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ELECTRONICS WORLD

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Measure EHT
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Extra-range PWM

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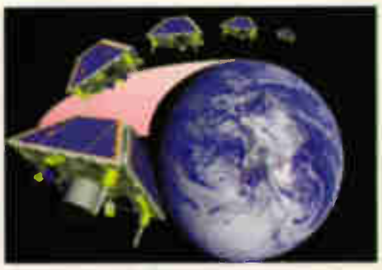
GPS in mobiles, cars and even watches is on the increase and it isn't stopping there, reports Melanie Reynolds. See page 765.

It seems logical to use digital techniques to store and analyse medical X-rays. Although some hospitals are equipped to do so, the take-up of the new technology is slow. Find out why on page 766.



Surrey Satellite Technologies receives the lion's share of DTI space funding. This and more news on page 756.

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REED BUSINESS INFORMATION

Baffling the speaker buyer

"Do loudspeaker makers design for performance or profit?" asks John Watkinson

It is an unfortunate fact that the great majority of loudspeakers are not designed for accurate sound reproduction. Instead they are designed to make a profit for the manufacturer and the retailer. This is what happens when a technology matures.

Figure 1 shows that when a technology is young, the industry is rich in inventors and investors and the rate of progress is rapid. Buoyant sales of a new technology fund further development.

However as a technology matures, the inventors have less to do and generally move on to do something more challenging. This leaves the industry run by management who only have a slight grasp of the technology. As they can't improve the technology, all they can do is make it cheaper.

In reality they are forced to make it cheaper because all of the startup companies that began the technology have grown to a size where the market isn't big enough to support all of them unless the price comes down.

This stage is called commoditisation. Fierce competition in an oversupplied market drives the price down further, forcing manufacturers to switch production to countries where labour costs are lower. In this climate, the management isn't about to take any risks and so the product design stultifies.

Inevitably new materials and technologies come along, which could improve the technology. Pure research throws up new and better understanding of the hearing mechanism, which could drive better designs.

Unfortunately though, the inventors who could put these ideas and technologies to use in the loudspeaker are long gone and the remaining finance-focussed management doesn't understand the old technology – let alone the new.

CAD reduces costs, but does little for performance

The result is that nothing happens. The new technologies are ignored and the new CAD system only lowers the cost of the design process without improving the product in any way.

In loudspeakers, the results are really quite depressing. Traditionally the hi-fi dealer has expected to make a significant mark-up on passive loudspeakers. This means that the price the manufacturer has to hit is significantly lower than the retail price.

Now the manufacturer has to abide by the same economics as everyone else. This suggests that the factory door price should be a certain multiple of the parts cost. Applying a simple calculation will show that even in a pair of hi-fi speakers retailing for several hundred pounds, the drive units in those speakers must have cost no more than about ten pounds each.

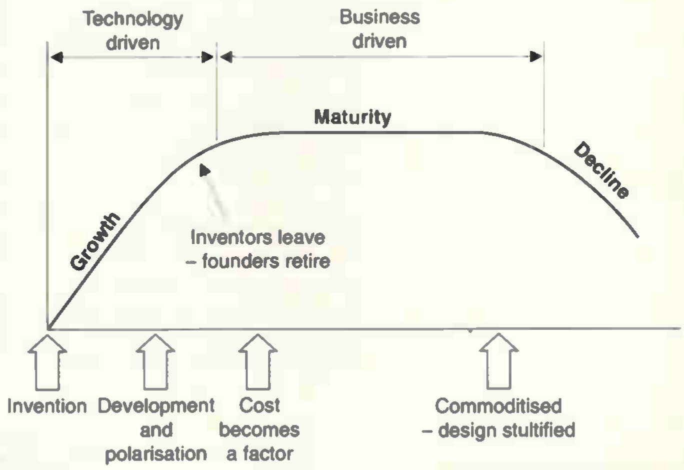

Fig. 1. The life of a technology.

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Were your speaker cables dearer than your speaker's drive units?

With drive units built down to a cost like this, the performance is not going to be special.

It is particularly saddening that the unwitting purchaser of such speakers is then likely to be gulled into buying a set of audiophile speaker cables which will probably cost more than the drive units they connect to. It doesn't need much thought to see that the overall result is that purchasing the cables simply gives the user even worse value for money.

Loudspeaker drive units are associated with electronics, but they are not themselves electronic. Every one knows that the cost of performing a certain electronic process continues to fall. It's formally known as Moore's Law and it applies particularly to computers and digital audio devices. By association, most people think it applies to loudspeaker drive units. Unfortunately they are wrong. **Figure 2** shows the consequences.

Some time ago when electronic equipment was expensive, a balanced audio system required roughly the same amount to be spent on the loudspeakers as on the amplifier. Today, the cost of the electronics has fallen according to Moore's Law. But the cost of the speaker hasn't. As a result to get a balanced system it is now necessary to spend about ten times as much on the speakers as on the amplifier. Unfortunately this isn't widely known.

The result is that the majority of today's audio systems are imbalanced. Generally the amplifiers are producing signals that are far too good for the drive units. In fact a good definition of an audiophile is someone who spends freely to eradicate an insignificant problem while neglecting a major drawback.

It would be very easy to mutter, "*caveat emptor*" and shrug one's shoulders. After all, if people are foolish enough to buy products on this basis, the industry should be allowed to make them.

The trouble is that the mediocre speaker is no longer confined to the consumer. Moore's Law has ensured that the cost of most of the technical equipment needed to make a recording studio has fallen. Large recording studios have fallen on hard times because the cost of equipping a personal studio is within the reach of the enthusiast. Excepting, of course, the loudspeaker. As a result the home studio owner generally settles for loudspeakers that are inferior to the rest of the equipment.

In professional audio, the quality of the sound produced cannot exceed the quality of the monitoring loudspeakers. If the speakers aren't accurate enough to reveal a defect, the defect goes unnoticed – however good the listener.

The equipment manufacturers have the same problem. How is it possible to test a piece of high-grade audio equipment without high-grade speakers?

What does 'inaudible' mean?

The consequences are all around us. The widespread use of digital audio compression or bit-rate reduction is largely on the assumption that it doesn't affect the sound quality. Unfortunately, testing audio compression with poor loudspeakers is an invalid test. Where the information capacity of the speaker is the limiting factor, the use of compression won't make the sound any worse.

The listener will wrongly assume that the compression is inaudible and a source of impairment will enter service. With impaired speakers, the compressed result will sound indistinguishable from a compact disc and the result will be dubbed 'CD quality'.

This term is meaningless. On precision loudspeakers, rather than commoditised speakers, all of the popular compression formats produce audible artifacts in comparison to direct CD playback. It is easy to compare MiniDisc and CD because the same titles are released in both formats, which have come from the same digital master.

If an obvious loss of quality isn't apparent when playing the MiniDisc version, the loudspeakers aren't accurate enough. ■

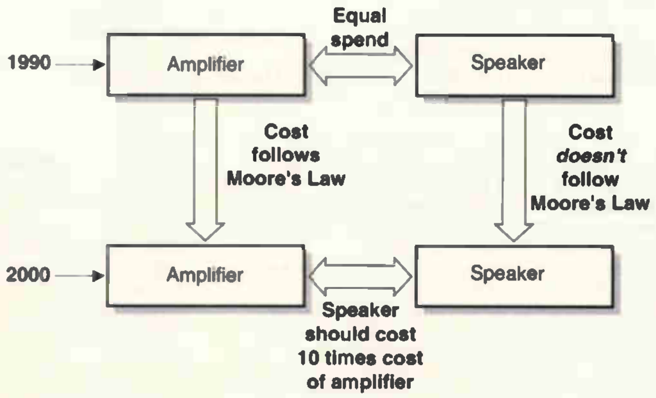


Fig. 2. Balanced system is a moving target. The cost of an amplifier has fallen, in accordance with Moore's Law, but the cost of speakers hasn't.

CIRCLE NO. 106 ON REPLY CARD

Spectrum auctions could hinder wireless services roll-outs

Industry body the Federation of the Electronics Industry (FEI) fears that a proliferation of radio spectrum auctions as a result of a recent report to the government will impact the roll-out of new wireless services.

The FEI's concern centres on the roll-out of services in the 2.4GHz band which includes fixed wireless access services and even the deployment of Bluetooth-enabled equipment.

This follows the publication of a document on the 2.4GHz band commissioned by the Spectrum Management Advisory Group (SMAG), the body that advises the government on radio spectrum issues.

"The treasury will probably continue to encourage spectrum auctions. We, however, don't believe that auctions are the most effective way of making use of spectrum," said Simon Wilson, director responsible for radio issues at FEI.

Expected at the end of August, the SMAG 2.4GHz report, which has been carried out by spectrum engineering consultancy Aegis Systems, is specifically on licensing "policy" according to Aegis MD Paul Hansell.

This is significant because it contrasts with last year's technical study by Aegis on technical issues such as interference at 2.4GHz.

Hansell would not comment any further on the new document but the content of other recent studies such as the 'Future of Convergence and Spectrum Management' (produced for the Radiocommunications Agency), suggest that the FEI's concerns may be well founded.

This latest report, which has particularly worried the FEI, seems to indicate the use of frequency auctions to actually determine how the spectrum will be used in the absence of any official policy. For example, the report recommends the use of spectrum auctions, "where the best configuration for spectrum is unclear".

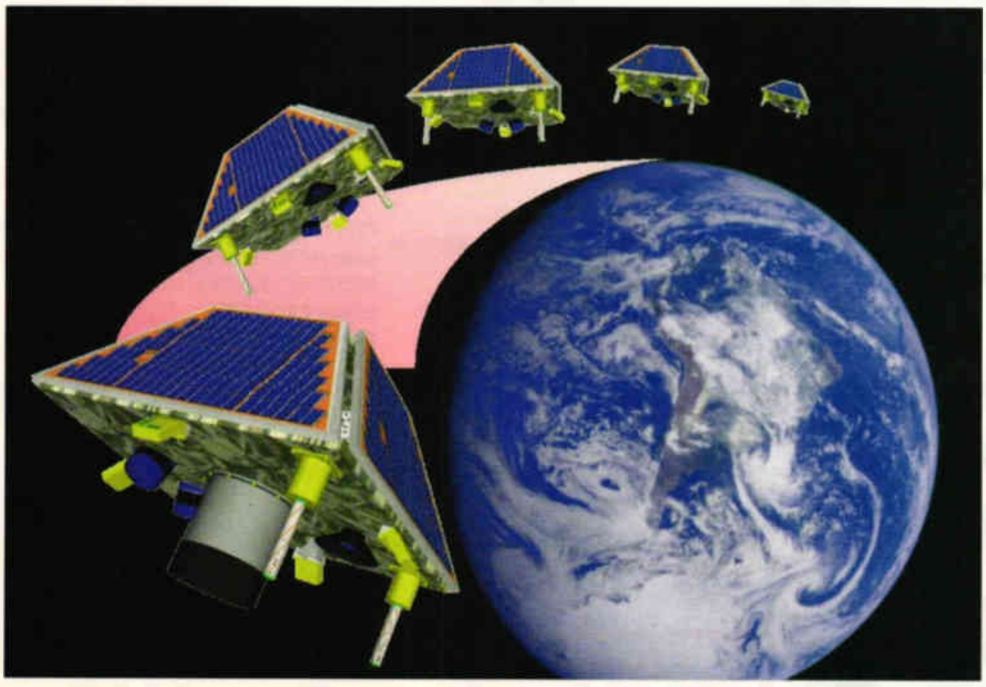
As a result auctions could impact many of the newer wireless applications. These include Bluetooth, 802.11 wireless LAN, HomeRF and public radio fixed access services, currently operating in the 2.4GHz band.

Bulk of space funding goes to Surrey Satellites

Surrey Satellite Technology (SSTL) is to get the lion's share of £15m support the government has announced for the UK's space programme.

The Guildford-based firm will use the cash, which has been made available by the British National Space Centre, to develop its mini and micro satellites. "We have got £11m – the first funding that we have ever had," said an SSTL spokeswoman.

SSTL is 95 per cent owned by the University of Surrey and has previously been self-funding, ploughing profits back into the University.



SSTL has to find £11m to match the DTI grant, which should not prove too difficult. The organisation has, according to the spokeswoman, a £36m order book.

The money is to be split between three projects including TOPSAT, a Defence Evaluation and Research Agency (DERA) led project to provide low-cost high-resolution images of the ground. SSTL will be providing the 'bus' – the chassis and flight systems – with the cameras coming from DERA.

SSTL is also prime contractor for the GEMINI project to build a geostationary satellite, the first attempted by SSTL, weighing under 400kg that will provide telecommunications in Nigeria.

The third project called Disaster Monitoring Constellation (DMC) is described as a new concept in satellite management. DMC will involve satellite to satellite message passing, groups of five to seven satellites will co-operate (pictured).

I Internet cattle auctions could be a feature of farming in the future, said science minister Lord Sainsbury.

He was announcing £2m of funding for the S@TCOM programme, which aims to help UK companies exploit satellite comms and navigation.

Steve Bush

Ericsson redirects its mobile phone business

Ericsson is carrying out a strategic review of its mobile phone handset business. Disappointing results could see the Swedish firm selling other manufacturers' hand sets under its own name for the first time.

"To restore profitability, a strategic review is ongoing and a number of firm actions have been initiated," said a company statement. One of the things being considered by Ericsson seems to be the badging up of lower cost hand sets.

The company said it would, "address the entry-level segment with a special purpose organisation and externally sourced products, concentrate the product portfolio, streamline manufacturing, and increase R&D for mobile Internet to expand leadership in GPRS and 3G."

Externally sourcing products would be a significant move for Ericsson, which is one of the world's top three mobile-phone technology firms.

Contrasting with the disappointing handset business, Ericsson benefited from strong business in building mobile phone networks, which helped first half pre-tax profits to jump by more than 300 per cent.

"These results confirm Ericsson's leading position in mobile infrastructure," said Kurt Hellström, president of Ericsson. "So far, we have captured over 50 per cent of both the GPRS and the 3G market – solid proof of our unmatched position as market leader."

But investors see worrying signs in the mobile-phone handset business which experienced a profit cut in the same period.

Sales of mobile phones increased by 40 per cent, but Ericsson said the consumer products business unit – as handset are called – went into loss, "due to component shortage from a key supplier and an unfavourable product mix."

Pressure is on the firm to restructure its handset business, particularly to take advantage of the booming low-cost consumer end of the mobile phone market.

Long wait ahead for mobile exposure data

UK mobile phone users will not be able to access information on the amount of energy absorbed by the human body while using a mobile phone until next year at the earliest.

This is despite a recommendation from the UK government's mobile phone health review earlier this year that such information should be made available to users.

Absorption of energy in the body is considered a crucial issue in assessing exposure levels. The amount of energy absorbed is given by the SAR (specific energy absorption rate) of a phone but the problem at the moment is there is no commonly agreed method of measuring it. European standards body CENELEC is working on this but the earliest date a standard is expected to be ready is the end of this year.

"Some companies measure them [SAR] in different ways," said a Department of Health spokesman. "It's important if the consumer is to have information on this that everybody's measurements are done in exactly the same way."

A leaflet stating the recommendations of the report is currently being finalised and is expected to be distributed this summer.

In March 1999 the European Parliament recommended mobile phones should carry labels giving information on the fields generated as a function of distance. If accepted the recommendations would be implemented in January 2001.

Leading mobile phone suppliers, including Nokia, Ericsson and Motorola, are working together to agree a common set of rules for including SAR figures on their phones. However, it is unlikely that any mobile phones in the UK will actually include such radiation information until the official European SAR standard is finally agreed early next year.

This information was recommended to be made available to consumers in the *Report on Mobile Phones and Health* by the Independent Expert Group on Mobile Phones in May this year.

The telecoms US authorities have already taken a lead on the SAR issue and introduced a requirement for mobile phone makers to disclose SAR figures for all new products submitted for product certification.

Melanie Reynolds

Schools warning 'low key'

The significance of a letter to schools warning about the possible dangers to the health of children from mobile phone use has been played down by the Department of Education.

The letter sent by the Department of Education to head teachers states the findings of the recently published Independent Expert Group on Mobile Phones, known as the Stewart report, into mobile phone health risks.

"All we've done is make sure schools are more aware of the Stewart report," said an Education spokesman.

"We're not saying mobile phones are dangerous or that they shouldn't be used. We're letting schools decide their own policy," he continued.

World chip Top Ten see sales jump 37 per cent

STMicroelectronics has jumped two places in the world rankings in the first six months of the year according to Arizona analysts IC Insights.

ST is now the seventh largest semiconductor company in the world. The other big mover in the rankings is South Korean firm Hyundai which, by its acquisition of LG Semicon, jumped six places in the rankings to join the top ten for the first time at No.8 having grown 232 per cent.

Korea's biggest semiconductor manufacturer Samsung has climbed a place from No 6 to No 5, almost doubling its sales on the back of a stabilised DRAM price and set to hit $9bn for the year.

Although Philips Semiconductors grew sales by 36 per cent, it dropped out of the top ten.

STMicroelectronics' growth came from its flash memory, analogue, and mixed-signal products. ST is spending $3bn on capital additions this year which should keep its growth continue strongly into next year and 2002.

Overall, sales of the top ten companies rose 37 per cent in the first half.

IC Insights expects a 17 per cent increase in the worldwide semiconductor market in the second half of 2000 compared to the first half of the year, with the full-year 2000 market forecast to be 35 per cent larger than in 1999.

"Strong second half sales of PCs, cellular phones, and consumer systems are expected to give the semiconductor producers their best year since 1995," said IC Insights.

Electronics skills dearth addressed

Electronics companies suffering skills shortages have a been given a fast-track route to the Government's ear to voice concerns over training.

At the request of Malcolm Wicks, Minister for Lifelong Learning at the Department of Education and Employment, a new industry working group has been set up specifically to advise the minister on training and skills issues within the engineering industry.

"I think he genuinely wants advice," said Dr Michael Sanderson, chief executive of the Engineers and Marine Training Authority (EMTA) which is home to new working group.

To be called the Minister's Working Group for Skills, Education and Training Policy Advice, the group will represent all of the engineering community. But, according to Sanderson, electronics will be a major concern. "I will be supporting growing industries. Electronics is right at the top of this list. I am less interested in declining industry."

Speaking about the shortage of skilled staff in the electronics industry, Sanderson said: "We [the EMTA] have been worried for a long time. The top end, high graduate level, is more or less adequately served.

"It is the senior technician level, with NVQ level 3 or 4, where the problem is. We are working with the Federation of the Electronics Industry to encourage employers to invest in work-based training."

The EMTA is home to the National Training Organisation (NTO) Group for all engineers except civil engineers.

STM moves into FPGAs

Ten people are standing around an elephant asking: "What is this?" The person standing by the tail says: 'This is a rope'. One standing by a leg, says: 'This is a pillar'. One by the body says: 'This is a wall'. The one by the trunk says: 'This is a snake'. One by a tusk says: ' This is a dagger'. Then they sit down together to draw a picture of an elephant, and they get a wall with a rope and a snake hanging off it standing on a pillar."

The story is told by Rahul Sud, something of a legend in the semiconductor industry, who is master-minding STMicroelectronics' move into FPGAs. The story is designed to illustrate why every big company has failed in FPGA, but why ST will succeed.

Sud came to prominence in the 1980s as the designer of the only commercially successful product Inmos ever had when it was independent – a 16K NMOS SRAM which became an industry standard and made hundreds of millions of dollars.

He then achieved fame as the founder and first president of Lattice Semiconductor, the successful programmable logic house. For some years he's been in charge of new projects at ST.

The elephant story signifies that, in FPGA, the whole is greater than the sum of its parts. "The big companies who tried it failed because they didn't see the whole picture", Sud told *EW*, "they totally underestimated the scale and complexity of such a project."

The fallers at the FPGA fence are a roll-call of the industry's biggest and best: Intel, Texas Instruments, Motorola, IBM, Philips, Toshiba and AMD.

"Most managements who have tried to do FPGA wanted to do it quickly. They've said: 'Let's take a licence'," says Sud, "so they've taken a licence, and got a mask-set and a software tool-set and, and run wafers and got into the business and, after six months, along comes a software update and they're in trouble. Because they don't have a deep understanding. There's no such thing as a quick entry; no such thing as a quick understanding."

"There is a common thread why they all failed," says Sud, "they all grossly underestimated the software challenge. Primarily FPGA is a software challenge not a hardware challenge. To succeed in FPGA you need to set up on the model of an EDA company not a silicon company. You need to start with customers' software requirements and the third party interfaces needed, not by designing a piece of silicon."

He points out that Pilkington Microelectronics – whose design was licensed by Motorola and Toshiba – did things the wrong way round by first coming up with, "a beautiful silicon architecture" and then asking the software people to write software for it.

"To be successful you need a portfolio of pre-programmed, pre-packaged IP," says Sud, "because people don't have the design resources to do it."

Sud has, for some years, been building a formidable resource for ST - a large and growing group of software developers in India, his home country. Being able to straddle both Indian and Western culture he has devised a system of developing architectures in the West, for which the code is written in the East.

"Just try hiring 100 IC designers in the US or the UK. First of all there's the recruitment challenge, then there's cost. So I've decided to go to the UK for the architectural development because it's the nearest thing there is to Silicon Valley while being closer to home".

Sud is currently looking for architectural talent in the UK for the project.

How soon before there are any products? Sud is resolutely not going to underestimate the difficulties. "Realistically," he replies, "it's going to take two to three years before we can enter the market seriously." ■

Distortion ...the facts

Misconceptions on the subject of distortion abound. Ian Hickman looks at a selection of them, and presents some practical results illustrating the true state of affairs relating to this emotive subject. None of these results is new, but some may surprise you.

Hi-fi enthusiasts range from those who are highly competent technically, familiar with the detailed performance figures for their equipment, to those who scorn such technicalities and rely solely on their golden ears.

I belong to neither camp. Both seem to spend more time listening to their equipment, and its failings, real or imagined, than to the music itself.

My hi-fi equipment has varied over the years, from a Williamson style amplifier with a very nice Gardner's output transformer, through a variety of home-brew solid-state amplifiers, to some fairly run of the mill commercial stuff at the moment. But articles on the subject in this, and other, magazines continue to hold great interest for me.

It is a pity that the very interesting recent article[1] did not include some measured performance parameters. In this article, I present some of my recent thoughts on sound reproduction, and in particular to its impairments due to amplifier problems.

Seconds versus thirds – and others

There is continuing controversy over amplifier architectures. Some advocate class A as the only possible route to hi-fi. Others insist that push-pull class B amplifiers can be engineered to be perfectly satisfactory.

Yet others insist that semiconductors will never give an acceptable performance, and that valves are the only serious contenders. Some even insist that a large single-ended triode – or paralleled triodes – is the only answer.

The triode camp points out that such amplifiers are substantially characterised by second-harmonic distortion, which is less unacceptable than third. But, of course, while a single-ended valve amplifier may provide notably second harmonic distortion, if driven near to its limits, it must run out of available swing on both positive and negative peaks. No amplifier is inherently immune from third-harmonic distortion, simply by virtue of being single-ended.

Many who approve of push-pull amplifiers only give their blessing to class A designs, due to the harmonics associated with cross-over distortion. The argument goes that crossover distortion introduces high orders of harmonic and intermodulation distortion. This is much more objectionable than a little second, or even third order non-linearity.

I must confess that I thought so too until a while ago. It is easy to see how this serious misconception can arise, as the following makes clear.

The anatomy of cross-over distortion

To illustrate the point, I made up a low power model of a push-pull output stage with bad cross-over distortion, shown in **Fig. 1**. The two devices were selected for matched current gain, although the match was admit-

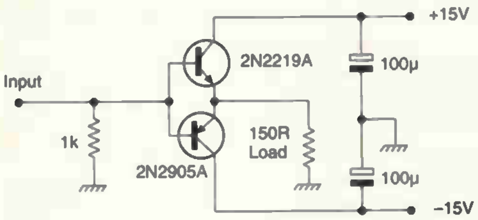


Fig 1. Circuit of a small low power push-pull output stage with bad cross-over distortion. Cross-over distortion appears at around the mid-point in the signal excursion, where one transistor takes over from the other.

tedly checked at a few milliamps only.

Figure 2 shows it handling an input at 440Hz, concert pitch A above middle C. The input is 12V pk-pk, while, due to the cross-over region's 'dead space', the output is only about 10.4V pk-pk.

Measured total harmonic distortion is 10%. **Figure 3** shows the residual signal output from the THD meter. This is just the distortion products, the fundamental having been suppressed entirely.

It seems obvious from the peaky nature of the residual that the distortion consists mainly of high-order harmonics, These are likely to be most deleterious to music reproduction, or indeed any signal.

But does the residual mainly comprise high-order harmonics? A THD meter does exactly what it says, and no more. It measures total *harmonic* distortion; it does not measure *total* distortion.

Looking again at Fig. 2, you can see that the amplitude of the fundamental has itself been reduced. One can only define distortion as any departure of the signal from what it should be. In the case of a nominally unity gain stage, such as Fig. 1, this means that any departure of the shape of the output signal from the input signal is – by definition – distortion. This was measured by probing both the input and output signals, and displaying trace A plus trace B, with trace B inverted, i.e. a differential measurement.

Figure 4 shows the result, together with the input signal. The smaller trace is the distortion, and it is simply a squarewave with sloping edges. It is in fact, the missing slice out of the middle of the input, due to the crossover 'dead space'.

Clearly, this distortion includes a component at the fundamental frequency. This does not appear in the residual output of the THD meter, due to the notch at the fundamental.

The spectrum of the distortion will be similar to that of a squarewave, except that due to the sloping edges, the higher frequency harmonic components will actually roll off rather faster than in a true squarewave.

Figure 5 shows just the harmonic components of the distortion, together with the output signal; the 'dead space' in the latter is clearly visible. I measured the distortion differentially as before, but the gain of one channel has been trimmed back, producing the same waveform as the THD meter output in Fig. 3.

It is not possible to 'off-tune' the THD meter to produce a residual showing the same waveform as Fig. 4, as any mis-tuning grossly shifts the phase of the fundamental relative to the residual, long before its amplitude changes significantly.

In case anyone should still think that crossover distortion results in predominantly high harmonics, **Figs 6** and 7 should dispel the notion. Figure 6 is the spectrum of the input 440Hz sinewave shown in Fig. 2; it is supplied by a 10Hz to 10MHz video oscillator of modest purity. The second and third harmonics are about 0.07% and 0.05% respectively, the others all being around 80dB down or more relative to the fundamental, giving a THD of about 0.1%.

Figure 7 shows the spectrum of the signal after it has been through the output stage. The fundamental is about 1dB lower than the input, while the level of the second harmonic is little changed. This reflects the good matching of the devices, although higher order even harmonics are more prominent than in the input signal.

However, the third, fifth and other odd harmonics indicate the substantial odd-order distortion due to the crossover effect. Clearly, they roll off in much the same way as the spectrum of a squarewave, although they are all well down on the fundamental. This is of course because, as Fig. 4 shows, the 'squarewave' distortion component is itself much smaller than the fundamental sinewave.

Of course, a practical class B output stage will not show the severe crossover distortion of Figs 1 to 5 – even open loop. But a class B output stage will evidence some crossover distortion, especially after a few years use, due to drift of component values. Where this is such as to leave the stage slightly over forward-biased, the phase of the crossover distortion will be inverted relative to that shown in Fig. 4.

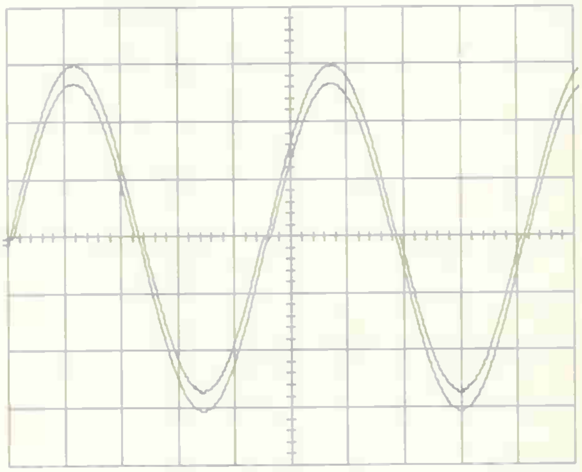


Fig. 2. Showing the input signal of 440Hz at 12V pk-pk, larger trace, and the output signal, smaller trace, at 2V/div. vertical, 500µs/div. horizontal. The output signal is smaller than the input due to crossover switching error.

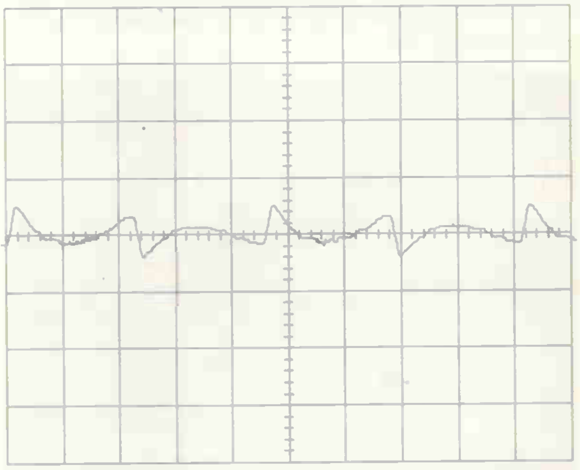


Fig. 3. Showing the residual signal from the simple class-B output stage, after removal of the fundamental. This residual was obtained using a THD meter.

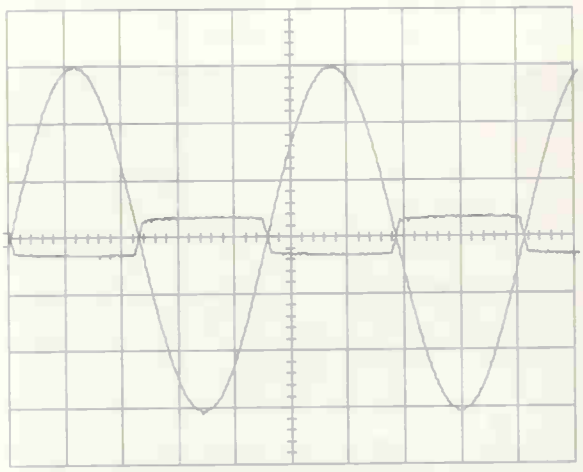


Fig. 4. The larger trace is the input signal. The smaller is the distortion, measured by probing both the input and output signals, and displaying trace A plus trace B, with trace B inverted, i.e. a differential measurement.

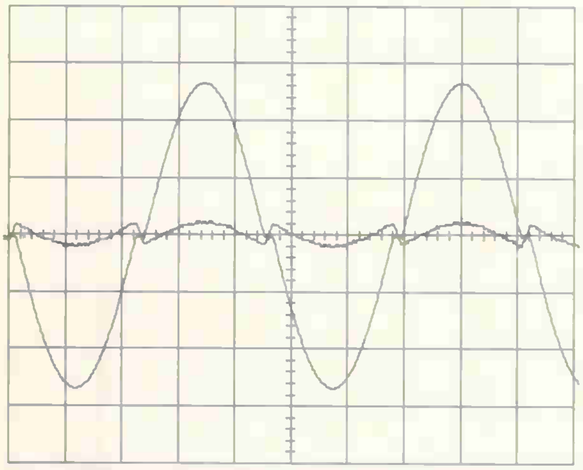


Fig. 5. The larger trace is the (distorted) output signal. The smaller is the distortion, measured as before, but the gain of one channel has been trimmed back, producing the same waveform as the THD meter output in Fig. 4. (The dead space steps in the output waveform are more clearly visible in this view than in Fig. 2.)

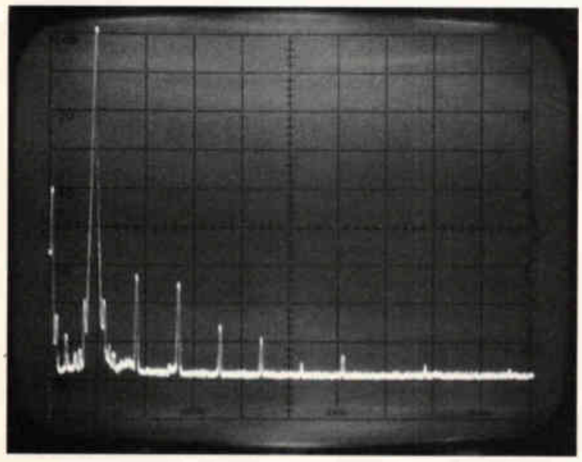


Fig. 6 The spectrum of the input signal, 440Hz. Spectrum analyser settings: span 0 - 5kHz, resolution bandwidth 10Hz, sweep speed 10s/div., display smoothing off.

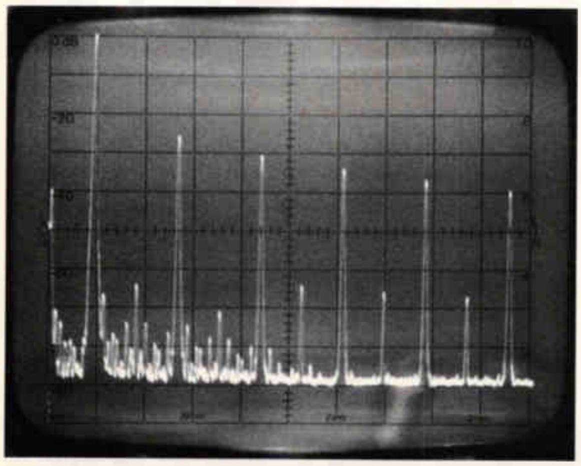


Fig. 7. The spectrum of the output signal, analyser settings as in Fig. 6, showing that crossover distortion is basically a squarewave.

Harmonics and intermodulations

While crossover distortion produces a range of odd harmonics of decreasing amplitude, a class-A amplifier will typically show far fewer harmonics of significant amplitude. These are principally second harmonic in the case of a single-ended amplifier and third in the case of a push-pull amplifier.

Two *identical* single ended amplifiers, harnessed together in bridge-tied load push-pull, will show zero second harmonic distortion, given a perfect phase splitter to drive them. However, the degree of distortion suffered by a pure sinewave is, of itself, of purely academic interest.

Sound produced by a solo musical instrument will normally contain many harmonics, both odd and even – except in the case of a clarinet, where the third harmonic greatly exceeds any second. As Dr Leakey, a one time colleague of mine and researcher into stereo reproduction once remarked, 'God never intended us to listen to sinewaves'.

However, it is important to remember that harmonic distortion is produced by non-linearity in the transfer characteristic of an amplifier. This same non-linearity has other, more dire, effects. When two tones are present, non-linearity will produce from them, other tones, not present in the input signal.

A degree of second-order non-linearity, typical of a single-ended amplifier, will produce – in addition to second harmonic products of the two tones – two extraneous tones not present in the original signal. These are the sum and difference frequencies.

For example, in the case of 1000Hz and 1333Hz – a major third – 333Hz and 2333Hz will appear. For some pairs of tones, such as a perfect fifth or a major third, the sum and difference tones (intermodulation distortion – IMD) will themselves be harmonically related to the original two tones, so they will not sound particularly discordant.

For more abstruse intervals though, the products become more discordant. When many tones are present, as in orchestral music, the opportunities for infelicitous products multiply alarmingly.

Where substantial third order non-linearity exists, the situation is even worse. As explained in chapter 4 of ref. 2, a two-tone input again results in extraneous products. But this time, not only do you have distortion components at the two fundamental tone frequencies themselves, and third harmonics of the original tones, but also the simple sum and difference products are replaced by third-order intermodulation products.

These products related to original frequencies f_1 and f_2 by the expressions $2f_1–f_2$, $2f_2–f_1$, $2f_1+f_2$ and $2f_2+f_1$. In the case of 1000Hz and 1333Hz, the third-order difference intermodulation products are 666Hz and 1666Hz. These appear either side of the two original tones, while the third-order sum intermodulation products appear between the two third-harmonic components.

Again as it happens, in the case of 1000Hz and 1333Hz the third-order difference products are not discordant. But on typical music, the effect is even worse than second-order intermodulation, which may itself also be present in a poorly balanced push-pull system. When substantial non-linearity is present, the result is a generally muddy sound, as commonly encountered on an old black-and-white pre-war movie.

Figure 8 shows how second-order non-linearity – the major non-linearity in a conservatively rated single-ended class-A amplifier – affects the transfer characteristic. Non-linear component 'd' is shown as 0.1, or 10% of the linear component.

A second-order parabolic curve added to a linear characteristic must, for large values of the abscissa (X-axis value), head off to plus infinity for both positive and negative values of X. You can see this in the upper plot. But over a restricted range, such as $-1<X<+1$, the expression can closely model a practical amplifier, as shown in the lower plot.

In fact, the characteristic looks pretty linear. But if you look at the page edge on, squinting along the line of the characteristic, it is clearly non-linear. Similarly, a third-order component can model the way the gain of a class-A push-pull amplifier departs from true linearity. The effects of both types of non-linearity, in terms of both harmonic and intermodulation distortion, are summarised later.

Which do I measure?

From time to time, there is discussion as to what measurements represent meaningful results when evaluating amplifier performance[3]. Should an amplifier be evaluated by the looking at level of its harmonic distortion, e.g. using a THD meter? Or should some other measure such as intermodulation performance be used?

From some articles, one might almost get the impression that a design trade-off is possible between THD and IMD. Would that this were so! If it were, one would surely choose to minimise or completely suppress the latter, at the expense of increasing the former.

But both forms of distortion are produced by one and the same mechanism – non-linearity of the transfer characteristic.

Given the degree of, say, third-order non-linearity of an otherwise distortion-free amplifier, the level of third-harmonic distortion at rated full drive is fixed, as is the level of third-order IMD due to a two-tone signal whose peak envelope voltage again equates to rated full drive.

Take for example a class-A push-pull amplifier whose transfer characteristic is given by $y=A(x-0.1x^3)$, where y is V_{out}, x is V_{in} and the non-linearity factor d is 0.1. For simplicity, let A=1 (unity gain as in Fig. 1) and let rated full drive be V_{in}=2V pk-to-pk. If V_{in} is 1×sin(ωt), where ω is 2π×440 radians/second, i.e. 1V peak at 440Hz, then – and I will spare you the maths – V_{out} is 0.925V at 440Hz plus 0.025V at 1320Hz.

Thus there is in addition to the third harmonic, a distortion component at the fundamental, reducing its amplitude by 7.5%. In the case of purely second-order distortion on the other hand, the amplitude of the fundamental is unaffected. But in addition to second harmonic, a small dc offset is produced. The offset is usually of no importance, being blocked either by a transformer in a valve amplifier, or a large electrolytic in a solid state amplifier.

These effects are tabulated in a separate panel, where again I have spared you the maths and simply quoted the results. However, as my maths was always prone to unreliability at the best of times, if any good reader checks the results and finds an error, perhaps he or she would be kind enough to inform the editor.

THD or IMD measurement?

It is clear from the preceding that a given non-linearity factor, d, determines both the harmonic distortion and the intermodulation distortion of an amplifier. Either type of measurement can therefore in principle be used to characterise an amplifier. In practice, on the other hand, there are various pros and cons to each, and these are summarised below.

THD measurement. Equipment needed to measure THD is a low distortion AF oscillator, a THD meter and, to take full advantage of the equipment, an oscilloscope to view the distortion products made available at the meter's residual output socket.

Commercial low-distortion oscillators and THD meters are readily available, or can be constructed for about £100 each from designs presented in this magazine. In any laboratory, a suitable oscilloscope should be available. Thus THD measurement is simple and straightforward, and relatively inexpensive in terms of equipment.

Since a THD meter can only notch out and suppress a single input signal, it cannot be used directly for intermodulation measurements. However, where one type of non-linearity predominates – second, or third, say – a measurement of percentage distortion permits the corresponding factor d to be determined. From this factor, intermodulation performance can be deduced if required.

Where several orders of non-linearity are significant, for example crossover effects, a THD meter provides a convenient single measure of distortion. Monitoring the THD meter's residual output on an oscilloscope, while adjusting the output stage bias on a class-B amplifier, is particularly revealing. ■

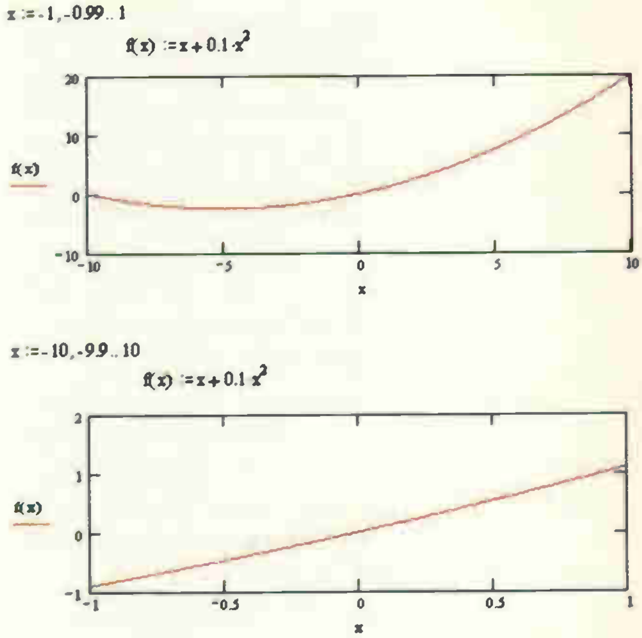


Fig. 8. Showing how second order non-linearity, the major non-linearity in a conservatively rated single ended class A amplifier, affects the transfer characteristic. Over a limited range, lower, it looks fairly linear.

Appendix

The transfer characteristic of an amplifier can be characterised as $f(x)=A(x+d_2x^2+d_3x^3+...)$ where x is the input voltage V_{in}, A is the gain and $f(x)$ is the output voltage V_{out}. To make the analysis convenient, I normalised A to unity, and took x as the rated full drive voltage, likewise normalised to unity. Relative values of the fundamental and distortion components in V_{out} were then calculated in terms of the second-order and third-order distortion parameters d_2 and d_3, and tabulated.

Second-order non-linearity

a) harmonic distortion:

V_{in}	V_{out}		
f	dc	f	$2f$
1	$d_2/2$	1	$d_2/2$

b) intermodulation distortion:

V_{in}		V_{out}						
f_1	f_2	dc	f_1	f_2	$2f_1$	$2f_2$	f_2-f_1	f_2+f_1
0.5	0.5	$d_2/2$	0.5	0.5	$d_2/4$	$d_2/4$	$d_2/4$	$d_2/4$

Third-order non-linearity

a) harmonic distortion:

V_{in}	V_{out}			
f	dc	f	$2f$	$3f$
1	0	$(1-3d_3/4)$	0	$d_3/4$

b) intermodulation distortion:

V_{in}		V_{out}								
f_1	f_2	dc	f_1	f_2	$2f_1-f_2$	$2f_2-f_1$	$3f_1$	$3f_2$	$2f_2+f_1$	$2f_1+f_2$
0.5	0.5	0	$(0.5-3d_3/16)$	$(0.5-3d_3/16)$	$3d_3/32$	$3d_3/32$	$d_3/32$	$d_3/32$	$3d_3/32$	$3d_3/32$

IMD measurement. The equipment required to measure intermodulation distortion includes two AF oscillators, an intermodulation-free means of combining their outputs into a single two-tone signal, and an AF spectrum analyser.

The specifications of the two oscillators are slightly more relaxed than those needed for THD testing. If new, however, the spectrum analyser is likely to be very expensive, and in the range £1000 to £1500 even if second hand.

Where one type of non-linearity predominates – second, or third, say – a measurement of corresponding intermodulation distortion permits the factor d to be determined. From this factor, it is possible to deduce harmonic distortion. More simply, one tone can be turned off and the harmonic distortion measured directly.

Where several orders of non-linearity are significant, for example in crossover effects, a single measure of distortion is not readily available, due to the large number of significant products at different frequencies.

Due to the limited on-screen dynamic range of an AF spectrum analyser, meaningful measurements on very low distortion amplifiers may simply not be possible.

In summary

The performance of hi-fi amplifiers may be evaluated either by THD measurements or by IMD measurements. Each method has its advantages and disadvantages, though these are sometimes overstated.

For example, the suggestion that THD measurements are meaningless and worthless is clearly an exaggeration, perhaps based on the misconception that THD and IMD are unrelated.

While it is true that intermodulation distortion is more disagreeable than harmonic distortion, you cannot trade one for the other. As the separate panel shows, both are directly related to the distortion parameters d_2, d_3, *etc.*

Far from being useless, a THD measurement at 10kHz, or even 20kHz, is very revealing. An amplifier whose full drive THD at 10kHz is little, if any, worse than at 300Hz clearly is not slew rate limited, and will therefore be substantially free of transient intermodulation distortion.

It is precisely to permit testing at 10kHz or 20kHz that THD meters have a response extending to 50kHz, 80kHz, or even in some instruments to 250kHz.

A very powerful method of measuring the performance of amplifiers with very low distortion is to use a spectrum analyser attached to the residual output of a THD meter. After notching out the fundamental, the THD meter may provide 40dB or even 60dB gain to the harmonics, extending downwards the effective range of the spectrum analyser by that amount. By this means, products down to around –120dB relative to the fundamental may be measured, limited only by the system noise floor[4]. ■

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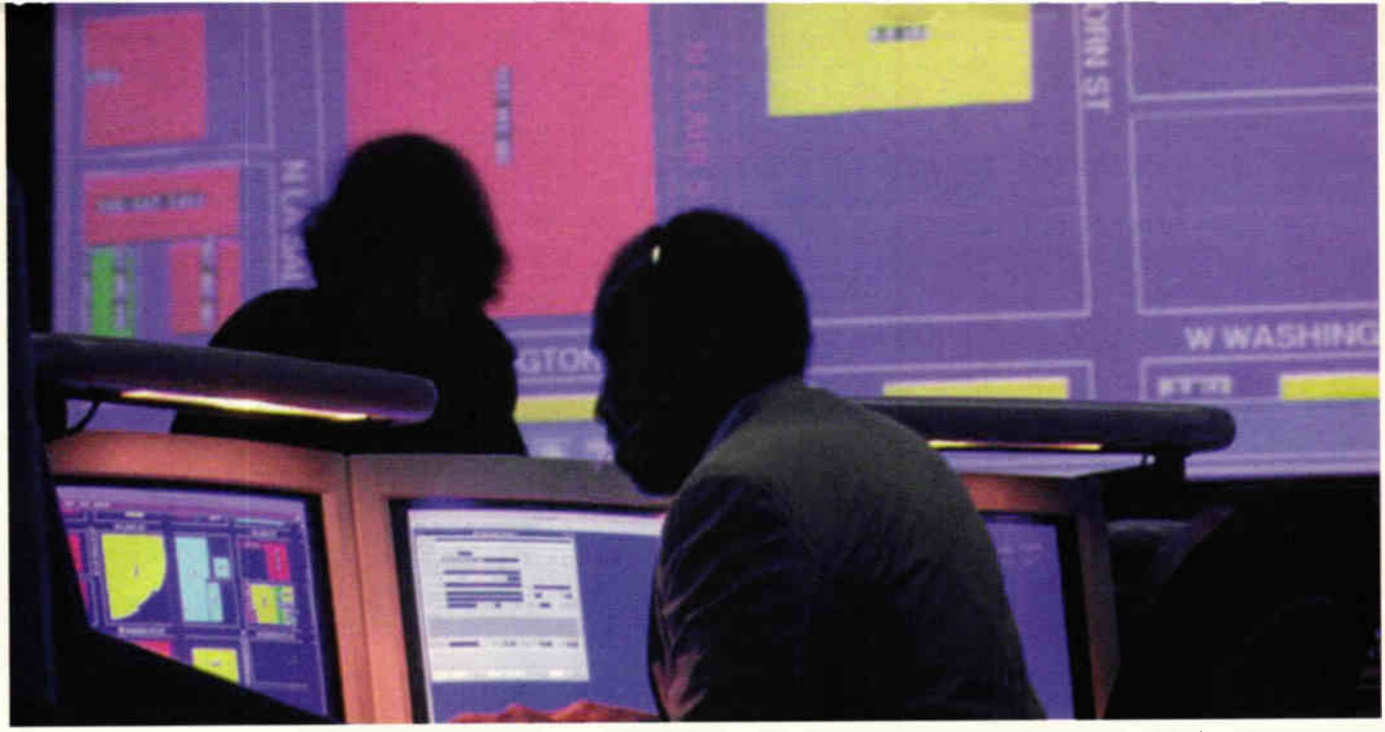


You are here... If your car is equipped with the RAC Trackstar vehicle security service then it can be located easily in case of theft or emergency. The system was developed by Trafficmaster and combines GPS satellite tracking from Trimble Navigation with a GSM wireless communications module. Together these allow the vehicle to be located and let the driver talk with the control centre.

Global positioning system comes of age

GPS in mobiles, cars and even watches is on the increase and it isn't stopping there, says Melanie Reynolds.

The number of devices that are able to determine fairly precisely where you are is increasing and, as the cost of the technology falls and its use becomes more valued, this number can only increase. Automobiles, hand-held computers, mobile phones, digital cameras and watches are just the beginning.

"We've found that as technology becomes available, people see what it will do for them," comments David Hall, vice president of mobile positioning and timing technologies at Trimble Navigation.

According to Hall, the future is even more appealing. "When you start adding in the concept of the Internet and instant access to information, because of the wireless communications link, which is tied to a location, then it becomes an awesome tool."

Hall sees the biggest growth market being in asset management. He sees assets ranging from trucks through diamonds to the immeasurable worth of human donor organs or an Alzheimers patient. These can all be kept track of – no matter what, no matter where.

"The number of packages moving around this world on an annual basis are about ten billion," estimates Hall. "We see a huge opportunity for managing those assets utilising GPS."

There are other areas where GPS comes into its own.

Architecture, engineering and construction can all make use of GPS's precision for their measurement purposes. Hall says the technology is accurate enough to be able to pinpoint a coin in a field.

GPS (global positioning system) is the satellite system most commonly used to determine position. Its satellites are financed, deployed and maintained by the United States.

There are 24 satellites in the constellation and three spares. The spares can be used to replace a faulty unit at any time. Eight new satellites are sent up at around five year intervals and eight are then taken out of service.

As you can imagine, this is not cheap. The system cost $13-15bn to launch and costs for maintenance are ongoing.

The Russians also have a system deployed which is called GLONASS (Global Navigation Satellite System) and Europe is planning its own known as Galileo.

The frequency spectrum for the European satellite positioning system was agreed last month at the World Radiocommunications Conference and it is tentatively scheduled to go into operation in 2008.

But is there room for another system?

According to Hall the answer is yes, because signals from the satellites in either systems can be used for the benefit of the end-user.

"We're intrigued with Galileo because we like the idea of interoperability. We're not tied to GPS so we're working with the people at Galileo now to understand the standards and the spectrum they intend to use," explains Hall. "We fully expect, and intend, to incorporate the capabilities of Galileo into what we do and merge the two."

The intentions of the Galileo creators are not yet clear, but Hall says the only potential problem with satellite systems will come if people start charging for the services.

The GPS system is free, and the US Government has said it will continue to be so for at least another ten years. "As we continue to get these pervasive applications protecting so much and saving life, how are you going to charge for it?" asks Hall. ■

X-ray specs

X-rays in brown envelopes are one of the most common sights in hospitals. But all that is set to change as digital imaging techniques that can store X-rays on computer take over.

For decades, hospitals have captured X-ray images on photographic film, moved them around on trolleys, and then filed them away in big brown envelopes.

But in the age of the digital camera, hospitals are now beginning to consider switching their radiology departments to solid-state imaging techniques that can capture an X-ray picture direct to computer. Such pictures could be enhanced and manipulated by image processing algorithms; computer-analysed for diagnostic information, as is done now with cervical smears; shared between specialists in different departments across a network; and archived on optical disk or tape when finished with.

This process – especially the archiving – is potentially much cheaper and more reliable than storing X-ray films. Films often go missing when they are urgently needed – even when film is only being transported within a hospital. Up to 13 per cent of films can go missing, according to statistics from University Hospital Wales. This means they have to be re-taken, implying not just extra expense at 50p a shot, but also higher radiation doses for the patient.

Besides this advantage, direct-to-digital radiography also delivers some benefits in terms of image quality. How, then, does it work?

Like all X-ray images, digital X-rays are created by shining radiation through the human body directly onto a light-sensitive plate. Because X-rays can't be focused, this detector plate has to be approximately the same size as the object being imaged. Traditionally the detector is ordinary photosensitive emulsion, but in digital radiography, it is replaced with either a semiconductor sheet or a phosphor screen.

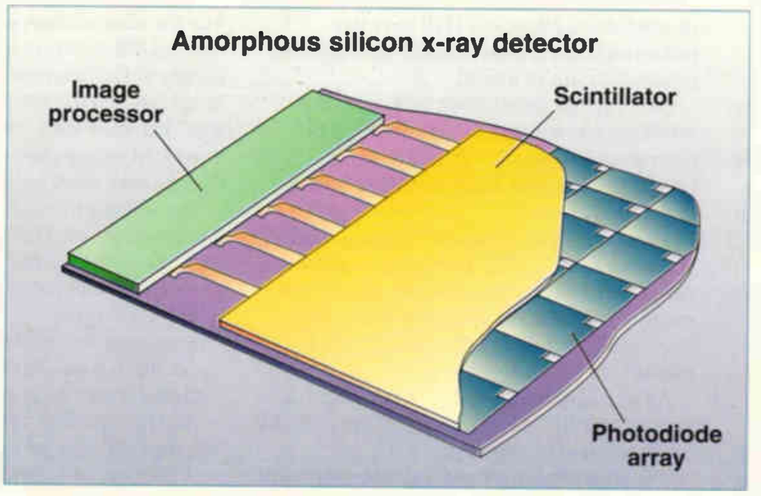
Most digital X-ray cameras adopt the semiconductor sheet approach, typically based on a large piece of amorphous (glassy) silicon. This panel is coated with a 'scintillator' material, typically caesium iodide. On the back of the sheet is printed an array of thin-film transistors (TFTs).

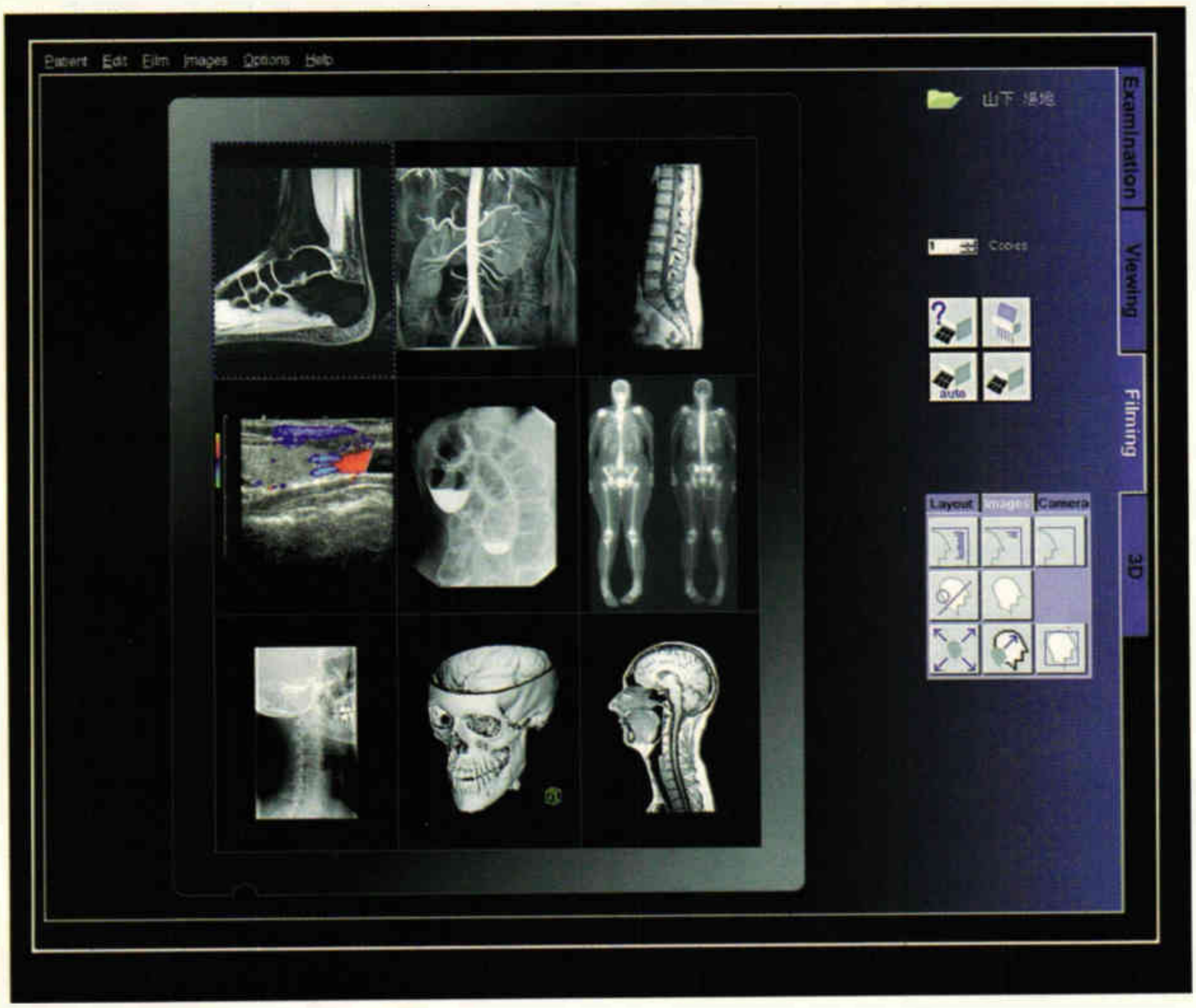
When an X-ray photon hits a point on the screen, the scintillator layer emits a photon of light into the semiconductor, creating a tiny, localised current, just like in a discrete photodiode. This current is amplified by the nearest TFT in the microcircuit array on the back of the screen. The array is constantly scanned by the instrument's control circuitry, which records the exact position of each such photoelectric event – thus building up a digital image of the amount of X-radiation incident at every point.

The difficulty has been to make the silicon sheets large enough. After at least ten years of research, the technology can now go up to several centimetres square. Several companies have developed instruments based on this technology. They include well-known medical instrumentation companies like GE, Siemens, Philips and Sterling.

Their products work at equivalent dose rates to those needed by ordinary X-ray film cameras, with less distortion from magnetic fields or background light, and the images are stable for a long time – unlike film.

Other companies, including Fischer and Swissray, have





developed an alternative method whereby the X-rays can be detected on a phosphor screen. The light from this is focused into a smaller CCD array, which accumulates the image over a period of up to 20 seconds. This two-stage approach, though apparently more complex and less sensitive, is cheaper because its components are pretty standard, unlike the large semiconductor panels. It has gained Swissray a market lead on its competitors.

So far the whole market has been slow to take off, despite its evident long-term prospects. Capital cost is no doubt one reason – though the running costs of digital systems are claimed to be lower, with no chemicals needed for development and far less space needed for storage. Another is the reluctance of some radiologists to move away from film – which is, when all is said and done, a very well understood and reliable technology.

But the main reason appears to be that, to get the most out of digital radiography, you need a digital hospital. To view the images, you need very high resolution workstations. To pass them around, you need a network – one that won't grind to a halt when people start circulating hundreds of 9Mbyte to 16Mbyte JPEG image files every day. You also need something called a "picture archiving and communications system," or PACS, which is a non-trivial combination of hardware and software.

Such infrastructure is neither cheap nor easy to install. For this reason the hospitals that *are* adopting digital X-ray technology tend to be either completely new building projects, or large multi-site trusts that have recently been merged and are thus faced with communication problems that they have to solve anyway.

Most installations so far are in the US. Recently, Data General announced it had won a contract to supply a PACS, with Swissray digital radiography, to New Jersey's biggest hospital, St. Joseph's. Swissray simultaneously announced orders for four more systems at big medical centres in Houston, New York and Philadelphia.

Direct digital radiography has barely touched the UK yet, but several hospitals have taken the first necessary steps – namely, converting their conventional film X-rays to electronic form by digitising them, so they can be managed with a PACS.

Two notable exceptions are the new Diana Princess of Wales Hospital in Birmingham and University Hospital of Wales (UHW) in Cardiff – whose Agfa digital X-ray system produces over 250 000 radiological images a year. Thirty more NHS hospitals are now in the early stages of procuring PACS, using funds provided through the government's £1bn NHS modernisation plan.

Once these PACS and associated networks are in place, the economics of going direct to digital will become much more attractive. It looks like the end is in sight for those dog-eared brown envelopes. ■

When sound was cylindrical

Joe Pengelly describes his quest for high-quality reproduction of cylinder recordings and his efforts to replay them using modern techniques.

For many years now, the high quality of sound being extracted from old 78 records, thanks to modern technology, has been truly amazing. Little of that technology, however, has been directed towards reproducing the surprisingly good sound signal to be found lying in the grooves of cylinder recordings.

Why is this? Well, the quality of the acoustic sound coming from the assorted horns of acoustic cylinder machines is poor enough to be rejected by even the most undemanding of listeners. Further, the duplication of spurious resonances in both the recording and acoustic replay processes, for the hi-fi buffs, is anathema.

Then there is the criticism that the repertoire on cylinder only embraces the popular or sentimental culture of its day. Hence it is of little consequence and not worthy of restoration efforts. This is a misconception though. Many of the classical greats of the early twentieth century were more than willing to commit their audio efforts to the cylinder format.

Certainly, such recordings are in the minority, but they are there, lying in cylinder grooves awaiting quality realisation. These considerations, along with the difficulty of finding efficient, smooth running contemporary machines for replaying cylinders has no doubt inhibited would-be experimenters in this field. The youngest traditional cylinder player is at least 70 years old.

It was my conviction, however, that a sound replay system that was feed-screw driven, merely floated a sapphire stylus in a hill-and-dale grove and not employing steel needles as used with laterally recorded 78 discs where the needle was progressed by groove walls, was worth investigation. Further, the hill-and-dale groove recording system obviated blasting both in the record and replay process. To this end I constructed a breadboard set-up, as in **Fig. 1**.

The set-up uses a spring-driven Edison cylinder 'Concert Opera' machine, minus its horn. Since this machine not only revolves a cylinder on its mandrel, but also traverses it, I was able to use a stationary high-quality pickup head and cartridge with a diamond stylus of the correct dimensions. This is a high-quality component made by Expert Pickups. Even with such a crude set-up the results were encouraging.

A modern cylinder replay machine

A Leverhulme research grant enabled me to commission Plymouth University to build an all-purpose electrical cylinder replay machine. Built to my specifications, this machine is capable of playing every known speed, size and groove configuration of cylinder, **Fig. 2**.

The machine is servo speed controlled from 1 to 250rev/min. It can run backwards so that the stylus impinges on the grooves in a fresh way. Also, the pickup head can be angled in relation to the groove over a very wide range.

Further, with its stylus merely floating in a cylinder groove, the continued maintenance of the pickup arm at right angles to a cylinder at all times is confirmation that the correct feedscrew value is being used. Above all, the pickup head plays with all the delicacy of that of a high-grade long-playing record deck.

All credit must go to Mike Stringer of Plymouth University for constructing this superb piece of machinery from what was merely my general concept. With such a machine the whole cylinder world was made available to me.

Unwanted bumps

Having this remarkable cylinder replay machine though, I soon ran into problems with the fidelity of the sound signal being extracted from cylinder grooves. Pre-eminent amongst these problems was the 'bump' rendered with unwelcome fidelity. This is caused by the pickup head bobbing up and down in a groove due to a lack of roundness in many cylinders.

My colleague, David Lane, has addressed this problem and all but removed the offending bump, together with a great deal of attendant rumble.

Electronic enhancements

The replay signal from a cylinder recording is an extremely complex one. Alongside the wanted information, there are high levels of broadband noise, impulsive noise – or clicks – and rumble. The first stage of amplification must be very linear if all these components are not to intermodulate with the wanted signal.

Additionally, the first stage must not clip on even the largest amplitude 'click'. If it does, truncation will spread the spectrum, making it more difficult to reduce subsequently by filtering.

A Burr-Brown OPA604 was chosen as the first amplifying device because of its very low distortion performance. It can also operate on ±22V rails, ensuring adequate headroom for even the largest of spurious noises.

Fig. 1. Something old, something new – a spring-driven Edison cylinder 'Concert Opera' machine, with its horn replaced by a modern light-weight arm and cartridge.

For a constant-amplitude recording, a magnetic pickup cartridge, because it is velocity operated, will produce an output that rises at 6dB per octave with increasing frequency. Of course, there was no 'standard' recording characteristic for acoustic cylinder recordings, but we felt it advisable to counter the rising output with frequency from the magnetic pickup.

A passive equalisation network achieved this compensation. It provides a 6dB/octave rise in the bass, being +3dB at 200Hz and a 6dB/octave fall in the treble, being –3dB at 6kHz.

Loss due to this network was made up by another OPA604 gain stage before passing the signal to conventional Baxandall-type bass and treble controls and a Quad-style variable slope low-pass filter, also based on a OPA604. At the normal output level of 1V rms, the THD between 100Hz and 10kHz is below the noise. Not until the output is raised to 5V rms does the THD reach 0.01%, indicating the good linearity of the design.

Rumblings

A particular problem with cylinder recordings is the high level of rumble. The high rotational speed of 160rev/min – when compared with LPs at 33-$^{1}/_{3}$rev/min and the 78rev/min of coarse-groove records – pushes the rumble components further up the passband.

Fortunately, there is little programme information present in a cylinder recording below 200Hz, making it relatively easy to remove the rumble with a steep cut filter.

In this case a 15-pole Butterworth high-pass filter with a –3dB point at 160Hz was used. This filter eliminates almost all traces of rumble. As a bonus, it greatly reduces the 'bump' that occurs once per revolution because of a lack of roundness in cylinders that have become deformed over the years.

Alternative replay methods

David Lane has significantly reduced many of the noise problems associated with the electrical replay of cylinders, reducing the 'bump' to an occasional, more benign, 'thump'.

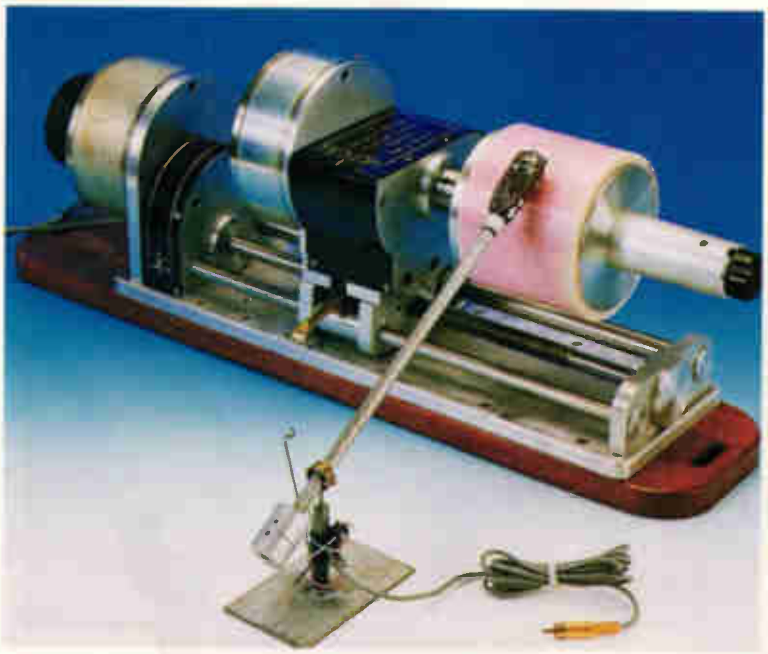


Fig. 2. Built at Plymouth University, this all-purpose replay machine is capable of playing every known speed, size and groove configuration of cylinder.

However, the lack of roundness in many cylinders prevents any method other than acoustic replay being used. A heavy acoustic sound box alleviates the bobbing up and down in a traditional player. Similarly weight also reduces the bump experienced when using a modern lightweight pickup.

It is the limited weight that can be imposed on the stylus bar of a modern electrical cartridge that remains the bugbear. Hence my letter of enquiry in the May 2000 issue of *Electronics World*, in which I asked if anyone could help me with a non-contact replay system involving, say, a light beam.

A side advantage of such a system would be its capability to be focussed even as a signal is being extracted from a groove, in order to achieve the best sound quality.

Whether such a light beam replay system will obviate wow and flutter in cylinder replay, resulting in a certain sourness of sound, remains to be seen. Not only are many cylinders out of round. They can have groove depressions within their ovality, compounding the problem. This depression defect causes a stylus to further accelerate and decelerate in a groove, resulting in an additional sourness of sound.

Wax cylinders

In the main, *wax* cylinders do not suffer from lack of roundness with its attendant problems. However, because of their soft composition, they are eroded significantly with every playing.

It is for this reason that all but one of the tracks on the CD that I have produced has originated from hard-wearing celluloid cylinders. When I have ventured to play wax cylinders electrically however the results have been startling. Such a track can be heard at the conclusion of the CD, in the production of which credit must go to Peter Cox and David Lane. ■

TiePieScope HS801 PORTABLE MOST

ABRITARY WAVEFORM GENERATOR-
STORAGE OSCILLOSCOPE-
SPECTRUM ANALYZER-
MULTIMETER-
TRANSIENT RECORDER-

Reliability

- The HS801: the first 100 Mega samples per second measuring instrument that consists of a MOST (Multimeter, Oscilloscope, Spectrum analyzer and Transient recorder) and an AWG (abritary waveform generator). This new MOST portable and compact measuring instrument can solve almost every measurement problem. With the integrated AWG you can generate every signal you want.

- The versatile software has a user-defined toolbar with which over 50 instrument settings quick and easy can be accessed. An intelligent auto setup allows the inexperienced user to perform measurements immediately. Through the use of a setting file, the user has the possibility to save an instrument setup and recall it at a later moment. The setup time of the instrument is hereby reduced to a minimum.

- When a quick indication of the input signal is required, a simple click on the auto setup button will immediately give a good overview of the signal. The auto setup function ensures a proper setup of the time base, the trigger levels and the input sensitivities.

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

- Measured signals and instrument settings can be saved on disk.This enables the creation of a library of measured signals. Text balloons can be added to a signal, for special comments. The (colour) print outs can be supplied with three common text lines (e.g. company info) en three lines with measurement specific information.

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

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

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

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

Web: http://www.tiepie.nl

CIRCLE NO. 109 ON REPLY CARD

Professional video multiplexer

Known commercially as the VMUX-8x1, Emil Vladkov's multiplexer selects one out of eight video sources and can be set to cycle round the inputs. Controlled via simple RS232 commands, and relatively easy to implement, the multiplexer can be used at home, or stacked for use in studio and cable TV applications.

Technical Specifications

Video input	
Video type	composite AC coupled, back porch at ground
Impedance	>200kΩ, unterminated
Number of inputs	8
Level	1 V pk-pk nominal +1.5 V to 0.5 V
Video output	
Number of outputs	1
Impedance	75Ω AC coupled
Frequency response	<±0.5dB to 50MHz 3dB bandwidth 110MHz
Gain	1±0.05 dB
Differential phase	<0.03°
Differential gain	<0.04%
Tilt	<0.1% – AC coupled mode
Crosstalk	<–66dB worst case @ 4.43MHz <–77dB typ. @ 4.43MHz
S/N – inc. hum	>70dB unweighted wideband >76dB unweighted to 10MHz
Feedthrough	<–63dB @ 10MHz
Switching period for circular mode	0.2s to 54min 36s
Delay in stack for circular mode	0.1s to 54min 35.9s
Interface	Serial RS-232, 9600 baud

This video multiplexer has eight inputs and one output. Having extended features, it is a versatile device for a broad spectrum of applications.

I developed the multiplexer for studio use, so its performance and the reliability conform to professional standards. This does not limit the use of the device for non-professional use by enthusiasts and hobbyists though. It is also low cost and reasonably easy to implement.

For example, the system can be used in home surveillance systems or for routeing video sources in home entertainment systems. Of course all useful features of the VMUX-8x1 become notable in the studio environment, where I use the system described as a carousel for selecting from 48 cable channels on a 6-input multi-image system.

Circuit details

Figure 1, the multiplexer's complete circuit diagram, shows that the heart of the device is IC_1 – a MAX440 eight-into-one video multiplexer. This device incorporates an output buffer to drive 75Ω loads directly.

Switching time is minimised and transients are suppressed via the circuit consisting of POT_1, R_{26} and C_{28}, connected to the V_{NS-} input of the IC. The analogue input voltage must never be more negative than the voltage on this pin. Resistor R_{26} provides the necessary protection.

Resistor network R_9, R_{10} provides the necessary amplification (×2) to drive a 75Ω load through the 75Ω output impedance.

Inputs of the multiplexer are high impedance – at nearly 270kΩ – due to biasing resistors $R_{1\text{-}8}$. As a result, the input signals can be relayed to other devices, such as RF-modulators. In this case the video multiplexer will not load the 75Ω line, but will multiplex the input signals.

Of course if no external device is connected daisy-chained after the multiplexer, external 75Ω loads have to be connected to the transit outputs, shown as Transit video outputs on **Fig. 2**, of the multiplexer to match the video signal.

The video signal is AC-coupled to the multiplexer IC, so the DC voltages

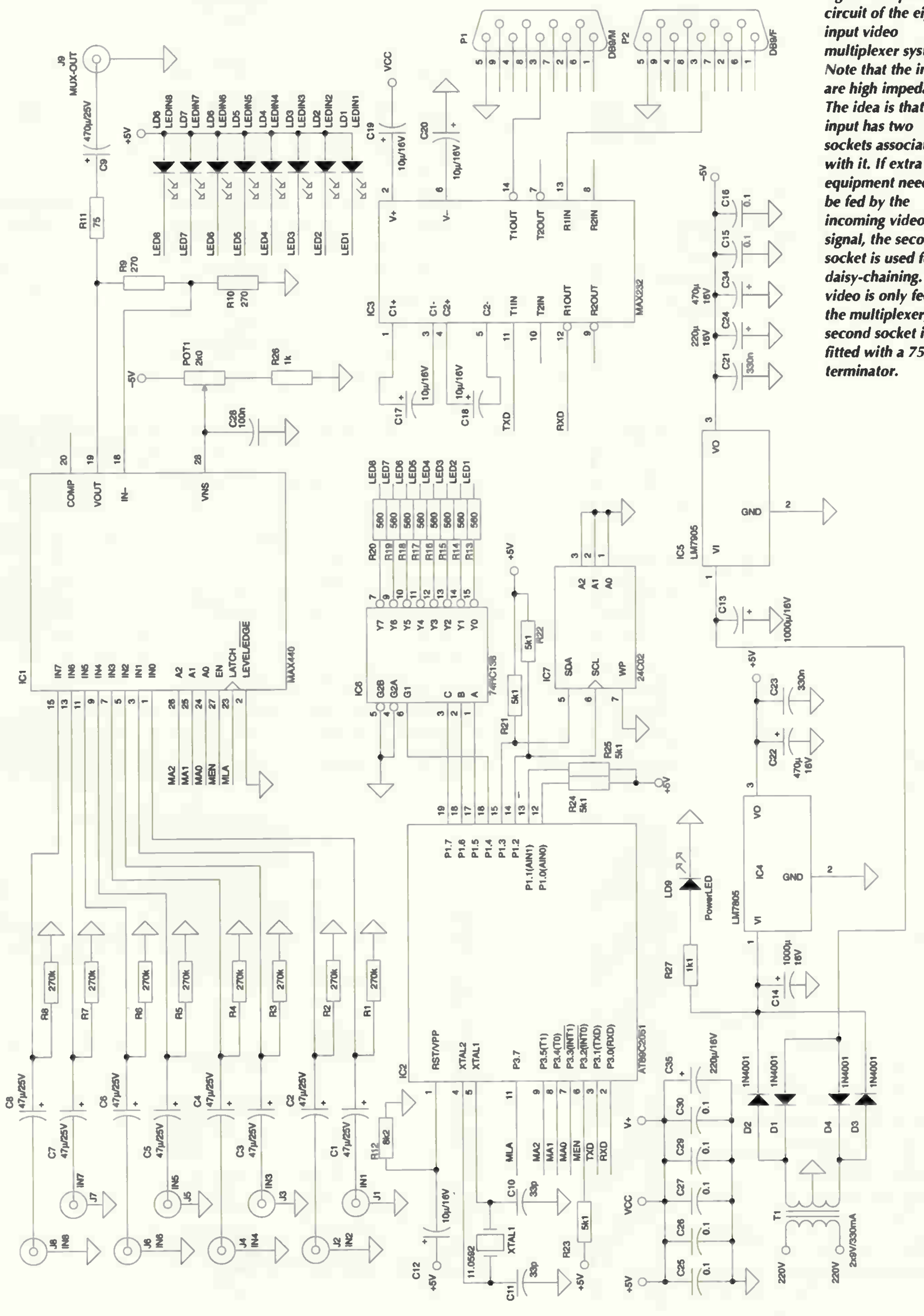


Fig. 1. Complete circuit of the eight-input video multiplexer system. Note that the inputs are high impedance. The idea is that each input has two sockets associated with it. If extra equipment needs to be fed by the incoming video signal, the second socket is used for daisy-chaining. If the video is only feeding the multiplexer, the second socket is fitted with a 75Ω terminator.

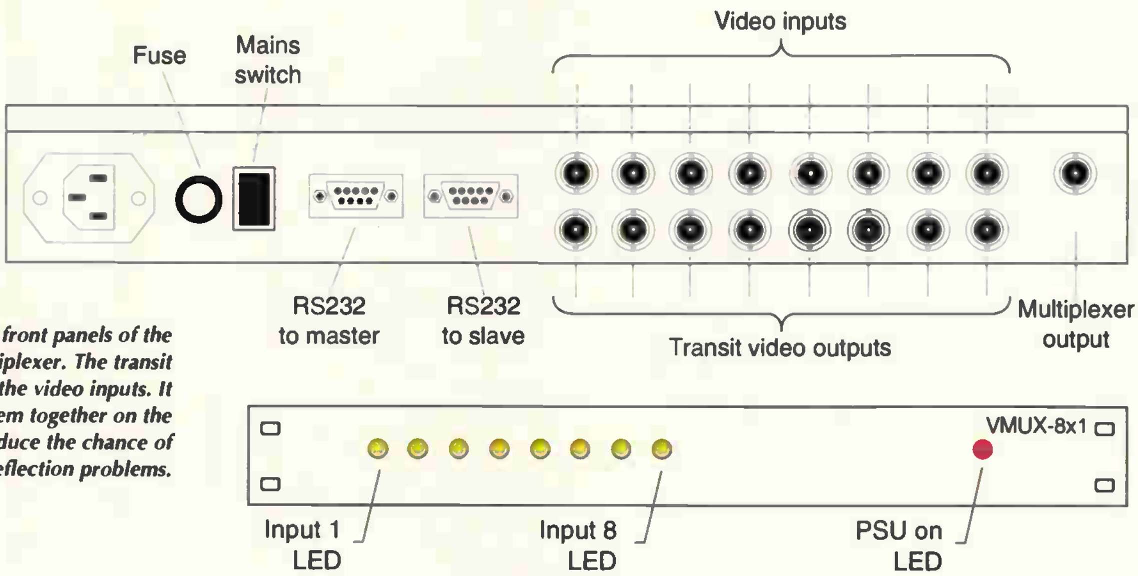


Fig. 2. Back and front panels of the author's multiplexer. The transit outputs connect to the video inputs. It is best to connect them together on the PCB though to reduce the chance of reflection problems.

present on the video lines are blocked. The AC coupling is performed by capacitors $C_{1\text{-}9}$. Of course if DC-coupling is essential for the success of the design, these capacitors can be shunted or omitted.

Microcontroller functions

An AT89C2051 8-bit microcontroller, IC_2, controls the multiplexing system. This device device has 2K of flash memory, where the program code for the system is stored.

Through lines P3.3, P3.4 and P3.5, tied to the address inputs of the multiplexer, the microcontroller selects which input is connected to the output of IC_1. The output amplifier of the system has an enable function, so the output of IC_1 can be enabled or disabled via P3.2. When P3.2 is high the device is enabled.

All digital control lines are latched in the internal registers of the multiplexer IC on the rising edge of the latch input. This input is connected to P3.7 of the microcontroller.

The level/$\overline{\text{edge}}$ input of the MAX440 has to be tied low to enable edge-triggered, rather than level-controlled, latching.

The multiplexer system has a feature that stores a start-up configuration. This determines which connection becomes immediately active after power is applied to the device. This information has to be nonvolatile, so it is stored in the EEPROM, IC_7. This eprom can be a 24C02 or similar device with 256 bytes of storage.

Address lines A0, A1 and A2 of EEPROM IC_7 are normally used in systems where many memories share the same serial bus. As, in this case, only one EEPROM is used, all address inputs are tied low.

Lines P1.2 and P1.3 of the microcontroller form the serial bus, which has serial clock and serial data lines designated SCL and SDA respectively. The protocol of the serial bus is generated in software.

Synchronising switching

The proposed implementation of the multiplexer has no sync. input capability, which means that switching

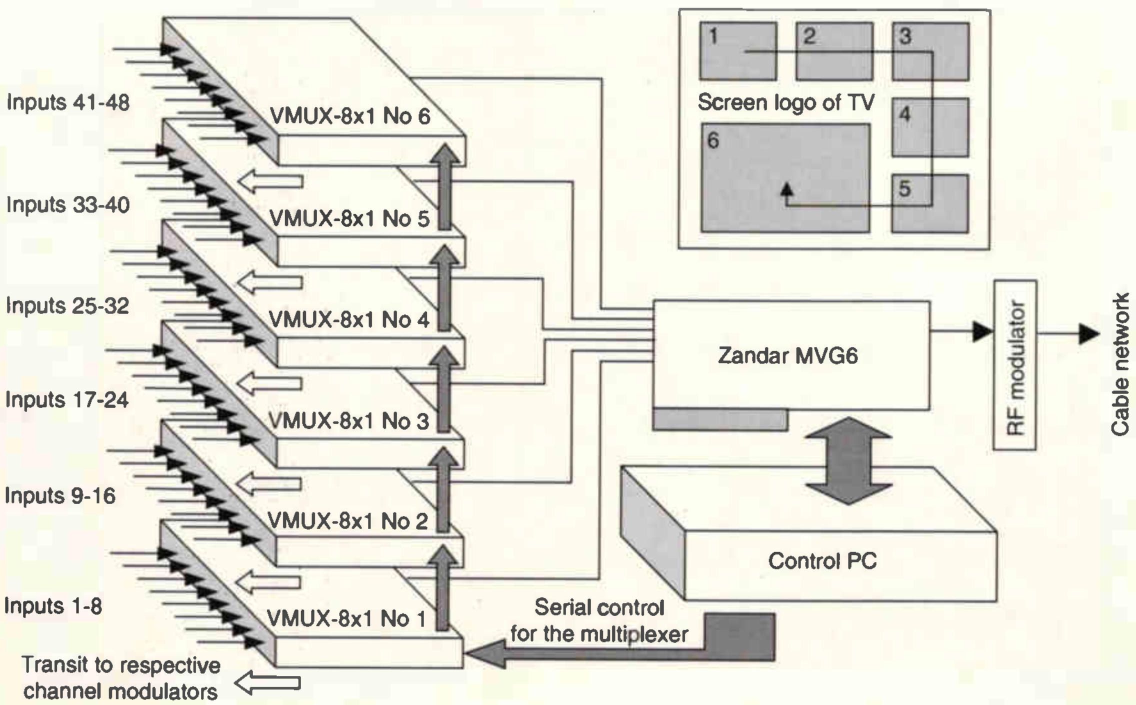


Fig. 3. Application example for stacking multiplexers. The Zandar unit is a commercial device that syncs all the video sources so that there's no switching flicker. It does this using frame buffers – but a much more cost effective idea from the author is described in the main text.

occurs not exactly on the vertical sync pulse of the video signal. This was not a problem in the case of my multi-image display system, where six video multiplexers are used, **Fig. 3**.

The PCI-card of the six-input multi-image system from Zandar Technologies that I use resynchronises all inputs through internal elastic store buffers. As a result, little or no switching transients are visible in the display.

If the vertical sync. capability is a must, than the problem can be easily solved if the analogue comparator inputs of the microcontroller, AIN1 on P1.1 and AIN0 on P1.0, are used. One of them can be tied to ground and the other can monitor the video signal, so if a negative-going sync. pulse with the right duration is detected in software, the switching can be performed synchronously to this event.

Although a visual indication of the active connection is not vital for the design, it can be extremely helpful in debugging and use in studio environment. This visual indication is accomplished by LEDs $LD_{1\text{-}8}$.

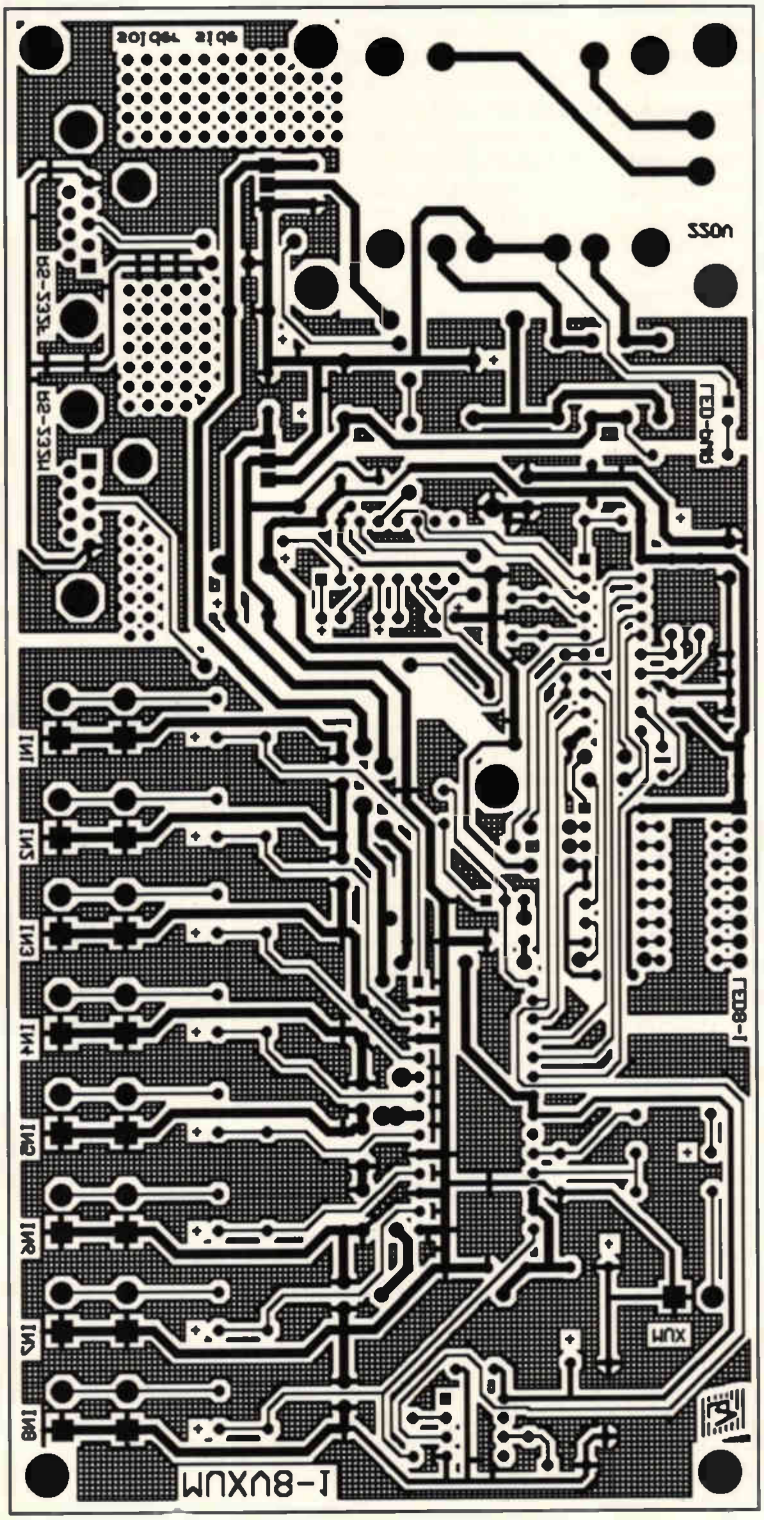
I used 5mm yellow flat surface LEDs. These connect through current-limiting resistors $R_{13\text{-}20}$ to the outputs of the address decoder IC_6, which is a 74HC138.

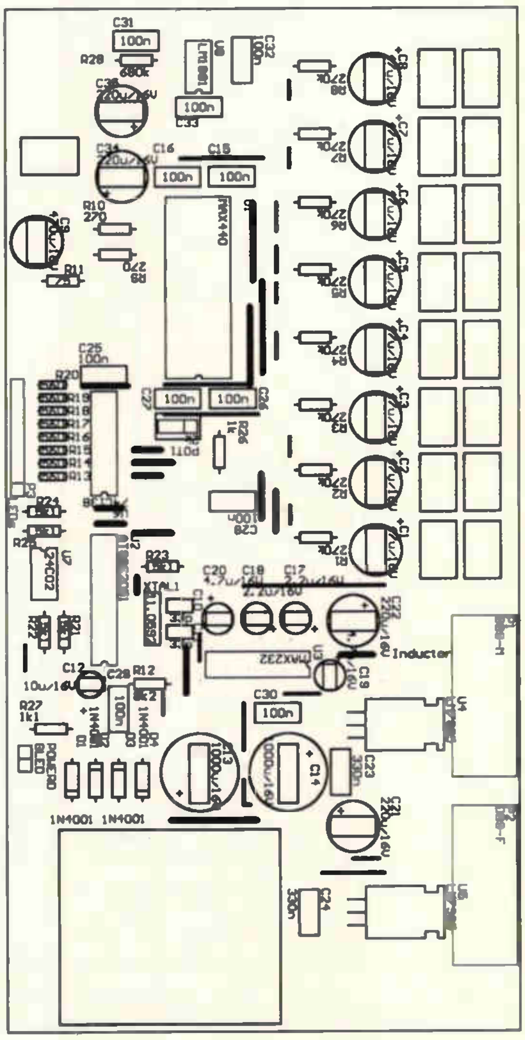
One connection is assigned to each LED. If the output amplifier is disabled, all LEDs are turned off. The status of the LEDs is controlled through lines P1.4 to P-1.7 of the microcontroller.

Controlling the system

The multiplexer system is controlled entirely via a PC through ASCII-commands on a Windows GUI interface. An asynchronous serial RS-232 connection is formed by the lines P3.0 and P3.1 of the microcontroller, which become RXD and TXD respectively.

The micro has a build-in asynchronous serial interface. Settings for the serial connection are: no parity, 9600bit/s, 8 data bits, 1 stop bit, i.e., N,9600,8,1. Levels required for prop-





Figs 4a/b). PCB layout and component placings. The design can be implemented on a single-sided board, greatly reducing the amount of work involved to repeat the design.

List. 1. Complete object code for the video multiplexer's Atmel microcontroller. Instructions for programming the controller can be downloaded from Atmel's web site as a PDF file.

```
:030000000200807B
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:08001B003232323232323232324D
:080023001201BF323232323209
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:10018000D11205C440F41206A979807A5A1207BE2A
:10019000E4C3D2D11205C440F479907A5A1207BE52
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:0307F0006B6F76B6
:00000001FF
```

er RS232 operation are generated by the IC_3 interface chip.

Stacking multiplexers

As mentioned earlier, the system can be used on its own, or stacked, as in Fig. 3. The stacking ability needs a special arrangement for the RS-232 ports.

Receive line RXD connects to P_2 on the female DB9 D-type connector. It performs the connection to the PC or to the master device. This line receives commands from the controlling device or PC. The command is then decoded, changed – or not – and retransmitted through the TXD line to the next slave device, P_1 male on the DB9 D-type connector.

Because of the analogue nature of the processed signals, a linear power supply is used in the VMUX-8x1. It consists of a mains transformer, bridge diodes D_{1-4} and linear regulators IC_4 and IC_5. These regulators provide the +5V and –5V, necessary for the video matrix and the logic circuits. C_{25-30} are power supply HF blocking capacitors.

The indication for the power-on state of the whole device is accomplished by the LED LD_9. I used a 5mm red flat surface type, with R_{27} providing current limiting.

Circuit board

I used the single-sided printed circuit board with the layout shown on **Fig. 4a)**. Figure **4b)** shows the component side, with the jumper-wires shown as thick lines.

All LEDs are combined on the connector, P_3 because they are mounted on the front panel. A ribbon cable can be used to connect them to the PCB.

I used a dual 9V/330mA mains transformer. It is mounted on the PCB. The video inputs and output on the board connect to the rear panel-mounted BNC-type connectors via short pieces of coax cable.

Mechanical considerations

My prototype is in a 19in 1RU enclosure. I recommend that you use a metal case that is properly grounded.

The indicator LEDs are flat-surface 5mm types mounted in the same plane as the front panel. BNC-type connectors are used for the eight video inputs, for the eight transit video outputs and for the multiplexed output.

Each transit output connects directly to the corresponding video input internally. I recommend that you make the connection not on the rear side of the back panel, but on the multiplexer's PCB, to avoid transmission line problems.

I put the master and slave D-type connectors on the back panel of the enclosure. The master is female and the slave male.

Code for the on-board micro

The object code for the on-board microcontroller 89C2051 is presented in **List 1**. This code is loaded into the 2Kbyte flash of the microcontroller with a programmer for Atmel's series of microcontrollers.

If you don't have such a programmer then visit the Atmel web-site and download the pdf-file with instructions how to build one.

The proposed code fits exactly in the 2K flash memory of the device, but if you want to modify and extend it, then the use of a device with more memory on board will be necessary

List 2. The multiplexer is designed to respond to these commands, sent over an RS232 link at 9600 baud. Each command is terminated with a carriage return only – no line feed.

Command	Description
`SW1`	Connects i/p 1 to the VMUX-8x1 o/p
`SW2`	Connects i/p 2 to the VMUX-8x1 o/p
`SW3`	Connects i/p 3 to the VMUX-8x1 o/p
`SW4`	Connects i/p 4 to the VMUX-8x1 o/p
`SW5`	Connects i/p 5 to the VMUX-8x1 o/p
`SW6`	Connects i/p 6 to the VMUX-8x1 o/p
`SW7`	Connects i/p 7 to the VMUX-8x1 o/p
`SW8`	Connects i/p 8 to the VMUX-8x1 o/p
`SW1D|Hbyte|Lbyte|`	Switch i/p 1 to o/p with delay, inserted before. The delay is given in 50ms increments. *HByte* and *LByte* are ASCII-expressed bytes in hexadecimal form. So you can enter SW1D0F8B, where 0F8B is the hex expression of the delay in 50ms increments.
`SW2D|Hbyte|LByte|`	Switch i/p 2 to o/p with delay, inserted before.
`SW3D|Hbyte|LByte|`	Switch i/p 3 to o/p with delay, inserted before.
`SW4D|Hbyte|LByte|`	Switch i/p 4 to o/p with delay, inserted before.
`SW5D|Hbyte|LByte|`	Switch i/p 5 to o/p with delay, inserted before.
`SW6D|Hbyte|LByte|`	Switch i/p 6 to o/p with delay, inserted before.
`SW7D|Hbyte|LByte|`	Switch i/p 7 to o/p with delay, inserted before.
`SW8D|Hbyte|LByte|`	Switch i/p 8 to o/p with delay, inserted before.
`SW1S`	Stores the connection i/p 1 – o/p as StartUp
`SW2S`	Stores the connection i/p 2 – o/p as StartUp
`SW3S`	Stores the connection i/p 3 – o/p as StartUp
`SW4S`	Stores the connection i/p 4 – o/p as StartUp
`SW5S`	Stores the connection i/p 5 – o/p as StartUp
`SW6S`	Stores the connection i/p 6 – o/p as StartUp
`SW7S`	Stores the connection i/p 7 – o/p as StartUp
`SW8S`	Stores the connection i/p 8 – o/p as StartUp
`SWOFF`	Disables the output amplifier of the VMUX-8x1
`SWOFFS`	Disabled output is the StartUp configuration
`CSW|B|E|PerH|PerL|`	Synchronous Circular Switching Command. Switches from begin input (*B* is a number between 1 and 8) till end input (*E* is a number between 1 and 8, but greater than *B*) with wrap-around. The interval between switching events is given in *PerH* (high byte – 2 ASCII symbols, representing the hex representation of the byte) and *PerL* (low byte).
`DCSW|B|E|PerH|PerL|DelH|DelL|`	Delayed Circular Switching Command. Switches from begin input (*B* is a number between 1 and 8) till end input (*E* is a number between 1 and 8, but greater than *B*) with wrap-around. The interval between switching events is given in *PerH* (high byte – 2 ASCII symbols, representing the hexadecimal representation of the byte) and *PerL* (low byte). The switching command SW1D... is retransmitted to the next stacked multiplexer after a delay, given in *DelH* (high byte – 2 ASCII symbols) and *DelL* (low byte).
`SCSW|B|E|PerH|PerL|`	Saves the Synchronous Circular Switching Command as the StartUp-command of the system.
`SDCSW|B|E|PerH|PerL|DelH|DelL|`	Saves the Delayed Circular Switching Command as the StartUp-command of the system.

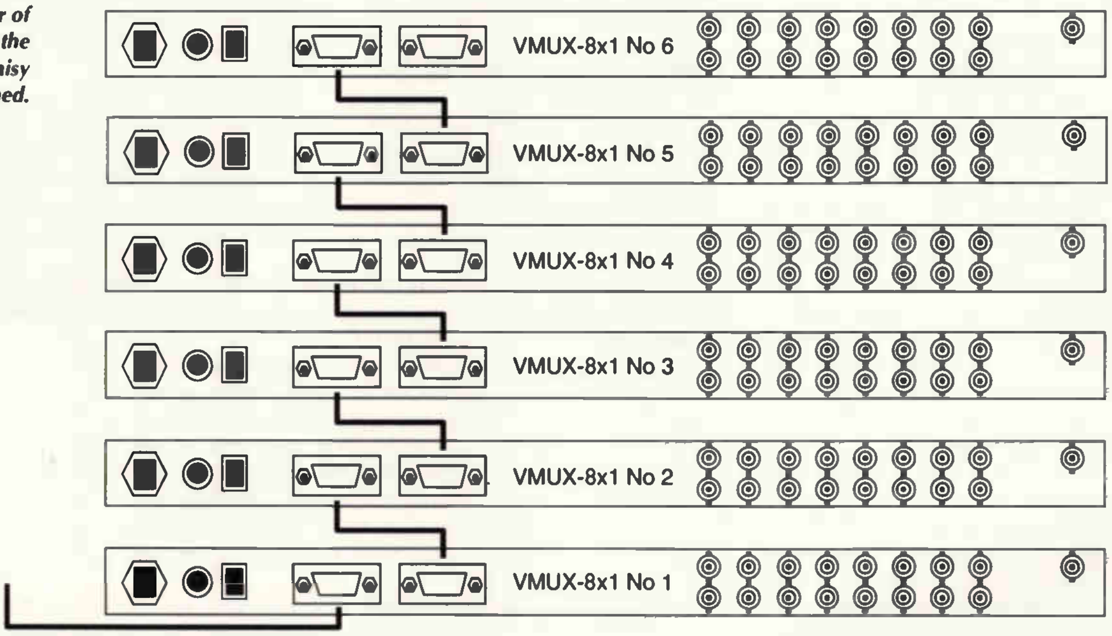

Fig. 5. When a number of multiplexers are stacked, the RS232 ports need to be daisy chained.

Such a device is the 89C4051 with 4Kbyte flash memory.

The code has the following functions in the system:

- Performs all necessary initialisations of the system components;
- Receives commands from the host-device – a PC or master-VMUX-8x1-system – decodes and executes them;
- Generates the specified delays for circular switching and stacked switching with delay, discussed later;
- Addresses and loads the MAX440 video switch with data;
- Turns on and off the LEDs conforming to the specified configuration;
- Writes and reads the serial EEPROM of the system, where the start-up-configuration is stored. After power-up loads the start-up configuration from the EEPROM. All commands stored in the EEPROM are backed up several times in the memory for increased reliability.
- Retransmits immediately, or with a delay, the received commands to the next stacked device.

Controlling the system

I have produced a graphical user interface for the controlling the multiplexer, running under Windows '95.

For those of you who prefer to write your own software, the ASCII-commands that the multiplexer responds to are given in **List 2**. These commands can be entered from a terminal emulating program running on the PC.

All commands end with a carriage return – not with carriage return plus line feed – so take care not to enter both in the command line.

How does stacking work?

Stacking allows multiple multiplexers to be connected together in a serial manner.

The multiplexers can be controlled from one PC, or from the first multiplexer, running in circular mode. In this mode, the inputs are sequentially switched to the output with a switching period specified by the user of the system.

To configure the system to accept stackable commands, the devices have to be daisy chained through standard serial cables as depicted on **Fig. 5**. I have used 6 devices in a multi-image system. They are numbered 1 to 6.

Multiplexer No 1 is the master of the carousel system and is connected to the PC to accept configuration information. The slave-output of No 1 connects to the master input of the next slave device No 2.

In the same manner device No 2 connects to device No 3 and so on. Theoretically there is no limit to the number of devices that can be stacked, but the delays in the serial transmission of the commands between devices have to be taken into account if the number of stacked multiplexers is too big.

Operating in circular mode

The circular mode of operation is an interesting feature of the multiplexer. It allows you to switch automatically between the input channels.

The start and end input channel for this mode is specified in the command or through the GUI interface. The only restriction is that the end channel has to be greater than the start channel.

You can specify channel three as the first and channel six as the last, for example. The switching is performed with a switching period, specified in the direct command, or through the Windows application software.

After the last channel is reached, the mode wraps around to the start channel – in the case of the above example from input six to input three.

There is another mode of circular operation, namely circular mode with delay. In the first circular mode, the command for switching the corre-

Software options

For a copy of the object code as a text attachment, e-mail Jackie Lowe at jackie.lowe@rbi.co.uk with the subject heading 'Multiplexer software please'.

A copy of the Windows GUI and object code in software form plus the Protel PCB file is available for £9 inclusive. If we receive a large number of requests, we may have to send this software to you by post rather than e-mail. The source code plus GUI and PCB file is available for £25 fully inclusive.

For payment, send a postal order or cheque payable to Reed Business Information, or send your credit card type, number, expiry date, card holder's address and a telephone number.

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sponding channel is transmitted from the master-device to all slave devices, so that all slaves perform circular switching synchronous to the master-device.

In delayed mode, a switching command with delay is transmitted to all slaves. The process is illustrated in **Fig. 6**.

The master multiplexer performs the switching with a period $T_{sw.period}$. On every switching event in the master, a command 'SWnD...', as detailed in the command list, is transmitted to the first connected slave.

The slave multiplexer receives and decodes the command and waits the time, T_{delay}, specified in it, before performing the switching event for itself. On the switching event for the first slave the delayed switching command is retransmitted to the second slave, and so on.

As you may have noticed from the diagram, an interesting effect is created in this mode of operation. The circular mode of switching is shifted for every following slave device with the step T_{delay}. This can be used to create interesting visual effects when the multiplexer system is applied in multi-image carousel systems.

Creating interesting effects

The configuration used for my carousel multi-image system is shown in Fig. 3. The multiplexers connect through serial cables and the master device is connected to the control PC.

The same PC is used as a host for a multi-image PCI card known as the Zandar MVG6. This has six inputs so there are six multiplexers in the stack.

Video output from the card is applied to an RF modulator that feeds the cable network. Altogether, 48 video sources are used as inputs to the multiplexers. They are routed via eight video outputs on the rear panel of the multiplexer to the respective channel modulators of the cable system

It is assumed that the modulators represent a 75Ω load, because the multiplexer does not terminate the video sources, as mentioned earlier. Every group of video sources connected to a multiplexer is displayed on a fixed window on the output signal of the card. A possible configuration of the six video windows on the screen, representing the output of the system is shown on Fig. 3 too.

The number of the window corresponds to the number of the multiplexer. The arrow shows the sequential change effect in the pictures of the windows if the circular mode of operation with delay is implemented, as shown in Fig. 6.

In synchronous mode, all windows are changed simultaneously. All the parameters of the single windows – position on the screen, dimensions, etc., – and the text and appearance of the logo of the cable TV provider (or organisation) are configured through the application software of the MVG6 system and are not subject to our discussion.

Graphical user interface

My Windows 95/98 application lets you control the system from a friendly graphical environment, Fig. 7. I have given the control elements a number to aid this explanation.

Configuration field – 1. 'Manual Switch' lets you control the system manually. This means that you should select the appropriate connection and then press the corresponding button, 'Load' or 'Save StartUp' in the field, to enable the selected configuration.

All stacked multiplexers are affected by the manual mode performing identically.

'Load' – 2. This command button lets you load the selected configuration to the device.

'Save StartUp' – 3. This command button saves the selected connection as a StartUp one.

'Disable' – 4. The disable option disconnects – i.e. disables – the output of the multiplexer. The 'Load' button has to be pressed to activate the command.

'Circular Mode' – 5. Configures the two modes – circular mode (synchronous for all stacked devices) and circular mode with delay.

'Direct' – 6. Contains the command buttons for the synchronous switching (the same for all multiplexers) in circular mode.

Further information

MAXIM MAX440/MAX441 high-speed video multiplexer/amplifier data sheet. Rev. 1; 7/95. www.maxim-ic.com.

Atmel 8-bit microcontroller with 2Kbyte flash, AT89C2051 Data Sheet. www.atmel.com.

Atmel 2-wire serial EEPROM AT24C02, data sheet, www.atmel.com.

Atmel application note, 'Interfacing AT24CXX with AT89CX051 Microcontrollers', www.atmel.com.

Zandar Technologies Omni Video Series Multi-Video Window Display System MVG6, www.zandar.com.

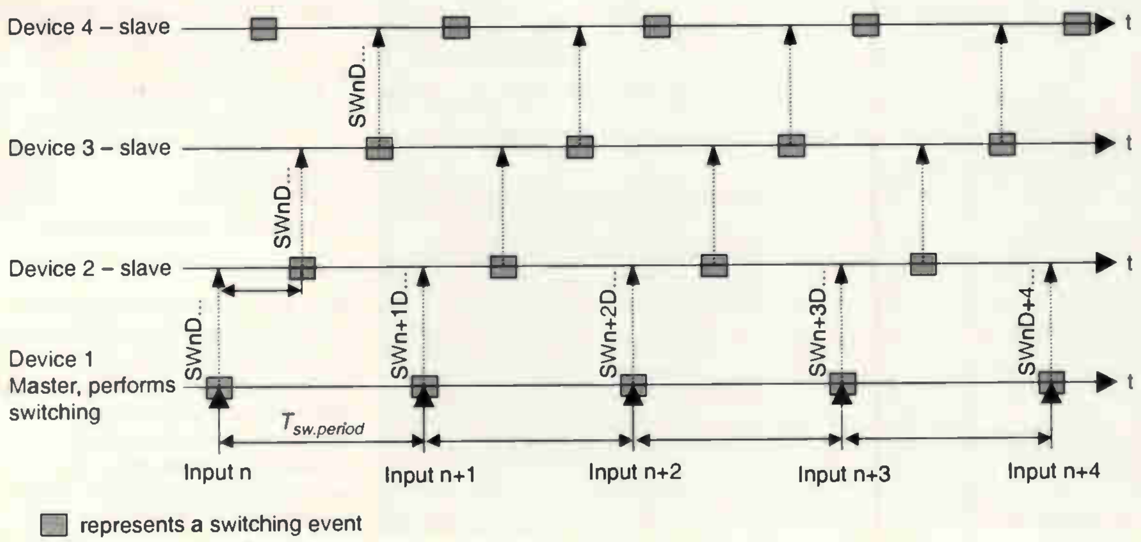

Fig. 6. When using multiple, stacked multiplexers in delayed mode, a switching command is transmitted to each slave sequentially with a delay between each transmit.

'with Delay' – 7. This field contains the corresponding buttons for the circular mode of operation with delay. There is a 'Delay' text box in this field. With it, you can specify the necessary delay in seconds, inserted between retransmission of the commands in the stack of devices, Fig. 6. The valid range of values for the delay are 0.1 seconds to the switching period minus 0.1 seconds.

'Preview – 8'. The 'Direct' and 'with Delay' subfields contain a 'Preview' button. The function of this button is to activate a simulation of the work of the stacked devices with the given parameters.

The simulation is performed on the graphical representation of six multiplexers – shown in area 11 on Fig. 7 – and is switched off by pressing the same button a second time. The preview does not affect the work of the multiplexers connected to the control PC. It is intended to give the user the possibility to optimise the system parameters.

Switching period – 9. The switching period in seconds can be entered by the user in the associated text box. Valid values are in the range 0.2 to 3276.0 seconds, but the selected value is always limited by the delay value from the lower side, which has to be greater by at least 0.1 seconds.

'to Input' – 10. Valid values for the start channel for circular mode are 1 to 7 and for the end channel 2 to 8, but the end channel has always to be greater in number than the start channel.

Preview – 11. The preview area shows six stacked multiplexers, every one with eight yellow LEDs, representing the active channel.

The user interface has the following menus:

- 'Communications' – the user can select the COM-port of the PC, connected to the system. The software supports COM1 to COM4.
- 'Lock' – the application software is locked and unlocked with this menu. When locked, the user can change parameters and perform previews, but can not download the selected configuration to the target device. So an accidental switching and possible errors and program drops are prevented.
- 'Language' – the application software supports 2 languages – English and my native language, Bulgarian. Of course if another foreign language is desired, the necessary changes can be made in the source code.
- 'SysTray' – the application is minimised as an icon in the system tray and can be activated every time the need arises. The source code for this option is created by E. Spencer (elliot@spnc.demon.co.uk) and is public domain.
- 'About' – Gives information about the current version of the software.

Fig. 8. Photos of the author's stacked multiplexers.

In summary

My six VMUX-8x1 multiplexers are shown in **Fig. 8.** I designed the system to perform well when working 24-hours a day, 7 days a week. This means that the hardware is reliable and the code for the microcontroller is well thought over to avoid locking and errors. In some cases it is written with redundancy to account for every unexpected situation.

The video multiplexer presented here has been in uninterrupted use since December 1999 by the greatest Bulgarian cable TV provider 'EVROCOM' – which caters for 500 000 households – and it works fine. ■

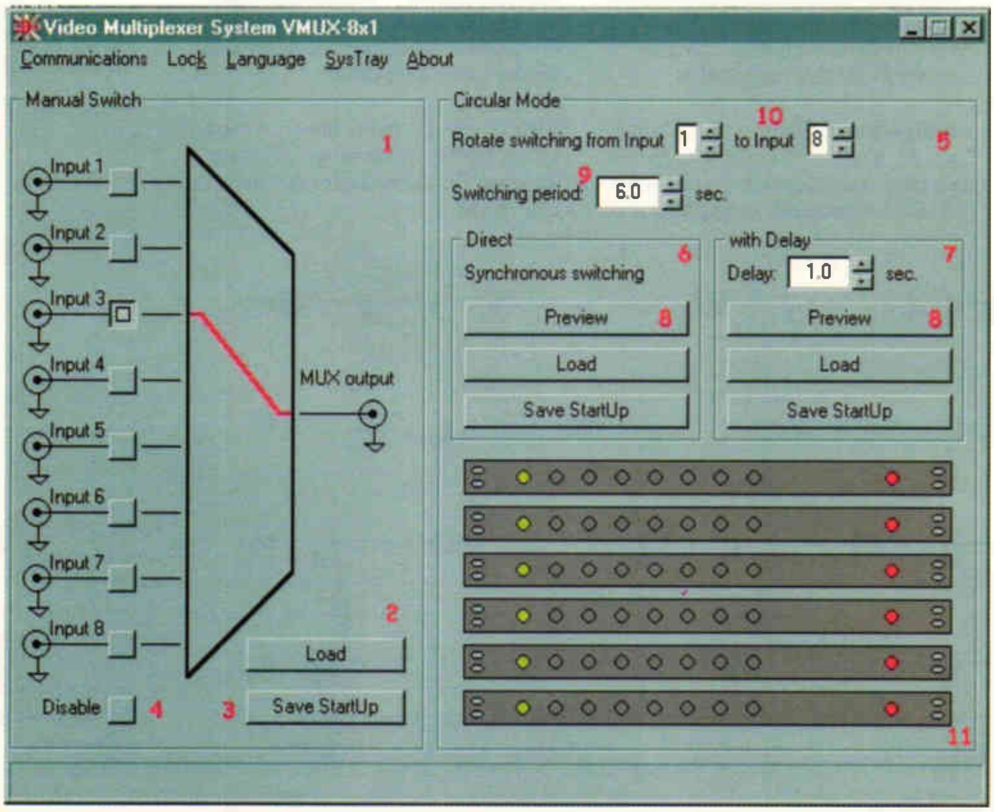

Fig. 7. You can write your own basic routines to control the multiplexer since it is controlled via simple ASCII commands down a an RS232 port. The author went a step further and produced a GUI, whose screen presentation is shown here. The numbers in red are applied to aid the description in the main text.

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

Total playing time 72.09

21 tracks – 72 minutes of music.
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One-stage receivers

Single-stage RF receivers have a number of advantages over their more complex sisters, conventional superheterodyne receivers and they save on costs. Richard Ball finds out if zero-IF is best for wireless.

Direct conversion, zero-IF, even superhomodyne – call it what you will – the single-stage RF transceiver architecture is flavour of the month. It has already appeared in set-top boxes and global positioning system (GPS) handsets, and is starting to be designed into other devices such as mobile phones.

Any wireless product, particularly battery powered ones, can benefit from direct conversion. Among the first products to do so was the humble pager. A traditional superheterodyne receiver's filters would have been the biggest components in terms of both size and cost in the small device.

Direct conversion receivers use quadrature mixing to go straight to baseband. The superheterodyne receiver converts the RF to one or more intermediate frequencies.

The major disadvantage of the superhet is that a mixer and filters are needed at each stage, increasing power consumption, component count and hence cost.

"Lots of applications have used superheterodyne and it's well understood, but it carries this extra filter and mixer," says Doug Grant, director of business development for RF and wireless systems at Analog Devices.

Direct conversion, with just one mixer, results in fewer filters and other discrete components. In fact in some applications, the whole receiver can be implemented on a single chip.

Overall up to 50 components could be lost from a typical mobile-phone design. This could result in cost savings beyond that just due to component count reduction. A single-chip on a motherboard is much cheaper than a network interface module or the tin can assemblies often seen today. It also reduces the effort required by board layout engineers – RF is the devil's own work when it comes to PCB design.

Bill of materials cost could be reduced by \$3 to \$4. It doesn't sound like much, but over a few hundred million phones it adds up.

Despite its advantages, direct conversion has some problems, including $1/f$ noise, time varying DC offsets, self mixing of strong signals, and a requirement for good anti-aliasing filters. Such issues, though, have not stopped the major semiconductor manufacturers from embracing direct conversion.

Analog Devices is one of the first firms to use the architecture in mobile phones with its Othello chip set. Two devices implement the direct conversion architecture and demodulator.

"We wanted to get the radio back onto a Moore's Law type of curve for cost reduction and integration," says Grant.

"Direct conversion's been around for a long time. This is the first time it's been applied to more complex modulation schemes." This includes tri-band mobile phones.

One of the problems encountered by Analog Devices was leakage of the voltage-controlled oscillator signal into the antenna. Because the VCO output is the same as the carrier, this can affect nearby handsets. The firm solved it by using a different frequency, around 1 350MHz, and dividing it or multiplying it to reach the required 900 or 1 800MHz at the mixer.

The problem of DC offset also affects the mobile-phone chip set. One solution is to AC couple the baseband signal, but this does not work when the signal has DC components, as in GSM.

Part of Analog Devices' solution is to carry out some clever processing of the baseband signal to remove the effects of the DC offset.

Analog Devices has now completed a reference design that integrates a complete GSM module on a single sided board half the size of a credit card. For data-only GPRS applications, such as in a PDA or PC Card with no keyboard and display interface, a two sided board is just one inch square.

By keeping the transmit and receive circuits switched off when not in use, standby time of a mobile phone can reach a staggering six weeks.

STMicroelectronics has begun using direct conversion in its set-top box and GPS chips.

From ST's point of view, the main problem area is phase noise. "We have to take care with the phase noise of the PLL," says Thierry Abraham, applications lab manager for satellite and terrestrial at ST. "This is the most critical point."

Perhaps ST's great strength is that it can produce the receiver in CMOS rather than BiCMOS. It has now started integrating the receiver with other parts of the tuner.

Its next generation set-top box chips will integrate the receiver, tuner and demodulator for digital video broadcasting. This brings its own set of problems, notably ensuring there is adequate isolation between the various blocks in the chip.

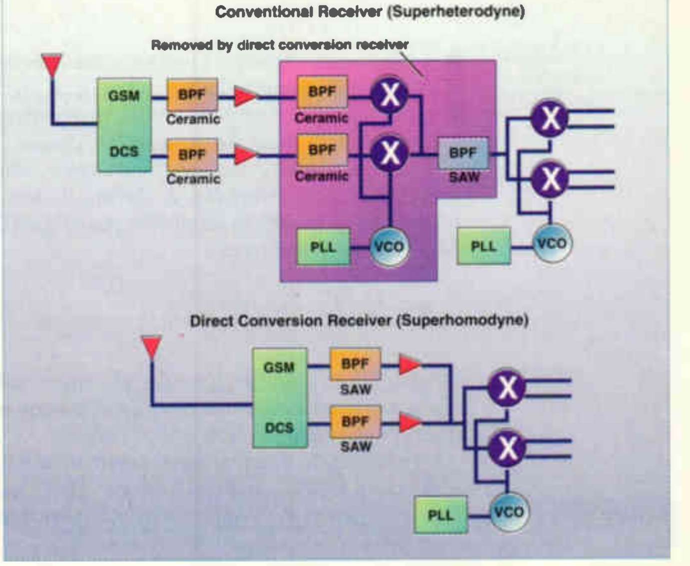
LSI Logic, Mitel and Philips are among other firms switching to direct conversion.

All three are keen to take a stake in the digital TV market, whether in set-top boxes, integrated digital TVs, or indeed both.

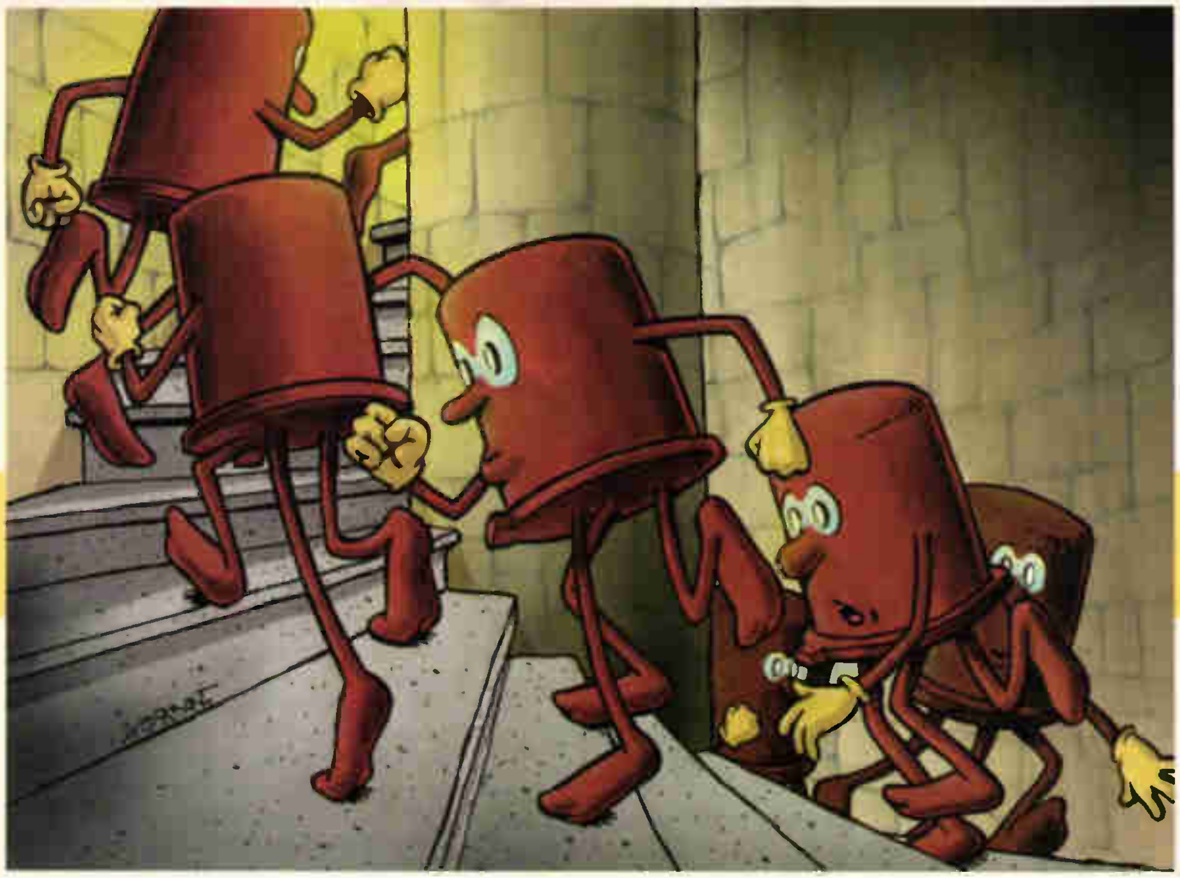
"We have a zero-IF tuner chip for satellite which has been successfully deployed," says Kenroy Francis, director of digital TV marketing at LSI. "What we're doing now is refining that even more by making it even more highly integrated."

Mitel Semiconductor has a single chip tuner for satellite TV using direct conversion.

"A lot of manufacturers are talking about it [a single chip], but not many are delivering the required phase noise and linearity," says Paul Fellows, marketing manager for digital TV technology. ■



Cut down architecture... These diagrams show a simplified version of Analog Devices' direct conversion architecture for mobile phones compared to the superheterodyne circuit. A substantial amount of circuitry and external components are saved – more than offsetting the extra complexity of direct conversion.



Better buffers II

In this second article looking at complementary compound emitter followers, Dave Kimber discusses short cuts for analysing the complementary feedback pair – including methods for determining distortion.

David P Kimber B.Sc.

In last month's article.[1] I looked at the complementary compound emitter follower, as in **Fig. 1**, and found that it can provide better performance than a plain emitter follower provided that the value of R is chosen correctly.

The critical factor is the transconductance, which varies with current as shown in **Fig. 2**. Having chosen a quiescent current that is enough to supply the peak signal current, then R can be found from,

$$R = \frac{25 \times \beta_2}{I_{c2}} \qquad (1)$$

Although β_2 is a poorly controlled parameter, this does not matter too much because in this region the transconductance is only changing slowly with current.

A variant of the complementary compound emitter follower, or CCEF, can provide some voltage gain, **Fig. 3**. This is known as the complementary feedback pair.

Feedback-pair operation

One can regard the feedback pair as a CCEF that has had its negative feedback reduced by the voltage divider R_1–R_2. The voltage gain is then approximately given by,

$$gain \approx \frac{R_1 + R_2}{R_2} = 1 + \frac{R_1}{R_2} \qquad (2)$$

However, the input transistor Tr_1 now has a non-negligible resistance in its emitter circuit, formed by R_1 and R_2 in parallel. This reduces transconductance; it is necessary to explore whether or not this reduction is significant.

Sometimes the feedback pair appears in the modified form shown in **Fig. 4**. Voltage gain determines the ratio of R_1 and R_2 while the DC conditions determine the effective resistance seen at the collector of Tr_2. If these two factors remain the same, then higher values for R_1 and R_2 will be needed than for the simpler circuit of Fig. 3.

Transistor Tr_1's emitter thus sees a higher resistance and so transconductance will be lower. The result will probably be higher distortion. This form of the circuit may have advantages if the feedback network is modified to include some frequency dependence. Otherwise it seems inferior to the simpler circuit. Let's stay with Fig. 3.

Evaluating feedback pair performance

The feedback pair has the same three operating regions as the CCEF. These regions – low, medium and high-current, were discussed last month.

One major difference is that in the low-current region – where Tr_2 is not conducting – the circuit degenerates into an attenuated emitter follower. As a result, it is even more important to avoid this region.

In the other two regions, the transconductance of Tr_1 is reduced by the parallel combination of R_1 and R_2 causing emitter degeneration. The position and size of the regions, in terms of current, will be nearly the same as for the CCEF because these are determined by what is happening in the base circuit of Tr_2 and this has not changed. A minor difference is that Tr_1 current does not directly contribute to the output.

So,

$$I_{c1(min)} = \frac{600}{R} \text{ (mA)} \qquad (3)$$

and,

$$I_{c2(m-h)} = \frac{25\times\beta_2}{R} \text{ (mA)} \qquad (4)$$

Operation in the medium-current region

In this region, the transconductance is that of Tr_2 multiplied by the voltage gain of the circuit around Tr_1. Current through Tr_1 will be only slightly higher than the low-medium transition current.

First find Tr_1's collector-circuit resistance $R_{c(Tr1)}$ and its emitter circuit resistance $R_{e(Tr1)}$.

$$R_{c(Tr1)} = \frac{R}{1+\frac{I_{c2}}{I_{c2(m-h)}}}$$

$$R_{e(Tr1)} = \frac{25}{\left(\frac{600}{R}\right)} + R_1 \| R_2$$

Transconductance is now,

$$g_m = 40\times I_{c2}\times\frac{R_{c(Tr1)}}{R_{e(Tr1)}} \text{ (mA/V)} \qquad (5)$$

Here, $R_1 \| R_2$ is the parallel combination of R_1 and R_2. This resistance will almost always be much larger than $R/24$ so some simplification in the transconductance equation can be carried out by making.

$$R_{c(Tr1)} = \frac{R}{1+\frac{I_{c2}}{I_{c2(m-h)}}}$$

$$R_{e(Tr1)} = R_1 \| R_2$$

$$g_m = 40\times I_{c2}\times\left(\frac{R}{R_1 \| R_2}\right)\div\left(1+\frac{Ic_2}{Ic_{m-h}}\right) \text{ (mA/V)} \qquad (6)$$

The transconductance rises linearly at first, then levels off as the current approaches the medium-high transition.

Design example

As for the CCEF design example in the previous article, I will take R as 1kΩ and assume that β_2 is 240. Then $I_{c1(min)}$ is 0.6mA, and $I_{c2(m-h)}$ is 6.0mA. Also assume that $R_1=R_2=1$kΩ, i.e. a voltage gain of approximately 2, and that I_{c2} is 2mA.

Quiescent output voltage is now 4.6V and an input bias of around 3.2V is needed. Resistance $R_1 \| R_2$ is 500Ω. Also, $R/24$ is 41.7Ω so the simplification in equation (6) gives an error of about 8%, which is acceptable.

$$R_{c(Tr1)} = \frac{1000}{1+\frac{2}{6}} = 750$$

$$R_{e(Tr1)} = 1000 \| 1000 = 500$$

so now,

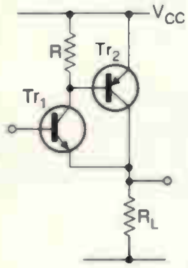
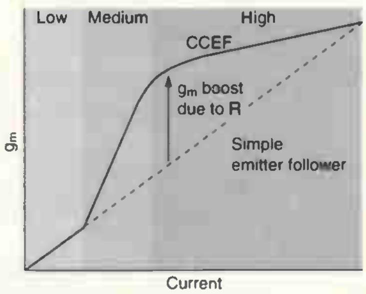


Fig. 1. The complementary compound emitter follower can provide better performance than a plain emitter follower – provided that the value of R is chosen correctly.

Fig. 2. When working out the performance of a complementary compound emitter follower, the critical factor is transconductance, which varies with current.



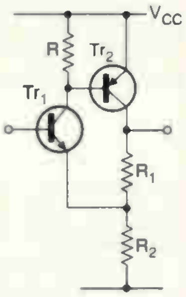
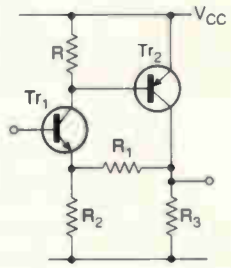


Fig. 3. The complementary feedback pair is a variant of the complementary compound emitter follower that provides a degree of voltage gain.

Fig. 4. Modified complementary feedback pair. This circuit may have advantages if the feedback network is modified to include frequency dependence, but otherwise it appears inferior to its simpler counterpart.



$$g_{m(2mA)} = 40\times 2\frac{R_{c(Tr1)}}{R_{e(Tr1)}} = 120 \text{ (mA/V)}$$

This is much lower than for a similar CCEF, which would have a g_m of 1464mA/V.

Evaluating distortion. To work out the signal distortion, assume a 2V output – i.e. 1V input – which will cause the current to swing from 0.59mA to 3.41mA.

$$g_{m-(0.59mA)} = 43\text{mA/V}$$

$$g_{m+(3.41mA)} = 174\text{mA/V}$$

Second-harmonic distortion, from g_m, is,

$$= \frac{1}{4}\times\frac{(g_{m+}-g_{m-})}{(g_{m+}+g_{m-})} = 0.15$$

Distortion of the circuit is,

$$= \frac{0.15}{\textit{feedback factor}} = \frac{0.15}{\left(\frac{1}{2}\times 120\times 2\right)} = 0.125\%$$

Here the feedback factor consists of the negative feedback division, namely 1/2, the g_m of 120mA/V and the output resistance of 2kΩ.

The distortion is higher than for an emitter follower producing the same signal as you will see from the previous article. It is also much higher than that of a CCEF. This is mainly because of the lower transconductance, but there is also a factor of 2 from the reduced negative feedback.

High-current region

The transconductance in this region comes from the transconductance of Tr_1 plus its emitter degeneration multiplied by the current gain of Tr_2. As for the CCEF, you need to include a correction for the signal current stolen from Tr_2 base by R.

$$g_m \approx \beta_2 \times \frac{1000}{\frac{R}{24}+R_1\|R_2} \div \left(1+\frac{I_{c2(m-h)}}{I_{c2}}\right) \text{ (mA/V)} \qquad (7)$$

As before, assume that $R_1\|R_2$ will be sufficiently large that this will dominate transconductance of Tr_1. Then,

$$g_m \approx \frac{\beta_2}{R_1\|R_2 \times \left(1+\frac{I_{c2(m-h)}}{I_{c2}}\right)} \text{ (mA/V)} \qquad (8)$$

where $R_1\|R_2$ is now in kilo-ohms. This looks quite different from equation (6) for the medium current region, but they are in fact equivalent so either can be used. Equation (8) shows that at high currents, the transconductance will approach an asymptote of $\beta_2/R_1\|R_2$.

Is *R* really necessary?

Initially, it seemed as though the complementary feedback pair could be considered as a variant of the CCEF. Now it is clear that there is a significant difference between the two.

For the CCEF, adding R boosts transconductance in the middle current region by increasing the current in Tr_1; this more than compensates for the loading effect of R. For the feedback pair Tr_1 transconductance is dominated by $R_1\|R_2$, so R appears to simply add loading and so reduce overall g_m at lower currents. Can we simply get rid of R?

The answer is, probably not! This is because the simplifying assumption that the g_m of Tr_1 is dominated by $R_1\|R_2$ is only true when Tr_1 has sufficient current. This only occurs when either R is present, and not too large, or β_2 is quite small.

So what is an optimum value for R? To find out, rewrite equation (7) in a way that makes its dependence on R explicit:

$$g_m \approx \beta_2 \times \left(\frac{1000}{\frac{R}{24}+R_1\|R_2}\right) \div \left(1+\frac{25\times\frac{\beta_2}{R}}{I_{c2}}\right)$$

$$= \beta_2 \times 1000 \div \left(\left(\frac{R}{24}+R_1\|R_2\right)\times\left(1+\frac{25\times\beta_2}{R\times I_{c2}}\right)\right)$$

$$= \beta_2 \times 1000 \div \left(\left(R_1\|R_2+\frac{25\times\beta_2}{24\times I_{c2}}\right)+\frac{R}{24}+\frac{25\times\beta_2\times R_1\|R_2}{R\times I_{c2}}\right)$$

For maximum g_m, you need to minimise the messy denominator i.e. we need to minimise,

$$denom = a + bR + \frac{c}{R}$$

where,

$$a = R_1\|R_2 + \frac{25\times\beta_2}{24\times I_{c2}}$$

$$b = \frac{1}{24}$$

$$c = 25\times\beta_2\times\frac{R_1\|R_2}{I_{c2}}$$

Differentiating *denom* gives,

$$b - \frac{c}{R^2}$$

which will be zero when,

$$R = \sqrt{\frac{c}{b}}$$

i.e.,

$$R = \sqrt{600\times\beta_2\times\frac{R_1\|R_2}{I_{c2}}} \qquad (9)$$

So for the above example, where R_1=R_2=1kΩ, β_2=240 and I_{c2}=2mA you will find that the optimum value for R is 6kΩ. Given this optimum value for R, the denominator simplifies slightly:

$$denom = a + bR + \frac{c}{R}$$

$$= a + b\times\sqrt{\frac{c}{b}} + \frac{c}{\sqrt{\frac{c}{b}}}$$

$$= a + 2\times\sqrt{b\times c}$$

$$= R_1\|R_2 + \frac{25}{24}\times\frac{\beta_2}{I_{c2}} + 2\times\sqrt{\frac{25}{24}\times\frac{\beta_2}{I_{c2}}\times R_1\|R_2} \qquad (10)$$

Now transconductance is 213mA/V – almost double the figure when R was 1kΩ. Note that you still have to slog through the full messy version when calculating g_m for other currents – for example at signal peaks – because equation 10 only applies when equation 9 is true, i.e. the quiescent state.

Has the distortion figure changed?

It is now possible to recalculate the distortion of the example circuit with R at 6kΩ instead of 1kΩ.

$$I_{c2(m-h)} = \frac{25\times 40}{6000} = 1\text{mA}$$

$$g_{m-(0.59mA)} = 119\text{mA/V}$$

$$g_{m+(3.41mA)} = 247\text{mA/V}$$

$$\text{Distortion (2nd, from } g_m) = 0.087$$

$$\text{Distortion (circuit)} = 0.041\%$$

This is about a third of the previous value, so optimising R has been worthwhile. The improvement comes from a higher transconductance, and less variation with current.

If R is removed altogether then the distortion becomes 0.22%, which is worse than the figure for the sub-optimal value of R=1kΩ.

A rule of thumb

As for the CCEF, the best value for R depends on β_2 which is poorly controlled. But the variation of transconductance with R is quite shallow in this region.

If R is chosen for a central value of β_2, then a factor of two in either direction should not hurt too much. Provided the value of R is near enough correct, then this will still be better than leaving it out altogether.

Equation 9 suggests that it might be possible to find a simple 'rule of thumb' formula for R that will be good enough for most non-critical applications. For a small-signal transistor, beta is usually somewhere in the range 100-500, so assume a value of 200.

Quiescent output voltage will be around 5V, so I_{c2} is $5/(R_1+R_2)$. You can now put these values into equation 9 – taking care with the units!

$$R=\sqrt{600\times 200\frac{R_1\times R_2}{R_1+R_2}\div\frac{5000}{R_1+R_2}}$$

$$=\sqrt{600\times 200\times\frac{R_1\times R_2}{5000}}$$

$$=\sqrt{24}\times\sqrt{R_1\times R_2}$$

but $\sqrt{(R_1 \times R_2)}$ is the geometric mean of the two resistor values. For small ratios, which are likely here, we can approximate the geometric mean as 0.8 times the arithmetic mean. So:

$$R\approx\sqrt{24}\times 0.8\times\frac{R_1+R_2}{2}\approx 2\times(R_1+R_2) \qquad (11)$$

So for the example circuit, this gives a value of 4kΩ for R, which is quite close to the correct value of 6kΩ.

Output impedance

Although the complementary feedback pair provides voltage gain, you have seen that its transconductance, and hence open-loop gain, is much lower than for a similar CCEF. This results in a higher output impedance, which is further increased by the lower overall negative feedback.

$$R_{out}=\frac{voltage\ gain}{transconductance} \qquad (12)$$

So for the example circuit, with R optimal at 6kΩ,

$$R_{out}=\frac{2}{213}=9.4\Omega$$

This is not too bad, but one well-respected textbook gives an example feedback pair with an output impedance of 500Ω! This arose partly from a higher voltage gain of 10, but mainly from a rather low value of 620Ω for R which is slightly worse in transconductance terms than omitting R altogether.

In a third article, I will be looking at using a pair of CCEF as a Class B output stage. ■

In last month's article, the figure 2400 in equation 15 should have read 24 000. Also, there was a small error in Fig. 2. The emitter of the Tr_1 should have been wired as in Fig. 3 of this article. Apologies.

Reference

1. 'Better buffers,' Part 1 – Introduction and Class A, *Electronics World*, September 2000, p.725.

NEW PRODUCTS

Please quote *Electronics World* when seeking further information

Power amplifier for CDMA

Agilent Technologies has announced that Samsung Electronics has selected a version of its single-band dual-mode power-amplifier module for its SCH-850 CDMA and AMPS phone. This pocket phone will serve US markets through service provider AirTouch. The amplifier module helps the phone achieve a battery life of up to 130hr on standby and 4.7hr of talk using the 1600mAh battery. Samsung will also use a commercially available Agilent module, the QCPM-9801 dual-band, dual-mode power-amplifier, in its soon-to-be-introduced SCH-8500 CDMA flip phone. It provides two power amplifiers in one package.
Agilent Technologies
Tel: 00 49 6441 92 460

Intelligent GSM network tool

Autonomy – an intelligent GSM network performance and optimisation tool, which eliminates the need for skilled personnel in the field – has been introduced by Freshfield Communications. It lets network operators remotely access network performance from a subscriber's viewpoint through a central data storage system incorporating user data evaluation and analysis software. It evaluates network problems, performs network quality audits, validates the performance of new cells within a GSM network and compares competitive networks. For vehicle mounting – trains, buses, taxis, vans and so on – the Roam data acquisition hardware that forms part of the package uploads data to a centralised database, which can be accessed from a control centre by multiple users. Up to three networks can be logged simultaneously.
Freshfield Communication
Tel: 0208 408 8110

PC/104 power supply with flash

The Micronix PV-5124 power supply module from Micro Technic has either a 0, 2 or 4Mbyte flash solid state disk drive (SDD), which emulates a standard flash disk drive (FDD), providing a way to supply power and disk capacity to PC/104 bus products. The UPS function suits remote battery-operated systems with its CPU controlled and automatic shutdown for battery discharge protection. It is a DC to DC converter with 30W output capability. It is for embedded vehicle and remote battery operated systems. Temperature range is –20 to +70°C. Power supply output is +5 and ±12V. SSD seek time is less than 1ms and efficiency is up to 85 per cent, claims the supplier. It has a 16 to 30V DC input range, low quiescent current and measures 95 by 90mm.
Micro Technic
Tel: 00 45 66 15 3000

Async interface analyser

DB Broadcast has introduced the B051 battery-powered handheld device that analyses DVB asynchronous serial interface (ASI) transport streams. It provides real-time indications on the state of the ASI data link, and carries out basic checks on the DVB 270Mbyte ASI transport stream. It measures 157 by 70 by 210mm and weighs 1kg.
DB Broadcast
Tel: 01353 661117

Wire-to-board SM connectors

Flint has introduced two 2.0mm pitch surface-mount connectors from JST. Crimp and IDC models are available for top and side-entry headers, with both receptacles being compatible

Fibre-optic chip-set multiplexes 16 independent T1/E1

Semtech has announced a fibre-optic access chip-set that can multiplex up to 16 independent T1 or E1 serial data channels over a fibre-optic cable. The ACS411 has an aggregate bandwidth up to T3, E3 and OC-1 rates (51.84Mbit/s) on a single channel reaching up to 90km without repeaters. It is for long-haul broadband access applications in cellular phone networks when base-station transceivers, controllers and switching centres may be at any distance from each other. It operates over twin fibre cables or a single fibre cable with WDM. The chip-set comprises the ACS4110 mixed-signal device that provides 16-channel mux and demux logic, clock and data recovery and two ACS9020 analogue transceivers, which include a transmitter with laser and LED drivers and a PIN receiver with a TIA and post amplifier.
Semtech
Tel: 01592 773520

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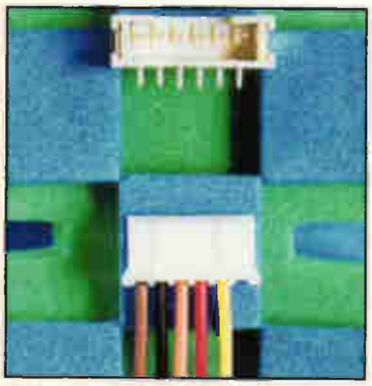


with either header type. The PH crimp style has top or side-entry connections with two to 14 circuits respectively. It is 7.5mm high and 4.5mm wide. Wire-strip holding length is 2.6mm. Crimp contacts are for 32 to 24 AWG wire, rated at 100V, 2A AC or DC for 24 AWG. Connector housings are made from heat-resistant resin compatible with reflow soldering. The solder side of the header has a retention mechanism to prevent floating during soldering.
Flint Distribution
Tel: 01530 510333

Smartcard connector

For automotive and consumer applications, the V02 smartcard connector from FCI has a height of 3.2mm, including a stainless steel cover. The connector accepts cards with thicknesses from 0.68 to 0.84mm and can be mounted into a through-hole PCB, reinforced by four metal latches on the cover. The cover prevents static discharge and provides mechanical protection. The single piece, moulded body has an integral, double-action blade switch. The normally open switch detects insertion and extraction of the smartcard and has a 2.5μm wiping action for self cleaning. Insertion and extraction forces are 7 and 5.3N respectively. It meets ISO7816 and GSM1111. Copper alloy contacts are plated with gold over nickel in the contact area and the maximum contact resistance is less than 200mΩ. Automotive applications include dashboard GSM telephones, GPS equipment, ignition security, electronic road tolls and tachographs. It operates from –20 to +70°C.
FCI
Tel: 01582 814800

Fibre optic connector

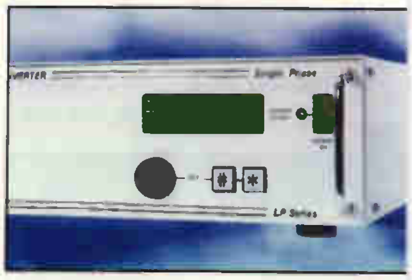
Amphenol has introduced hermaphrodite fibre-optic connectors for harsh environments. Hermaphrodite coupling allows for daisy chaining, which eliminates the need for polarising the assemblies and the use of in-line adapters. Markets include military, energy, marine, rail, and oil and gas exploration and mining.
Amphenol
Tel: 01227 77320

Rail-to rail quad op-amp

Maxim has introduced the Max4403 quad, rail-to-rail op-amp in a 14-pin SOIC or TSSOP package. This device consumes 320μA supply current per amplifier, while achieving an 800kHz gain-bandwidth product. Operating from a 2.5 to 5.5V supply, it is suitable for portable and battery-powered applications. The output architecture achieves an open-loop gain of 110dB and can drive a 2kΩ load to within 55mV of both supply rails. Total harmonic distortion is 0.009 per cent and it is unity-gain stable with capacitive loads up to 400pF. Temperature range is –40 to +125°C.
Maxim Integrated Products
Tel: 0118 930 3388

Frequency converters

From Magnus Power are three microprocessor-controlled frequency converters, the LP300, LP600 and LP1000, with outputs of 300, 600 and 1000VA respectively. Each of these single-phase output units has its output frequency variable from 45 to 440Hz in 1Hz steps. Output voltage is variable from zero to maximum in two ranges, usually 0 to 135V and 0 to 270V, with output current capability doubled on the lower of these ranges. The output is a clean, conditioned pure sinewave. There is automatic selection of input voltage. The output is controlled by one multi-turn knob and two pushbuttons. These are operated with a two-line alphanumeric display. The chosen output is microprocessor-stored when the converter is shut down so the setting is immediately available the next time it is switched on.



Smart label identification

RFID Components has introduced the Discovery kit for prototyping and evaluating smart label RFID identification systems based on Texas Instruments' Tag-It smart label technology. The kit lets users evaluate performance for baggage handling, product and parcel tracking, and warehouse and logistics tracking. Other applications include plastic access control cards, brand protection for high value products vulnerable to counterfeiting, container tracking and tracking important files or documents in finance, medical and legal offices. The kit comprises a Tag-It Series 6000 reader, housed antenna, smart label inlays, power supply, serial PC communications cable, Navigator software and manuals. With the software and PC connection, users can read and write data to the label. The 13.56MHz inlays can be embedded into flexible labels compatible with print-on-demand thermal printers. The transponders can be read electronically without any line of sight at any orientation and inside packaging. Multiple labels can be read simultaneously.
RFID Components
Tel: 01234 840102

If fixed output parameters are required, for example 115V 400Hz, a version can be supplied with a preset output.
Magnus Power
Tel: 01424 853464

WLAN chip-set

Intersil has announced an enhancement to its Prism II wireless LAN chip-set. Called Prism II.v, the chip-set reduces the number of chips from five to four and has a USB interface option in addition to its PCMCIA interface.
Intersil
Tel: 001 321 729 4928

PCB-to-fibre connector

The blind-mate backplane LC connector from Molex joins active devices terminated on a PCB to fibre-optic components on a backplane. It has an RJ45 latching interface. The blind-mate plug-in interface allows insertion and removal of the daughter card without disrupting I/O ports and associated cabling. Axial float lets the connectors self-align inside the adapter. Mechanical latching provides an isolated connection to the daughter card. Doors on each port stay closed when not in use to prevent dust and laser hazards. It is available with two, four and eight ports.
Molex
Tel: 01252 720720

60V MOSFET for automotives

A power MOSFET from Toshiba with active clamping provides stable operation in ABS and other safety-critical applications over the automotive temperature range. Based on 60V trench gate process technology, the 2SK3311 has an on-resistance of 15mΩ and a

maximum leakage current of 1μA. Built-in Zener diodes provide active clamping and protection respectively. It comes in a TO-220AB package and has temperature clamped voltage ratings from 33 to 43V.
Toshiba Electronics
Tel: 00 49 211 52960

Sealed connectors

Deutsch's DT sealed connectors are for cable-to-cable applications on engines, transmissions, chassis or in the cab. Thermoplastic housing withstands operational knocks and they are protected by silicone seals. No special tools are required for installation and there's a click and visual alignment indicator for the assembler. All contacts are rated to 13A and there is the option of two, three, four, six, eight or 12 pins. The DT13 and 15 PCB headers mate with DT06 plugs for heavy duty

Single-slot PCI radar input card

Primagraphics has introduced the Virgo PCI board, a single-slot PCI radar input card for its Metro PC radar acquisition and display systems. The board allows the connection of various radar types to a standard PCI workstation. These include ACP, ARP, Synchro, ASDE-3, parallel, phased array and the US Navy's Radds radar display and distribution system. The card provides video and azimuth processing, including a 2D video amplifier parameter table that compensates for radar characteristics and clutter. It has three input channels and a 32-bit PC interface for output data. Proprietary or non-standard azimuth formats can be supported via a custom plug-in module. Radar input gain and offset are programmable functions of range and azimuth. The signal is sampled at 50MHz to 8-bit precision. The samples are mixed with the four thresholded data inputs and remapped to 8-bit by a programmable lookup table. The 8-bit data can be sub-sampled using peak detection to produce a user-specified sampling rate, or an external clock can be used.
Primagraphics
Tel: 01763 952222

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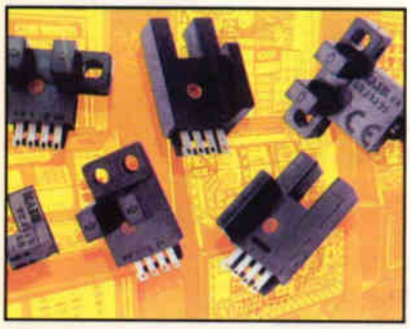
electronic sensor and small module applications.
Deutsch Industrial
Tel: 01424 852722

Slot sensors

UZJ3 photoelectric slot sensors from Matsushita come in sizes down to 10.5 by 13.4 by 12mm.



There are 17 configurations or shapes and 29 types including cabled and connector versions. All have two independent transistor outputs for light on or dark on operation and the response time is 20μs – version specific.
Matsushita Electric Works
Tel: 01424 852722

ESD protectors

Circuit protection devices using voltage-variable material (VVM) to protect electronic equipment against electrostatic discharge are available from CPS. Made by Littelfuse, the Pulseguard range includes connector array modules and surface-mount devices. Each device is based on a polymeric VVM that becomes electrically conductive when subjected to high voltages. This lets the device conduct away ESD spikes to ground while remaining electrically transparent to the lower signal voltages encountered during normal operation. The range consists of array modules to fit the connector sockets in PCs and peripheral equipment, such as printers, router hubs and scanners, with surface-mount devices for PCB mounting in digital communications equipment. The connector arrays are for use with standard nine, 15, 25, 37 and 50-pin connector sockets and are for press-in fitting with no soldering.
CPS
Tel: 01252 816192

Low dropout regulators

Three low-dropout voltage regulators from Rohm deliver stable typical output voltages down to 1.5V and maximum output currents up to 1A from a TO252 package with built-in protection. For hard-disk drives and DVD equipment, the BA25BC0FP and BA18BC0FP accept 3V inputs and provide stable output voltages of 2.5 and 1.8V respectively. The BA00BC0WFP also accepts a 3V input, but is a variable output version that can be configured to provide stable output voltages between 1.5 and 15V. A switch lets the device be turned on and off. Output accuracy for all three is ±2 per cent.
Rohm Electronics
Tel: 01908 282666

CCD board camera

A black and white CCD board camera is available from Premier Electronics. It uses a 0.85cm interline type CCD solid state image sensor, up to 437,664 pixels and CCIR or EIA interface. Measuring 42 by 42mm, the VPC-4651495 has auto iris outputs and a 2:1 interlace scanning system. Features include internal and external synchronisation and a horizontal resolution up to 570 TV lines. Sensitivity is down to 0.2lux and signal-to-noise ratio is over 50dB. Full screen BLC is provided with an electronic iris, which ranges from 1/50 to 1/100,000s. It has on-off selectable AGC, weighs 25g and provides lens options of 2.5 to 25mm. A glass lens is fitted. Options include C-mount, CS-mount, D-mount, auto-iris, pinhole lens and high or low resolution. Applications are from large military operations to undercover surveillance.
Premier Electronics
Tel: 01992 634652

Multiband software radio

The Series 4400 M3SR is a Rohde & Schwarz multiband, multimode and multifunction surface radio for navy, air force and tactical applications. The

radio is suitable for voice and data transmission in fixed frequency or EPM (ECCM) mode. It will operate with Nato and non-Nato communication standards such as Saturn and Secos. Incorporated is a reprogrammable software radio to increase flexibility. Changes in waveforms, modulation, interfaces or frequency bands can be done by downloading software or using already stored software. The hardware radio modules have plug-and-play functionality. Additional communication lines of up to four per radio can be set up without hardware alignments.
Rohde & Schwarz
Tel: 01252 811377

2.5V micropower LDO reference

The LT1790 is a micropower, 2.5V, SOT-23 series reference with an initial accuracy of ±0.05 per cent maximum and a temperature coefficient of 10ppm/°C maximum. The 0.1V maximum dropout voltage and 20V maximum input voltage lets it be directly powered by various supplies from a lithium ion battery to a wall adapter. The 60μA maximum supply

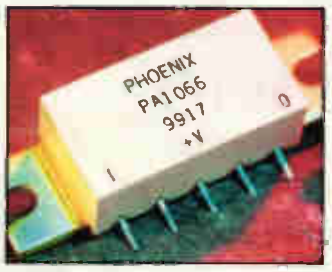
MPEG transport-stream generator for multimedia

An MPEG transport stream generator from Yokogawa is for multimedia and digital TV test equipment. The VT3000 provides reusable MPEG streams for testing signal-to-air interfaces and set-top boxes. The instrument can capture, record and replay any transport stream data. Transport streams can be captured on an internal hard disk via a DVB parallel port, and up to three independent channels of data can be played back simultaneously. Each channel is provided with a clock for multiplexing tests. It can be operated over a TCP/IP network (10/100baseT), so transport-stream files can be sent to a remote unit for playback over the air or captured remotely for local analysis.
Yokogawa Martron
Tel: 01494 459200

current suits portable or battery-powered applications. Noise is 12µVpp from 0.1 to 10Hz and 33µV rms from 10Hz to 1kHz. Applications include wireless base-stations and precision handheld instruments. Hysteresis is 40ppm typical and operating range –40 to +125°C. Long-term drift is low.
Linear Technology
Tel: 01276 677676

Linear RF amplifier modules to 4W

Linear amplifier modules have been launched by Phoenix Microwave for cellular base-stations, microcells and wireless LANs. Available from Sematron, the LPA models provide output power levels up to 4W. Frequency ranges are 800 to 960, 1800 to 2000 and 2000 to



2200MHz, which will shortly be extended to 2500MHz. Input and output impedances are 50Ω. Units can be cascaded and are complete amplifiers that include DC bias, RF blocking and bypass capacitors. They have either one or two stages of amplification using medium to high power GaAs FETs. Third order intercept point is up to 20dB above the 1dB compression point.
Sematron
Tel: 01256 812222

SMPS 1GBT

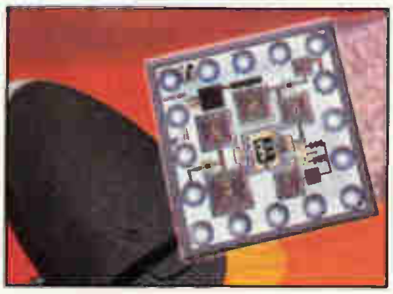
Intersil has introduced a switch mode power supply IGBT for the 24 x 7 power management demands of internet infrastructure. The 600V transistor is a MOS gated switching device combining the high input impedance of a MOSFET and the low on state conduction loss of a bipolar transistor.
Intersil
Tel: 001 407 729 5973

Real-time debug

Noral Micrologics has launched a background debug mode tool that can debug in real-time embedded applications based on Motorola's OSEK/VDX RTOS. For use with Motorola 68HC12 processors, the Flex-BDM/68HC12/OSEK debugger is for OSEK-based automotive designs and can be used instead of in-circuit emulation techniques.
Noral Micrologics
Tel: 01254 295800

Miniature low-noise amplifier

Intarsia has introduced an amplifier module for wireless and RF applications. The module was developed using



BOOK TO BUY

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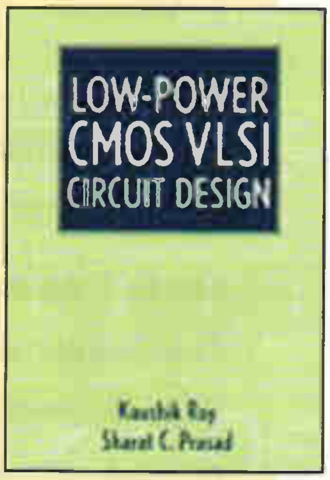
A comprehensive look at the rapidly growing field of low-power VLSI design



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Low-power VLSI circuit design is a dynamic research area driven by the growing reliance on battery-powered portable computing and wireless communications products. In addition, it has become critical to the continued progress of high-performance and reliable microelectronic systems. This self-contained volume clearly introduces each topic, incorporates dozens of illustrations, and concludes chapters with summaries and references. VLSI circuit and CAD engineers as well as researchers in universities and industry will find ample information on tools and techniques for design and optimisation of low-power electronic systems.

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NEW PRODUCTS

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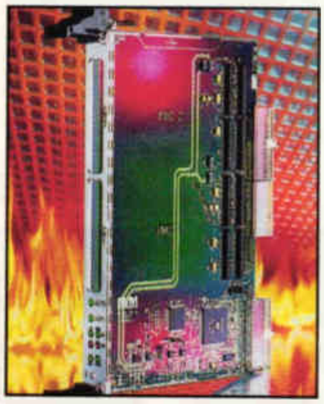
the firm's Passport design palette for thin-film-on-glass microcircuits. The Phemt pseudomorphic high-electron-mobility transistor module comes in a 4 by 4 by 1.2mm surface-mount direct-module-attach package that includes biasing and matching circuits. Noise is 1.1 to 0.8dB from 1.5 to 2.7GHz.
Intarsia
Tel: 01608 819161

Development tools that switch

Hitex in-circuit emulators can be operated with Panta – an intuitive development framework from Hiware. Before, users of Hiware tools had to change the user interface when switching from simulation



to test with an in-circuit emulator. This is no longer necessary, since the real-time debugger component of Panta co-operates with Hitop, the native debugger of the Hitex emulators. Although the user exchanges simulation on a host with real-time emulation on the target system, the handling remains the same.
Hiware
Tel: 024 7669 2066

Control rely for automotive uses

A motor control relay for automotive use has been launched by Omron for applications such as electric windows, central locking, intermittent wiper control, flasher units and electric sunroofs. Measuring 14mm long, 7.2mm wide and 13.3mm high, the G8N relay weighs 4.1g. The 12V DC relay can switch up to 30A over a range of –40 to +80°C, and withstand inrush currents up to 70A. As well as the standard relay, there are three versions for flasher operation, high sensitivity applications and high temperature environments to 105°C.
Omron
Tel: 0208 450 4646

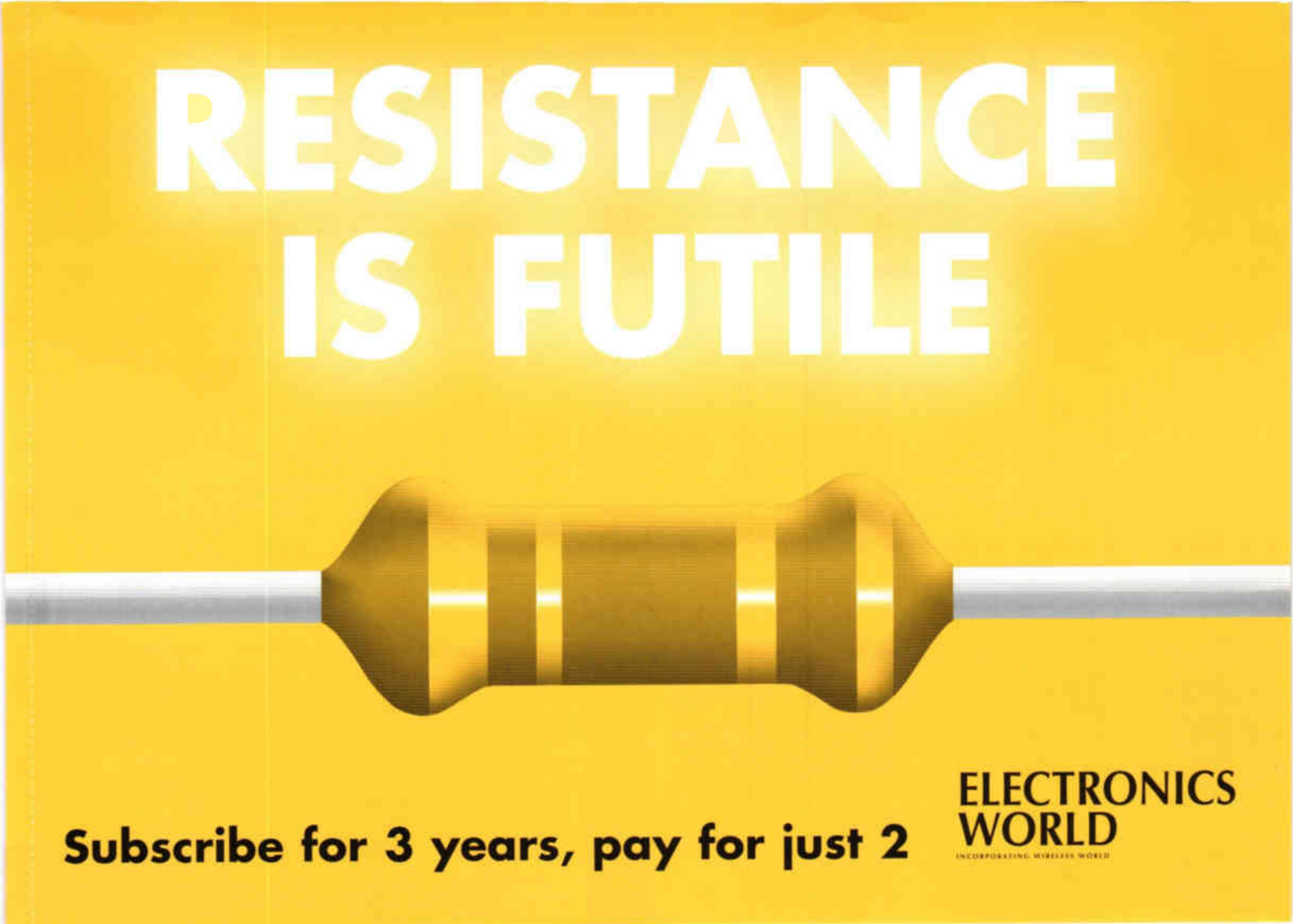
Nickel-coated spring wire

Sandvik Steel has launched a nickel coated spring wire. The Gusab Nicoat A wire is for use in the electronic industries in Europe, the US and south east Asia. It is available with a coated or a bright surface finish. The coated surface consists of a lubrication layer made of mainly sodium and calcium compounds and almost free from chlorides. Any coating remaining on the finished spring can be removed using alkaline agents.
Sandvik Steel
Tel: 01922 728800

Power MOSFETs in FlipFET packs

International Rectifier has introduced HEXFET power MOSFETs in packages in which all the terminals are on one side of the die. The FlipFET package is aimed at

portable applications. The first two products using this technology are the 20V p-channel IRF6100 single and IRF6150 bidirectional dual. The IRF6100 measures 1.52mm^2, and the IRF6150 3.05mm^2. ■
International Rectifier
Tel: 0208 938 4620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Letters to the editor

In the dark ages of reliability

"It's crashed again!" That phrase is synonymous with PC users all over the world. Unstable operating system, incompatible hardware leading to lockups, badly behaved applications, which crash your system.

The symptom is the same. your work comes screeching to a halt.

Perhaps designers in a large organisation will have an IT/IS person to sort their problems out. The rest of us, on the other hand have to battle on with this millstone of unreliability.

How is it that we have all been duped into allowing such unreliability back into our labs, offices and homes? I remember the early days of electronics being like this. Tube (valve) tellies and hi-fis broke down a couple of times in their expected lifetime and there was a chain of repair shops to solve the problems.

Now, with the MTBF of semiconductor electronics in excess of the expected life of a product, and with the production techniques employed, there is usually no need – and often no way – to service a product. This has spelt the death of the electronics service industry. This is a shame. It was a great place for a start in electronics.

Now, as a design engineer, I find the PC has become an essential tool. CAD, modelling, documentation and programming are all excellent applications of a PC in the electronics design process– that is, until the next crash or lockup.

As I see it, these are tools. Just that: tools. A drawing board is seen as old, difficult to use and messy, but seen in the light of reliability, it could be considered equal.

With a drawing board you don't have some unknown force that erases the last 10 minutes of your work when you're not looking, or insists on printing your drawing with the last page mysteriously lost.

As for PC based instrumentation, when has any of us had to re-start our Tek oscilloscopes just when making some important measurement? Or not been able to turn off a signal gen without losing or corrupting its settings?

My signal generators and scopes have always been 'there' and worked when I needed them. If anything, some stopped working due to some component failure, which I can (or can have) repaired.

No such parity with PC based tools. I do not see a chain of PC repair shops that are willing or able to re-configure my PC and make it run better – or at all.

Until PC based test equipment shows such ability to work come rain or shine, I'm sticking with my 'blue boxes'.

This might sound like a kick in the pants for designers of soft test instruments. Not so. They also must sigh at the unreliability of a platform that has become so prolific.

Some applications software houses are keen to support their product. Labcenter for example has very good technical support. But even that company is battling against forces it cannot control: hardware, operating system/other badly behaved applications corrupting their files.

So on behalf of all electronics engineers – and probably any PC user – lets give the PC industry a kick up the pants to get developing and implementing industry wide standards to bring the reliability of the PC out of the dark ages.

We need stable, reliable platforms on which to run our tools – just as reliable as the electronics that we design will do! Now is that asking too much?

Michael Jeanes
Design Engineer
Advanced Media Design
Calabasas
California

Comme une bombe!

Reading about French radios in the February and April issues reminded me of something that has long puzzled me.

During a French lesson in the 1960s, someone was reading a passage in which there was a description of a radio being turned on. It was said that the radio remained silent for some time, and then the sound came on, "comme une bombe".

The French master explained that the sound from French radios really did come on like a bomb, reaching the set volume instantly. A British-made radio on the other hand would build up gradually when the valves had warmed up.

I have never heard an explanation for this. Was the application of the HT somehow delayed until after the valves had warmed up?

Hugh Mirams
Via e-mail

Beginners' corner

Although not an engineer with academic qualifications, I worked in the electronics field for around 40 years before retiring, a keen experimenter and a reader of *Wireless World* since the early 1940s and its reincarnation, *Electronics World*, still continuing to enjoy its articles.

The current 'Beginners' Corner' articles by the always excellent contributor Ian Hickman are a welcome addition to the wide spread of contents in the magazine – especially in view of the increasing demand of electronics engineers (July issue). They may also possibly attract more younger readers to the magazine.

Today there are few practical electronics magazines available. Those of the halcyon fifties and sixties having now mostly disappeared.

Norman Smith
Stoke-on-Trent

List of software

I am compiling a list of electronic-engineering software, and would appreciate the assistance of fellow *Electronics World* readers. If you know of any software that is applicable to electronics design, then I would appreciate hearing from you.

Our intention is to make this list available via the links section of our company's web-site.

There are plenty of CAD and PCB packages around, but there must be some I haven't seen yet, so include them. But I am more interested in electronics-specific stuff – and especially 'freeware' or 'shareware'.

For instance, I recently found a free program that displayed serial port traffic live on screen – very useful at the time.

Please send me the company URL where the program can be obtained. A one line description of the program function would help to sort the data. I'm not too interested in programming tools such as compilers etc., as there are plenty of them, but other than that I'd be glad to hear from you.

Thanks in anticipation.

Mike Bull
mike@msbdesign.co.uk

Everything's relative

I read the article 'Special Relativity right or wrong,' from the September issue with astonishment. I must say that, before publishing such a 'scientific discovery,' it would be useful to consult a specialist. The conclusions emerging from the article are, of course, wrong!

Unaware readers may be trapped by the ingenuity of the author to discover new laws of physics. He should have known that special theory is a subset of the general theory, the latter being able to explain what he claims to be nonsense.

I would recommend the author a very nice book, namely, 'The elegant universe,' which may bring some light at its relativistic speed into his discoveries.

Finally, I would recommend anyone who reflects on these theories to start thinking differently from he/she usually thinks for non-relativistic experiences.

As Einstein said, "Imagination is more important than knowledge." And some other physicist – I think Bohr – said many years ago that only a few scientists understand the theory of relativity.

The published article proves this latter reflection.

Sorin Stan, Ph.D.
Philips Optical Storage
Optical Recording
Predevelopment Laboratory
Eindhoven, The Netherlands

Al replies...

Sorin Stan says that my article has "ingenuity", which I take to mean that he cannot find a flaw in the reasoning.

He should read Hasselbach & Nicklaus (Phys. Rev. A, 48,143-51,1993). They list 21 different attempted explanations of the Sagnac effect. They say "this great variety (if not disparity) in the derivation of the Sagnac phase shift constitutes one of the several controversies that have been surrounding the Sagnac effect".

Having considered the eight proposed Special and General Relativistic explanations, they conclude, "the classical kinematical derivation has the advantage of yielding the correct first order result in a very simple and intuitive way".

Why should it be necessary to invoke General Relativity to explain a phenomenon that is evident at speeds of 10m/s? That is like saying that it is necessary to invoke G. R. before we can calculate the time of the rolling of a child's ball down a slope!

The Sagnac phase shift on the Michelson & Gale test is 3 000 000 000 000 000 times larger than calculated by Relativity theory.

Einstein wrote to Schrôdinger, "All the other fellows do not look from the facts to the theory, but from the theory to the facts; they cannot extricate themselves from a once accepted conceptual net, but only flop around in it in a grotesque way."

We must always give preference to well proven experimental results.

Al Kelly, Ph.D.
Dublin

100W Class-B

I found the article by Russell Breden in the June 2000 issue – 'A new 100W Class-B topology,' interesting, but less than convincing.

In a nutshell, the main problem he addresses is the reduced gain at high frequencies, leading to less effective feedback and making it very difficult to reduce cross-over-distortion – which of course contains many high-frequency components.

I can agree with his analysis that a better adaptation of the voltage amplifier stage could improve matters, because this would increase the available feedback gain. However, I cannot see why introducing feedback over the output and voltage amplifier stages should do this.

In a given design, there is a certain amount of excess gain (open-loop gain minus the required closed-loop gain) that can be used for distortion-reducing feedback. It doesn't make a difference how this is applied. It can be in the form of a single global feedback loop, or split between two or more nested feedback loops.

Nested feedback in combination with a modified frequency response of the network(s) *can* make a difference if it leads an increase in available feedback gain – in particular at high frequencies. However, there is no indication that this is the case in this design.

In addition to the nested feedback loop, Mr Breden tries to increase the voltage amplifier's input impedance by using a Darlington configuration, as in Fig. 3, $Tr_{11,12}$. This would indeed increase the available excess gain.

However, the input stage is also loaded by the inner feedback network, $R_{12,13}$ and $C_{5,6}$, which at mid frequencies reduces to approximately R_{13}, i.e. 68kΩ.

What is not immediately apparent though is that the output voltage at one end of R_{13} is in opposite phase to the driving signal at the collector of Tr_5. That divides the effective value of R_{13} by the closed-loop gain of the output *cum* voltage amplfier stage, negating the advantage of the Darlington.

In short, there is no evidence that Mr Breden's topology performs any better in the areas he addresses than the generic design.

Jan Didden
Consultant
Hees Belgium

I cannot accept Russell Breden's promotion of "the VFET's negative temperature coefficient".

I agree that the gain of VFETs usually has a negative temperature coefficient, but so does the threshold voltage. These two mechanisms compete with each other to produce a positive temperature coefficient. This means that for a given gate source voltage, the drain current will increase as the device temperature increases over most of the device's operating current range.

In fact, some devices, like the IRF632, exhibit this behaviour for all currents up to and in excess of their rated maximum continuous drain current.

This effect is quite evident from graphs of drain current versus gate-source voltage for various temperatures. The curves for higher temperatures are further up the current axis.

Unfortunately, I was not able to find data from any manufacturer for a device with part numbers 10N20 or 10P20*. Data from Thompson for an STM10NB20 shows that the threshold voltage – the gate to source voltage for which the device begins to conduct – does indeed decrease with increasing temperature. However the net effect – with decreasing transconductance – is not plotted. For clear graphs of this phenomenon, device data from Siliconix or International Rectifier is better.

Inevitably there is a drain current at which the device exhibits zero temperature coefficient. For all the VFETs whose data that I have studied, this current was always in excess of half the rated maximum continuous drain current. Hitachi made power MOSFETs like the 2SJ50 and 2SK135 whose zero temperature-coefficient current was quite low, at about 200mA for a 7A device.

However Hitachi made it quite clear in their application data that these were different from D series vertical power MOSFETS. They were lateral S series types. They also exhibited lower gain and higher saturated on resistance.

The first power MOSFET I tried was a vertical type. I literally got my fingers burnt. Luckily the FET survived and I suffered no permanent scarring. But I did learn a few lessons about MOSFETs and believing what I read.

I have since tried a variation of the amplified diode circuit, using a MOSFET instead of a BJT, but I was unimpressed with the performance. I suspect there is not good matching between devices, even of the same type – unlike BJTs.

One might ask if this is really important. On reflection, I have decided that for a Class-AB output stage it is important, since it impacts upon bias stability and probably cross over distortion.

In a design such as Russell's which seems to have no quiescent current – i.e. Class-B – it would seem at first glance to be unimportant. This would also be so if BJTs were used.

However, if the device is sufficiently heated – for example by running at high power for extended periods of time in a poorly ventilated audio-rack at the height of a Sydney summer – it is conceivable that the threshold voltage might be lowered sufficiently for the output stage to conduct some quiescent current. Then his amplifier becomes a Class-AB amp, like most audio power amps. I have neither the time nor the interest to investigate this.

Another point worth mentioning is the positive temperature coefficient of the power MOSFET's saturated drain resistance. This is more important when the device is used as a saturated switch, but it is also an important factor – along with the supply voltage – which will determine the maximum output swing a power amp can deliver before clipping.

As the device heats up its saturation voltage increases, which reduces the output of the amplifier. When used as a saturated switch, the current is largely set by the rest of the circuit. As the device heats up, its resistance increases, which in turn will increase its dissipation. So the temperature – and hence the resistance – increase.

This is hardly a scenario that can be described by a negative temperature coefficient, and if the device has inadequate heat sinking, thermal runaway may well occur.

The suggestions that VFETs are negative temperature-coefficient devices, and are therefore inherently thermally stable, are bunkum. Power MOSFETs are great devices with some nice properties, but they should be used with due design diligence given to their particular thermal behaviour.

Phil Denniss
Dept. of Plasma Physics
University of Sydney
NSW Australia

*Sorry, I should have spotted this before the article went to press. The 10N20 and 10P20 are 10A, 200V devices from Maplin. Ed.

More letters on page 825...

Dealing with mains interference

Joe Carr looks at the problem of interference from the mains, describing ways to avoid it – and ways to remove it.

Over the past several decades I've seen a lot of situations where 50/60Hz interference disrupted all manner of circuits and instruments. It's basically a pain.

The best solution, by far, is to prevent it from occurring. The usual methods involve shielding, or using a location that is free of 50/60Hz fields. Lots of luck! Have you noticed those lights you are using to read this article?

A practical example arose when I was employed as an electronics technician (later as an engineer after I got my degree) at George Washington University Medical Centre in Washington, DC. A physiologist was using a Wheatstone bridge transducer to measure the contractions of some muscle or another.

The output of the sensor was very tiny. Its sensitivity factor was 5μV/V/gf (1gf, or 1gram-force is a force of 980dynes). With a 6V battery, and gram-forces between 0.5 and 2, the output of the contraction pulses were from 15μV to 60μV. The bridge had a differential output, so had to be processed in a differential amplifier.

Figure 1 shows the system used. The two sensor output wires were passed to a very long – about 3.5 metres – shielded two-wire cable, to a differential input on a strip-chart recorder. That recorder is shown here as a differential amplifier.

Unfortunately, the differential input was not truly differential. It was – like an oscilloscope – a two-channel device, with channels A and B, that permitted an 'A-B' input to function as pseudo-differential.

Even though the cable was shielded, the slight difference in the input amplifiers produced a common-mode voltage that was seen as a valid input signal by the recorder. The 50/60Hz fields radiated by the building power wiring induced 50/60Hz signals into the sensor circuit. And with the tiny input signals, any 50/60Hz at all was fatal. The scientist's tracings were corrupted with 50/60Hz interference.

My first solution was to tell him to buy a differential amplifier plug-in for the recorder. Wrong answer! It seems that research grants for junior scientists are not all that generous. He simply didn't have the money to buy the correct amplifier. So I decided to build a differential amplifier at the sensor end of the cable.

Figure 2 shows the resultant system. A gain-of-100 dc differential amplifier, using a CA3140 operational amplifier, was built and installed right at the sensor. A single-conductor coaxial cable was run to one of the single-ended amplifier inputs of the recorder. This eliminated the A-B pseudo-differential problem. The 50/60Hz completely disappeared.

Figure 3 shows the device. The Grass FT-3 strain gauge sensor had a Cannon connector on it. I obtained the mate and mounted it on a Pomona box that contained the amplifier. A seven-pin Amphenol 126-series connector on the other end carried DC and grounds –

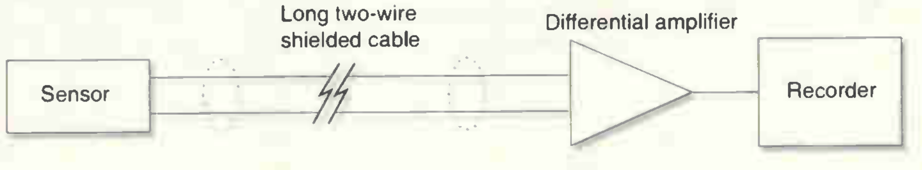


Fig. 1. Original configuration of sensor and amplifier was sensitive to hum.

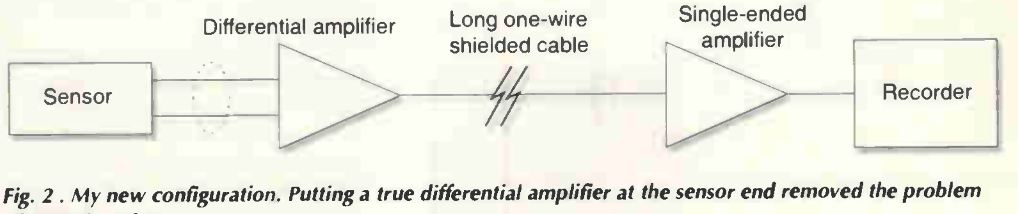


Fig. 2. My new configuration. Putting a true differential amplifier at the sensor end removed the problem of 60Hz interference.

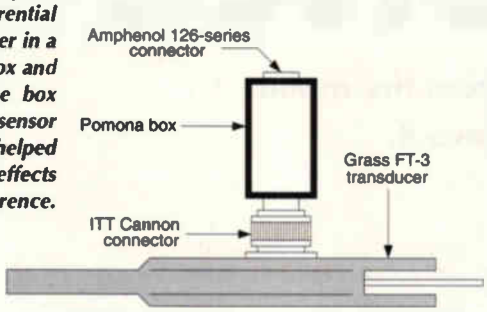
one each for DC and signal – to the box, and carried signal back to the recorder. Cost? About $25 in 1975 bucks.

One humourous aside. I met the scientist a day or so later and he asked me for a written 'technical description' of the project. OK, so I sent him a couple paragraphs. He then said he had something on the order of five to ten pages on the project. So I wrote a piece about the length of a magazine article that started with the theory of op-amps and ended in my two original paragraphs.

About a year later he handed me fifty reprint copies of 'our article.' It seems that he had submitted my words along with about as many of his own to a physiology journal 'tech notes' section.

A bit shocked, I told him that any second semester electronics technology student at the local community college could do the same. He replied: "Yes, but a physiologist couldn't... and would have to pay $975 for the correct plug-in amplifier."

Fig. 3. Putting the extra differential amplifier in a 'Pomona' box and mounting the box local to the sensor also helped reduce the effects of interference.



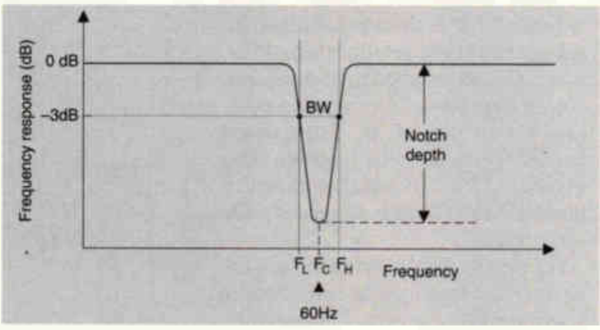


Fig. 4. Frequency response of notch filter, whose bandwidth, BW, is the bit between the -3dB points of the notch.

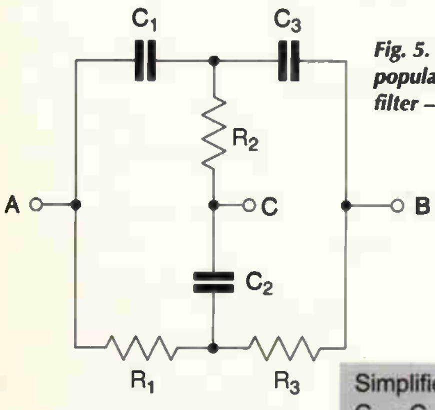


Fig. 5. One of the most popular forms of notch filter – the twin-tee.

Simplified expression
$C_1 = C_3 = C$
$C_2 = C$
$R_1 = R_3 = R$
$R_2 = R/2$

If you can't remove it, you'll have to filter it

Unfortunately, it's not always possible to prevent 50/60Hz from occurring. In that event, we need to find a solution involving filtering out the 50/60Hz stuff.

If the input signal permits, then a high-pass or low-pass filter can be used. Any signal has harmonics, and these must fall within the bandpass of the filter. For very slow signals, with a maximum frequency content of 20Hz or less, then a low-pass filter will sufficiently attenuate the 50/60Hz signals. Similarly, if the signal frequencies are much higher than 50/60Hz, then a high-pass unit will be needed.

Notch filters. The usual solution to unwanted in-band frequencies is the notch filter. The frequency response of a notch filter is shown in **Fig. 4**. These filters are similar to another class, i.e. bandstop filters, but the band of rejection is very narrow around the centre frequency, F_c.

The bandwidth of these filters is the difference between the frequencies at the two -3dB points, when the out-of-notch response is the reference 0dB point. These frequencies are F_L and F_H, so the bandwidth is F_H–F_L.

The 'sharpness' of the notch filter is a measure of the narrowness of the bandwidth, and is specified by the 'Q' of the filter. The Q is defined as the ratio of the centre frequency F_c to bandwidth,

$$Q = \frac{Fc}{BW} \qquad (1)$$

For example, a notch filter that is centred on 60Hz, and has -6dB points are 58 and 62Hz (4Hz bandwidth) has a Q of 60/4 or 15.

The notch filter does not remove the entire offending signal, but rather suppresses it by a large amount. The notch depth, see again Fig. 4, defines the degree of suppression, and is defined by the ratio of the gain of the circuit at an out-of-notch frequency – for example F_{ob} – to the gain at the notch frequency.

Assuming equal input signal levels at both frequencies – which has to be checked, most signal generators have variable output levels with changes of frequency! – the notch depth can be calculated from the output voltages of the filter at the two different frequencies,

$$Notch\ depth = 20\log\frac{V_{fc}}{V_{ob}} \qquad (2)$$

Twin-T notch filter networks. One of the most popular forms of notch filter is the twin-tee filter network, shown in **Fig. 5**. It consists of two T-networks, consisting of $C_1/C_3/R_2$ and $R_1/R_3/C_2$. Notch depths of -30 to -50dB are relatively easy to obtain with the twin-tee, assuming proper circuit design and component selection. Very good matching and selection of parts makes it possible to achieve -60dB suppression.

The centre notch frequency of the network in the generic case is given by,

$$F_c = \frac{1}{2\pi}\sqrt{\frac{C_1 + C_3}{C_1 C_2 C_3 R_1 R_3}} \qquad (3)$$

You can simplify the expression above by adopting a convention that calls for the following relationships,

$C_1=C_3=C$
$R_1=R_3=R$
$C_2=2C$
$R_2=R/2$

If this convention is adopted, then we can reduce the frequency equation to,

$$F_c = \frac{1}{2\pi RC} \qquad (4)$$

In these expressions, F is in hertz, R is in ohms and C is in farads. Be sure to use the right units when working these problems: 0.001µF is 1×10^{-9} farads.

In calculating values, it is usually prudent to select a capacitor value, and then calculate the resistance needed. This is done for two reasons: one is that there are many more standard resistance values, and second, potentiometers can be easily used to trim the values of resistances, but it is more difficult to use trimmer capacitors.

One of the problems of these filters is that the depth of the notch is a function of two factors involving these components. First, that they are very close to the calculated values, and second, that they be matched closely together. For example, a 60Hz notch filter was built using the 0.15µF and 17 684Ω values, the 0.15µF capacitors were selected at random from a group of a dozen or so 'mine-run' capacitors of good quality, while the resistors were 18kΩ, 5% metal-film types. The notch depth at 60Hz was only 10dB, but at 58Hz it was 48dB. The mismatch caused a significant shift of notch frequency.

A second filter was built using the same values. In this case, the 0.15μF capacitors were selected from about 20 on hand. Precision components are difficult to obtain.

In order to match them as closely as possible, the capacitance of each was measured using the capacitance tester function on a sub-$100 digital multimeter. The order of priority of selection was to find those that closely matched each other, and only incidentally how close they come to the calculated value. Errors in the mean capacitance of the selected group can be trimmed out using a potentiometer in the resistor elements of the twin-tee network.

When selecting a frequency source, choose either a well-calibrated source, or use a frequency counter to measure the frequency. Keep in mind the situation described above where only a 2Hz shift produced a 38dB difference in notch depth! Alternatively, use a 6.3V or 12.6V ac filament transformer secondary as the signal source – after making sure that its potentially lethal primary is properly insulated and protected.

Adjustable-frequency notch filters can be built using the twin-tee idea, but none of the usual solutions are really acceptable. One implementation requires three ganged, matched potentiometers or three ganged capacitors. Unfortunately, in either case, at least one of the variable sections must be of different value from the other two, causing a tracking problem. You might not notice a tracking problem in some circuits, but in a high-Q notch filter it can be disastrous.

Active twin-tee notch filters. Active frequency selective filters use an active device such as an operational amplifier to implement the filter. In the active filter circuits to follow, the 'twin-tee' networks are shown as block diagrams for sake of simplicity, and are identical to those circuits shown earlier; the ports 'A,' 'B,' and 'C' in the following circuit are the same as in the previous network.

The simplest case of a twin-tee filter is to simply use it 'as is,' i.e. use the filter circuits shown above. But the better solution is to include the twin-tee filter in conjunction with one or more operational amplifiers.

There is one solution in which the twin-tee network is cascaded with an optional input buffer amplifier and a mandatory output buffer amplifier. Such amplifiers tend to be non-inverting op-amp follower circuits.

The purpose of buffer amplifiers is to isolate the network from the outside world. For low-frequency applications, the op-amps can be 741, 1458, or similar. For higher-frequency applications, i.e. those with an upper cut-off frequency above 3kHz, use a non-frequency compensated device such as the CA-3130 or CA-3140.

A superior circuit is shown in **Fig. 6**. In this circuit, port-C of the twin-tee network – the common point – connects to the output terminal of the output buffer amplifier. There is also a feedback network consisting of two resistors, R_a and a capacitor, C_a. The values of R and C in the twin-tee network are found from the equation above, while the values of R_a and C_a are found from,

$$R_a = 2RQ \qquad (5)$$

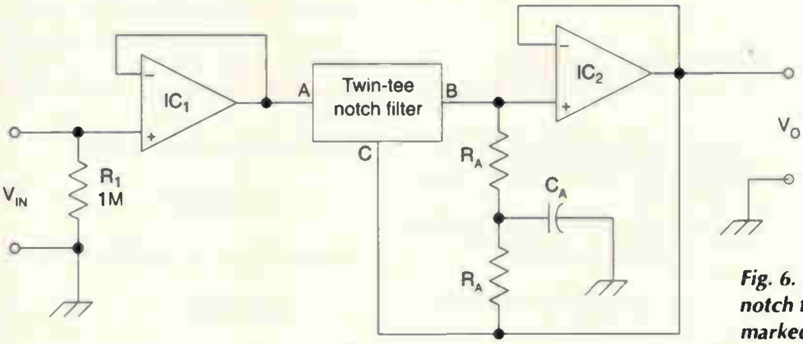


Fig. 6. Active twin-tee notch filter. The box marked Twin-tee notch filter is simply Fig. 5.

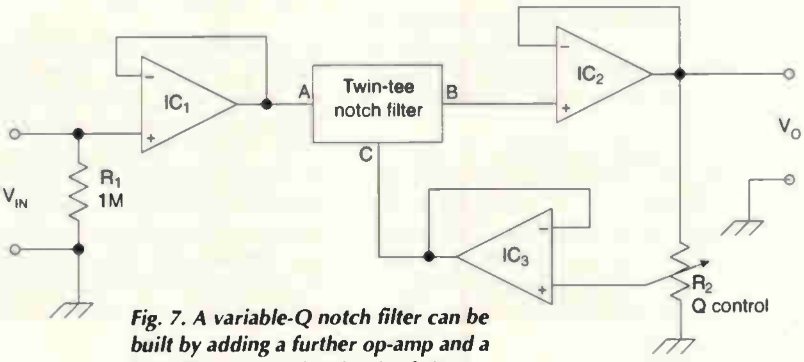


Fig. 7. A variable-Q notch filter can be built by adding a further op-amp and a potentiometer to the circuit of Fig. 6.

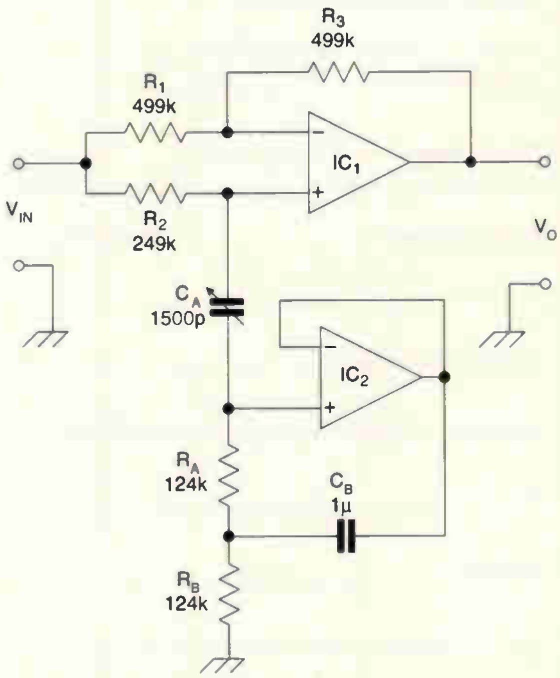


Fig 8. Gyrator notch filter. The 1500pF variable can be made up of a fixed capacitor and a trimmer in parallel.

and,

$$C_a = \frac{C}{Q} \qquad (6)$$

Filter design example

Design a 60Hz notch filter with a Q of 8.

1. Select a trial value for C of 0.01µF
2. Calculate the value of R from the equation: 265 392Ω
3. Calculate $R/2$: 265 392/2=132 696Ω
4. $C_2=2C=(2)(0.01\mu F)=0.02\mu F$
5. Select R_a;R_d=$2QR$=(2)(8)(265 392Ω) =4.24MΩ
6. Select C_d=C/Q=0.01µF/8=0.0013µF.

When Fig. 6 was built using these values in Joe's Basement Lab, using the twin-tee network the null was close to –48dB deep using components at hand.

A variable-Q control is shown in **Fig. 7**. In this circuit, a non-inverting follower, IC_3, is connected in the feedback loop in place of R_a and C_a. The Q of the notch is set by the position of the 10kΩ potentiometer, R_2. Values of Q from 1 to 50 are available from this circuit.

Another approach to notch filter circuits is shown in **Fig. 8**. This circuit is sometimes called the gyrator or active inductor notch filter (it's also sometimes called the virtual inductor notch filter).

The notch frequency is set by,

$$F_c = \frac{1}{2\pi\sqrt{R_a R_b C_a C_b}} \qquad (7)$$

Equation 7 can be simplified to,

$$F_c = \frac{1}{2\pi R\sqrt{C_a C_b}} \qquad (8)$$

If the following conditions are met,

$$\frac{R_3}{R_1} = \frac{R_2}{R_a + R_b} = \frac{R_2}{2R} \qquad (9)$$

It is possible to use any one of the elements, C_a C_b R_a, or R_b, to tune the filter. In most cases, C_a is made variable and C_b is a large value fixed capacitor.

The 1500pF variable capacitor can be made by parallelling all sections of a three-section broadcast variable, with a single small fixed or trimmer capacitor. Alternatively, since most applications will require a trimmer rather than a big honkin' broadcast variable, it is also possible to parallel one or more small capacitors and a trimmer.

For example, a 100pF trimmer can be paralleled with a 1000pF and 470pF to form the 1500pF capacitance required. Make sure that you use low-drift, precision capacitors. You can match them using a digital capacitance meter.

Be careful when using any filter to remove components from a waveform. If the filter is not a high-Q type, then too much of the signal might be removed. In medical electrocardiograph (ECG) systems the signal has components from 0.05 to 100Hz, so 60Hz is right in the centre of the range! Oops.

To make matters worse, the leads have to be connected to the human body, so are unshielded at their very ends. Interference from 60Hz is almost guaranteed unless care is taken. But filtering can take out components that assist the physician in making diagnosis, so is only used when it is unavoidable. On medical ECG amplifiers the filter is usually switchable so it can be either in or out of the circuit. ■

BOOK TO BUY

The definitive biography of the century's godfather of invention–from the pre-eminent Edison scholar "Israel's meticulous research and refusal to shy away from the dodgier aspects of Edison's personality offers a fresh glimpse into the life of the inventor."– New Scientist

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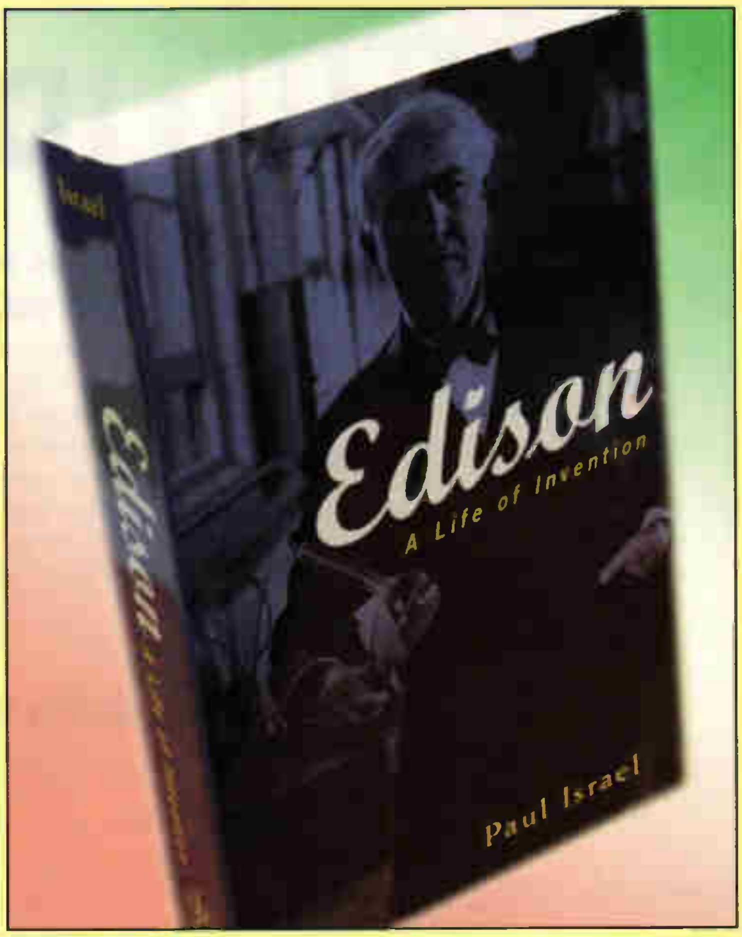

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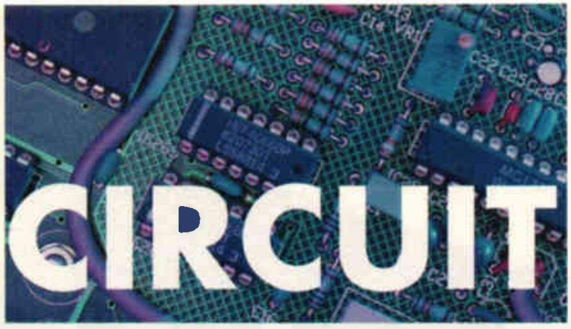


CIRCUIT IDEAS

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

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

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

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

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

Half-duplex data direction indicator

If serial data is to be monitored in half duplex data transmission network, it is essential that the data direction is known before the interleaved bi-directional character strings can be separated.

Where the protocol is not known, or is too complex to enable any synchronisation characters to be detected in real time, one alternative is to determine the data direction using hardware. The diagram is a circuit that extracts a data direction signal, which can then be used to route the serial data to the appropriate channels for monitoring. Values of resistors R_1 to R_5 are suitable for operation with CMOS/TTL levels, and have been tested successfully with Smart Card (ISO7816) serial data communications up to 9600 baud.

Circuit operation is as follows. When data direction is from I/O port A to I/O port B, if a logic 1 is being transmitted, current flow through R_3 will be from A to B. Port A will be current sourcing, and the receiving load will be current sinking.

Resistor R_5 provides some additional load to ensure adequate current flow through R_3 when driving high-impedance loads such as CMOS gate inputs. When a logic 0 is being transmitted, then current flow through R_3 will from B to A. Port A will be current sinking and the receiving load will be current sourcing, this time assisted by R_4.

The converse is true when the data flow is from I/O port B to I/O port A – with R_1 and R_2 contributing to the load. Thus, the logic level 'C' from the comparator indicates the direction of the current flow, and when this is XOR'ed with the data level 'B', indicates the direction of the data flow. The function is similar to that of a synchronous demodulator.

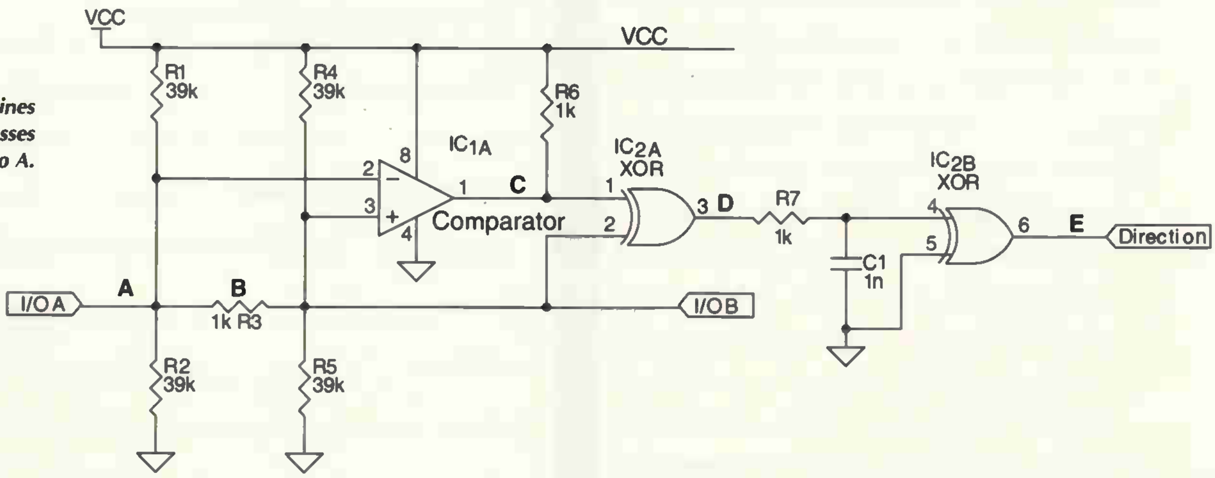
Glitches at 'D' caused by propagation delays are filtered out by R_7 and C_1 with IC_{2B} buffering the output 'E'. A high-value feedback resistor could be connected between the output of IC_{2B} and the input on pin 4 to provide some hysteresis if needed.

The direction signal can be used to route the serial data stream to the appropriate serial channel or to TTL to RS232 converters for monitoring using programs such as CardMon, Comwatch or Windows Terminal etc.

The principle is equally applicable to half-duplex RS232 and RS422/485 etc., however some modification to the values of the resistor $R_{1\text{-}5}$ and to the comparator supply voltages would be required.

Alan Kitching
Middlesbrough
E6

This circuit determines whether data passes from A to B or B to A.



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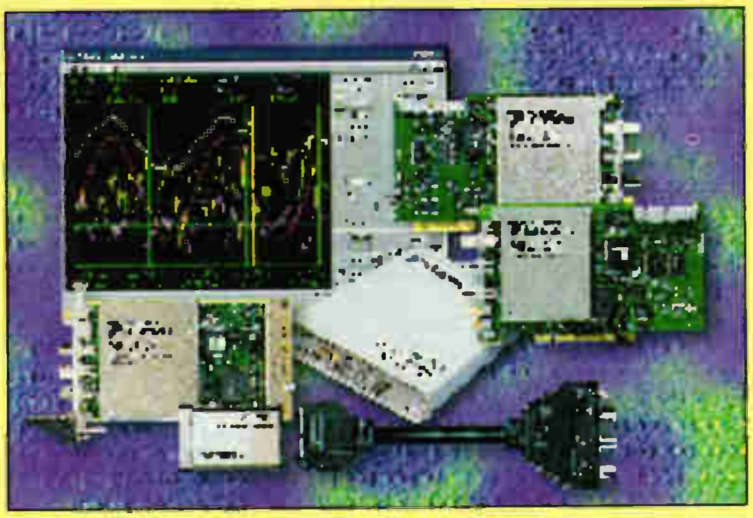

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*All published circuit ideas that are not eligible for the prizes detailed here will earn their authors a minimum of £35 and up to £100.

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Simple keyless entry lock

Where controlled access to a room or premises is required, a keyless lock is often the most convenient solution. Additional 'keyholders' can be authorised at any time, newly cut keys being unnecessary, and the key can be easily changed, by reallocating a few wires, should this become necessary.

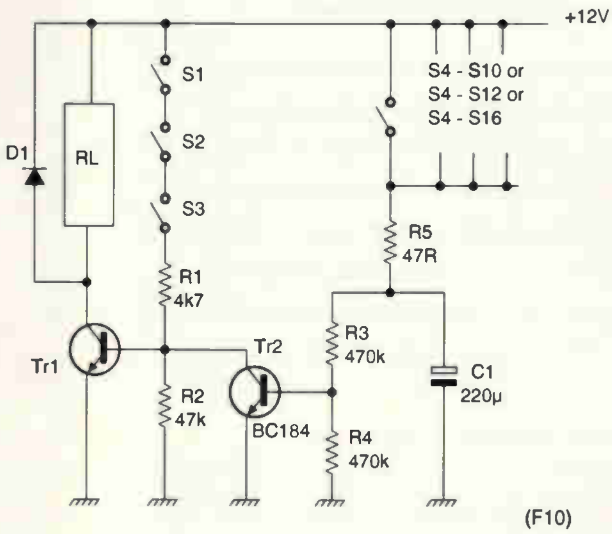
The diagram shows one such a scheme, using a ten, twelve or sixteen switch keypad. Where the solenoid-controlled latch release is 12V DC operated, it can be driven directly by Tr_1 as shown.

In this case, Tr_1 should be a Darlington transistor, such as a TIP121 fitted with small heatsink, and D_1 a 1N4001. Otherwise, *RL* can be a relay controlling the latch supply, and Tr_1 and D_1 can be a BC182 and 1N4148 respectively.

Switches $S_{1\text{-}3}$ can be any three of the contacts of the keypad. To gain access, it is necessary to press all three simultaneously. This is likely to foil the unauthorised, as it differs from the usual arrangement, where a set sequence of presses is required.

Furthermore, pressing any other key at any stage will charge up C_1. This will disable the lock for about a couple of minutes, adding further security and making unauthorised entry quite impossible in practice.

Ben Sullivan
Waterlooville
Hampshire
F10



(F10)

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Power supply noise-reduction circuit

Periodic radiated and conducted electrical noise at a spot frequency can be reduced in power by spreading the spectrum of the noise signal. The technique was used on an adapted power supply circuit with good results.

This circuit alters the switching period on a per pulse basis. Due to its complexity, it may find use as a final line of defence against electrical noise where the addition of further filtering and screening had proved ineffectual.

Where the power supply period of operation is set by the charging of a capacitor through a fixed resistance, this circuit feeds an error current into the capacitor. The error current changes per power supply cycle.

Gate drive from the power supply is used to clock a pseudo random noise generator formed by a shift register with tapped feedback. The weighted output of all 8 taps is fed to the power supply timing capacitor circuit. Attenuation of the this error signal gives reduced deviation of the power supply switching frequency.

On this circuit 5V power was taken from the regulated 5V output of the power supply switching *IC*, (UC3843 50mA output max.). The advantage is the generation of the power on reset signal to the shift register.

Frank Serrels
Birmingham
E7

£50 Winner

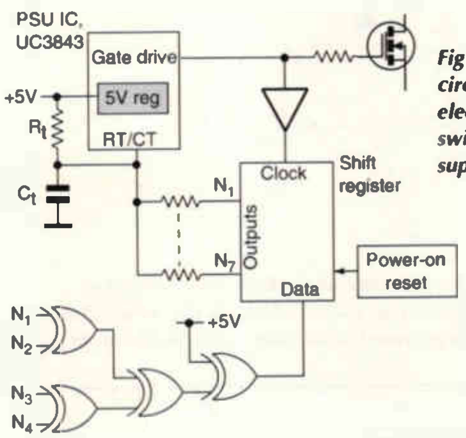


Fig 1. Block diagram of circuit for reducing electrical noise from a switching power supply.

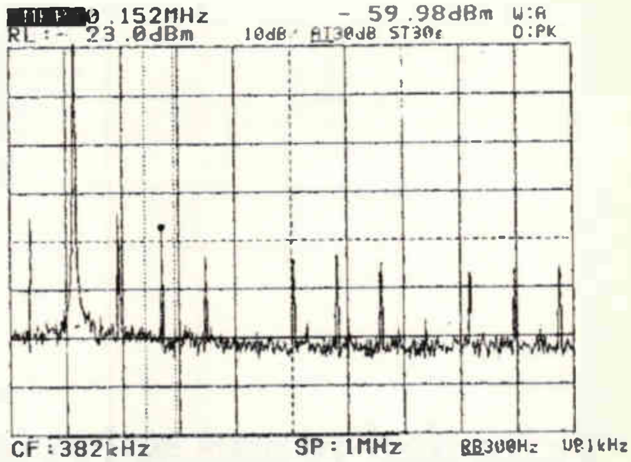


Fig. 3. Spectrum of power supply output, 5V in this instance, circuit disabled. DC – i.e. 0Hz frequency – is shown for reference.

Fig. 2. Circuit diagram of scheme for reducing electrical noise from a switching power supply.

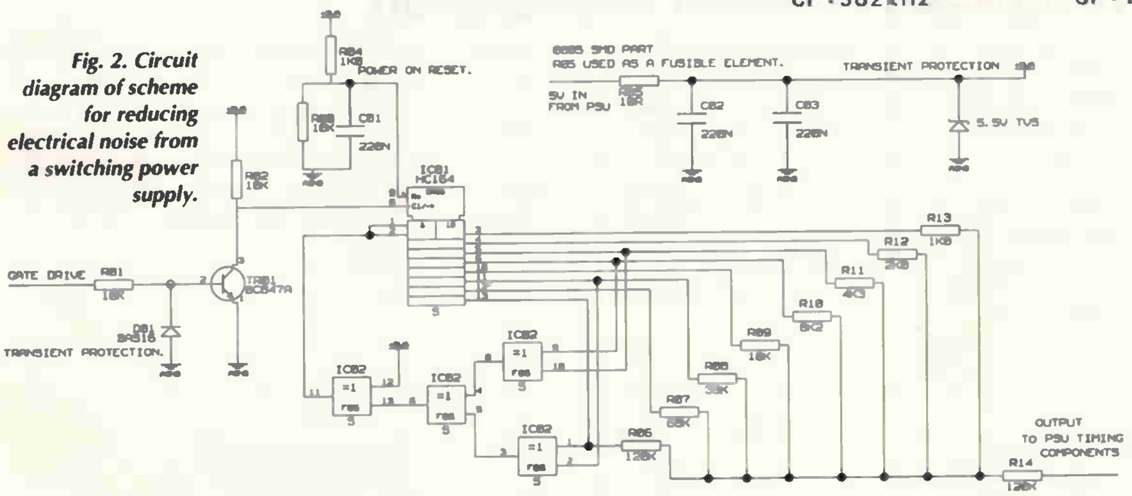
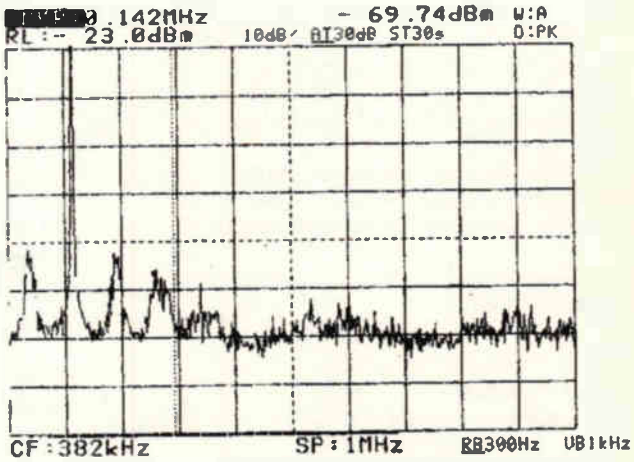


Fig. 4. As Fig. 3, circuit enabled.



Comparison of results taken with and without the power supply noise reducer in circuit.

Harmonic	Frequency (kHz)	Power unmodified (dBm)	Power modified (dBm)	Change (dBm)
	76	–57	–65	–8
2	152	–60	–70	–10
3	228	–66	–74	–8
4	304	<–80	<–8	
5	380	–67	<–80	–13
6	456	–67	–78	–11
7	532	–68	<–80	–12
8	608	<–80	<–80	
9	684	–70	<–80	–10
10	760	–69	–79	–10
11	836	–69	<–80	–11

Switch a PC CD-ROM writer's power supply

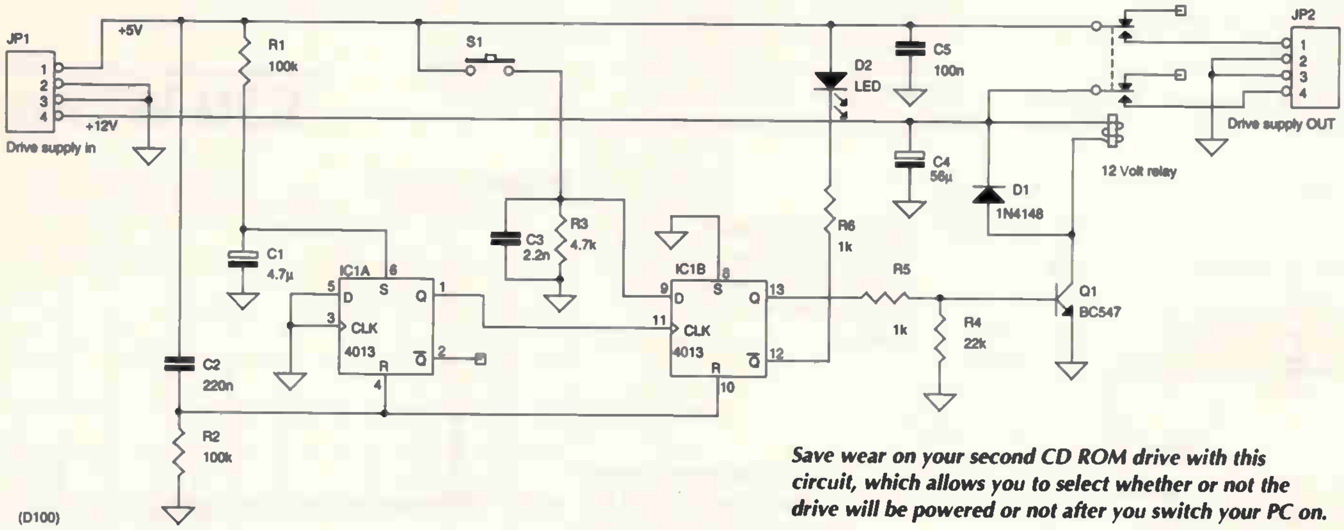
This simple 'error-free' switch powers a second SCSI CD-ROM writer only when needed.It is impossible to change the CD state, on or off, after the system boots. When powering on the PC, the ICs are reset by a single pulse provided by C_2-R_2 network.

After a little delay provided by R_1 and C_1, IC_{1B} is clocked by the rising level of IC_{1A}'s output. The logic level present on input *D* of IC_{1B} control the relay: If S_1 is pressed during switch on, the CD is powered.

Capacitors C_4 and C_5 provide supply bypass and C_3 acts as a debouncer. Because IC_{1B} flip-flop is clocked only when the PC is turned on, the circuit inhibits erroneous operation.

Both S_1 and D_2 are placed on the PC's front panel, near the main power switch. To use CDRW, keep S_1 pressed while turning on the PC until the led lights, The switch is then released and the CD writer is available for use.

Marco Spada
Bari
Italy
D100



Save wear on your second CD ROM drive with this circuit, which allows you to select whether or not the drive will be powered or not after you switch your PC on.

Extended-range pulse width modulation

In normal 555 applications, the capacitor voltage is limited between 1/3 and 2/3 V_{cc}. This can create problems when using this *IC* for speed-control using PWM, as in most cases the speed setting is a voltage level, between for example 0V and 10V.

It is however possible to create a circuit which creates an offset voltage between the capacitor voltage and the voltage sensed by IC_1, thus making it possible to build a PWM regulator with the capacitor voltage within the desired voltage control range.

By inserting R_1 and R_2 between pin 2 and C_2 the capacitor will discharge to a lower voltage level before retriggering IC_1. Resistors R_1 and R_2 are not only used to charge the capacitor; they also create the necessary offset between the real capacitor voltage and the voltage sensed by IC_1.

As the resistors are connected between C_2 and pin 2, the threshold level isn't affected. In the circuit drawn the capacitor starts charging at 1.8V while the discharge cycle starts at 10V

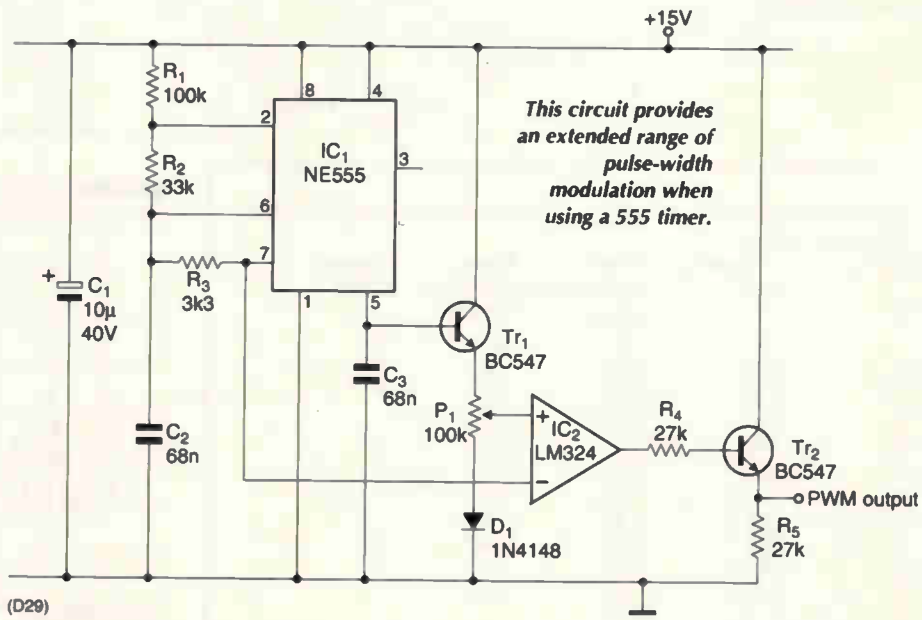
As it was necessary to generate PWM with a minimum duty-cycle of 5%, the capacitor voltage isn't extracted directly at the capacitor, but on pin 7 of IC_1. This allows the inverting input of IC_2 to remain near GND level during C_2's discharge-cycle.

If the input potentiometer remains higher than 0,6V, the output of IC_2 will stay at *+Vcc*. By placing a short circuit at the place of D_1, the minimum PWM signal disappears and the output -remains low during the discharge cycle of C_2

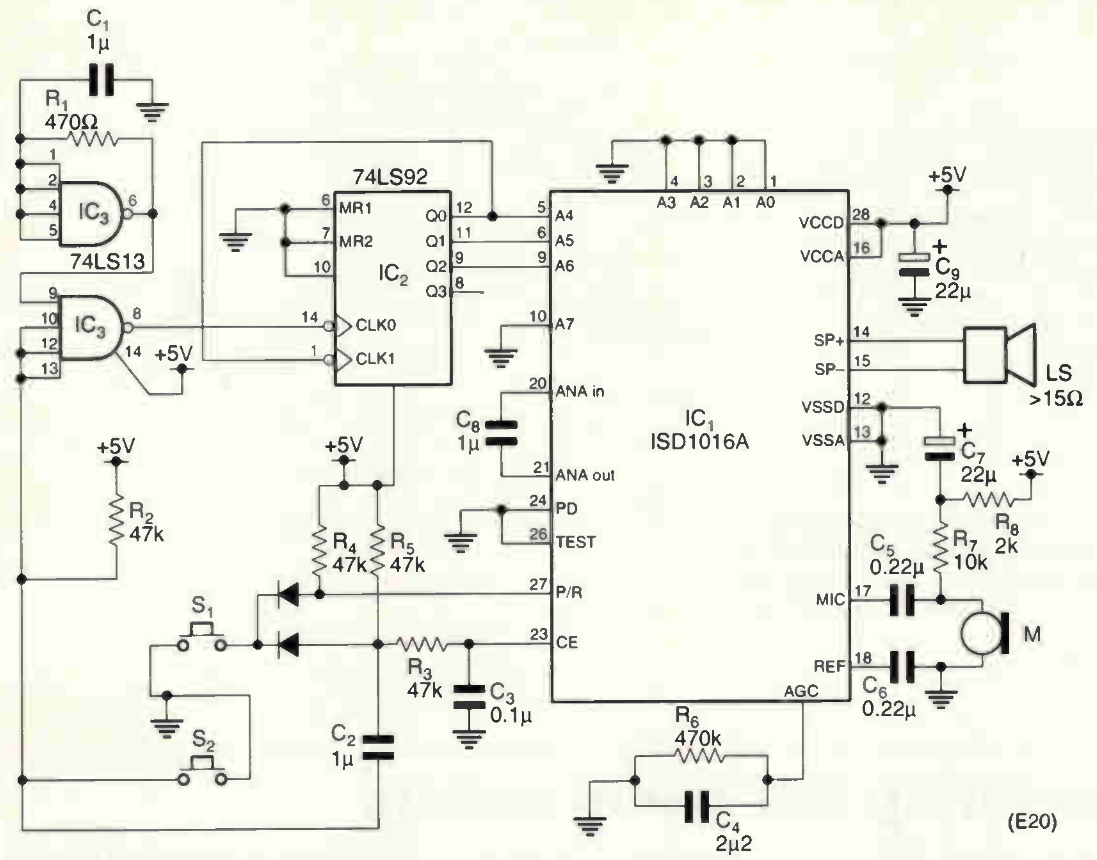
As there is a internal reference in IC_1 which is used to generate the capacitor voltage, it is possible to use this voltage as a reference for the potentiometer. Using a transistor makes it possible to drive the low impedance of the potentiometer from the control-voltage pin of IC_1 without affecting the voltage level>

Bernard Van den Abeele
Evergem
Belgium
D29

Specifications	
f	108Hz
Duty, minimum	5%
Duty, maximum	89%
Duty cycle at U_{P1}=5V	33%



This circuit provides an extended range of pulse-width modulation when using a 555 timer.



Press my S_2, and I speak the throw of the die.

Speaking die

This speaking die circuit is suitable for use by people with sight impediments, as it audibly outputs the result of a 'throw' via a small loudspeaker, in the user's language.

Squarewave oscillator IC_3 (upper) drives divide by 12 counter IC_2, of which the most significant output is not used, via *NAND* gate IC_3 (lower). Counter outputs Q_0, Q_1 and Q_2 are connected to address lines A_4, A_5 and A_6 of voice chip IC_1.

Pushing S_2 – 'throwing' the die – closes the *NAND* gate, stopping the counter and triggering IC_1 via C_2. The speech chip IC_1 then outputs the message recorded at the address at which the counter stopped.

For use as a die, the messages recorded at the six locations are 'one' through 'six'. The frequency of the oscillator is high enough that it is impossible to predict the count at which S_2 is closed, and the resulting word is therefore effectively randomly chosen.

Recording is achieved by speaking into the microphone, *M*, with S_1 closed. Releasing the button stops recording. The message is recorded at the location defined by A_4, A_5 and A_6, which must therefore be frozen during recording. Addresses can be frozen by pressing S_2 repeatedly until the counter stops at the desired address, keeping S2 depressed thereafter, to prevent the counter advancing.

Address lines A4, A5 and A6 divide the voice chip memory into five segments of 1.6s each, and one of almost eight seconds. Each slot can thus easily contain a numeral and slot six also 'bells, whistles and fanfares' to celebrate a good throw, if desired.

There is no limit to the number of times the message can be changed, and the device can thus be modified for other purposes. For example, if three slots contain 'yes' or 'heads' and the other 'no' or 'tails', the device is like tossing a coin. The author's prototype was built in a box 8cm by 5cm by 3cm, powered via a 78L05 regulator. This in turn can be powered by a battery of a plug-top mains-powered supply.

Heikki Kalliola
Helsinki
Finland
E20

One microsecond trigger pulse generator

The two circuits shown both generate a one microsecond trigger pulse on the rising edge of an input clock pulse. The first circuit came from the 'Art of Electronics' and is a very useful little circuit that can use spare logic gates on a circuit.

The second circuit came from trying to implement

Continued over page...

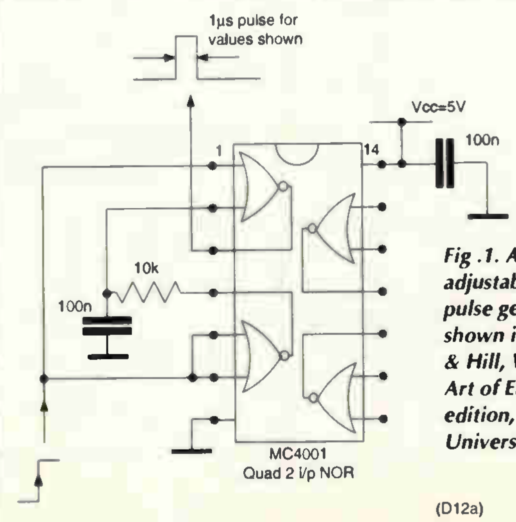


Fig .1. A Simple adjustable rising-edge pulse generator as shown in Horowitz, P., & Hill, W., (1995): The Art of Electronics, 2nd edition, Cambridge University Press, p557.

(D12a)

the first circuit, but discovering that I would need an extra chip to do it. However I noticed that I had a spare flip-flop and was able to generate the requisite 1ms trigger pulse using it. It is the most basic of circuits, but it may be of use to someone caught in a similar bind.

Richard Weir
via e-mail
D12

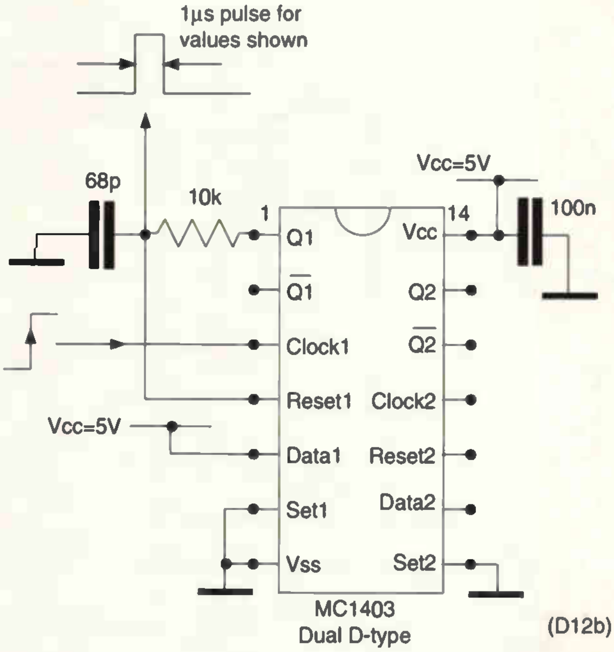


Fig. 2. A simple adjustable rising-edge pulse generator constructed from the spare side of a D-type bistable device, saving the use of an additional chip in a tight spot.

Measuring EHT beam current

Measuring beam current in electron-beam evaporation equipment, sputter coaters and similar devices presents difficulties, due to the problems of insulation of panel mounted meters. This circuit represents a good workable solution, using a small transformer with good insulation. The principle of operation of the metering circuit developed by Milman Consultants is shown in the diagram.

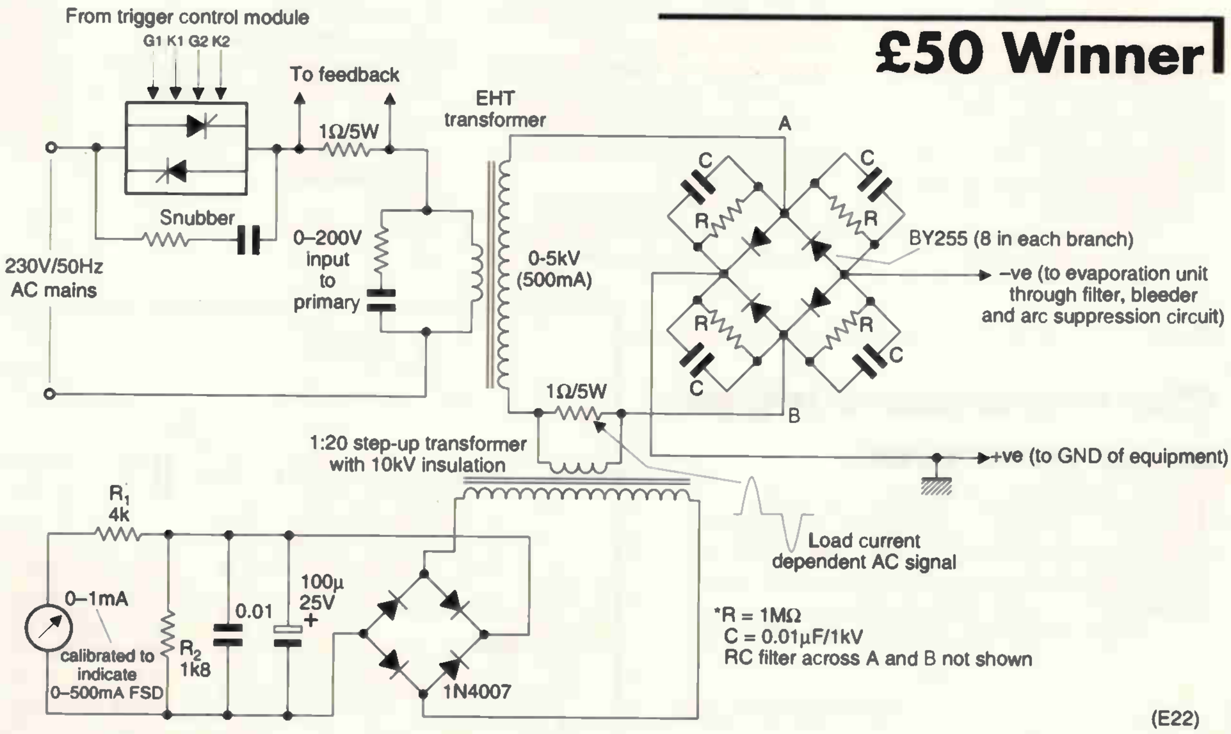
EHT load current is sensed before rectification by taking the voltage drop across a 1Ω, 5W resistor connected in series with the secondary of the EHT transformer. This feeds a miniature 1:20 step-up transformer with adequate insulation.

Stepped up voltage is rectified, filtered and fed to a 1mA meter through a series resistor. Actual values of R_1 and R_2 can be adjusted to suit the metering range etc. Since the power handled by the transformer is so small, its size is dictated by the necessary between winding and to-ground insulation.

Before incorporating the metering circuit and EHT supply into the host equipment, the meter was calibrated using a dummy load. A plot of the meter reading against the actual load current indicated a linear relationship throughout the range. Maximum error at the extreme ends of the range was less than ±5%, which was acceptable for all applications. ■

Dr. G Acharya & R Ombase
Pune
India
E22

£50 Winner



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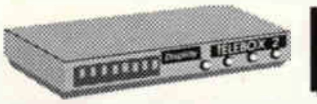

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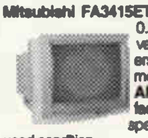

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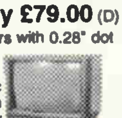

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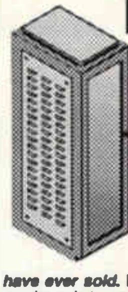

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CIRCLE NO. 122 ON REPLY CARD

Measuring true power

As circuit miniaturisation increases, knowing the temperature of small components becomes ever more important, and ever more difficult to determine. Cyril Bateman has been searching the Net for alternative ways of assessing power dissipation.

Today's emphasis on physical size reduction requires similar reductions in the actual power dissipated in a circuit, if excessive component temperatures are to be avoided. With miniature components, sensible measurement of temperature can prove difficult.

For several years I have measured component temperatures using miniature naked-bead, 0.2mm-wire, thermocouples. However even with a size 1206 resistor or capacitor, due allowance for temperature reduction due to the heat conducted along the thermocouple wires is needed. With smaller components this becomes impracticable.

Alternative temperature-measurement methods

One possible method is to estimate component temperatures using thermochromic liquid crystals. I have used this technique with some success, using the Redpoint Spectratherm liquids. These cover the temperature range 60°C to 115°C. You can find more details on this technique at the 'Electronics Cooling' site[1].

Another technique uses an infrared thermometer to measure surface temperatures. This method is most successful for high temperatures and low-cost thermometers are readily available. These however 'view' much too large an area to be useful when measuring small components.

Extremely small components can be measured using microscope-based infrared measurement methods, but equipment for such techniques is very expensive.

Consequently, while voltage, current and frequency are the most common measurements made when

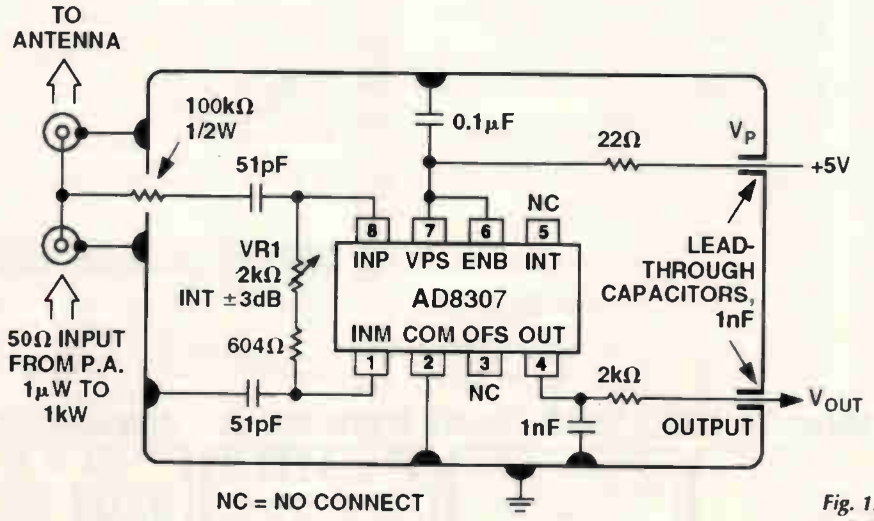


Fig. 1. A low cost, easily applied solution for 50Ω RF power measurements from 1µW to 1kW.

Bugs

The recent spate of e-mail based virus attacks[A] leaves the impression that other security problems for Internet users have disappeared. Not so. Default installations of many operating systems/browsers frequently expose your computer to potential damage.

Having just completed the assembly of my new Athlon based machine, I decided to subject it to a practical security test. With this computer, I can access Internet using either OS/2 or Windows 98SE, so I configured both operating systems/browsers to ensure Internet security.

Using Windows 98SE with Netscape 4.72 then IE 5, I visited the 'Shields Up'[B] for some practical tests. In a few seconds my port 139, or NetBios port, was found to be completely 'open', providing an easy route into my data. This revealed just how quickly a hacker could verify your machine's Internet presence and penetrated it[B].

Fortunately, better results were obtained using OS/2 with Netscape4.61. Shields Up reported all ports were correctly closed and my system had not been interrogated. Using the same hardware I have also found that OS/2 produces faster Internet download results, than when using Windows 98SE with either the above browsers. So with this improved security as well as extra access speed, I will continue using OS/2 as my main communication operating system.

Improved security can be attained using a 'firewall'. While the traditional hardware approach may not be practicable for a single-user system, low-cost software-based firewalls for Windows9x and NT systems, are now available[C].

One that has received favourable review, costs only $19.95 for business use and nothing at all for personal or non-profit use, Fig. A. In addition to its firewall function, the current version also includes 'MailSafe' which stops dead any e-mail borne, visual basic script worms, like the recent 'I Love You' virus. At only 1.6Mbyte, 'Zone Alarm 2.1'[D] can be quickly downloaded.

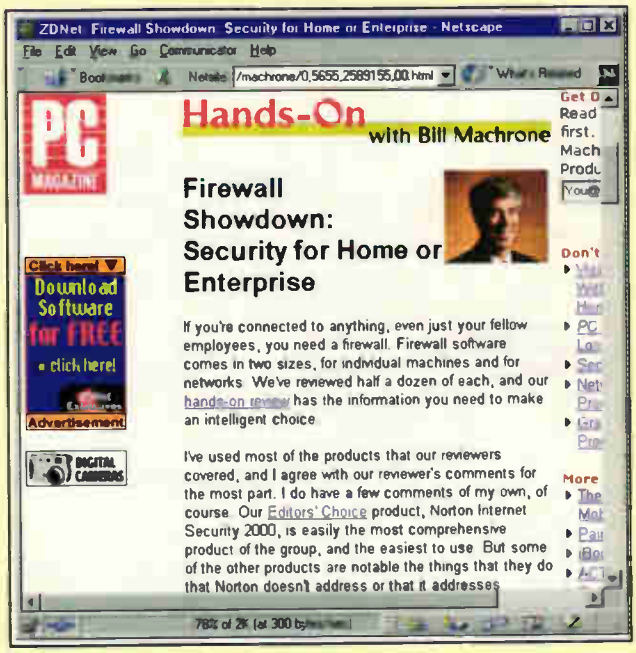


Fig. A. Perhaps you cannot ensure your computer ports are properly controlled and safely closed, or you prefer belt and braces. A software firewall can significantly enhance your safety – especially when you are using a popular operating system.

Bugs references

A. Melissa's Back in Town	http://www.zdnet.com/zdhelp/stories/main/0,5594,2578293,00.html
B. Shields Up!	http://grc.com
C. Firewall Showdown Security for Home or Enterprise.	http://www.zdnet.com/pcmag/stories columnists/machrone/0,5655,2589144,00.html
D. Zone Labs	http://www.zonealarm.com

optimising a design, I now find myself needing to regularly check actual component power levels.

Power measurement

With DC only conditions, simply measuring voltage and current, current and resistance or voltage and resistance provides accurate answers. When AC dissipation is involved, such simple measurements may not suffice.

In a resistive circuit at low frequencies, power level is quickly established. Simply measure voltage and current. Instantaneous power, which is the product of instantaneous voltage and current, however is of little practical value. Apparent power, the product of the RMS values of voltage and current is more useful, especially given small phase angles. What you really need is true power, which applies to all impedances or phase angles.

True power is the mean value of the instantaneous power modified by the actual voltage/current phase angle. True power then is equal to RMS voltage multiplied by RMS current multiplied by the cosine phase angle.

As for DC, AC true power is fundamentally defined by its heating effect. Heat rise measurement is still the optimum method where accuracy is important and especially so for very high frequency measurements. AN64-1 from Agilent Technology, titled 'Fundamentals of RF and microwave power measurements', describes the various techniques which can be used.

For more information...

1. Find chip temperatures with thermochromic liquid crystals	http://www.electronics-cooling. com/resources/EC-Articles/JAN97/jan97_02.htm
2. Analog Devices Inc	http://www.analog.com
3. A Simple Digital Power Meter	http://www.warc.org.uk
4. RMS-to-dc converter is accurate and stable	http://www.ednmag.com/reg/1995/091495/19di4.cfm
5. Make a low-cost benchtop power meter	http://www.ednmag.com/DesignIdeas/September11,1998
6. Real Time Power Measurement of a Transistor	http://www.elantec.com/AN-7

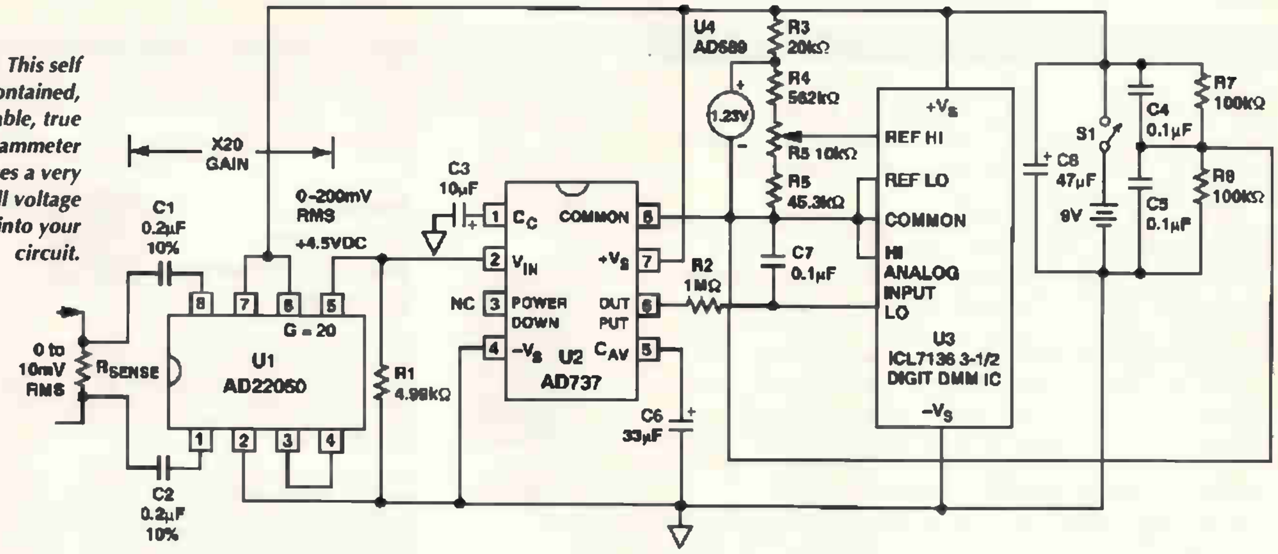
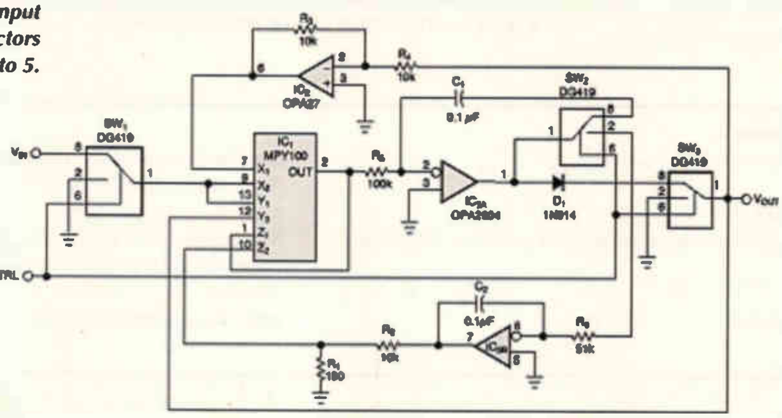


Fig. 2. This self contained, portable, true RMS milliammeter introduces a very small voltage burden into your circuit.

Fig. 3. Self calibrating for time and temperature induced offsets, this true RMS meter can handle up to 2V input and crest factors up to 5.



Alternatively, since the impedance of any AC circuit can be expressed in the form $|Z|=R+jX$, then knowing the value of the 'R' term, true power can be defined as $P=I^2\times R$, where I is the RMS value of the current. Since AC RMS voltage is more easily measured than AC current, you may prefer to use V^2/R.

One frequent need is to measure power in a 50Ω circuit. For this, you need only measure the RMS voltage developed across the 50Ω termination. Many published circuits rely on measurement of voltage developed across a known resistance, both for 50Ω and indeed other impedance values.

1μW to 1kW 50Ω power meter

In the August issue, I mentioned the AD8307 logarithmic amplifier from Analog Devices[2]. It can measure signals from audio frequencies to 500MHz over a 92dB range. This IC, together with five resistors and six capacitors, can be assembled in a small, screened enclosure. With the scaling and components shown, this circuit measures transmitter power in a 50Ω system from 1μW to 1kW at frequencies above 10MHz. This power range represents an input voltage range from 7.07mV minimum to 223V maximum, **Fig. 1**.

A fully worked example power meter[3], usable up to the 70cm band, illustrates this approach. Based on two separate building blocks for ease of assembly, its measuring head is based on this AD8307 logarithmic amplifier. The circuit's display unit uses a PIC14C000 controller with a digital read out.

Measuring RMS

When the signal being measured has a defined waveshape, such as a sinewave, almost any AC meter can be used, but with non-standard waveforms a true RMS responding meter should be used.

One such circuit, which uses the AD737 to measure the true RMS value of AC currents up to high audio frequencies, can be found in Analog Devices' AN-268 publication[2]. This circuit develops a voltage burden of 10mV using a 50mΩ sense resistor to measure 100mA, **Fig. 2**.

Key to the above circuits is the measurement of true RMS voltage or

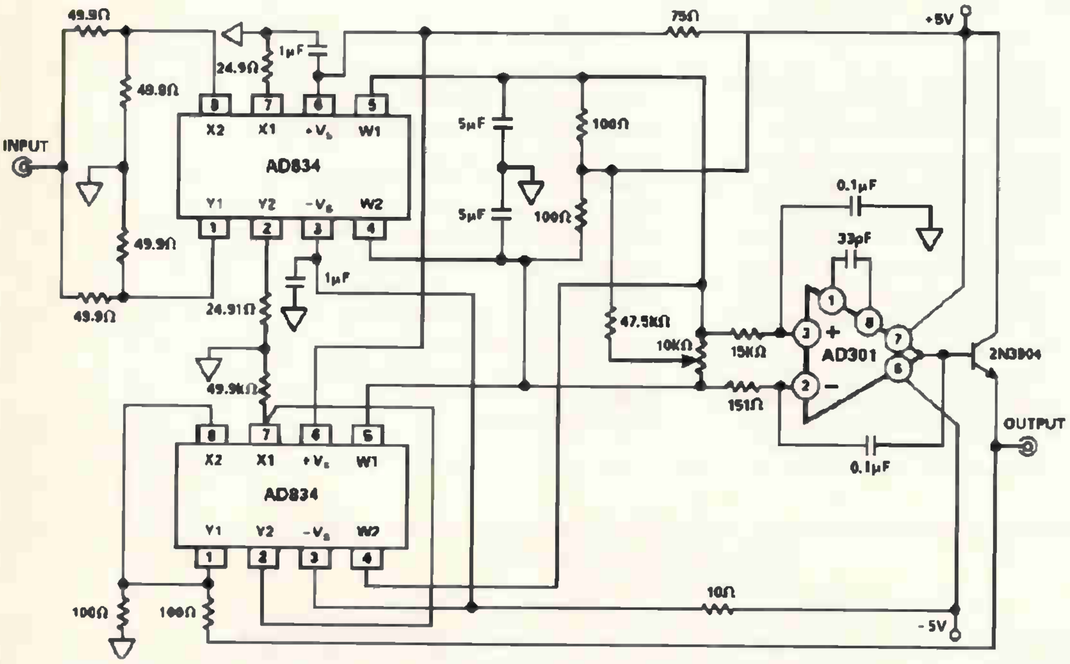


Fig. 4. Using two very fast multipliers both configured as a 'squarer', to measure the true RMS voltage of RF signals, up to +15dBm level, in a 50Ω system.

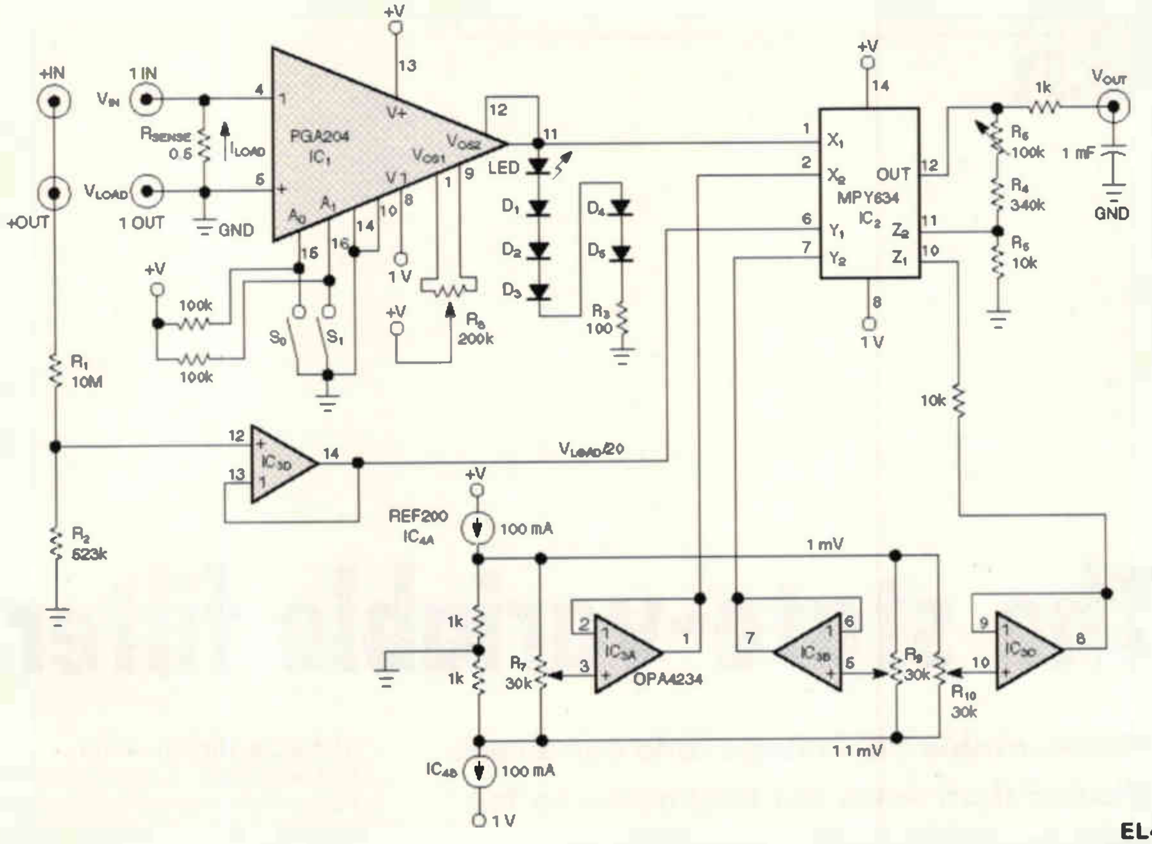


Fig. 5. Described as a 'low-cost bench power meter' this design can measure true power of signals ranging from a few milliwatts to 50W. It is easily calibrated following the published instructions.

current. But many commercial dedicated RMS to DC converter circuits are severely frequency limited. Other approaches based on multiplying chips to first square the input then calculate the mean square root, can be used. With this approach care may be needed to minimise errors caused by offset voltages, which can change with temperature and time.

Sergey Velichko used a sample-and-hold auto calibration technique to eliminate these offset errors, with an MPY100 X, Y, Z-input multiplier[4]. This circuit is able to measure AC signals in the range 0 to 2V with crest factors up to 5. Periodic auto calibration compensates for time and temperature induced drifts. **Fig. 3**.

The AD834 from Analog Devices is a four-quadrant multiplier that can be used with ±1V full scale differential inputs and from DC to above 500MHz[2]. Using two AD834 circuits, both configured as a 'squarer', together with an AD301 integrator, produces a very wideband true RMS meter. With such high frequency capability, component inductive paths must be minimised and an effective ground plane is essential. **Fig. 4**.

True power

Many simple circuits can be used to measure power levels, assuming circuit current and voltage remain in phase or have acceptably small phase differences. Using a multiplier circuit introduces the possibility of measuring true power in circuits when current and voltage phases differ significantly. In other words a multiplier can solve for the expression RMS voltage multiplied by RMS current multiplied by cosine phase angle. This remains valid for any phase angle.

For example one could measure the power output from a 100W audio amplifier by inserting a low-value resistor, say 0.1Ω, in the earth return of the test load, to sense load current. Voltage developed across the test load, multiplied by the load current using a multiplier, calculates the true power dissipated in this load.

Since true power is being measured, the load can be the usual non-inductive resistor or even a reactive test load, such as a multi-element crossover, multiple loudspeaker system.

One worked example[5] demonstrating this technique uses a PGA204 programmable gain in-amp to sense the load current. A unity-gain op-amp senses the voltage across the load. These outputs are then multiplied in an MPY634 multiplier to produce an output, from 10mW to 10W per volt, depending on the PGA204's gain selection, **Fig. 5**.

Using similar techniques it is possible to measure in real time, the true power dissipated in a transistor while it is in normal operation. For this the transistor current is sensed either using a current transformer, or more simply a low value sense resistor in its emitter lead.

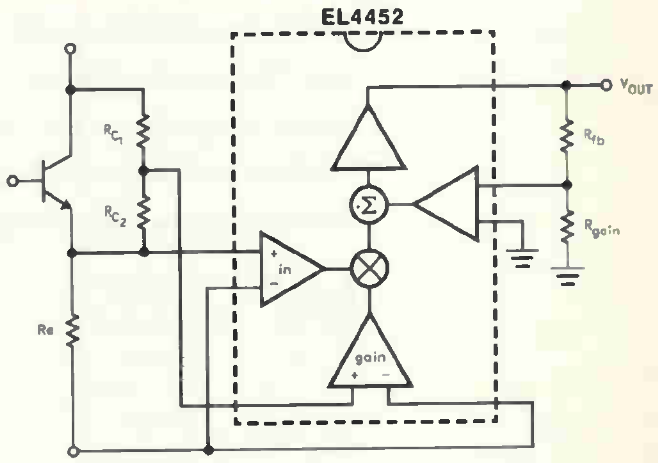


Fig. 6. With a bandwidth of 50MHz, this circuit is able to monitor the actual power dissipated in most power transistor applications, while the transistor is functioning in your design.

The voltage dropped across the transistor is scaled down, if necessary, using a resistive potential divider connected between its collector and emitter. This collector-emitter voltage is multiplied by the transistor's current, using a voltage multiplying IC. With suitable scaling to account for the current sense resistor and any potential divider used, the output represents the true power dissipated in the transistor[6]. See **Fig. 6**.

Having reviewed a few established techniques, my next article examines circuit simplification possible using some of the most recent integrated circuit developments for power measurement. ■

BEGINNERS' CORNER

The state-variable filter

State-variable filters are a little more complicated than most, but they make up for their complexity with versatility. As Ian Hickman shows, the simplest form has three outputs – low, band and high-pass – and a few extra components give you notch and all-pass responses too.

While filters for use at radio frequencies are generally realised with inductors and capacitors, or crystals or surface acoustic wave (SAW) devices, none of these is very convenient for use at audio frequencies.

A series resistor followed by a shunt capacitor, **Fig. 1a)** provides a low-pass response with a 6dB per octave roll-off in the stop-band. The 'corner frequency', where the pass-band ends and the attenuation is 3dB, is given by $F_c=1\div(2\pi CR)$. This is a 'single-pole' response, so called because the transfer function shows a single pole in the complex plane, as explained in reference 1.

A faster rate of descent into the stop-band, at 12 per octave, can be obtained by cascading two such circuits, as in Fig. **1b**), to give a two-pole response. Note that it is advisable to buffer the second stage with a unity gain buffer, for example, an op-amp as shown, to prevent it loading the first.

But the attenuation at the corner frequency F_c is now 6dB. Cascading N buffered sections to obtain $6\times N$dB/octave cut-off in the stop-band would produce $3N$dB attenuation at F_c. So at audio frequencies it is common to use active filters incorporating one or more op-amps. This way, it is possible to design a two-pole circuit to give a 12dB/octave roll-off, but still with only 3dB attenuation – or even less – at F_c.

Higher-order filters, giving 6NdB/octave roll-off but still only 3dB or less attenuation at F_c, can also be designed.

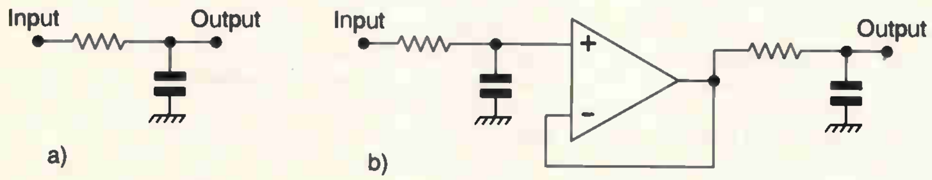


Fig. 1. In a) is a simple passive low-pass – i.e. 'top-cut' – circuit that gives a 6dB/octave roll-off in the stop-band.

Fig. 1b). Circuit b) has two such cascaded sections giving 12dB/octave roll-off, but now attenuation at F_c is 6dB.

The state-variable filter

One of the most important forms of basic two pole active filter section is the state variable filter, or SVF for short. This is perhaps a little profligate in its use of op-amps, requiring three, whereas many other active two-pole filter sections need only two – or even just one. However, the SVF is extremely versatile, and also, unlike some other types, it is 'canonic'. This means that it uses only two frequency-determining *RC* pairs to obtain its two pole response.

The basic SVF provides three outputs, giving low, band and high-pass responses. With a few extra components it will furnish band-stop (notch) and all-pass (phase equaliser) responses as well.

The op-amp integrator

Since the SVF uses two op-amp integrators, it's useful to know how they work.

An integrator is like a water butt; if it rains steadily, the level rises at a constant rate, but it tells you how much rain has fallen in total, since it was empty, whether the rain is steady or comes in bursts.

Similarly, the voltage level V on a capacitor C tells you how much charge Q has flowed into it, the charge being the water and the charging current I the rate of rainfall. So $V=Q\div C$. Capacitance C is in the denominator because the bigger the 'water butt' C, the less the (voltage) level for a given amount of charge (water).

Figure 2 shows the usual sort of

op-amp integrator. together with what happens for the component values and input voltage shown. The op-amp's non-inverting terminal is earthed, and the negative feedback via *C* from the output to the inverting terminal will cause the latter to be at – or at least, very close to – earth potential also.

There is an important oddity to note, however. An ideal integrator with a constant +1V input would result in the output voltage rising at a constant rate. But with the circuit arrangement shown, a constant positive input results in an output voltage going steadily more *negative*. Thus the circuit is an 'inverting integrator', due to the input being connected to its inverting input.

Non-inverting integrator circuits exist, such as the De Boo integrator, but they are a little more complicated than the circuit shown.

Figure 2 also shows what happens when the input is a sinewave at the frequency $F_c=1\div(2\pi CR)$, where the reactance of the capacitor, in ohms, equals that of the resistor. Since the current flowing in the capacitor is equal to the current in the resistor, the gain is unity, but the output sinewave is at 90° – i.e. in quadrature – to the input. In an ideal integrator, it would be lagging by 90°, but due to this circuit's inverting property, it appears to be leading by 90°.

In practice, integrators can be tricky things to handle. Any unintentional input, such as the op-amp's own input offset voltage, will be integrated for ever more. The output thus has an annoying tendency to wander off to one supply rail or the other. while one's back is turned. So integrators are frequently used within an overall negative-feedback loop to prevent this, and the SVF is an example of just such a circuit.

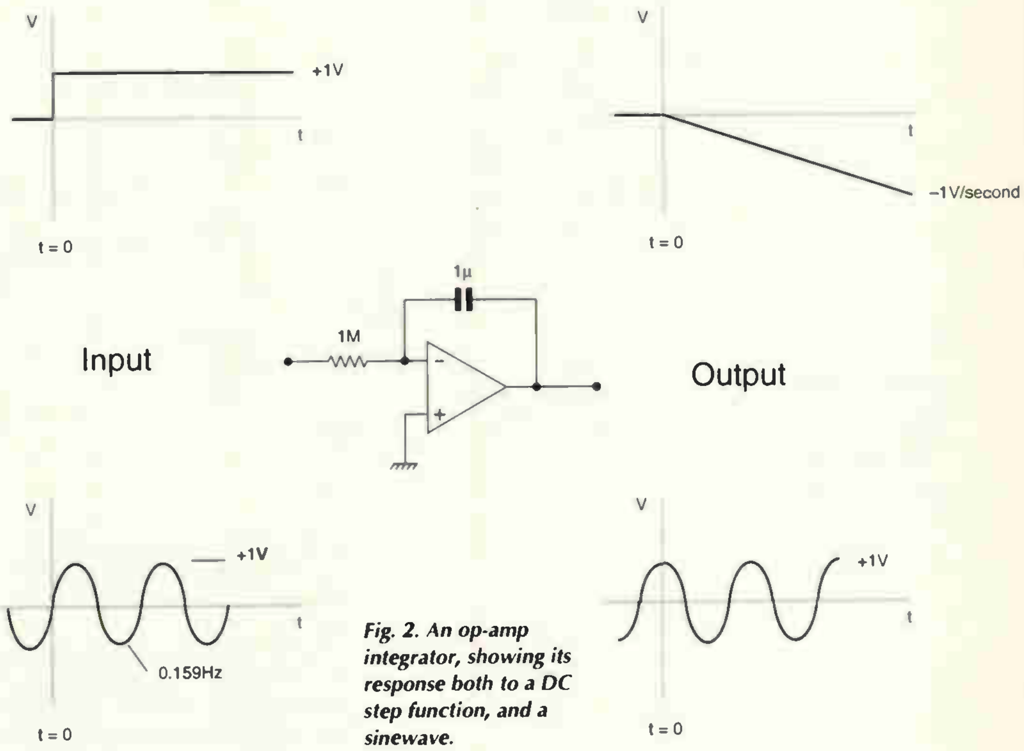
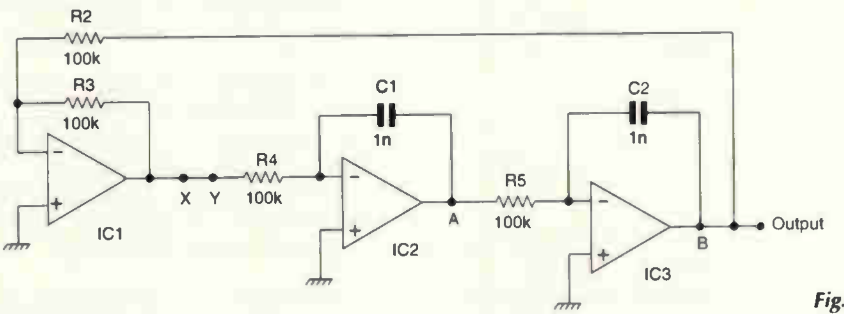


Fig. 2. An op-amp integrator, showing its response both to a DC step function, and a sinewave.

Filter or oscillator?

Figure 3a) shows two integrators connected in a loop with a unity gain amplifier. As all three circuits are inverting, the overall feedback is negative, and all three outputs shown will sit at 0V earth potential – on average.

Figure **3b)** shows the vector diagram. This diagram assumes that there is a sinewave of 1V peak amplitude and frequency $F_c=1\div(2\pi CR)=1.59$kHz, at point



Y. For the moment, imagine that it is disconnected from point *X*. It turns out that there is an identical voltage at point *X*, so the circuit would be capable of supplying its own input.

The same argument applies, whether you assume a voltage at *Y* of 2*V* peak, or of zero volts. The circuit is in fact either an oscillator, or a filter with infinite *Q* – the same thing. In practice, due to residual excess phase shifts in the circuit, it will almost certainly oscillate, the amplitude building up until the peaks of the sinewave hit the supply rails.

By making R_2 and R_4 equal to 91kΩ instead of 100kΩ, the clipping will take place in IC_1, resulting in rather a pure sinewave at the low-pass output, output of IC_3, assuming

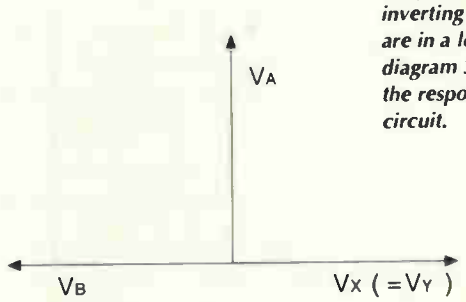
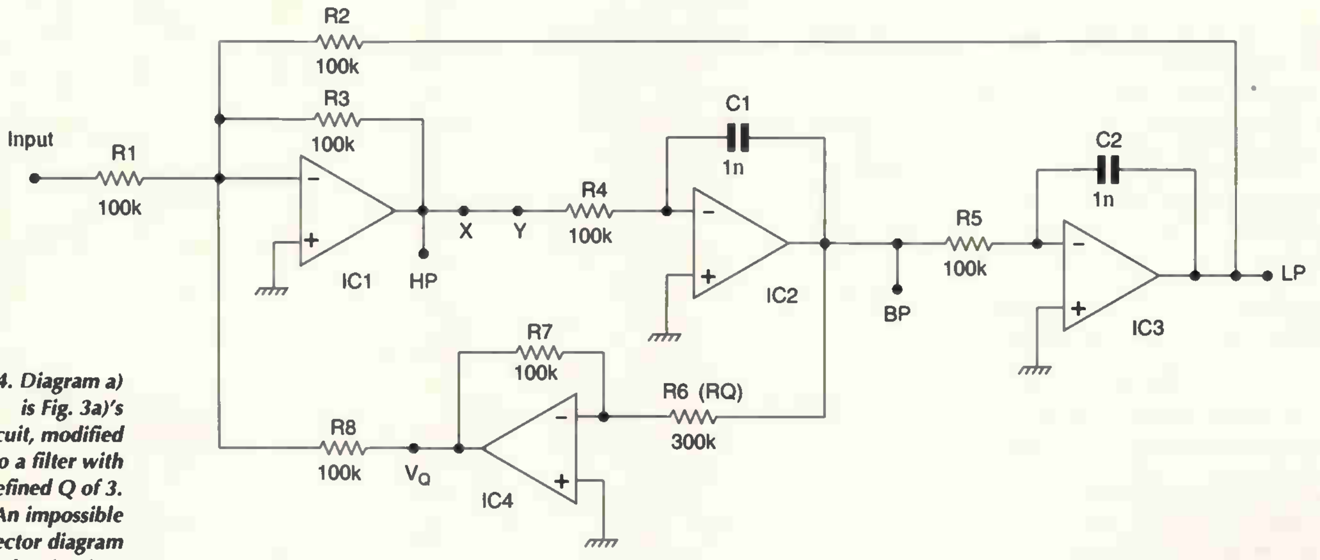


Fig. 3. In a), two integrators and an inverting amplifier are in a loop. Vector diagram 3b) shows the response of the circuit.

C_1 really does equal C_2. If the output of the third op-amp is still clipped, try interchanging C_1 and C_2.

Making it stable

Adding just three resistors can turn the circuit into a sanitary. designable filter of defined *Q*. But



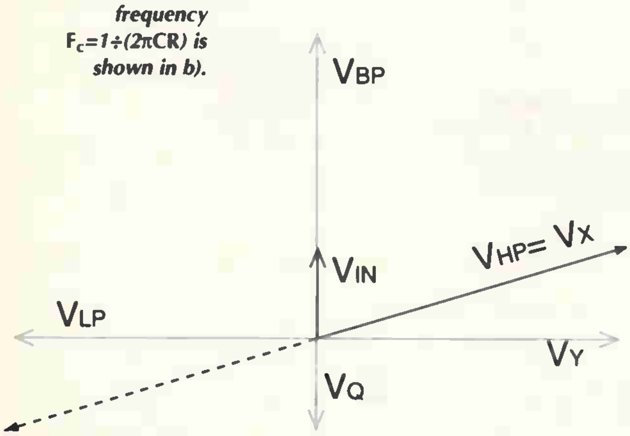


Fig. 4. Diagram a) is Fig. 3a)'s circuit, modified into a filter with defined Q of 3. An impossible vector diagram for the circuit at frequency $F_c=1\div(2\pi CR)$ is shown in b).

as op-amps come four to a package, the spare one can be put to good use as in **Fig. 4a)**, which incidentally makes it simpler to explain just how the circuit works.

Now, besides a band-pass output, there is a 1/3-scaled inverted band-pass signal also available. Feeding this back into the summing amplifier provides a defined Q for the filter, as illustrated in the vector diagram, Fig. **4b)**, where, with the value of R_Q shown, the Q is 3.

In Fig. 4b), I have deliberately shown a vector diagram which is impossible, if you imagine for the moment that V_{in} is zero. Assuming as before that there is a sinewave V_Y of 1V peak amplitude and frequency $F_c=1\div(2\pi CR)=1.59$kHz, at point Y, then V_{BP} and V_{LP} are as shown. But there is also V_Q – an antiphase version of V_{BP}.

The amplitude of V_Q is, with the circuit values shown, only one third that of V_{BP}. Added to V_{LP} (the dashed vector shows the sum) it is applied to the unity gain inverting amplifier IC_1, resulting in an output $V_{HP}=V_X$.

This is manifestly not the same as V_Y, so the vector diagram cannot be right, as X and Y are the same point. But imagine that an input V_{IN}, equal in amplitude but opposite in phase to V_Q, is applied. Now, V_{IN} and V_Q cancel each other out, so the dashed vector is coincident with V_{LP}, and so $V_X=V_Y$.

The higher R_Q, the smaller V_Q and so the smaller the input relative to the three outputs – i.e. the higher the gain at F_c. The result would have been the same if R_6 and R_8 had been interchanged, and in fact that is how the four op-amp SVF is usually designed.

Well below F_c the low-pass output gain is unity and the high-pass gain rolls off at 12dB/octave as the frequency falls. Conversely, well above F_c the high-pass output gain is unity and the low-pass gain rolls off at 12dB/octave as the frequency increases. The band-pass gain rolls off at 6dB/octave either side of F_c except in the vicinity of F_c, where the roll-off is steeper, the higher the Q.

Combining the high- and low-pass outputs provides a notch filter, since at $F_c=1\div(2\pi CR)$, these two output are equal in amplitude, but in antiphase. The higher the Q, the narrower is the notch. Combining the high-pass and low-pass outputs with 1÷Q multiplied by the bandpass output provides an all-pass filter. Its gain is independent of frequency, but the output changes from leading the input by 180° at low frequencies, through in phase at $F_c=1\div(2\pi CR)$, to lagging by 180° at high frequencies. The higher the Q, the more rapid the phase change with frequency in the region of F_c.

Experimenting

Try making up an oscillator as in Fig. 3a); hint: in the unlikely event that it does not oscillate, try connecting a capacitance of a few 'puffs' (picofarads, or pFs) in parallel with R_3.

This circuit can be used as an audio-frequency source. Now make up a tuneable filter, as in Fig. 4a) but using 1.5nF capacitors, and a 100kΩ twin gang potentiometer, used as two variable resistors, in place of R_4 and R_5.

Connect the output of the oscillator to the input of the filter. Monitor the filter's band-pass output with an oscilloscope, and adjust the two-gang potentiometer to 'tune in the signal'. You will find the filter overloaded, with the output waveform severely clipped.

Increase the input resistance R_1 of the filter to 1MΩ, and you should now find that the filter handles the signal without overloading. Try the effect of varying R_6 (R_Q), and see how the centre frequency gain increases as Q is increased. This is one facet of the SVF that can be less than useful.

Another related filter design, the biquadratic filter, handily provides a centre frequency gain that is independent of Q, although the down side is that it provides no high-pass output. You should be able to find details of this and other active filters in most good electronics textbooks, such as reference 1. ■

References

1. Analog Electronics, 2nd Edition 1999, Ian Hickman, Butterworth-Heinemann, ISBN 0 7506 4416 8.

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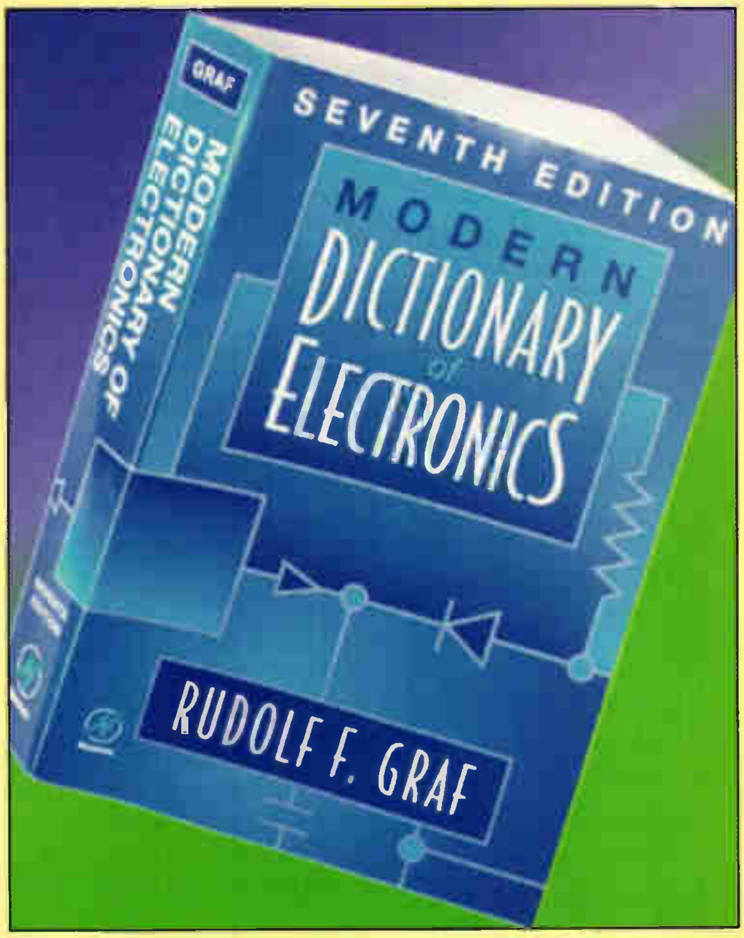

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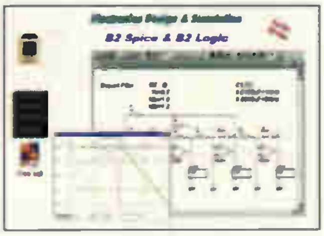

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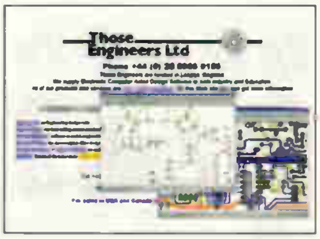

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Following on from the newsgroup discussion last month there is a UK Email group for TV technicians where you can send an Email to everyone in the group. There's just over 30 people in the group at present. For more details and how to register look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

REED CONNECT

http://www.reedconnect.net/

Another free internet access site, this time from Reed Business Information. However the site possesses a useful UK People and Business Finder, with an e-mail search. There's also business news and local information, and some good links to directory sites.

REPAIRWORLD

http://www.repairworld.com

Repairworld is a sophisticated US based fault report database which is updated bi-weekly. It operates on a subscription basis and describes itself as an "affordable solution for all technicians". There is apparently no minimum number of months for which you have to subscribe. You can see some samples of the material for free, monitors, VCR, DVD and Camcorders being of particular relevance to UK users. The site even provides a "chat room" where you can talk via your keyboard to others "in the room".

Letters to the editor

Letters to "Electronics World" Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS e-mail jackie.lowe@rbi.co.uk using subject heading 'Letters'.

IMD versus THD

I enjoyed Mr New's article, 'THD is meaningless,' in the June issue questioning the relevance of total harmonic distortion as a figure of merit. Nonetheless, any suggestion that intermodulation distortion is a more objectionable distortion mechanism than harmonic distortion is misplaced, since the two simply measure the same phenomenon: the deviation of the amplifier from linearity.

Suppose we accept, for simplicity, a memoryless non-linear model for the amplifier. In this model, we attempt to write the output signal as a polynomial function of the input signal. If we inject a pure sinusoid at the input, measure the gain of the amplifier and the strengths of the different harmonic overtones, then we can determine the coefficients of our polynomial model.

From this model, we can accurately predict the intermodulation terms that will result from injecting two sinusoids at the input.

It works the other way around too: on measuring the amplifier gain and all the intermodulation products in the two-tone experiment, we can again determine the coefficients of our polynomial nonlinearity. Therefore we can predict accurately the harmonic spectrum that will result from injecting a single sinusoid.

This is to say that harmonic distortion and intermodulation distortion are not distinct mechanisms, but rather different techniques to measure the very same mechanism, which is amplifier nonlinearity.

Therefore, harmonic and intermodulation distortion measurements are equally valid, or equally irrelevant, depending on one's position.

Mr New is quite right in defending psychoacoustics as a legitimate consideration in audio reproduction. This is not to be confused with so-called 'subjectivist' views that are derided regularly by Mr Self and his disciples.

Harmonic and intermodulation distortion, for example, are based on steady-state amplifier behaviour to a sinusoid, or two, at the input, i.e., after the transients have settled. These measurement techniques are therefore 'deaf' to transient response features of an amplifier.

There can be little doubt that transient detail is essential in high fidelity audio reproduction: drums and percussion, for example, don't resemble sinusoidal test signals. Neither do the initial attack waveforms from a competent violinist or guitarist.

Indeed, it is unclear how suitable the sinusoid is as a test signal for music reproduction, given that no major musical works seem to have been composed for the sinusoidal oscillator.

Transient response features and time-domain behaviour, regrettably, are harder to quantify in audibly meaningful terms.

The relationship between impulse response and frequency response is valid only if the amplifier is indeed linear.

Rise-time or slew-rate figures are of some use, but cannot constitute a complete description of an amplifier's transient response.

Such figures do not explain, for example, whether microphonics in vacuum tubes – however well controlled – have any impact on musical transients. Nor do they explain whether nonlinear Miller capacitors in the form of reverse-biased p-n junctions alter the trajectory of a transient waveform.

Sinusoidal or 'steady-state' tests are deaf to transient influences, but are we sure that listeners are?

Finally, I should point out that the memoryless polynomial non-linearity referred to above, although a good 'mid-band' approximation to a real amplifier, is ultimately a white lie.

One may show that any amplifier having a finite bandwidth cannot be memoryless, i.e., the output signal at a given point in time depends on the past history of the input signal.

Any amplifier exhibiting a non-zero rise time must likewise have memory.

A treatment of nonlinear models incorporating memory is beyond the scope of this letter. That would take us into Wiener, Hammerstein, and Volterra models, among others.

Note, however, that a single sinusoid – or even two – is generally insufficient as a test signal for determining the parameters of a nonlinear model with memory, no matter how accurately one could measure the amplifier output.

Therefore, knowledge of the behaviour of a practical nonlinear amplifier in response to a sinusoid is ultimately insufficient to predict its sonic qualities.

Two amplifiers with the same harmonic spectra can respond to transients in distinct manners. That would not be the case with memoryless nonlinearities.

To further the debate on what is audible versus inaudible distortion, I might suggest you commission some articles describing the MPEG layer III audio coding scheme, and why its sound quality is fairly 'good' despite its distortion.

Professor Phillip Regalia
Institut National des Telecommunications
Evry
France

I liked Anthony New's article very much, but it led to a very odd conclusion when I read the assertion that to add a few percent distortion to a note from an instrument is merely to brighten its timbre.

The little known fact is that instruments have true harmonics. Take the clarinet for example. Its reed modulates an airstream in a particular waveform, so that harmonics are locked to the fundamental.

Others, such as the piano, have strings with different modes of vibration that are slightly above the harmonic frequencies pitchwise. So the tuning sounds better if octaves are 'stretched' a little.

Tests have shown that listeners expect these overtones to be out of tune in the lower bass notes at least. Now if the note has harmonic distortion added by an amplifier, the genuine harmonics generated for the fundamental frequency will beat with the overtones to give very low frequencies, thus fulfiling the definition of a discord.

No classical composer writes a third or a fourth down below the bass clef, since the discord is as bad as a semitone discord.

Bernard Jones
London

In his article on THD, Mr New states that he finds it hard to believe that anyone can hear small amounts of distortion at 10kHz. I believe that even a large amount is inaudible at that frequency. If you doubt me, carry out the following simple experiment.

Connect an AF signal generator to an amplifier's input and a speaker to its output. Set frequency to 1kHz and the waveform to square. Adjust the volume to a normal listening level, switch to sine waveform and note change in timbre.

Increase frequency in increments of 1kHz and repeat the test. You will reach a frequency at which it is impossible to distinguish between sine and square wave.

We read in the books that distortion sounds worse at HF, but this experiment would tend to to indicate the opposite,

This seems to confirm Mr New's theory that intermodulation is the most important factor.

John Ellis
Birmingham

I read Anthony New's article with interest, as it is always stimulating to have someone come in from outside your speciality and tell you you've got it all wrong.

I have probably spent more of my life doing THD tests than most people, and it would be disconcerting to learn at this stage that it was all labour lost and work in vain. I need not have worried.

The point that seems to have

totally eluded Mr New is that no-body claims that THD is a direct measurement of audio degradation. It clearly isn't. We do not listen to sinewaves for our entertainment, though they would probably be more interesting than some television programs.

The point about THD is that it is a highly effective measurement of the linearity of an amplifier. It needs relatively simple equipment to determine, and gives a visual output that often permits linearity problems to be diagnosed at a glance.

If an amplifier has low THD across its working frequency range, then it must be highly linear. It will also have low intermodulation products. I must admit that I thought this was too obvious to need spelling out.

The real beauty of the THD test is that the residual of the output signal – i.e. what is left after you have notched out the fundamental – is basically the difference between perfection, i.e. no residual, and whatever damage your circuitry actually does. This residual's waveform provides a wealth of information on what the problem is.

Listening to music is another matter altogether; you don't get one frequency, you get lots at once.

A solo instrument is likely to have at least 50 harmonics in the audible range. A small musical ensemble may provide four instruments and two voices, which gives us about 300 frequency components.

The possibilities for intermodulation are clearly great. I recall reading many years ago (though I cannot, I'm afraid, locate the reference) that in general the level of intermodulation products was likely to be some 50 times above the harmonic components.

As a quick thought experiment, consider how the extra components generated build up as the situation gets more complex.

Two sinewaves applied to a circuit with second-order nonlinearity produce two second harmonics, and one intermodulation product. Total: 3 components.

Three sinewaves applied to a second-order nonlinearity produce three second harmonics, and three intermodulation products. Total: 6 components.

Three sinewaves applied to a third-order nonlinearity produce three third harmonics, and 16 intermodulation products. Total: 19 components.

The number of intermodulation components clearly builds up quickly. It looks quite plausible that their level could exceed that of the harmonics by a factor of fifty in a real situation with many instruments and more complex non-linearities.

Maybe I should write an article about RF technology.

Doug Self
Via e-mail

The June 2000 issue features a polemical rant by Anthony New, which alleges that THD measurements are irrelevant and irrational. Mr New then enlightens us as to what IMD is, its importance, and how to measure it.

Although a few readers may find that this is all startling news, it has to rankle even novice audio engineers to be told about measurements that they routinely perform, and the value of them, by an RF engineer who writes as if he had just discovered them.

Far from unusual, IM distortion plots are even ***de rigueur*** in such primarily subjectivist audio publications as *Stereophile*[1]. The idea of applying IMD measurements to audio is not New's – or news.

Mr New acknowledges that THD can be considered a subset of IMD. He also goes on about the distribution of harmonics and the lack of such information in the single THD number. Again, even the glossy popular magazines are showing complete spectra of harmonic distortion these days. This is also not news.

The utility of THD is in its ease of measurement, and that it tells us about nonlinearity and its dependence on levels and frequencies. It is by no means useless.

With the exception of pathological cases, a signal with a given spectrum that is input to a nonlinear system will produce an output with many additional spectral components. If we see significant THD, even only as a single number, we know that there is work to do. Knowing the time and frequency domain characteristics of the distortion can help to tell us where to begin.

One point that Mr New does contend about the, "apparent subjective differences," among amplifiers I can heartily agree with: that they are mostly due to differences in frequency response.

He may be surprised to know that these have been frequently referred to as 'linear distortions' in the audio literature. He does not mention what is typically responsible for such variations, namely the interaction of the frequency-dependent load impedance with the output impedance of the amplifier.

Valve amplifiers are notorious for high output impedance compared to sound solid-state Class-B designs. Now that some switching amplifiers are getting audiophile attention, it is tempting to conjecture that the interaction of the load with their output filters – filters required for reduction of RF emissions – accounts for the reported difference in sonic character.

It is a human trait to hear 'different' as 'better', unless the difference is simply awful ("...the amplifier revealed nuances and details in selection X that I'd never heard before...").

Furthermore, knowing in advance that the amplifier under evaluation is unusual in some way to which one is favourably predisposed – for example, 'all-digital' – almost guarantees a rave review.

I suppose that it is a good thing that Mr New's article may help to educate those hitherto unfamiliar with intermodulation distortion. But he displays ignorance and arrogance when he implies that the, "parochial and fashion-conscious world of audio," does not comprehend or appreciate it.

Brad Wood
Chatsworth
California

1. *Stereophile*, Jan 1997, Vol. 20 No 12.
2. I'm referring here to the small signal use of feedback in gain blocks and equalisers. Power amplifiers driving non-linear reactive loads seem to be the bone of contention.
3. Total harmonic and intermodulation distortion.

Anthony New argues that it is hopeless to describe an audio amplifier by specifying its total harmonic distortion. He proposes that instead intermodulation distortion (IMD) be used.

One is left wondering just what test procedure to follow. After all, the results of an IMD test will surely depend on the frequencies and amplitudes of the two – or more – input signals. For an IMD test to be useful in comparing amplifiers, some kind of standard will be needed.

Help is at hand. A book by Terman and Petit – namely Electronic Measurements, McGraw-Hill, 2nd ed. 1952 – has a section on nonlinear distortion in audio frequency amplifiers. According to Terman and Petit, of the many combinations of signals that can be used for an IMD test, two in particular merit attention.

One is the method used in the movie business, standardised by the Society of Motion Picture and Television Engineers (SMPTE). Here, a low-frequency signal, at frequency f_1, and a high-frequency signal, at f_2, of somewhat smaller amplitude are applied. The IMD is then expressed as the square root of the sum of the squares of the modulations m_1, m_2 and so on, where m_1 is the degree of modulation of f_2 at frequency f_1, and m_2 is the degree of modulation at $2f_1$, and so on.

The values used are 60-100Hz for f_1, and above 5kHz for f_2, with the amplitude ratio set to 4. The authors say that the actual value of the higher frequency is not critical, and the IMD will vary with f_1 if this is in a region where the amplifier response is falling off.

Another way of specifying the IMD from the same test is offered, based on the amplitudes of the sum and difference frequencies produced. In this method, the amplitudes of the various frequency components are used. This approach might be more straightforward in this day of FFT.

In another method of performing an IMD test, the two test signals are equal in amplitude, and relatively high in frequency. This is the method of the CCIF (International Consultative Telephone Committee, but in French).

The distortion is described by the ratio of the amplitude of the difference frequency signal in the output to the sum of the amplitudes of the signals at the input frequencies. Again, this would seem to lend itself to simple calculation.

The frequencies are not specified, but it is pointed out that the difference frequency should fall in the range of the amplifier (Terman and Petit suggest between 60 and 200Hz). The method can be used to explore the IMD-producing nonlinearity of an amplifier's HF response.

Terman and Petit go on to compare the various methods. For a single-ended amplifier (those were the days!) operated in a region where there is little distortion and the frequency response is flat, they say that the SMPTE IMD figure will be 3.5 times the THD figure, and the CC IF IMD figure will be half the THD figure.

It is also stated that when the nonlinear distortion is different at different frequencies, these simple ratios do not apply, and the correlation with listening tests is lost.

Terman and Petit go on to say that no one test correlates with listening tests over a wide variety of conditions. If a single test is to be made, an intermodulation test is usually more useful than a harmonic distortion test. Some examples are given: THD of about 1% can be detected by listening tests, but is not serious until about 10% (these were the days of valves, remember)

Even larger harmonic distortion can be tolerated if the frequency is below 100Hz or above 4000Hz. Intermodulation distortion cannot be detected by the SMPTE method until about 10% if the LF tone is below 100Hz. If the LF is a few hundred hertz, 3 or 4% can be heard.

Interestingly, IMD by the CCIF method can be heard at levels of a small fraction of a percent, and becomes objectionable at values of 3 to 4% when the difference frequency lies in the range 400 to 5000Hz – where the ear is most sensitive.

For those of you interested in tracking down the original material, try the following:

'Intermodulation testing,' by John K. Hilliard, *Electronics*, Vol. 21, p 123, July 1946.

'Distortion Tests by the Intermodulation Method,' Proc IRE, Vol. 29, p. 613, Dec. 1941.

'Instrument for Intermodulation Measurements,' *Electronics*, Vol. 21, p. 134, March 1948.

'Measurement of Nonlinear Distortion,' *Gen. Rad. Expt.* Vol. 25, March 1951.

'American Standard Method of Making Intermodulation Tests on Variable-density 16-millimetre Sound Picture Prints,' *J. SMPTE*, Vol. 46, p. 303, April 1946.

'Audible Audio Distortion,' H. H. Scott, Electronics, Vol 18, P 126, January 1945 (covers CCIT method).

'An Analysis of the Intermodulation Method of Distortion Measurement,' W.J. Warren and W.R. Hewlett, Proc. IRE, Vol. 36, p. 457, April 1948 (and yes, that is Bill Hewlett of H-P fame!)

It certainly seems that author New is right, and IMD tests would be more meaningful than THD. I hope to see some results of such tests applied to modern amplifiers published in the pages of *EW/WW* in the not-too-distant future.

Harold Kirkham
Jet Propulsion Laboratory
Pasadena
California

Anthony replies...

I seem to have stirred up a hornet's nest with my views on the value of THD as a measurement tool – perhaps not surprisingly!

Although some may disagree with my conclusions, I think the resulting discussion is a good thing if it forces those inside the amplifier design industry and those outside it to reassess their methods and yardsticks.

I am well aware, by the way, of the use of IMD figures in the better hi-fi reviews over a very long period; an earlier draft contained references for this, which I edited out for brevity.

However I believe my comments are justified so long as hi-fi reviews still quote as an important figure of merit THD figures at levels below 0.01% – which are completely meaningless except as a proxy for something else, and so long as designers produce amplifiers in which frequency response errors cause more distortion in practice than are claimed in theory.

The adherents of THD conveniently ignore that THD – under conventional single-tone sinewave tests at least – does not measure the difference between the output and input signals directly. They also ignore the fact that that there is significant evidence of important differences.

If any of them chooses instead to design multi-carrier RF amplifiers, one or two might be deeply shocked to find that their customers expect the performance to be tested and certificated by representative multiple-carrier signals rather than convenient but unrepresentative sinewave ones.

Why should the discerning hi-fi buyer accept anything less?

Let us not forget that linearity is not an end in itself, and that our choice of measurement method itself needs to be validated if the conclusions are to stand scrutiny.

Anthony New
Wireless Systems International
Bristol

Blowing a trumpet for the horn

I have been following John Watkinson's series of articles on loudspeakers with some interest. His comments on speaker polar pattern and placement have been particularly interesting.

Interesting that these recommendations invalidated much of the wisdom of the 70s and 80s on speaker placement, which advocated the 'extended headphone' model. This involved placing speakers directly in front of the listener to minimise reflections and improve imaging while playing down the role of reverberant energy as a useful attribute.

I have recently reread two papers on modulation distortion in loudspeakers by P. W. Klipsch, *AES*, the second dated February 1970.

Traditionally Klipsch has advocated the horn-loaded loudspeaker as the only practical way of achieving realistic sound-pressure levels and very low levels total modulation distortion* and wide dispersion. Mr Klipsch also advocates the use of corners in a room mainly for his low-frequency horn design – the Klipschorn – but recommends it for any loudspeaker to improve bass.

These horn-based designs seem to fulfil the requirements proposed by Mr Watkinson regarding polar patterns and the generation of reverberant energy which he suggests sets good and bad designs apart.

Interestingly, on this point, a mid eighties American review commended the Klipschorn for sounding so realistic and natural when listened from the room next door. This was attributed precisely to the reverberant energy produced by its placement and polar pattern.

Notwithstanding all this, horns have not attracted a general following. They are certainly not commonly found in contemporary high end products – very true in the British scene with match box living rooms. Why is this?

The usual answer is that horns sound, well, like horns. This is too simplistic an answer and not true in my experience.

In the above paper Klipsch shows that the horn can produce at least ten times less total distortion than a direct radiator at at least 10dB higher sound-pressure level, obviating the need for high-power amplification.

Indeed, only about 3W of audio power is all you need with the Klipschorn to produce near live sound levels in a domestic setting.

Given the all merits of the horn, direct radiator designs are paradoxically too prevalent. The cost/size argument would only apply to bass reproduction. Horn mid-range speakers and tweeters are of comparable cost and size.

Can I ask Mr Watkinson for an update on the horn versus direct radiator issue? ■

George Evans
Lymington, Hampshire

*Frequency plus amplitude modulation distortion

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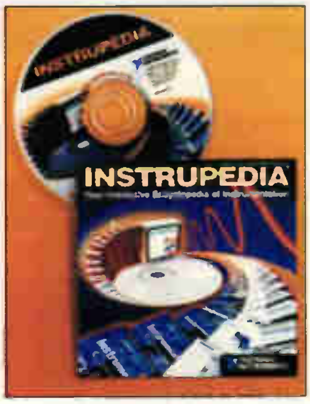

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