody was very co-operative, everybody was keen to re-establish peacetime values and this released an enormous amount of energy," remembers Wilkes. In consequence, he asserts, "there couldn't have been a more favourable time to establish an ambitious project of that sort."

Wilkes was then 33 and the whole of his life seemed to have been a preparation for the task that lay before him. As a schoolboy he had tinkered with electric batteries, lamps and bells.

After the advent of broadcasting in 1922 he became a keen constructor of crystal sets and receivers, reading *Wireless World* – the main source of his knowledge of electronics. "As a boy," he said, "I used to lap up *Wireless World*." During his sixth-form years, Wilkes became a radio ham, building his own rig.

Wilkes went on to study mathematics at Cambridge, achieving first class honours in his finals in 1934. He stayed on to study for a PhD, joining the radio group at the Cavendish Laboratory, where he undertook research into the passage of long wave radio in the ionosphere.

During this period Wilkes attended a lecture by Professor Hartree of Manchester University on their differential analyser. He used a machine modelled on the analyser for his own research at Cambridge.

This work gave Wilkes an insight into mechanical calculation and the demand for such facilities among university researchers. He was appointed to a

Wilkes and Wireless World

In his autobiography 'Memoirs of a Computer Pioneer', Wilkes pays tribute to the part that *Wireless World* played in the development of his interest in electronics.

Some time in the mid 1920s, he says he, "began to take regularly the *Wireless World*." This excellent journal published constructional articles containing complete designs for receivers... and tutorial articles on the more theoretical aspects of the subject. It was through reading *Wireless World* that I laid the foundation of my knowledge of electronics".

junior position in the new Mathematical Laboratory established in 1937.

The War years

For many, the war years marked a hiatus in their careers. For Wilkes, although physically removed from Cambridge, the war years marked a further stage in his progress as an electronics engineer capable of applying his knowledge to a variety of problems.

Early on he was introduced to the new technology of radar, briefly operating on radar sites, but later undertaking research into the development of radar for a range of wartime applications.

Wilkes returned to Cambridge as Acting Director of the Mathematical Laboratory in 1945. He soon came into contact with Professor Hartree who had recently visited the US. While there, he had seen the Eniac computer which was under development at the Moore School of the University of Pennsylvania.

In the following year, Wilkes read the 'Draft report on the Edvac' written by John von Neumann on behalf of the group at the Moore School. It laid out the principles on which the digital computer was to be based, prompting Wilkes to comment that, "I recognised this at once to be the real thing, and from that time on never had any doubt as to the way computer development would go."

This certainty was reinforced later in 1946 when Wilkes visited the US to attend a course on electronic computers at the Moore School. There he saw Eniac – a monster containing 18 000 vacuum tubes, 70 000 resistors, 10 000 capacitors and 6000 switches. It occupied three sides of a 40 feet by 20 feet room.

However, Eniac lacked a memory. Programs were set up manually by inserting plugs in sockets and via three function tables, large vertical panels bearing some hundreds of switches on which numbers could be set.

Wilkes was impressed by the scale of Eniac, but he realised that it was already yesterday's machine. The future lay with the stored-program principle.

Even before Wilkes left the US, he began to sketch out the design of the machine that finally became Edsac. He acknowledged that this would follow the ideas of Eckert and Mauchly as set out in von Neumann's Edvac report. Even so, much had still to be decided – not least the actual design of the all-important memory.

However, Wilkes was well prepared for the task that lay before him and could claim in his memoirs that, "with my experience of radio and radar behind me, I was entirely confident in the design of electronic circuits and knew exactly what it was possible – and not possible – to do with them."

Edsac, circa 1949 – the first computer to run a program read into memory.

Developing the memory

The only part of the proposed computer that called for significant technological innovation was the memory. Although not specified in the Edvac report, the type of memory envisaged by Eckert and Mauchly was a delay line that depended on the transmission of ultrasonic pulses through a column of mercury.

Fortunately for Wilkes he met Tommy Gold in 1946. Gold was a research student at the Cavendish Laboratory who had worked on the design of mercury tanks during the war. The devices looked like 'tubes' but the word 'tank' was used to avoid confusion with vacuum tubes. With the help of Gold's experience, Wilkes soon designed and built his first mercury tank.

Writing for a technical journal in 1948, Wilkes described the delay system as follows:

> *'Ultrasonic waves in liquids travel at speeds which are slow on the time scale on which electronic events can be made to occur, so it is possible to delay a train of pulses for a comparatively long time by converting them into pulses of ultrasonic sound and passing them down a column of liquid (mercury) a few feet long. The conversion is done by means of an X-cut quartz crystal and an exactly similar crystal is used at the far end of the tube to convert the ultrasonic pulses back into electrical pulses.'*

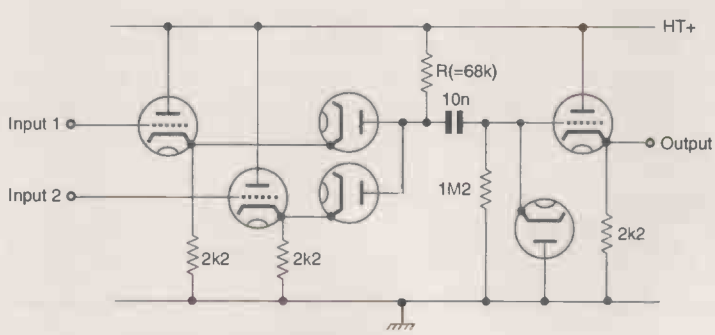
When the pulses appear at the output terminals, they are amplified and routed back to the input. They are thus kept in circulation for a definite length of time and can be taken out when required.

Principles and prototypes are all very well, but converting them into an engineered product draws on different skills. When asked which aspect of the Edsac project had given him most satisfaction Wilkes replied, "I certainly enjoyed designing the memory batteries, enjoyed it very much. It was a bit of mechanical design, properly stressed and designed so that it could be made."

Electronic memory – 1.6m long

A single mercury tank 1.5 metres long would hold only 16 words of 32 bits. It was decided that in order to give a reasonable amount of memory, 32 tanks would be required. These all had to be the same length to within a very close tolerance. Moreover the crystals at the opposite ends of each tube had to be aligned very accurately, within a few minutes of arc.

Rather than building each tank as an independent unit, Wilkes said, 'I



Typical logic gate circuit. A pulse train was fed to input 1 and a square wave to input 2. According to the description, "Pulses appear at the output for as long as the square wave lasts. They cease to pass as soon as the grid of the valve at input 2 returns to zero."

Maurice Wilkes in his office at AT&T's research laboratories, March 1999.

thought it would be much nicer to build these tanks in batteries of 16 and to have each battery made with sufficient precision so that when it was bolted up solid everything was in line and the crystals were the correct distance apart'.

Short mercury tanks, an inch or so long and each holding only one short or long number, were used for the registers in the arithmetic unit, the accumulator and the multiply register. Short tanks were also used for the registers for holding instructions that had been called.

In all, Edsac contained around 3000 vacuum tubes. A typical gate used for switching comprised two cathode followers that applied inputs to the cathodes of two diodes the anodes of which were connected together. There was a pull-up resistor, marked *R* on the circuit diagram, and the output was fed into an amplifier.

Wilkes says, "I was very proud of my discovery. After quite a lot of experimental work, I found that the ideal sort of amplifier for that purpose was a cathode-coupled type." Dual diodes, namely EB34s, were used for the rectifiers and EF54s (RL7s) connected as triodes were used for the cathode followers.

The machine was AC-coupled. To keep the level baseline in a fixed position, DC restoring diodes were used ; "We had to put these things everywhere and make sure they worked," commented Wilkes. Consequently, DC restoring diodes made up an appreciable proportion of the total number of valves in the machine.

The circuit assemblies were housed in purpose-built racks with long chassis to make access easier, rather than the more familiar 'Post office' racks.

"I never liked 19in racks"

Wilkes says, "I never liked 19-inch racks. Why something that appeals to people who design telephone exchanges should have exercised such a grip on the electronics industry, I have always failed to understand."

With the self-confidence of a born engineer, Wilkes has never been afraid to question conventional wisdom if he could see a better way to achieve his objectives. Following Edsac, he went on to demonstrate the concept of microprogramming in Edsac 2.

Wilkes also become involved in the time-sharing developments of the early 1960s, and to conceive the idea of the Cambridge ring – an early local area network – in the late seventies.

Even so, it will be for the work which culminated in the running of that program one day in May fifty years ago that Maurice Wilkes will be remembered. ■

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Professionally designed and manufactured printed circuit boards for Giovanni Stochino's no compromise 100W power amp are available to buy.

These high-quality fibre-glass reinforced circuit boards are designed for Giovanni Stochino's fast, low-distortion 100W power amplifier described in the August 1998 issue. Layout of the double-sided, silk screened and solder masked boards has been verified and approved by Giovanni.

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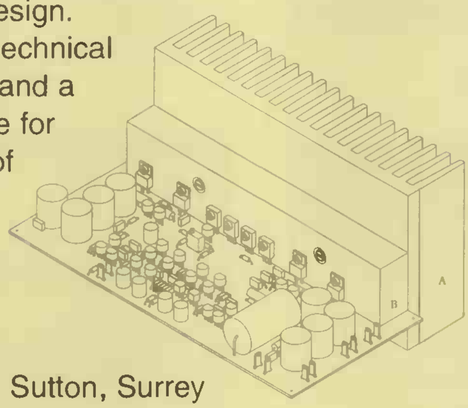
You can receive a copy of the technical notes that accompany the pcbs and a copy of Giovanni's original article for £3.50, refundable on purchase of one or more pcbs. Send a postal order or cheque payable to Reed Business Information to 'Board notes,' Electronics World Editorial, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. Don't forget to include your address.



Giovanni's high-performance power amplifier mounted on its heat sink.

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Specifications

Power into 8Ω load	100W
Small-signal bandwidth before the output filter	20Hz (–0.1dB), 1.3MHz (–3dB)
Unity gain frequency before the output filter	22MHz
Output noise (BW=80kHz, input terminated with 50Ω	42μV rms
Measured output offset voltage	+32mV

Distortion performance

V_{out}, pk-pk	1kHz	20kHz
5	0.0030%	0.0043%
10	0.0028%	0.0047%
20	0.0023%	0.0061%
40	0.0028%	0.0110%
80	0.0026%	0.0170%

Slew rate

Positive slew-rate	+320V/μs
Negative slew-rate	–300V/μs

Hotter Spice II

In this second article presenting a breakthrough in power mosfet modelling for audio design, Ian Hegglun adds compensation for mosfet electrothermal effects and answers the question of whether subthreshold conduction is needed for simulating crossover distortion.

The macromodels presented in the May issue do not show the effect of junction heating with time. This occurs when a curve tracer generates manufacturer data sheet curves.

Trying to fit models to published power mosfet curves will cause significant errors in the region where V_{DS} and I_D are both high and the sweep time is not short enough to limit junction temperature rise to only a few degrees.

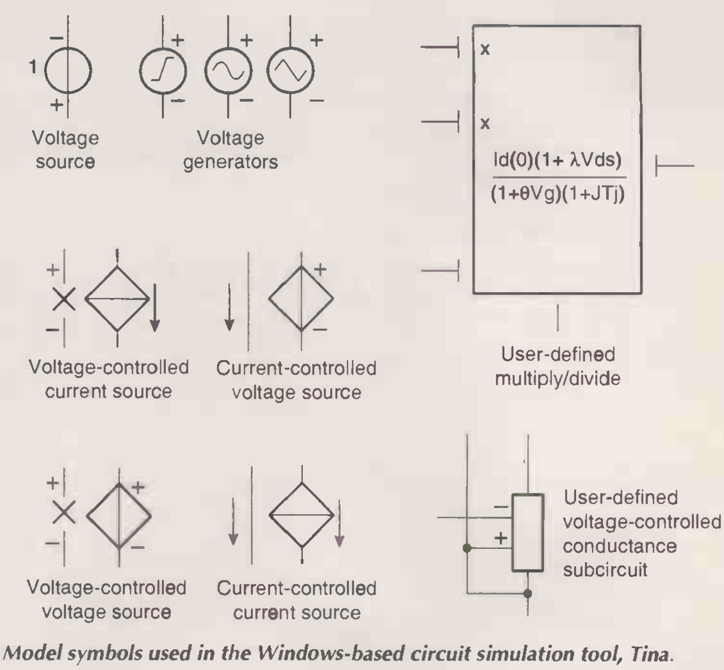
The only way to simulate mosfets subjected to large cyclical junction temperature changes is to use an electrothermal model. The effect of junction temperature can be modelled to a first approximation using,

$$\beta = \frac{\beta T_0}{1 + J\beta(T_J - T_0)}$$

$$V_{TH} = V_{TO} + J_{VTH}(T_J - T_0)$$

where J_β is the temperature coefficient for β, typically –0.5mV/°C for vertical mosfets,[1] and J_{VTH} is the temperature coefficient for the threshold voltage, typically –5.5mV/°C.

Spice fet models that include electrothermal feedback are rare. The Parker Skellern level 4 MESFET model is one. This is now included as standard in some of the latest versions of Spice. It can be adapted to behave like a mosfet by



Model symbols used in the Windows-based circuit simulation tool, Tina.

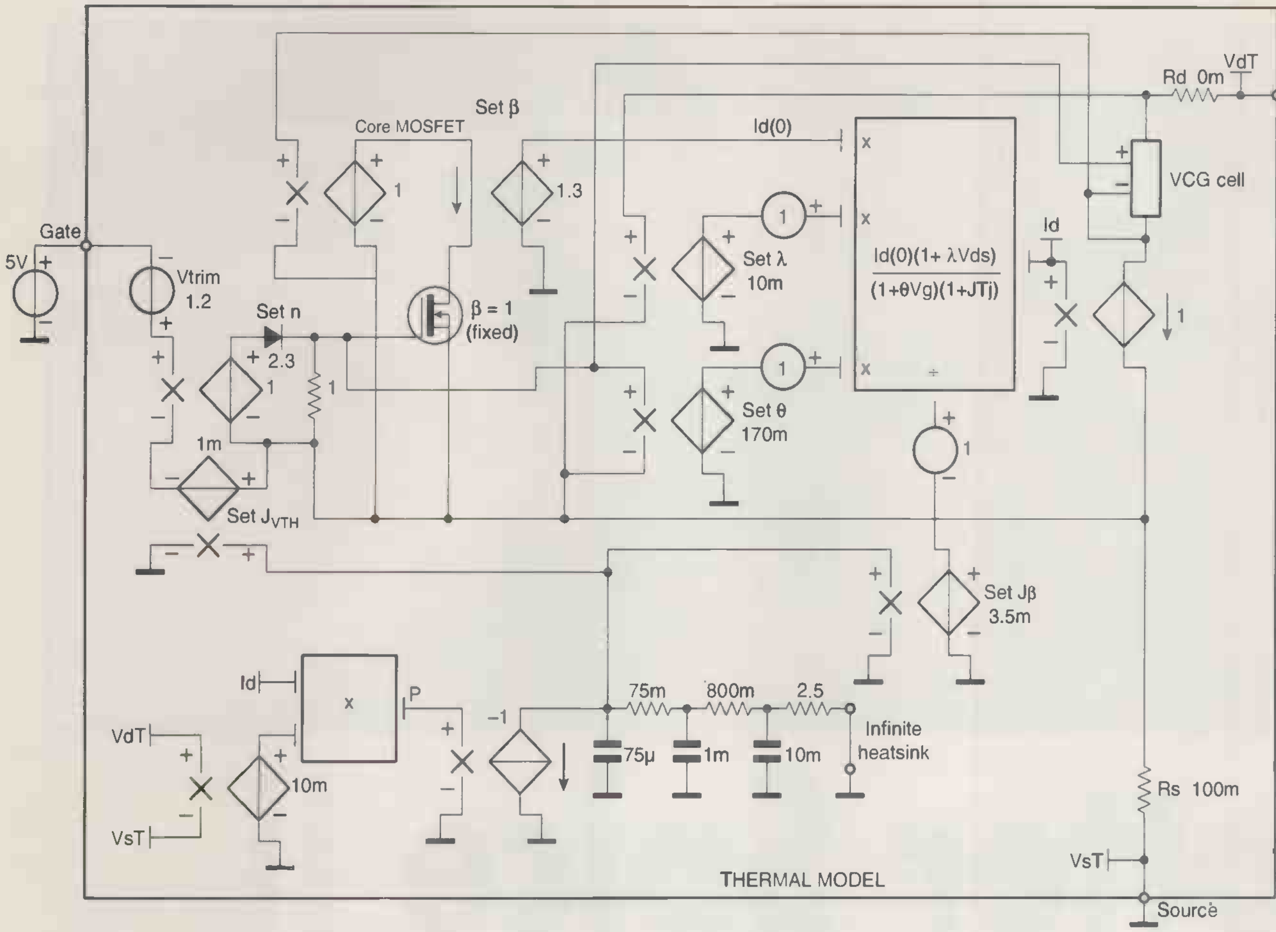


Fig. 1. An electrothermal model including subthreshold conduction, velocity saturation and smooth pinch-off. A two-quadrant multiplier/divider subcircuit is used to modify the current from the core mosfet and generates the overall drain conductance.

disabling the gate diodes, by setting IS=1e–30 and N=10. This also models subthreshold conduction and is reported to run fast.[2]

Figure 1 shows an electrothermal model, which includes subthreshold conduction, velocity saturation and smooth, pinch-off. A two-quadrant multiplier/divider subcircuit, Fig. C in the 'Maths' panel, generates velocity saturation and the overall drain conductance.

A four-quadrant multiplier, Fig. D, is used to calculate the instantaneous power dissipated by multiplying $V_{DS} \times I_D$, which allows for power dissipation in drain and source termination resistances.

The $V_{DS} \times I_D$ product is passed to a thermal model using a three-stage *RC* low-pass network that models the thermal impedance of the die and mounting. The resulting junction temperature appears at the input of the *RC* network, which then changes the threshold voltage and β parameter. If a V_{CG} cell is included in the drain path the parameter λ of the core mosfet must be set to zero as previously mentioned.

To set up the model, first fix the junction temperature to 25°C by inserting a fixed voltage source in place of the output of the *RC* thermal network. Fit values for β, θ, V_{TH}, λ and β_0, then raise the temperature setting to that of the data sheet and vary the temperature coefficient for β and V_{TH} until correct curve is achieved, as **Fig. 2a)**. Finally remove the fixed source and reconnect the thermal model.

An oscilloscope in X-Y mode allows thermal looping of the I_D/V_{DS} curve to be viewed Fig. **2b)**. Even at 2.5μs (100kHz) the junction temperature rises by 20°C. Data

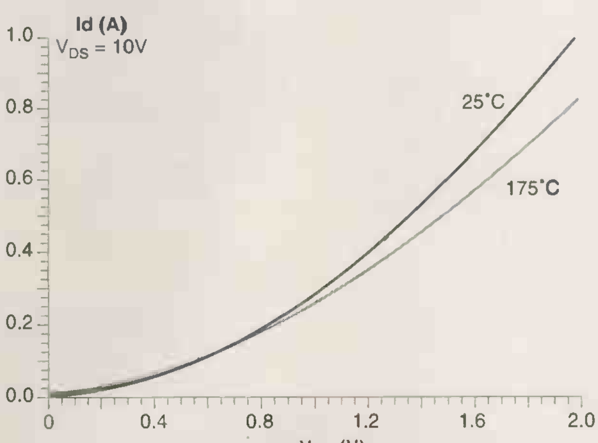


Fig. 2a). An I_D/V_{GS} plot with a fixed T_j for a 2SK134.

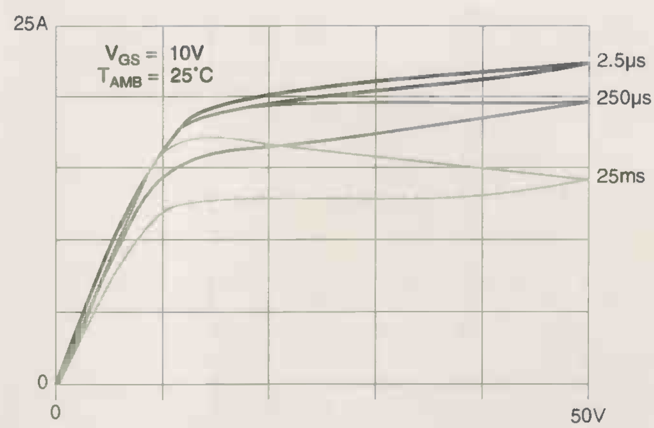


Fig. 2b). Thermal looping oscilloscope plot for different sweep rates when V_{GS} = 10V and T_{amb} = 25°C.

sheet curves do not usually show the return trace for 20μs pulses. It is significantly different from the forward trace at high currents.

Gain variations due to junction heating invariably affects large signal linearity in a power amplifier. As the frequency falls the amount of gain variation during each cycle will increase causing distortion to increase as frequency falls.

Note the peak transconductance depends on pulse duration for currents above 2A for V_{DS} of 30V. Hence, large signal linearity in a power amplifier will depend on frequency, with deviation increasing as frequency falls.

In most cases this type of nonlinearity is swamped by crossover nonlinearity and the increasing amount of negative feedback as frequency falls will offset the thermal distortion increase. However, the longer term thermally-induced changes in the bias point for class AB circuits usually lead to a detectable increase in distortion, particularly after significant change in the output power level.

Lateral audio mosfets such as the *2SK134* are almost unaffected by thermally induced bias point changes because the zero temperature coefficient point lies in the region for optimum bias. But not all lateral mosfets have this property – the *2SK400/2SJ114* pair being one exception.

The thermal model

A three stage *RC* low-pass model is used for the *IRF510* for accuracy down to 10μs pulses based on the data sheet. The roll-off slope of thermal impedance for the *IRF510* falls by one decade every two decades of pulse duration reduction, or half the rate of a simple *RC* integrator.

A three-stage network has its time constants staggered over six decades of time. The slowest *RC* time constant is assigned the value of 200s, one decade higher than the devices time constant of 20ms as seen on the log $R\theta_{JC}$/log(t_{PULSE}) graph, Fig. 3. As a result, the shortest pulse is placed at 20μs, where the value of $R\theta_{JC}$ is read as 75mΩ. This value is used for the first resistor closest to the junction.

The next resistor is the value of $R\theta_{JC}$ at 2ms, two decades to the right, read as 750mΩ. The final resistor value is $R\theta_{JC}$ at dc less the sum of the other resistors, i.e. 3.5–0.825=2.675Ω. The first capacitor is assigned the same numeric value as the first resistor, ie 75mF, and the second 0.75F. The last capacitor is the impedance extrapolated out to the slowest time – 200ms — giving 10°C/W or 10F.

An infinite heat sink is simulated using shorted output at the case. This keeps the thermal time constant to a minimum to reduce the simulation time to approach the final temperature; around 50ms to reach 90% of the final value in this case.

All diode capacitances and time delays should be set to zero to avoid instability. Wherever possible, the internal circuitry has been level shifted to ground reference most of the circuit in an attempt to minimise convergence problems.

In some cases it was necessary to enable the 'calculate operating point' option when performing a transient simulation so the simulation can get started. When simulating very low frequencies with the electrothermal model, the first two capacitors in the thermal model can be removed or set to zero.

Simulating crossover distortion

Distortion simulations are run in transient mode with a sine source with 500 time steps per cycle and applying Fourier analysis to one complete cycle.[3]

For most audio-frequency simulations, mosfet gate termination resistance, lead inductances and junction capacitances can be set to zero or left out to speed up simulations. Also, open loop simulations for the output stage only are faster and more likely to reach convergence than closed loop.

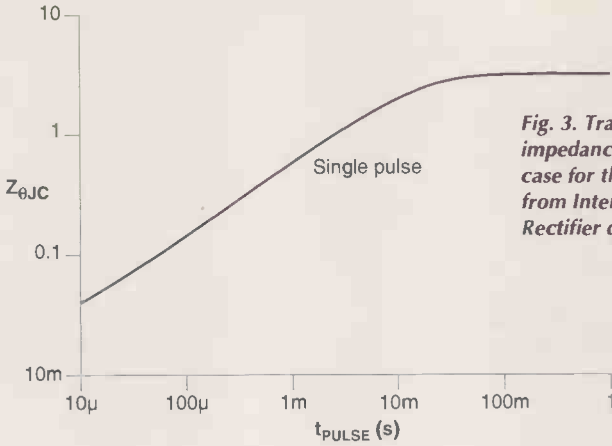


Fig. 3. Transient thermal impedance junction to case for the IRF510, *from International Rectifier data.*

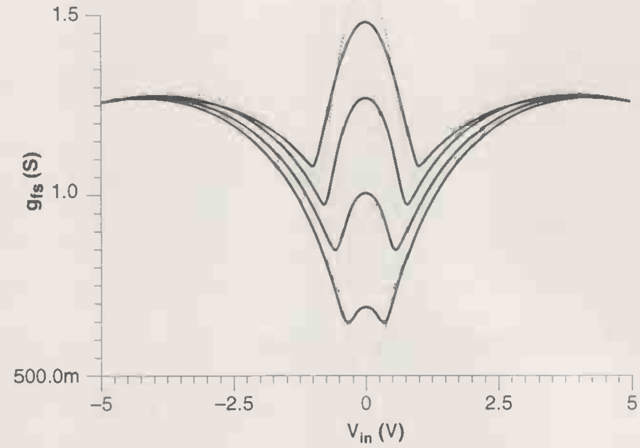


Fig. 4. Comparison of 2SK134/2SJ49 *with subthreshold conduction (solid) and without (dotted). Fig. 2's macromodel from my last article is used here with a level 1 core mosfet (β=1.15, λ=10m, R_s=400mΩ). Diode* n*=2.3, I_s=100pA and V_{trim}=1.2V. The load is 8Ω and supply rails are 55V. Bias voltage is stepped from 0.4 to 1.0V with subthreshold conduction and 0.2 to 0.8V without subthreshold conduction.*

In closed loop, distortion of the entire circuit may be hard to measure due to limited numeric accuracy. With the Tina[4] circuit simulator, Fourier series analysis resolves to 0.0002% thd – sufficient for most closed-loop simulations.

Figure 4 shows results from a class AB simulation of a simple output stage using a matched *2SJ49/2SK134* pair tested open loop with an 8Ω load and 55V supply rails, Fig. 5. When subthreshold conduction is included, using the macromodel in Fig. 2 of my previous article, gain curves for various bias settings turn out to be surprisingly similar to the reference level 1 models once the 200mV difference in the bias settings is removed. Although I initially tried the level 3 mode, I could not get it to run.

With subthreshold conduction the curves are smoother. Thd results are similar – around 1% for 50Vpk swing using a similar bias.

In reference 6 of my previous article, the bias voltages were fixed at the bench test values. This suggests that the

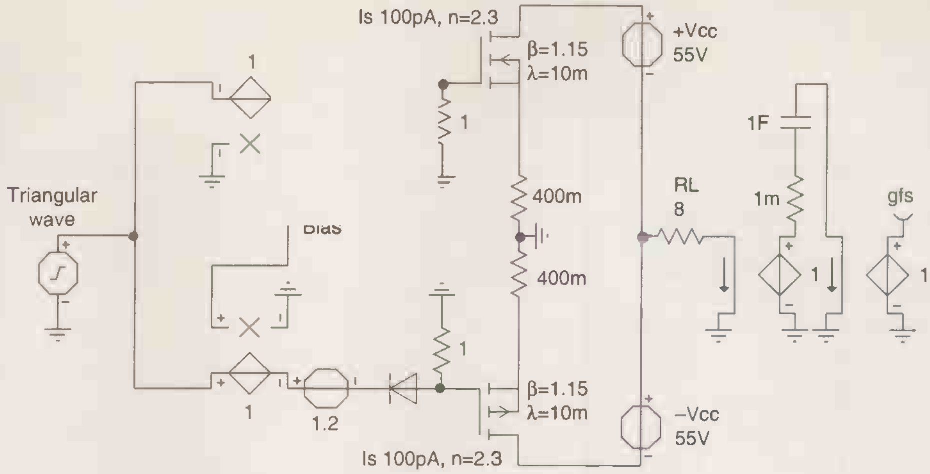


Fig. 5. Circuit used to generate the open-loop gain curves of Fig. 5 using Tina. Closed-loop gain obtained by connecting the load to the source and moving the earth to the supply centre tap. The bias generator can be stepped to generate a family of curves.
Note: Source resistors are part of the macromodel for generating velocity saturation.

simulated thd values are significantly different from the measured thd because of significantly different biasing points due to the choice of the V_{TH} parameter rather than the choice of model.

This seems to answer my question of whether sub-threshold conduction is needed for simulating crossover distortion. In most cases it seems the simulated thd can be predicted reasonably well with a simple level 1 model if quiescent current is the reference parameter.

For cases where error correction is used to reduce distortion, the sensitivity to model error is increased. One way to discover whether the extra parameters are needed or not is to run a simulation with and without these effects.

If, when adding or dropping an effect, the results change significantly, then the parameter changed should be included in the model. Otherwise it can be left out to improve the simulation time.

Maths functions for Spice

The Schockley equation can be used to make log and anti-log subcircuits. By setting a user defined diode with zero series resistance, I_s=1A and the diodes emission coefficient parameter n to 27.88 – ie, $1/V_T$ at 27°C – gives $I_D=exp(V_D)-1$. Converting I_D to voltage and adding 1V gives $V_{out}=exp(V_D)$, **Fig. A.**

The natural logarithm can be generated by applying a voltage controlled current and measuring the diode voltage giving $V_{out}=\ln(V_D+1)$. Subtracting 1V from the input gives $V_{out}=\ln(V_D)$, **Fig. B.**

A multiplier can be constructed by adding logs then taking the antilog. Negative inputs can be handled by offsetting an input by a positive amount and the additional expansion terms are removed by subtraction. **Figure C** shows a combination two quadrant multiplier and divider.

Input 1 may vary in the range of –99.9V to several kilovolts positive while inputs 2 and 3 must be greater than zero. Output voltage is given by $V_{out}=V_1 \times V_2/V_3$.

A four-quadrant multiplier can be constructed as in **Fig. D** while **Fig. E**, also a four-quadrant multiplier, uses mosfets. In Fig. E the output is valid when $|V_1|+|V_2|<V_{offset}$.

Simulators may only allow V_{GS} to go to 30V but this restriction can be overcome by scaling the input to allow a much higher input level and then scaling the output to compensate. The mosfet drain supply also needs to be greater than the highest gate voltage for square law operation.

Some versions of Spice 2 allow the voltage-controlled voltage source, or

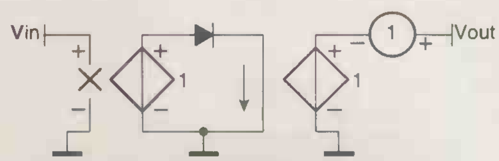


Fig. A. Implementing the exponential function using an ideal diode.

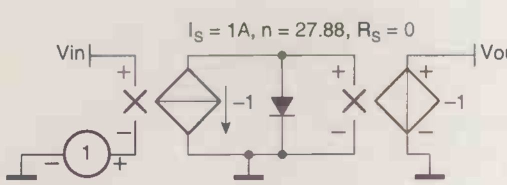


Fig. B. Implementing natural logarithms.

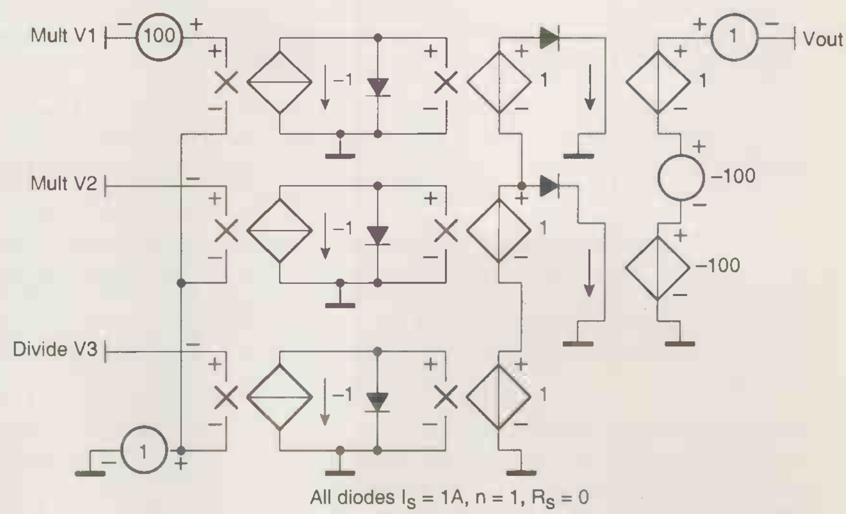


Fig. C. Implementing a two quadrant multiplier/divider with diodes.

Getting started

Anyone who has used a circuit simulator before should not have too much difficulty translating these macromodels to suit their version.

Those beginning from scratch are best advised to choose a simulator with schematic capture; TINA and Electronics Workbench are two already mentioned.

There are several others available free as demos or student versions on the net, although most are too crippled to be used for these macromodels. A number of text based versions are not crippled but require more learning to get going.

In Electronics Workbench, subcircuits are called custom parts and can be placed into a circuit more than once, but in TINA there is no facility for subcircuits. Instead the copy and paste clipboard must be used to duplicate macromodels.

When defining circuit blocks in TINA it is best to avoid using the interconnect pins because the labels get copied and must be changed one by one, leading to possible errors. Instead run wires for all connections.

Test each subcircuit first and save it before copying. If you want to simulate a complex amplifier circuit, do it in stages by getting the output stage running first then add the other circuitry.

Anyone using Spice will sooner or later discover non-convergence problems. Tina's help file under 'Set Parameters' in the 'Analysis' section may help. More information can be found from references 1 and 5.

Simulators often require the same diagnostic skills as for real circuits just to get them to run properly. Once you get to know their limitations and shortcomings, you you should find it possible to live with them. ■

Ian's first article on this topic was presented in the May issue. As the two articles are closely linked, this one should have been presented in the June issue. Sorry for the gap. Ed.

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4. C. Ouseby, 'Design Lab Bytes', *Electronics World*, September 1996, p663-665. http://www.eece.napier.ac.uk/~tina/
5. T. L. Quaries, 'Analysis of Performance and Convergence Issues for Circuit Simulation', University of California, Berkeley, ERL memo, M98/42.

VCVS, to be configured as a multiplier. This can be done using the POLY x x 0 0 0 0 1 statement, as described in the book 'Spice: Practical Device Modelling' by R. Kielkowski (McGraw Hill ISBN 0-07-911524-1). This statement simplifies some macromodels as well as making them run faster.

Some of the graphical-based simulators such as *Electronics Workbench* include a multiplier, but note that the multiplier in *Workbench V4* cannot be used in negative feedback arrangement. Workbench version 5/EDA has overcome this problem.

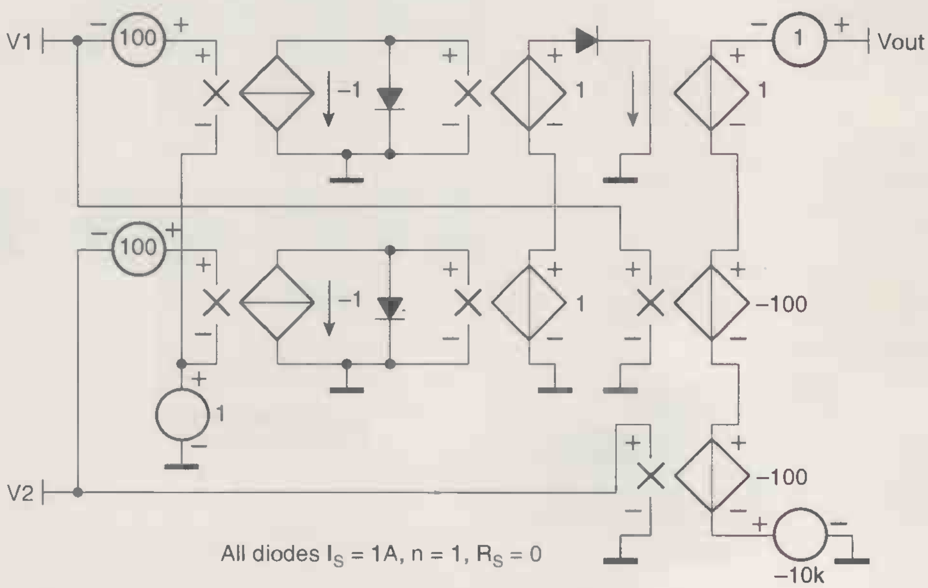

Fig. D. A four-quadrant multiplier using logs and antilogs

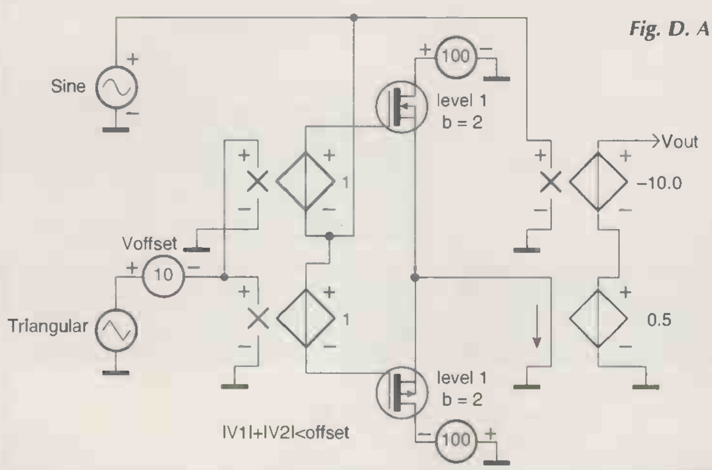

Fig. E. A four-quadrant multiplier based on the difference of two square laws.

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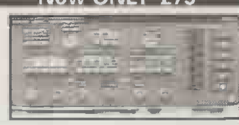

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Early applied

Bryan Hart illustrates the importance of the 'Early' effect by describing its influence on four common circuit configurations – common base, common emitter, current mirror and long-tailed pair.

Last month I looked at the theory behind the 'Early' voltage, and explained its importance in the design of high-gain, small-signal amplifiers working at low frequencies. This article looks at the effect of the Early voltage on real discrete transistor configurations including the current mirror and long-tailed pair.

In the numerical examples that follow, it is assumed that, β=100, so βV_T=2.5 V, V_{BE}=0.7V and V_A=100V.

The common-emitter amplifier

Figure 1 shows a common-emitter amplifier stage with necessary DC bias components omitted. They set the quiescent output voltage at $V_O=V_{CC}-V_R$, but play no role in small-signal calculations based on the equivalent circuit of Fig. **1b**). The input signal is v_s and output signal v_o. By the voltage divider principle,

$$G = \frac{v_o}{v_s} = -\frac{r_\pi}{r_x + r_\pi} \times \frac{\mu R_c}{R_c + r_o} \qquad (1)$$

Since $r_\pi=\beta V_T/I_{CQ}$, you can ignore r_x in comparison with r_π if you operate with,

$$I_{CQ} \leq \frac{\beta V_T}{r_x}$$

For design purposes this can be interpreted as,

$$I_{CQ} \leq \frac{\beta V_T}{10 r_x}$$

Provided $r_x \leq 50\Omega$, this means $I_{CQ} \leq 2.5$mA – a condition I will assume from now on.

The maximum magnitude, $|G_m|$, of G is achieved with $R_C >> r_O$: then, $|G_m|=\mu\approx 4000$. This value is only approached with an active load, one example of which is the output circuit of a current-mirror – dealt with later – using p-n-p transistors.

However, employing an active load necessarily involves the use, also, of a DC feedback network connected from a second or later stage to the input circuit of the stage in question, in order to guarantee a value for V_O in the linear amplifying region. This is the case with some operational amplifier designs.

For the present case of a resistive load, $R_C=V_R/I_{CQ}$

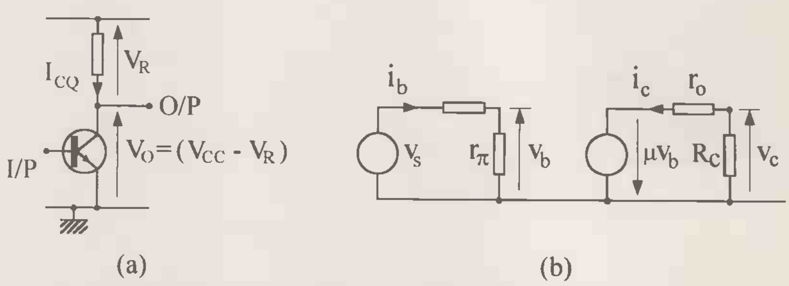


Fig. 1. Common-emitter amplifier stage, a), and its equivalent circuit, b).

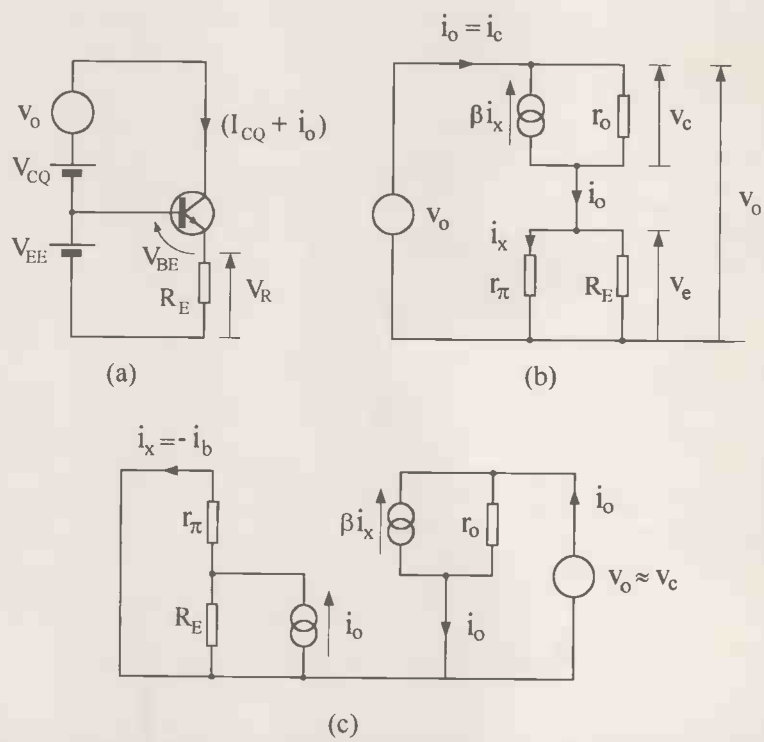


Fig. 2. A common-base current generator, a), an equivalent circuit, b), and a simplified form of b) emphasising feedback action, c).

Substituting for R_C and r_o in eqn 1 then gives,

$$G=-\mu\frac{V_R}{V_A+V_R}$$

Since V_A>>V_R, this reduces to,

$$G\approx-\frac{V_R}{V_T}$$

Assume that V_{CC} is 10 V and suppose the bipolar-junction transistor is biassed for maximum symmetrical output voltage swing, $V_R \approx V_{CC}/2$, then G is approximately –250.

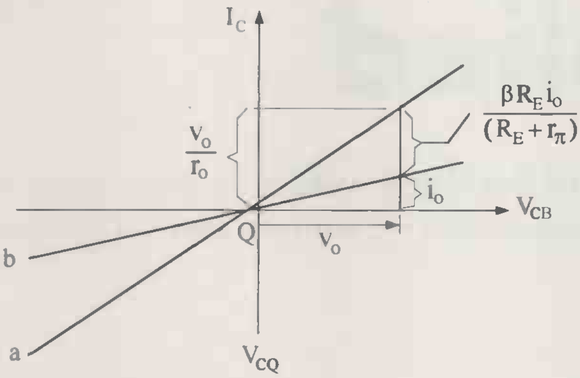


Fig. 3. Relating the common-base output characteristics with constant R_E, a), to that of a common-emitter stage, b). Not to scale.

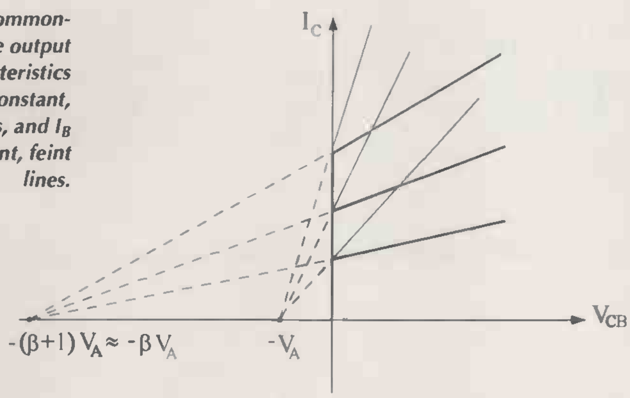


Fig. 4. Common-base output characteristics with I_E constant, bold lines, and I_B constant, feint lines.

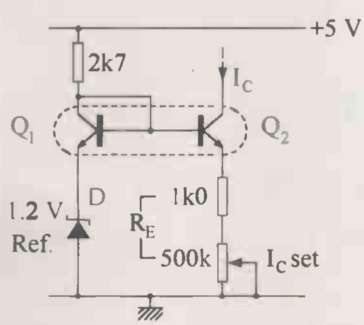


Fig. 5. A practical circuit solution of a current generator design problem in the text.

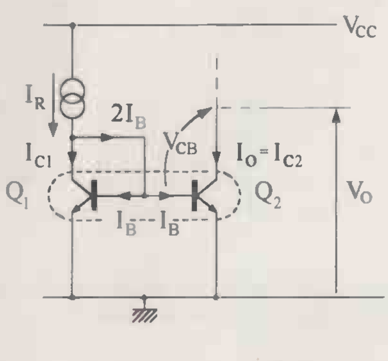


Fig. 6. Basic current-mirror.

The common-base stage

Figure 2a) shows a common-base stage intended to operate as a current generator and Fig **2b)** shows a small-signal equivalent circuit of it. The problem is to develop a design formula for the output resistance R_o for a given operating current.

By inspection, v_c>>v_e because r_o>>r_π and the current in r_o is much greater than that in r_π. Consequently, r_π, R_E can be omitted from the output circuit, but they cannot be omitted from the input circuit. This is because the fraction of the output current, i_o, flowing in r_π controls the current generator βi_b.

Figure **2c)** – a simplified and redrawn version of Fig 2b) – emphasises the negative feedback action that leads to a resistance transformation at the collector of the bipolar junction transistor.

$$i_o=i_c=\beta i_b+\frac{v_o}{r_o}$$

But,

$$i_b=-i_x=-i_o\frac{R_E}{R_E+r_\pi}$$

$$\therefore i_o\left[1+\beta\frac{R_E}{R_E+r_\pi}\right]=\frac{v_o}{r_o}$$

and,

$$R_o=\frac{v_o}{i_o}=\left[1+\beta\frac{R_E}{R_E+r_\pi}\right]r_o \tag{2}$$

Figure 3 is a graphical interpretation of eqn 2. It shows how the output characteristic for a fixed value of R_E, line b, is related to the characteristic for constant I_B, line a.

For fixed values of I_E, you obtain the common-base characteristics by making R_E infinity in eqn 2. These are shown in **Figure 4**: now, $R_O=(1+\beta)r_o$. The Early voltage may be regarded as increased by a factor β.

For design purposes, you should make the substitutions $R_E \approx V_R/I_{CQ}$, $r_\pi=\beta V_T/I_{CQ}$ in eqn 2. Then,

$$R_o=\frac{V_A}{I_{CQ}}\times\left[1+\beta\frac{V_R}{V_R+\beta V_T}\right] \tag{3}$$

Problem: Design a simple temperature-insensitive current generator having an output current that can be set precisely at 1mA and an output resistance exceeding 3MΩ. The supply rail is 5 V.

Solution: Substituting figures in eqn **3**, you will find that V_R is greater than 1.02V. **Figure 5** is one possible solution to the problem. The V_{BE}s of $Q_{1,2}$ – part of a matched array – balance and track with temperature. This is because they operate at approximately the same current, so V_R is equal to the voltage provided by the 1.2V reference source.

The potentiometer permits precise setting of I_C.

The current mirror

For the basic current mirror shown in **Fig. 6**, Q_1 and Q_2 are assumed to have identical characteristics. The relevant circuit equations are,

$$I_B=\frac{I_{C1}}{\beta_0}$$

$$I_R=I_{C1}+2I_B$$

$$I_O=I_{C2}=I_{C1}\left(1+\frac{V_{CB}}{V_A}\right)$$

From these we obtain,

$$I_O = I_R\left[\left(1+\frac{V_{CB}}{V_A}\right)\div\left(1+\frac{2}{\beta_0}\right)\right]$$

Using the binomial theorem, then multiplying out and neglecting second and higher order terms, gives:

$$\lambda = \frac{I_O}{I_R} \approx 1-\frac{2}{\beta_0}+\frac{V_{CB}}{V_A} \qquad (4)$$

As β_0>>1 and V_A>>V_{CB},, λ is close to unity: also, R_O=r_o because ΔV_{CB}=ΔV_O if V_{BE} is constant.

Both λ and R_o are improved using the scheme in **Fig. 7a**), which is the 'Wilson' current mirror with an additional transistor.[1]

The V_{BE} of Q_4 balances that of Q_3, so Q_1 and Q_2 operate at the same V_{CB} and the third term in eqn 4 is zero. The second term also effectively disappears because the current $2I_B$ is supplied from the emitter of Q_3 rather than by I_R. Consequently, λ is unity with a tolerance that depends on the V_{BE} matching of Q_1 and Q_2 and can be as small as 1%.

Figure 7**b**) is an equivalent circuit for the calculation of R_O. It has been simplified using the same reasoning that was employed to derive Fig 2c) from 2b),

By Kirchhoff's current law, i_o=$2i_x$. Furthermore,

$$i_O = -\beta i_x + \frac{v_O}{r_O}$$

$$R_O = r_O\left(1+\frac{\beta}{2}\right) \approx \beta\frac{V_A}{2I_{CO}}$$

Experimental results attest to the added benefits of the quad-transistor current mirror. **Figure 8** shows the output characteristics using a CA3045 matched bipolar-junction transistor array. The near-vertical line close to the origin was obtained with a collector-base strap on Q_3 and denotes a boundary line for saturation.

A word of caution though. If source I_R is replaced by a resis-

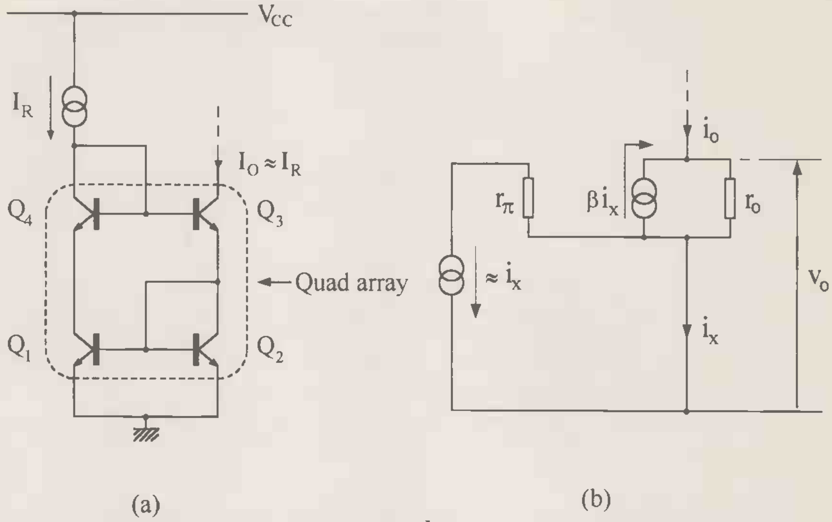


Fig. 7. A quad-transistor current mirror, a), and a simplified equivalent of a) for finding the incremental output resistance, b).

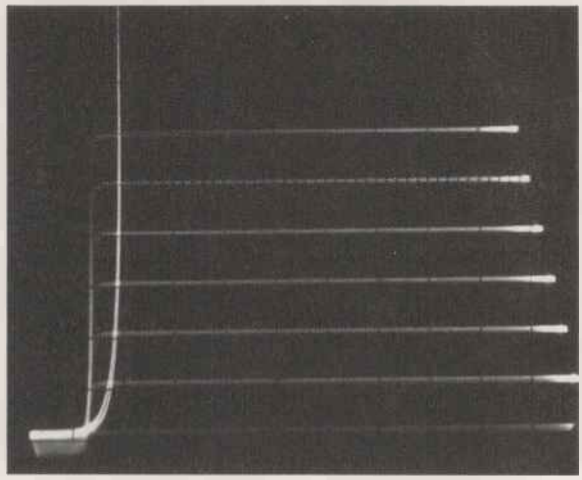


Fig. 8. Output characteristics for circuit of 7(a) using a CA3045 bipolar-junction transistor array. Horizontal scale: V_{CE}=0.5V/div. For V_{CE}>1.5V, the vertical spacing between each characteristic is 100μA.

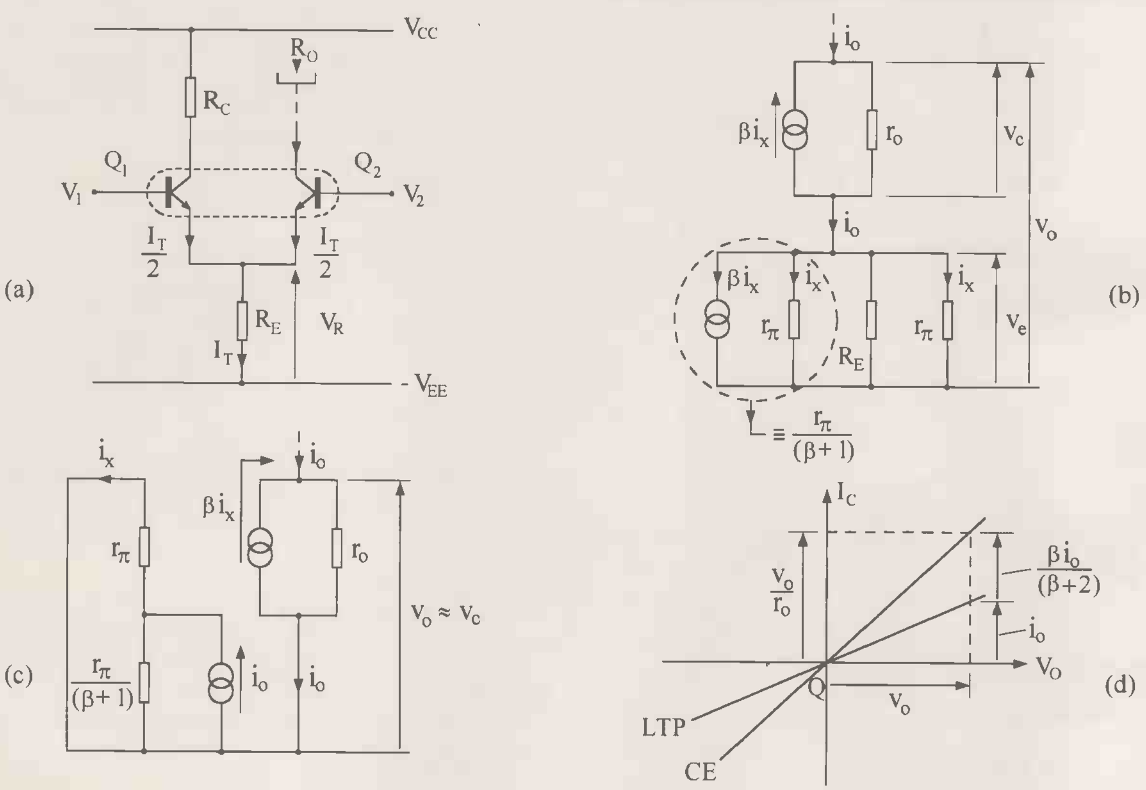


Fig. 9. Section a) shows a long-tailed pair, b) shows an equivalent circuit for finding R_o. Section c) shows a simplified form of b) assuming R_E>>r_π/(β+1) while d) shows how long-tailed pair output characteristics are derived from those of a common-emitter stage.

tor R_S across which the potential difference is V_R it can be shown, after some tedious algebra, that,

$$R_O \approx \beta \frac{\beta T_O V_R}{2V_R + V_T} \tag{5}$$

If V_{CC}=5V, R_o as calculated from eqn (5) is only some 75% of what would be obtained using an ideal current source at the input.

The long-tailed pair

To calculate R_O for a long-tailed pair, **Fig. 9a)**, both bases are set at AC ground on the equivalent circuit, Fig. **9b)**. The base currents of $Q_{1,2}$ must be equal because they both experience the same change in emitter base voltage, v_e.

A simplified form of Fig. 9b) is shown in Fig. **9c)**: it assumes v_c>>v_e as with the common-base stage, and that,

$$R_E >> \frac{r_\pi}{\beta+1}$$

$$i_x = \frac{i_o}{\beta+2}$$

and,

$$i_o = -\beta i_x + \frac{v_o}{r_o}$$

so,

$$R_o = r_o\left(1 + \frac{\beta}{\beta+2}\right)$$

or, as β>>1,

$$R_o \approx 4\frac{V_A}{I_T} \tag{6}$$

Since $R_E=V_R/I_T$ and $r_\pi=2\beta V_T/I_T$, the condition,

$$R_E >> \frac{r_\pi}{\beta+1}$$

is met if V_R is much greater than $2V_T$. This is always the case in practice.

Figure **9d)** shows how the output characteristic for a fixed R_E is related to that for a fixed I_B.

Problem: Design a long-tailed pair having an I_T of around 2mA and R_o greater than 10MΩ at either output point.

Solution: It is clear from eqn 6 that if I_T is 2mA then R_o is 200kΩ, so a straightforward long-tailed pair is unsuitable.

However, the required R_o can be achieved with the added cascode scheme shown in **Fig. 10**. For this, eqn 2 applies with R_E>>r_π. So $R_O=2V_A(1+\beta)/I_T$, i.e., R_o is approximately 20MΩ.

Tail current I_T can be supplied by a circuit such as that in Fig. 5.

In conclusion, consider a long-tailed pair with a current mirror as an active load, **Fig. 11a)**. Referring to the equivalent circuit of Fig. **11b)**, the collector current change in Q_{N1} is equal in magnitude, but opposite in sign, to that i_y of Q_{N2} and this is reflected by the current mirror so that,

$$i_o = 2i_y + \frac{v_o}{r_{op}}$$

But,

$$i_y = \frac{v_o}{R} = \frac{v_o}{2r_{on}}$$

$$\therefore R_o = \frac{v_o}{i_o} = r_{on} // r_{op} \tag{7}$$

In these equations, r_{on}, i.e. V_{AN}/I_C, is the collector resistance of Q_{N2} and r_{op}, which is V_{AP}/I_C, is the collector resistance of Q_{P2}. Hence,

$$R_o = 2V_{AN}\frac{V_{AP}}{I_T \times (V_{AN} + V_{AP})} \tag{8}$$

■

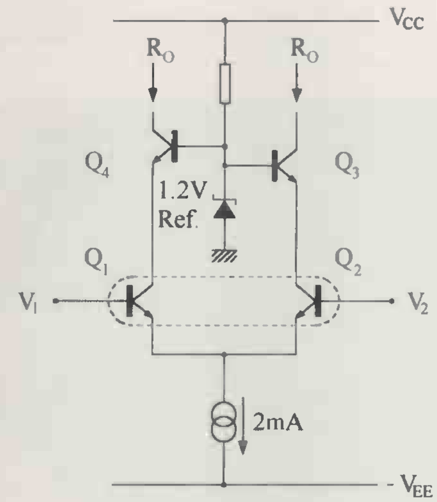


Fig. 10. A circuit solution to a current-mirror design problem described in the text.

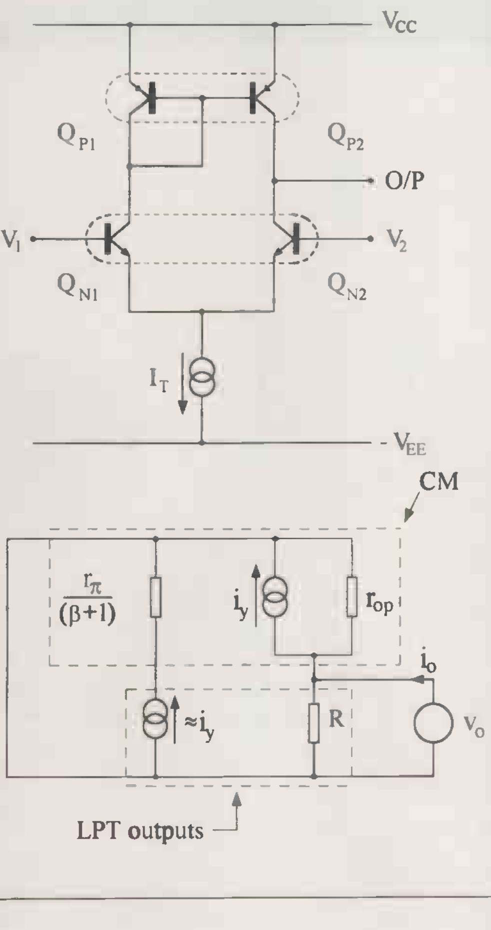


Fig. 11. A long-tailed pair with active load is shown in a) and an equivalent circuit for finding R_o is shown in b).

Reference

1. Hart, BL, and Barker, RJ, 'DC matching errors in the Wilson current source,' *Electronics Letters*, Vol. 12, No 15, pp. 389-390, July 1976.

Brighter driver for EL lamps

Andy Wolff looks at a new electroluminescent lamp driver chip that produces 220V pk-pk from a couple of penlight cells with only five additional passives.

Andy is Applications Manager, Display Products at IMP Incorporated in San Jose.

The popularity of electroluminescent lamps is shown by their almost universal use for backlights, signs and keyboards by the makers of pagers, cellphones and digital notepads.

IMP has made electroluminescent lamp drivers a major technology focus. By combining high-voltage process technology and clever analogue circuitry, IMP has become a major EL driver innovator in the last year.

The company's newest product is the *IMP528*, an eight-pin surface-mount device packaged in both SOIC and MSOP that generates 220V pk-pk from a battery or from fixed inputs in the range 2 to 6.5V dc.

An internal feedback circuit makes battery operation especially efficient using several techniques. One of these, shown in **Fig. 1**, regulates the output by skipping pulses in its step-up oscillator when batteries are fresh, and restoring them as they age. This has the effect of making brightness appear constant almost to the point when the battery quits.

Figure 2 shows a generalised glimpse of the innards of the device, **a)**, and the actual circuit connections, b). These diagrams show how the boost converter section charges C_S to high voltage of 110V dc and the high-voltage switching bridge that applies it to either side of the lamp at 220V pk-pk. The two oscillators are individually tunable via R_{SW} and R_{EL} for optimal efficiency and lamp colour.

While a single supply is the most common operating scheme, certain applications may separate the chip supply, V_{DD}, from the main supply to the inductor, V_L. This is frequently done in automotive applications – 5V for V_{DD}, 8-12V for V_L – and pagers – 1.5V for V_L, doubled to 3V for V_{DD}. IMP also supplies drivers that operate directly from one cell and require no doublers.

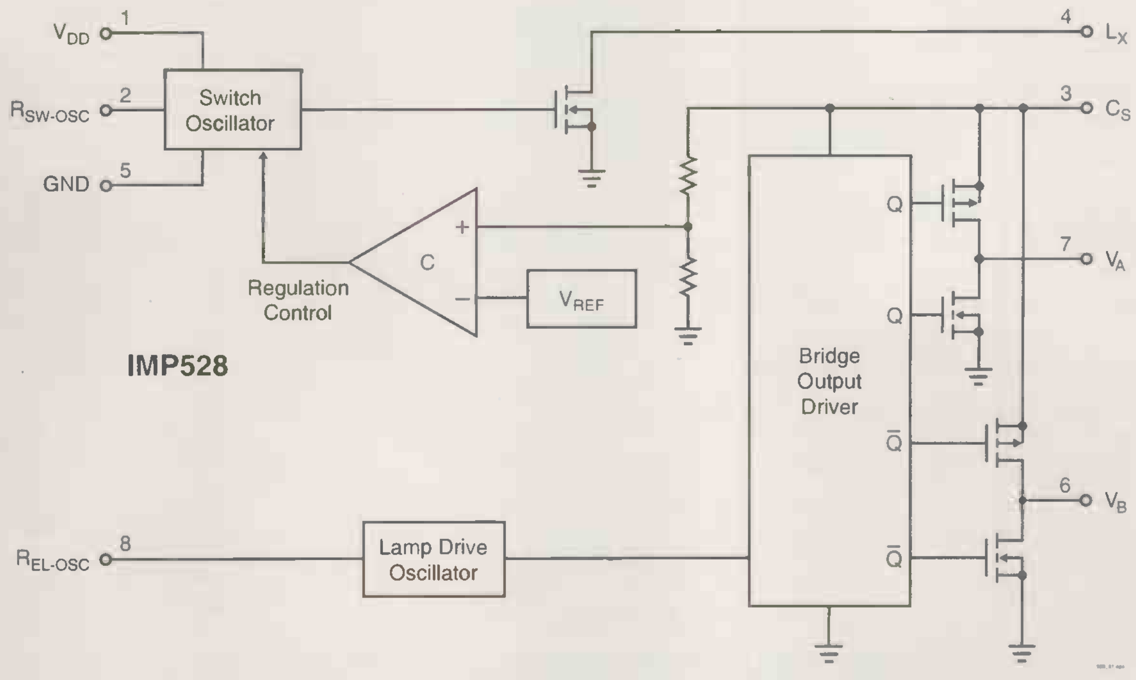
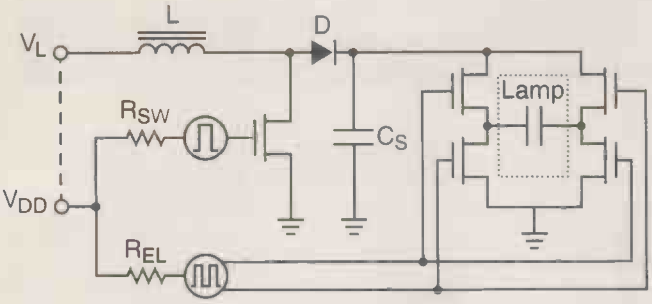


Fig. 1. IMP528 internals showing the feedback control that keeps light intensity almost constant as battery voltage falls. The device drives lamps at high brightness up to 50nF capacitance – or more but with decreasing light output.



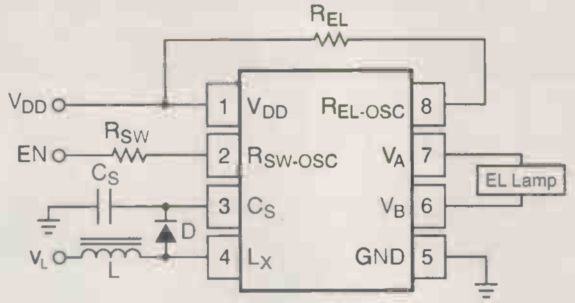
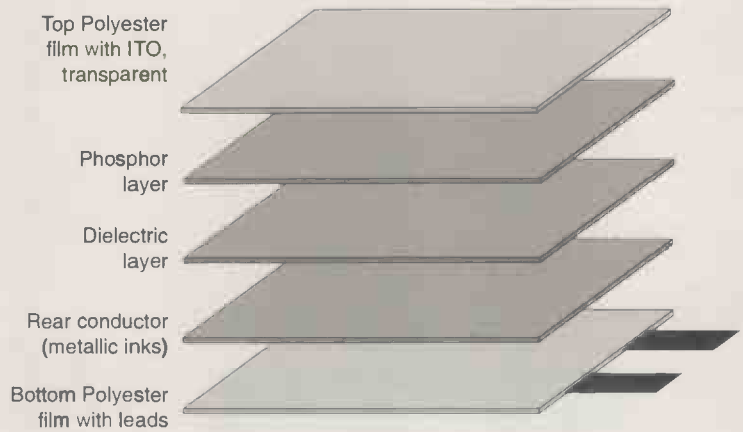


Fig. 2a) Generalised schematic of the integrated electroluminescent lamp driver and b), the complete circuit. This high-voltage CMOS chip eliminates the need for an external protection resistor in series with the EL lamp. Typical V_{DD} current consumption is 420μA.

You can see in Fig. 2**b**) that an 'enable' function can also be achieved by connecting R_{SW} to V_{DD} or ground. The ability to optimise each component that contributes to efficiency and performance is the hallmark of these devices. Values for L can range from 100μH–1mH, and for C_S, 10-100nF.

Final selection is made on the basis of the characteristics of the lamp selected – its size, colour and desired brightness. The high boost converter frequency used of around 70kHz permits the use of tiny components on the board. ■



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Comprising a sensor mounted in a small casing, a CD ROM containing software development kit and comprehensive documentation, and all the necessary cables, the FingerTIP Evaluation Kit from Siemens Semiconductors is intended as an introduction to biometric fingertip recognition technology. The evaluation kit's sensor can be plugged directly into the parallel port of a PC and can be powered directly from a PC mouse port or via the separate power adaptor supplied. The FingerTIP Software Developer's Kit contains all the software needed to easily integrate the system into any application.
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Tel: 0990 550500
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PCMCIA cardkit for DECT phones

Robinson Nugent has developed a custom PCMCIA cardkit for use in wireless data communications based on the DECT digital cordless telephone standard.
It has a type III cardkit as the basic housing. In order to integrate with the DECT system a transmitter/receiver and antenna is built into the back of the frame assembly. When a signal is received by the antenna identifying a local base station a LED module, which is built onto the end of the PCMCIA card frame, is activated, the light being reflected through the transparent LED housing.
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Four-channel, real-time, 100MHz digital oscilloscope

Tektronix's TDS224 is a four-channel, real-time, 100MHz digital oscilloscope. It is the latest member of the TDS200 series. The 60 and 100MHz scopes are designed for digital design engineers and technicians working with 8-bit microcontrollers. Using oversampling technology, all three models, the TDS210, TDS220, and the TDS224, sample at 1Gsamples/s, providing full-bandwidth single-shot acquisition simultaneously on all channels. This technology enables users to capture non-repetitive signals that would be invisible to analogue oscilloscopes and distorted by DSOs without sufficient sample rate.
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Digital audio for set-tops

Micronas Intermetall has added a set-top box interface to its multistandard sound processor (MSP). In addition to all analogue TV audio standards, the MSP 34x8G also supports the digital interface with digital or hybrid (analogue/digital) set-top boxes and TV receivers.
The audio devices are also capable of processing digital audio signals such as those delivered by the digital TV broadcasting decoders on the basis of the standards DVB (Europe) or ATSC (US).
Micronas Intermetall
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Gigabit Ethernet Fibre Channel ICs

Vitesse Semiconductor has announced a family of ICs for Gigabit Ethernet, Fibre Channel and serial interconnect applications at speeds up to 1.25Gbit/s. Initial products include the VSC7123 and VSC7133 10-bit transceivers. These two fifth generation devices provide the serialiser/deserialiser function for 10-bit, 8B/10B encoded data between 1.0 and 1.25Gbit/s. They feature low jitter generation, cable equalisation, reliable signal detection and reduced sensitivity to power supply noise and reference clock jitter.
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Win CE platforms

Advantech has announced a system-level solution, which includes a compact chassis and a palm-size single board computer with an on-board solid-state disk. The disk comes pre-loaded with Windows CE v2.1 to reduce system integration time. Also available are boards with a disk loaded with Windows CE. According to the supplier, one difficulty in dealing with Windows CE is that it is very hardware dependent. There may often be times when no drivers are available for the desired device. Secondly, lack of both familiarity and support for Windows CE development tools may cause difficulties. One must also spend time looking for a suitable hardware platform on which to run Windows CE.
Advantech
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SM fuses

Bussmann has extended its range of surface mount circuit protection components, with the addition of a time lag fuse. The 6125TD Brick Fuse is rated at 125V AC or 60V DC, making it suitable for use in primary circuits and to replace glass tube fuses in certain applications. It is

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designed for applications where it must withstand transient current surges, yet open safely when subjected to continuous fault currents. It is designed to carry 200 per cent of its rated current for a minimum of one second, ensuring it does not open prematurely due to harmless transients. It measures 6.1mm x 2.5mm x 2.5mm.
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Enq No 508

Comms adapter

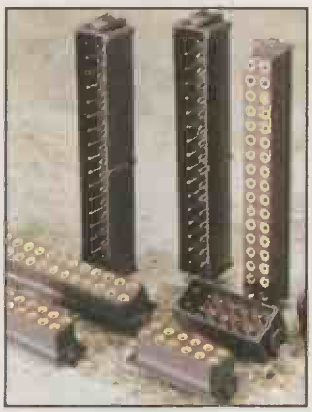
Motorola has announced the MPMC860 communications adapter in PCI mezzanine card format for T1 and E1 connectivity. Powered by the PowerQUICC MPC860 quad integrated communications controller, the adapter is for use in high-bandwidth voice and data switching systems. An optional controller allows connection to an SCbus switching fabric to provide a high-bandwidth TDM telephony bus. The MPMC860 PMC board has two T1 or E1 interfaces connected via RJ45 connectors on the front panel or via a rear transition module. A carrier board with two adapters can provide up to four E1 and T1 interfaces in a CompactPCI or VME slot.
Motorola
Tel: 0171 386 1499
Enq No 509

Flat panel controller

Microchip has introduced what it claims to be the world's smallest 8-bit ROM microcontroller in an 8-pin package. The PIC12CR509A features 1024 words of ROM program memory, 41 bytes of user RAM, six I/O pins with on-chip clock oscillator, 33 single-word instructions, full speed 1μs instruction cycle at 4MHz, seven special function hardware registers and a two-level deep hardware stack.
Other features include an 8-bit real time clock/counter with 8-bit programmable prescalar, watchdog timer, direct LED drive, 2.5-5.5V operating voltage and less than 2mA at 5V, 4 MHz power consumption. The PIC12CR509A is available in 8-pin PDIP and SOIC packages.
Microchip
Tel: 0118 921 5858
Enq No 509

SM connectors

Harwin has launched surface mount connectors on a 2mm pitch. The Datamate connectors are for use in medical, computer and aerospace industries. They come with four to 44



positions and have a board stacking height of 7.85mm maximum. The termination, which incorporates a surface mount pad, keeps the solder point within the profile of the connector moulding, allowing for more tracking space outside the connector profile. The connectors are packaged in a tube as standard with tape and reel available as an option.
Harwin
Tel: 01705 370451
Enq No 510

16Mbit 1.8V flash

AMD is offering the Am29SL160C 1.8V only, 16Mbit flash memory device. The device offers read access times of 100ns.
The flash device features a power-management system, which conserves power by automatically putting the flash device into sleep mode during inactive periods. There is no latency/wake-up time when the system subsequently accesses the device.
In addition it features electronic serialisation, accelerated programming mode and hardware write protection.
AMD
Tel: 01276 803100
Enq No 511

Configurable enclosure

Rittal's Quickrack enclosure can be configured to specification. Dismantling and reassembly takes less than five minutes because of the design of the multi-folded frame sections. Supplied fully screw fixed, it comes with a glazed front door, sheet steel frame, changeable door hinge and single-pane safety glass window. The vented rear door is made of sheet steel. Also included are front 482.6mm mounting angles that are adjustable on depth stays to which a rear pair of mounting angles can also be fastened. Metric 535mm mounting angles can also be installed, and slide rails and component shelves can be attached to the punched holes of the mounting angles.
Rittal
Tel: 01709 704000
Enq No 513

PC Card reader on PCI bus

Elan Digital Systems has introduced PC Card (PCMCIA) card reader/writer for PCI Bus applications which offers support under Windows95/98 and NT for all PC Cards, including those requiring interrupt control such as ATA, modem, LAN and DAQ. Model P128 allows a single PCI Bus motherboard slot, with two rear expansion slots, to be used for 2 PC Cards of any Type I, II and III simultaneously. This means that two large ATA cards may be inserted at the same time. A software driver is supplied with each unit and full Plug-n-Play features are available.
Elan Digital Systems
Tel: 01489 579799
Enq No 512

RF amp for CDMA mobiles

Hewlett-Packard has announced a new GaAs PHEMT (pseudomorphic high electron mobility transistor) RFIC amplifier in a four-lead SOT-343 package. It is aimed at CDMA handsets for mobile phone networks operating at 900 and 1800MHz. The MGA-72543 is the first release of the CDMA dual-band chipset. HP intends to introduce the remainder of the chipset - including power amplifier, upconverter, downconverter, modulator and demodulator – later this year. The chipset, which is fabricated with silicon bipolar and GaAs PHEMT technology, is designed for use with existing digital baseband ICs, filters and duplexers.
Hewlett-Packard
Tel: 00 49 0 64 41 92 4646
Enq No 514

Miniature SM oscillators

A series of low-profile surface mountable digitally temperature compensated, voltage controlled crystal oscillators has been introduced by C-MAC Frequency Products. Intended for personal digital cellular, CDMA, GSM and other mobile communications devices, the IQDTCVCXO-91 range offers frequency stability to ±1.0ppm

over a temperature range from –30°C to +80°C. In addition to digital temperature compensation, the device incorporates an AFC function that allows the output frequency to be trimmed by at least ±5ppm through a voltage input of 1.5V±1.0V. Both functions are integrated within an SM ceramic package measuring 7 x 9 x 2mm. Frequencies offered include 13MHz and 19.2MHz. Ageing is specified to within ±1ppm for the first year (at 25°C). Designed to provide a 1V pk-pk output into a 10kΩ/10pF load, the devices operate from a 3V supply.
C-MAC Frequency Products
Tel: 01460 74433
Enq No 515

SC1105/6 PWM controller

Thame Components has announced the SC1105 and SC1106 from Semtech. The SC1105/6 is PWM controller designed for advanced graphics port (AGP) power supply applications. The device switches between two different voltages when

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the TYPEDET# is toggled. The SC1105 is set to a fixed 1.5V while the SC1106 regulates to a voltage programmed by the user. The SC1105/6 provides a more efficient solution than the conventional linear regulator, claims the supplier. For a typical AGP circuit, the current requirement is 2A, which, for a linear solution implies undesirable heat sinking requirements. The SC1105/6, a switching solution, does not require heat sinking for the MOSFET or the rectifying diode.
Thame Components
Tel: 01844 261188
Enq No 517

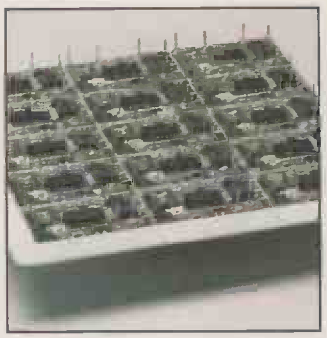
Panel meters

Made by Sifam, Select Din sized digital panel meters are available for voltage-current, process, counting-

totalising, rate-frequency, temperature, and strain gauge measurements. They have up to six digits on the LED displays. The case meets IP66 and Nema Four standards. For security, the setup can be restricted using password protection and there is a removable front bezel option to prevent unauthorised button access. Features include menu driven configuration, programmable smoothing and averaging features, display scaling, and remote control access via computer.
Sifam
Tel: 01803 407710
Enq No 518

Thick-film conductors

Emca-Remex has introduced silver Multisyst conductors for automotive, telecoms and RF circuitry. There is a choice of leach resistance, green strength and thermal-aged adhesion. Multisyst 2075D thick film is a cadmium-free, mixed-bonded, 6:1 AgPd, screen-printable conductor material for automotive circuit applications requiring heavy aluminum wire bonding. It incorporates powder and vehicle technology, providing thermal-aged and temperature-cycled adhesion at 175°C on alumina and the 7001D and 7015D dielectrics. Recommended solder is 10Sn/88Pb/2Ag. Other solders with a solidus above 250°C can also be used. Applications include automotive and telecoms hybrids. The 2060D is a mixed-bonded, Ag:Pd conductor for plug-hole applications in laser-drilled alumina substrates. It provides



adhesion to laser-drilled holes in 96 per cent alumina. It reduces processing steps by letting ceramic through-holes be completely filled and fired only once at 850°C.
Emca-Remex Products
Tel: 001 908 685 5148
Enq No 519

Comms test set

Racal has launched a portable radio communications test set for GSM mobile repair and maintenance applications. The 6104 is claimed to let a typical GSM phone be checked in under 30 seconds. For GSM 900, 1800 and 1900 radios, the tester

provides manual and automatic operation and has an IEEE488 remote control capability. It can run either predefined or user-defined test sequences. Test results can be stored in a database via the RS232 or IEEE 488 port and two PC card memory ports. This also provides a way to download operating code and custom test sequences into the instrument.
Racal Instruments
Tel: 01344 388000
Enq No 520

Single-in-line resistor network

An SMT single-in-line network from Siemens can be positioned vertically on the circuit board by automatic placement machines. The network makes two fusing mechanisms possible – mechanical clip-in fuses and PTC. It allows tailoring of the circuit protection area and is suitable for protecting against fault currents, overvoltages, lightning strikes, mains disturbances and switching problems. The heat conductivity of the substrate materials makes it possible to have different current strengths in different power ranges. The network can be assembled fully automatically using either grippers or vacuum nozzles and can be soldered in a reflow oven. The variable tripping characteristics provide protection for currents up to 350mA and mains disturbances up to 600W. It is also triggered by a rise in temperature to 300°C. Resistance drift for a surge voltage is 0.1 per cent.
Siemens
Tel: 00 49 89 636 53849
Enq No 516

Plastic circular connector

A plastic circular connector from Channel Electric uses rapid push-pull locking. The 544 connector works by sliding the locking collar forward on

the two-part body, pulling out the socket to open, inserting crimped or soldered terminations into the plug, mating the two components, and reversing the collar for a secure lock of plug and socket.
Channel Electric Equipment
Tel: 01635 864866
Enq No 521

Sensor stand

The UZZ2 universal sensor-mounting stand is adjustable in height and angle when used to mount photoelectric and fibre-optic sensors. Available in four assembly sets – basic, lateral arm, reflector and fibre – the stand can be combined with other Matsushita mounting sets, letting two or more sensors be used on one assembly. The stand can be mounted in horizontal and vertical planes and is for use on conveyors and

production machinery.
Channel Electric Equipment
Tel: 01635 864866
Enq No 523

Ceramic chip caps

TDK has announced Ni electrode multilayer ceramic chip capacitors that meet the COG temperature characteristic standard. They are available from 100 to 10,000pF. Temperature coefficient with capacitance change is 0±30ppm/°C from –55 to +125°C. Provided in a surface-mount package, the devices are suitable for flow, reflow and high temperature soldering.
TDK
Tel: 00 49 0 211 90 770
Enq No 524

Noise generator

The ANG broadband noise generator from Atlantic is for the broadcast, wireless, satcom, radar and avionics industries. For laboratory and field operation, it generates white Gaussian noise over multi-octave bands at frequencies from 10Hz to 18GHz. Noise output can be varied over a 111dB range in 0.1dB steps by front panel step attenuators with the options of continuously variable control or remote digital control. The ANG6109 version provides typically +10dBm output power, equivalent to –80dBm/Hz noise density, over the 100Hz to 1GHz band with flatness better than +2dB into a 50Ω load. The ANG8109 has 1W, +30dBm output and – 55dBm/Hz noise density from 1 to 300MHz. Fundamentally generated noise output is available up to 18GHz but, with up-conversion, this can be extended to 40GHz. Options include a signal combiner, which lets the user add the desired signal to the generated noise in making receiver sensitivity tests, and output filtering. The instrument measures 370 by 110 by 300mm and is also available in a 2U x19in, rack mounted version. Standard input power is 80 to 240V, 50 to 60Hz but operation from a DC supply can be specified. ■
Atlantic Microwave
Tel: 01376 550220
Enq No 525

8-bit microcontroller

Fairchild has announced a 2kbyte reprogrammable EEPROM version of its arithmetic controller engine (ACE) microcontroller with dedicated functions to improve central processor efficiency by offloading some coding operations. The ACE1202 is for battery powered applications including automotive security systems, home security systems and portable devices. As well as the EEPROM for code storage, it has 64byte of parameter storage EEPROM and 64byte SRAM on chip. It has a 16-bit multifunction timer with a dedicated difference capture function for independently capturing pulse widths to a resolution of 1μs. EEPROM data is memory mapped to the rest of the controller, eliminating the need to write instructions through a protocol, typically I²C. Features include idle timer with watchdog and programmable low battery detection.
Fairchild Semiconductor
Tel: 01793 856831
Enq No 522

Cool solutions

Alternatives to thermally-conductive grease continue to be developed to overcome not only thermal management issues but also production issues. James Stratford gives an overview of the the options available from thermal product specialist Bergquist.

James is Sales Manager with The Bergquist Company

Think of how to dissipate heat in an electronic circuit and most of us think of heat sinks and fans – strap a large lump of aluminium to the hot component and blow air over it. This remains a popular method of dissipating heat – but it's certainly not the only method.

Although there are innovations in electronics to reduce the wasted energy from active components, thermal management is an essential part of any modern circuit design. Moving the prototype from the R&D department and into production requires careful thought about how to build a product reliably and in volume.

As the demands for size and cost reduction increase, so traditional technologies for cooling become less plausible. Bulky heat sinks can also make up a noticeable cost within a product. Exacerbating this are the costs and time delays associated with manual heat sink assembly following PCB construction.

Production issues are best overcome by implementing good design practices to begin with. It is essential that engineers appreciate the thermal considerations of power devices as well as their electronic parameters. In this way engineers can explore complementary technologies to fans and extruded aluminium.

When is a fan undesirable?

When size constraints are placed on a design, heat dissipation becomes more problematic. Laptop PCs, for example, have a number of hot components critical to the operation of the product, but limited space for dissipating heat.

Forced convection using miniature fans is not desirable due to noise, excessive power consumption and reliability. As a result, hot components need to be laid out carefully on the PCB to ensure placement is close enough to an acceptable thermal transfer component such as a chassis.

Ensuring that a component as large and critical as a microprocessor remains in permanent contact with a suitable heatsink is difficult. Simple flexing of the board, or thermo-mechanical stress could result in the component being separated from the heatsink. The tolerances that have to be worked to are stringent.

To remedy this Bergquist has developed *Gap Pad*, a conformal material with high thermal conductivity. During final assembly, the material is sandwiched between the hot device and heatsink, its shape conforming to take up the air gap and ensuring a uniform thermal transfer.

There's more on this material in the panel.

Why not use the circuit board as a heat sink?

Why stop at creating a uniform thermal transfer from a component to a heat spreader? Why not mate the PCB to the heat spreader and use surface mount components?

The common PCB material FR4 is a poor thermal conductor. New package types such as D-PAK can exceed their safe operating temperature when mounted on FR4. The increasing use of surface mount components, driven by size constraints and simpler production techniques, leaves heat dissipation problems.

To accommodate the use of surface mount components, FR4 is typically manufactured to include thermal vias designed to carry the heat to the underside of the PCB where a heatsink can be added relatively easily. In many applications, this arrangement is adequate, but for more demanding thermal requirements, Bergquist has developed an insulated metal substrate material called *Thermal Clad*, **Fig. 1**.

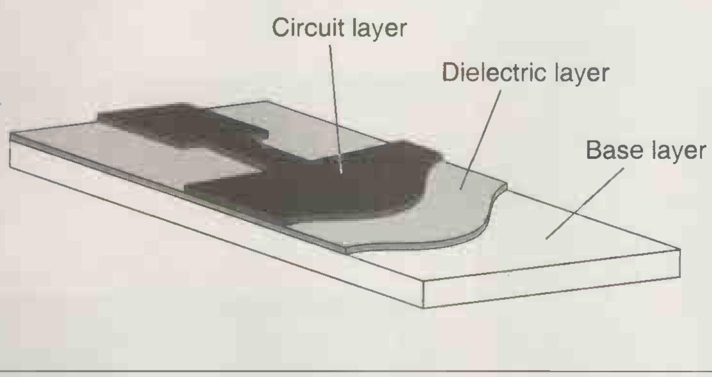


Fig. 1. Thermally conductive pcb material. The base is a metal such as aluminium or copper while the thermally conductive dielectric layer provides galavanic insulation. The circuit layer is 1, 2 or 3oz copper.

Continued on page 606

Void-filling cooling pads

In power inverter capacitors, constant charging and discharging generates heat that has to be removed in the interests of reliability.

Forced air cooling requires noisy, relatively costly fans which, being electromechanical parts, introduce an additional reliability issue. They also need power to drive them.

Convection cooling removes the need for a fan, but it involves large open areas in which air can flow. Both forced-air and convection require additional enclosure space.

Conduction is the most efficient method of dissipating heat, but large can capacitors are, by nature, irregular in shape. Consequently mechanically pressing the capacitor against the casing leaves air gaps. This results in poor thermal transfer.

Dimensions of can capacitors of the same type can vary by up to 2mm either way. Expansion and contraction of the capacitor during temperature cycling exacerbates this variation. This forces engineers to resort to fans or convection cooling rather than the preferred thermally-efficient option of conduction.

Grease compounds are sometimes used, but they are undesirable due to mess, migration and application. Grease can also present significant problems should the assembly require re-work.

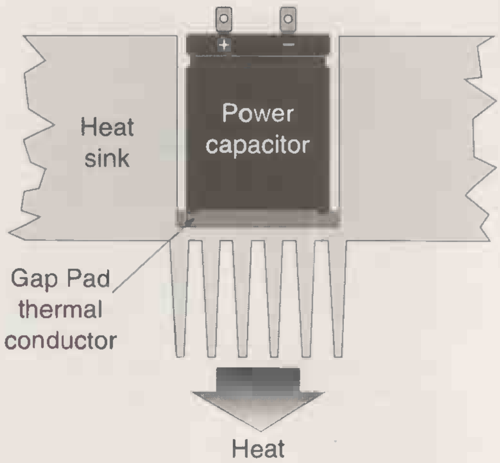
To remedy this, Bergquist has developed a conformal pad material with high thermal conductivity. Called *Gap Pads*, the range of products includes a number of alternative thermal performance versions as well as a pressure injectable material called *Gap Filler*. These products have a thermal conductivity of between 0.8W/mK and 3.0W/mK.

Gap Pad materials work by changing to conform to the required shape. This prevents the thermal interface from being broken, regardless of the fluctuations in physical shape or size.

Used with capacitors, *Gap Pad* conforms to the irregular shape of the can. Even with movement of the component relative to the case, *Gap Pad* conforms to maintain the thermal interface between the two.

Gap Pad is also available in standard, soft and gel-like modulus formats to accommodate low stress applications or where very irregular shapes require thermal interfacing.

Thicknesses of these products range from 0.5mm to 6.0mm in 0.5mm increments.



Gap Filler cures at room temperature to form a gel-like modulus that does not move from the interface, yet conforms with temperature cycling.

This material is made up of three layers – a base plate, typically aluminium or copper, a thermally efficient dielectric layer which electrically separates the base plate from the active circuit layer.

This copper circuit layer is etched using standard PCB fabrication practices and surface mount components assembled, soldered, cleaned and re-worked using common equipment. Heat from components passes directly through the dielectric layer and into the base plate with minimal interruption. Mounting a heatsink to the metal base plate completes the assembly forming a thermally efficient product.

In high-frequency applications such as DC/DC converters, switch mode power supplies or motor controllers, *Thermal Clad* exhibits unwanted capacitance, affecting the operation of the circuit. If these capacitive effects are important there is a two layer version of the product with a buried layer that can be used as a Faraday shield, as well as additional tracking as required. This shield significantly attenuates unwanted capacitance.

If you must use a fan...

In some instances the heatsink and the fan are the only practical solutions, but even here there are practical considerations for engineers to take into account.

Traditionally, grease and/or mica is used to sandwich a hot component to a heatsink. As a cheap electrical isolator, mica does the job, but it is brittle and easily damaged.

A reasonable heat transfer can be achieved with grease providing it is laid down in an even layer.

However, grease cannot be applied to partially assembled PCBs as it contaminates solder and cleaning is not possible because of the risk of washing out the thermal interface. This means it is only possible to apply grease at the final assembly stage after the PCB has been soldered and cleaned. Grease is also difficult to clean once applied, making re-work after testing difficult.

The problems associated with grease are not limited to production; reliability is also in question. During development, engineers ensure the application of grease is even. However, during production inconsistencies in greasing up components are inevitable.

Too much grease or an uneven application changes the thermal characteristics of the circuit, resulting in stressing of the component. An air hole in the grease layer may result in a hot spot on the component, also causing stress. Alternative silicon based greases can also deteriorate over time; migrating silicon molecules dry the grease out and contaminate the assembly.

Alternatives to thermally conductive grease

Grease-replacement materials overcome these inconsistencies by forming a uniform layer of thermally conductive, electrically insulating material. They also mean less mess. One of Bergquist's grease alternatives is *Sil-Pad*. This range of high-performance dielectric films combines a tough carrier including fibreglass and silicone rubber as a binding material.

Silicone rubber has a low dielectric constant, high dielectric strength, good chemical resistance and high thermal stability making it good for conducting heat away from the component. Supplied as a sheet, or pre-cut to suit common package formats and custom shapes, *Sil Pad* comes in a number of formats to suit the application, including a version with an embedded copper layer for shielding purposes.

These materials exhibit 'cold flow', which excludes air from the interface as it heats and conforms to the mating surfaces. The films achieve thermal resistance of down to 0.2°C/Watt and a breakdown voltage of typically 6kV.

The uniformity and stability of the material allow the product to be reliably manufactured to the same standards time after time. Re-work of faulty products is far simpler with such materials as the pads are simple to remove and replace.

And for high-volume applications

Greasing up components during production also introduces a delay and a break in the manufacturing process. Typically, this is a manual job and so inevitably impacts on production costs – especially as production volumes increase. Even with dielectric films such as *Sil-Pad*, a manual process is usually required.

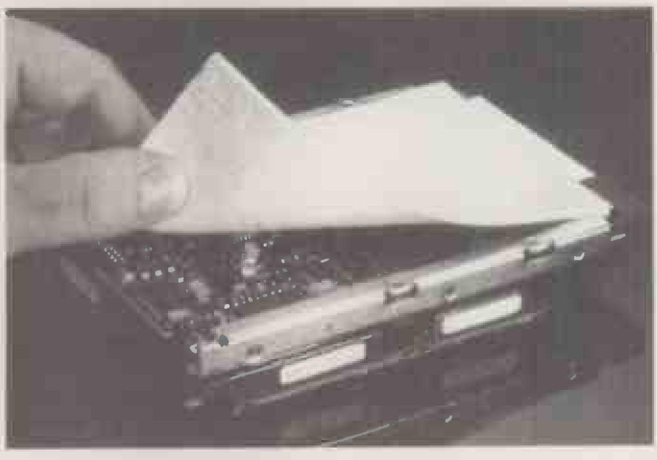
For such high-volume applications, Bergquist has developed *Softface*. Supplied as a film, this interface material is transferred to the desired surface using commercial hot stamping equipment.

To apply the material, the heatsink or case is pre-heated. Next, the interface material is coated on in any pattern as required. The heatsink can then be supplied to the production line for final assembly.

The material softens at 43°C to form a uniform thermally conductive layer. Its thixotropic characteristics prevent the material migrating out of the interface, and the material can be reused after re-work should it be required.

Unlike grease, *Softface* is also able to withstand solder baths and cleaning equipment, allowing partial or full assembly without the danger of contamination or wash out of the thermal layer. The combined benefits make the material an ideal grease replacement for medium and high volume manufacturing.

Ultimately, the thermal management requirements of a product depend on the design requirements. With the number of techniques designers have at their disposal to overcome thermal management issues, they don't have to totally rely on traditional heatsink arrangements. ■



This recent addition to the Gap Pad range has increased thermal conductivity relative to its counterparts in the same range. Available in thicknesses from 0.02 to 0.25in, Gap Pad 1500 has a thermal conductivity of 1.3W/mK and operates in temperatures ranging from –60 to 200°C.

Free with this issue

Gap Pad – an innovation in thermal management

The thermally-conductive pad on the cover of this issue isolates up to 6kV and has an operating temperature range of –40 to 200°C. But its really useful property is that it is flexible.

Being capable of conforming to its mating surfaces, the Gap Pad is clean and easy to apply and needs no grease – just a light pressure to ensure integrity of the joint. As a bonus, the Gap Pad doesn't dry out like grease so any voids that might occur due to thermal cycling and the like are filled as they occur.

This sample, known as Gap Pad VOSoft, has a thickness of 1.5mm and a thermal performance of 0.8W/mK, depending on applied pressure. To use the pad, simply assure a slight pressure between the hot component, its cooling surface, and the Gap Pad between them. Other products in the range cover 0.8W/mK to 5W/mK in thicknesses from 0.5mm to 6.0mm in 0.5mm increments. There's also a liquid version that can be pumped into a void, where it sets, providing effective thermal conduction between irregular surfaces.

Design engineers needing more information should call Bergquist on 01908 263663.

Thermal management specialist Bergquist

The Bergquist Company is a privately held family owned business started by Carl Bergquist in the sixties. The company began distributing electronic components in the upper Midwest of America and soon after developed their first proprietary product, *Sil-Pad* thermally conductive insulators.

This very successful product line has evolved into the Thermal Products Group which provides solutions to engineers for controlling and managing heat in electronic assemblies and printed circuit board designs. The other two main divisions of The Bergquist Company are Bergquist Switch and Bergquist Distribution, which includes the power cord product line.

The Thermal Products Group's 84 000 square foot manufacturing facility is in Cannon Falls, Minnesota. Thermal Products has additional facilities in Germany and the United Kingdom. A new facility in Prescott, Wisconsin houses the Thermal Clad printed circuit board operations and the new Touch Screen manufacturing operations.

Bergquist products are used by many of the world's largest OEMs in a variety of industries including automotive, computer, military/aerospace and telecommunications to name a few.

Over 260 materials have been developed to fit many different applications for transferring heat in electronic assemblies. Many of the new materials developed are the result of customer requests for a specific material that can perform for their particular application.

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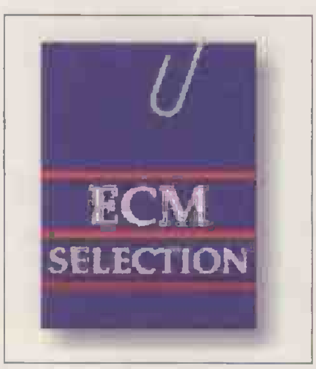

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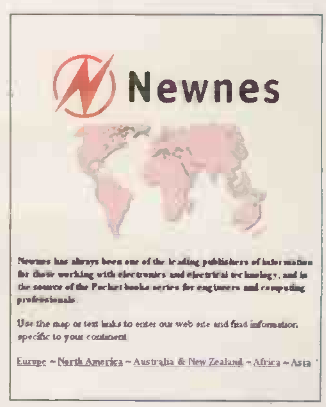

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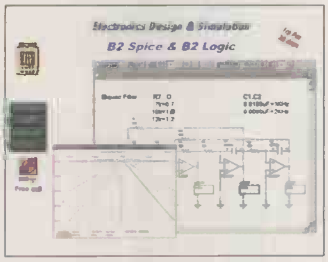

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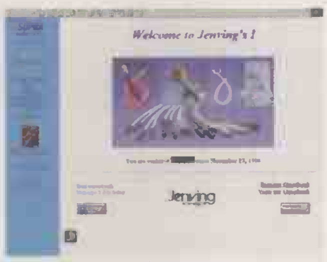

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Programming the eeprom HC11

Peter Topping's step-by-step guide for programming the LW radio-code clock presented last month demonstrates how easy it is to program and develop a modern microcontroller. The software is freeware and the hardware is simply an RS232 interface.

A suitable method of programming the *MC68HC811E2* is to use the PCbug11 software package.

PCbug11 is a free program that runs on a PC. It talks to a small piece of code embedded in the HC11, or to code downloaded into its RAM using serial communication. The compact – in the case of the LW clock, downloaded – code is interrupt driven and is referred to as the *talker*.

This combination facilitates a debugging environment that can be used to download application code and then debug it by allowing the code to be modified and executed, breakpoints to be set, RAM locations to be inspected etc.

I used PCbug11 to develop the LW clock application described in last month's issue, but the only feature discussed here is its ability to download code into the EEPROM of the *MC68HC811E2*. This can only be done once communication has been established between the PC and the *HC11*.

The RS232 chip shown in the circuit diagram is required to establish communication. The LW clock digital board can be used by simply adding a small board containing this extra circuit.

Only three wires, Tx, Rx and ground, connect the extra board which can use the *MC145407* shown, or a *MAX232*. Both these chips contain charge pumps so they do not need the additional ±12 volt power supplies normally required for an RS232 interface. Connection to the PC's COM1 port with a default baud rate of 9600 is assumed.

Peter is an MCU applications engineer at Motorola's East Kilbride plant.

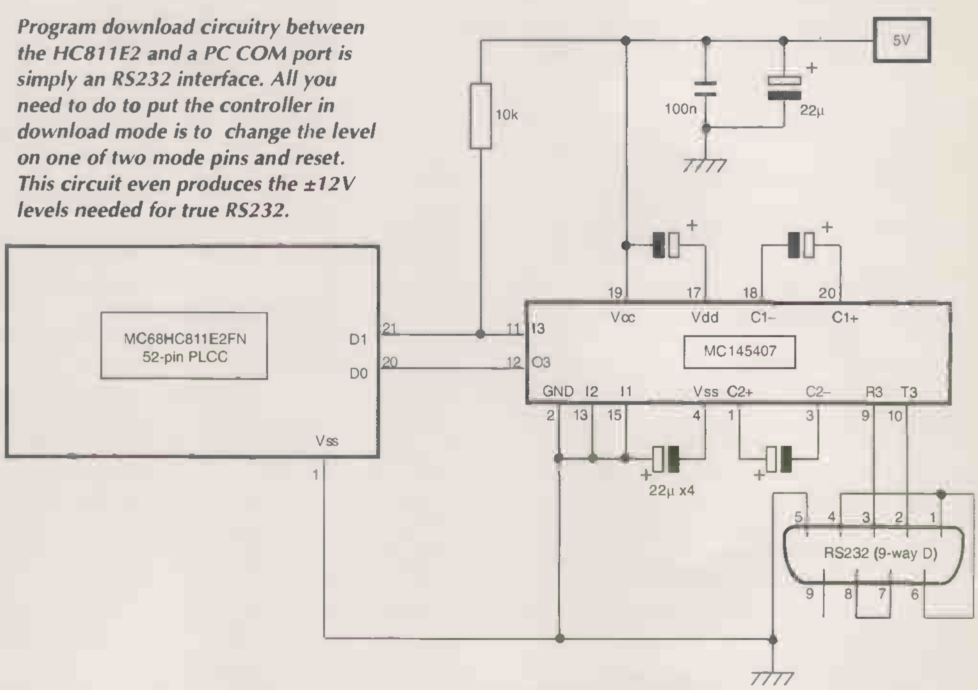
There are two modifications that must be made to the *HC11*'s circuit on the clock's digital board. These are the change of the crystal to 8MHz and the selection of a different mode.

To download simply short one pin

The circuit shown in last month's article uses single-chip mode by holding the MODA pin low and the MODB pin high. For PCbug11 the special bootstrap mode is used requiring both mode pins to be low.

As the MODB pin is held high with a pull-up resistor it can simply be shorted to ground to select the appropriate mode for PCbug11. With this done, a reset will force the *HC11* to enter it's special bootstrap mode. In this mode it runs pre-programmed ROM code that downloads data into on-chip RAM via the serial SCI interface. The interface is implemented via by pins D0 (20) and D1 (21).

When using PCbug11, the talker is first downloaded. Once downloaded, the talker can be used in conjunction with the PCbug11 program on the PC to download



Program download circuitry between the HC811E2 and a PC COM port is simply an RS232 interface. All you need to do to put the controller in download mode is to change the level on one of two mode pins and reset. This circuit even produces the ±12V levels needed for true RS232.

and debug the user's application code. This code can be loaded into RAM, EPROM or EEPROM. But for the purposes of the LW clock, the download is only into the *MC68HC811E2*'s 2Kbyte EEPROM.

The application has already been fully debugged so PCbug11's many other features – breakpoints, code modification, RAM inspection etc. – are not used.

To use PCbug11, I recommend that the entire contents of the floppy disk available from *Electronics World* is copied into a new DOS directory – say pcbug11 – on your hard disk. You should then run with this as the default directory. A real implementation of DOS is required. Emulated DOS running within Windows NT will probably not work correctly.

Application code download into EEPROM is in S-record format, as is the listing printed last month on page 453. An S-record is in ASCII and contains the following information.

The first two characters on each line are the record type – S0: header, S1: data, S9: terminator. The next two characters indicate, in hexadecimal form, the number of bytes of data that follow. The next four characters contain the 16-bit hexadecimal address of the first data byte and the data follows in hexadecimal, at two characters per byte, the last byte being a checksum. ■

In last month's LW clock article, the penultimate line in Table 4 should have read Y:3/5(1999)W:13. Apologies.

Programming and LW-clock software

The disk available from *Electronics World* contains the following PCbug11 files:

pcbug11.exe	PCbug11 executable program
codes.p11	Mnemonic tables for PCbug11
offsets.p11	Addressing mode offsets for mnemonic tables
pcbug11.hlp	PCbug11 help information
talk88.boo	Downloaded talker for the MC68HC811E2

As well as its executable file, PCbug11 requires the two .p11 files and the talker for the MC68HC811E2. The disk also includes a help file, which is accessed by typing "help *topic*" within PCbug11. For a list of commands simply type "help". This disk also contains these LW clock application files.

lwrd.s11	Source code (Introl assembler switch setting: -TI).
lwrd.lst	LW clock listing output from assembler.
lwrd.doc	LW clock listing (Word format).
lwrd.bat	Batch file for use with Introl relocatable linker.
lwrd.ld	Address locations for use with linker.
lwrd.0	Object code in S-record format.

The lwrd.s11, lwrd.bat and lwrd.ld files are only relevant if modification and reassembly are required. The assembly directives included are for the Introl assembler and may need modification for other assemblers. For programming the LW clock application described in last month's magazine only the S-record object code file (lwrd.0) is used.

The object-code listing shown in last month's article can be obtained as a text attachment to a text file by e-mailing jackie.lowe@rbi.co.uk. The annotated source code can be obtained on disk for £10 to cover copying, administration and postage. Send a postal order or cheque payable to Reed Business Publishing Group to Precise Clock, Jackie Lowe, Electronics World, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.

Please note that the source-code listing will not be e-mailed due to network bandwidth limitations.

Procedure for using PCbug11 to download the LW clock program:

1. Connect the serial board to the PC and to the LW clock digital board. Power up in special bootstrap mode (8MHz crystal, both mode pins low).
2. On PC enter "pcbug11 -88"
3. If communication is correctly established – i.e. no error message – then continue. If communication between the PC and the HC11 is not established this will be intimated by an error message. If this occurs then the problem should be rectified before attempting to continue. The most common problem is incorrect connection of the Tx (D1) and Rx (D0) lines or the lack of a pull-up on the Tx line. This pull-up is missing from the circuit in some editions of the PCbug11 user's manual. Communications can be checked by typing ctrl R. This will get a response of "communications synchronised" or display an error message. After a communications error a retry should be carried out by exiting PCbug11 (quit y), resetting the HC11 and restarting at step 2. This is required as the talker is downloaded into RAM and if this step has failed then typing ctrl R will not fix the problem as it doesn't retry the download.
4. Once communication is functioning correctly enter the following PCbug11 commands :

control base hex	(allow subsequent commands to specify addresses and data in hex)
eeprom f800 ffff	(specify address range to be written as EEPROM, i.e. not RAM)
ms 1035 00	(enable EEPROM writing)
loads \pcbug11\lwrd.0	(download S-record code into EEPROM, this takes 2 minutes)
verf \pcbug11\lwrd.0	(verify that EEPROM contains the correct code)
quit y	(quit PCbug11)

5. If everything has occurred correctly the EEPROM will now contain the LW clock code. To convert the hardware back to run the clock: power down, remove the additional board containing the serial interface to the PC, replace the crystal with the application frequency of 2MHz, remove the short to ground on the MODB pin and connect the LW clock analogue board.
6. On power-up the LW reference clock should now be operational.

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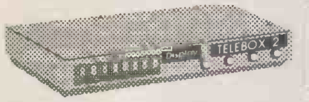

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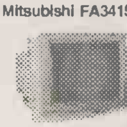

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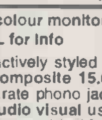

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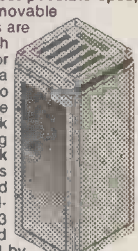

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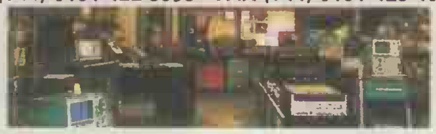

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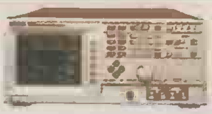

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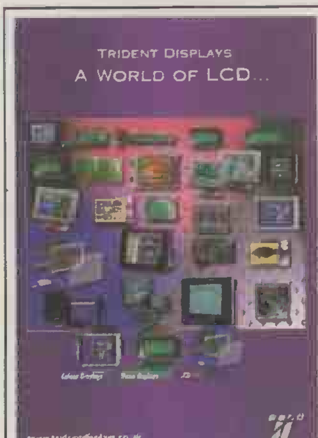

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