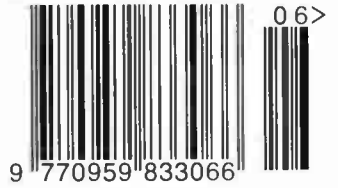


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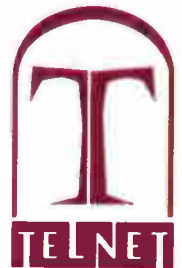
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CONTENTS

433 COMMENT

Something more than a virtual future?

435 NEWS

- Scottish microdisplay initiative
- IBM chips go a third faster
- Maps on your mobile phone
- Mobile-mast caution
- E-mail phone
- Copper IC processing

438 DOTTY.COM...

Dot.commers have become such a constant source of fun and amusement that we will never be able to take them seriously again **David Manners** reports.

440 THD IS MEANINGLESS

Looking at audio amplifiers from an RF designer's perspective **Anthony New** argues that THD figures are "irrelevant, irrational, and completely spurious." So what's the alternative?

448 RS232 RADIO LINK

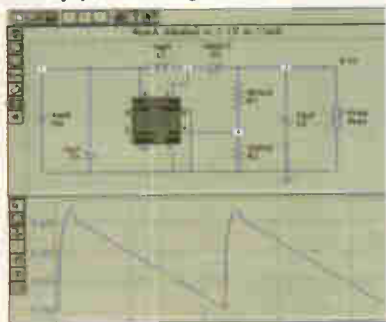
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456 CIRCUIT IDEAS

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- Two transistor audio-visual alert
- Mains flasher
- Multichannel amplitude discriminator
- Alternative neon tester
- Simple FM broadcast receiver

460 EFFICIENT BATTERY POWER SUPPLIES

Cyril Bateman shows how important subtle capacitor parameters are in gaining maximum efficiency from battery-powered regulators.



466 BEGINNERS' CORNER

Having come across electronics students that had had no experience of making and troubleshooting electronic circuits, **Ian Hickman** decided to show them how. Here's where he began.

469 NEW PRODUCTS

New product outlines, edited by **Richard Wilson**

480 JLH A LIFETIME IN ELECTRONICS

John Linsley-Hood recalls the emergence of the IC and his first experiences with PLLs, the synchrodyne and cassette recorders.

486 A NEW 100W CLASS-B TOPOLOGY

In a conventional Class-B amplifier, distortion rises with frequency. But it's at higher frequencies, where the ear is most sensitive, that you want the best performance to suppress the undesirable influences of cross-over switching.

Russel Breden believes his reconfiguration 100W Class-B design solves that problem.

492 BECOME A TRAPEZIUM EXPERT

There's any number of circuits for generating square, sine and triangular waveforms, but how often do you see a circuit for producing trapezium waveforms? **Anthony Smith** explains not only how to make trapezoidal waveforms, but also reveals why they can be so useful.

499 CALIBRATOR FOR 4-20MA INTERFACES

After reeling at the price of a calibrator for 4 to 20mA loop interfaces, **Darren Heywood** decided to look into designing his own.

502 WEB DIRECTIONS

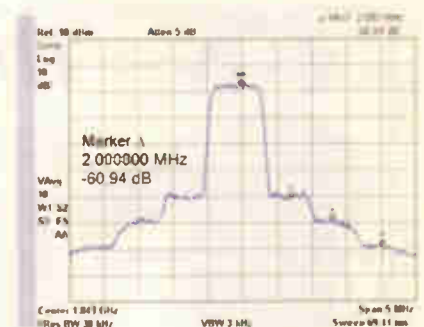
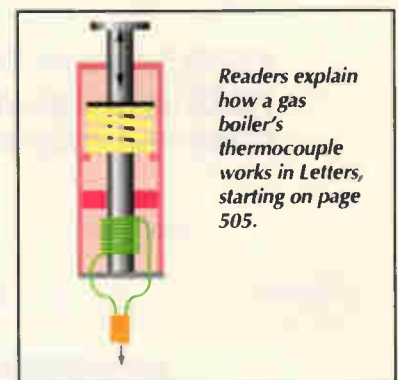
Useful web addresses

505 LETTERS

Audio power analysis. Easily-bared ends, Photodiode sensing. In defence of privatisation, Domestic thermocouples, Blumlein.

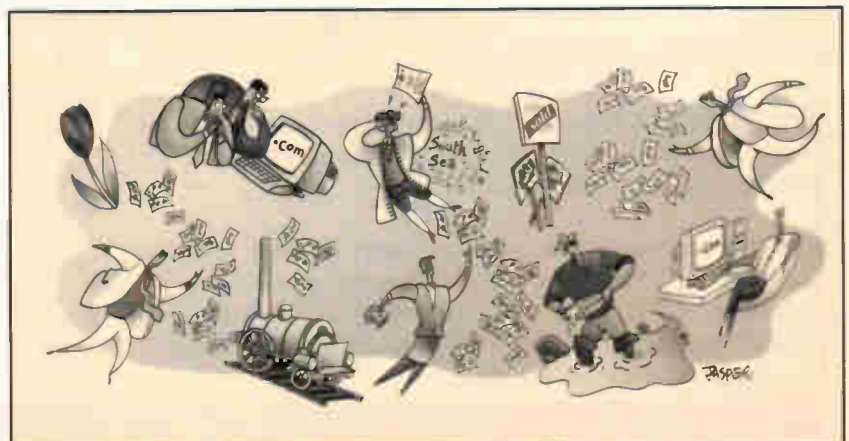


Photography : Mark Swallow



As a means of judging an audio amplifier's performance, THD is pretty useless argues **Anthony New**. So what's the alternative? Find out page 440.

Are dot.com companies taking the mickey out of the financial world? **David Manners** says they are, on page 438.



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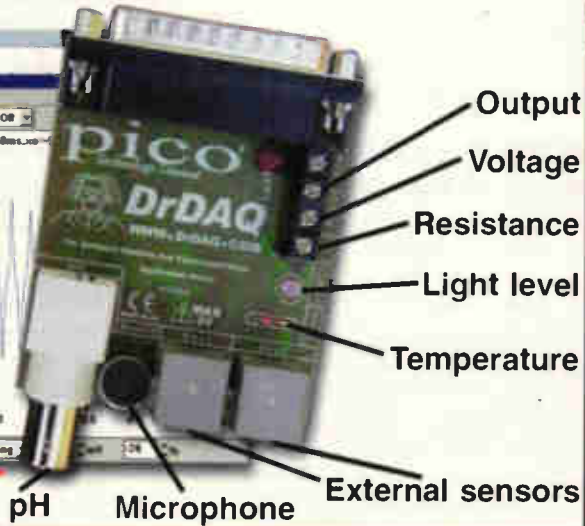
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Something more than a virtual future?

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The UK electronics industry is making a welcome return to a practice it was once particularly good at. That is manufacturing ground-breaking products. This news that the industry is returning to its manufacturing roots could not have come at a more opportune moment.

The electronics sector has been dragged into the much-hyped dot.com stock market investment phenomenon. That's good news for those individuals lucky enough to find themselves turned into an over-night millionaire on the strength of a stock market flotation. But it is not doing much for the longer term strength and competitiveness of the electronics industry as a whole.

Indeed the more thoughtful entrepreneurs in the Cambridge start-up belt fear that the credibility of the high-tech start-up could eventually be damaged by the misplaced faith of ill-advised City investors in the dot.com phenomenon. The good news is that behind all the smoke and mirrors created by dot.com mania there are encouraging signs of something more permanent growing in the grass roots of the industry's manufacturing sector.

Just before Easter an Oxfordshire-based firm Bookham Technology was valued at £2.9bn in its initial public offering (IPO) on the London Stock Market. Nothing too surprising about that in the current climate, except that Bookham is an electronic component manufacturer.

Bookham's value as a business is based on a technique for manufacturing low cost opto-electronic components used in optical fibre communications networks. It is not surprising that a group of UK-based engineers should come up with a world-class semiconductor component technology.

However, it is surprising that a UK-based start-up has actually made a success of manufacturing its products and is selling them to blue-chip multinationals around the world. What is even more surprising is that the City investors have recognised the potential in a company like Bookham Technology, without being frightened away by the inevitably high up front investment required for any new manufacturing operation.

Even the *Financial Times* pointed out that Bookham Technology, despite making a loss this year, has a more solid under-pinning than many recent high-tech stock market launches.

Investing in electronics manufacturing is a long term venture and happily it is starting to happen once again in the UK.

Another recent example of this 'grass-roots' investment is particularly pertinent and welcome. The government is to put £40m into a new electronics manufacturing collaborative programme between industry/universities.

The scale of this investment is obvious when you

remember that the government put a mere £3m into its three year Microelectronics in Business (MiB) electronics design programme.

Such is the relative size of the government's new manufacturing initiative that some of the civil servants are worried that the achievements of the MiB programme will be dwarfed and forgotten in comparison.

This is great news for UK electronics. At last the government has realised that it must tackle the skills shortage and opportunity bottleneck at the manufacturing end of the high-technology revolution.

Another established electronics manufacturer, Marconi is investing £40m in a new graduate and engineering skills programme in partnership with Cambridge University.

Microsoft, Rolls-Royce and AT&T have all invested in Cambridge University, and last year British Aerospace, another big engineering trainer in the 1970s, launched plans for its own 'engineering university'.

For so many years the firms have been warning and about the shortage of skilled designers, but few were prepared to make the sort of financial commitment needed to address the issue.

What is significant about Marconi's collaboration and investment in Cambridge University is that it shows one major electronics employer is prepared to do something about the skills shortage.

What makes the initiative all the more impressive is that Marconi is unlikely to see any tangible return on its investment for three, maybe five years.

You can have all the dot.com flotations you like, but if Tony Blair wants to encourage new industries to create the wealth once taken for granted from industries like car-making and ship-building then he must rebuild some industrial foundations for the high-tech sector. That means being able to manufacture the electronic products which spring from the innovative ideas being produced like crazy on the regional science parks.

At different ends of the corporate scale that is exactly what companies like Bookham Technology and Marconi are doing.

It would be foolish to believe that manufacturing is fashionable again. Too much has been done in this country in the last 20 years to undermine that belief. The digital revolution continues to be driven by fabled design house like ARM and IPO obsessed dot.com individuals.

But at least Bookham's successful market flotation and this latest manufacturing initiative from the government signals that the rebuilding of the UK's electronics manufacturing industry is well on the way and is entering a new phase of optimism.

Richard Wilson

Electronics World is published monthly. By post, current issue £2.65, back issues (if available) £3.00. Orders, payments and general correspondence to **L333, Electronics World, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS**. Tlx: 892984 REED BP G. Cheques should be made payable to Reed Business Information Ltd **Newstrade**: Distributed by Marketforce (UK) Ltd, 247 Tottenham Court Road London W1P 0AU 0171 261-5108.

Subscriptions: Quadrant Subscription Services, Oakfield House Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 01444 445566. Please notify change of address. Subscription rates 1 year UK £36.00 2 years £58.00 3 years £72.00. Europe/Eu 1 year £51.00 2 years £82.00 3 years £103.00. ROW 1 year £61.00 2 years £98.00 3 years £123

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USA mailing agents: Mercury Airfreight International Ltd Inc, 10(b) Englehard Ave, Avenel NJ 07001. Periodicals Postage Paid at Rahway NJ Postmaster. Send address changes to above.

Printed by Polestar (Colchester) Ltd, Filmsetting by JJ Typographics Ltd, Unit 4 Baron Court, Chandlers Way, Southend-on-Sea, Essex SS2 5SE.

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Scottish microdisplay initiative could see the UK featuring in the bigger picture

Scottish universities are aiming to put a solid academic base under the UK microdisplay industry.

The proposed initiative is called CUPID, for Combined Universities Participating in Displays. "The group forming CUPID is essentially on the East Coast of Scotland, but we are hoping to make it a UK-wide initiative," said David Vass, Professor of applied physics at the University of Edinburgh, "We are keen to promote the UK microdisplay industry."

At the moment there are five universities expressing interest. "We are all speaking to each other, although we do not have a formal agreement yet," said Vass. The universities are: Edinburgh, Napier, Heriot-Watt, Abertay Dundee and Dundee and the aim is a long term collaboration. "The idea is to bring together expertise in microdisplays over the next 10 to 15 years," said Vass.

Among these, Abertay brings experience sometimes overlooked in

Possible partners at a glance

- University of Edinburgh – LCD microdisplays since the early eighties.
- Napier University – chemistry of LCD and light emitting polymers, and optical assessment.
- Heriot-Watt University – applications of display and the interface architecture.
- University of Abertay Dundee – human factors and performance characteristics
- University of Dundee – amorphous silicon field emitters.

technology-based research projects: ergonomics. "One of the major advantages of is the human factors element," said Vass.

Microdisplays, displays under 25mm across, offer a way of presenting highly detailed visual information to mobile users without the bulk of conventional displays. Optics focus the image in such a way that when the display is held, or worn, close to the eye, the image fills the visual field in the same way a large TV does.

Usable resolution can be far

higher than existing PDAs and other conventional mobile information devices.

Industry interest in microdisplays is such that the Society of Information displays, an international organisation, is making them the main theme of its 2000 conference.

Microdisplays take many forms. Specifically CUPID will be looking at "Active backplanes driving LEDs or light-emitting polymers, based on standard semiconductor processing, which can be obtained from various ASIC foundries," said Vass.

IBM chips go a third faster on low-k

IBM Microelectronics is to boost its chip speeds by a further 30 per cent when it starts using a low-k dielectric next year.

The material is used as an insulator between layers of metal interconnect in a chip. A lower value for k means reduced capacitance between wires, leading to increased speed and reduced crosstalk.

"It's a very large improvement compared to what most people use which is FSG," said Michel Rivier,

a technical specialist at IBM.

Licensed from Dow Chemical, the low dielectric constant material has a k of 3, significantly better than that of FSG, the most common material used today, with a k of 4.

The first process to use the material is called Cu-11, which uses copper for interconnect (see page 46). Using 0.13µm lithography, the process will result in transistor channel lengths down to 0.08µm.

"Right now it's in pre-production," said Rivier.

Maps on your mobile?

Yeoman group, the mobile navigation firm, has entered an agreement with Ordnance Survey to co-develop standards for mobile navigation systems.

Yeoman said it is developing an operating architecture for mobile navigation systems, which should allow consumers access to map databases via WAP phones.

"This is a significant step towards reaching a complete mobile naviga-

tion offering," said Vincent Geake, Yeoman's chief technical officer.

Yeoman also said it has made an unsolicited offer for UK firm Laser-Scan of one new Yeoman share for every Laser-Scan share with a cash alternative of 42.9p per share. Laser-Scan has developed software to support a mapping database.

"The Yeoman board believes that the enlarged group will significantly enhance value for shareholders of



Welcome medicine for sick mines... The Japan Alliance of Humanitarian Demining Support (JAHDS) has presented the HALO Trust with equipment specifically designed to support mine removal efforts at Angkor Wat in Cambodia. Mine Eye was a joint development by companies including sensor-designer Omron. JAHDS was started by Hiroshi Tomita after he discovered that 'butterfly' mines in Angola were deliberately shaped to entice children to play with them.

both companies because their technologies in the area of mobile navigation and geographical information are complementary," said Yeoman in a statement.

Scotland urges mobile-mast caution

The Transport and Environment Committee of the Scottish Parliament has recommended that a precautionary approach regarding health issues should be adopted when siting mobile phone masts.

If its recommendations are accepted then full planning control would be introduced for masts.

In its 'Inquiry into Telecommunications Developments' report admits there is no conclusive

scientific evidence of a health risk, but it believes the level of public concern justifies a precautionary approach.

This would mean schools, hospitals and residential areas would be considered unsuitable sites.

It also wants the environmental impact to be more carefully considered, with initiatives such as mast and site sharing used. The possibility of requiring a national roaming agreement to be made is also

an option.

Similar recommendations could soon appear before the UK government. A report from a similar inquiry is expected to be presented any time now by the Independent Expert Group on Mobile Phones.

The group was set up to look at concerns about the health effects of mobile phones, assess existing research and give advice based on that knowledge.

Amstrad rings in the e-mail phone

Amstrad has finally unveiled the mass-market Internet product which the company has been working on for over a year.

The 'e-m@iler' is intended to bring e-mail to the mass market without the need for a PC. The unit was developed in collaboration with BT and consists of a phone unit with keypad and LCD screen.

The unit can send and receive e-mail, holds up to 700 contact details and has automatic e-mail notification and collection. It also provides a digital answering machine and fax facilities.

"I see the e-m@iler becoming the 'all-in-one communications centre' in the home. It is the blockbuster product Amstrad has been working on for the last eighteen months and which the market has been waiting for," said Sir Alan Sugar, Amstrad's chairman.

The e-m@iler is being sold at a subsidised price of £79.99 and is apparently available in the High Street already.



Solutions waiting for a problem

European electronics companies are flocking to join intellectual property Web site yet2.com.

Started in February by 30 US firms, yet2 is a shop window for intellectual property that companies have invented, but have no use for.

The most recent additions from Europe are BT, Bosch, Philips Electronics and Siemens, together with Japan's Toshiba.

"Large corporations are sitting on huge and growing reserves of great ideas that never see the light of day or are used once only and never again," said Chris de Bleser, CEO of Yet2.com.

US founding sponsors of yet2 include 3M, Dow, DuPont, Honeywell, Polaroid, Rockwell and TRW.

Doubts cast over benefits of copper IC processing

Using an all-layer copper process to make a chip showed 'no difference' in performance compared to the same chip made in aluminium, says TSMC's top scientist.

The finding could affect the widespread use of copper processing for the upcoming

0.15µm generation of process.

According to the main proponents of copper processing, IBM and Motorola, copper gives an added performance advantage of 20 per cent over aluminium because of copper's lower resistivity.

Asked if TSMC had made demonstrator chips to compare the performance of copper with aluminium, Dr Shang-Yi Chiang, vice-president for R&D at TSMC, replied: "A customer who ordered copper saw no difference in performance."

Dr Chiang emphasised that the process involved, which was TSMC's 0.15µm all-layer copper process, did not use low-k dielectrics. TSMC's low-k dielectrics process is currently in product qualification, which is due to be completed in June.

"Also, the problem was that it was the same design - the same

layout," said Chiang, "you have to optimise the design to take advantage of the copper process."

Accordingly, Chiang believes: "We do not expect a very large demand for copper until the 0.13µm process when customers have learnt how to optimise their designs to use copper."

Copper processing has been promoted by IBM as the answer to limitations in traditional aluminium processing. Motorola is in the forefront of copper processing and licensed its process to Chartered Semiconductor of Singapore.

AMD says it will use copper processing on some layers of its Athlon microprocessor. Although Intel has dubbed its latest generation of microprocessors 'Coppermine' it does not use copper to make them and has said it does not think it will use copper until the 0.13µm generation.

Engineers' pay settlements hit by strong pound

Engineering pay settlements have remained at a historically low level of 2.4 per cent for three months in a row.

The latest survey findings from the Engineering Employers' Federation (EEF) shows that nearly one in eight settlements were pay freezes in the three months to the end of February 2000.

The EEF said this situation is due to the continuing high level of sterling.

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

Dot.commers have become such a constant source of fun and amusement that we will never be able to take them seriously again. David Manners stifles his giggles.

Smirk at those who ramble on about dot.com fortunes – often the same ones who talk about house prices. Giggle at the discomfort of life-long money-grubbers seeing youngsters making instant fortunes. Ridicule the moans of those who didn't get shares, or enough shares, in the latest IPO. These Sad Acts give us a good laugh.

Not since Screaming Lord Sutch have we had so much fun as we're having with the dot.com companies.

His late Lordship took the mickey out of the political world; the dot.coms are taking the mickey out of the financial world.

Seeing solemn money-men trying to add their patina of logic and justification to the dottiness of the dot.com world is hilarious. In future they are going to find it hard to convince us of the authority of any of their analyses.

The only rational explanation for the dot.com phenomenon is that they resurrected the greed/fear frenzies of the past: 'Tulipmania' in the 1630s, when Dutchmen paid the price of a house for a tulip bulb; the South Sea Bubble of 1720; the 1849 Gold Rush; the 19th century boom/bust in railway shares; the 1970s Australian Mining Boom led by Poseidon shares; the 1988/9 UK Housing Boom; Japan's 1980s 'Bubble Economy' when Tokyo land prices were so high that the Imperial Palace's gardens were worth more than the State of California.

In all these greed/fear frenzies, the fear of being left out made people take leave of their senses and buy pigs-in-pokes at crazily-escalating prices, and greed made people over-borrow to buy assets they could not afford, gambling on massive profits from the expected price rises, and ruining themselves and their families in the ensuing crash.

The same is happening with the Internet and the mania to invest in dot.com companies. But, unlike some of the popular frenzies of the past, the dot.com scenario has a calculated, professional element as traditional financial interests seek to get their share.

We are encouraged to think that dot.com companies are started by sparky young people with nothing except a 'good idea'. How far from the truth that is.

One of the backers of the recently floated lastminute.com was Intel, whose PR initiatives

helped create lastminute's high public profile in the months preceding the launch.

The founders are not the innocent young techies operating from a garage of an earlier generation - lastminute's founders are highly articulate, well-funded Oxford graduates with well-honed skills in making sophisticated financial presentations.

Venture capital - once jealously hoarded by high-tech start-up companies for innovative product development - tends to be spent by dot.coms mostly on publicity rather than on developing a service or a product.

Some of the dot.com companies are spending on publicity at the rate of £1m a month - the money coming from venture capitalists wanting to make a quick killing through an early public offering on the stock market.

The message of the venture capital-backed dot.com founders is usually simple - grab the IPO money and run.

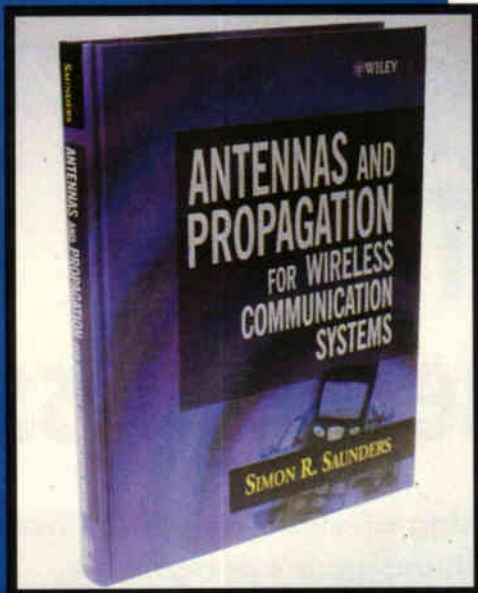
For all those who are not overwhelmed by the fear/greed frenzy of the dot.com phenomenon, it can be a fun thing to watch.

Smirk at those who ramble on about dot.com fortunes - often the same ones who talk about house prices. Giggle at the discomfort of life-long money-grubbers seeing youngsters making instant fortunes. Ridicule the moans of those who didn't get shares, or enough shares, in the latest IPO. These Sad Acts give us a laugh.

There are, of course, many fine and worthy Web sites, some delivering wonderful things, but the venture capital-backed dot.com is often a greedy, flaky beast to be ridiculed and exploited.

What they are good for is: 1) To make a quick killing; 2) To create an inflated share valuation which can then be used to take over proper companies with revenues, employees, assets and profits; 3) To exploit their capital-raising abilities to provide useful products and services. ■

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"THD is meaningless"

Looking at audio amplifiers from an RF designer's perspective, Anthony New argues that THD figures are, "irrelevant, irrational, and completely spurious." He believes that intermodulation distortion figures are far more relevant, and, unlike THD figures, reflect how an amplifier 'sounds'. Anthony also explains what 'IMD' is and how to measure it.

I think I can explain why the standard definition of 'THD' is completely meaningless as an indication of what it purports to measure and how it is utterly irrelevant to the uses to which it is generally put; I have however no idea why the many engineers with far greater experience of amplifier design than myself should continue to use the term at all, let alone attach so much importance to it. Yet they do.

So what are my objections to it? The problems fall into several categories.

'Total' harmonic distortion

Firstly, the concept of 'total' harmonic distortion is spurious because it sums a great many separate components which are not equal in kind or effect. Anyone who has experimented with waveform generation will appreciate that for example, 1% second or third harmonic distortion on a reasonably pure tone has a quite different sound from 1% seventh or ninth harmonic, and is much less audible, Fig. 1. In fact 1% of second or third harmonic distortion is not only not unpleasant but is sometimes positively preferred by those who like 'valve sound', whereas early transistor amplifiers producing a great deal less than 1% of higher-order harmonics sound pretty awful on any challenging music.

Secondly, of course, many of these harmonics will be outside the range of human hearing anyway. It is common practice to include distortion figures at frequencies as high as 5 or 10kHz, but of what possible significance are they? As a young man I could (just) hear loud tones as high as 20kHz, and found the common TV line oscillator whistle at 15.625kHz acutely painful. But I very much doubt whether anyone can hear the third or fifth harmonic of a 10kHz tone

A recent article by Ian Hickman showed how to measure total harmonic distortion, or THD, down to levels below 0.001%. This achievement is worthy of applaud for its technical challenges, yet I cannot help marvelling at the enormous waste of effort that has been made over the years on such an irrelevant, irrational, and completely spurious figure as THD.

Irrelevant? Irrational? How so? And how can a figure used so frequently in audio design be spurious? The latter is a very good question, and one which I have not been able to answer.

— even a loud one — and certainly not one of amplitude below 1% of its fundamental.

I contend therefore that the practice of adding all distorting harmonics together to give a sum total, without any weighting factors, is quite arbitrary and not indicative of the audibility of any harmonic distortion produced by an amplifier.

Since the audibility of a given THD figure depends heavily on its actual makeup, the figure is also pretty useless even as a purely theoretical comparison of two or more amplifiers, since no acoustic model of audibility is included.

THD and the ear

However there are yet worse flaws in the THD concept which make the above problems almost academic. These concern the very nature of harmonic distortion itself.

In my view one of the central problems with traditional audio amplifier design is the insistence on considering the device as a piece of electronic equipment devoid of any psycho-acoustic considerations. The extreme of this was the concept of 'a straight piece of wire with gain' which is fortunately unattainable, as its gain and bandwidth would make it seriously less than optimal and possibly quite unusable in a real system.

This is not to say I fall into the 'subjectivist' camp in audio criticism — far from it. I have listened attentively to the debates those of this persuasion have had with such luminaries as Douglas Self and have been mightily impressed with Self's clear — and seminal — analysis of amplifier distortions.

The problem I have with these debates is that neither side seems particularly interested in what the other is saying. On the one hand we are told 'all the distortions have been correctly analysed', on the other 'a difference can be heard'.

It seems to me that if these opposing views are to be reconciled the answer must lie at least partly in psycho-acoustics, that is — as far as I am concerned here — the study of how we perceive sounds.

I don't claim to any professional qualifications in this field, but one thing stands out about the current discussion of distortion in audio systems, namely the lack of any auditory model. It is as if in designing seats and seat belts for cars, nobody was prepared to test a human body — or even a dummy model of one.

I can certainly understand how an engineer is tempted to subtract the input signal to an amplifier from a linear proportion of its output and declare — by definition — any difference to be distortion. The problem I have with this view is that, traditional THD testing methods only look at one small part of this difference, and so far as I can see, *harmonic distortion* isn't perceived by the ear as *distortion* at all.

What is the effect to a listener of adding a few percent harmonic distortion to the waveform of a musical instrument or group of instruments? It is to brighten the *timbre* of the instrument.

Since most of the 'distortion' products will already be present in the undistorted signal, a similar effect may be obtained by adjusting the tone controls. Those of you who have spent much time siting microphones in the recording

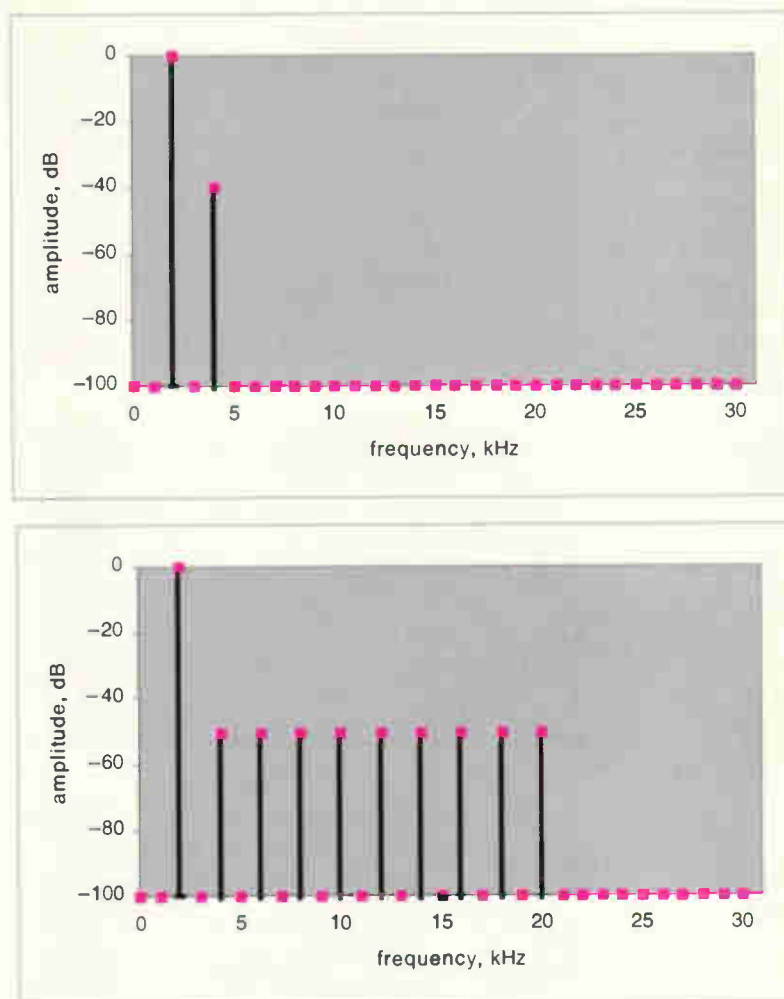


Fig. 1. Subjective audibility of THD: which sounds worse — 1% of purely 2nd harmonic distortion as in (a) or 0.5% of mixed harmonics as in (b)? Probably the latter though its thd specification is better. All the spectra that follow have a logarithmic Y-axis (amplitude) and a linear X-axis (frequency) even where this isn't shown.

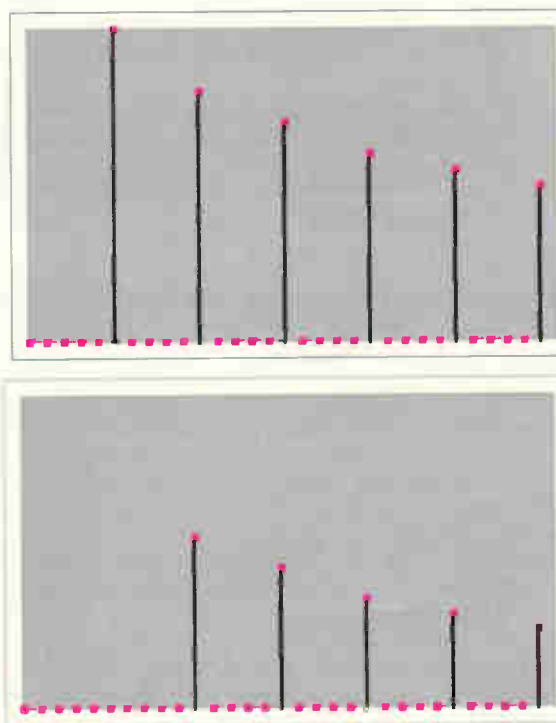
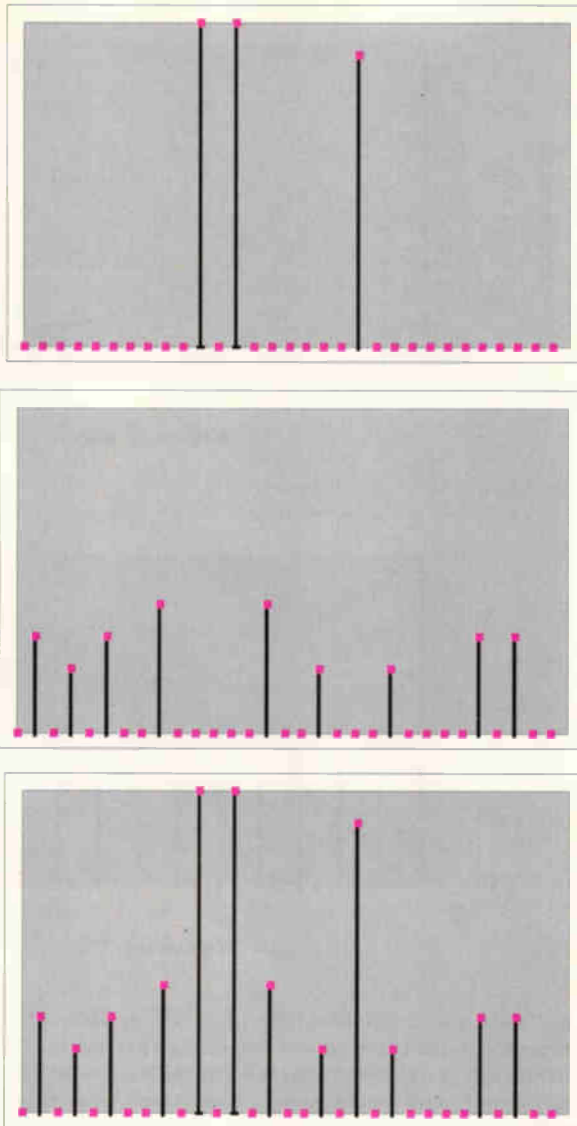


Fig. 2. Auditory masking of harmonic distortion. (a) Typical spectrum of real signal with many harmonics; (b) Nominal distortion products at 0.01% each; (c) Error products due to frequency-response non-linearity on original signal; note that these are on the same frequency as the distortion signals and at much higher level, masking the actual distortion.

Fig. 3.
Intermodulation distortion showing the lack of auditory masking.
 (a) Pure multi-tone signal at input to amplifier;
 (b) 3rd order IMD products produced by amplifier;
 (c) Amplifier output signal: IMD products are at different frequencies from either input tones and therefore not masked.



industry will be aware how critical their exact placement is to recording balance – and I don't simply mean relative loudness.

You will also be aware of the dramatic change in both *subjective* sound and *objective* frequency response obtained by alterations in these positions. Even small movements can have effects far more noticeable than minute levels of THD in the recording or playback medium.

Golden ears

It seems quite possible to me that when those with 'Golden Ears' say they can hear a difference with such-and-such change in the equipment they may be right.

When I was younger and my ears were sharper I listened to many excellent loudspeakers. Very few sounded as good as a live performance. I only heard one – namely the Quad electrostatic – that could actually fool me into thinking the performer was present in the room. The illusion was so strong that I was convinced the performer was hiding behind a curtain until I looked.

Even now in any hi-fi demonstration, the difference in sound between different loudspeakers in the same room – even those produced by the same company – is so marked as to make a nonsense of the claim that many of them can really be 'low distortion' in the 'blameless' sense that Self used for amplifiers. It also makes a nonsense of the idea that state-

of-the-art amplifier distortion could be significant compared with it.

The point is, that 'being able to sense a difference' is *not* equivalent to 'sensing distortion' in any meaningful sense. Nor is it an indication even that one of the items being compared is *necessarily* better or worse than another.

Any real musical instrument – including electronic ones such as keyboards – produce sounds which, when converted into analogue electrical signals, contain possibly many discrete tones. Usually they also contain many harmonics of the tones, the relative amplitudes of which strongly influence the 'sound'. The relative levels of these harmonics – both perceived and measured – vary with many factors including auditorium response and the distance between source and listener. Further factors occur due to the room where the sounds are replayed. Even the shape of the ear itself has an enormous effect, and the presence of hair or hat!

Consequently, even for a particular note played there is no absolutely 'right' or 'wrong' quantity of any of these harmonics. A *slight* alteration of the levels of these does not correspond to an unpleasant 'distortion' of the sound but to a slight change in perceived distance, position, or playing by the instrumentalist.

Furthermore, such slight changes in these levels may be correctable – or at least adjustable in part – by variation of the user's tone controls. In addition, the recording engineer may already have done this to a considerably greater extent prior to or after mixing the output of several microphones.

Frequency response

I also contend that most of the apparent subjective differences that still exist between different audio amplifiers are *not* due to distortion at all but to slight differences in frequency response. This point should receive far more attention during design than it generally does.

Any deviations from a flat response are likely to have a greater impact on the level of high-frequency harmonics present in the amplifier output than the tiny harmonic *distortion* products. If noticeable and uncorrectable with tone controls, these can also contribute to listener fatigue.

The Human ear/brain combination is also very good at correlating impressions over time, so even slight bumps in the frequency response can become noticeable and even irritating eventually.

Since these real-world variations in the levels of a signal's harmonics dwarf any likely distortion products in a correctly operating amplifier of moderately good quality, it seems perverse in the extreme to use any measure of these tiny 'distortions' as a useful figure of merit.

I also note that conventional methods of measuring amplifier performance don't really satisfy the traditional definition of distortion – output relative to input. Distortion tests use only a single frequency source and cannot monitor either non-harmonic distortions or frequency-response errors. Also, the frequency response tests are done differently and are far less sensitive.

For example, when did you last see an amplifier's frequency response flatness specified to 0.01%? Plus or minus 1dB is more usual, which is 12%, and even 0.5dB is still 6%. Of what possible significance is the 0.001% harmonic distortion of an amplifier when its frequency response contributes an error in harmonic content of several percent?

Since the harmonic distortion products will also lie on existing signal frequencies they will be effectively masked from audibility, Fig. 2.

Intermodulation distortion

Does this mean that I join the subjectivists in eschewing measurement completely? Not at all. It just means I favour using a *sensible* measure of distortion instead of a *senseless* one.

Fortunately one is conveniently to hand.

Outside of the parochial and fashion-conscious world of audio, most amplifier designers have long since given up measuring or even talking about harmonic distortion and use instead *intermodulation distortion*, or IMD for short.

Measuring IMD has three particular virtues over THD. One is that, unlike THD, IMD is always a measure of distortion in-band. No weighting is needed for audibility at different frequencies. The second is that it really *does* degrade performance of a system. It does so regardless of whether it is measured *objectively* by such quantities as BER (bit-error rate), SVE (signal-vector error) or spectral spread or regrowth, or *subjectively* by intelligibility of communication.

A third advantage is that unlike the case of harmonic distortion, intermodulation distortion is quite easily measured by standard laboratory equipment. Figs 4, 5.

At a stroke the problem introduced earlier of distortion of 10kHz tones is solved. If two tones at, say, 9kHz and 10kHz are supplied to a good but not perfect amplifier, it is not the *harmonic* distortion that is audible but the *intermodulation* distortion.

The non-linearity in the amplifier produces new tones, not present in the original, such as, in this case, 1kHz, 8kHz, and 11kHz. Fig. 6. Although the audibility of IMD depends on the type of music, in general it is much more audible than any harmonic effects precisely because the distortion produced is not harmonically related to the signals of interest.

Intermodulation distortion typically makes music sound muzzy and indistinct. The worse case of this is usually heard on old car loudspeakers where the cone is broken or the voice coil rubs on the pole pieces, but it can be heard in very much more expensive and well cared-for equipment.

This is the reason why a blameless amplifier must be linear – the harmonic distortion measured is a complete red herring.

What is IMD?

Although IMD has not been as much discussed in the design of audio amplifiers compared with THD, there is a considerable literature on IMD in general and its application to RF amplifiers. For this reason, I will give a simple overview of IMD and point out a few implications for its use in audio design.

In general, IMD is produced whenever two or more signals with distinct frequencies F_1 and F_2 pass through a device – be it an amplifier, filter, or other circuit – that possesses an amplitude non-linearity of the form,

$$Y = A_1 X + A_2 X^2 + A_3 X^3 + A_4 X^4 + A_5 X^5 + \dots$$

where A_1 is the nominal gain of the amplifier and the higher powers of X correspond to the various non-linearities that may be present. I have ignored phase non-linearities here for simplicity.

This non-linearity produces IMD products at the following frequencies.

$$nF_1 + mF_2$$

where n and m are non-zero integers. Note that if you put $n, m = 0$, you get purely harmonic distortion rather than IMD, which indicates that harmonic distortion is a special case of a more general phenomenon.

The order of the IMD products is defined as,

$$k = |n| + |m|$$

so that the 'third-order' products that often dominate are of the form,

$$F_1 \pm 2F_2, \quad F_2 \pm 2F_1 \quad \text{and} \quad 3F_1, \quad 3F_2$$

In RF amplifiers a further restriction often applies. Even in many 'wideband' products the overall bandwidth of the amplifier is less than an octave, and so only those odd-order

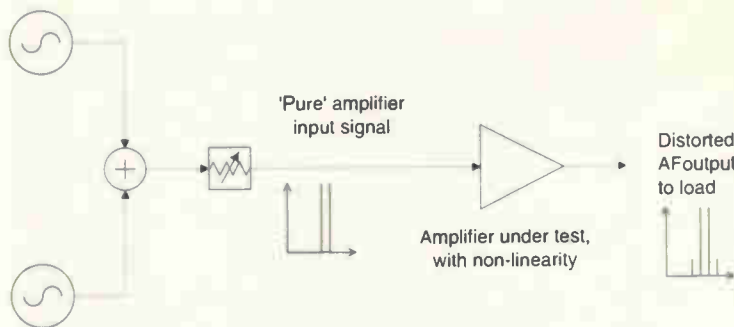


Fig. 4. Producing multi-tone test signals with standard sinewave signal generators. Harmonic output of the generators is not critical. The passive combiner and attenuator should not affect the measurement linearity – their IP3 can be measured in principle by increasing the generator output level beyond what the amplifier requires, allowing the effective generator IMD to be calculated at the lower levels for the amplifier test.

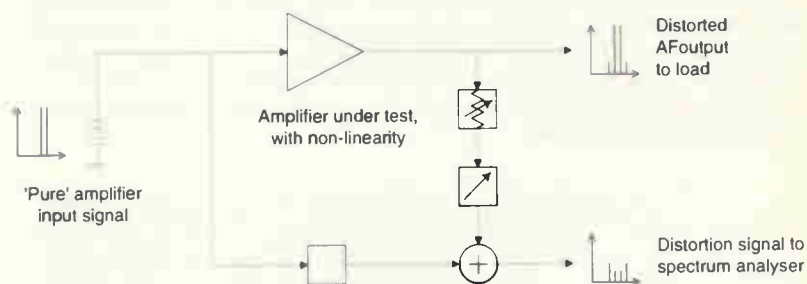


Fig. 5. Possible distortion measurement circuit. The attenuator, phase shifter, and time delay are first adjusted on a network analyser to cancel the input signal as well as possible across the whole audio range. This reduces the dynamic range of the distortion signal for spectrum analysis. The residual input tones also reveal the gain flatness of the amplifier over frequency, which contributes to the amplifier's output errors. The input tones may be swept across the frequency range with a constant difference frequency.

products with,

$$|n| - |m| = \pm 1$$

are 'in-band' and of concern: consequently even-order non-linearities – second, fourth, etc. – which produce no odd-order products are usually ignored.

However in a multi-octave device such as an audio amplifier, this restriction will not apply. The most common and usually most important non-linearity is however still a third-order non-linearity of the form:

$$Y = A_0 X + A_3 X^3$$

which will result in IMD products of third order only, namely $\{2F_1 - F_2, 2F_2 - F_1, 2F_1 + F_2, \text{ and } 2F_2 + F_1\}$. The first two represent the classical IMD products and the other two are higher-frequency IMD products, at roughly three times the fundamental frequencies when these are close together.

The spectrum of Fig. 6 shows these third-order products of a two-tone signal, in addition to higher-level, second-order products: note that only the third-order products are close in frequency to the input tones.

Figure 7 shows an idealised spectrum of four-tone test sometimes used with RF amplifiers. If the power input to the device is varied, the output levels of the IMD products will vary too. This is shown in Fig. 8, from which you can see that if the input level increases by 10dB, the third-order IMD products increase in absolute level by 30dB. Their level relative to the wanted output signals also increases by 20dB.

If the straight lines are extended to the right you will see that they all meet at a single point. For obvious reasons, this

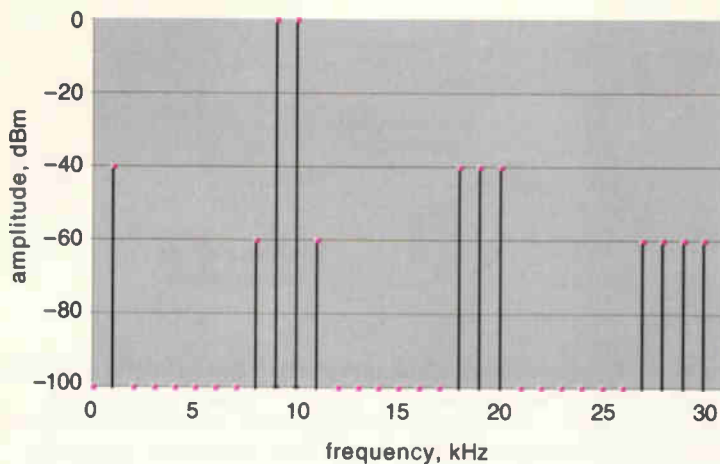


Fig. 6. Frequency spectrum of two-tone signal showing expected 2nd- and 3rd-order products due to intermodulation distortion. The lowest frequency component is the 'beat' frequency between the tones, the two small components next to the two main tones are the 'in-band' 3rd-order components, and the rest are a mixture of harmonic and higher-frequency non-harmonic products. In an audio amplifier all these – and more – may be audible for some pairs of tones, though they might be out-of-band in a typical RF amplifier.

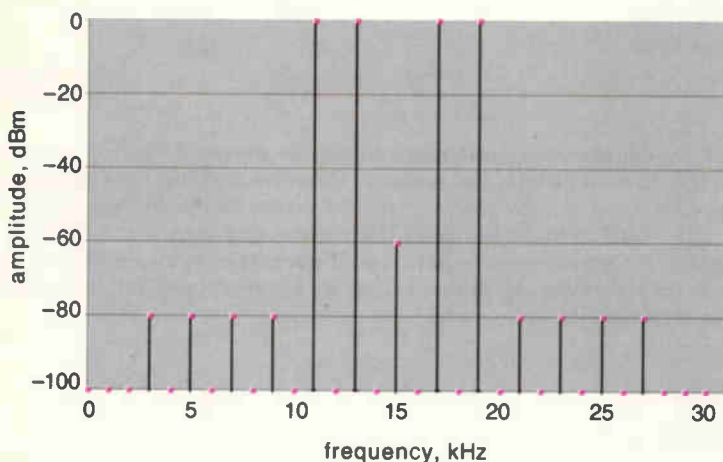


Fig. 7. Idealised spectrum of four-tone test sometimes used with RF amplifiers: the four main tones are harmonically related, phase-locked and phase-peaked to maximise the peak value of the signal envelope in the time domain. This – in an RF amplifier at least – is likely to maximise the visible 3rd-order IMD products, particularly the central one between the two pairs of tones which thus makes an easy frequency component to check. With zero IMD this component would be completely absent.

Fig. 8. Amplitude response of amplifier displaying 'classical' 3rd-order IMD. Amplifier input signal level is displayed on the X-axis, and output level on the Y-axis; both axes are logarithmic. The straight line through the origin represents ideal linear response. The straight line at a steeper angle shows the theoretical level of IMD products, which change three times as quickly with input amplitude as the signal itself. The point where the straight lines meet is the '3rd-order IMD output intercept point' or IP3. The curved lines show the likely real characteristics as the amplifier begins to clip, however for sensible operating points well below the IP3 the straight lines are a fairly good match for a single-stage class-A RF amplifier without any special linearisation techniques. The IP3 concept is also useful for other devices such as mixers which also display IMD. For any input signal level on the X-axis, the upper line will show the nominal output level and the vertical separation between the two straight lines will show the expected linearity in dBc. When high-level multi-tone signals are concerned this figure – rather than the noise figure – usually represents the dynamic range of the signal, since it indicates the relative level of interfering products.

is known as the 'output intercept point'. Strictly in this instance, it is the 'two-tone, third-order output IMD intercept point' or IP3.

For any signal below this point, the level of IMD products can be estimated by subtracting the output signal level from the IP3 to give a figure in decibels, and doubling this to give a figure in dBc. This represents the IMD relative to the 'carrier' i.e. wanted signals, assuming them to be similar in level.

Real signals

It is highly unlikely that a real device could be operated anywhere near its IP3 point. This point is useful for calculation and reference only.

Also, a real device is likely to show IMD at other orders, particularly fifth, when it is driven at all hard. As these will reduce by the fifth power of the signal level instead of the third though, they are likely to be lower in level.

However a fifth-order non-linearity will also produce some third-order IMD. This may even cancel out some or all of the third-order IMD produced by the third-order non-linearity, resulting in the fifth-order IMD product dominating at some output power.

A similar situation exists for higher-order IMD products, but the actual levels are generally both lower than third and fifth-order IMD products and rather less predictable.

When more than two large signals are sent through the same amplifier at the same time, the number of IMD products grows rapidly. Figure 9 shows spectra of a real, albeit RF, amplifier with real multi-tone signals.

Two tones produce two close-in IMD products, in addition to the other distant ones shown in Fig. 6, but nine products are visible with three tones, Fig. 9c).

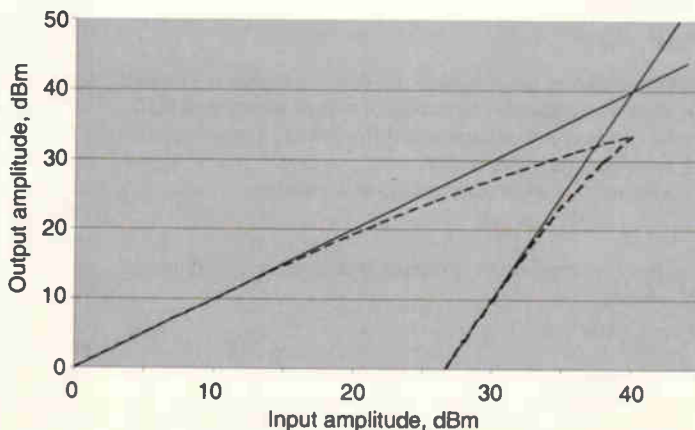
With four tones the number increases again, Fig. 9d), though in practice some of these may be co-incident. When a complex modulated signal is used rather than a set of CW tones, the IMD products occupy a bandwidth rather like a noise spectrum.

Since these IMD products are not harmonically related to their causative signals, they behave like noise, too, reducing the intelligibility of speech or data transmissions to a measurable degree. It is not possible to filter them out, since although the bandwidth they occupy increases with their order of distortion, Fig. 10, their bandwidth always includes the original signal.

So how can IMD be measured?

As I commented, one of the benefits of IMD over THD is the relative ease of measurement due to the distorted products being not harmonically related to the original signals.

A typical setup will consist of a pair of signal generators – or, often, a dual-output generator – a linear combiner, possibly resistive, and a spectrum analyser, Fig. 4. The analyser display will then look something like Fig. 6 if the analyser



span is sufficiently wide, or like Fig. 9b) in the more usual narrow-band case.

The IMD products may be much lower in level but are easily seen provided the analyser has enough dynamic range. If the analyser has appropriate delta markers, the relative distortion can simply be read off the screen display. If not a little mental arithmetic is required.

Note that with IMD tests it is not necessary to use especially low-distortion oscillators since the harmonics produced will not normally interfere with the measurement process. However the linearity of commercial spectrum analysers is rarely much better than 80dBc – or 0.01% in voltage terms – and may be poorer. For the best amplifiers some additional filtering may be needed to notch out the pure signals, or a coherent subtraction method used as shown in Fig. 5.

Another test commonly used is the four-tone test illustrated in Fig. 7. Here, four tones at, say, 3, 4, 6, and 7kHz are produced by four phase-locked generators and the analyser tuned to look for the missing 5kHz component, which can only arise from a non-linearity.

What figure-of-merit is needed?

A figure-of-merit commonly used in RF amplifier design is the intercept point, in dBm, Fig. 8. The higher this is for a given power level required from the amplifier, the lower will

Decibels, dBm, dBW and dBc

It is common to specify amplifier distortion in terms of percentage, with the understanding that voltage ratios are intended. However where loudspeakers have to be driven it is power that is more relevant.

For constant-impedance systems with a wide signal range a convenient logarithmic measure is the decibel or dB. This is strictly a ratio of two quantities with the convenient feature that 10dB corresponds to an increase in signal power by a factor of ten, and 20dB corresponds to an increase of ten in voltage and ten in current, making one hundred in power. Specifying an increase of 20dB is then unambiguous, regardless of whether the speaker is thinking in terms of power or voltage.

Where an absolute level is needed, the terms dBm, i.e. dB relative to one milliwatt, and dBW, i.e. dB relative to one watt, are commonly used. In specifying levels of distortion a further measure is useful, namely dBc. This refers not to a noise-suppression scheme but to decibels relative to the carrier – i.e. the main signal.

Where multi-tone signals are present there is however a further possible confusion between dBm/tone, dBm mean, and dBm peak.

Fig. 9. Spectrograms of real signals. For convenience these have been taken at RF, though similar spectra could be observed at audio. (a) 2-tone signal, no visible distortion; (b) 2-tone signal plus obvious 3rd-order IMD; (c) 3-tone signal; and (d) 4-tone signal. Note that each extra main tone produces many extra IMD products. In (d) the tone frequencies have been deliberately chosen to make as many products visible as possible; usually several would either overlap or appear to do so within the resolution of the spectrum analyser. However in a complex musical signal the large number of signal tones would cause the hundreds of IMD products to merge into a noise-like background which reduces the clarity of the signal.

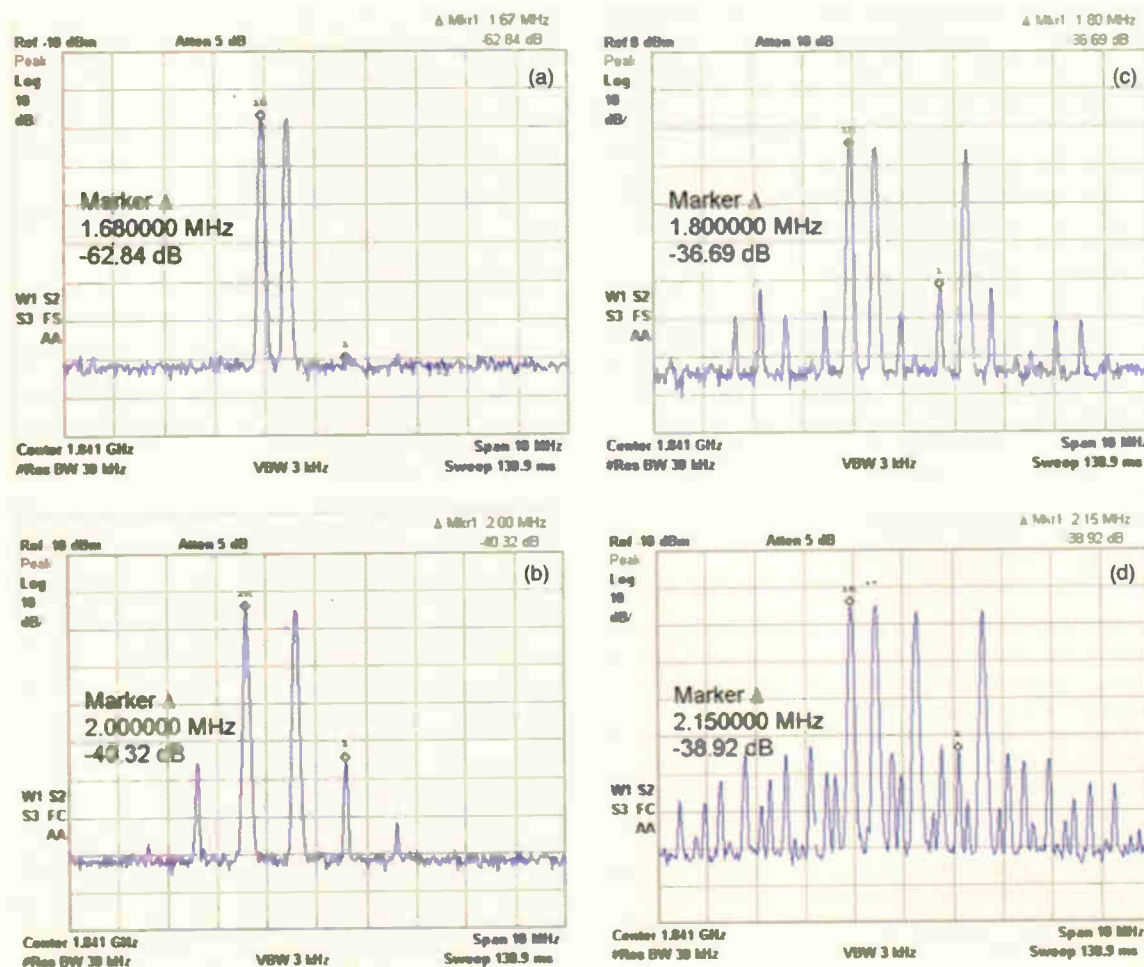
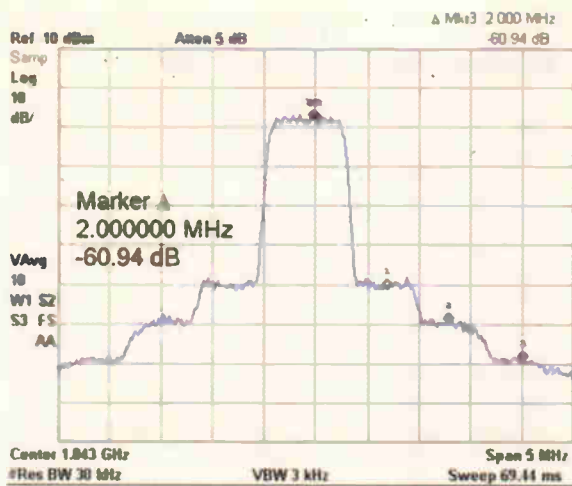


Fig. 10. Example of a real modulated signal with IMD – here the IMD products are not discernable individually but serve effectively to raise the noise level in steps – each step corresponds to a particular order of IMD. The first step on each side is produced by 3rd-order IMD, the next by 5th, the next by 7th, and so on. Note that the IMD products occupy more bandwidth than the original signals; the higher the order, the more bandwidth occupied. The step pattern might however not be visible with the less-ordered spectrum of a typical music signal.



be the distortion produced, and from this IP3 figure it is quite easy to calculate how much distortion is likely at any given power level.

However at this point I should comment that one of the many differences between RF and audio amplifiers is that RF amplifiers are usually operated somewhere near their continuous peak power rating. Alternatively they are at least backed off from this by a consistent amount. Also, they do not often use feedback to achieve good linearity. Consequently, they may have an IMD response which approximates to a classical curve over most of their useful power range and for which a single IP3 specification is a useful measure of linearity in any application.;

Where real audio amplifiers are concerned I feel that typical responses are unlikely to be so simple over the wider range of signal levels encountered, particularly in a 'blameless amplifier' where all of the distortion mechanisms have been separately identified and reduced to a low level by various means.

Furthermore audio music signals can have a very high peak-to-mean ratio. It is common practice to specify amplifiers with a power handling greatly in excess of what is normally required. As a result, much of the time they will be operating at a very small fraction of their nominal power output, where the real distortion produced is somewhat different from the 'classical' third-order model.

It is likely, therefore, that rather than a single calculated IP3 figure, a curve of measured IMD levels versus signal level is more appropriate, rather like the waterfall spectrographs sometimes used. It would also give a far better indication of the order of distortion produced than a single figure, even an IP3 figure.

Nevertheless the level of intermodulation produced by an amplifier is, as I have shown, absolutely critical to its quality as an audio device. Any useful specification for its linearity should reference this. I therefore propose that the specification should run something along the lines of:

'Two-tone third-order IMD performance:
better than -70dBc over 0.1W to 30W and 50Hz to 20kHz'

or something similar. In practice, it may be necessary to limit the tests to a set of standard test tones, for example 3.5kHz and 4.5kHz. Intermodulation products would be looked for at 1kHz, 2.5, 5.5, 11.5, and 12.5kHz.

Harmonic products at 7kHz and 9kHz might also be present but could be due to the signals sources themselves. It would also be possible to repeat the test at additional low and high frequencies to test IMD performance there, such as 350/450Hz, and 15/16kHz.

Of course modern lab equipment is capable of performing swept measurements and down-loading the results to a PC for analysis and printing, so a swept measurement may be acceptable as a standard.

The author
Anthony is an electronics engineer at Wireless Systems International, currently working on high-linearity RF amplifiers for mobile base-stations.

Second-order non-linearity

The third-order function discussed earlier was selected to represent a typical amplifier non-linearity. What would happen with an amplifier having a second-order non-linearity?

This is a rather interesting case study, as it helps to explain the difference between 'valve sound' and 'transistor sound' which used to convey such emotion many years ago and in some circles still does.

A second-order non-linearity such as that often found in a thermionic valve produces second-harmonic distortion – which is not unpleasant in moderation. And it only produces even-order IMD, namely zeroth and second-order, at low level. It produces *no* odd-order IMD products of the form $(F_1 \pm 2 \times F_2)$.

A narrow-band amplifier produces no in-band IMD at all. Thus the absence of any second-harmonic cancellation in a class-A configuration has no impact on the IMD present, as suspected by those who prefer this configuration.

For much of the music program, the loudest frequencies present in the signal will often be harmonically related. Many of these extra distortion terms, of the form $F_1 \pm F_2$, will fall on, or close to, existing signal frequencies at much higher level and may be reasonably effectively masked.

Provided the levels of distortion are not excessive the result will probably not be particularly unpleasant, and may give the effect of a warm colouration to which one can become accustomed. Note that this form of distortion reduces markedly as output level drops, so that soft passages may be portrayed quite realistically; loud passages are likely in any case to have a richer harmonic texture which hides the IMD more effectively.

In contrast the chief distortion mechanism of early transistor amplifiers was *not* large-signal output device non-linearity but crossover distortion. This generally becomes increasingly noticeable at low volume settings.

Large amounts of feedback were often added to cure this and other problems, though the designers perhaps did not always appreciate how much the loop gain dropped in the crossover region. Consequently transistor amplifiers tended to suffer from less high-amplitude *low-order* non-linearity and more low-amplitude *high-order* distortion.

The effect of this on the reproduced sound was quite distinctive. Gone was the warm coloured but fairly clean sound familiar to many who hadn't perhaps experienced the best valve amplifiers. Even by the 1960s, these could boast less than 0.1% THD, most of that being the relatively benign low orders. In its place was a cold, muzzy (and sometimes hissy, but that's another issue) sound which could be particularly noticeable in solo piano works. I think the worst commercial design I ever heard was the 'Sinclair 2000', which was pretty, but built down to a price.

It has been said that the unthinking application of negative feedback around an amplifier can often merely transform large amounts of low-order distortion into small amounts of high-order distortion. It is also true that any crossover-induced IMD remaining after the application of nfb is still present at low signal levels rather than diminishing with volume.

It would have been nice had designers appreciated the futility of their policy of measuring distortion in terms of THD at full output. Instead they concentrated in reducing it, albeit with some success. To those concerned with measurable THD, the trade-off seems worthwhile, but the high-order IMD products were usually spread far away in frequency from any masking tones in the signal, and were thus very audible.

On complex music containing many strong frequency components, the large number of high-order products degenerate into a background noise. This noise is signal-dependent and hides any subtle details from the ear, Fig. 10. It is no accident that the 'clarity', often regarded as the highest accolade in audio, is the direct result of an absence of IM products.

TIM and other factors

Distortion in phase can also occur in an amplifier that causes changes in pulse response; real amplifiers also usually display some am-to-pm and pm-to-am conversion too.

I have deliberately avoided discussion of these since there is considerable doubt whether modest phase effects are audible at all. However it is less contentious to say that over some of the audible range at least, differences in phase response between channels will at the very least degrade or alter the stereo image presentation and are therefore undesirable.

There is another form of intermodulation distortion that has been discussed in audio design, namely transient IMD or TIM. This is the distortion said to occur when a part of an amplifier suffers slew-rate limiting. For a brief period of time, the amplifier is unable to follow the input signal at all. During this time the amplifier gain is zero.

The usual remedy is to ensure effective low-pass filtering prior to any stage that suffers slew-rate limiting. But since the event is transitory, the distortion may not show up in steady-state measurements – particularly the continuous-sine waveforms generally used in total-harmonic distortion measurements.

With a suitable input signal though, (one with a high peak-to-mean ratio perhaps) this should show up in an intermodulation distortion test.

Other test waveforms are often used with amplifiers, for example square waves to show load stability. These may well continue to be necessary, though it may be sufficient – and perhaps preferable – instead to measure the IMD performance with a range of realistic load impedances, since it is the distortion we are primarily interested in.

In summary

I have shown here that current testing methodology fails to test the performance of audio amplifiers adequately in a manner that relates to audible performance. It fails to measure what it purports to do, namely the difference between a representative complex and time-varying signal input to an amplifier and the actual output from it.

Instead, testing concentrates on an extremely narrow interpretation of 'distortion' that the ear doesn't actually hear as distortion at all; it makes no attempt to measure important types of distortion that certainly are audible; and it does not apply the same rigour to frequency-response issues that it does to THD.

Tests for load stability are also generally done separately to other tests. This is presumably done on the assumption that variations in loads can't possibly affect other aspects!

In my view, any real test of an amplifier should apply a representative complex signal and compare this with the actual output of the amplifier under a range of likely loads. This could be done in many ways, with real or artificial sources and measured over frequency or time.

A suitable and relatively simple means exists which is already used in other fields, namely IMD measurement under multi-tone conditions. The exact format of these tests could, and should, be adjusted to maximise their relevance to the particular case of wide-band ultra-low distortion audio amplifiers. These or other measurements should also be capable of measuring frequency response to a far greater level of accuracy than is current.

When appropriate and psycho-acoustically relevant tests are available, then perhaps we can better assess audio amplifiers objectively and better relate their objective performance to subjective tests.

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The system has a wide range of applications in remote control, security, radio-linked message distribution and personnel paging.

How it works

Parallel-to-serial encoders, serial-to-parallel decoders, radio transmitters and receivers are used in the system.

Inside a transmitter unit, the HT640

encoder converts an 18-bit parallel data into a serial data. The first 10 bits of data represent address and the other 8 bits represent data.

The encoded serial data is fed into a radio TX2 transmitter, in which the serial data modulates a 418/433MHz-carrier signal using the FM modulation scheme. The radio signal is then transmitted to the surrounding area through an antenna, Fig. 2a).

Inside a receiver unit, the radio signal from the antenna is demodulated by an RX2 radio receiver module. Demodulated serial data is fed into the HT648L or HT658 serial-to-parallel decoder, that converts the serial data back to the parallel data (10-bit address and 8-bit data), Fig. 2b).

The address is compared with the pre-set address of the decoder. If they match, the 8-bit is placed to the output. If the address does not match, the decoder ignores the present data reception. As a 10-bit binary data has 1024 possible combinations, the maximum number of receiver's addresses is 1024.

Transmitter and receiver modules

The radio transmitter and receiver modules make the digital radio link so easy to be implemented. They are surface acoustic wave (SAW) controlled FM radio transmitters and receivers specially designed for radio telemetry and tele-command applications.

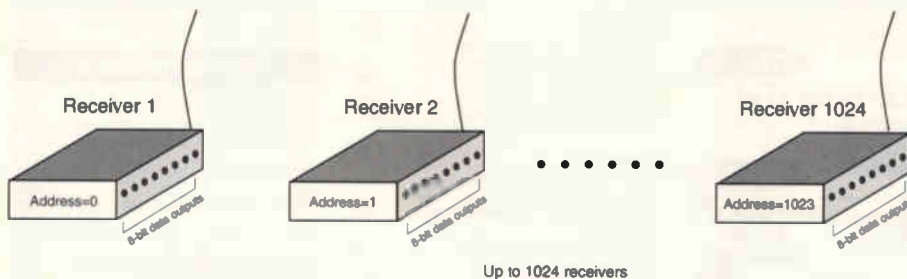
Each module is type-approved to the Radio-communications Authority in the UK and in Europe. This means that there is no need to submit the project for type approval. For details of UK MPT1340 and European ETS300-200. Details of the modules are described in the data sheets in reference 1.

A variety of TX2 and RX2 modules can be used with this design, as you will see from Table 1.

Transmitter module. Pin functions of

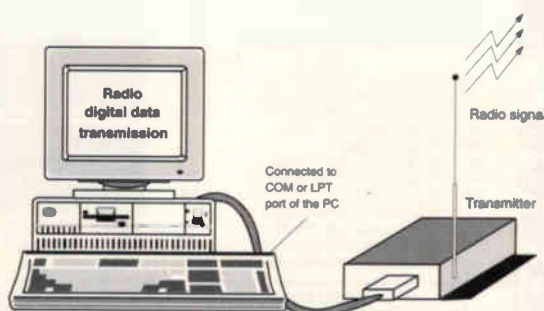
Table 1. TX2 and RX2 radio link module options.

Parameters	Description	Example
Frequencies	418.00MHz for UK use 433.92MHz for European use	TX2-418: 5V TX, -6dBm TX2-433: 5V TX, 9dBm
Supply voltages	5V (4-6V for TX2 and RX2) 3V (2.2V-4V for TX2, 3-4V for RX2)	
RX data rate	-A: 7kHz baseband BW, slow data up to 14kbps -F: 20kHz baseband BW, fast data up to 40kbps	RX2-433-3V RX2-418-A-3V RX2-433-A-3V



Up to 1024 receivers

Fig. 1. This radio data transmission design allows one transmitter to send an 8-bit data to 1024 receivers. The transmitter connects the RS232 port of a PC.



the transmitter are given in Fig 3a). For the +5V and 433MHz version, the operating voltage ranges from 4 to 6V DC. At 5V, typical current consumption is about 10mA.

For the +3V and 433MHz version, a supply between 2.2V and 4V DC is needed, with a typical current consumption of 6mA at 3V. Digital data to be sent – which should be at CMOS logic level at the same power supply voltage – is fed to pin 5. An antenna connects to pin 2. Figure 4a) is a block diagram of the transmitter.

The transmitter's antenna can be a helical, loop or whip type. Fig. 5. Of these, the helical antenna is most compact, but it needs to be optimised for the exact wavelength in use. The loop antenna can consist of a loop of PCB track, which is tuned by a variable capacitor. The whip-type antenna is a wire, rod, PCB track, or a combination of them.

Wireless receiver. Figure 3b) gives the pin functions of the receiver. For the +5V version, the operating voltage ranges from 4 to 6V DC. Typical current consumption is about 13mA at 5V. For the +3V version, the operating voltage is between 3 to 4V DC and it needs around 13mA at 3.5V.

Received output data at CMOS logic levels appears at pin 7. RXD. Pin 3 is a carrier-detect output. It can be used to drive an external p-n-p transistor to obtain a logic level carrier detect signal. If not used though, it should be connected to +5V. The block diagram of the receiver is shown in Fig. 4b).

Any types of the antenna previously described in the transmitter section can be used with this module.

Encoding and decoding

The HT640 and HT648L/HT658 are CMOS LSI encoders and decoders designed for transmitting and receiving digital code, Fig. 6. Typical applications are given in Fig. 7. Details of the modules are given in reference 2.

HT640 encoder. The HT640 converts 18-bit parallel data into a serial data. It transmits the serial data on receipt of a low-to-high transition at the transmit-enable pin, TE. The 18-bit data comprises 10 bits of address, A0 to A9, and 8 bits of data, D0 to D7.

The chip has an on-board oscillator that relies on an external 5% resistor connected between pins 10 and 11. It has a wide operating voltage from 2.4V up to 12V with a typical standby current of 1µA for a 3V supply.

Figure 8a) is a flowchart of the device's operation. Timing of the encoder is shown in Fig. 9. Initially the encoder is in stand-by mode. On

Technical support

A designer's kit is available from the author. It includes PCBs, components and VB5 software. Please direct your enquiry to Dr Pei An, 11 Sandpiper drive, Stockport, Manchester SK3 8UL, Tel/fax/answer: 44-(0)161-477-9583. E-mail: pan@intec-group.co.uk

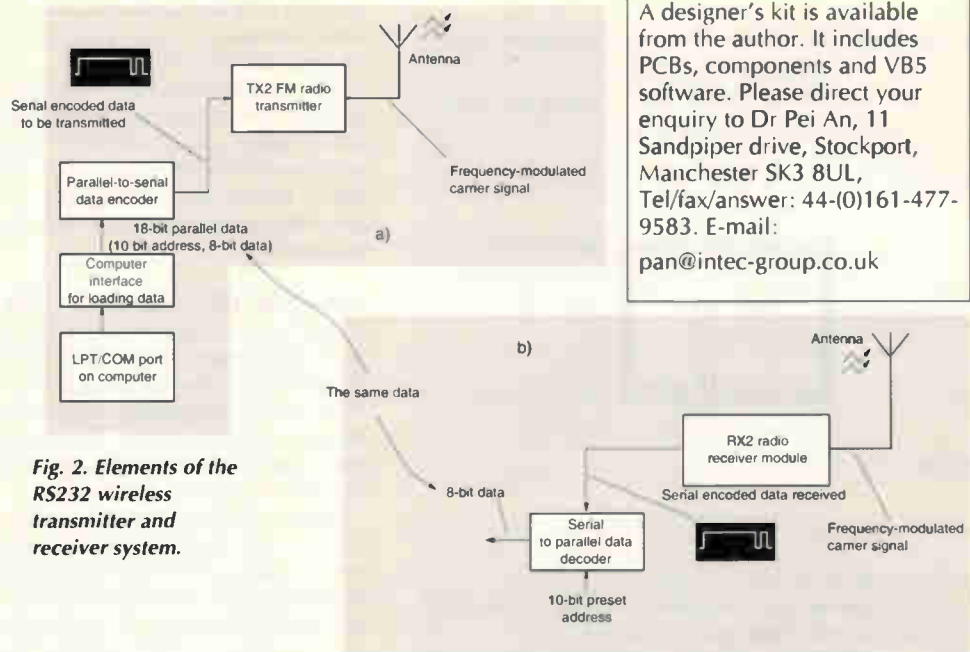


Fig. 2. Elements of the RS232 wireless transmitter and receiver system.

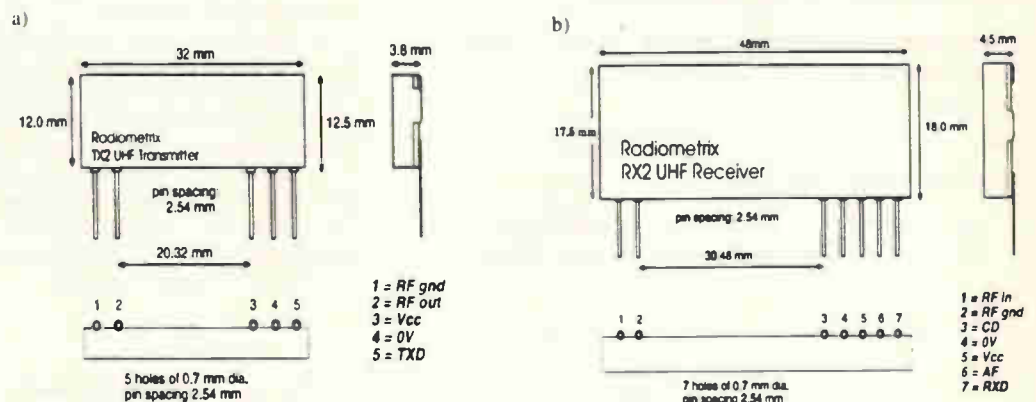


Fig. 3. Pin-out of the TX2 radio transmitter and RX2 receiver modules. 418MHz and 433MHz versions are available. They are type-approved by UK and European radio communication authorities.

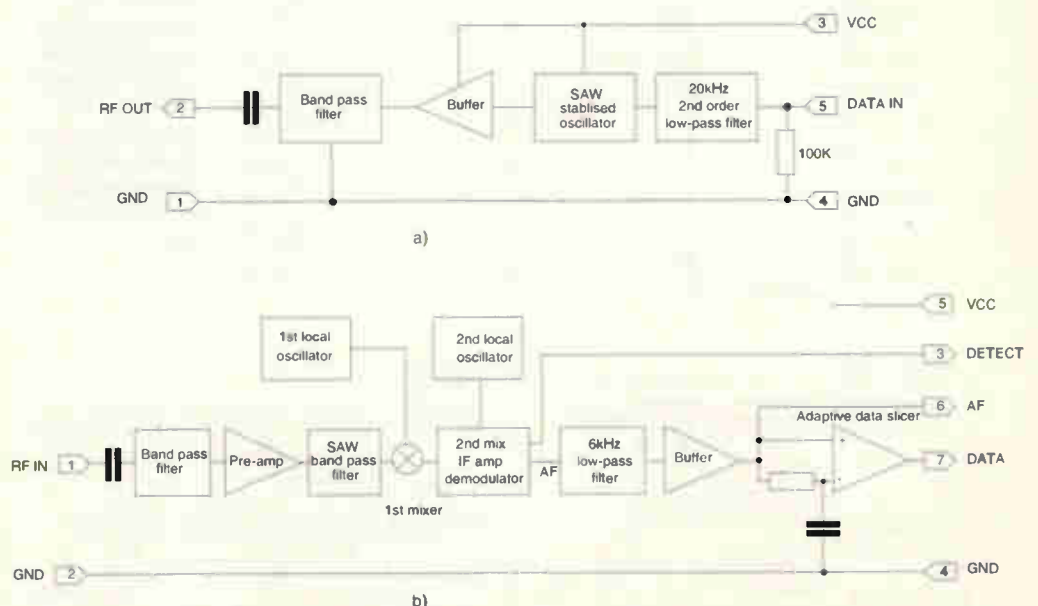
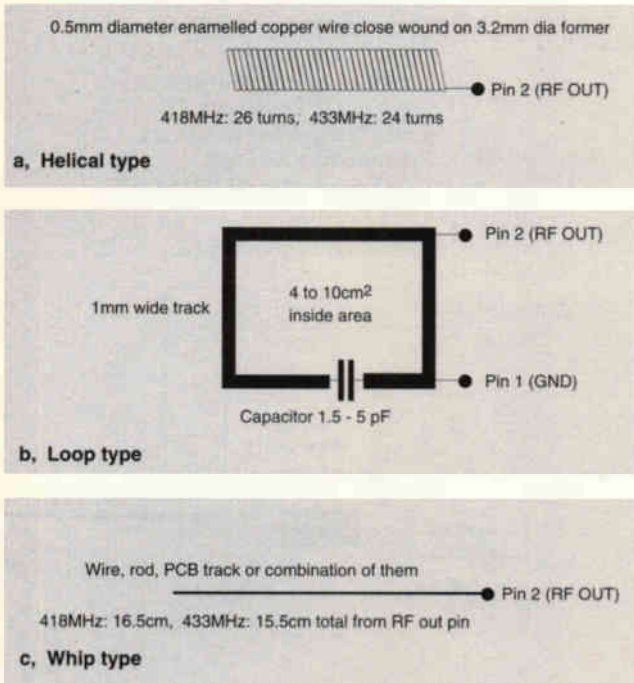


Fig. 4. Internal block diagrams of the TX2 and RX2 modules.



Antenna performance chart	Helical	Loop	Whip
Ultimate performance	✓✓	✓	✓✓✓
Ease of set-up	✓✓	✓	✓✓✓
Size	✓✓✓	✓✓	✓
Immunity to proximity de-tuning	✓✓	✓✓✓	✓

Fig. 5. Three types of antennas can be constructed and used with the TX2 and RX2 modules – helical, loop and whip.

receiving a TE signal, it begins a three-word transmission cycle and repeats the cycle until TE goes low.

Each word contains four periods: the pilot code period, synchronisation code period, address period and data period as shown in Fig. 9b). Logic levels '0' and '1' are encoded as in Fig. 9c). An open state can be also encoded, but it is not used here.

HT648L/HT658 decoder. The HT648L or HT658 receives the 18-bit word and interprets the first 10 bits as the address and the last 8 bits as data. When the received address matches the decoder's pre-set address, the valid-transmission output, VT, goes high and the 8-bit data appears at the output.

The device operates on supplies from 2.4V to 12V with a typical standby current 0.1µA at 5V. Figure 8b) shows its encoding flow and Fig. 9d) its timing. Initially the encoder is in stand-by mode. A signal on DIN activates the oscillator, which in turn decodes the incoming address and data.

The decoder interprets the first 10 bits as address and the last 8 bits as data. Each decoder checks the received address twice continuously. If all the received addresses match the address of the decoder, the data are output to the output pins and the VT pin goes high to indicate a valid transmission. That condition lasts until the address is incorrect or no signal is received.

The HT648L has latched outputs. Valid data appears at the outputs during a valid transmission, and is latched until the next valid transmission. The HT658 has momentary outputs. Data only appears at the outputs during a valid transmission and then resets.

UCN5833 serial latch. The UCN5833A is a 32-bit serial-input latched driver, Fig. 10. It has 32 bipolar Darlington open-collector drivers. Each is capable of driving 150mA with a maximum control voltage of 40V.

The IC consists of a data latch for each driver, two high speed 16-bit shift registers and control circuitry. It is controlled via four CMOS digital input lines, which can be driven directly by outputs from a computer. The maximum data input rate is 3.3MHz.

Timing for the latch is shown in Fig. 11. A serial data bit present at the input is shifted into the shift register on the transition from 0 to 1 of the clock input. On subsequent clock pulses, the registers shift data towards the serial data output. Serial data must be stable at the input prior to the rising edge of the clock input.

Data bits stored in the 32 registers are transferred into output latches when the strobe input is high. The latch continues to accept new data as long as the strobe is high.

Data is latched at the high-to-low transition of the strobe. When the output-enable input is low, all the output buffers are turned off. When it is high, the status of outputs is controlled by the contents of the latches.

Transmitter circuitry

Figure 12 is the transmitter's circuit diagram. Three lines from the pc's RS232 port control data loading into the UCN5833A. The DTR line controls the CLK input; RTS line controls the DATA and TD line controls the STROBE.

Lines from the RS232 port are clamped to +5V by zener diodes D₁ to D₃. Details of the RS232 port, Fig. 14, and how to use it are described in reference 3. It is also possible to connect the transmitter to a computer's parallel printer port.

On the UCN5833, OUT1 to OUT10 supply an address to the encoder and OUT11 to OUT18 supply the data. Output OUT19 controls the transmit-enable of the encoder.

All the lines are pulled up to the +5V supply rail by a 10kΩ resistor from one of the resistor arrays, RL₁, RL₂ or RL₃. Outputs OUT25 to OUT32 are pulled up by a further 10kΩ resistor array, RL₄. Output OUT25 switches the power supply to the encoder and radio transmitter on or off.

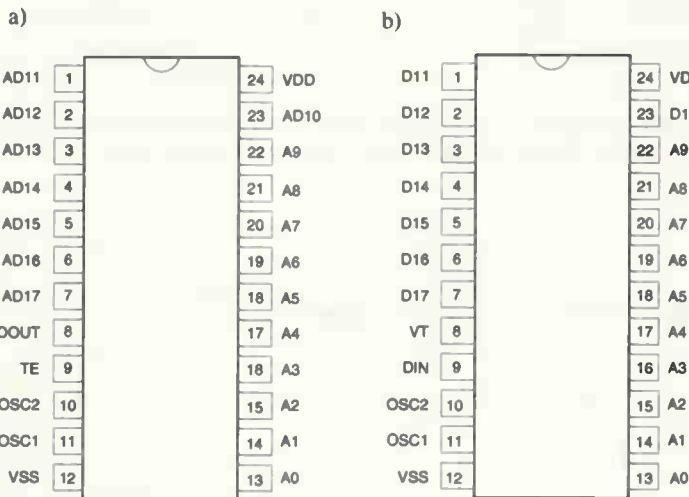
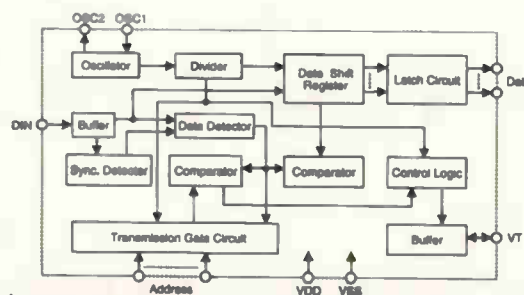
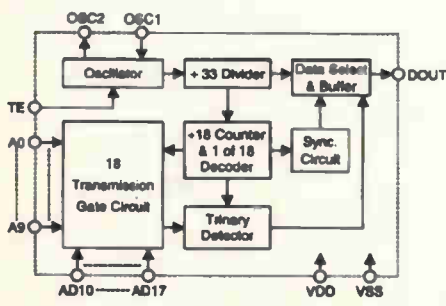


Fig. 6. Pin-outs and internal block diagrams of HT640 encoder and HT648L/HT658 decoders. The encoder is able to encode 18 bits of parallel data into a serial data. The decoder interprets the first 10 bits as address bits and the last 8 bits as data bits.



For the HT640 encoder, the value of the external resistor is chosen as 150kΩ at 1%, resulting in oscillation at 200kHz. At this rate, each data transmission takes about 0.075 seconds.

Power supply to the UCN5833 is generated by a low power +5V low drop-out voltage regulator, namely a TC55RP0052EZB. An LK115D05 voltage regulator produces the +5V power supply to the HT640 and the radio transmitter.

This regulator has an on/off control pin. When the pin is low, the LK115D05 switches on the +5V supply. When it is high, the supply is

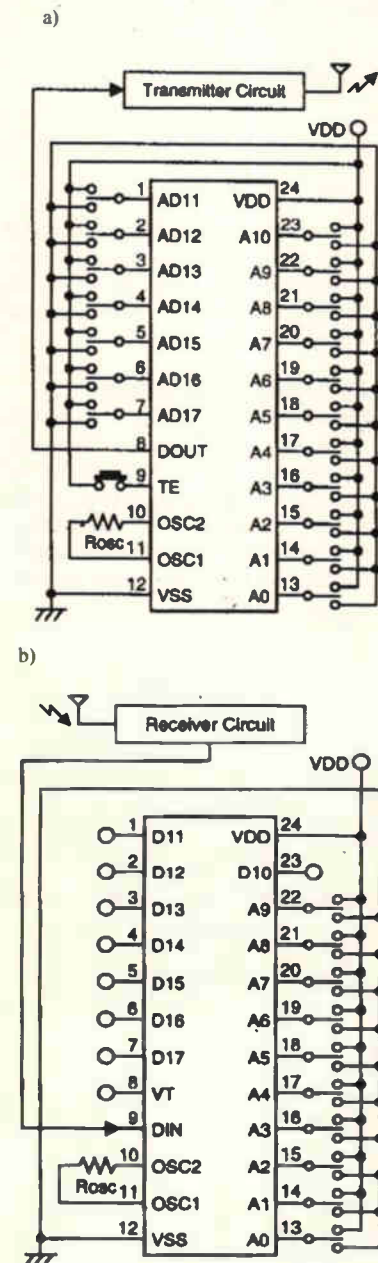


Fig. 7. Typical application circuits for the HT640 and HT648L/HT658. In the present application, the radio link is used to transmit data from the encoder to the decoder.

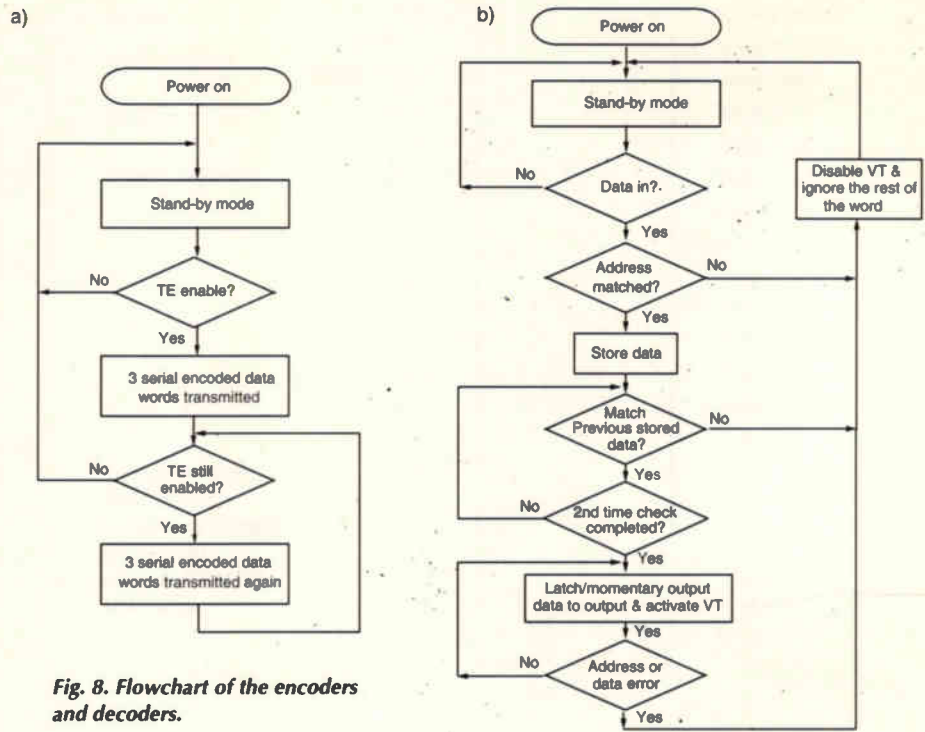


Fig. 8. Flowchart of the encoders and decoders.

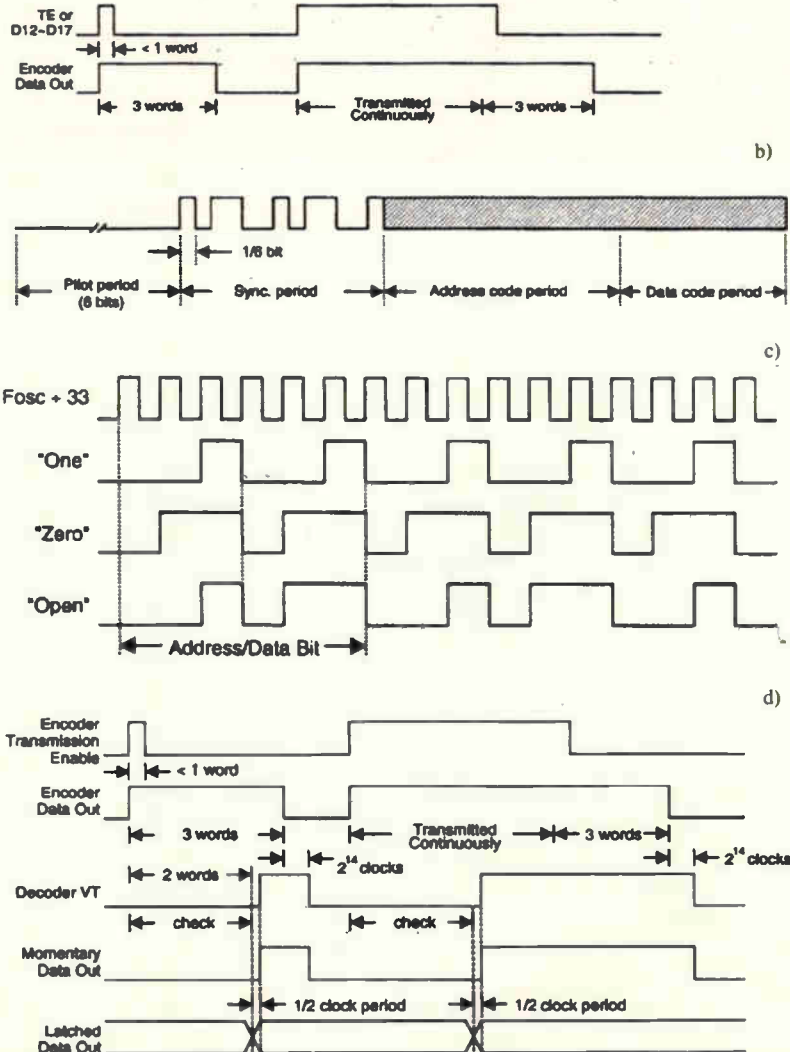
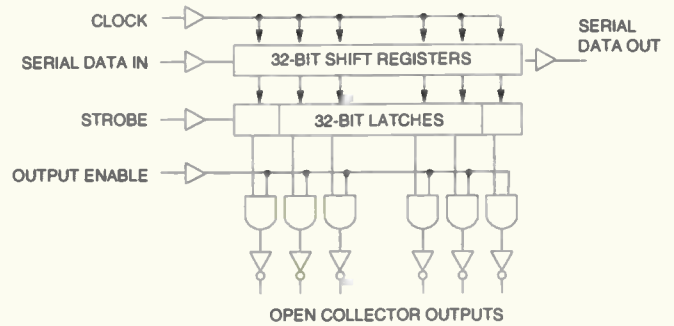
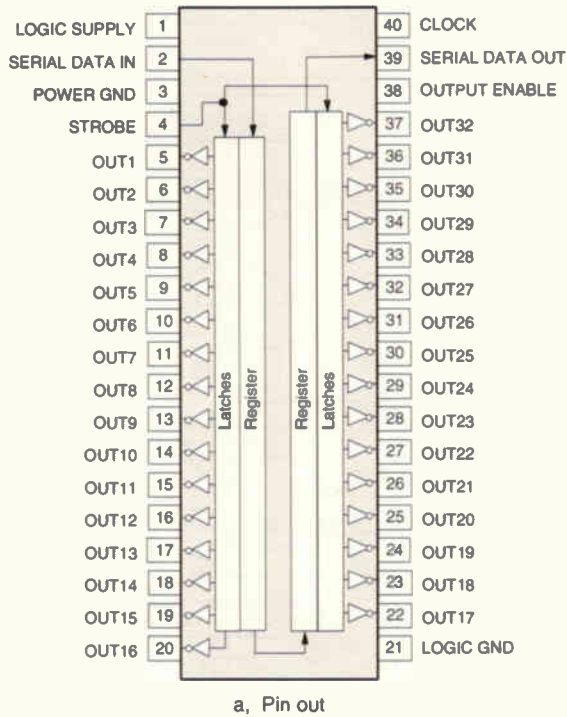
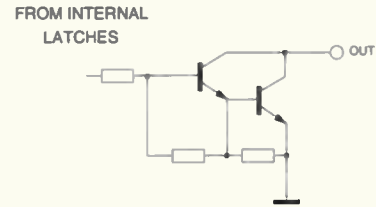


Fig. 9. Encoder timings are shown in a), b) and c) while d) shows that of the decoder.



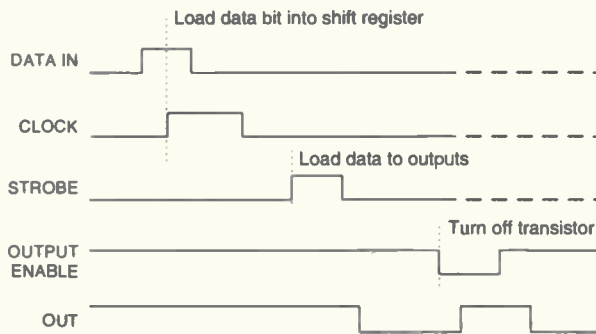
b. Internal block diagram



c. Output stage

Fig. 10. Pin-out, internal block diagram and output stage of the UCN5833A serial data latch.

Fig. 11. Timing sequence of the serial data latch. Three input lines – Data in, Clock and Strobe – are required to control data latch into the IC.

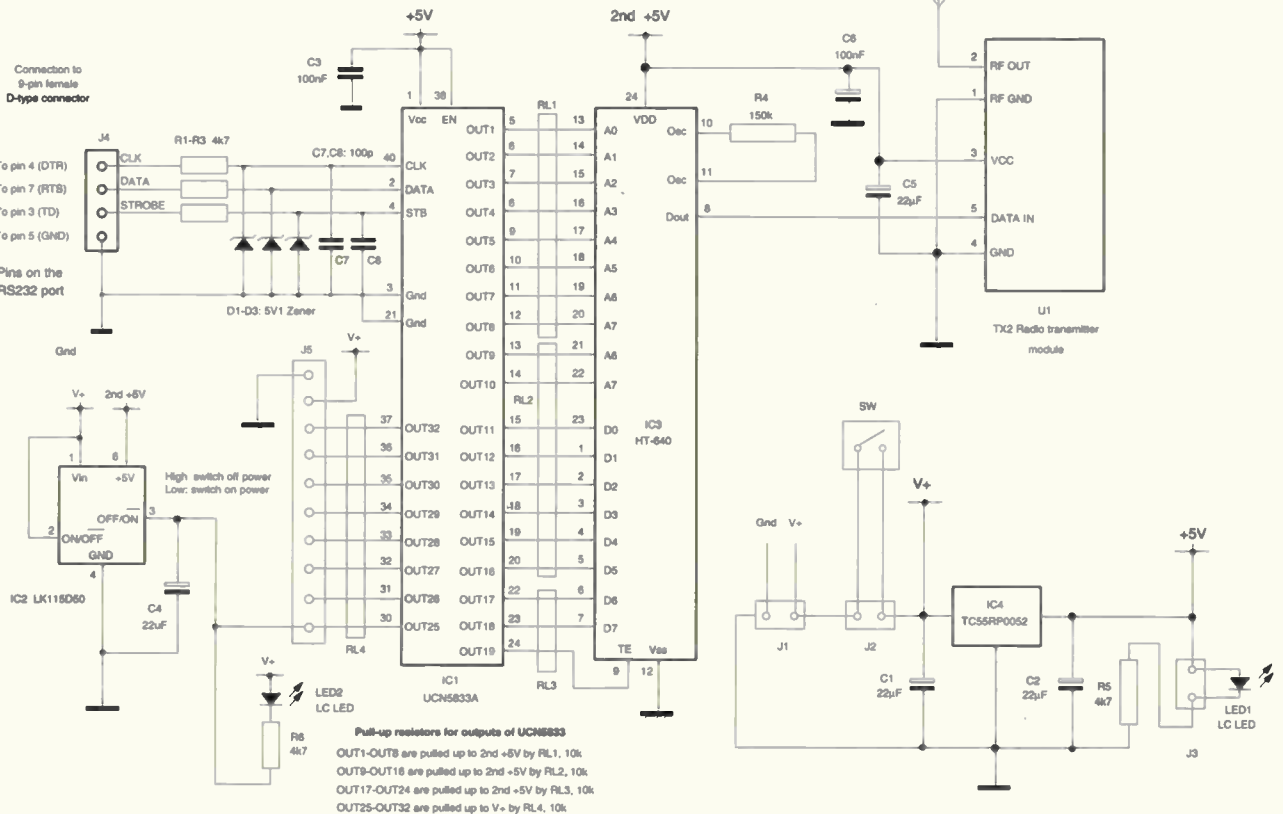


switched off. This scheme is adopted in order to save power when the transmitter is idle. In this mode, the current consumption of the board is around 1.5mA.

Receiver circuit details

Figure 13 is the receiver's circuit diagram. This module receives the transmitted radio signal from its helical antenna. Demodulated output is then fed into the decoder. For the decoder, the value of the external resistor is chosen as 140kΩ at 1%, which gives an

Fig. 12. Circuit diagram of the transmitter unit. The serial data latch is used as a computer interface and HT640 is the encoder. The transmitter, TX2, connects to the RS232 port of a computer. It is also possible to connect it to a computer's parallel port.



Pull-up resistors for outputs of UCN5833
 OUT1-OUT8 are pulled up to 2nd +5V by RL1, 10k;
 OUT9-OUT16 are pulled up to 2nd +5V by RL2, 10k;
 OUT17-OUT24 are pulled up to 2nd +5V by RL3, 10k;
 OUT25-OUT32 are pulled up to V+ by RL4, 10k.

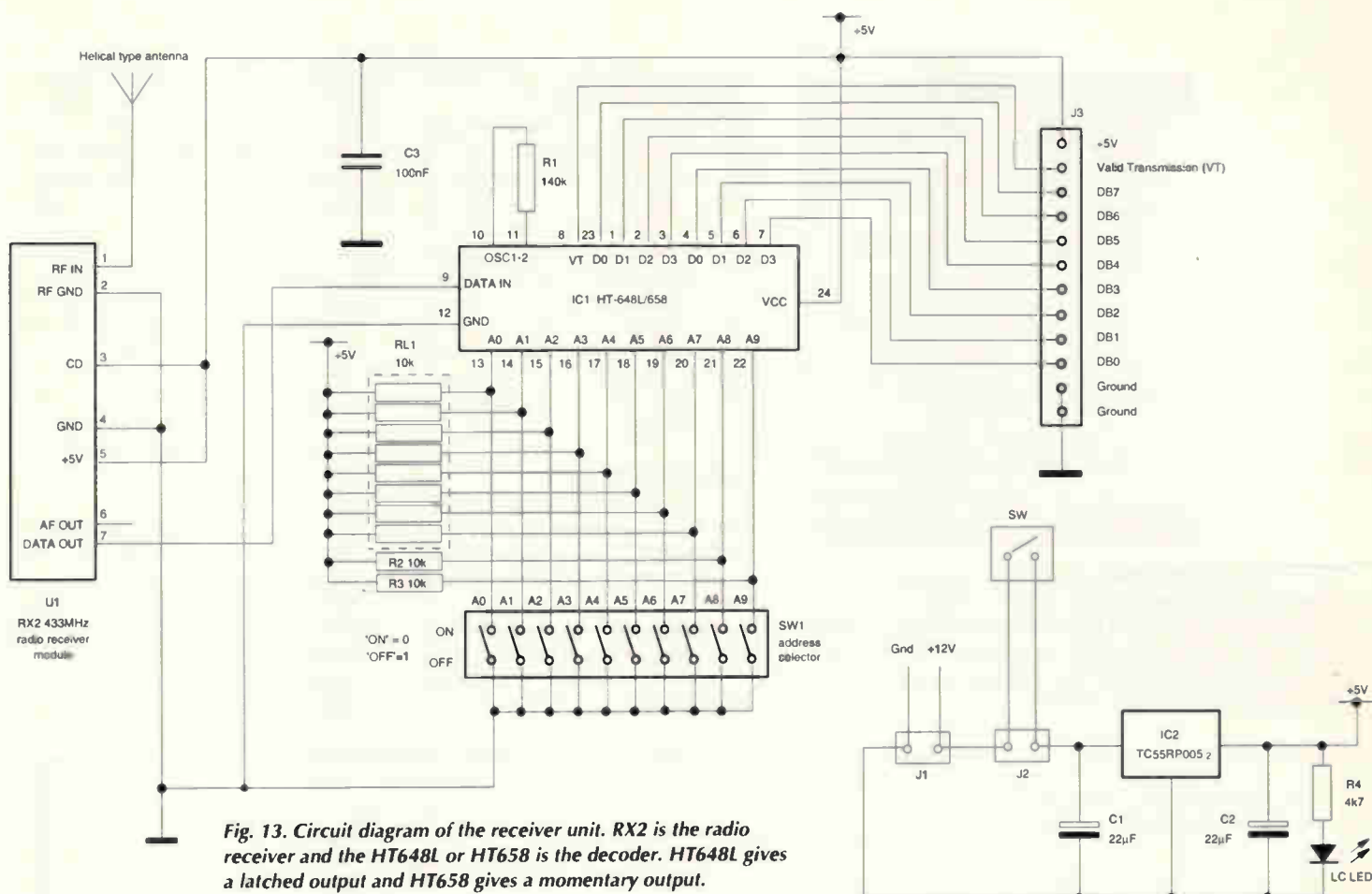


Fig. 13. Circuit diagram of the receiver unit. RX2 is the radio receiver and the HT648L or HT658 is the decoder. HT648L gives a latched output and HT658 gives a momentary output.

oscillation frequency of 200kHz.

The address of the decoder is set by a 10-way dip switch. Outputs from the decoder and VT line are available from J3. A low-power, low drop-out voltage regulator, the TC55RP0052EZB, produces +5V power supply.

Programming

Operation of the transmitter is controlled by the DTR, RTS and TD lines of the RS232 port. Data bits to be loaded are put onto the RTS line and are shifted into the shift register of the UCN5833 at the low-to-high transition of DTR, connected to the clock line.

After 32 clocks, 32 bits are loaded into shift registers. Next a low-to-high-then-low pulse is applied to the TD line, connected to the Strobe line, to latch the data to the outputs OUT1 to OUT32.

In idle mode – i.e. with the transmitter not transmitting any data – OUT25 is at logic high to switch off power supply to the encoder and radio transmitter to save power. A data transmission comprises the following procedure.

Firstly, OUT1 to OUT10 are loaded with address bits; OUT11 to OUT18 are loaded with data; OUT19 is loaded with '1' to enable the TE line.

Next, OUT25 becomes low for 0.2

second. During this period, the HT640 and the radio transmitters are activated to transmit the encoded data. After this, OUT25 is brought high again to stop data transmission and to enter power save mode again.

A demonstration program has been developed for testing the functionality of the system using the Visual Basic 5 language. This language is supplied with a serial port control called 'MSCOMM' that is used to control all operations of the serial port.

In the VB5 editor, if the name of the MSCOMM control is declared as 'MSCOMM1', the following commands can be used to control the logic status of the DTR, RTS and TD lines. For details of the MSCOMM control, have a look at reference 4.

To make DTR line high or low, use:

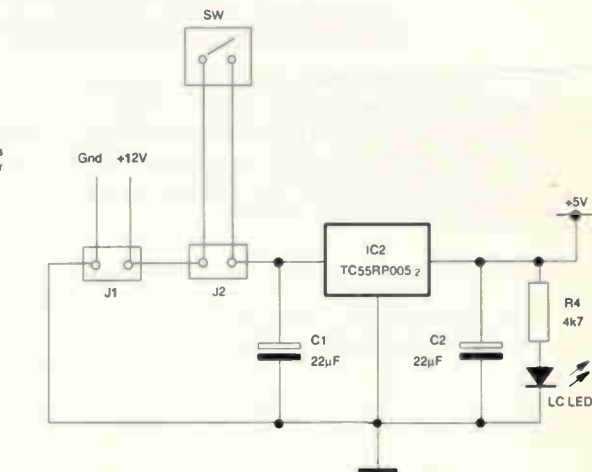
```
MSCOMM1.DTREEnabled=True
or
MSCOMM1.DTREEnabled=False
```

To make RTS line high or low, use:

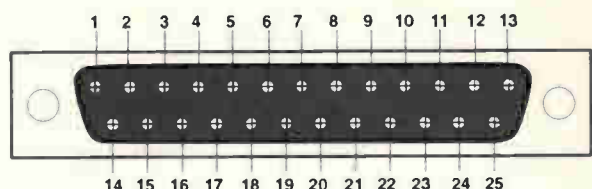
```
MSCOMM1.RTSEnabled=True
or
MSCOMM1.RTSEnabled=False
```

To cause TD line to generate a low-to-high-then-low pulse, use:

```
MSCOMM1.Output="0".
```



(a) 9-pin male socket viewed from the back of the computer



(b) 25-pin male socket viewed from the back of the computer

Pin functions of the RS232 connectors

25 pin	9 pin	Name	Direction (for PCs)	Description
1		Prot	-	Protective ground
2	3	TD	Output	Transmit data
3	2	RD	Input	Receive data
4	7	RTS	Output	Request to send
5	8	CTS	Input	Clear to send
6	6	DSR	Input	Data set ready
7	5	GND	-	Signal ground (common)
8	1	DCD	Input	Data carrier detect
20	4	DTR	Output	Data terminal ready
22	9	RI	Input	Ring indicator
23		DSRD	I/O	Data signal rate detector

Fig. 14. Pins and functions of the RS232 port. In this design, TD, DTR and RTS lines control data loading into the serial-data latch. The COM port bit rate should be set to 9600 baud, 8 data bits, 1 stop bit and no parity bit.



Fig. 15. User screen of the demonstration software, which allows you to send an 8-bit data word to a receiver specified by the address. If the 'Start testing' button is clicked, a square wave signal will be produced at the outputs of the specified receiver unit. The periods of logic high and low are determined by the value of 'Interval'.

Examples of how to use 'MSCOMM' control are given in List 1.

The user's screen of the demonstration software is shown in Fig. 15. From that screen, the user specifies an address and a data byte to be sent. Clicking 'Send data' causes the transmitter to transmit the data to the specified receiver. If the 'Start testing' button is pressed the specified receiver outputs a square wave signal. The periods of high and low state are determined by the interval — which is a value bigger than 500ms. Click 'Stop' to stop testing.

Application ideas

In a remote control application, the eight digital outputs from the receiver could control lights, heaters, motors, etc., using mechanical or solid-state relays.

In an information distribution or message display system, it would be possible to send a message to particular groups of receivers with their own dis-

play units. Because of the addressing capability, it would also be possible to set up a form of paging system.

Finally, I would like to thank Mr Kangyan from Radiometrix Ltd for his help with this design.

References

1. Data sheets for TX2 and RX2 are available from Radiometrix Ltd. Tel: + 44 (0) 20 8428 1220. Web site: www.radiometrix.co.uk
2. Data sheets for HT640, HT648L and HT658 are available from Holtek. Web site: www.holtek.com.tw
3. PC Interfacing – Using Centronic, RS232 and game ports, Pei An, Newnes, Butterworth-Heinemann, 1998, ISBN0240514483.
4. Real-world programming with Visual Basic, Anthony T. Mann, SAMS publishing, 1995, ISBN0672306190.

List 1. Demonstration program for the wireless RS232 link in Visual Basic.

```

Function loaddata_1(dataA As Byte, dataB As Byte, dataC
As Byte, dataD As Byte)
'Load 4 bytes (8 bits) into the UCNXX
'DTR = clock, RTS = data, TX = strobe
'If UCNS833/32 is used: RTS=true, output low (transistor
on),
'RTS=false, output high (transistor off)
Dim i As Byte
'load Port D
For i = 1 To 8
MSCOMM1.RTSEnable = 1 - (dataA And bitweight(9 - i))
\ bitweight(9 - i)
MSCOMM1.DTREnable = True
MSCOMM1.DTREnable = False
Next i
'load Port C
For i = 1 To 8
MSCOMM1.RTSEnable = 1 - (dataC And bitweight(9 - i))
\ bitweight(9 - i)
MSCOMM1.DTREnable = True
MSCOMM1.DTREnable = False
Next i
'load Port B
For i = 1 To 8
MSCOMM1.RTSEnable = 1 - (dataB And bitweight(9 - i))
\ bitweight(9 - i)
MSCOMM1.DTREnable = True
MSCOMM1.DTREnable = False
Next i
'load Port A
For i = 1 To 8
MSCOMM1.RTSEnable = 1 - (dataA And bitweight(9 - i))
\ bitweight(9 - i)
MSCOMM1.DTREnable = True
MSCOMM1.DTREnable = False
Next i
'Strobe data into UCN serial latch IC
Sleep (20)
MSCOMM1.Output = "0"
End Function

Function loaddata_2()
Dim outbuf(1) As Byte
'Load 32 bytes into UCNXX
outbuf(1) = 255
For i = 1 To 32
MSCOMM1.RTSEnable = 1 - DataBits(33 - i)
MSCOMM1.DTREnable = True
MSCOMM1.DTREnable = False
Next i
Sleep (20)
MSCOMM1.Output = outbuf
End Function

Function save_power()
Dim outbuf(1) As Byte
'Load 32 bytes into UCNXX
outbuf(1) = 255
For i = 1 To 32
MSCOMM1.RTSEnable = 0
MSCOMM1.DTREnable = True
MSCOMM1.DTREnable = False
Next i
Sleep (20)

```

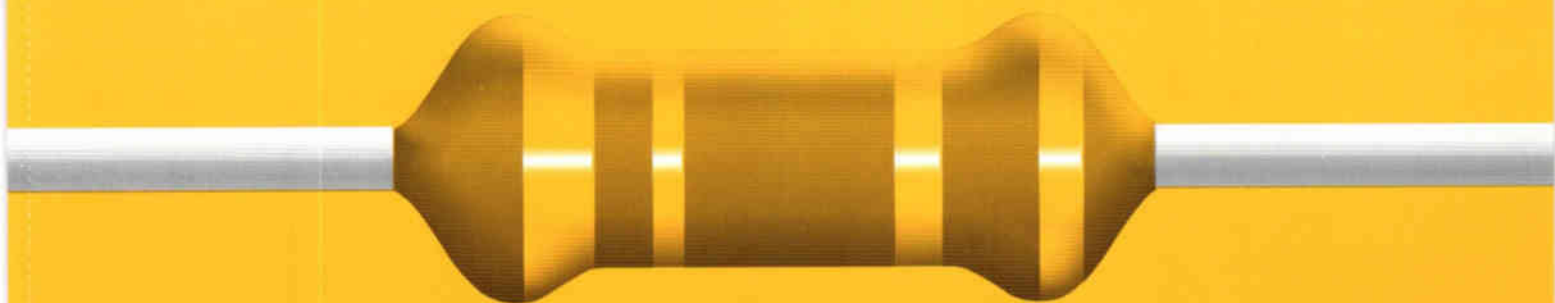
List 2. Module 1.bas.

```

Global bitweight(8) As Byte
Global done As Boolean
Global dataA As Byte, dataB As Byte, dataC As Byte,
dataD As Byte
Global DataBits(32) As Byte
Global Com_number As Variant
Declare Sub Sleep Lib "kernel32" (ByVal dwMilliseconds
As Long)

```

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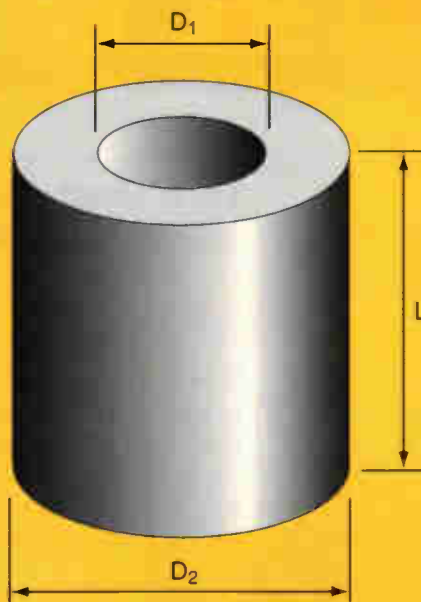
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218-003	1	3	5
219-001	2	4	6
219-002	1.3	4.5	8

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Transmitter – TX3

- Crystal-locked PLL, FM modulated at up to 50 kbps
- Operation from 2.2V to 10V @ 10mA
- Built-in regulator for improved stability and supply noise rejection
- 0dBm (1mW) RF output



Receiver – RX3

- Single conversion FM superhet with SAW front end filter
- Operation from 2.7V to 13V @ 9.5mA
- Built-in regulator for improved stability and supply noise rejection
- 50kb/s, -100dBm sensitivity @ 1ppm BER
- RSSI output with 75dB range

The TX3 and RX3 are miniature UHF radio transmitter & receiver modules designed for PCB mounting. They allow the simple implementation of data links at speeds up to 50kbps and distances up to 30m in-building or 120m over open ground.



Universal Evaluation Kit

This evaluation system can be used to evaluate a TX1 or TX2 or TX3 transmitter module together with matching RX1 or RX2 or RX3 receiver or a BiM module.

Range of facilities

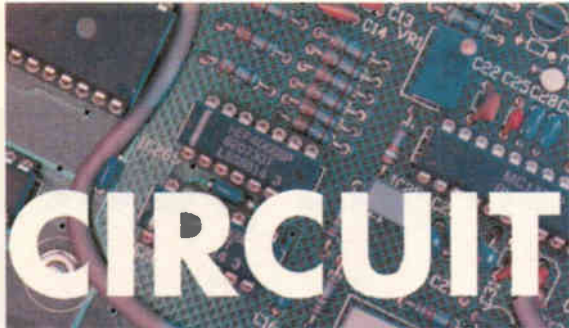
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CIRCUIT IDEAS

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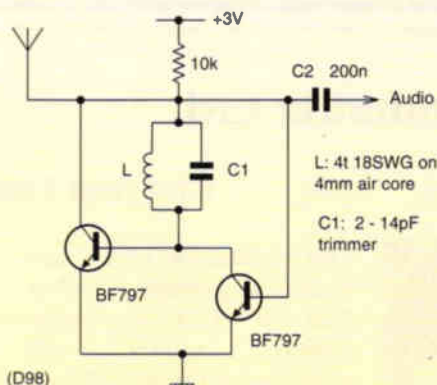
Your submissions are judged mainly on their originality and usefulness. Interesting modifications to existing circuits are strong contenders too – provided that you clearly acknowledge the circuit you have modified. Never send us anything that you believe has been published before though.

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.

Two transistor FM broadcast receiver

This simple FM radio receiver uses only two transistors. They function as a low-power oscillator,



This synchronised oscillator recovers the audio from a broadcast FM station.

the frequency being determined by L and C_1 .

When the frequency of oscillation is the same as that of the wanted signal, the recovered audio is available at the output. The audio signal is dc blocked by C_2 and fed to an audio amplifier.

Rajik Gorland
D98

This circuit was so intriguing that we had to see if it worked. It does. Thanks to Ian Hickman for trying it out. Ed.

Winner of the second National Instruments digital multimeter worth over £500

Simple phase-sensitive detector

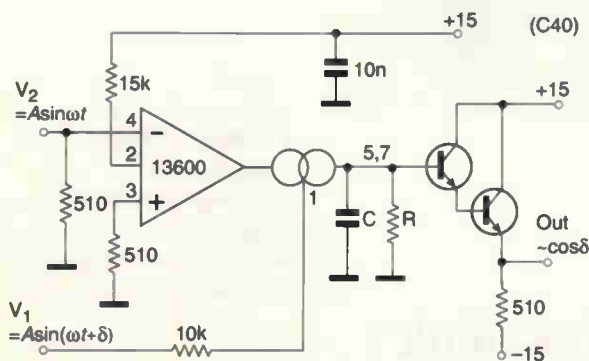
The XRL3600 is a transconductance op-amp with linearising diodes and buffer, similar to the LM13600. It can be used as a modulator or audio range mixer.

Here, it is shown connected as a phase detector, the output being proportional to $\frac{1}{2}\cos\delta$. The

This phase-sensitive detector can be used as a modulator or audio-range mixer and, having only one op-amp, it is easy to implement.

advantage of using this device is its simplicity, since the CR low-pass filter to suppress the carrier frequency ripple and its harmonics can be added 'inside' the op-amp, so that no other op-amp is needed. The cut-off frequency ω_0 of the low-pass filter is $\omega_0 = 1/RC$, where ω_0 should be much less than 2ω , the frequency of the input signals.

Kamil Kraus
Rokycany
Czech Republic
C40



National Instruments sponsors Circuit Ideas

National Instruments is awarding over £3500 worth of equipment for the best circuit ideas.

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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.

NI4050

The NI 4050 is a full-feature digital multimeter (DMM) for hand-held and notebook computers with a Type II PC Card (PCMCIA) slot. The NI 4050 features accurate 5½ digit DC voltage, true-rms AC voltage, and resistance (ohms) measurements. Its size, weight, and low power consumption make it ideal for portable measurements and data logging with hand-held and notebook computers.

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- AC Measurements: 20mV rms to 250V rms; 20mA rms to 10A rms;
- True rms, 20Hz to 25kHz
- Up to 60 readings/s
- UL Listed
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LabVIEW

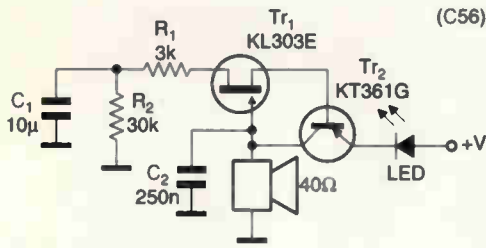
LabVIEW is a highly productive graphical programming environment that combines easy-to-use graphical development with the flexibility of a powerful programming language. It offers an intuitive environment, tightly integrated with measurement hardware, for engineers and scientists to quickly produce solutions for data acquisition, data analysis, and data presentation.

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This beeper/blinker operates using a simulated injection field transistor.

(C56)

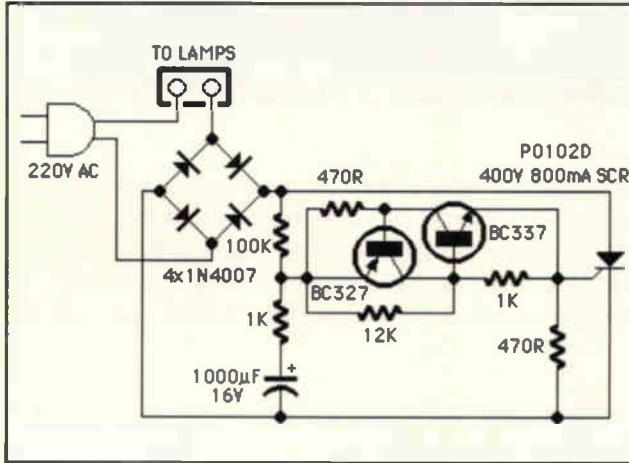
Two-transistor AV alert

A simple audio-visual alerting circuit is described here. Such a circuit can be arranged using an injection field transistor, but these are no longer common items. Instead, the 'beeper' shown emulates an injection field transistor with the combination of an n-channel FET and a p-n-p transistor. These provide a current-voltage characteristic with a negative dynamic resistance. Switching on the beeper generates short synchronous sound and light signals.

Working in conjunction with C_2 , the value of R_1 determines the duration of the sounds; R_2 defines the pause between them.

Capacitor C_2 provides the characteristic colouring or the sound. The circuit uses a 40Ω low resistance phone. At a supply voltage of 6 to 15V, average current consumption is 1.5mA.

Mickael Shustov
Tomsk
Russia
C56



Frugal flasher

Two transistors and an SCR form a reliable replacement to thermally-activated switches used for lamp-flashing. Unlike many similar circuits, this one needs no high-power resistors or high-voltage capacitors.

Timing is provided by 100kΩ and 1kΩ resistors with a capacitor. Best performance is obtained with a capacitor from 470 to 1000µF and with the resistor over the emitter and collector of BC327 set at around 12kΩ.

For proper operation it is essential to use an SCR with a very sensitive gate. If you are unable to find one, you can use triacs such as the TIC206M rated at 600V, 4A. Note that you will still need the diode bridge.

Flavio Dellepiane
Genoa
Italy

Multichannel amplitude discriminator

Amplitude discriminators are used in engineering measurement systems, to route signals into separate channels, on the basis of amplitude. The amplitude of a signal is easily enough indicated by multi-comparator chips driving LEDs, such as the UAA170/180, UL1970N/80N, A277D¹ etc.

The circuit shown uses a UAA180 to direct a signal to one or other of several separate outputs, depending upon its amplitude.

The input is connected to the inputs of twelve CMOS analog switches, and also to a voltage doubler type rectifier circuit. Resultant dc level is smoothed

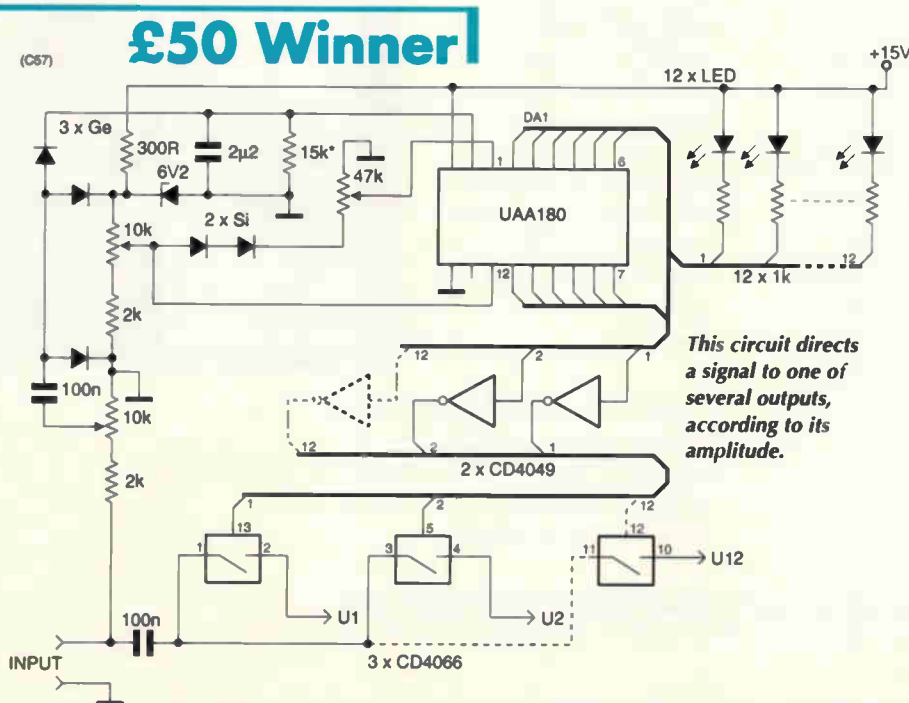
by an electrolytic capacitor and applied to pin 2 of the UAA180.

Upper and lower limits of the 12-step indication range are set by potentiometers connected to pins 3 and 16 of the device, the drop across two silicon diodes setting the minimum range which can be set.

The UAA180 indicates the level of the signal present by lighting the appropriate LED. Via one section of two hex inverters, it also closes one of the analogue switch sections, routing the input signal to the corresponding output channel.

In addition, the device can form the basis of a multichannel analogue quasi-filter, by introducing a frequency-to-voltage converter¹. By summing the output signals via controlled dividers you can synthesise a multiband equaliser, to realise a rejecter circuit, bandpass or other filter.

Mickael Shustov
Tomsk
Russia
C57



This circuit directs a signal to one of several outputs, according to its amplitude.

Reference

1. Shustov M. A. 'Application of polycomparator chips in engineering of radio communications,' *Radioamateur* (Byelorussia), 1997 No 6 pp. 13-15.

Alternative neon tester

It is important that you read the warning panel on the right before going any further.

This idea is an alternative to the traditional screwdriver mains tester incorporating a miniature neon indicator in its handle. It uses a complementary pair of transistors connected as a multivibrator.

When the 510k Ω probe touches a 220V live line, a current not exceeding 400 μ A charges the capacitors via the bridge rectifier, returning through the user (sensor). The device then begins to generate short flashes of light and sound impulses (clicks), at a frequency of up to five per second.

Using high-frequency diodes and an antenna, connected instead of the 510k Ω probe, the device is capable

of remotely registering the presence of high-frequency fields of high strength, e.g. fields of transmitters, horizontal sweep transformers, etc. ■

Mickael Shustov

Tomsk

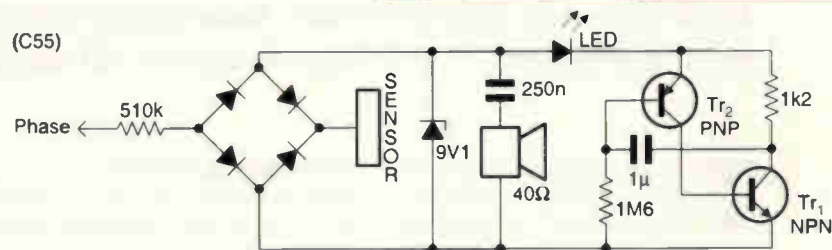
Russia

C55

Warning

Do **not** use this circuit as prescribed here by its author, i.e. do not use the circuit to test the mains or any other potentially lethal voltage source.

The idea is fine, but the circuit is not isolated. If the 510k Ω resistor fails, the circuit immediately becomes lethal if used on mains and high voltages. Ed.



Alternating voltage tester with audible and visible indication.

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TABLE OF CONTENTS

Applications
Applications by description
Applications by part numbers
Company addresses

Books
Circuit Ideas
Information
Subject Index
Analogue Design
Audio
Avionics
Broadcast
Communications
Components
Computing
Consumer Electronics
Control Electronics
Digital & DSP Design
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Efficient battery power supplies

Using the battery-powered regulator he designed for his $\tan\delta$ meter, Cyril Bateman demonstrates how important subtle capacitor parameters are in gaining maximum efficiency.

The ever increasing demand for small, lightweight and efficient battery powered equipment has resulted in an explosion in the number and variety of dedicated power supply integrated circuits. Many of the latest and most efficient designs are only available in minute surface mount packages. These encourage designers to use physically small capacitors and inductors, which can have high losses.

Portable equipment that doesn't need a lot of current can be provided with a stabilised +5V, or lower, rail using only a low drop-out linear regulator, a decoupling capacitor and a PP3 battery. If required, a negative supply can be produced using a charge-pump converter.

Provided the required current is small, this combination provides a low noise, low first-cost system and acceptable battery life. However, the conventional alkaline PP3 size battery, typically rated at 550mAh, exhausts very quickly when providing the 50mA that may be needed to support an efficient $\pm 5V$ supply at 20mA.

Switching alternatives

With increasing load current, battery life using a linear regulation system becomes unacceptable, and a switched-mode alternative becomes essential.

While new battery technologies are now available, many designers choose to use AA-sized alkaline cells when their load cannot be supplied using a PP3 type. To ensure acceptable battery life with a 5V supply, four such cells may be needed in series.

Generating a +5V stabilised supply from four AA batteries poses the difficulty that with fresh batteries, the circuit must reduce the battery voltage. As the batteries discharge, the circuit must automatically change over to boosting the battery voltage.

Capable of very high efficiency, the switching process takes current from the battery supply in bursts. Batteries possess internal resistance, so these current pulses impose fast transients on the supply voltage and result in significant output noise levels. These transients may be minimised by choosing the capacitors and inductors carefully.

There's more on this in the panel entitled, 'Batteries'.

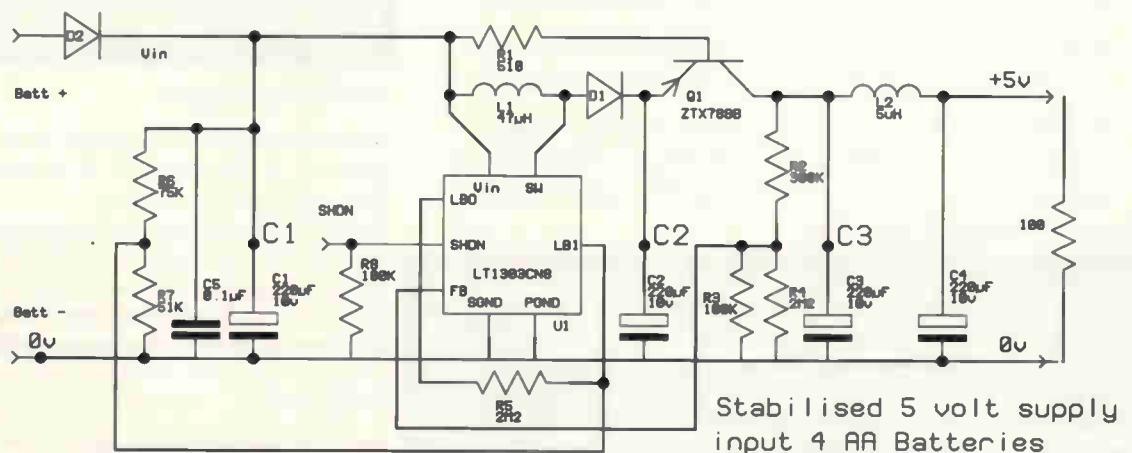
I experienced this problem first hand when designing the +5V switched-mode power supply featured in my $\tan\delta$ meter circuit¹. In this design, the LT1303CN8 integrated circuit² boosts the battery voltage just enough to enable the transistor linear regulator to provide the required +5V output.

Using this circuit as an example, I will demonstrate to you the change in transient voltage levels that I measured using different capacitors in the C_1 , C_2 and C_3 positions, Fig. 1. These measurements illustrate the importance of careful component selection.

A switched-mode power supply draws 'pulses' of current from the battery. This intermittent current creates transient voltage drops due to the internal resistance of the battery, the 'reversed battery' protection diode, the switch and circuit wiring. These momentary voltage drops generate significant noise, degrading power supply regulation and battery life, Fig. 2a).

These current peaks can be min-

Fig. 1. Switched mode power supply used in the $\tan\delta$ meter. For clarity the shutdown timer and low-battery indicator have been removed. In this article, circuit performance is studied using three different capacitor combinations for C_1 , C_2 and C_3 .



imised by using an input storage capacitor, C_1 , adjacent to the switched-mode power supply integrated circuit. This reduces and smooths out the peaks of current demand, the battery then supplying a more steady current into this capacitor.

Other capacitors are needed to act as reservoir, C_2 , and perform smoothing, C_3 , following the power supply switching rectifier and the linear regulator.

Values and voltages required for these capacitors depend on your chosen power supply integrated circuit. Most data sheets recommend makers' part numbers, as well as capacitance values and voltage rating. Unfortunately many of these specific devices are not easily obtained in the UK – especially in the smaller quantities needed for prototype development.

Effective capacitors

For a +5V, 100mA output supply using 4 AA batteries, the LT1303 data recommended capacitor types and values which were not immediately available. They suggested 33 μ F for C_1 and C_2 with 220 μ F for C_3 .

For many years my stock capacitors have been chosen from the Philips 037 sub-miniature general purpose type and the company's 135 low-impedance, high-ripple ranges³. The small capacitors needed for this power supply are not available in the 135 style.

A worked example

Using 47 μ F, 50V and 220 μ F, 10V Philips 037 radial-lead aluminium electrolytics, I assembled a prototype power supply. Powered from four AA batteries and using a 100 Ω resistive dummy load, this prototype was noisy and inefficient.

I redesigned the PCB to decouple the noise input to a capacitor from its smoothed output, using four-terminal

track routings¹. This revised PCB, together with its 100 Ω load, was used to provide the oscillograms and measured results used in this article. To avoid variations in battery performance, the circuit was powered from my bench supply, set to 4.5 volt.

I chose this voltage for two reasons. First it is the median usable voltage from four AA alkaline batteries. Combined with the small voltage dropped in the 'reversed battery' protection diode, it ensures the power supply works only in its 'boost' mode.

Using two 47 μ F, 50V 037 capacitors for C_1 and C_2 , with a 220 μ F, 10V 037 for C_3 , the supply drew 86mA. Output was 5.1V, giving 51mA into the 100 Ω load. The circuit was noisy with excessive output ripple voltages, Fig. 2.

In this series of plots, to ensure the very fast transient spikes can be seen, the Y amplifier settings for C_1 , C_2 and C_3 differ, but were kept consistent with change of capacitor, by capacitor reference. I used 200mV/cm for C_1 , 100mV/cm for C_2 and 50mV/cm for C_3 . In each case my 250MHz oscilloscope probes were switched to divide by 10. Switched to unity, most of this fast transient detail was missing.

The LT1303 IC features 'burst-mode' operation, adapting its switching speeds to suit circuit conditions. For each photo, the X time base was set to 5 μ s/cm.

To obtain a stable trace for a photograph, the actual sweep rate was slowed using the vernier control. Hence the trace speed was not controlled. The important point here is the change in peak transient and ripple voltages with change of capacitor. The Y scaling is consistent for each capacitor number.

I then replaced both 47 μ F 50V capacitors with two more 220 μ F, 10V

037 types. The 47 μ F 50V and the 220 μ F 10V capacitor case sizes were 6.3mm by 11mm.

Efficiency was slightly improved; the supply current measured 80.1mA. The LT1303 chip has changed mode, now producing three transient bursts per oscilloscope cycle. Using the same 'Y' oscilloscope settings for each capacitor reference as for Fig. 2, the transient

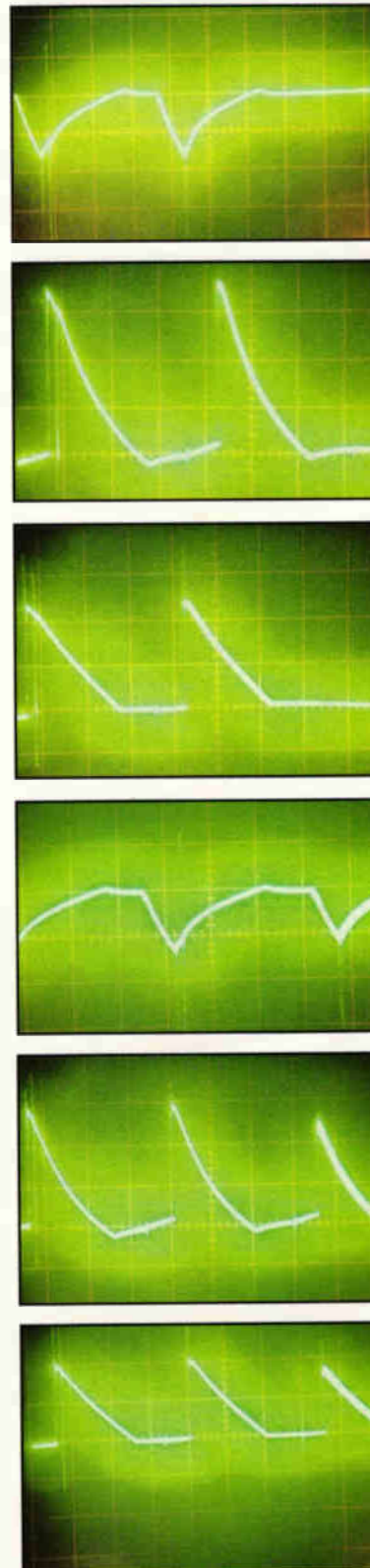


Fig. 2. Ripple waveforms together with superimposed fast transient voltage spikes, measured using Philips 037 capacitors. Capacitors C_1 and C_2 were 47 μ F 50V with 220 μ F 10V for C_3 .

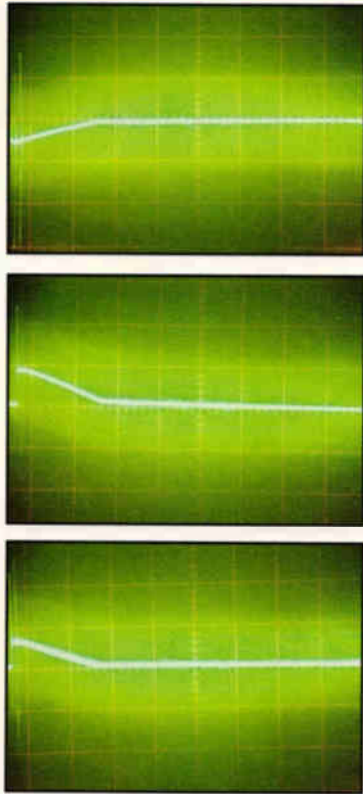
- a) C_1 ripple voltage is 0.32V pk-pk, transient spikes are 1.14V pk-pk.
 b) C_2 Ripple voltage is 0.36V pk-pk, transient spikes are 0.38V pk-pk.
 c) C_3 Ripple voltage is 0.12V pk-pk, transient spikes are 0.25V pk-pk.

Fig. 3. With conditions set as for Fig. 2, these photos show the marginal improvement gained by substituting 220 μ F, 10V 037 capacitors for C_1 and C_2 . Nominal capacitance is much increased, but ripple and transient spikes improve little.
 a) C_1 : Ripple voltage 0.26V pk-pk, transient spikes 1.02V pk-pk.
 b) C_2 : Ripple voltage 0.3V pk-pk, transient spikes 0.39V pk-pk.
 c) C_3 : Ripple voltage 0.1V pk-pk, transient spikes 0.235V pk-pk.

Table 1. Measured parameter values for one sample of each capacitor style only.

Unit/parameter	10kHz	30kHz	100kHz	300kHz	1MHz
47μF/50V					
Philips 037					
IZI (Ω)	1.151	1.05	0.993	0.938	0.878
ESR (Ω)	1.106	1.043	0.94	0.93	0.87
Capacitance (μ F)	58.3	42.6	22.6	7.87	1.99
220μF/10V					
Philips 037					
IZI (Ω)	0.824	0.785	0.738	0.706	0.671
ESR (Ω)	0.818	0.783	0.74	0.71	0.67
Capacitance (μ F)	129	68.4	27.1	9.1	2.0
220μF/10V					
Rubycon YXF					
IZI (Ω)	0.404	0.372	0.341	0.322	0.308
ESR (Ω)	0.394	0.372	0.34	0.32	0.32
Capacitance (μ F)	153	94	44	20.7	9.2

Fig. 4. Using three Rubycon YXF 220µF, 10V capacitors, low impedance types with the same physical size as the 037. Although nominal capacitances are unchanged from Fig. 2, the ripple voltages are much smaller.



- a) C₁ Ripple voltage is 0.12V pk-pk, transient spikes are 0.98V pk-pk.
- b) C₂ Ripple voltage is 0.09V pk-pk, transient spikes are 0.36V pk-pk.
- c) C₃ Ripple voltage is 0.035V pk-pk, transient spikes are 0.21V pk-pk.

spikes and ripple voltages were little changed. With a nearly fivefold increase in nominal capacitance for C₁ and C₂, why had these waveforms not improved more? See Fig. 3.

I removed these capacitors from the printed board. After allowing sufficient time for them to cool to room temperature, I measured impedance at 100kHz. The 47µF capacitor measured 0.993Ω, the 220µF slightly less at 0.738Ω.

Board area and height above board being limited, the capacitor case size

could not be increased to fit the Sanyo 'Oscon' type 220µF 10V size of 10mm by 11mm⁴. The Rubycon 220µF 10V YXF style case size was acceptable, and it claimed reduced impedance⁵.

At 100kHz I measured 0.341Ω impedance, a notable reduction, so three of these types were fitted. My measurements were repeated, using exactly the same 'Y' settings as before, Fig. 4.

Ripple voltage is substantially reduced, but the transient spikes remain almost unchanged. Using the same PCB, components and set up, except for these three capacitors, current drawn from the supply had reduced to 75.6mA, improving efficiency.

To reduce these very fast transients, I inserted a 5µH inductor L₂ and an additional 220µF, 10V YXF capacitor C₄, as shown in the schematic. These reduced the transients and the ripple voltage to acceptable levels.

The protection diode was removed from circuit and a 5V supply applied. Now the current measured 66.5mA – representing an efficiency of 78%. I could now complete the design of the power supply for my tand meter.

My curiosity had been aroused, so when time permitted, I determined to retrace these steps in order to quantify why these differences occurred and write this article.

Capacitor parameters

One extremely useful tool I used during the initial design phase of this power supply, was the Spice 2G based 'Micropower SwitcherCAD' v2 software, available from Linear Technology². This models the integrated circuit and provides a selection

of capacitors from its database.

Other capacitors can be used provided you input values for ESR and capacitance. I suspect the ESR values used relate to a frequency of 100kHz and the capacitance is the catalogue value. I have not been able to confirm this from the help files or user manual, Fig. 5.

This raises the question of whether, in practice, ESR at 100kHz is the only relevant parameter? I believed other characteristics were equally important, but a study of capacitor makers' data did not help.

While some makers provided full data, it was not possible to discover how ESR at 100kHz had been measured. It was not always clear whether quoted values were typical or maximum values. Some capacitor makers provided no high frequency data at all, apart from a few selected impedance graphs. There's more on this in the panel entitled, 'Modelling Capacitors'.

I decided to measure the 037 and YXF capacitors using only general-purpose laboratory equipment and methods. While these were not precision measurements, they gave an adequate comparison of ESR and capacitance/inductance and can be made without investing in extremely expensive, high frequency precision LCR meters, such as the Agilent Technologies HP4284.

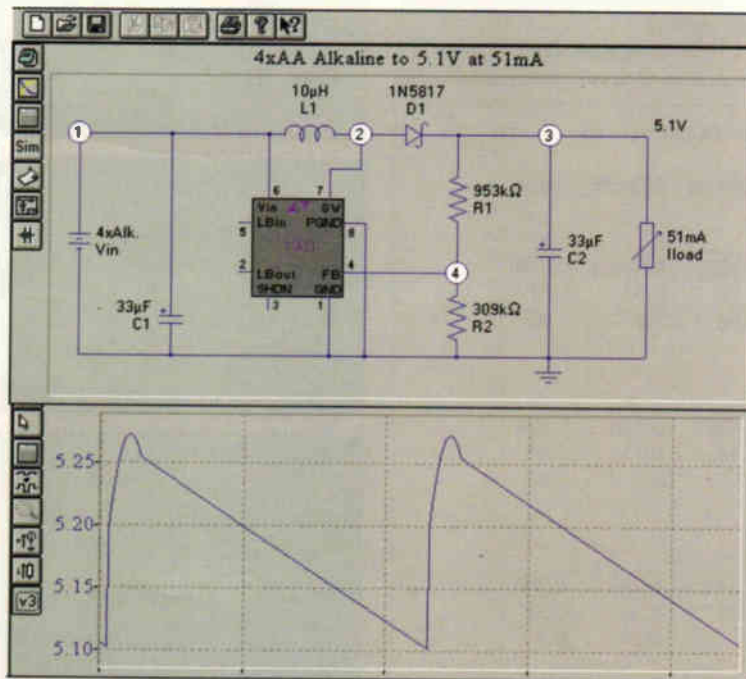
Using basic volt/amp impedance measurement techniques, I measured samples of the three capacitors for impedance by frequency from 10kHz to 1MHz⁶. Then from phase measurements, I calculated ESR and capacitance at these frequencies.

The accuracy of my impedance measurements was confirmed simply by measurements of 1Ω and 0.047Ω metal-film resistors. The true DC values of these had been confirmed by measurement of voltage drop, while passing a direct current of 100mA.

Values given in the tables for ESR and capacitance however depend totally on the accuracy of my phase measurements. Using my phase meter⁶ with a 4.5-digit DMM for increased resolution of 0.01°, its accuracy for the angles measured is good up to 100kHz. As the capacitor approaches self-resonance and phase angles less than 1° must be measured, accuracy reduces.

Philips' data for its 037 style is quite clear, specifying a maximum |Z| at 10kHz for both capacitance values, at 2.0Ω The Rubycon YXF data available when I was developing the tand power supply simply claimed 0.4Ω at 100kHz for the 220µF, 10V capacitor. Current data from Rubycon clearly states that maximum impedance at 20°C is 0.4Ω, increasing to 1.6 Ω at -10°C. The above

Fig. 5. Using Micropower SwitcherCAD software from Linear technology. Simply enter your chosen capacitor into a dialogue box. The schematic is automatically drawn then simulated. The lower trace shows that the voltage simulated at C₂ approximates that measured, but the fast transients are missing.



measurements confirm both claims.

From these calculated, apparent capacitance values the similarity in my circuit's ripple and transient voltage performance between the 47 μ F, 50V and the 220 μ F, 10V 037 types is explained. At 100kHz and higher, both capacitance values are similar. The 220 μ F, 10V 037 capacitors I used offered 25% less impedance but little extra capacitance.

The measured capacitance for the Rubycon YXF style, reduces less with frequency. Across this frequency band its ESR is much smaller.

Ignoring for now the very fast transient spikes, this increased capacitance coupled with its reduced ESR has more than halved the ripple voltage compared to that measured using the 220 μ F, 10V 037 capacitors.

The fast transient voltage spikes, less than 50ns duration, result from capacitor and circuit-board inductances. To evaluate whether this inductance is contributed by the 220 μ F capacitors or by the circuit board, a different measurement is needed. I mounted both types as shunt loads to ground in test jigs. They were inserted into a correctly terminated 50 Ω system.

Application of a fast rise-time pulse generator revealed these two capacitors exhibit dramatically different behaviour when subject to a fast rise time pulse. For Fig. 6a), both capacitors were mounted in individual, but otherwise

Modelling capacitors

The most problematic part in a switched mode power supply simulation is the capacitor model used. Because a switched-mode power supply operates essentially in the time domain, most designers will use a time-domain simulator, usually derived from the public-domain Spice series.

As you have seen, capacitor parameters are essentially frequency dependent. As a result, the simplistic ideal capacitor model provided in Spice simulators cannot

adequately represent a practical electrolytic capacitor.

Certain improved capacitor models have been proposed. At 'CARTS 95'. John Prymak of Kemet proposed one method which he applied to a number of tantalum capacitors⁸. His approach was reported by Intusoft in the company's Wescon/95 Preview, and its November 95/January 96 newsletters.

Subsequently a selection of improved capacitor models were made freely available from Intusoft's site⁹. I am not aware that any have been incorporated into the model

libraries supplied with other simulators.

A further difficulty is that these models use the variable 'Freq' for frequency to calculate frequency dependent parameters. Frequency of course is not a time domain parameter, so these improved models should not be used. They can only apply to frequency-domain simulation runs.

If such models are used for a time-domain or 'transient' simulation, because the 'Freq' variable then defaults to zero, then the model assumes an unduly large value for ESR.

identical, 50 Ω impedance printed circuit test jigs. These jigs were each connected to my pulse generator via a 10dB isolating attenuator, to minimise line reflections that might affect the generator output.

Both oscilloscope 'Y' channels were carefully adjusted for equality, prior to inserting the test jigs and capacitors into the system. The jig-mounted capacitors attenuate the pulse generator signal, as measured by the oscilloscope.

The expected change in pulse height can be seen. But while the Rubycon capacitor provides better attenuation, it

also contributes positive and negative going inductive overshoot at the pulse edges. Inductive overshoot is not visible using the Philips 037 capacitor.

This explains why the very fast power supply transient spikes are so similar for both makes of capacitor. The Rubycon YXF has not attenuated the fast spikes significantly more than has the Philips 037, even though its has more capacitance and less ESR at 100kHz. Fig. 6a).

Using exactly the same test set up, I then applied a 100kHz sinewave to both capacitors, which visually confirmed their relative impedances as

A capacitor's impedance

The quality of many high-frequency components, such as RF inductors and very low-loss capacitors, is often defined by their 'Q' factor. This is the result of dividing a component's measured AC reactance by its AC resistive losses.

The reciprocal of 'Q' is $\tan\delta$, which is defined as the capacitor's AC resistive losses (ESR) divided by its capacitive reactance, X_C , at that frequency. $\tan\delta$ is used to describe the quality of almost all general purpose capacitors.

$$\tan\delta = \frac{ESR}{X_C} \text{ where } X_C = \frac{1}{2\pi FC}$$

Or alternatively,

$$ESR = X_C \times \tan\delta$$

and,

$$\tan\delta = ESR \times 2\pi FC$$

$$|Z| = ESR + jX_C = \sqrt{ESR^2 + X_C^2}$$

As you may have noticed, $\tan\delta$ has no upper limit; it can, and frequently does, exceed unity. Particularly at high frequency, ESR can greatly exceed the capacitor's reactance X_C . At frequencies above or below the capacitor's self-resonance frequency, ESR

must always be smaller than the measured impedance $|Z|$.

As frequency increases, the above two-element model becomes invalid. The capacitor must then be viewed as a three-element device. In principle, it is possible to extract these three components, even at low frequency. In practice, unless the value of this inductive element is known, measurements several octaves above and below the capacitor's self resonant frequency are needed.

$$|Z| = \sqrt{ESR^2 + (X_C - X_L)^2}$$

Here, X_C is capacitive reactance and X_L the inductive reactance at the measured frequency¹⁰.

At the self-resonant frequency of the capacitor, X_C and X_L have equal and opposite reactances and cancel each other. Measured impedance $|Z|$ then equals the capacitor's ESR, but only at that particular frequency. Phase angle then measures zero.

An LCR meter calculates capacitance and inductance using the two-element model from its measured values for magnitude of $|Z|$ and phase angle. At frequencies just below self resonance, the inductive reactance decreases the capacitor's measured

reactance, and thus increases its apparent capacitance, as displayed on an LCR meter.

At frequencies just above self resonance, the capacitive reactance decreases the capacitor's measured reactance, reducing the apparent self inductance, measured on an LCR meter.

These two effects result in the sharp resonance null observed when measuring ceramic and film capacitors. Usually, with aluminium and tantalum electrolytics, their ESR is substantially greater than these reactances, so a null cannot be observed.

With aluminium-electrolytic capacitors, the impedance curve can appear flat bottomed over a wide frequency band. In this case the resonant frequency can only

be determined by measuring the capacitor's phase angle. This zero-phase frequency need not coincide exactly with the frequency of minimum impedance.

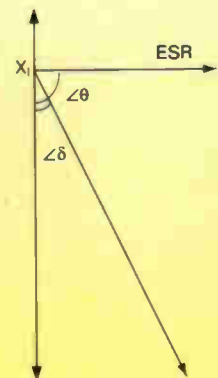
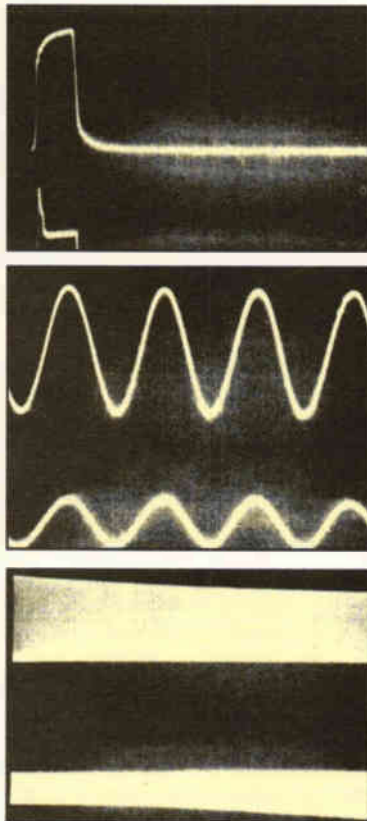


Fig. 6. Comparison measurements of the 220µF, 10V Philips 037, top trace, versus the 220µF, 10V Rubycon YXF, lower trace, subjected to identical waveforms. Both measurement channels were pre-calibrated to ensure equal responses with no capacitor.



a) Stressed with a fast pulse, the 037 capacitor exhibits less capacitance and higher impedance, while the YXF style shows inductive ringing. 'X' time base was 0.2µs/cm.
 b) Same set-up as 'A' except using 100kHz sinewave generator and slower time base to show the relative impedances of the two capacitors.
 c) Relative impedances of these capacitors, with sine wave sweep from 1MHz to 10MHz. The 037 shows no visible increase, but the YXF capacitor shows self inductance increasing impedance above 2MHz.

measured for the table, Fig. 6b). Both capacitors were the same size and mounted on identical printed-board test jigs. Both had their can decks pressed hard against the circuit board, ensuring identical lead lengths. I expected they would exhibit similar self inductances. To investigate whether their inductances differed, using the 'shunt' jigs, I applied a sinusoidal frequency sweep from 1MHz to 10MHz. The Rubycon YXF style is becoming inductive around 2MHz. The Philips 037 capacitor impedance continues to fall, remaining capacitive to 10MHz. The Philips capacitor has less self inductance of the two types, Fig. 6c).

You can see from this that knowledge only of ESR or impedance at 100kHz is simply not sufficient. Capacitance and ESR by frequency and a value for self inductance are all needed when choosing capacitors for use in switched-mode power supplies. These parameters can be obtained from simple measurements. Have a look at the panel entitled 'A capacitor's impedance'.

ESR of aluminium electrolytics is also strongly dependent on temperature. Increase in temperature reduces ESR so 'improving' the capacitor. However, reduction in ESR can allow the inductive effects to become more dominant.

Should your application be required to work above or below room temperature, it is advisable you also perform capacitor comparison measurements at your highest and lowest working temperatures.

In summary

These simple, easily-performed tests explain why my original tanδ power supply behaved as reported when its capacitors were changed. Details of these changes provide a methodology easily adapted for other designs.

Many other capacitor makes and types are available, some of which may be more suitable than the ones I used. Repetition of these tests, using capacitor values and voltages appropriate to your design, will assist your final selection.

The optimum choice of capacitor is essential when designing an efficient power supply. Equipped with these test methods, a designer can use test frequencies and temperatures suited to the end application, rather than basing choice only on 100kHz, room temperature, values.

Apart from my test jigs, I used no specialist or expensive measuring

equipment. The above tests are easily replicated using only conventional laboratory test instruments with suitable test jigs and methods.

In my next article I will describe the test jigs, together with test methods and calculation formulae used for this article. Those of you interested can then measure your own capacitor choices.

I will introduce calculation methods that can be used to translate measurements of impedance and phase, into the three-element capacitor model – inductance, ESR and capacitance in series – needed for accurate simulation.

In a final article I will describe a more advanced impedance measurement method that gives a direct read-out of impedance. There will also be a circuit diagram for a dedicated, direct-reading impedance meter with test jig, usable from 10kHz to 10MHz. ■

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Batteries

Batteries are non-linear devices having significant internal resistance. For a given type, their capacity varies some 10% to

15% from cell to cell. This internal resistance increases with load current, battery ageing and reduced ambient temperature. Consequently precise modelling to gauge battery life is not practical.

One approach to simulating common battery types, devised by Steve Hageman⁷, uses PSpice analogue behavioural modelling techniques. Steve used a large capacitor to represent the ampere-hour capacity of the cell, with look-up tables of cell-voltage versus charge state. A discharge rate normaliser represents capacity loss at higher discharge rates and a variable cell resistor to characterise the internal resistance of the battery. A discharge circuit completes his

simulation model.

The pulsing current demand from switched-mode power supplies complicates battery modelling. Hageman suggests using the RMS average of the pulsed current to avoid convergence problems and excessive simulation run times.

A similar approach is used in the Micropower SwitcherCAD software available from Linear Technology². This uses a simplified battery model and simulates a constant power drain from the battery. This power level is calculated from the results of a burst mode simulation cycle for the switching IC used.

The simulator then outputs a graph of cell voltage versus discharge hours to approximate battery life, Fig. A.



Fig. A. Typical plot of battery life, using the Micropower SwitcherCAD software, shows battery voltage by time with a 292mW load.

Beginners' corner

Ian Hickman has produced a number of projects for members of the University of

Portsmouth RF Club to help students build, troubleshoot and test circuitry, in connection with the RFEE Initiative, described in last month's issue. The circuits are graded in complexity - the earlier ones not even directly concerned with radio frequencies. But all are instructive. Here, Ian introduces the very simplest, suitable for those with little, if any, prior experience of constructing and trouble-shooting hardware.

In earlier generations - such as my own - students of light-current electrical engineering would usually have had some exposure to practical electronics, prior to embarking upon their degree course. Conversations with fellow students and, later, with colleagues, showed that my own route into electronics was typical of many in that era.

Starting with a fascination, as a lad, with clocks and all things mechanical, I progressed through Meccano to messing about with torch bulbs, old No 8 batteries and scruffy bits of double-cotton covered wire.

Eventually, I rescued a scrapped alarm clock, and repaired it. With the aid of a relay and other bits and pieces, I turned into a potentially lethal time-switch, to turn on my bedside light in the morning.

Later I progressed through crystal sets and then 2V battery valves until, in the 6th form, I constructed a home-brew three waveband mains superhet. This used components that I bought from the various government surplus radio shops to be found in London's Lisle Street in the early fifties, all later replaced by sex shops* and later still incorporated into Soho's Chinatown.

By contrast, it appeared that few, if any, of the present-day students for whom I produced these little exercises in electronics build-and-test had had any previous experience in making up

circuits, and getting them working.

So it was inevitable that they would metaphorically stub their toes on a hundred minor problems, and learn a great deal in the process. And that is the whole point of this little one-IC project, and others to be described later.

The IC

The circuit uses that ubiquitous and versatile integrated circuit, the 555 timer. Produced originally in bipolar technology many years ago, it is now available from many manufacturers also as a current-frugal CMOS device.

Both versions are packaged in an 8-pin housing, while a 'dual 555', namely the 556, is available in a 14 pin package. Figure 1 shows the internal workings of the beast, which include three equal resistors connected across the supply pins, two comparators, and a flip-flop with reset input.

The comparators, connected to the resistive potential divider, control the bistable device, which is also known as a flip-flop. This in turn controls two output stages, one low impedance, the other an open collector n-p-n transistor.

How the circuit works

Figure 2 shows the 555 timer IC connected as an astable multivibrator, or oscillator. It can also be connected as a monostable multivibrator, as a frequency divider, or as any of various types of modulator¹.

Operation is as follows. Capacitor C_2 charges up towards the +15V supply voltage V_{cc} , via R_2 and R_3 . When the voltage at pin 6 reaches the threshold voltage at the other input of the associated comparator, the flip-flop is reset. This sets the low impedance output at pin 3 'low', i.e. connects it to 0V ground instead of V_{cc} , extinguishing the light-emitting diode. It also turns on the open collector 'discharge' transistor at pin 7, clamping the junction of R_2 and R_3 to 0V ground.

Capacitor C_2 now discharges towards 0V via R_3 , until the voltage at the 'trigger' input, pin 2, reaches the threshold at the other input of its associated comparator. This then sets the flip-flop, turning off the discharge

transistor at pin 8, and setting pin 3 high.

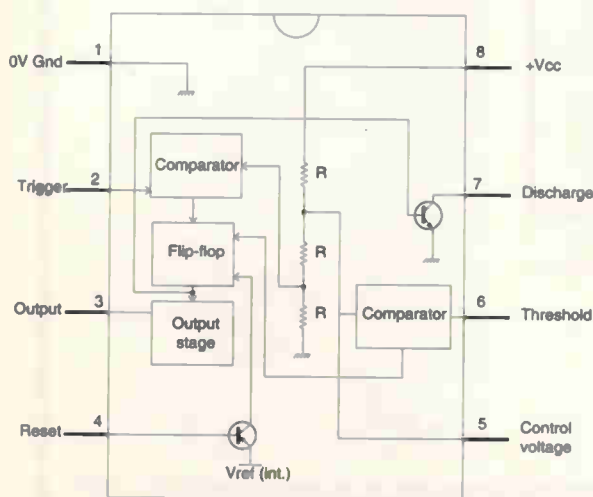
At this point, the LED lights again, and the cycle repeats, with C_2 charging up towards V_{cc} once more. As the three resistors, R , forming the internal potential divider chain are all equal, the voltage at pin 6 cycles up and down between 1/3 and 2/3 of V_{cc} .

Build it

The circuit can be built up in various ways. A scrap of 0.1 inch matrix copper strip-board, cut from RS stock number 433-595 or 433-602 can be used; the same material, in different-sized sheets, is available from all the usual electronic components catalogues.

Alternatively, a printed-circuit board layout can be produced. Nowadays, this is always done on a personal com-

Fig. 1. Showing the internal arrangement of the 555 timer IC.



* I wonder how he knows? Ed

puter, IBM compatible or MAC, rather than the old way with a light-box, film and black tape. But producing a pcb is time consuming, and assumes the university's or college's pcb facility is available for use.

The students I was working with met after hours, in their own free time, so the pcb facility not an option. Instead they made the circuit up on 0.1in matrix plug-type prototyping board, after the style of RS 488-618 or 488-933.

With a circuit operating at low frequencies such as this, layout is unimportant. Component leads were simply bent as required, to plug in wherever convenient. This produced some weird and wonderful layouts, but they all worked, eventually, the LED winking away encouragingly. It also meant that the circuits could be disassembled at the end of the evening, the leads straightened out, and the components returned to the appropriate compartments in a set of drawers. This had been purchased, like the prototype boards and indeed the components themselves, especially for use by the university's RF Club, with funds provided by the RFEEL.

Try to get it working...

The students worked either alone or in small groups, whichever they preferred. Predictably enough, quite a few of the circuits did not work first time. In a few cases there were accidental misconnections, which could have destroyed the IC or the light emitting diode, had not the advice of a seasoned circuit developer been followed. I showed the club members how to ensure that the first-time power-up was safely achieved by using a power supply with variable current limit.

The output voltage of the power supply unit was first set to +15V with the prototype circuit disconnected. The variable current-limit control was then set to minimum, fully anticlockwise, causing the output voltage to collapse to zero.

Next, the circuit was connected and the power supply's output switched on, advancing the current limit control cautiously while keeping a close eye on the current meter. If the current did not exceed the expected ten milliamps or so as the voltage rose to the preset +15V level, then all was well. Even if the circuit was not actually working, at least it was safe to leave on while trouble-shooting.

If on the other hand, the current increased alarmingly while the supply voltage was still only a volt or so, it was time to switch off and recheck the circuit connections. Of the circuits that

did not work first time, the causes were many and various, but all mysterious to the uninitiated.

...and if it doesn't

One group checked and rechecked their circuit, but it still refused to work. Asked to assist, I naturally checked that the power was actually connected.

Yes, the PSU was on, and red and black wires ran to the appropriate coloured terminals at the end of the prototype plug-board. But my pocket DVM registered nothing between pins 1 and 8 of the IC. This picked up its +15V supply from one of the rows of holes along the edge of the plug-board, with a red line alongside them. So it was apparent that the red terminal was not meant for a lead from the PSU, but for a wire to pick up on one of the red-strip holes! The lead from the PSU was meant to have a 4mm banana plug, to engage with the central socket of the isolated red terminal post.

Another group, finding the circuit did not work, concluded that the IC must be faulty, and changed it – to no avail. Of course, 999 times out of a thousand, if a circuit does not work, it is not the fault of a component; it is much more likely that of the circuit developer.

Fortunately, most modern semiconductors prove very hardy indeed, and survive accidental abuse beyond what one would imagine possible.

Features of the 555

Having got their circuits functioning, the club members were keen to investigate the workings of the circuit. Oscilloscopes were to hand, and so the various waveforms could be investigated.

Pin 3 of course showed a 0 to +15V squarewave – except that it was not 'square'. The reason appeared on viewing the 'sawtooth' waveform at pin 6, for which purpose, C_2 was reduced in value to 100nF. This permitted a faster, flicker-free 'scope trace to be used, and showed the two segments of the exponential charge/discharge cycle, as in Fig. 3, lower trace, and the output, upper trace.

Capacitor C_2 charges from 1/3 to 2/3 of V_{cc} via R_2+R_3 , and discharges in the opposite direction, between the same two voltage levels via R_3 alone. Thus these two resistors set the on:off duty cycle. Whatever their value, this can never be 1:1. If they are equal, as in Fig. 2, then the on:off ratio at the output is 2:1.

The on time is given by $t_{on}=0.693(R_2+R_3)C_2$ and the off time by $t_{off}=0.693(R_3)C_2$. So the frequency f in hertz is given by $f=1/T$, where $T=(t_{on}+t_{off})=0.693(R_2+2R_3)C_2$.

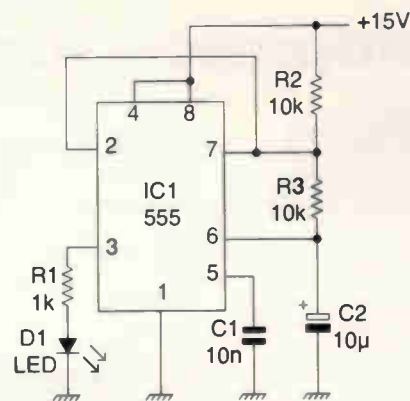


Fig. 2. Astable LED flasher circuit.

$$\text{Hence } f=1.44/(R_2+2R_3)C_2.$$

If the supply voltage V_{cc} changes, say increases, the charge/discharge currents increase and the voltage across C_2 changes more rapidly. But as the two threshold voltages produced by the IC's internal chain of three resistors change by exactly the same percentage, the frequency is completely independent of supply voltage.

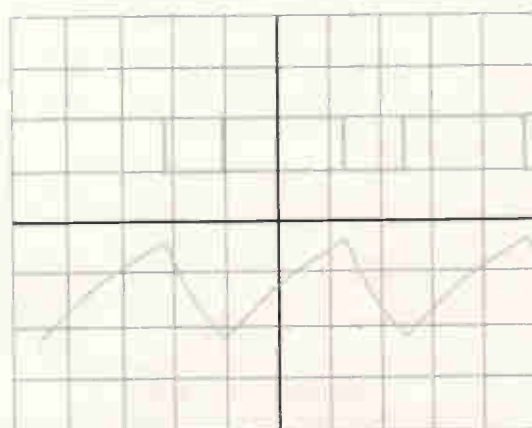
A negative-going edge at the reset input, pin 4, resets the internal flip-flop. As this function is not required in the astable mode, pin 4 is tied to pin 8 and V_{cc} . In monostable mode, it may be used to terminate the output pulse early, if so required.

Pin 5 connects to the junction of the upper two resistors of the IC's internal reference chain, and is shown decoupled to ground in Fig. 2 via a 10nF capacitor. In the present application, this is not essential. But if the circuit were part of a larger system, sharing its V_{cc} with other devices, in the absence of C_1 , noise or ripple on V_{cc} could affect the timing accuracy, causing jitter on the output waveform. ■

Reference

1. See for example the application data in the LM555/LM555C Timer details in the National Applications Specific Analog Products Databook, from National Semiconductor.

Fig. 3. Upper trace, 555 output at pin3; lower trace, waveform at pin 6.



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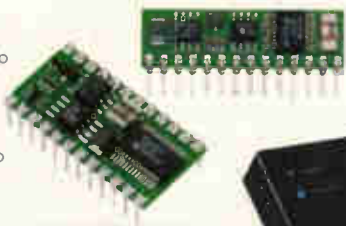
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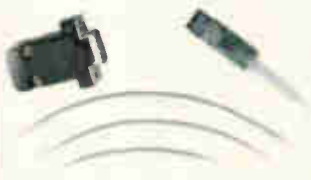
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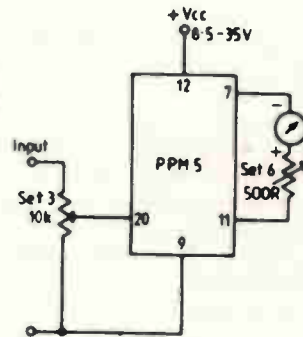
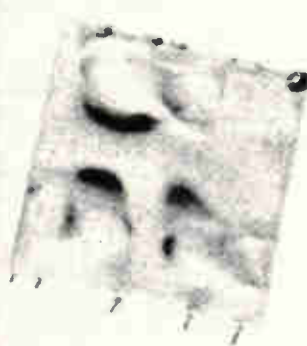
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Waferscale Integration
Tel: 001 510 656 5400
Enq No 501

DIMM sockets

Molex is now manufacturing double data-rate DIMM sockets for high-bandwidth memory systems in workstations, servers and desktop computers. Features include dual latches and a positive latch that produces an audible click when secured in the open and closed positions. A module prealignment feature reduces misalignment and

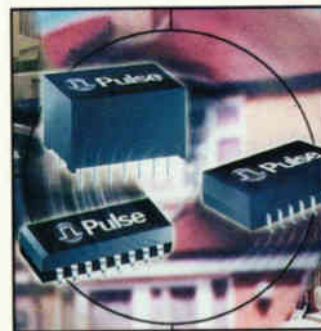


terminal stubbing. The 184-circuit version has the same profile as its 168-circuit DIMM socket, and the 200-circuit unit has the same profile as the 200-circuit standard.

Molex
Tel: 01420 488488
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Filter modules

Pulse has launched three filter modules that meet the Home Phone Networking Alliance 1Mbit/s specifications and are UL1950 certified for supplementary insulation requirements. They provide the



bandpass filtering for home phone-line networking products such as network-interface cards, modem add-on cards and set-top boxes. They also stop the home networking signal from interfering with telephone or Internet access. The B6001 is a 14-pin, through-hole device incorporating 10baseT filtering and isolation. The B6003 and B6005 are 16-pin surface mount parts. The B6005 also has 10baseT filtering and supports the 79C978 chipset from AMD. The bandpass filter portion ensures the 6 to 9MHz operation for home networking products has a nominal insertion loss of 3dB. These parts offer a return loss of 12dB within the home networking portion and 18dB for 10baseT. They include protection circuitry, isolation and EMI filtering.

Pulse
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Enq No 504

Power metering

IFR has added statistical analysis functions to its Gigatronics 8650A universal power meters for designing and testing wireless communications systems. They include histograms, cumulative distribution function, complementary cumulative distribution function and strip-chart capabilities. The meters are available in single and dual channel configurations and can measure the peak and average power of TDMA, GSM and CDMA signals. They can also measure CW and pulse modulated signals with NIST traceable accuracy from 10MHz to 40GHz between -70 and +47dBm.

IFR
Tel: 01483 772172
Enq No 505

DSL network processor

Sequoia is making available a network processor platform combining silicon with networking software. The integrated product from Basis Communications includes a BC6911 DSL network processor chip, with service protocol stacks, real-time operating system and open API for developers to add IP-based applications. The platform is for customer equipment such as DSL modems, routers and bridges for Internet access. One chip comprises ATM segmentation and reassembly processor, ARM7TDMI Risc core with 4kbyte of local SRAM, 8kbyte of internal SRAM, PCI interface, Utopia interface and peripheral block. As well as performing ATM adaption layer-five segmentation and reassembly, the ARM CPU controls the end device, supports network signalling and provides local management. It uses Wind River's Vxworks and runs TCP/IP stacks and other IP based applications for bridges and routers. HTTP, DHCP, NAT, PAT, SNMP and



Microcontroller starter kit

NEC starter kits are available for evaluating applications on the firm's V850/SA1 microcontroller in real time. The Startware kits offer in-device emulation and operate at the maximum performance of the microcontroller. They will run programs from a PC running Greenhills Multi embedded development kit. Programs will run without wait states if they are executed from internal flash memory or external SRAM and require one wait state if external flash memory is used. The kits contain elements to evaluate I/O-functions such as DIP switches and a seven-segment-display. The 256kbyte internal flash of the microcontroller is available so a kit can be used as a stand-alone device or as a ready-to-use system in a user's application. Half the flash memory runs hardware test functions and a debug monitor. The debugging mode uses the monitor program in the internal flash memory of the microcontroller. There is also a stand-alone mode when a user program can be in the internal or the external flash memory. User application programs can be downloaded into 512kbyte external flash memory or SRAM. Chip select logic lets RAM and flash swap their memory locations.

NEC
Tel: 01909 691133
Enq No 502



Please quote *Electronics World* when seeking further information

MIB can be added to the Vxworks RTOS. Local SRAM is used to execute timing-critical driver code.

Saquoia
Tel: 01189 769000
Enq No 506

Multichip module

From GD Technik is White Electronic Designs' WEDC3C750A multichip module combining a 32-bit PowerPC 750 Risc processor with 1Mbyte SSRAM L2 cache. The 200MHz version of the PowerPC is embedded into the module but Arthur (200 to 300MHz) and Conan (300 to 400MHz) upgrades are available. User-programmable power-saving modes are incorporated including doze, nap and sleep states. The two 128k x 36 synchronous pipeline SRAMs have a maximum cache frequency of 100MHz. Internal bus frequency is 66MHz. The processor and two SSRAMs are flip-chip attached to a 255 CBGA or optional CCGA. The module measures 21 by 25mm. I/O count is 255. Embedded applications include power and fire control, navigation, guidance and aerospace systems.

GD Technik
Tel: 001189 342277
Enq No 507

Linux DSP board

Ixthos has introduced a CompactPCI DSP board for telecoms voice processing. Features include more than 26000Mips of DSP, 81.3 Specint95 of Risc and the firm's Champ common heterogeneous architecture for multi-processing. The board is for processing high voice and data throughput channels from DTMF and vocoder to voice recognition and generation. Running Linux, the board can process more than 1000 broadband integrated voice and data channels in a 6U CompactPCI slot. It uses the Motorola PowerPC AltiVec Risc microprocessor and the SMP real-time extensions from Linux. The board supports two PMC expansion modules. These expansion sites can be used for T1 and E1, ATM, frame relay, OC3 and OC48 Sonet, 10 and 100baseT and Gigabit Ethernet applications.

Ixthos
Tel: 001 703 779 7800
Enq No 508

8-bit micro

Philips has announced an 80C51-based 51LPC microcontroller with integrated four-channel, multiplexed 8-bit a-to-d converter on-chip. Applications include battery chargers and sensors. The 87LPC767 has



onboard programmability for last-minute adaptation to the application. Features include brownout detection function, power-on reset and watchdog timer. The device also includes an on-chip oscillator to provide five user-programmable modes that let the user tune the performance and power consumption. When using the oscillator, two additional pins can be released for use as I/Os, increasing the I/O pin count from 16 to 18.

Thurlby Thandar
Tel: 01480 412451
Enq No 510

ADPCM processor

For speech compression in wireless PBX and digital cordless telephones, the AT1008 from Steadlands is an eight-channel (octal) full-duplex ADPCM processor for cordless PBX



and T1 and E1 switching. The chip is compatible with ITU-T G.726 at 40, 32, 24 and 16kbit/s and can operate on 16 channels of PCM to ADPCM compression, 16 channels of ADPCM to PCM decompression or eight channels of full-duplex operation in an 8kHz frame. Available in 28-pin DIP or skinny packages, it can be configured for setting input or output for PCM clock, PCM frame sync, ADPCM clock and ADPCM frame sync for system use by a three-wire serial port. It can be programmed with algorithms, data rate and time slot assignments for individual channels on-the-fly.

Steadlands
Tel: 01670 361261
Enq No 512

Dual motor-driver IC

Allegro Microsystems is supplying a motor-driver IC, the A3976, which is designed to drive both windings of a bipolar stepper motor or bidirectionally control two DC motors. The device features two H-bridges, both of which are capable of continuous output currents of up to ± 500 mA and operating voltages to 30V. Free-wheeling, substrate-isolated diodes are included to provide output transient suppression when motors or other inductive loads are being switched. For each bridge, a phase input controls load current polarity by selecting the appropriate source and sink driver pair. The enable input, when held high, activates the respective output

H-bridge. When both enable pins are held low, the device will enter sleep mode, when it consumes less than 100 μ A.

Allegro Microsystems
Tel: 01932 253355
Enq No 511

LED module

Lascar Electronics has launched the DPM 340 LED module. The 3.5 digit unit has a panel cut-out of 38 by 18mm and an individual digit height of



7.6mm. It comes in a carrier with integral snap-in bezel. The module operates from a 5V supply and has auto-polarity, auto zero and ± 200 mV full scale reading. A PCB socket strip is provided with each module for connection to the target instrument.
Lascar Electronics
Tel: 01794 884567
Enq No 539

Battery backup supplies

Kingshill is stocking BS30 24V battery-backup power supplies for mains-failure applications. The series covers outputs from 2 to 40A. Though designed for sealed lead-acid batteries, the units can be adjusted

for NiCd systems. Hold-up times are determined by battery size. The use of linear technology makes the units suitable for fire and security alarms, process control and radio communications.

Kingshill Electronics
Tel: 01634 821200
Enq No 513

Videophone IC

Toshiba has launched a single-chip MPEG-4 videophone system-level IC for the IMT-2000 digital mobile phone system due for introduction in Japan next year. The IC conforms with standard video-and-speech compression and integrates an MPEG-4 video codec, speech codec, audio and video multiplexer and a 16Mbit DRAM. It supports MPEG-4 image compression and decompression. Using 0.25 μ m CMOS technology, it measures 10.84 x 10.84mm.

Toshiba Electronics
Tel: 0208 938 4644
Enq No 514

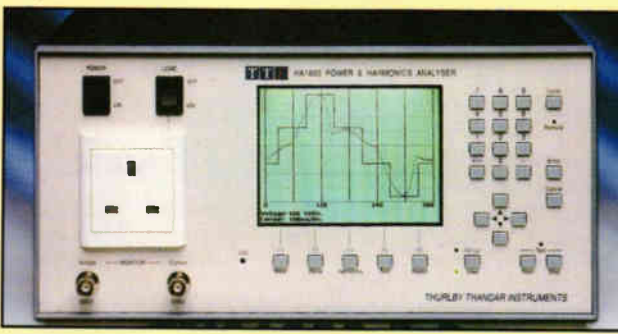
PMC analyser

VMetro has announced the PBTM-515 PMC bus analyser. Sampling speed is up to 66.7MHz. For debugging, testing and validation of PCI chips, boards and systems, the module contains a complete logic analyser for the PMC bus. The analyser can capture and display all bus activity in PMC motherboards with trigger and store qualifiers, and has statistics functions to measure PCI performance. It operates on PCI buses up to 64-bit wide. The user can control the analyser via USB and RS232. It can be expanded with

Mains/harmonics analyser

The HA1600 from Thurlby Thandar is a mains and harmonics analyser with graphical display to test compliance with forthcoming EEC directives relating to the harmonic content of the current waveform. It is capable of continuous real-time analysis of voltage and current. As a general-purpose mains analyser, it can measure square wave, VA, root-mean-square voltage and current, peak voltage and current, crest factors, total harmonic distortion, power factor, frequency and inrush current with a rating of 16A rms continuous. As a harmonic analyser, it is for precompliance measurements using normal mains supplies. Its shunt resistance of 3m Ω lets it make compliance measurements to IEC1000-3-2 with a suitable power source. Capabilities include real-time class D evaluation and visual display, continuous harmonic calculation of harmonic limits to IEC1000-3-2, inrush current analysis and timed test sequences with analysis of fluctuating harmonics.

Thurlby Thandar
Tel: 01480 412451
Enq No 509



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ECC85	3.00		PL508	3.00	6BW7	3.00	13E1	85.000
ECC85	3.00		PL509/519	10.00	6BX7GT	7.50	5Z2B	27.50
ECC88	6.00		PL802	4.00	6BZ6	3.00	805	45.00
ECC808	15.00		PY500A	3.00	6C4	2.00	807	7.50
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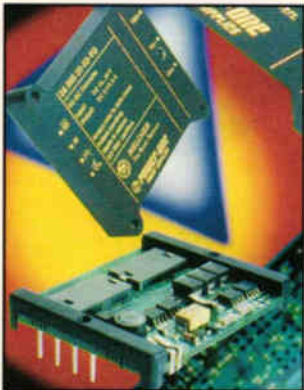
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piggyback modules or daughter cards, such as an exerciser and a protocol and timing violation detector.
VMetro
 Tel: 00 47 22 106090
Enq No 515

25W DC-to-DC converter

Available from Melcher, the IMS25 is a 25W PCB-mounting DC-to-DC converter for processor-based applications including telecoms, data-processing, networking, test equipment and industrial automation. Housed in a 50 by 40mm package



with a profile of 8.5 or 10.5mm, it is based on a forward converter topology, using all-surface-mount components and planar magnetics to provide a conversion efficiency up to 91 per cent. A synchronous rectifier provides its three fixed output voltages of 2.5, 3.3 and 5.1V. No external circuitry is required as the input and outputs are internally filtered to reduce ripple and noise and comply with EN55022. Isolation complies with EN60950 for the rated input voltage range, so it can be used in IT and office equipment. Two versions are available. The open-frame version has the 8.5mm profile for protected environments. The ruggedised full-case version has lugs for screw mounting and a coating for protection from harsh environments.
Melcher
 Tel: 001 425 474752
Enq No 516

Low-noise amplifiers

A series of low noise amplifiers starting with a 14dB gain variant with a bandwidth from 10 to 1000MHz has been introduced by Pascal. It has parameters of IPO3 at 52dBm and noise figure less than 3dB over most of the frequency range. Units are powered by 15V DC and come in



SMD trimming pot

Bourns has introduced an SMD trimming potentiometer, The Trimpot 3361 has a plastic housing and rotor, and can withstand infra-red radiation, convection solder reflow and high-pressure wash systems. The component's rotor adjustment slot has been designed for high-speed automatic machine interfaces.
Bourns
 Tel: 00 41 41 768 5555
Enq No 517

rugged, screened aluminium housings.
Pascal
 Tel: 01983 817300
Enq No 518

Audio and video connectors

Dubilier has announced audio and video connectors for mixing desks, industrial audio equipment and



measurement devices. Available in 6.35 and 3.5mm sizes, the range includes RCA phono plugs and sockets, headphone and earphone jacks and battery connectors.
Pascal
 Tel: 01371 975758
Enq No 519

Terminal and chip card ICs

Infineon is offering two chips for contactless chip card applications – the SLF9000N secure-terminal contactless-logic IC and SLF9611 security access module. They are for public transport applications, security access and secure RF identification



cards. Controlling communication and security functions between chip card, terminal and background system independently from each other, they let the terminal read all chip cards with an ISO14443 contactless interface. The SLF9000N enables the terminal's communication with a chip card. The device supports type A (amplitude shift keying 100 per cent) and B (ASK 10 per cent) contactless modulation methods for transmission in the RF band and offers anti-collision methods. The SLF9611 enables secure authentication between the terminal and the chip card and handles the card's communication with the background system. It allows the background system's online transactions for administration, maintenance and software updates.
Infineon Technologies
 Tel: 00 49 89 234 24497
Enq No 520

150MHz receiver

For remote utility metering, Micrel has introduced two Qwikradio 150MHz single-chip RF receivers – the MICRF004 and F044. They take RF directly from the antenna, and provide

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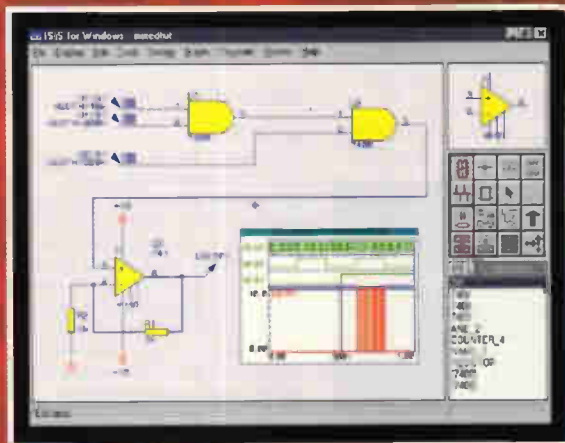
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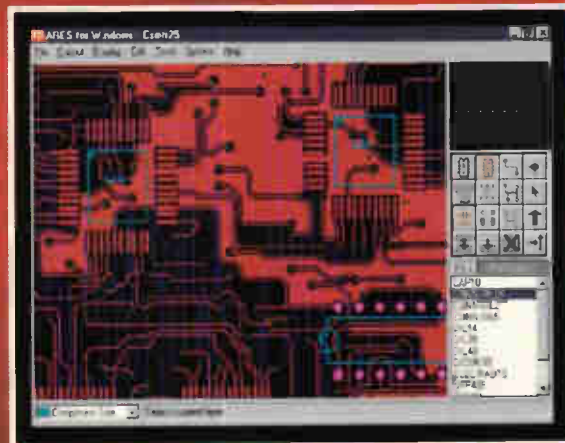


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a digital data-stream output. The F004 comes in SOIC-16 packaging and the F044 in a SOIC-8 package. Power consumption is 2.4mA and can be reduced by duty cycling; a 10:1 duty cycle reduces it to 240µA.

Micrel Semiconductor
Tel: 01635 524455
Enq No 521

Power MOSFETs

Intersil now manufactures two n-channel power MOSFETs using its Ultrafet technology. Rated at 100V at 75A with a maximum $R_{ds(on)}$ of 14 and 8mΩ, the HUF75645 and 75652 respectively are for uninterruptible power supplies, DC-to-DC converters, load switching, motor controls and AC/DC power supplies. The firm claims they can withstand high peak currents and energy in the avalanche mode for switching inductive loads.

Intersil
Tel: 01344 350250
Enq No 523

Drive cooler

Sight Systems has introduced a drive cooler for use with its industrial chassis systems. The SR-DC1 cooler fits in front of the 5.25in drive housing. Based on a Papst 8000 12V DC fan with external rotor motor, the cooler has integrated commutation electronics and protection against



reverse polarity, blocking and overload by PTC resistor, partially impedance protected. The fan delivers airflow at about 34m³/hr and incorporates a dust filter accessible from the front panel of the computer.

Sight Systems
Tel: 01903 242001
Enq No 524

Triple reset generator

Linear Technology has introduced the LTC1726 triple-supply monitor, which lets the user monitor three supply voltages with ±1.5 per cent threshold accuracy. It has an open-drain reset output with an adjustable delay so supervisory functions can match the application. The reset and watchdog time-out periods are adjustable using external capacitors. The monitor is configured for 5 and 3.3V or 3.3 and 2.5V with the third supply adjustable down to 1V. It comes in SO-8 and eight-pin MSOP packages.

Linear Technology
Tel: 01276 677676
Enq No 526

Portable GSM measurement

Rohde & Schwarz's new TS55-C3 portable GSM measurement system is for GSM900, 1800 and 1900 use indoors and outdoors. The system includes a triggering circuit and a TS55-RX three-channel RF receiver plug-in card inserted into the PCSP coverage measurement unit. Each channel on the plug-in card can be set individually and assigned with a frequency of any GSM band. The GSM900 band also covers GSM-R (railway) and GSM-E (extended). The system is controlled by the firm's Romes measurement software. It has four RS232 serial interfaces for connecting external equipment such as GPS receivers or GSM test

modules. When used outdoors, the measurement system is supported by a navigation system for dead reckoning, recording position data even without a GPS signal. For this, a D-GPS-compatible 12-channel GPS receiver and an inertial navigation system are integrated. This lets the user perform distance triggered field measurements without adaptation to the vehicle. The firm claims the complete system will fit in an attache case.

Sight Systems
Tel: 00 49 89 4129 3779
Enq No 525

Comms analyser

Tektronix has introduced the CSA8000 communications signal analyser for transmitter designers, manufacturing test engineers and technicians. It can test 10Gbit/s transmitters and can handle multi-rate optical communications testing. The



user-configurable modular architecture and various optical plug-in modules support conformance testing to multiple standards. Short-term trigger jitter is typically less than 1ps and timebase stability less than 0.1ppm.

Tektronix
Tel: 01344 392000
Enq No 527

Microwave materials

Frequency Products has launched microwave ferrites and dielectric materials shaped as rods, blocks, discs, truncated triangles, hexagons and substrates. For microwave communication applications, the ferrites include garnets and spinels. Garnets can be supplied in yttrium-aluminium, yttrium-gadolinium aluminium iron and yttrium gadolinium iron. Spinel is available in lithium-titanium zinc iron, nickel-chromium



zinc-iron and magnesium-manganese-aluminium-iron. Dielectric materials include temperature-compensated types in zirconium-titanate. Typical characteristics include dielectric constant of 37 ±1, Q of 5000 at 9.4GHz, loss tangent less than 0.0002 and temperature coefficient of -3 to +12ppm⁻¹ ±1. Schaefer plastic dielectrics are also available. These comprise a finely divided ceramic filler dispersed in a polystyrene matrix. They come in 114 by 114mm blocks up to 30mm thick. Typical characteristics are dielectric constant from 3 to 20, loss tangent less than 0.0009 and operating temperature from -55 to +100°C.

Frequency Products
Tel: 01460 256300
Enq No 528

Embedded vision processor

The Coreco Mamba from Pinnacle Vision is an embedded vision processor based on the Pentium II for machine vision and medical imaging applications. Data transfer speeds on the 200Mbit/s auxiliary bus are higher than on the host PCI bus. Developers can use the Mamba with the firm's Viper-RGB for colour applications, the Viper-Digital for applications that require cameras with multiple digital inputs, and the Viper-Quad for simultaneous acquisition from up to four cameras.

Pinnacle Vision
Tel: 01784 473990
Enq No 529

Internet enabler processor

Atmel and Aplio have launched the AT75C310 Internet appliance processor IC with embedded Linux operating system, VoIP, audio application software and an application development platform for Internet phones, e-mail phones and MP3 appliances. The VoIP application software delivers telephone sound quality using Packetplus technology. The Linux layer supports DSP functions including modules for a V34 modem, G723.1 and G729A voice codecs, silence compression and echo cancellation.

Atmel
Tel: 001 408 436 4229
Enq No 530

Snap-action switch

Introduced by Matsushita is the CS snap-action switch with built-in connector. No crimp-blade or screw terminals are needed and no soldering is required. It has a dust prevention



One-chip telephone ICs

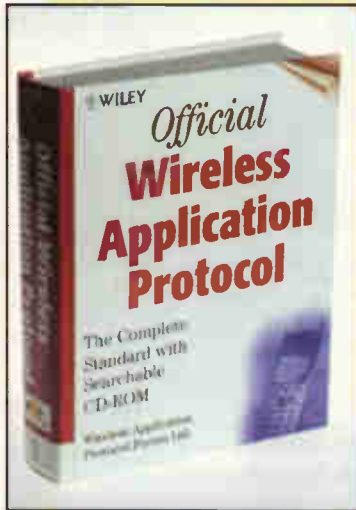
Philips has announced two one-chip telephone ICs that combine several onboard functions of speech, dialler and ringer devices. The UB2050 and 2051 let corded phone makers replace three chips with one. They come in the 28-pin SO package and integrate features such as DTMF and pulse dialling, last-number redial and repertory dialling of 13 numbers with up to 21 digits. Also onboard are an integrated earpiece amplifier with gain boost facility, microphone amplifier and programmable ringer with four-level volume control and up to four melodies programmable via the keypad. Pulse and DTMF dial settings are adaptable to different parts of the world via an external resistor and the dialler has two access pause intervals of 2 and 3.6s. Line current is from 11 to 140mA and they operate at DC line voltages down to 1.45V. Automatic gain control provides line-loss compensation for the microphone and receiving amplifiers.

Philips Semiconductor
Tel: 00 31 40 272 2091
Enq No 522



Official Wireless Application Protocol: The Complete Standard with Searchable CD-ROM

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guard and gold-clad double-layer contacts. The contact arrangements are optionally available in normally open and normally closed variants, and are rated 5V DC at 1mA to 30V DC at 0.1mA. It is available in standard and backward lever positions and as a pin plunger type. These options allow fine-tuning for stroke setting. There is a choice of four actuators. The switch is UL and CSA approved and is suitable for vending machines, photocopiers, printers and pinball machines.
Matsushita Electric Works
Tel: 01908 231555
Enq No 531

Inverter modules

Converter-brake-inverter modules for AC input voltages from 110 to 550V three phase have been introduced by Ixys and are available from GD Rectifiers. The converter section uses



1200 to 1600V planar glass-passivated rectifier diodes. The IGBT inverter stage consists of six NPT-IGBTs, with breakdown voltages of 600 and 1200V and current ratings from 4.5 to 25A and 2.5 to 17A respectively at 90°C case

temperature. Turn-on losses are reduced by platinum diffused fast-recovery diodes with soft recovery characteristics. The integrated IGBT brake-chopper with its associated flywheel diode can be used with an external resistor for dynamic braking in the regenerative mode. An NTC temperature sensor lets the user measure the temperature inside the module and on the surface of the DCB substrate. Dimensions are 82 by 37.4mm and they can be mounted with two screws to a heatsink.
GD Rectifiers
Tel: 01444 243452
Enq No 532

Relay for AC

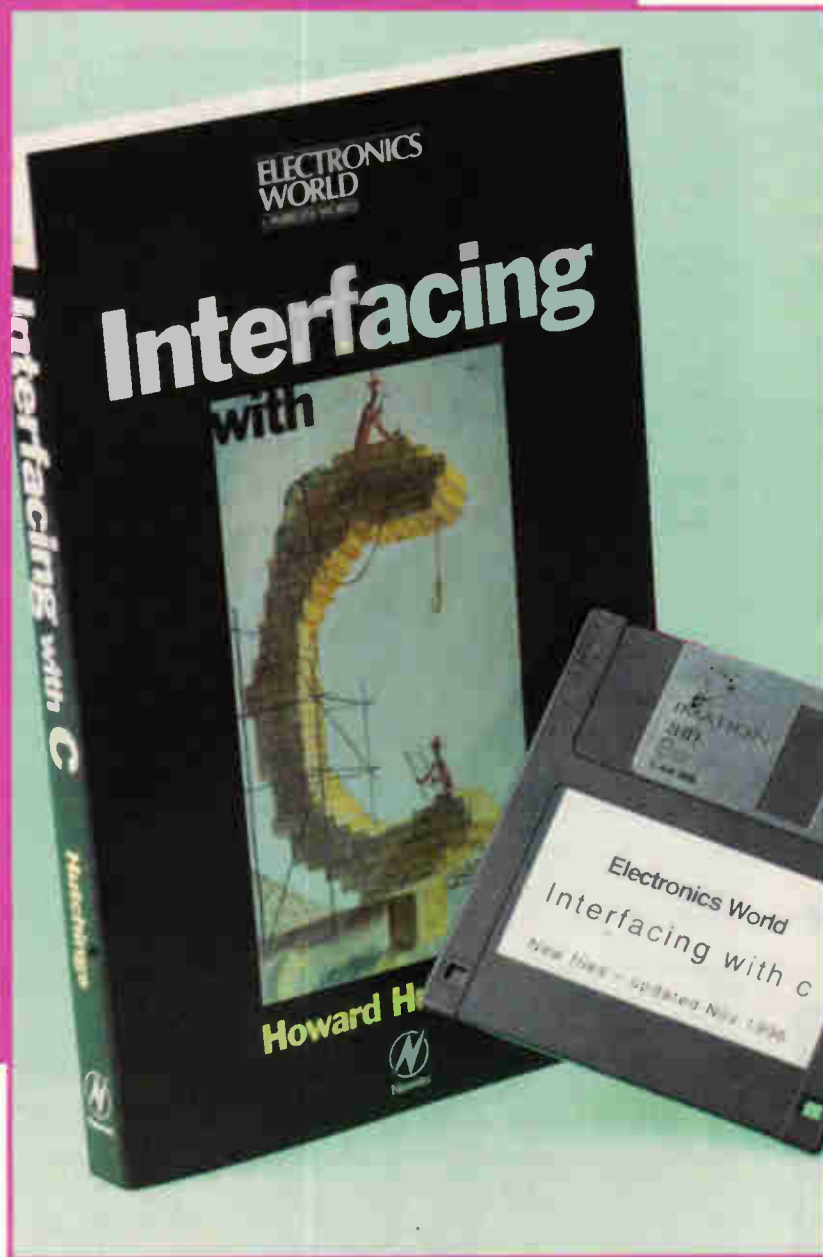
Teledyne has introduced a PCB and surface mountable AC solid state relay using its Powertherm system. The LR1200480D40 DC controlled relay can control up to 40A at a line voltage of 480V AC. Features include zero voltage turn-on through back-to-back SCR output switches capable of 1200V peak transients, logic compatible control and 4kV rms isolation between input and output. Voltage and current ratings are from 240 to 480V AC and 25 to 40A. A triac output version rated at 10 and 25A is also available. There are three control types – 90 to 280V AC input, 3 to 32V DC input and random turn-on DC control for phase control and PWM use. Applications include heater controls, light dimmer controls and process controls. It comes in a plastic package measuring slightly less than 3.5 by 2.8cm.
Teledyne
Tel: 01236 452124
Enq No 533



Placement routing tool

Zuken-Redac has announced a version of PR Editor, its placement and routing tool for Mentor Graphics' Board Station. Operating under Unix and Windows NT, it can cross probe between logical and physical components to provide autorouting directly from a logical schematic. The splash and route feature automatically pushes aside components and reroutes them as they are moved. The tool can automatically route through dense areas such as connectors and ball-grid arrays by switching to free angle mode. It can create partial power planes of arbitrary shape and automatically eliminate isolated areas. Constraints defined by the engineer in Board Architect are automatically read and obeyed. Interactive tools can solve high-end routing problems such as matched delays and crosstalk. It supports area fills, micro-vias and HDI designs.
Zuken-Redac
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Especially useful for students, the original *Interfacing with C*, written for Microsoft C Version 5.1, is still available at the price of £7.50. Phone 0181 652 3614 for bulk purchase price.

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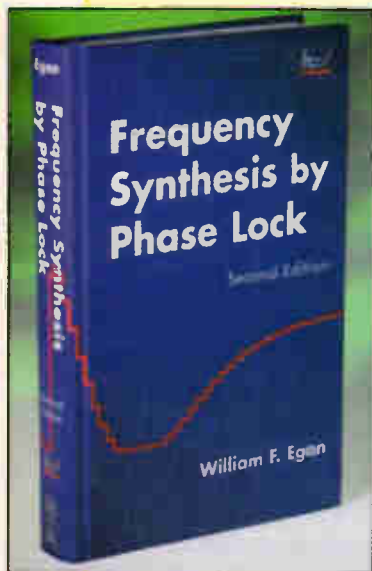
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Frequency Synthesis by Phase Lock

Frequency synthesis is an important element in the design of all communications equipment, but has taken on new life recently with the advent of new hand-held wireless devices. This technology not only allows wireless transmitters to change frequencies quickly, but also gives high reliability and security in transmissions. Thus, mobile devices such as cell phones can utilise this technology to change frequencies until a suitable one is found for the location in which it is being used.

- Emphasises the fundamentals of frequency synthesis
- Based on a course that Dr. Egan has been teaching for over 20 years at Santa Clara University
- Provides a link to the Wiley ftp site for the use of associated MATLAB exercises
- Taken together with Phase Lock Basics by the same author, the two books provide readers with complete coverage of the field.

CONTENTS Introduction; The Elementary Phase-Locked Synthesizer; Modulation, Sidebands and Noise Spectrums; Frequency Dividers; Phase Detectors; Higher Order Loops; Sampling Effects; Architectures; Large-Signal Performance, Natural Acquisition; Acquisition Aids; Spectral Purity; Computer Aided Engineering.



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Chip for battery management

Dallas has announced a battery-management chip. The DS2438 stores battery-specific data and tracks battery parameters, including temperature, voltage, current and remaining charge. Once inside the battery pack, it can identify the pack and configure itself to charge and monitor the battery. Each chip gives its pack a unique 64-bit net address so multiple battery packs can be wired to one bus but addressed individually.

Dallas Semiconductor
Tel: 001 972 371 6085
Enq No 534

Prototyping adapter boards

Emulation Technology has announced ball-grid array and chip-scale package prototyping adapter boards. They let designers add 1.27, 1 and 0.8mm pitch BGAs to prototype breadboards. For testing, designers plug the adapter onto a prototyping board with a 2.54mm centre grid and surface mount the IC to the board to make the prototype ready to operate. Extra jumper pads are included on both sides for custom wiring and additional circuitry. Using it as a test socket, engineers can remove a problem BGA or CSP component from the production board, install the component on the adapter and test

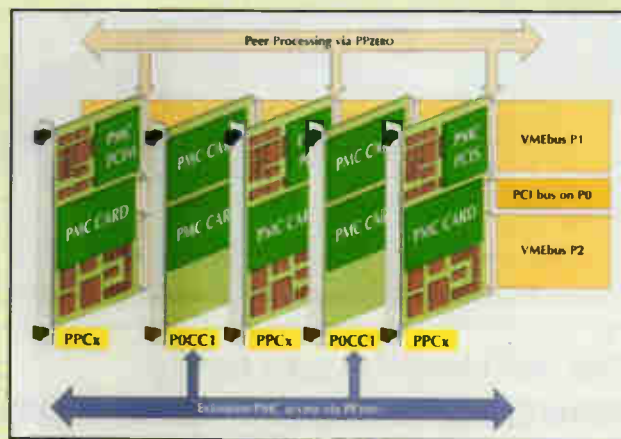
the component outside the circuit. This is done by attaching a test socket to the prototyping adapter and wiring the adapter to a test instrument.

Emulation Technology
Tel: 001 408 982 0660
Enq No 535

Dual hot-swap controller

Linear Technology has introduced the LTC1647 dual hot-swap controller with independent inputs to control the supply to modular components from the same supply or to handle supply sequencing of multiple voltages. Each channel accepts supply voltages from 2.7 to 16.5V and provides inrush-current limiting, electronic circuit breaker and a fault flag. The controlled turn-on from independent inputs gives the flexibility to control device bay applications or multiple supply systems such as disk-drive arrays. The device is for multiple loads or multiple supply applications and comes in SO-8 or 16-pin SSOP packages. The SO-8 parts have two on pins for controlling two loads from the same input supply. They support automatic retry or latch the supplies off if there is an overcurrent fault. The 16-pin SSOP version separates the fault and on pins to allow automatic retry or latch off the supplies.

Linear Technology
Tel: 01276 677676
Enq No 537



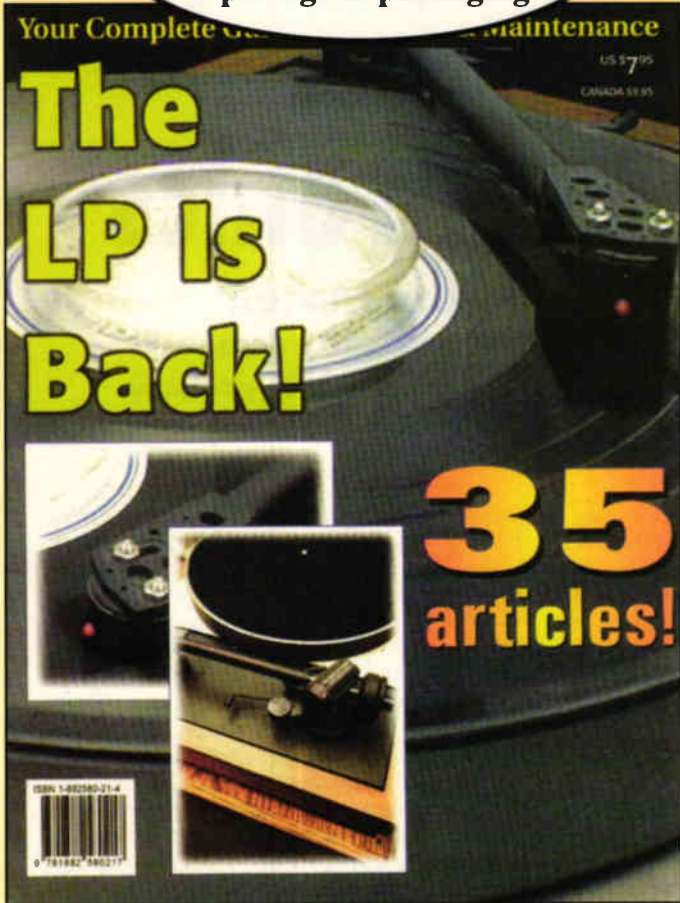
Backplane architecture

Radstone has announced PPzero, an architecture that provides peer multi-processing between VME boards via a PCI secondary bus concurrently with VME data transfers. The hardware and software components complement the firm's PowerPC single board computers, and can be retrofitted to products in the field. Hardware components extend the PCIbus from a PowerPC board, via the standard VME P0 connector, to the backplane letting multiple VME boards communicate via PCI. The 6U PMC carrier cards also interface to PPzero. Software components maintain the VME backplane driver interface standards for PCI transfers between peer processors. Vxworks support is available and Lynxos support is planned. Many Cots software drivers, developed for PCI desktop systems, are directly applicable to PMC format products, giving support for integrators. Custom system functions can also be absorbed into PMC cards.

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a lifetime in electronics

John Linsley-Hood recalls the emergence of the IC and his first experiences with PLLs, the synchrodyne and cassette recorders. Rounding off this final article, John also shares his thoughts on how digital technology is changing the face of electronics.

There is a story, probably apocryphal, about a semiconductor manufacturer in the USA that made components for electronic calculators, and also made the complete calculators.

It is said that the company was annoyed to find that other manufacturers in areas of low labour costs were buying their components, assembling them, and then selling the final calculators at a lower price than they could make them for themselves.

The company's response was to redesign their calculator so that it just consisted of one calculator IC and a display device. They would have put the display on the IC as well but for the fact that it would then be too small for the user to read.

In a nutshell, that is the story of the IC – that the smaller the chip the lower its manufacturing cost, and the more one can get on the chip the lower the associated labour costs will be. The impact of this was brought home to me recently by the announcement in a technical brief that all the electronics for a complete colour TV could now be provided on a single IC.

This is both good news and bad news for the electronics enthusiast wanting to do a bit of DIY design. On the credit side, there is a large number of general purpose ICs such as op-amp gain blocks and voltage regulators, as well as a host

of other 'application specific' devices. If one of these will do the job, fine. It will be well designed and bug-free.

On the other hand, most ICs have pin connections at 0.1 inch spacings and this makes for a very congested board layout. Bread-board layouts based on copper strip-board drilled at tenth inch spacings tend to look pretty scruffy and may not work as well as they should – especially if HF signals are involved.

An answer to this problem, though this may appear somewhat daunting to the amateur, is to equip yourself with the wherewithal to print, etch and drill your own printed circuit boards. Doing so will also allow you to make up circuits such as 10.7MHz FM IF amplifiers, where stray capacitances and inductances would otherwise lead to instability and impaired performance

A phase-locked loop FM tuner

With the advent of small, low-cost, plastic-encapsulated junction transistors, the design of simple and drift-free FM tuners was a much easier task than it had been in the late 1940s. A growing interest in the 'phase-locked loop' prompted me to have another look at designing an FM system. The PLL seemed to me to be ideal for use as a low-distortion FM demodulator, Fig. 1.

Most of the early FM demodulators, such as the 'slope detector' – a fancy name for a slightly off-tuned circuit – or

the ratio detector or the Foster-Seeley discriminator, suffered from a fair measure of distortion. They produced between 0.5% and 2% depending on how well one had done the tuning. Admittedly this was better than the typical 2-5% figure for the average AM demodulator, but it wasn't very good in comparison with the 0.01% figure beginning to be expected from the audio amplifier.

I played with phase-locked loops and their variants over a period of some years, making gradual improvements in one way or another. I found that the crucial factor in FM radio applications is the linearity of the voltage-controlled oscillator. I showed two very linear designs in *Wireless World* in 1975 and 1979.

The first of these designs used a multivibrator layout. It had a splendidly linear control voltage/output frequency relationship, but – as is usually the case in RC oscillator layouts – it was somewhat noisy due to frequency jitter.

The later design used a linearised 'varicap' LC circuit and was quieter. Both of these circuits gave less than 0.1% distortion at $\pm 75\text{kHz}$ deviation. This was not quite in the same league as the better audio power amplifiers of that period, but a lot better than the average run of FM tuners.

A PLL stereo decoder

A further use for the phase-locked loop is to extract a single frequency sinusoidal signal from a noisy background which might otherwise swamp it.

A very good example of this kind of use is in the GE/Zenith technique for decoding a stereo-encoded FM signal into its separate left and right-hand components. In order to do this it is necessary to regenerate a small amplitude 19kHz 'pilot tone' from which a further 38kHz 'sub-carrier' can be constructed.

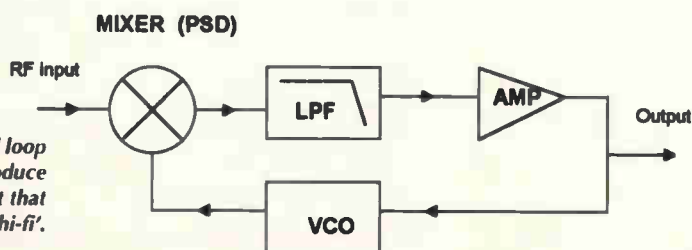


Fig. 1. The phase-locked loop allowed designers to produce FM radios with output that could be classed as 'hi-fi'.

The signal carried by this modulated sub-carrier can then be recombined with the original 'mono' mixture to give a pair of stereo signals.

An elegant, albeit somewhat complex, circuit for doing this, which used a PLL to extract the 38kHz carrier, was described by Portus and Haywood in *Wireless World* in September 1970. This was obviously a good thing. It wasn't long before an IC that could do the whole lot on one chip – along the lines proposed by P and H – was available off the shelf. That IC was Motorola's MC1310P.

Discrete ideas

My interest in this topic was aroused because I had designed a low distortion PLL FM receiver for my own use and I wanted a stereo decoder to go with it.

Motorola's IC version of the P and H design had clear benefits. However, I had a sneaking feeling that the performance of the decoder might be improved if the signal channel could be handled entirely by discrete components. In particular, I wanted to use a 'sample-and-hold' decoder arrangement using junction FETs as the switches in the sampling circuit.

The final layout was described in *Electronics Today* in 1987 and 1990. It used a steep-cut input filter of my own design to remove the noise components which might otherwise be demodulated by harmonics of the 38kHz sub-carrier.

Overall input-to-output linearity was better than 0.05% at 1V RMS. For those who like 'tweaking', I also provided a phase adjustment control that allowed more precise synchronisation of the switching waveform in relation to the composite audio signal.

In this circuit – or in any other adjustable stereo decoder – the best stereo channel separation can be found by adjusting the circuit while listening to audience applause. Ideally, the applause should appear to subtend an arc slightly greater than that between the loudspeakers.

If a dual-trace oscilloscope is available, the same adjustment can be made, on any stereo signal, by displaying the L and R outputs on the X and Y axes of the instrument. In this case, the greatest separation is indicated by the maximum roundness of the resultant Lissajous figure.

MOSFET audio power

That my feet should take me into this field was partly as a result of technical curiosity, and partly as a result of requests from friends.

I first encountered the MOSFET as a very high input impedance device when wearing my 'nine-to-five' hat as a physicist cum industrial electronics

engineer. My need at the time was for a very high input impedance amplifier for use in an atmospheric pressure ionisation chamber. A prototype arrangement of such a chamber using a sub-miniature electrometer valve performed fairly well but the circuit was highly microphonic – a bad defect in any industrial equipment.

The best answer to this problem appeared to be to use an n-channel MOSFET solid-state triode. In 1960 or thereabouts, Plessey Semiconductors offered a commercial example of one of these. It had an extremely high input impedance and was mechanically robust.

However, these MOSFETs carried a warning that gate voltages in excess of ±15V would cause immediate destruction of the device. In reality though, this was not too difficult a constraint.

Nylon shirts and absent earths

Only one of my colleagues had a problem with blown gates. It turned out not to be due, as we had first supposed, to electrical discharges generated by his nylon shirt. Rather it was because he had undone the earth lead from his mains-voltage powered soldering iron for reasons that appeared good at the time. As a result of this, his soldering iron bit carried a gate-destroying potential of 120V RMS.

The MOSFET is typically a device in which a conducting layer, the 'gate', has been deposited on the surface of an 'intrinsic', i.e. undoped, layer of silicon, the 'substrate'. It is separated from it by a thin insulating layer.

If a voltage is applied to the gate it will induce electrostatic charges in this layer. Now current can flow across it from 'source' to 'drain'.

The speed at which conduction occurs following the application of a charge to the gate depends on how quickly current can flow along the gate

connecting lead, and how quickly the charge on the gate can change. These factors depend on the resistance, inductance and capacitance of the system.

From the viewpoint of the circuit designer, the effect of this is that MOSFETs are exceedingly fast in operation, and can burst into oscillation with very little provocation unless the circuit and its layout are carefully chosen.

The simple 30W integrated amplifier

My 75W amplifier had proved exceedingly popular. However, a number of my friends had commented that while it was undoubtedly a very nice amplifier, it was all a bit complicated. What they would really like was an 'integrated' – i.e. preamplifier plus power amplifier – design with an output of about 30W per channel.

A simplified layout using op-amps as the preamplifier gain blocks and power Darlington transistors as the output devices was published in *Hi-Fi News* in January to March of 1980. This was after my friends had tried out the prototype, and concluded that it met their needs.

Sadly, this design soon got into the hands of the hi-fi cognoscenti, who compared it with the very best they could find. They observed that the very best – at ten times the price – were actually somewhat better.

There were some things that could be done to improve the basic 30W design. Of these, the major one was to replace the output Darlington transistors with power MOSFETs. This made a very nice sounding amplifier. I have shown the revised output stage circuit in Fig. 2.

Using TL071 op-amps, instead of the original 741s in the preamplifier circuitry added a final touch. I sometimes thought that if some eccentric burglar were to steal all my audio amplifiers,

What is a phase-locked loop?

The PLL is beguilingly simple – superficially at least.

If a phase-sensitive detector is fed with two sinusoidal inputs, then the detector's output will be a composite 'sum and difference' frequency signal.

Generally, one of the PLL's input signals is derived from some external source and the other is derived from a voltage-controlled oscillator or VCO. The control voltage for the VCO is derived from the loop output.

The phase-sensitive detector output is filtered to remove the 'sum' frequency signal. An interesting situation arises if these two input signals should momentarily be at the same frequency. In this case, if the external signal is large enough, the loop will 'lock', and the VCO will be forced into frequency synchronism with the input signal.

If the oscillator has a linear relationship between the input control voltage and output frequency, then the PLL will provide a linear – i.e. distortion free – means of demodulating an input FM signal.

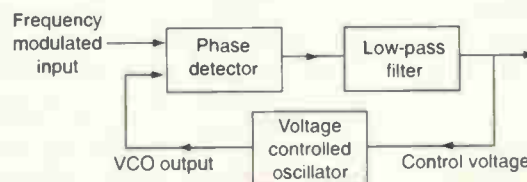
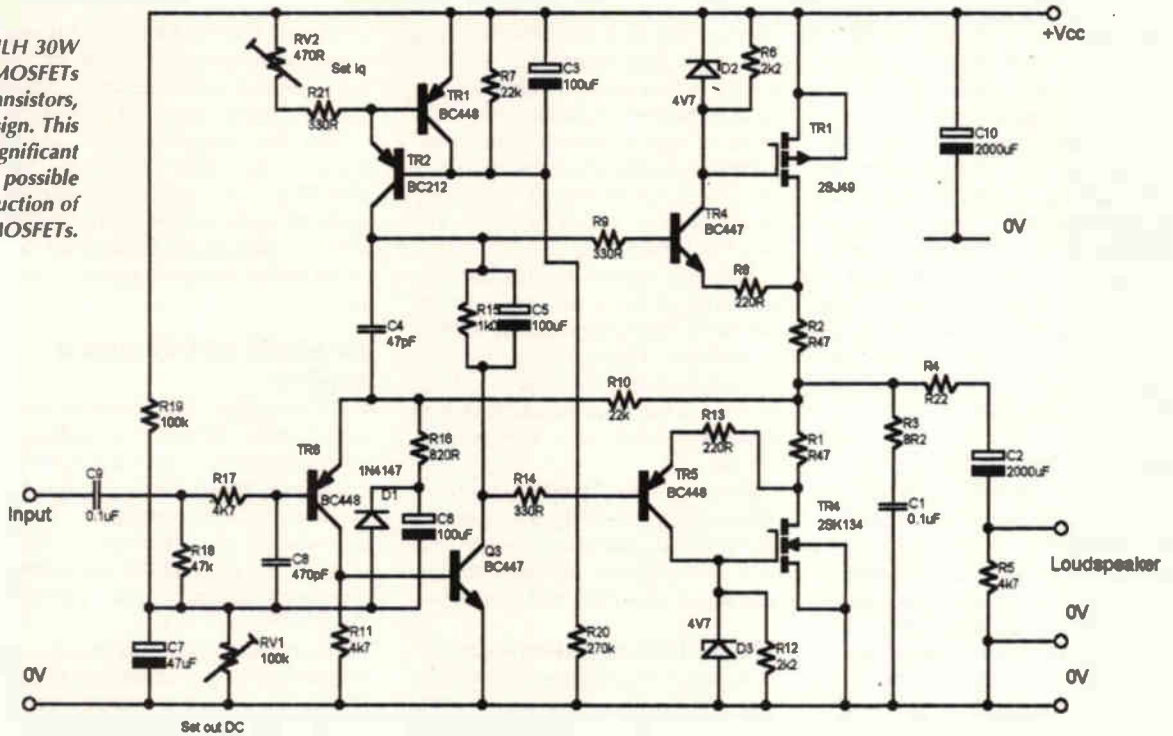


Fig. 2. Revised JLH 30W amplifier using MOSFETs instead of bipolar transistors, as in the original design. This represents a significant improvement, made possible by the introduction of MOSFETs.



this design – but direct coupled, and fed from a symmetrical pair of power supply lines – is what I'd make to fill the gap.

HF equalisation

Any amplifier system using closed-loop negative feedback is prone to instability unless the loop gain is very low, or the loop phase shifts are small. Much of the art of the amplifier designer is concerned with achieving adequate levels of loop negative feedback while preserving a sufficient loop phase margin.

The classic method of meeting these, sometimes conflicting, requirements is to slug the major gain element with a capacitor. This produces what is called single-pole 'dominant lag' HF compensation.

I prefer to do this compensation by applying a capacitor across two gain stages so that this part of the circuit acts as an active integrator. Unfortunately, this technique only works if the high-frequency gain of the circuit is high. This condition is more easily met by using power MOSFETs with effective gain transition frequencies of the order of hundreds of megahertz, rather than by bipolar junction transistors whose gain transition frequencies are less than a tenth of this.

In the hope of encouraging more of my peers to use power MOSFETs in their audio amplifier designs, I published several further circuits of this type in *Wireless World* namely in the April to August, 1982 and June 1993 issues. I also had designs in *Hi-Fi News*, December 1980, and in *Electronics Today*, June 1984, May 1989.

Cassette recorders

Some time in the 1960s, Philips introduced the 'Compact Cassette'. It was principally intended as a portable recording medium for secretarial and dictation uses.

In terms of simplicity in use, the tape cassette offered many advantages over the existing reel-to-reel tape recorder. However, it was not intended as anything approaching a hi-fi recording machine. This limitation was implicit in the tape track widths and recording speeds of 0.61mm and 4.76cm/s chosen for this new medium

These restrictions did not prevent circuit and equipment designers from exploring the cassette's possibilities as a music recording system though – with particular reference to in-car use. By 1974, cassette recorder mechanisms complete with record/replay and erase heads, and a simple electronic speed control system, were available for use by the DIY enthusiasts.

I had a perfectly satisfactory car radio, which I had no wish to replace, and a high-quality, commercial, 15in/s reel-to-reel tape machine – which was a bit expensive on tapes and awkward to use. However, the possibility of using a cassette recorder to save some of the splendid music now available from the BBC FM stereo broadcasts was an exciting one, so I decided I'd have a go.

Ignoring imperfections in the tape transport mechanisms – which you couldn't do much about, there were two main problems to be solved. One was to design a sufficiently low-noise replay system for amplifying the minute signal generated by the tape creeping slowly across the replay head.

The other was to provide a suitably modified record/replay response curve to give an overall frequency response that was somewhere near flat. The circuitry I used is shown in Fig 3.

An almost universal commercial answer to the need for a low-noise cassette replay system was to use 'Dolby B', which involves record pre-emphasis/replay de-emphasis. This technology was not available to the amateur though. You needed a Dolby licence. A Dolby IC was later offered by National Semiconductors however.

My answer, at least in the medium term, was to adopt 90µs equalisation, as later recommended for 'chrome' tapes. This pushed the standard ferric tapes a bit harder, but I felt the tape coating formulations then available were better than those around when the record/replay time-constants were decreed.

I was so pleased with what I'd done that I invited *Wireless World* to come and listen to the results. My invitation was accepted. Obviously the magazine liked my recorder design too. It was published in the May to June 1976 issues.

Mechanical problems

Having the article published was gratifying at the time, but any mechanical system is short-lived in comparison with any wholly electronic one.

In the case of the cassette recorder, the problems are general wear and tear on the heads and the tape transport mechanism. You could replace the heads, but the rest of the gear – especially with the relatively low cost mechanisms used at the time – began to show its age after a decade or so.

Also, with increasing expectations of performance, the better cassette recorders were now three-head designs. With these, the individual heads could be optimised for their respective functions.

Finally, the demand for better tape speed control, with less wow and flutter led to better drive mechanisms. They had twin drive-motor, 'dual-capstan' layouts. Even crystal-stabilised speed control was adopted for many of the better machines. Sadly though, this sort of electro-mechanical elaboration lies in the sphere of the large manufacturer rather than that of the enthusiastic amateur.

Alternative radio systems

In the 'superheterodyne' receiver, the incoming antenna signal is converted into a signal of a different intermediate frequency in order to obtain the necessary gain and selectivity. This is by far the most common technique used in radio receivers.

I have made a number of superhet designs for interest, to try out some new idea, or to try to cure one or other of the intrinsic drawbacks of the superhet technique. I am not alone in this quest. Over the past sixty years or so, a number of interesting ideas have emerged, one of which is the 'synchrodyne'.

In this, the local-oscillator frequency is chosen to be identical to that of the incoming signal. In this way, the output of the mixer - which will be the sum and difference frequency components of the two signals - will be a signal at twice the signal frequency, and one at the signal frequency itself.

If the 2f component is removed by filtration, then what is left is the wanted signal. The selectivity of the receiver, its ability to discriminate between the wanted signal and one at a closely adjacent frequency, can be achieved by AF filtration.

The most immediate snag with this scheme is that the local-oscillator signal has to be in frequency and phase synchronism with the input one. If it isn't at the same frequency, the output is a piercing howl, and this poses a severe demand for frequency stability.

The 'homodyne' attempts to solve this last problem by extracting the local-oscillator waveform from the incoming carrier by clipping off the modulation. But this ignores the additional requirement that the local oscillator signal must be at phase quadrature to the input and this is difficult to maintain.

My interest in the synchrodyne was prompted by the work I had done on phase-locked loops, as a means of demodulating FM signals. Feeling that this answered the synchrodyne problem, I designed a 'phase-locked synchrodyne' with an off-station muting facili-

ty to cut out howls. This was published in *Wireless World* in the January to March 1986 issues. Although this circuit worked as I intended, it was somewhat complicated to tune, requiring a two-knob (frequency and gain) approach.

Dad's loft

Clearing out my father's loft, I found my grandfather's old 1938 'Philco' four valve table radio, with three wavebands and a 3.5 watt output. This was still in quite presentable condition, but

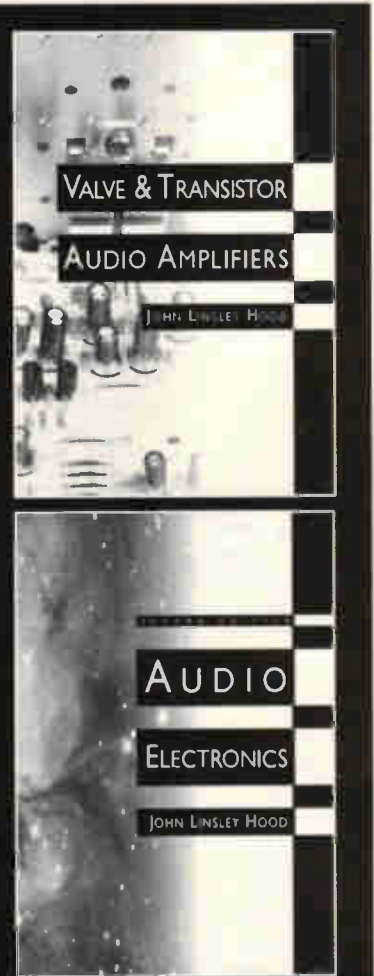
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long since defunct.

It struck me that it would be an interesting exercise to rebuild it using modern components, such as dual-gate MOSFETs for RF, mixer and IF stages, and a group of fast op-amps to provide low distortion AM demodulation and delayed amplified AGC.

A simple audio output stage gave 8.5 watts at less than 0.05% distortion. I was pleased with its performance, and I published the circuit details in the October 1986 issue of *Wireless World*. It is, I am happy to say, in daily use, mainly for Radio Four news broadcasts on 198kHz.

Although I have played with a number of electronic systems – mainly those with audio applications, and almost exclusively as an amateur – my real love is still radio, if only, perhaps, because it is so much more difficult to do it right. The absence of commercial

pressures means that you can do what you think best, without having an accountant or marketing person breathing down your neck.

The awareness of the difficulties involved in designing radios is not restricted to the perfectionist amateur in his shed. It has troubled some of the most prestigious of the broadcasting and research organisations in Europe.

One of the main difficulties is the poor performance of the VHF/FM broadcasting system when listened to in a car. Here, the target is that of matching the performance of the compact disc as an in-car entertainment medium. This has resulted in the 'Eureka 147' proposal.

Eureka was the name rather optimistically given to the ensuing joint research programme. The number '147' is not explained in the literature, but I suspect it refers to the number of

schemes they tried before they got one which worked! I'm giving a brief account of the system, if only because it marks – along with other digital systems – the ultimate parting of the ways between the amateur and the commercial manufacturer.

Thoughts on digital radio

The major benefit from digital audio systems is their greater freedom from interference. The trade-offs are rigid limits on bandwidth and distortion, and an enormous increase in the complexity of the means.

In the case of the digital-radio receiver, the signal is converted into digital form before broadcasting. It is then transmitted by a form of FM known as 'quadrature phase-shift keying' – chosen for its freedom from 'bit' errors.

The broadcast bandwidth allocated to the broadcaster is then divided into randomly-allocated segments. These are then re-assembled by the receiver to form the programme – or collection of data – wanted by the user. This process is called 'de-multiplexing'.

Since the signal had been converted into a digital form before it was broadcast, none of the radio receivers I have made, or have described above, will make anything of them. It is like expecting a CD player to extract music from a five-inch gramophone record. I now find myself on the same shore line, beached by the receding technical tide, as those who failed to adapt to the revolution of the solid state.

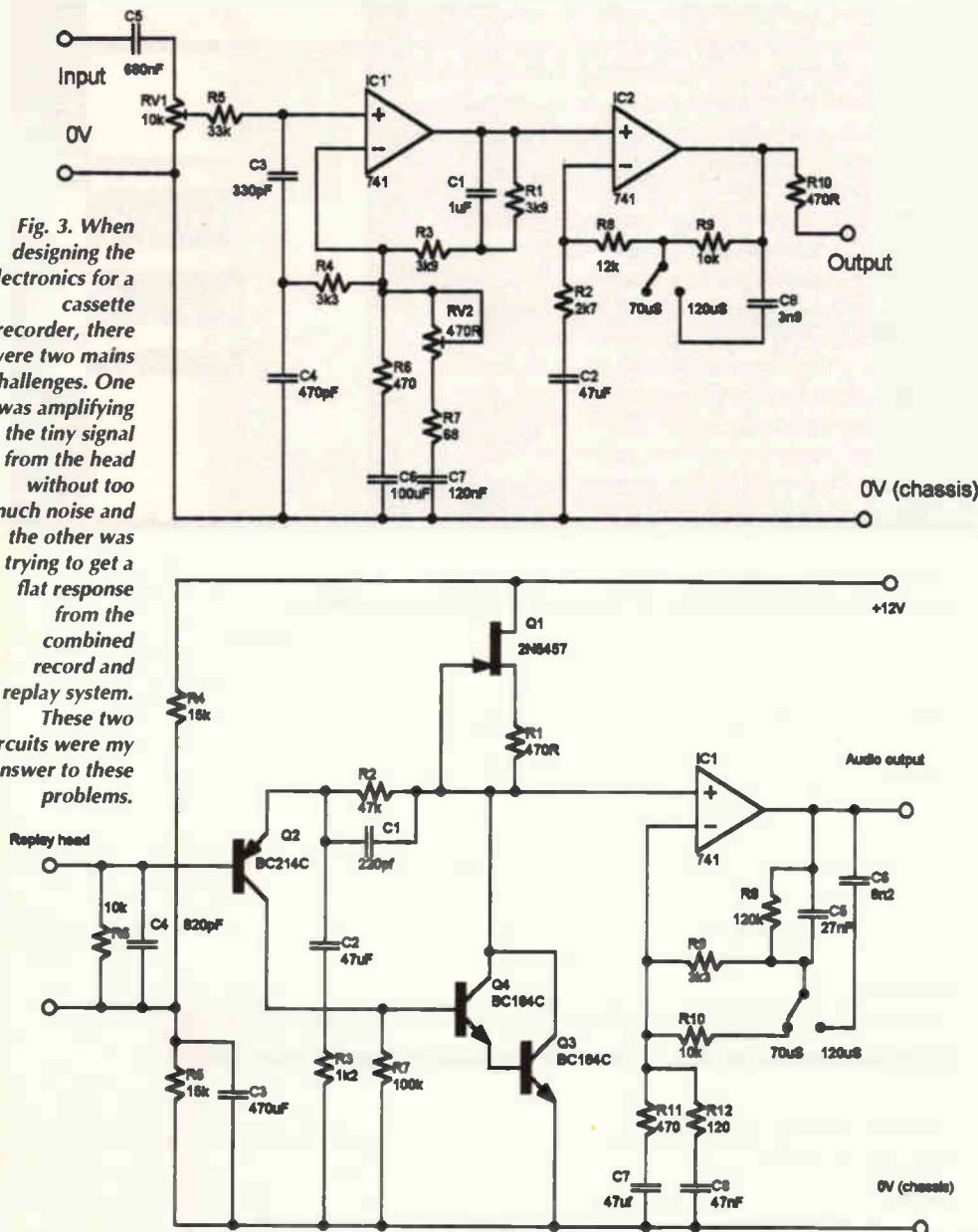
I shall, of course, get a 'digital' radio, but it will be of commercial origin, for the same reasons that I have a commercial CD player, rather than a DIY job.

Meanwhile, I afford myself a wry smile on seeing another digital radio system, described by Slifkin and colleagues in the October 1999 issue of *Electronics World*. In it, the incoming radio signal is amplified, bandwidth limited and converted into digital form, from when on the processing is done in the digital domain, by software.

Luckily, writing software is something that the young amateur can do with his 'PC' in his bedroom just as well as the big boys – apart from the fact that unscrambling the random multiplex of the Eureka coding may prove difficult.

I also note that while digital may be better, the analogue is often a lot cheaper. Take the compact cassette *versus* the recordable Mini Disc for example. I don't know how dear the Eureka 147 will be, but I would guess it will be a lot more expensive than the little 'trannie' in the garage or bedroom. So, maybe, analogue electronics is not quite dead, yet.

Fig. 3. When designing the electronics for a cassette recorder, there were two main challenges. One was amplifying the tiny signal from the head without too much noise and the other was trying to get a flat response from the combined record and replay system. These two circuits were my answer to these problems.





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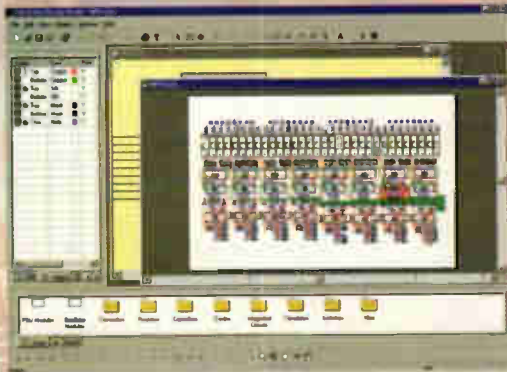
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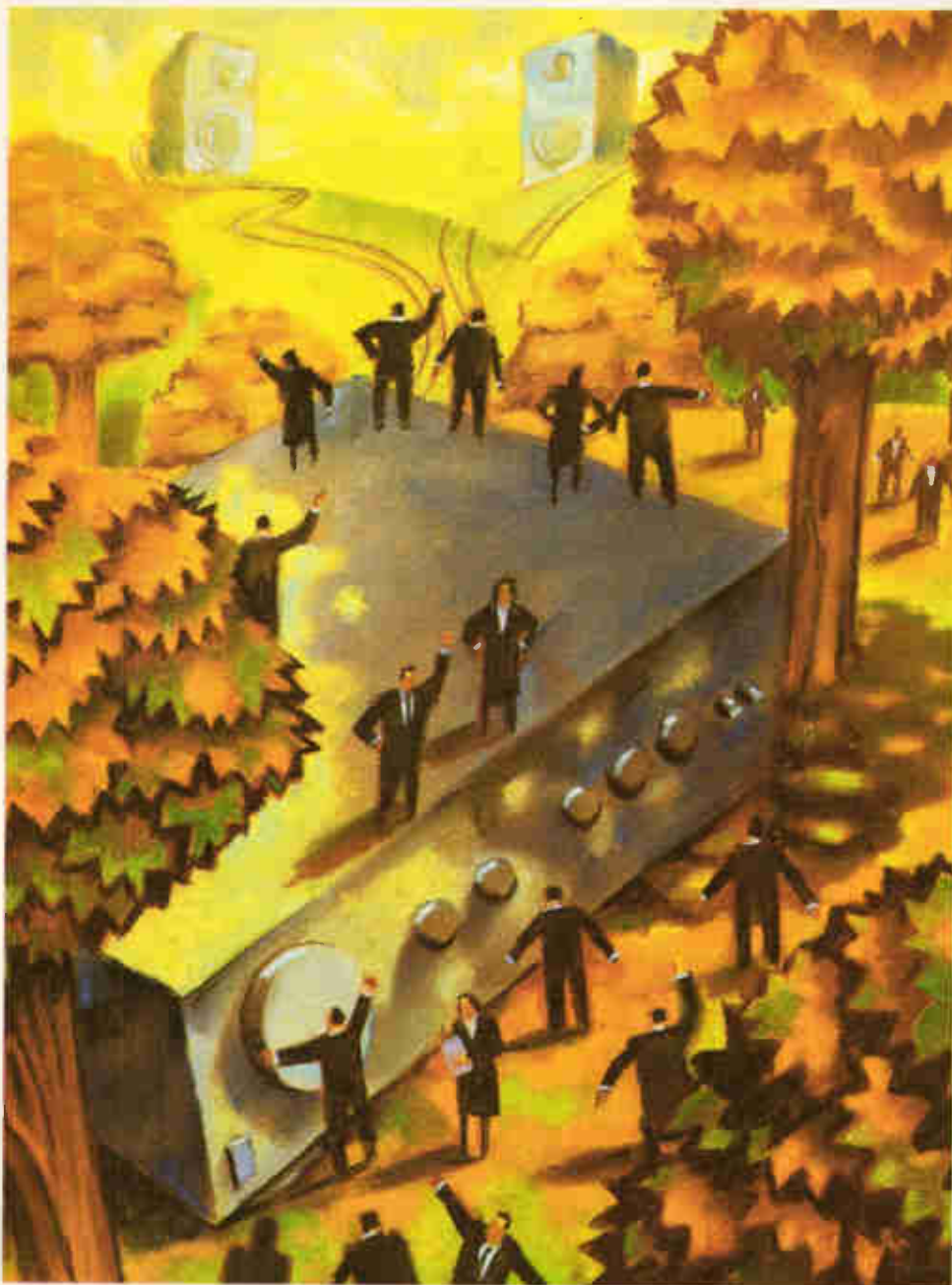
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CIRCLE NO. 118 ON REPLY CARD



A new 100W Class-B topology

In a conventional Class-B amplifier, distortion rises with frequency. But it's at higher frequencies, where the ear is most sensitive, that you want the best performance to suppress the undesirable influences of cross-over switching. Russell Breden believes his reconfigured 100W Class-B design solves that problem, and is far more effective at minimising crossover distortion.

Despite the recent advances in electronic circuitry, one perennial audio problem still remains. How do you make an economical audio amplifier that doesn't suffer from crossover distortion?

In this article, I outline a design method that pushes all crossover products below audibility. The method is illustrated using a specific design example for a 100W power amplifier. As a bonus, this improvement can be obtained by simple modification to the standard generic amplifier circuit.

I can hear a clear difference between Class-A and Class-B amplifiers, especially when they are operated at low levels. These differences disappear though with the design technique outlined. At low levels, where crossover effects make themselves most apparent, none of my hi-fi buff friends can tell the difference between a solid-state single-ended Class-A amplifier and the design presented here.

The problem is simple

The basic problem can be stated quite simply. In order to make an economically viable amplifier, you have to operate the output stage in Class-B. This requires output devices that conduct on alternate signal polarities.

Figure 1 shows the Class B output stage in its simplest form. It consists of a pair of complementary transistors, operated as emitter followers.

The signal is applied simultaneously to both transistor bases. When the signal goes positive it turns the upper transistor on which then provides the output voltage across the load. Similarly on negative excursions the lower transistor conducts and provides signal to the load.

The problem is that the signal traverses the non-linear portion of both transistors' characteristics. The aptly

named cross-over distortion that results contains a large number of harmonics which are added to the output signal.

The standard cure is to apply a bias voltage to the transistors so that a small quiescent current flows in the output stage. This is only partially effective because no two transistors have identical characteristics.

The logical solution is to apply negative feedback around the circuit in copious quantity to reduce crossover to inaudibility. Unfortunately, the nature of this distortion is such that most unwanted harmonics are at high audio frequencies.

In order to produce a stable amplifier, the open-loop gain of the circuit must be rolled off early. Global feedback is then applied to produce a flat response from the amplifier. However, the amount of loop gain, the difference between open-loop and closed-loop gains, is insufficient to totally eliminate crossover distortion.

The net result is the standard thd/frequency graph of a typical amp, which is near the noise floor up to 1kHz then rises rapidly with frequency. A further complication is that distortion products, although reduced by feedback, can go through the amplifier again producing even higher frequency distortion.

Because of the lowered loop gain at high frequencies these are not reduced by global feedback as well as one would like.

Driving down distortion

So what can be done about this? For reasons of economy, not to mention electricity bills, Class A is out for anything more than a few watts. Effectively we're stuck with Class B and its attendant non-linearity.

One logical solution would be to split the amplifier into two, the first stage being a high quality voltage gain stage. The output driver stage could be designed with high open-loop gain and massive local feedback to reduce the gain to unity.

This works well in practice but for a better solution global feedback can be applied over both gain blocks to reduce all non linearities well below audibility.

Such a system uses nested feedback loops. However what is not realised is that the generic Class-B amplifier typified by Doug Self's articles already uses nested feedback although it doesn't, as it stands, do anything to reduce output-stage distortion. This is the function of the global feedback loop.

Figure 2a) shows a model of the standard generic Class B power ampli-

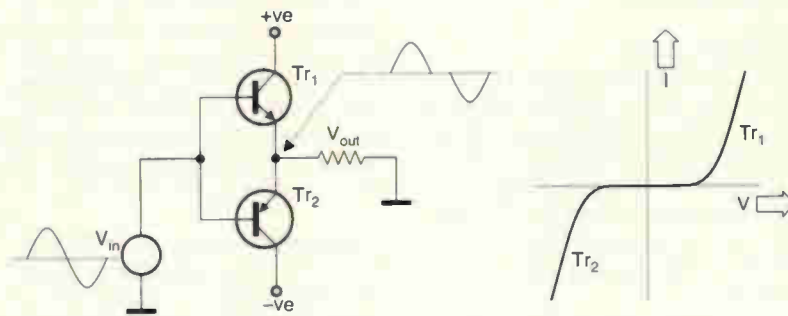


Fig. 1. The problem with the basic form of Class-B output stage is distortion caused at the cross-over point, where one transistor takes over from the other.

fier. It consists of three stages. First a transconductance amp turns input voltage signals into a current output. This current is then fed into the voltage amplifier stage, or VAS. Normally this is simply a common-emitter stage.

Herein lies one of the problems of the generic circuit. The low input impedance presented by the voltage

amplifier compromises open-loop gain by shunting the output impedance of the transconductance stage. This is probably only of academic interest in the standard amplifier, where no attempt is made to linearise output-stage distortion except by means of global feedback.

However for the circuit presented

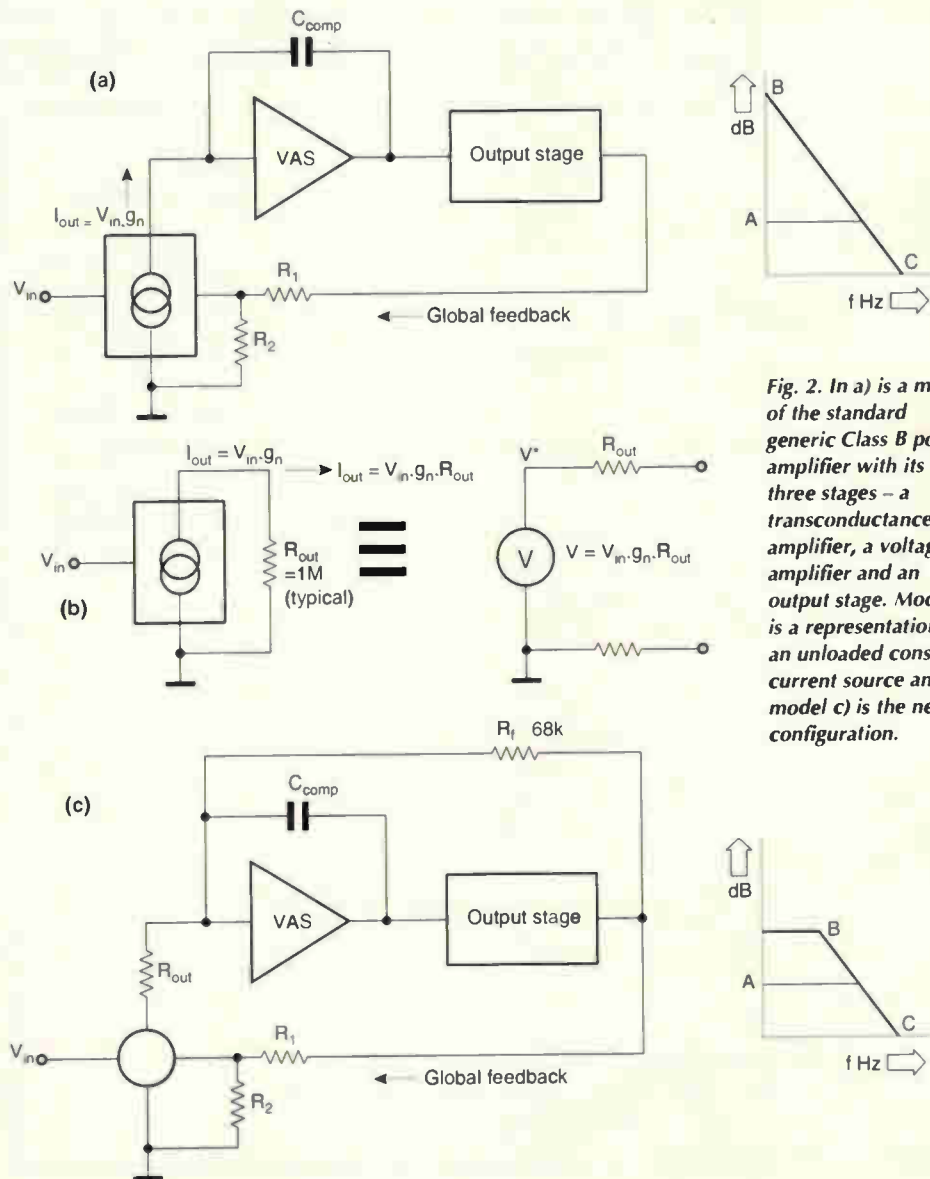


Fig. 2. In a) is a model of the standard generic Class B power amplifier with its three stages – a transconductance amplifier, a voltage amplifier and an output stage. Model b) is a representation of an unloaded constant current source and model c) is the new configuration.

here, and to make proper sense of the generic model VAS, input impedance needs to be high.

Stability issues

To ensure stability when the global feedback loop is closed, a capacitor, C_{comp} , is connected between collector and base. This turns the voltage amplifier into a shunt feedback amplifier or transresistance amplifier. Capacitor C_{comp} in conjunction with g_m also sets both the slew rate and unity gain frequency.

The voltage gain of this combined transconductance/transresistance stage is simply $g_m \times C_z$, where C_z is the impedance of C_{comp} at any given frequency. Thus, looking at the thumb line graph at the right of the figure the resulting open loop gain at VAS's output descends at 6dB/octave with increasing frequency.

Horizontal line A represents the gain after the global feedback has been applied via $R_{1/2}$. Note that because of the way that the open-loop gain decreases, progressively less feedback is applied at high frequencies. This is

just where the amplifier could use it most to suppress cross-over artifacts.

To illustrate what can be done to improve matters, taking a closer look at the transconductance stage is informative. Figure 2b) shows a representation of a constant current source. As shown, it is unloaded. The model requires just two pieces of information to specify its characteristics, namely g_m and R_{out} .

Transconductance stage design

In a perfect transconductance amplifier, the stage's output impedance, R_{out} , would be infinite. From Ohm's law you can deduce that a perfect unloaded transconductance amplifier would produce an infinite output voltage for any input signal. In practice, an R_{out} of $1M\Omega$ or more can be achieved with the rather simple circuit used in most amplifiers.

The other factor needed to specify the response of the transconductance stage is the transconductance, g_m . This is specified as the current output for a given voltage input and is usually expressed in mA/V. Finally we get to Fig. 2c). This shows the amplifier with

the voltage equivalent of the transconductance stage.

As before, the open-loop C_{comp} defines gain at high frequencies, the addition of R_f places the output stage within the local voltage amplifier's feedback loop. However notice the ratio of R_f to R_{out} . Provided that the input impedance of the voltage amplifier is high enough, the feedback factor approaches unity.

As far as the signal is concerned the voltage amplifier is acting almost as a unity-gain stage. Overall voltage gain is determined by g_m , R_f and is independent of the feedback factor of the circuit.

This implies massive local negative feedback which will reduce thd in the output to $<0.1\%$ – before global feedback is applied.

In addition to this, C_{comp} is still dominant at high frequencies so stability is assured. Indeed the prototype of this amplifier was used to drive a pair of Quad 63 electrostatics without stability problems. This is widely regarded as the most difficult load encountered in practice, proving the amplifier's unconditional stability.

Stability can only be ensured however by the inclusion of L_1 between the output and load.

Déjà vu?

Now if you have a sneaking suspicion that you've seen this circuit somewhere before, you're right. A circuit of this kind has already appeared in an article called 'Hot audio power' in the October 1995 issue.

In that article, the valve output stage was the transresistance amplifier, fed from a transconductance, solid-state phase-splitter. Consequently, the output valves operated as if they were cathode followers due to the local feedback applied.

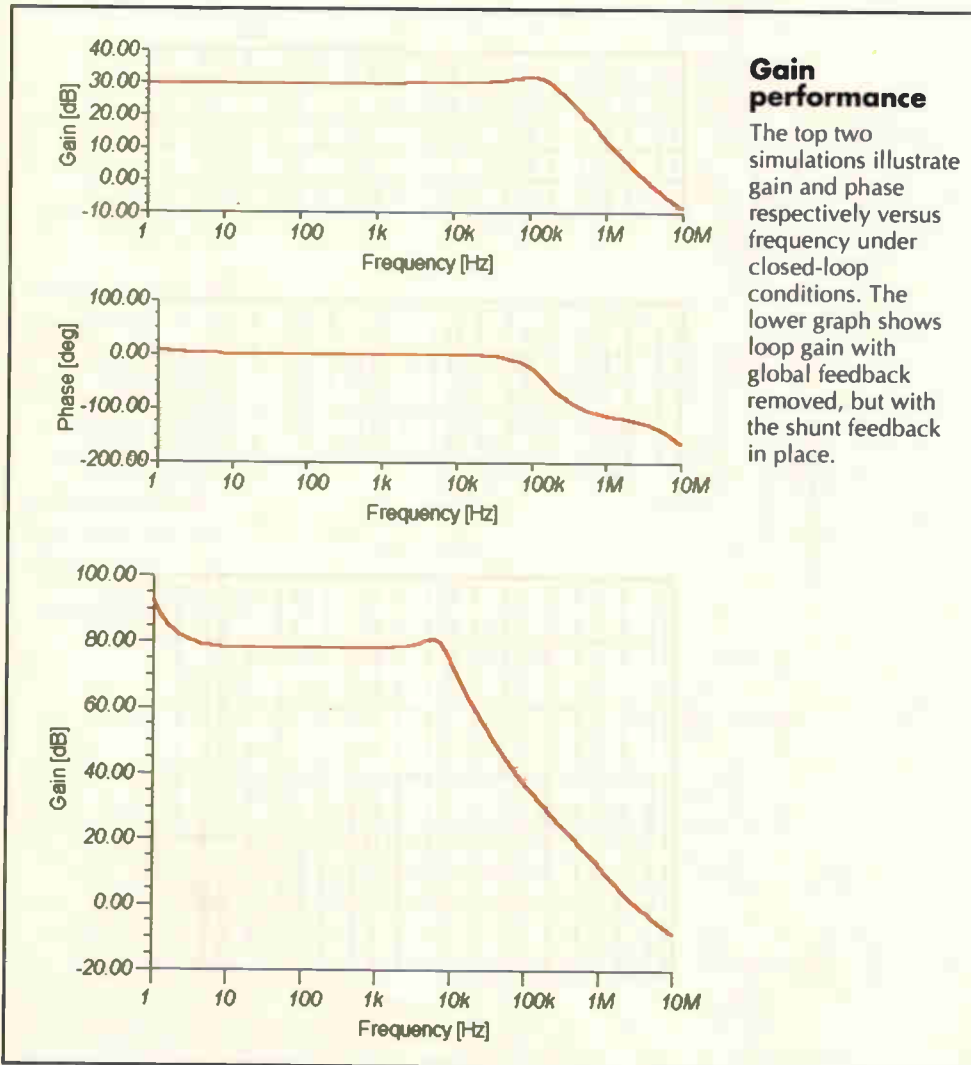
Cherry has suggested that the compensation capacitor C in Fig. 2a) could be connected between Tr_1 's base and the output. I've tried this but found stability problems. Instead I apply my local feedback via a resistor from the output stage to the base of Tr_1 , leaving C where it is. This produces a stable circuit. By manipulating g_m the global feedback still has plenty of loop gain to bite on to further linearise the circuit.

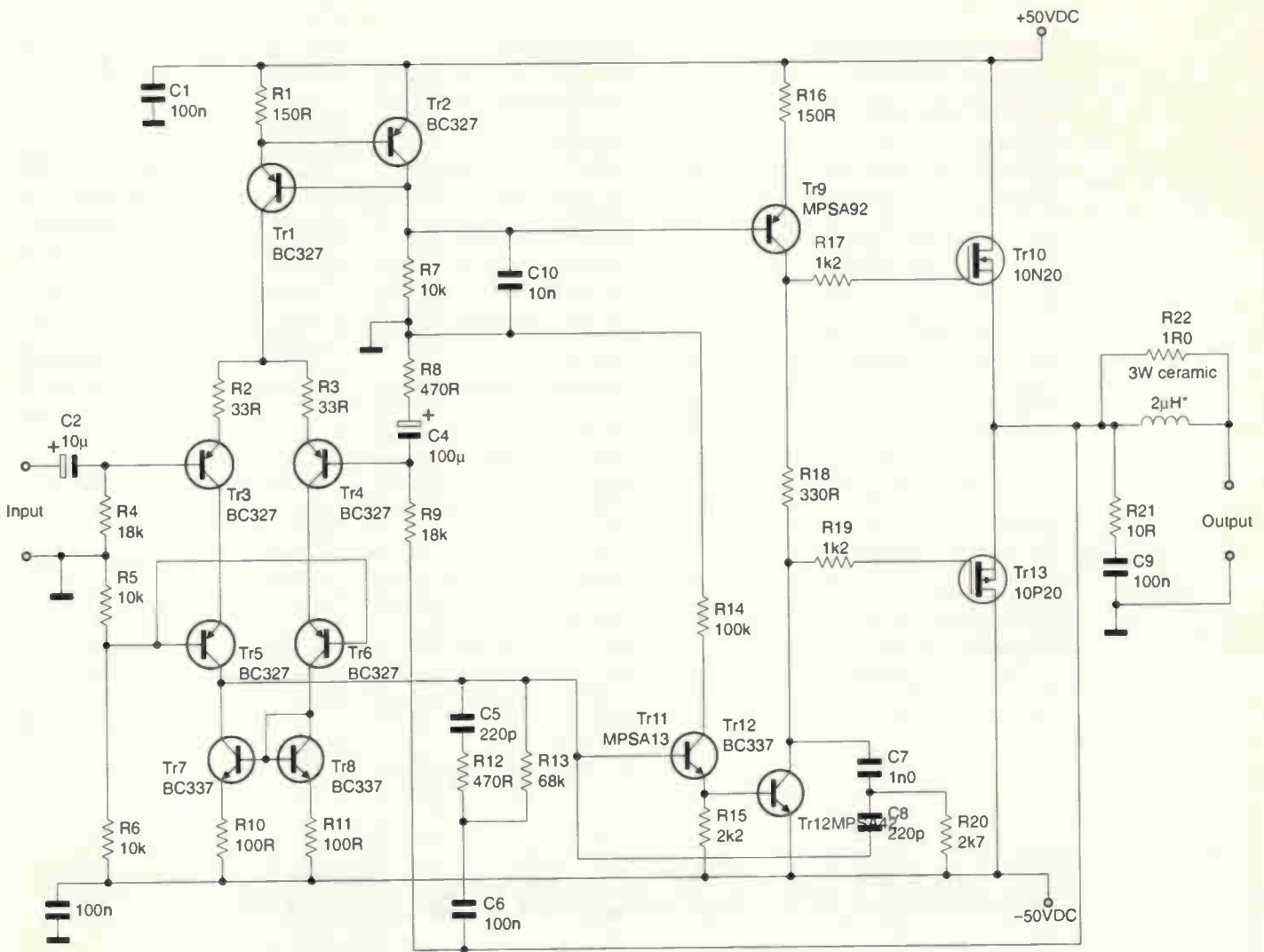
In amplifiers built this way, the crossover products are already greatly suppressed by local feedback and then further reduced by global feedback.

As I mentioned earlier no one who has heard this amplifier can tell the difference between it and a Class-A alternative. This great improvement has been wrought by the addition of a few components and a little re-jigging of the basic Class-B design.

Gain performance

The top two simulations illustrate gain and phase respectively versus frequency under closed-loop conditions. The lower graph shows loop gain with global feedback removed, but with the shunt feedback in place.





Circuit details

The circuit schematic is shown in Fig. 3. It consists of two sub-circuits, a transconductance amplifier and the VAS/output stage.

A fundamental requirement of the trans-shunt circuit is that the input voltage signal should be converted into a current to drive the transresistance output stage. Here *Tr*₁ and *Tr*₂ together with *R*₁ and *R*₇ form a conventional constant current source for the input transistors *Tr*₃ and *Tr*₄. These are operated as a differential pair. Input signals are coupled into *Tr*₃'s base via the dc blocking capacitor *C*₂. The input impedance is set by the value of *R*₄ that also refers *Tr*₃'s base to 0V.

To produce the high output impedance required for a transconductance stage, *Tr*₅ and *Tr*₆ are used cascode with *Tr*₃ and *Tr*₄. Emitter degeneration is used, *R*₁₀ and *R*₁₁, to linearise the stage. Transistors *Tr*₅ and *Tr*₆ are biased by the voltage divider *R*₅ and *R*₆.

The output impedance at *Tr*₅'s collector is in the order of a megaohm or so. This is sufficient to reduce the distortion of the driver/output stage to low

levels when shunt feedback has been applied.

To prevent the high output impedance from being compromised a current mirror *Tr*₇ and *Tr*₈ are used. High gain n-p-n transistors are used here, emitter degeneration being provided by *R*₁₀ and *R*₁₁.

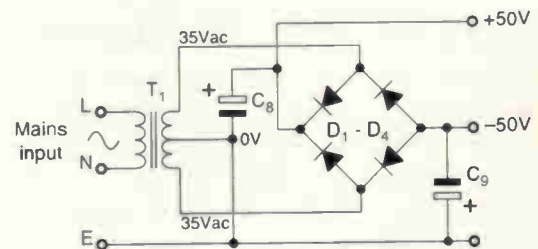
Driver stage design

Turning to the driver stage, in order to apply a large amount of shunt feedback around the output stage, the driver needs a high input impedance. Preferably this should be infinite.

The idea that at single common-emitter stage can provide this is a fallacy. The typical driver transistor has a low *H*_{ie}, leading to input impedances in the range of a few kilo-ohms. This shunts directly across the transconductance amplifier's output and effectively reduces the feedback factor of the circuit.

To avoid this a Darlington transistor, *Tr*₁₁ is used as an emitter follower buffer. The voltage amplifier stage proper is *Tr*₁₂ a conventional common-emitter circuit.

Applying masses of feedback around



the output stage is all well and good provided that the voltage amplifier has sufficient open-loop gain to make it viable. Single pole compensation could be used but to maintain maximum loop gain through the audio band double pole compensation is better. Doug Self described this technique in a previous issue¹.

Initially double-pole compensation starts to roll-off at 12db/octave. At hf however it reverts to a 6dB/octave roll-off to ensure stability. In this design the double pole network consists of *C*₇, *C*₈ and *R*₂₀.

Component values have been chosen so that *R*₂₀ doesn't load the voltage amplifier collector circuit unduly within the audio band.

Fig. 3. 100W power amplifier and its conventional power supply. Unusual features are the feedback path via *C*₆ and the network comprising *C*_{7,8} and *R*₂₀.

Choosing an output stage

The choice of output stage is always a fraught one for designers. Several possibilities present themselves.

During development of this design I have used pure complementary, quasi complementary and V-fet output stages. Because of the large amount of negative feedback applied I haven't noticed any sonic differences worth talking about. But a 100W amp is not a project for the squeamish!

Considerations of ruggedness and device longevity are paramount. In the end I chose V-fets for the output stage. Suitable n and p-type devices are readily available.

Furthermore the stage is immune to thermal runaway due to the Vfet's negative temperature coefficient and will shrug off a short circuit load. Don't try this one with bipolar transistors.

Staying with the output stage stability compensation for driving reactive loads is given by the network comprising C_9 and R_{21} and the output inductor L_1 . This latter component comprises 15 turns of 18 SWG enamelled wire wound around a 3W wire-wound resistor, R_{22} .

As described earlier local shunt feedback is applied around both the VAS

and output stage via R_{13} .

Capacitor C_6 prevents compromise of the amp's dc conditions. Components R_{12} and C_5 ensure hf stability and introduce a little more feedback at the high frequency end of the audio spectrum. This shunt feedback network reduces the distortion to below 0.1% before global feedback is applied.

Input impedance at Tr_9 's base is also reduced by this feedback to less than 100 Ω . Thus the trans-shunt circuit is returned by feeding the current drive from Tr_5 directly into Tr_{11} 's base.

Global feedback is applied through R_9 to Tr_4 's base. In conjunction with R_8 , R_9 sets the voltage gain and C_4 reduces the dc gain to unity while passing ac signals. This component also sets the -3db point of the amplifier at approximately 3.5Hz

Finally the power supply. This is entirely conventional in design. As the current drains of the various stages are defined with constant-current sources, the amplifier can operate from $\pm 20V$ to $\pm 50V$ supplies without modification. These supplies roughly represent power output ratings from 20 to 100W continuous into 8 Ω loads.

Heat-sink size must of course depend on the amount of power required. For

maximum, output a minimum 1°C/W per channel sink is recommended.

Putting it together

As far as implementation is concerned, the usual rules of good layout should be adhered to. Keep input wiring away from the output and use screened lead for the input. The easiest way to set the quiescent current is as follows.

Temporarily solder a 100 Ω resistor between Tr_{10} and Tr_{13} 's drains and the supply rails. Don't attach a load to the amplifier yet. Power up and check that the output is within 50mV of 0V. If there are any wiring problems at this stage, the 100 Ω resistors will be the only casualties!

Switch off, remove the 100 Ω resistors and reconnect the drains to the supply. The amplifier is ready to use.

In conclusion I suggest that this amplifier, with its low distortion and low power requirements, will fulfil the needs of most audiophiles. Having had mine operating for over six months, I haven't felt the desire to change it for a commercial model, regardless of price. ■

Reference

1. Self, Douglas, 'Distortion in power amps?' *Electronics World*, Feb 1994

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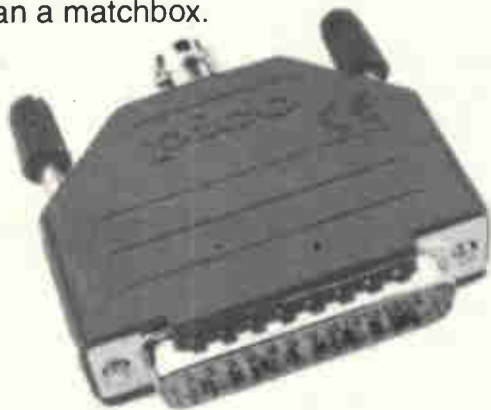
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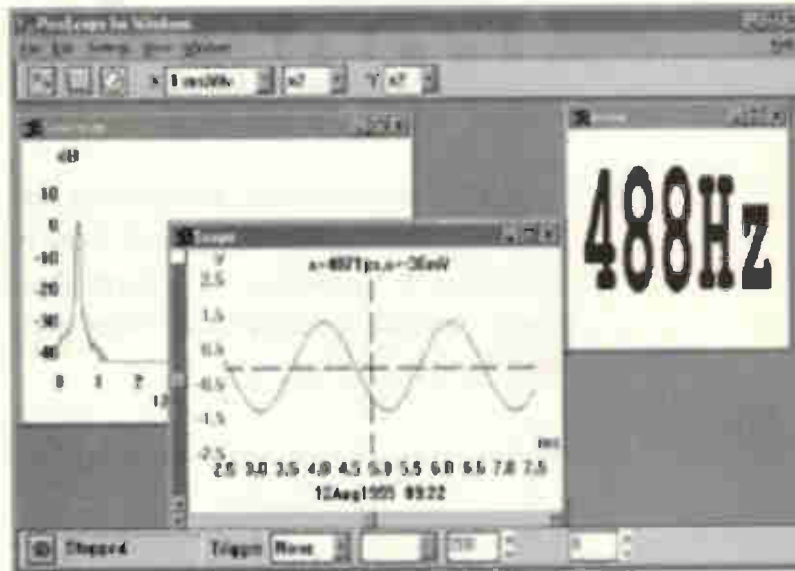
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Become a trapezium expert

There's any number of circuits for generating square, sine and triangular waveforms, but you rarely see anything on producing trapezium waveforms. Here, Anthony Smith not only explains how to make trapezium waveforms, but also reveals why they can be so useful.

You needn't search too hard if you want to generate a sine, square, triangular or sawtooth waveform: textbooks and cookbooks have all sorts of ideas for generating them. However, whether you consider it to be a clipped triangle or a slew-limited pulse, the trapezium wave seems to get scant mention in much of the literature. Applications for trapezium waveforms are surprisingly varied. I first encountered this versatile waveshape when working on Highway-Addressable Remote Transducer, or HART, communications systems. An outline of HART is presented in a separate panel.

Motor and power control

Trapezium waveforms find applications in active device testing, where they can be used to model real-world digital signals having finite rise and fall times.

Since the trapezium can form a good approximation to a sine wave, it is also used extensively in motor controllers for three-phase systems, induction motors, etc. Here, the relative ease with which the amplitude and frequency of the trapezium can be controlled makes it an attractive alternative to sine wave control.

As well as being used to provide drive for motors, the

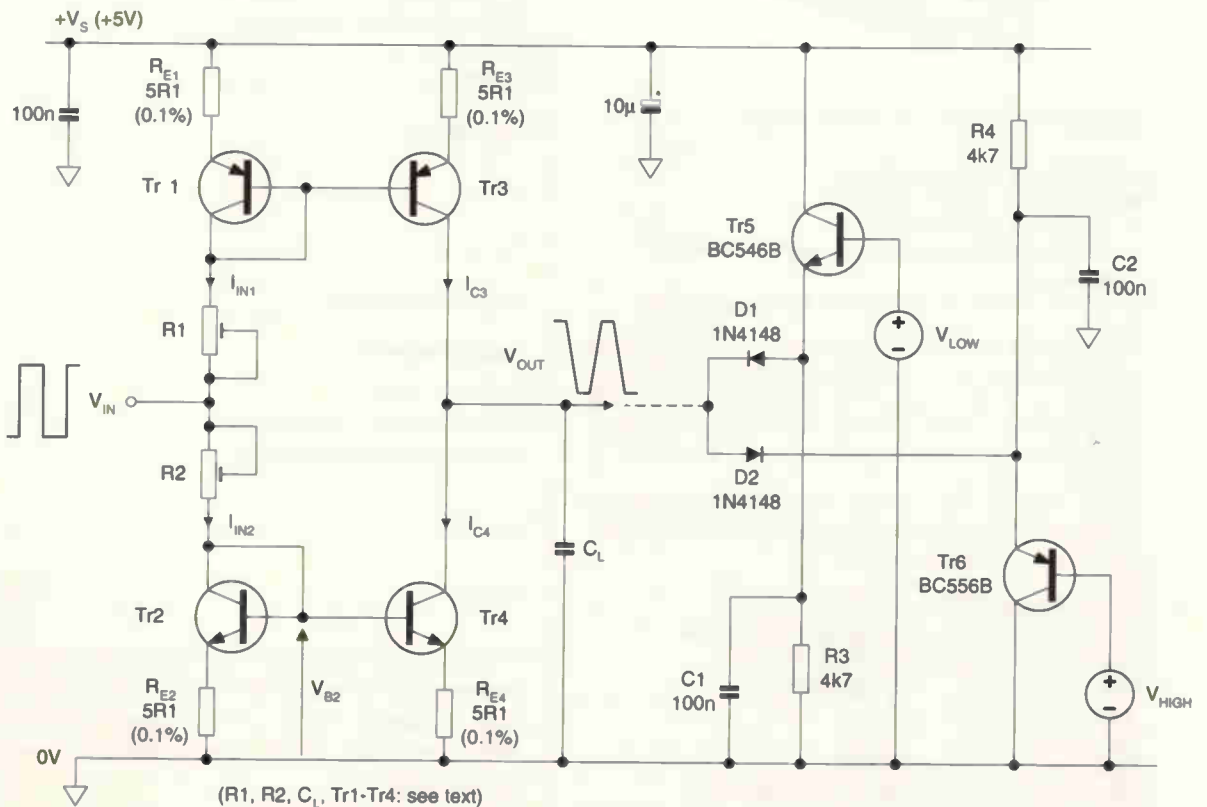


Fig. 1. Switched current mirrors and a simple diode clamp define the trapezium on C_L.

trapezium wave may also be used to control a motor's velocity, a typical example being the print-head drive motor of a dot matrix-printer.

Known as 'trajectory profile generation' and 'velocity contouring', such methods use the trapezium shape as a control signal. The rising and falling edges determine the acceleration and deceleration of the motor.

A separate panel entitled 'Trapezium waveform parameters' shows how the RMS value of a trapezium voltage may be varied from a minimum of $V_p/\sqrt{3}$ to a maximum of V_p simply by altering the slope of the rising and falling edges. Consequently, the trapezium lends itself to RMS regulation techniques, whereby the power to a load may be controlled either by adjusting the 'crest' width, t_c , while keeping the rise and fall times constant, or by adjusting only the slope of the edges. For systems where the load parameters are known and constant, the power may be controlled by sampling the peak load voltage and using pulse-width modulation to adjust the relative width of the trapezium crest.

Rings a bell...

The trapezium waveshape also finds applications in telecommunications systems such as dual-tone multi-frequency, or DTMF, telephone dialling circuits. Here it is used to approximate a sinusoid.

Dual-tone multi-frequency dialling signals make use of two simultaneous sinusoidal tones of different frequencies to represent each key on a telephone keypad. To ensure compatibility with DTMF receivers, the tones must have accurately defined frequencies. This may be achieved by generating low-pass filtered trapezium signals that are frequency locked to a crystal reference.

The signal that rings the telephone bell may also be trapezium shaped, or may be a pseudo-sinusoid formed by low-pass filtering a trapezium-shaped waveform. The Harris/Intersil HC55171 ringing Subscriber-Line Interface Circuit, or SLIC, for example, can produce sinusoidal or trapezium-shaped ringing signals.

Bellcore ringing specifications set limits on the ringing signal characteristics, particularly the crest factor, which is the ratio of peak voltage to RMS voltage, and the minimum RMS ringing voltage. Bellcore specifies a crest factor range of 1.2 to 1.6.

It can be deduced from the separate panel that the trapezium's crest factor ranges from unity to $\sqrt{3}$, i.e. 1.732. So by adjusting the trapezium's shape to provide a crest factor less than that of a sinusoid, which is 1.414, the RMS voltage available to the handset can be maximised. This results in more ringing power and an increase in the allowable loop length between the handset and the SLIC.

Slew-rate control

In high-speed datacomms applications, such as Low-Voltage Differential Signalling, or LVDS, where data rates up to 400Mbits/s are feasible, crosstalk and EMI are of particular concern.

Devices such as Texas Instruments' SN65LVDS31 LVDS drivers¹ allow the user to minimise these effects by connecting a capacitor across the differential output. This reduces the slew rate of the output pulses, resulting in a trapezium signal shape, thereby minimising the harmonic content of the transmitted signal.

Slew-rate limiting techniques are not new. They have been used to good effect in other devices such as National Semiconductor's DS3662 high-speed trapezium-waveform

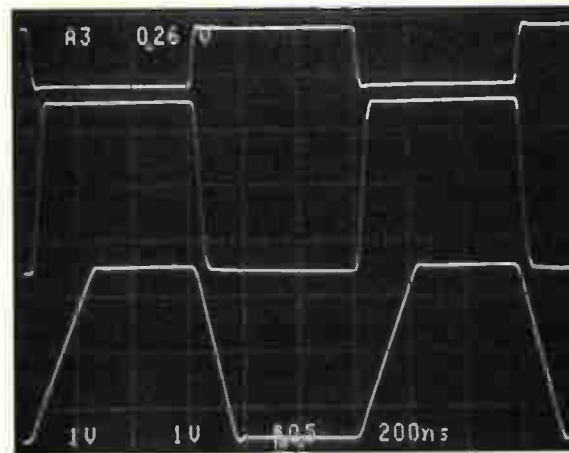
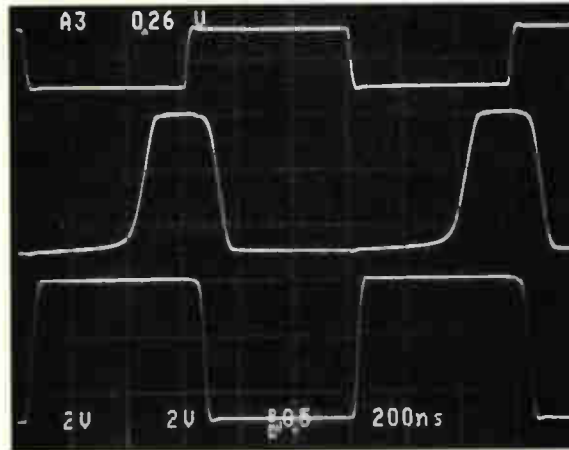


Fig. 2. Switched current mirror – unclamped response. The middle and bottom traces compare V_{OUT} using general purpose transistors and fast-switching transistors respectively, both at 2V/div. The top trace shows V_{IN} displayed at 5V/div, and the time-base was set to 200ns/div.

Fig. 3. Switched current mirror, clamped response using general-purpose transistors. The middle trace is V_{OUT} with C_L at 10pF and the bottom trace is V_{OUT} with a C_L of 230pF. Both are at 1V/div. The top trace is V_{IN} at 5V/div. The time-base setting was 200ns/div.

bus transceiver. Each of the DS3662's quad drivers generates a precise trapezium waveform having rise and fall times of 15ns.

This slew-rate limiting reduces noise coupling to adjacent bus lines: when used with the low-pass filtering inherent in the DS3662's receivers, the performance in terms of data rate *versus* line length can be an order of magnitude better than that achieved with other transceivers that do not have slew rate limiting².

Recently, new devices have appeared that offer pin-selectable slew-rate control. Linear Technology's LTC1535 isolated RS485 transceiver, for example, has a 'slow slew-rate select' pin. Via this pin, typical driver transition times can be increased from 20ns to around 500ns to

Trapezium therapy

The trapezium wave can benefit not only electrical systems, but neurological ones, too. Electrical stimulation has been used in nerve therapy where trapezium-shaped pulses are applied to a damaged nerve by way of electrodes attached to the patient's skin³. Having a frequency range of 60 to 250Hz and an RMS voltage of around 20V, the pulses can be used to stimulate damaged nerve tissue back to its normal, healthy function. Although it is not completely understood why pulses of a trapezium shape are so effective in this form of treatment, the technique can return a damaged nerve to near normal condition and is particularly successful in cases where nerve damage has resulted in hearing loss.

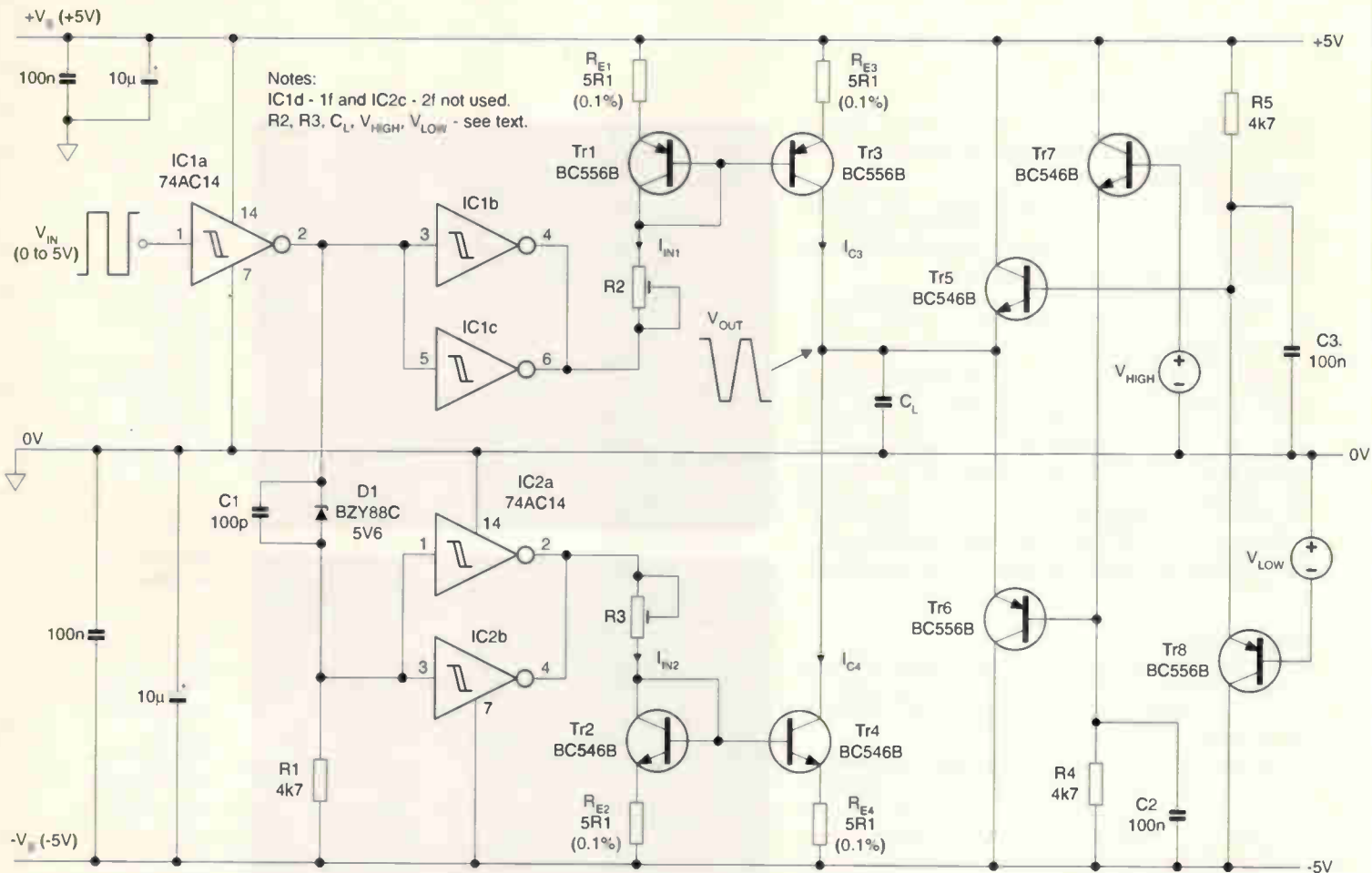


Fig. 4. Bipolar trapezium-waveform generator with improved clamping circuit.

minimise EMI and line reflections.

There is, of course, a price to be paid in terms of reduced bit rate and an increase in supply current during the slow slew-rate edges.

Practical circuits

What, then, is the best way to produce a trapezium waveform?

For general test purposes I needed to generate a trapezium waveform whose positive and negative peaks could be varied over a range of at least ±5V. I needed a broad frequency range – from a few hertz up to at least several hundred kilohertz – and the positive and negative slopes had to be independently adjustable, with an upper limit of at least ±10V/μs.

The need for fast slewing prompted me to discount d-to-a converters and digitally-controlled resistor networks: I wanted a solution that would provide inherently smooth, linear ramping between peaks – a desire that led, inevitably, to the charge and discharge of a capacitor.

A simple way of achieving this is shown in Fig. 1. Here, switched current mirrors formed by Tr₁ with Tr₃ and Tr₂ with Tr₄ provide charge and discharge currents for load capacitor C_L.

Simple diode clamping, comprising D₁, D₂, V_{LOW} and V_{HIGH}, buffered by emitter followers Tr₅ and Tr₆, defines the upper and lower peaks of the trapezium. A rectangular input signal, V_{IN}, that swings from 0V to +V_S, determines the current injected into each mirror.

For example, if you assume the n-p-n mirror is ideal, i.e. Tr₂ is perfectly matched to Tr₄, then the discharge current, I_{C4}, is given by,

$$I_{C4} = I_{IN2} = \frac{+V_S - V_{BE}}{R_2}$$

Clearly, the charge and discharge currents, I_{C3} and I_{C4}, and hence the rising and falling edges of the trapezium can be controlled by adjusting R₁ and R₂.

In practice, any mismatch between the mirror transistors can be mitigated to some extent by emitter resistors R_{E1-4}. These also impart a degree of temperature stability and increase the output resistance of the mirror⁴.

Prototype performance

I decided to test the circuit using both general purpose transistors (p-n-p BC177B; n-p-n BC108B) and fast, switching transistors (p-n-p ZTX510; n-p-n ZTX314).

Static performance was reasonable considering that the devices were not matched in any way: the worst-case current matching error for the general purpose devices was around 130%, whereas for the Zetex 'ZTX' parts, the error was a more respectable 48%. However, it was the dynamic performance that sorted the men from the boys.

Figure 2 shows circuit response with the output unclamped, i.e. Tr₃ and Tr₄ allowed to saturate. Resistors R₁ and R₂ were set at 1kΩ, and C_L at 10pF, representing the probe capacitance.

The middle trace illustrates the saturation effects of the general purpose devices: the p-n-p transistor, Tr₃, takes around 100ns to come out of saturation, whereas the n-p-n part, Tr₄, takes around 400ns. The bottom trace shows the circuit response with the general purpose transistors replaced by the Zetex parts: response from saturation is now almost instantaneous.

Since $Tr_{1,4}$ form a complementary current mirror, output transition time will depend on the turn-on time delay of a given output transistor combined with the turn-off delay of the complementary device. In general, the turn-off delay of a bipolar transistor tends to be much greater than the turn-on delay. Turn-off delay is dominated by the 'storage time', during which the device is saturated and an excess of charge builds up in the base region⁵.

It follows that switching performance can be optimised by selecting transistors specifically designed to minimise the storage delay: refer again to the bottom trace of Fig. 2, where it takes around 70ns after V_{IN} rises before V_{OUT} starts to fall (the p-n-p ZTX510 has a maximum turn-off time of 90ns), and roughly 10ns after V_{IN} falls before V_{OUT} starts to rise (the n-p-n ZTX314 has a maximum turn-off time of 18ns).

Alternatively, it should be possible to minimise the turn-off delay using general purpose devices by keeping the transistors out of saturation. The middle trace of Fig. 3 again shows the output response using only the general purpose transistors; this time though, D_1 and D_2 are connected to C_L , and V_{LOW} and V_{HIGH} have been set to clamp V_{OUT} at +1V and +4V.

Notice how the output signal responds immediately to

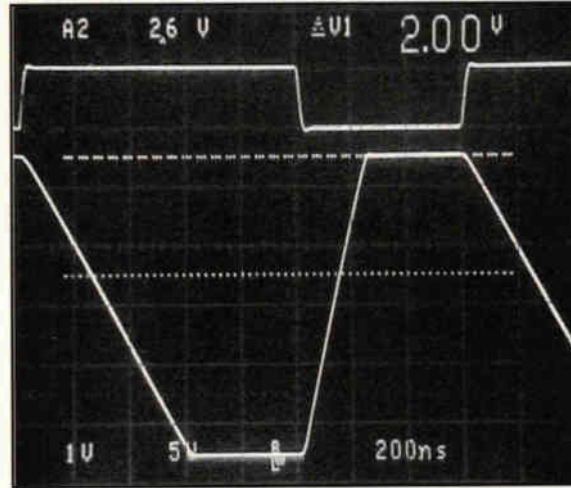


Fig. 5. Bipolar trapezium clamped at +2V and -3V with C_L at 110pF. The upper trace is V_{IN} at 5V/div. and at the bottom, V_{OUT} at 1V/div. The time-base setting was 200ns/div.

changes in V_{IN} - turn-off delay is almost completely absent.

The bottom trace of Fig. 3 shows the output response with C_L increased to around 230pF. The resulting trapezium slews cleanly between the clamp levels at a rate of about $\pm 12V/\mu s$.

Trapezium waveform parameters

This diagram shows an idealised case where a rectangular input signal results in a trapezium output waveform whose time and amplitude characteristics can be independently adjusted.

In this example, the positive peak, V_P , is greater than the negative peak, V_N , and the positive slope is steeper than the negative slope, such that the rise time, t_r , is shorter than the fall time, t_f . Clearly, the 'crest' time, t_c , can be adjusted by varying the width of the input pulse.

For the case of an amplitude-symmetrical waveform, i.e., when $|V_P|=|V_N|$ such that $(t_r+t_c+t_f)=T/2$, i.e., equal to half the period, the RMS value of the trapezium is given by,

$$V_{RMS} = V_P \sqrt{1 - 4 \frac{t_r + t_f}{3T}}$$

If you let $t_r=t_f$ and make $t_c=0$, we find that,

$$V_{RMS} = \frac{V_P}{\sqrt{3}}$$

which should be familiar as the RMS voltage of a triangle wave having peak amplitude V_P . On the other hand, if you make $t_r=t_f=0$, such that $t_c=T/2$, the RMS value reduces to $V_{RMS}=V_P$, representing the case of an ideal, symmetrical square wave.

An interesting case arises when $t_r=t_f=t_c=T/6$, resulting in,

$$V_{RMS} = V_P \sqrt{\frac{10}{18}}$$

which is very nearly equal to $V_P/\sqrt{2}$.

This shows that when the rise time, fall time and crest time of the trapezium are

all equal, the RMS value is very close to that of a sine wave with peak amplitude V_P .

Fourier series

If you consider an amplitude-symmetrical trapezium that is also time-symmetrical, i.e., when $t_r=t_f$, the waveform can be represented as a function with the following Fourier series,

$$F(t) = \frac{4V_P}{\pi K} \left\{ \begin{aligned} &\sin(K) \times \sin(\omega t) + \\ &\frac{1}{3^2} \times \sin(3K) \times \sin(3\omega t) + \\ &\frac{1}{5^2} \times \sin(5K) \times \sin(5\omega t) + \\ &\frac{1}{7^2} \times \sin(7K) \times \sin(7\omega t) + \dots \end{aligned} \right\}$$

where the higher terms have been omitted, and where $K=2\pi t_r/T$ and $\omega=2\pi/T$.

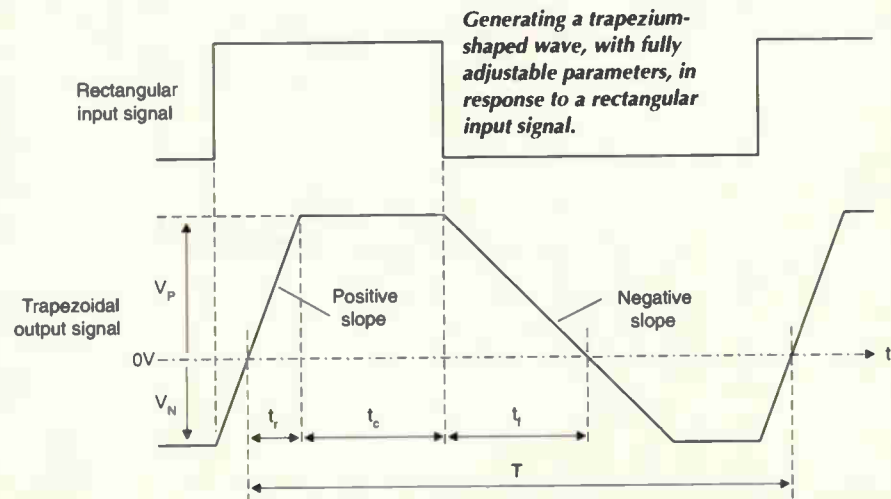
If you again consider the case where $t_r=t_f=T/6$, the series reduces to,

$$F(t) = \frac{12V_P}{\pi^2} \left\{ \begin{aligned} &\sin\left(\frac{\pi}{3}\right) \times \sin(\omega t) + \\ &\frac{1}{3^2} \times \sin(\pi) \times \sin(3\omega t) + \\ &\frac{1}{5^2} \times \sin\left(\frac{5\pi}{3}\right) \times \sin(5\omega t) + \\ &\frac{1}{7^2} \times \sin\left(\frac{7\pi}{3}\right) \times \sin(7\omega t) + \dots \end{aligned} \right\} \text{ which}$$

can be simplified to,

$$F(t) = V_P \left\{ \begin{aligned} &1.053 \times \sin(\omega t) + \\ &0 \times \sin(3\omega t) - 0.042 \times \sin(5\omega t) + \\ &0.021 \times \sin(7\omega t) + \dots \end{aligned} \right\}$$

Clearly, the $\sin(3\omega t)$ term drops out - as do all other harmonics divisible by 3 - and ignoring higher harmonics which are small in amplitude relative to the fundamental, the expression becomes $F(t)=1.053V_P \times \sin(\omega t)$, which is very nearly the same as a sine wave of amplitude V_P and frequency $f=1/T$.



Generating a trapezium-shaped wave, with fully adjustable parameters, in response to a rectangular input signal.

Bipolar swing

A bipolar version of the circuit is shown in Fig. 4 where C_1 , D_1 and R_1 form a crude level shifter which transfers IC_{1a} 's positive output pulses to $IC_{2a,b}$. Thus, R_3 is driven by a signal swinging from $-V_S$ to $0V$, which is in phase with the signal swinging from $0V$ to $+V_S$ that drives R_2 .

Complementary emitter followers $Tr_{5,6}$ form an improved clamp circuit: assuming the V_{BE} s of the n-p-ns are roughly equal to those of the p-n-ps, the upper and lower peaks of the trapezium will equal V_{HIGH} and V_{LOW} , respectively.

Figure 5 shows the high-frequency response, again using only general purpose transistors, and with C_L at $110pF$. Resistors R_2 and R_3 were adjusted to set IC_3 at $2.75mA$ and IC_4 at $0.92mA$, resulting in slopes of $+25V/\mu s$ and $-8.4V/\mu s$.

The clamp circuit performed well: the values of V_{HIGH} and V_{LOW} required to establish a swing from $-3V$ to $+2V$ differed by less than $50mV$ from the actual levels observed. However, the trapezium's peak-to-peak swing is limited by the base-emitter breakdown voltages of $Tr_{5,6}$ and must not exceed $V_{(BR)EBO} + V_{BE}$, where $V_{(BR)EBO}$ is the minimum breakdown voltage of either Tr_5 or Tr_6 .

For the devices shown, having a minimum $V_{(BR)EBO}$ of

around $5V$, the swing must be limited to less than $5.6V$ pk-pk. Nevertheless, provided this condition is met, V_{HIGH} and V_{LOW} can be adjusted in conjunction with R_2 and R_3 to create trapezium, triangular and sawtooth waveforms that lie anywhere between the $-5V$ and $+5V$ rails.

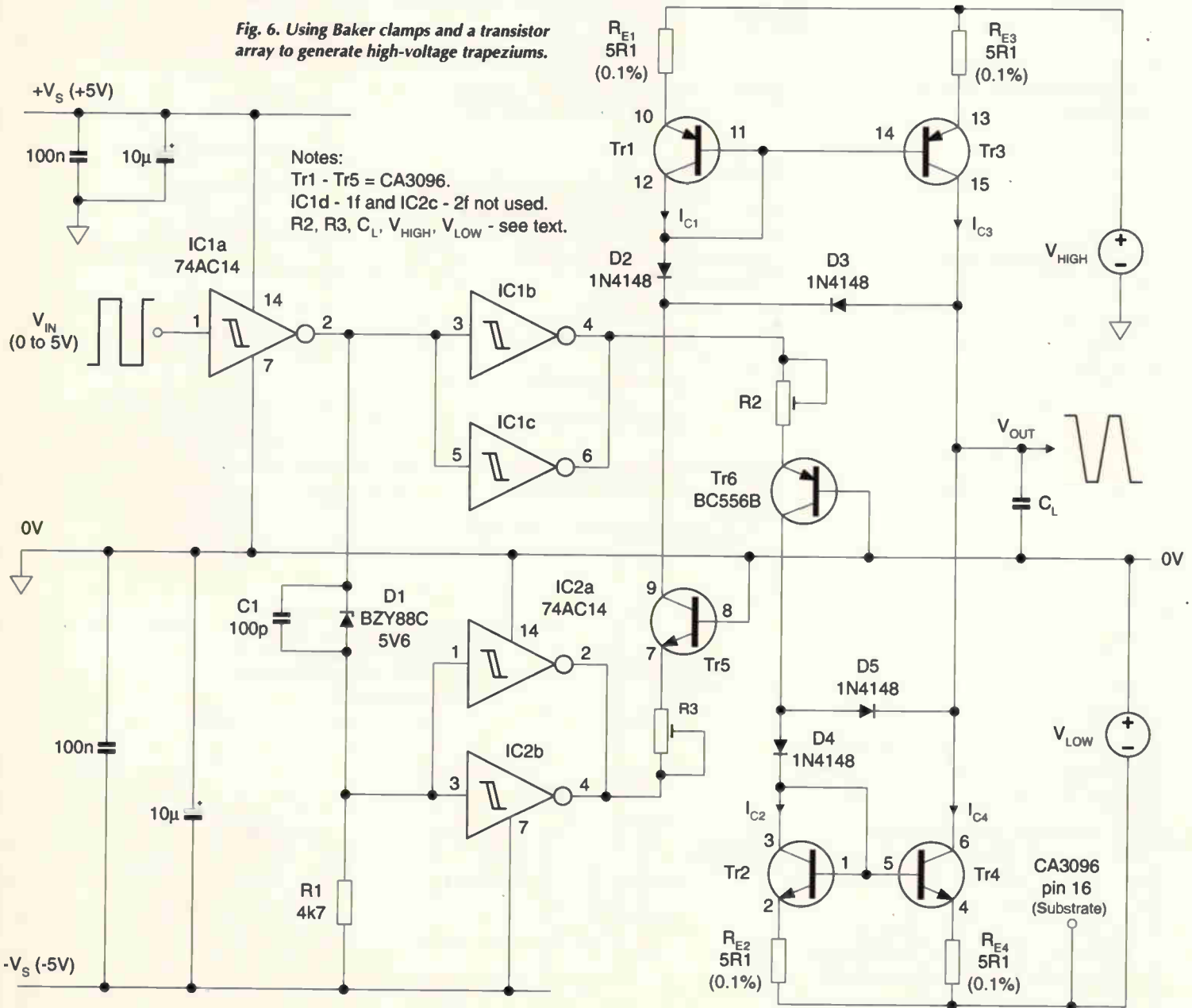
High voltage swings

A variation on the above theme is shown in Fig. 6, which is capable of generating high-voltage waveforms. The trapezium slopes are again controlled by R_2 and R_3 . These components determine the currents injected into the complementary mirrors via Tr_5 and Tr_6 .

Voltages V_{HIGH} and V_{LOW} are no longer used to clamp the waveform. Instead, they establish the upper and lower limits of the trapezium by varying the supply rails to the mirrors. However, for good high frequency response, it is still necessary to keep Tr_3 and Tr_4 out of saturation. The Baker clamps formed by $D_{2,3}$ and $D_{4,5}$ achieve this.

Baker clamps have been used extensively in switched-mode power supplies⁶. In such designs, the bipolar transistor that drives the magnetic components must be kept out of saturation in order to ensure fast switching and thus

Fig. 6. Using Baker clamps and a transistor array to generate high-voltage trapeziums.



maintain high efficiency.

In Fig. 6, D_2 and D_3 ensure that Tr_3 's collector-emitter voltage cannot become less than its V_{BE} drop. When Tr_5 is driven 'on' and sinks current through D_2 , the collector voltage of Tr_3 rises at a rate determined by I_{C3} and C_L until it is high enough to forward bias D_3 . At this point, D_3 's anode potential will equal that of D_2 - assuming similar drops across each diode - effectively clamping the collector of Tr_3 to the same potential as its base. Consequently, the trapezium peaks at a level just below V_{HIGH} , determined by Tr_3 's V_{BE} drop and the small drop across R_{E3} .

Diodes D_4 and D_5 provide a similar function with Tr_4 .

Transistor array

I decided to test the circuit using a Harris/Intersil CA3096 transistor array for Tr_{1-5} . This IC provides a degree of matching between the transistors in its n-p-n and p-n-p pairs - but beware that only the 'A' version offers guaranteed matching.

Static performance was impressive. With V_{HIGH}/V_{LOW} at $\pm 5V$, and I_{C1} and I_{C2} set to 2.0mA, currents I_{C3} and I_{C4} differed from this value by no more than 4%, although this error increased to 15% with V_{HIGH}/V_{LOW} at $\pm 15V$.

Figure 7 illustrates dynamic performance. Here, R_2 and R_3 were selected to yield slopes of $+25V/\mu s$ and $-50V/\mu s$ across a load capacitance of around 66pF. Levels V_{HIGH} and V_{LOW} were adjusted to swing the trapezium from $-10V$ to $+15V$.

For exactly the same conditions, Fig. 8 details the output response with the Baker clamps disabled, i.e. with D_3 and D_5 removed. You can see that the output signal now takes about 150-200ns to respond to transitions of V_{IN} , the sluggish response being due to saturation of Tr_3 and Tr_4 .

The peak-to-peak swing of the trapezium must be limited to a value less than the collector-emitter breakdown voltage, $V_{(BR)CEO}$, of Tr_3 and Tr_4 . Since the minimum $V_{(BR)CEO}$ for the CA3096 devices is 35V, a safe limit would be around 30V pk-pk. However, if high-voltage transistors were employed, such as MPSA44s for the n-p-ns, and MPSA92s for the p-n-ps, and provided D_3 , D_5 and C_L were also suitably rated, it should be possible to generate waveforms with amplitudes exceeding 100V pk-pk.

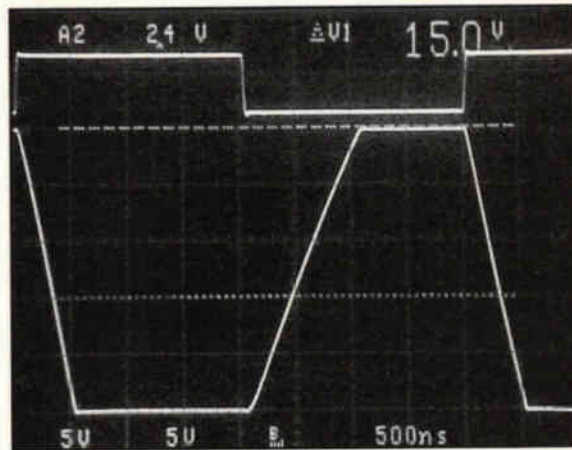


Fig. 7. Generating a $-10V/+15V$ trapezium with Baker clamps active. At the top is V_{IN} at 5V/div. and at the bottom is the output waveform at 5V/div. In this case, the time-base was 500ns/div.

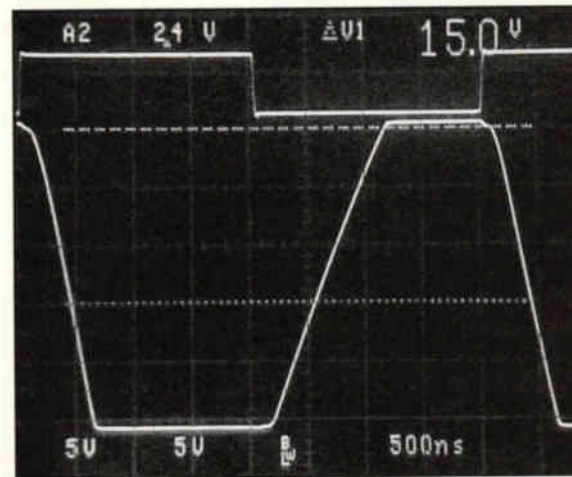


Fig. 8. Output response for same conditions as Fig. 7, but with Baker clamps removed. The upper trace is V_{IN} at 5V/div. and at the bottom, the output waveform at 5V/div. Time-base setting, 500ns/div.

Digital control of an improved mirror

The output resistance of the simple current mirror considered so far can be approximated by $R_O = V_A/I_C$, where V_A is the output transistor's Early voltage, and I_C is output current^{7,8}.

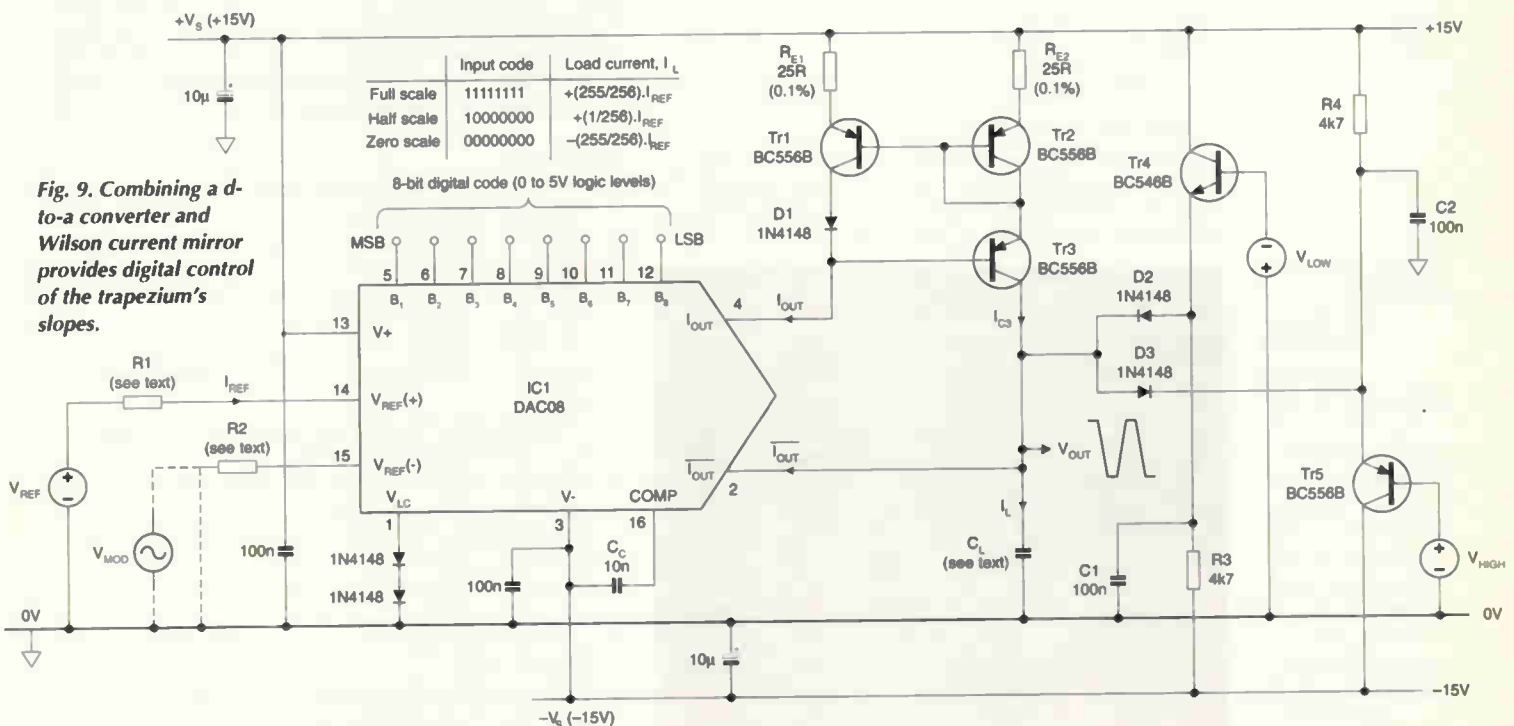


Fig. 9. Combining a d-to-a converter and Wilson current mirror provides digital control of the trapezium's slopes.

What is HART?

In the HART system, 'smart' devices, such as temperature transmitters and actuators, use the HART protocol to communicate with a central controller. They communicate by superimposing digital signals on a conventional 4-20mA current loop using phase-continuous frequency-shift keying at 1200Hz/2200Hz.

The protocol calls for a balanced AC signal with zero DC content that will not disturb the 4-20mA loop signal. A sinusoid would be the preferred waveshape, but generating a phase-continuous FSK sine wave is not so simple. Slew-limiting the edges of the square wave digital signal, however, can produce a trapezoid, relatively easily.

With suitable amplitude scaling, the resulting 'pseudo-sinusoid' is then coupled to the 4-20mA loop allowing a mass of information – such as tag numbers, range and span data, and diagnostics – to be communicated between Smart devices.

Rarely quoted on transistor data sheets, V_A might range from 50 to 100V, depending on the device used. Clearly, for I_C greater than 2mA, R_O will be less than 50kΩ – hardly an optimum value for a current source!

This relatively low output resistance causes slight changes in the output current as the voltage on C_L varies, an effect that manifests itself as a curvature in the trapezium's slopes. This non-linearity can just be discerned from Figs 7 & 8. Here, the slope of the rising edge, for example, varies by as much as 7% from its nominal, mid-slope value of +25V/μs.

Output resistance can be greatly increased by using a Wilson current mirror as in Fig. 9. The mirror's output current, I_{C3} , is controlled by the d-to-a converter. This circuit is a modified version of a circuit⁹ intended to provide precise, digital control of the bipolar current fed to an external load.

The high-speed, eight-bit d-to-a converter, IC_1 , generates complementary output currents I_{OUT} and $\overline{I_{OUT}}$. These actually flow into the device, the sum of which equals the full-scale current, I_{FS} , for all values of digital input code. The load current, I_L , flowing into C_L is given by,

$$I_L = I_{C3} - \overline{I_{OUT}}$$

Assuming the Wilson current mirror formed by $Tr_{1,3}$ to be ideal, then $I_{C3} = I_{OUT}$, such that,

$$I_L = I_{OUT} - \overline{I_{OUT}}$$

Since,

$$I_{OUT} + \overline{I_{OUT}} = I_{FS}$$

you will find that $I_L = 2I_{OUT} - I_{FS}$.

The magnitude of I_{OUT} depends on the input code and I_{REF} , where I_{REF} is the reference current given nominally by V_{REF}/R_1 . In particular, $I_{OUT} = (INPUTCODE/256) \times I_{REF}$. Since $I_{FS} = (255/256) \times I_{REF}$, you will find that,

$$I_L = I_{REF} \cdot \left\{ \frac{(2 \times INPUTCODE) - 255}{256} \right\}$$

Therefore, I_L is a bipolar current ranging from $-0.996I_{REF}$ to $+0.996I_{REF}$ as shown by the table in Fig. 9. By switching the input code between suitable levels, the voltage on C_L can be made to ramp up and down at precise rates, thereby allowing digital control of the trapezium slopes.

Diode D_1 with $Tr_{1,3}$ form a modified Wilson current mirror. The presence of Tr_3 increases the output resistance⁸ of the simple, two-transistor mirror by a factor of $\beta/2$, where β is the common-emitter current gain of the transistors used. Diode D_1 , which could be another p-n-p transistor, balances the V_{BE} of Tr_3 , forcing the collector potentials of Tr_1 and Tr_2 to be equal. This improves the matching of I_{C3} to I_{OUT} .

Linear response

With a total load capacitance of around 130pF – which includes C_L , probe capacitance, and the capacitance of Tr_3 , D_2 , D_3 and pin 2 of IC_1 – and with I_{REF} set to 2mA, I tested the circuit's response. I did this by switching the input value from 00000000₂ to 10101000₂ at about 160kHz. Levels V_{LOW} and V_{HIGH} were adjusted to clamp the trapezium's peaks at -7V and +10V, as in Fig. 10.

I found that the measured output current levels – and, hence, the trapezium slopes – differed slightly from calculated values, due, simply, to matching errors in the mirror. However, the rising and falling edges were perfectly linear, with no discernible 'curvature'.

Output swing must be constrained to prevent Tr_3 from saturating. It must be at least two V_{BE} s below $+V_S$. The negative swing is limited by IC_1 's negative voltage compliance¹⁰: this is around -10V with I_{REF} at 2mA and $-V_S$ at -15V.

Pulse-slope modulation

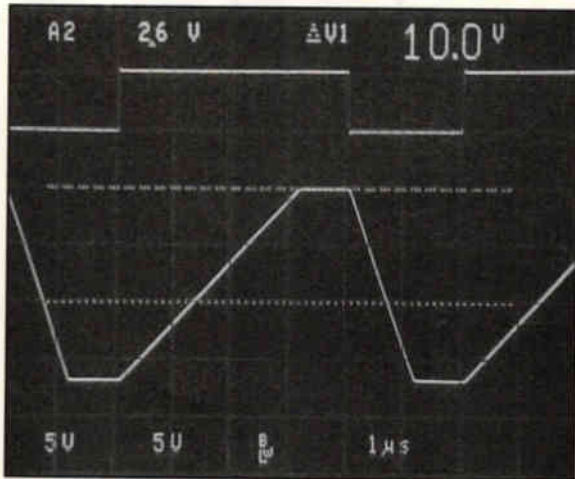
Usually, R_2 , which is connected to the inverting input of the DAC08's internal reference amplifier, is made equal to R_1 and tied to 0V. Alternatively, it may be connected to a variable reference voltage to trim I_{OUT} and $\overline{I_{OUT}}$.

However, by connecting R_2 to an alternating signal, the load current I_L – and, hence, the trapezium's slopes – can be modulated. For example, with R_2 fed by a low frequency sine wave, the trapezium's rising and falling edges became a blur as the dV/dt varied in proportion to the amplitude of the modulating sine wave. This fascinating effect suggests that the circuit could be used for 'pulse-slope modulation'. I'll leave it to you to think of an application for this! ■

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Fig. 10. Linear response of the d-to-a converter controlled trapezium generator. The top trace is the input signal applied to B₁, B₃ and B₅ displayed at 5V/div. All other bits are low. At the bottom you can see the output signal ramping from -7V to +10V, also at a setting of 5V/div. The time-base used was 1μs/div.



Calibrator for 4-20mA

After reeling at the price of a calibrator for 4 to 20mA loop interfaces, Darren Heywood decided to look into designing his own. He succeeded in making one that performs better than its commercial counterpart – and at a fraction of the cost.

The most important parameter of any instrument is its rate of output change with ambient temperature, hence if long term stability has been accomplished then all that remains is to calibrate to a known source.

I describe here a 4-20mA calibrator of the type widely used in process industries. Equivalent commercial 4-20mA calibrators with a price tag of

around £300 have a typical stability of $4\mu\text{A}/^\circ\text{C}$ at 4mA, best case.

My circuit has a staggering drift gradient of only $100\text{nA}/^\circ\text{C}$ at 4mA and $600\text{nA}/^\circ\text{C}$ at 20mA. This means it is some 40 times more stable than commercially available units.

Circuit overview

My unit runs from eight 1.5V standard AA batteries, i.e., 12V, the SG3524

PWM chip generates a 26V supply. This allows the unit to handle external load resistances up to $1\text{k}\Omega$.

Transistors Tr_1 and Tr_2 are essentially class C amplifiers that deliver a power pulse every half cycle. The supply can sustain 26V while sourcing some 50mA. The step up is accomplished without the need of a custom wound expensive transformer: instead two cheap easily available inductors are used.

The SG3524 runs at about 200kHz. Components Tr_1 and Tr_2 run very cool under normal conditions, which means step up conversion is very efficient – a

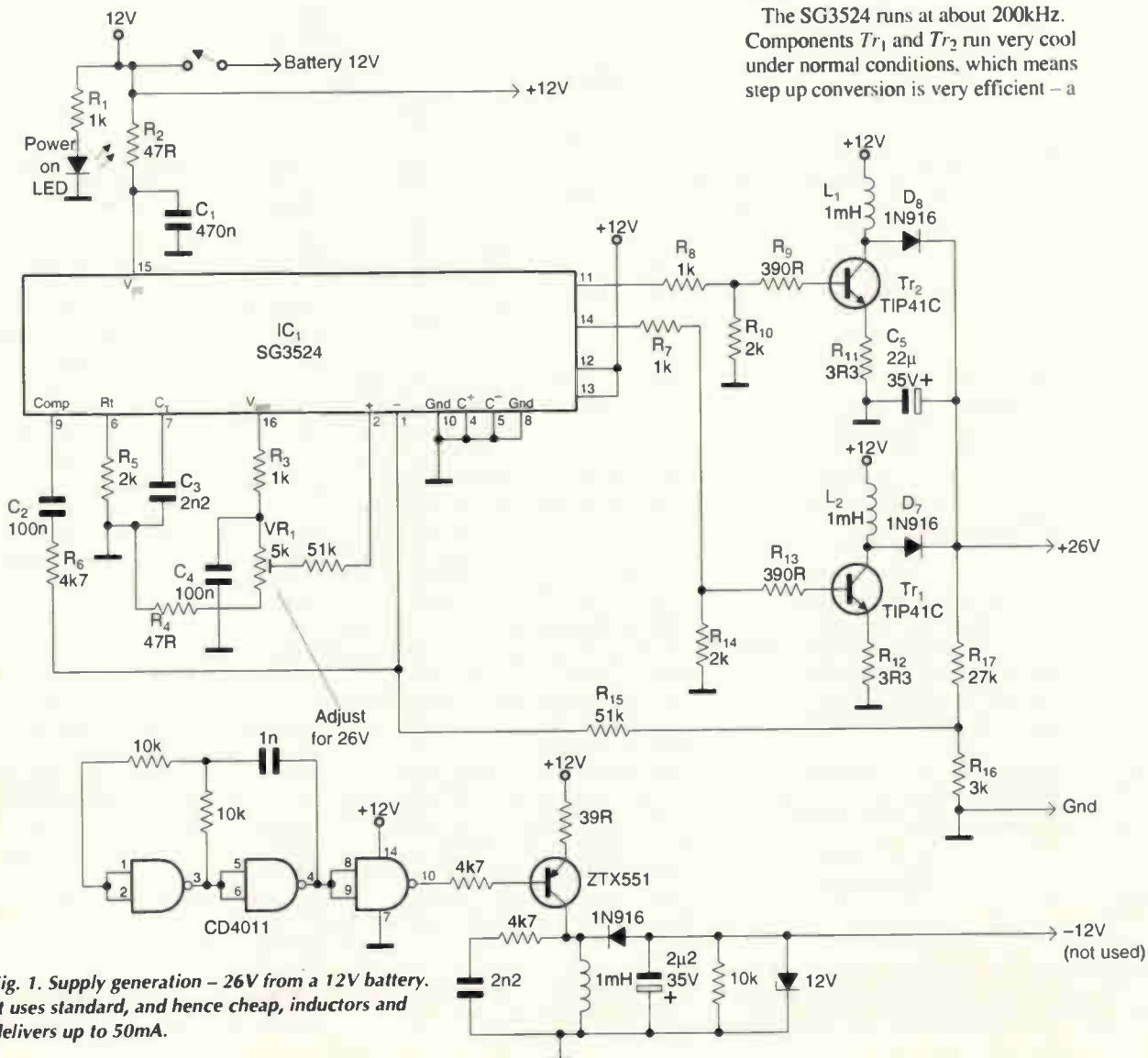


Fig. 1. Supply generation – 26V from a 12V battery. It uses standard, and hence cheap, inductors and delivers up to 50mA.

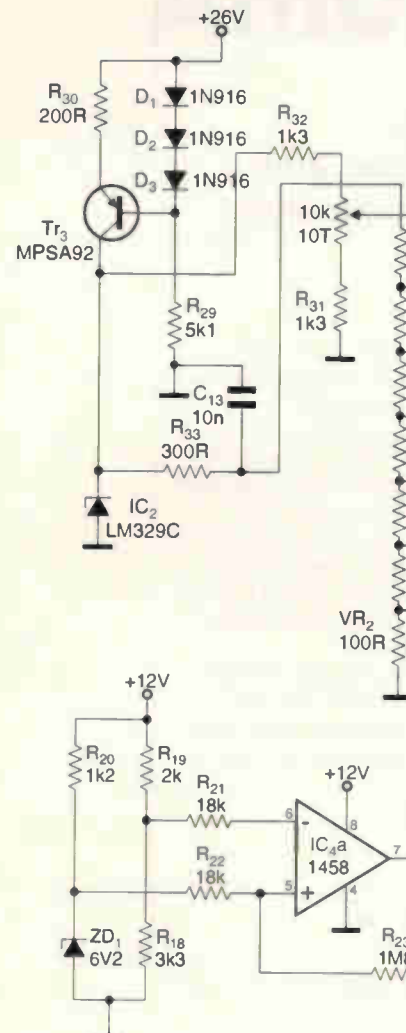


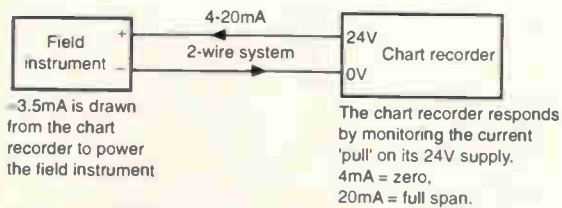
Fig. 3. Low battery monitor lights the LED when voltage falls to 10V.

pre-requisite for battery powered instruments.

The main control circuitry comprises an LM329C which is a 20 ppm/°C reference of 6.9V. Its output is stepped down via R₃₃ through R₃₉. You can select 4 to 20mA in steps of 20% increments or decrements.

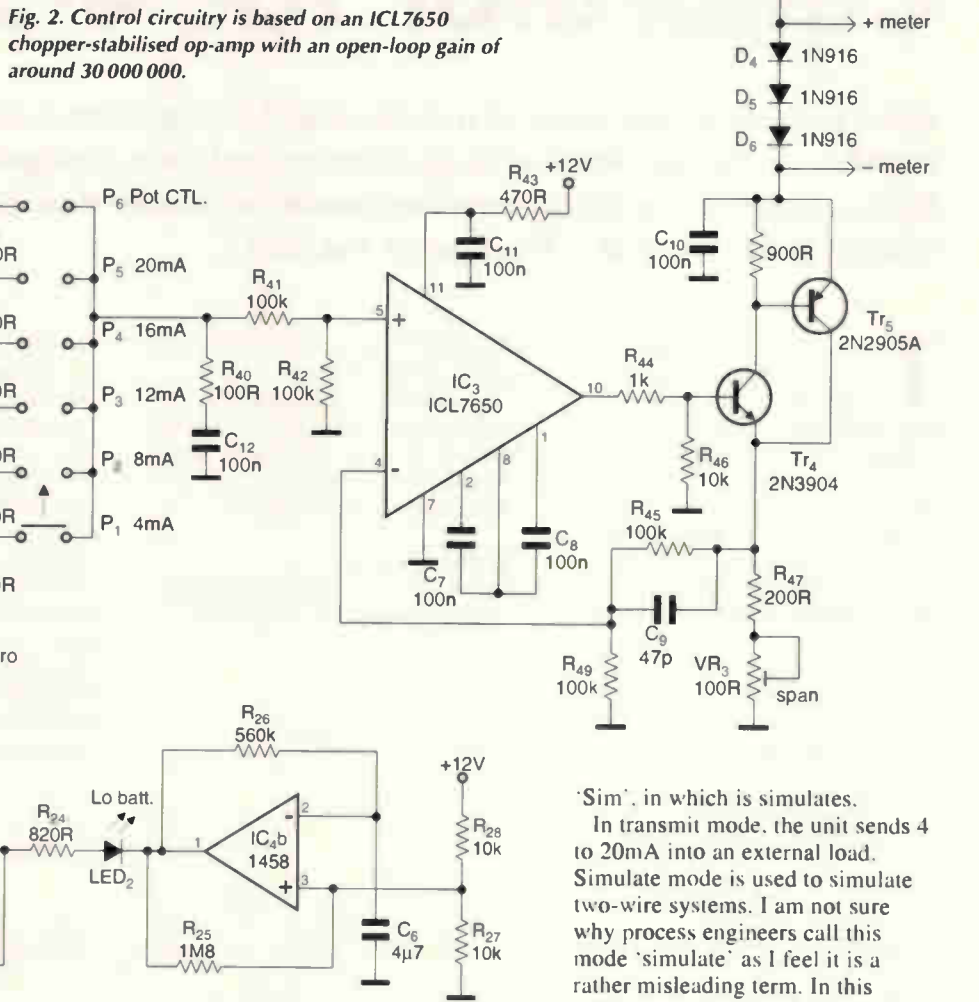
Position 6 allows you infinite resolution between 4 and 20mA with the aid of the ten-turn potentiometer.

The switched positions are fed into the excellent ICL7650 chopper-stabilised amplifier. This device has an open-loop gain of approximately thirty



The 4-20mA loop. Two-wire 4-20mA system use less cable than their four-wire counterparts and they give fewer ground-loop problems – but they pose more problems for the designer.

Fig. 2. Control circuitry is based on an ICL7650 chopper-stabilised op-amp with an open-loop gain of around 30 000 000.



'Sim', in which it simulates.

In transmit mode, the unit sends 4 to 20mA into an external load. Simulate mode is used to simulate two-wire systems. I am not sure why process engineers call this mode 'simulate' as I feel it is a rather misleading term. In this mode, my calibrator sinks a 4-20mA signal from the instrument under test's own 24V supply and the calibrator behaves or 'simulates' a two-wire 4-20mA field instrument. Current being drawn through the 24V supply is measured.

In the field, direct current measurement is possible by connecting a DMM across diodes D₄ through D₆ without disturbing or breaking the current loop.

Calibrating

When calibrating the unit, connect a 0.1% 15ppm 0.33W 250Ω resistor across the Tx terminals. With a 6½-digit bench multimeter switched to the volts range, measure the voltage across the 250Ω resistor. Currents of 4 to 20mA will correspond to a 1 to 5V drop across the 250Ω resistor. Switch to position 1 and adjust the zero potentiometer for 1V. Then switch to position 5 and adjust the span potentiometer for 5V.

Repeat this procedure until satisfactory results are obtained. ■

million, i.e. 150dB. This device retails for as little as £3.

Notice the floating offset correction capacitors C₇ and C₈. The only drawback IC₃ has is that its output impedance is quite high hence loading the output can gobble up much needed gain. To overcome this slight problem, Tr₄ and Tr₅ make up a localised current-feedback pair, providing a high input impedance.

During development, I included a buffer between IC₂ and IC₃. It totally spoilt the stability of the circuit and so I removed it. Resistors R₃₃ to R₃₉ inclusive should be 15ppm 0.1% types, but excellent stability is still possible with 50ppm 1% alternatives.

The unit is equipped with a low-battery indicator, IC₄. Should the battery fall below about 10V, LED 2 flashes on and off indicating that accuracy is no longer be guaranteed.

In use

The unit has two modes of operation, 'Tx', in which it transmits and

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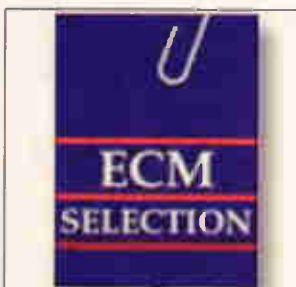
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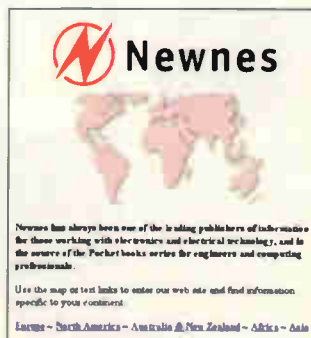


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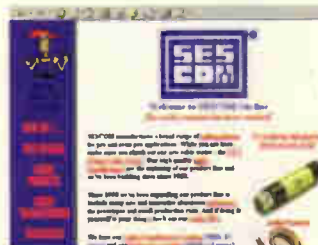
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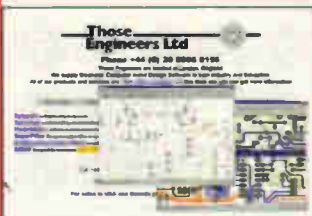
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Audio power analysis

I found Doug Self's article on audio power analysis in the December 1999 issue most interesting. It is always good to see quantitative data replace assumptions and conjecture.

While the analogue circuit described achieved the desired aim, it does have the stated disadvantage that multiple passes are required to obtain the required density functions. If the music is available in digital form – particularly if it is a digital data stream from a CD – there is a simple technique for obtaining the density function in one pass with 14-bit resolution for CDs. The procedure is as follows:

Initiate a block of memory sized 2 power 14 to all zeros. Smaller memory may be allocated by truncating the data.

For each data word, use the data value as a memory address offset and increment the word at that address.

The pseudo code would be as follows:

```
Variables:
MemSize = size of memory
           allocated
StartMem = address of first word
           in memory block
Offset = 14 bit value
; Initiate Memory
For x = 0 to MemSize
  ptr(StartMem+x) = 0 ;clear the
  memory block
Loop
Offset = CD data value (14bit)
;read in a data value from
the music CD
Increment(StartMem+Offset);
Increment the word at
address (StartMem+Offset)
Repeat Loop
```

When a run is complete each address will hold a value equal to the number of times a particular signal amplitude occurred. The stored values directly give the probability density function. The value at each address equates to the probability that a signal is in the range $(x, x+dx)$ where $dx=1$ bit.

Simple integration of the stored values gives the cumulative density function. Further code would be needed to check for overflow and to scale the stored values so that they all add up to equal one but this simple software procedure will provide the data required to produce density functions

in a single pass. Anyone who has a sound card or CD-ROM on his or her PCs would be able to use this technique.

Given that Doug has obtained quite satisfactory results using his analogue circuit, there is probably little incentive to implement this procedure, but it may be of some interest to other readers.

Darren Conway
Auckland
New Zealand

Doug replies:

I would like to thank Mr Conway for his kind words. The routine he gives is very close to that actually used in the DSP version I described in the original article. It is of course very much faster, but since it requires a digital-signal processor with a support platform, it is very much a 'maximum-hardware' solution, as opposed to the 'minimum hardware, maximum tedium' comparator approach.

Easily-bared ends

If, like me, you have spent half your life winding coils, toroids and so forth, scraping the enamel off 'magnet wire' is tedious chore. However, with a hand-held 12V PCB drill fitted with a 6-by-6mm grinding wheel, you can cut through even the toughest of varnish coatings and produce a perfect 'tail' for subsequent soldering.

RS sells a 12V drill (547-616) with collets up to 3mm diameter, and W160 grinding wheels (575-273/pack of ten). The latter are available individually from Farnell (700-9422).

CJD Catto
Cambridge

Photodiode sensing

I would not recommend readers use a tee attenuator in place of a high-value feedback resistor in a photodiode circuit such as Fig. 2 of the article 'Photodiode sensing', *Electronics World*, March 2000, page 212.

The purpose of using a high-value feedback resistor – 10, 100 or even 1000M Ω – is to produce a usable size output from a very small photodiode output, i.e. with the diode in very dim light. Provided the op-amp has very low current noise, the circuit gain can be increased

Can anyone throw any light on this?

With reference to Joe Pengelly's letter in the May issue, I would have thought that one could add a modern infra-red or visible light emitter and detector – i.e. LED and phototransistor – to any pickup headshell and contrive a stylus-carrier as well out of a redundant cartridge. Unlike many vinyl discs, I suspect that cylinders are not transparent to infra-red.

John Woodgate
Via e-mail

greatly by raising the value of the feedback resistor like this, without incurring a significant noise penalty.

With op-amps now available with bias currents of around a picoamp or less and very low current noise, the gain is thus limited only by practical values of feedback resistor, and the acceptable frequency response.

However, using a tee attenuator in the feedback network to simulate a high-value feedback resistor raises the noise gain of the circuit.

The easiest way to describe this is to consider the op-amp's input offset voltage V_{os} . Negative feedback drives the op-amp's output to provide a voltage across the attenuator's shunt leg of V_{os} . In Cyril's Fig. 2, the op-amp output will thus sit at 100 times V_{os} .

Now V_{os} is simply an unwanted input: it is just the dc component of the input 'noise'. Exactly the same argument applies to the ac noise at the op-amp's input: it too will be amplified by 40dB, rendering the extra gain to the signal of no use.

An alternative way of looking at it is to consider the star-delta transformation applied to the tee network. It gives the wanted 1000M Ω feedback resistor, plus a 20M Ω shunt resistor at the op-amp's output – of no consequence. It also gives a similar resistance shunting the op-amp's input – definitely not wanted!

You can find the subject simply but well treated in the data sheet for the LT1115 op-amp, from Linear Technology.

Ian Hickman
Waterlooville
Hants

In Defence of privatisation

Spitfire, Hurricane, Lancaster, Mosquito. The last time it really mattered, it was civilian innovation that delivered the goods.

Tanks, torpedoes, submarines, rockets, parachutes, planes, nuclear explosives, you name it; civilian innovation again, and all rejected by the establishment as 'of no military use'.

Even when the most brilliant of innovations come from right inside its own organisation, the natural response of the military is to reject it. Frank Whittle knew all about that.

As for 'security', virtually the whole of MoD's inventory is designed, built and maintained by private industry – including the bits so secret they don't even exist.

But all this is common knowledge, which makes Richard Wilson's editorial in the March issue so very odd. Anybody with any real concern for the quality of UK military equipment should be clamouring for DERA's privatisation.

And what has DERA done that so impresses Richard? What makes DERA, "so important to the UK"? Well, Richard first tells us about

This month's hot topic

We received a significant number of replies relating to domestic thermocouples. Apologies; there wasn't space to include all of them, but many thanks everyone that wrote in. Please note that only a Corgi-approved fitter is allowed to tamper with gas appliances. The 'bodge' mentioned below is potentially lethal. It is included only to make the writer's point.

I hate to contradict the editor of such an august publication, but the thermocouples used to control flame-failure devices on gas appliances are thermoelectric devices. If they were not, it would not be possible to bodge a failed appliance with an AA cell.

I do agree that the use of what is almost a lab-curiosity phenomenon to actually do something significant in the big chunky world is a cause for wonder.

Chris Bulman
Bedford

With regard to Graham Cox's letter in the May edition about domestic thermocouples, they are most definitely thermocouples rather than the capillaries or bimetallic systems that you would normally find in a thermostat. When you press the button to release the pilot gas valve, you are also closing the air gap between the solenoid and armature.

When the couple is hot enough, the tiny current generated is enough to hold the armature in place. The reason that a small signal can do this is that it only has to hold the valve open as you, the user have already opened it.

A number of manufacturers use a small brass fitting that intercepts this signal and routes it via an over-temperature switch. The Ideal Elan central

heating boiler is an example.

It is possible to test a valve using a D cell and a current limiting resistor. I won't go into details as we have come across systems that have been left being 'Tested' for some considerable length of time – most definitely not a good idea!

Gary Williams
Consultant Gas Engineers Ltd
Via e-mail

Thermocouples that work in the pilot flames of gas appliances really are thermocouples. Thermostats in fridges and boilers are not. Typical thermocouple voltage – excluding platinum-rhodium – is $40\mu\text{V/K}$.

Running at just below red heat (400°C) you get about 16mV . But this is a very low impedance generator, so you can get 10mA or so into a 1Ω load.

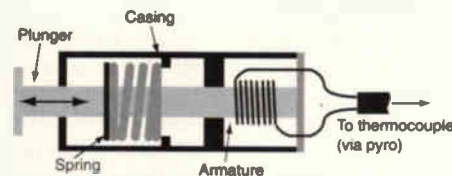
In a gas valve, this flows through a moving-coil motor – like a miniature loudspeaker – that holds open a diaphragm valve. You actually open it by twiddling the control knob when you light the pilot flame. This lets the incoming gas pressure open a bigger diaphragm valve which, in turn, lets the gas through to the burner.

It is a gas-pressure mechanical servo. If you take an old gas valve to pieces, you will find that it is very complicated, but the principle is as I've described.

John Woodgate
Via e-mail

Basically, a gas valve incorporates an electrical 'ratchet' comprising a coil wound around an armature, as in the sketch.

The gas valve is a soft-iron plunger, lightly spring loaded in its 'out' position. It has a flat plate on the 'outside' end,



Thermocouple-locked gas valve. The coil doesn't move the valve plunger; it simply holds it in place when it is open.

which can cut off the gas supply – the default condition.

When the plunger is depressed – which you do when you hold in the button with a flame logo on it while firing the pilot ignitor – a small plate on the 'inside' end is pushed into contact with the armature.

In operation, you have to hold in the button until the thermocouple has produced enough power to hold this plate on the armature unaided.

Incidentally, I investigated some faulty thermocouples and a new 'good' one this afternoon, and can confirm the following: Off-load, in a gas cooker flame, the good thermocouple produces a potential difference of about 25mV and has a resistance of less than 0.1Ω .

On-load I believe that it can easily produce $2\text{-}3\text{W}$. At $40\text{-}50\mu\text{V}/^\circ\text{C}$, this equates to a flame temperature of about $650\text{-}700^\circ$, allowing for the fact that the 'cold' junction is probably also rather warm.

This particular example in fact produces so much power that it has a secondary cutout – probably a bimetallic strip thermostat – wired in series with it as an additional safety cutout on the boiler it controls. This is not wired in pyro, but does not seem to interfere with the thermocouple/gas valve operation.

While experimenting with the good thermocouple, it was surprising how little heat was actually required to hold in the actuator plunger – just a quick waft in the gas flame for a few

seconds.

If you push the plunger in a few times while warming it, at first it gets 'sticky', then it just stays in. Very satisfying.

All of the faulty thermocouples had gone high resistance, about $4\text{-}5\Omega$. I suspect that the cold welds fail – possibly due to migration effects, but not temperature cycling fatigue, as by and large they are not cycled.

Perhaps somebody else can explain the failure mechanism? On reflection, I think, that it's not surprising that this query arose. The majority of electronic engineers are just not conditioned to consider that something normally used as a sensor can actually produce enough energy to do something useful – merely by obeying Ohm's law.

There is a moral, of course. For assistance, read Alexander Calandra.¹ There, now I've probably said too much...

Steve Garnett
Via e-mail

Reference

1. Williams, J. (Ed) 1991 'Analog Circuit Design – Art, Science and Personalities', Newnes: pp 3-4, 'Barometers and Analog Design.'

Just got the May 2000 issue. The effective source impedance of a thermocouple is proportional to junction area. Therefore, build a big enough thermocouple and you could get enough current to weld with – theoretically, anyway.

Chris Eccles
Via e-mail

some PC software they've designed for the Navy. Hmm... OK. He then goes on to tell us about a guided artillery shell – but then fails to mention that the technology was developed 10-12 years ago by private industry – at no cost to the taxpayer.

Oh dear. 12 000 employees and £1000 000 000 p.a? I think I'd have to back the treasury.

But why stop with DERA? Why not flog the lot off? Split it into four bits – Army, Navy, Air Force and odds & sods – and float it off. No seriously; think about it: contrary to what Richard Wilson says there is still an awful lot to sell off. MoD owns more land than all the privatised industries put together. It almost certainly owns more assets, so selling it would raise a truly enormous sum.

We could have an immediate cut in income tax – some 10-12 pence in the pound perhaps – because there would be no need for a defence budget.

The government would receive a truly massive cash windfall to spend on education, NHS, policing and welfare. Generals, admirals, air commodores and the like would love it – they'd all receive astronomical rises *à la* Cedric Brown.

The decision to use the military would still rest with the government. The four services would raise funds by selling defence insurance.

We'd have no more 'Air-borne Early Warning' type fiascos. But most importantly you would be empowered to support the service of your choice. You might think the Navy needs a brace of new

There's no mouse on my knee

Apologies to readers who have tried to find the company that supplies the mouse that sits on one's knee. There's two companies by the name of NMI and only one stocks mice. The one you want is NMI at 12 Lichfield Close, Newcastle upon Tyne NE3 2YW. tel. 0191 214 6704. www.NMI.ukf.net

carriers, perhaps, so you could invest in the Navy. I might choose to support the RAF.

Richard Wilson could invest in DERA – and show his concern by giving them even more than he does now.

Like Lady Thatcher said – it's about giving the consumer choice.

R M Burfoot Bristol

Blumlein line

The recent biography of A.D. Blumlein *Electronics World*, Jan 2000, makes fascinating reading for many reasons, and not least because of the interest shown in *EW/WW* for some years over the contrast between his many achievements and the former dearth of recognition and of a biography.

The biography covers a very large amount of his work in great detail – especially that of stereo reproduction and of television. But take even more heart for a small community in the UK and a larger community in the USA that have celebrated the great man outside electronics for at least 30 years. This appears to be little known in the wider electronics field, so perhaps I may be permitted in retirement to play the (faulty) ancient mariner.

Blumlein's patent specification 589127, Oct 10 1941 refers to a double transmission-line pulse generator for high voltages. It enables the full charge voltage to be outputted, compared with a single line in which only half the charge voltage outputs to the matched load.

In his specification, distributed (LC) lines are considered, but coaxial lines work identically. He was obviously interested in driving radar modulators.

In 1960 I worked for a short period for the late Charlie Martin at AWRE, Aldermaston near the start of his groundbreaking work on the generation of multi-megavolt

sub-microsecond pulses and relativistic electron beams. I do not remember the name of Blumlein being mentioned.

In 1967 I returned to pulsed work, joining Patt Flynn's team to design and build Charlie's then brainchild, a 5MV, 100kA, 70ns generator, which we called Eros. I believe that it is still operational.

By then Charlie had already enthused his own team, devised the giant double-concentric transmission line and enthused the Americans at Sandia, Livermore, Naval Research Labs, Physics International and elsewhere.

Someone with a better memory, or records, than myself may know the exact date and circumstances, but as you will have guessed, by 1967 the 'pulse-forming line' had become 'the Blumlein'. Succeeding generations of the pulsed-power community in the UK and USA have always known it by that name.

The size of the Eros Blumlein would have surprised the man himself – at about 30ft long and 11ft outside diameter of concentric steel tubes with 10in of oil insulation. For many years it was the largest in the UK, but was already well dwarfed by some American giants.

Independently, Roy Fitch and Vernon Howell at Aldermaston published a paper on novel forms of high-voltage pulse generation, extending the idea of vector inversion and adding on an idea inherent in Blumlein's patent; *Proc. IEE*, Vol.111, No

4,849, Apr 1964.

Fitch later went on to develop the ringing Marx generator at Maxwell, San Diego, and Howell developed the spiral generator (rudely christened the toilet roll) commercially for flash X-Ray generation.

Like Blumlein, Charlie Martin rarely published in the open literature, but on the rare occasion one of his famous internal notes was published in later years, *Proc. IEE*, Vol. 80, No 6, 934, Jun 1992. Blumlein was included, as he was also in Adler's Pulse Power Formulary, 1989: North Star Research Corporation. Perhaps I may claim one of the earlier published mentions of the Blumlein; *J. of Phys. E*, Vol.6, 1223, 1973.

The Blumlein principle is also well known among the users and makers of thyratrons, especially the experts at EEV-Marconi at Chelmsford and, as a final, but pleasant, irony, in recent years Harry Kitchin, of Bournlea Instruments has made a Blumlein-based high-voltage generator for the great man's old establishment at Malvern.

*E Thornton
Telbury
Gloucestershire*

The name of Alan Dower Blumlein is known to historians of electronics as one of the key figures in the formative years of electronic engineering.

It has long been a matter of regret that no comprehensive biography has been available – until now, when two books have

appeared within a few months of each other. Robert Alexander's, 'The inventor of stereo: the life and work of Alan Dower Blumlein' (Focal Press) has been widely publicised. I should like to draw your attention to Russell Burns' 'The life and times of AD Blumlein', which was published in January by the Institution of Electrical Engineers (IEE). This is a major scholarly, yet very readable, treatment of Blumlein's life.

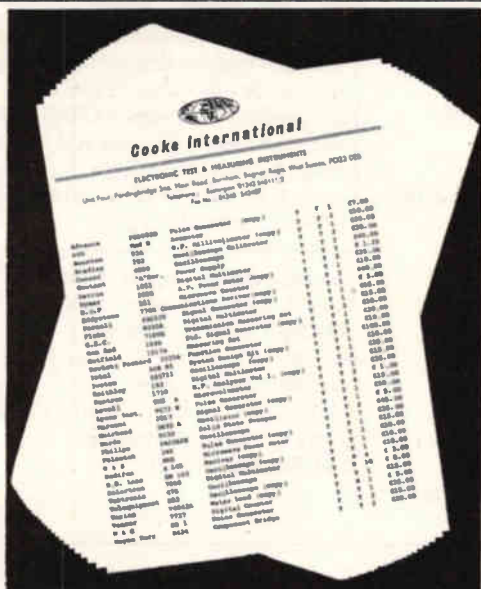
The author details Blumlein's work on transmission, mono and stereo recording and reproduction, television and radar. His writing is also informed throughout by Burns' close study of taped recordings (now in the National Sound Archives) of the reflections and recollections of Blumlein's wife, Doreen, and his best friend JB Kaye. Burns has worked in close contact with the Blumlein family and the book carries a Foreword by Blumlein's eldest son, Simon.

For those familiar with the history of the efforts to obtain a biography of Blumlein, note that both of these two new books are entirely independent of the efforts of the late Francis Thompson. It is not clear whether any of the Blumlein material collected by Thompson will be made public and, if it is, whether it will add to what Burns and Alexander have to tell us.

*Robin Mellors-Bourne
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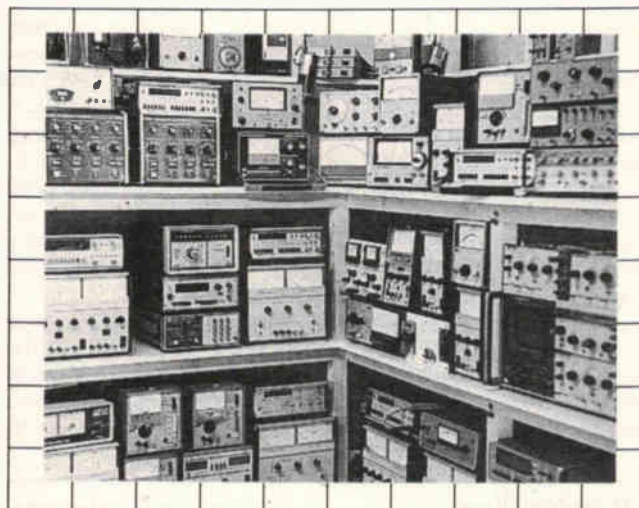
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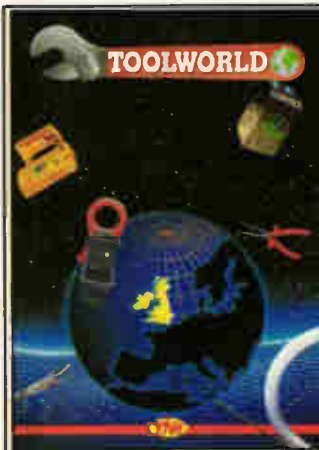
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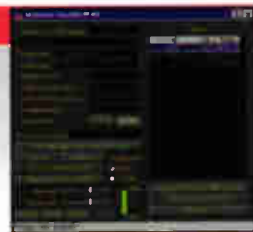
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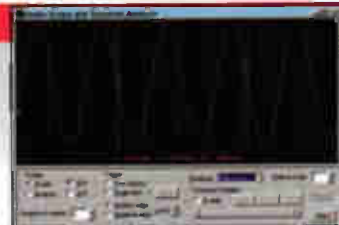
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