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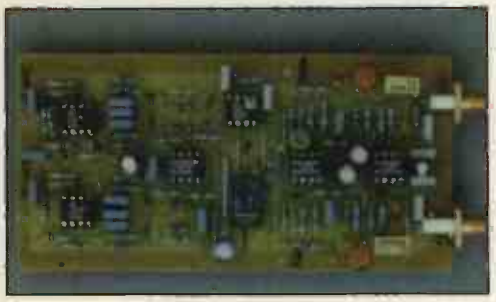

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February issue on sale 23 December

Cover photography Mark Swallow

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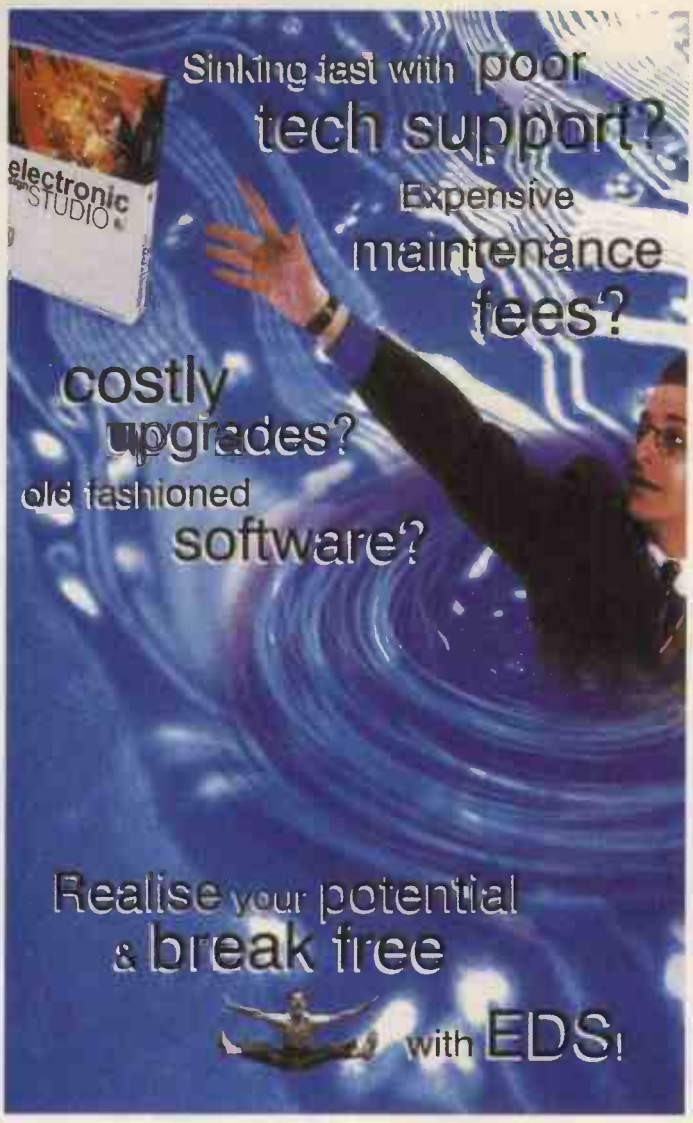
New sensor film indicates tactile force between surfaces – see News on page 8.

Will the telecoms industry make the UK the caveman of Europe? David Manners reports on page 64.



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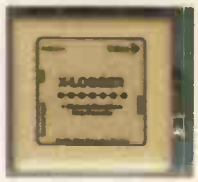

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Exploiting third world skills

So, the Labour government's plan to fill the IT skills shortage by allowing large numbers of suitably skilled immigrants – mostly from the Indian sub-continent – to by-pass the normal channels and enter Britain freely is now virtually in place.

Why should anyone worry about this? These people will certainly not be a burden on the British social security system. They will be paying taxes along with the rest of us, and being educated people they will most likely fit easily into our society.

Apart from the obvious need to use up a bit more green belt to provide more housing in this tightly packed island, everything should be fine, shouldn't it? But it isn't.

Firstly, extracting large numbers of the brightest and best skilled personnel from third-world countries like India strikes a considerable blow to the Indian IT infrastructure. There have already been protests from India about a first-world country like Britain pilfering a third-world country's intellectual assets in order to prop up its shaky IT infrastructure in this way. This is entirely understandable.

The situation is similar to the recent raid that the NHS made in the Caribbean for nurses to shore up its equally deficient ranks. An undesirable result of this is that many hospitals in the Caribbean are now short-staffed. But then, why should we worry about that? We're all right Jack!

In both these cases, the fact that Britain is a former colonial power and is still behaving so thoughtlessly and selfishly towards its former colonies has raised hackles.

Secondly, there is evidence that there are numbers of well qualified IT people in Britain that are being excluded from the jobs market *because* of their age. A letter in the 21 September issue of *Computer Weekly* from a 38 year old IT graduate highlighted this interesting problem.

It appears that industry does not just want IT workers – it also wants *cheap* IT workers. Older IT professionals commanding a higher wage are out.

You can bet that the immigrant workers will not be paid the going rate. The law of supply and demand will ensure this. The jobs market will be diluted sufficiently to depress the general wage But by importing workers from abroad the politicians are pandering to the industry in this respect, and in doing so, promoting ageism.

Thirdly, an injection of IT personnel from abroad is the very epitome of the despised policy of short-termism. It does absolutely nothing to address the roots of the problem – which is that we are simply not recruiting and training sufficient IT workers ourselves.

The problem is that not many school leavers are all that interested in IT at a technical level. They are acquiring some IT skills because that is almost unavoidable in schools these days. But they are much more worldly-wise than former generations.

School leavers know that the quickest route to a new BMW and that house in the country with a swimming pool is to be a Eurobond dealer, human rights lawyer or some similar parasitic trade along those lines. And who can blame them? Why be in the hot seat as an over-burdened and under-paid employee in the IT department of some demanding company, to be looked down on and patronised by a technically ignorant board of directors when more interesting careers with better rewards are offered elsewhere?

Below is reprinted an extract from a 1984 edition of *Electronics World* highlighting this very problem. It could have been written yesterday. And it highlights the total incompetence of our politicians regardless of which party they belong to.

Indeed, politics and not technology is once again at the very heart of this problem. Observers of the political scene – particularly those with a socialist leaning – will marvel at the fact that it is not an extreme right-wing anti-worker government doing this. It is a *Labour* government.

However, the more cynical of you will see it as just another example – as if we needed any more after the Dome fiasco – of the breathtaking moral bankruptcy of the current breed of career politician, whose main aim in life is to stay in power rather than advance any personal conviction. ■

Simon Wright

Extract from *Electronics World*, October 1984

To investigate, and make proposals to rectify, the country's shortage of skilled people in the information technology field, the DTI has set up an IT Skills Shortage Committee under the Chairmanship of Under-secretary John Butcher.

The Committee's first report looks at the shortage of graduate skills. The Alvey directorate has estimated a current shortfall of some 1500 graduates which is likely to escalate, if no action were taken, to 5000 by 1988.

Recommendations put forward by Committee are to increase the number of first degree places in the appropriate areas, increase the number of IT conversion courses for updating or upgrading a student's IT skills.

Areas identified as being in particularly short supply of the right technical skills are:artificial intelligence, large-scale integrated circuit design, and software engineering. There is also a great shortage of teachers of these skills.

In addition to increasing the number of university and polytechnic places, the committee has recommended the institution of the Great Switch by reducing the number of places in less productive disciplines (the education spokesman in the Committee pointed out a more than ample supply of historians and biologists).

The major strategy is to encourage the industry to provide equipment and expertise in the setting up of a new Partnership of collaboration with the Government and with academic institutions.

Another strategy is to attract brain-drain ex-patriots back with suitable tax and capital incentives...

...The need for urgency is paramount, especially as it takes five years from the time that a 16 year old chooses which subjects in which to specialise to the time that a 21 year old emerges with a first degree.

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UPDATE

Consumer Association questions mobile phone hands-free test methods

Which? magazine's second set of mobile phone hands-free kit test results wboth ere revealed recently.

The Consumers' Association (CA), publisher of *Which?*, tested radiation levels from hands-free kits (HFKs) in April and discovered they increased radiation into the brain rather than decreasing it.

This first set of results was rubbished by both the phone industry and the DTI, which reported on its own set of tests in August.

Criticism was of the CA's choice of test. It constructed one that it considered a good model of actual use. Industry standard tests for HFKs are adaptations of tests designed for phone handsets – which are now, somewhat inaccurately, grouped under the more general term 'SAR' tests.

The second set of *Which?* tests were initiated to discover why its initial test results differed so markedly from the SAR tests performed by phone companies, HFK-makers and the DTI.

These repeated and extended the first set – at test house ERA – and performed the SAR tests at SAR Test, the test house used by the DTI.

The CA concludes that handset SAR tests are not applicable to HFKs. "The tests are designed to test mobile handsets. They do not always allow for the appropriate position of the handset cable [when used to test personal hands-free kits]," according to CA principal scientist Roy Brooker.

Now all the test results have been published and the CA is convinced that the SAR test and the CA's test are not mutually exclusive, and that its tests are more representative of real HFK use.

Key findings at a glance

- All hands-free kits can push in-brain signal levels up or down.
- Signal level is mainly dependent on free wire length '*d*'.
- No rule-of-thumb can be made to minimise exposure.
- Specific industry tests for hands free kits should be devised.
- No kit/phone combination pushes signals above Euro-limits.

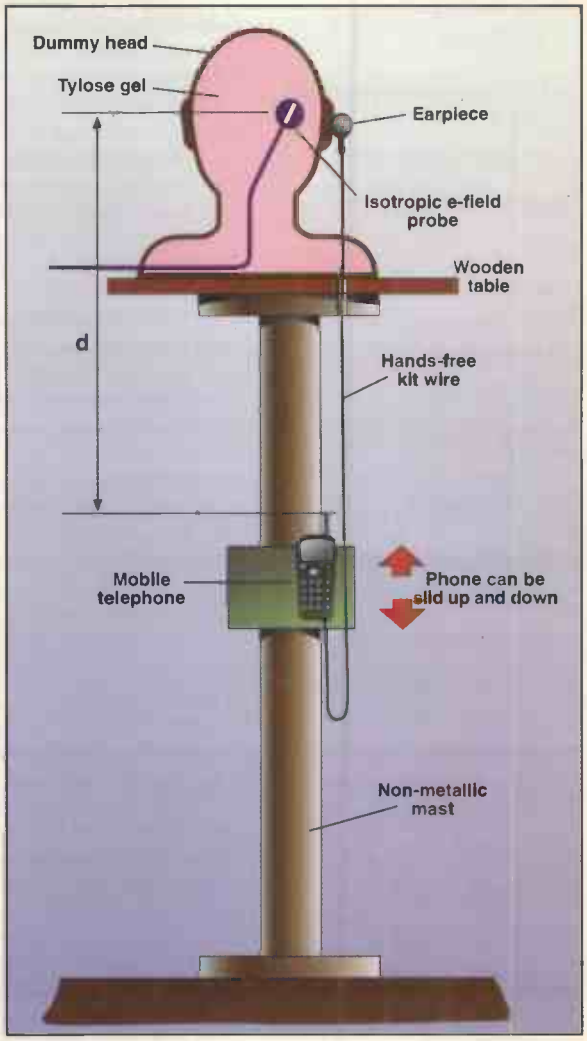
It claims that, in normal use, in-brain signal levels are mainly determined by the length of cable between the phone antenna tip and ear piece (shown as '*d*' in the diagram).

It also found the relationship between *d* and signal level always followed a notch-like curve. Level decreased for a narrow range of *d* and increased progressively either side of this. A movement of 4cm was enough to change from attenuation to gain.

The curve was similar, but not identical, to all HFK-phone combinations. Curve differences and steepness make a signal-reducing rule-of-thumb impossible.

Preliminary work, says the CA, shows that ferrite beads on the HFK lead can cut radiation levels. Some questions have been raised about the validity of the RF probe used in the test. A CA spokeswoman said the organisation was confident that the probe was valid.

Steve Bush



The* Which? *test rig. The antenna is always at the same place, inside the simulated head. Crucial, says the CA, is that the hands-free kit hangs down as it does in use – impossible in industry standard 'SAR' tests. In-head levels were 10 to 350 per cent of measurements with the phone held to the head.

UK parallel processing start-up is set for product launch

Aberdeen parallel processing start-up Axeon has completed a second round of funding, raising £3.5m to finance its product launch.

The Scottish firm plans to launch its 'learning processor' early next year. The design, used as a co-processor to a DSP, will enable 3G mobile phones to run at their maximum of 2Mbit/s, claimed Hamish Grant, CEO of Axeon.

"The current state of the art is potentially only going to deliver 56kbit/s in the 3G environment," Grant told *Electronics Weekly*. "They're talking about aspiring over the next three or four years to reach the 384kbit/s level."

Axeon's design is a parallel array of 256 Risc processors containing a patented algorithm, similar to a neural network. Unlike other arrays, each processor can operate independently using its own instruction set.

Its strength is solving problems that have uncertain data or uncertain data relationships. Channel equalisation in a mobile phone is a perfect example of this.

"If you integrate our component today, you will be able to get 2Mbit/s in an economic chip," said Grant.

The complexity of channel equalisation means it takes thousands of Mips on a conventional digital signal processor. Even if the task could be done conventionally, it would raise an unacceptable power figure.

Axeon's processor is orders of magnitude better than a DSP in terms of both overall performance and Mips per Watt, Grant said.

By the end of the year Grant hopes to have a fully working 3G prototype, while a team is working on a synthesisable version of the design.

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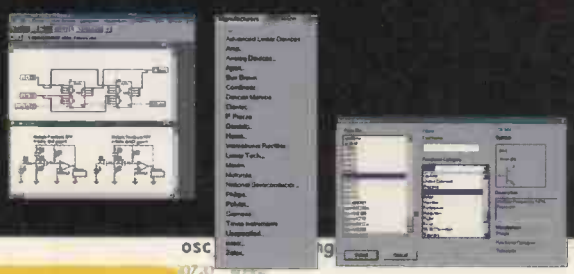

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CIRCLE NO.108 ON REPLY CARD

Government invests in 'Cybercops' to combat e-crime

The Government is to invest £62m to employ more than 80 specially trained 'Cybercops' to tackle crime on the Internet.

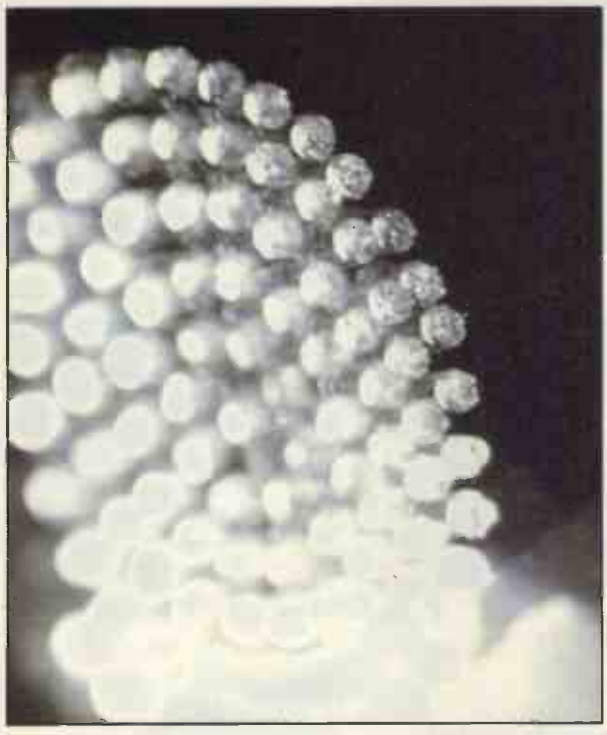
Home Secretary Jack Straw said the investment was a key part of the drive to make Britain the best place to do e-business.

A new National Management Information System (NMIS) will be set up for police forces in England and Wales using £37m of the money.

It will provide them with a comprehensive information management and analysis tool joining up data held on a variety of different police IT systems. It will present data in a consistent format so that the whole range of police business can be easily and reliable compared and analysed nationwide.

NMIS is likely to be rolled out throughout the criminal justice system and related organisations in the future. The other £25m will pay for up to 86 special police officers both nationally and locally.

There will be an extra 40 specially trained officers and the new high-tech crime unit to investigate Internet and other e-crime will begin work in April 2001.



The black art of white LEDs

The white LED is the least understood of all recent LED developments, according to Philip Logan, business development manager for supplier Marl International.

Consequentially this product is the one most likely to exhibit unexpected characteristics to the unwary designer.

The white LED is produced by a combination of a blue high intensity LED, and a phosphor material. The narrow band blue output characteristic is used to excite the phosphor, which in turn fluoresces, resulting in a wideband response, emitting a bluish-white colour.

"While this would appear to the naked eye to be white, it lacks much of the red spectrum, and has a radically different spectral output when compared with incandescent filament lamps," says Logan.

LEDs are being used to replace incandescent products in a wide range of retrofit applications, and in many instances, they are required to operate behind some form of optical filter. These optical filters are predominantly colour filters, used to remove certain bands of incandescent output to achieve particular colours.

"Because of the difference in spectral output outlined above, care needs to be taken when using a white LED to replace an incandescent bulb," says Logan.

For example, a red LED behind a red filter will give a better performance than using a white LED, despite the original light source being a white incandescent bulb. White LEDs have tremendous potential for illumination, rather than indication, and research is in progress to develop this type of application.

www.marl.co.uk

300mm wafers could give Taiwan world lead

Taiwan could take the lead in semiconductor manufacturing in 2002/3 as seven out of the fifteen wafer fabs being built to use the larger 300mm wafer size are either Taiwanese-owned or are a part of a Taiwanese joint venture.

Such an advantage in manufacturing, which reduces costs per chip by 30 per cent, could propel Taiwan into a leading position in the semiconductor industry.

According to the Arizona analysts company IC Insights, the seven fabs being built either by Taiwanese companies or in a joint venture with a Taiwanese company are: Trecenti – the Hitachi/UMC joint venture to build a 300mm fab in Japan due to come on-stream in January next year; ProMOS, the Infineon/Mosel-Vitelic joint venture which aims to start volume production in 2002; Powerchip, the Mitsubishi/ Powerchip joint venture which expects to start running 300mm wafers in 2002; Macronix starting 2002, UMC starting next year; and two fabs at TSMC starting 2001 and 2002.

Europe has an interest in three of the 300mm fabs being built: ST/Philips at Crolles starting in 2002; Infineon/Motorola at Dresden due to move into volume in 2002, and Infineon's ProMOS joint venture fab.

Japan has an interest in three fabs: Hitachi's interest in Trecenti; Mitsubishi's interest in Powerchip, and NEC's own 300mm fab in Hiroshima.

America is involved in four fabs: TI in Dallas due to start next year; Intel in Oregon and Arizona due to start production in 2002, and IBM in New York.

The move from 200mm to 300mm has taken much longer than expected. In 1996, equipment trade body SEMI said 300mm equipment was 'already available' and forecast that there would be nine 300mm pilot lines and two volume production lines running in 1998.

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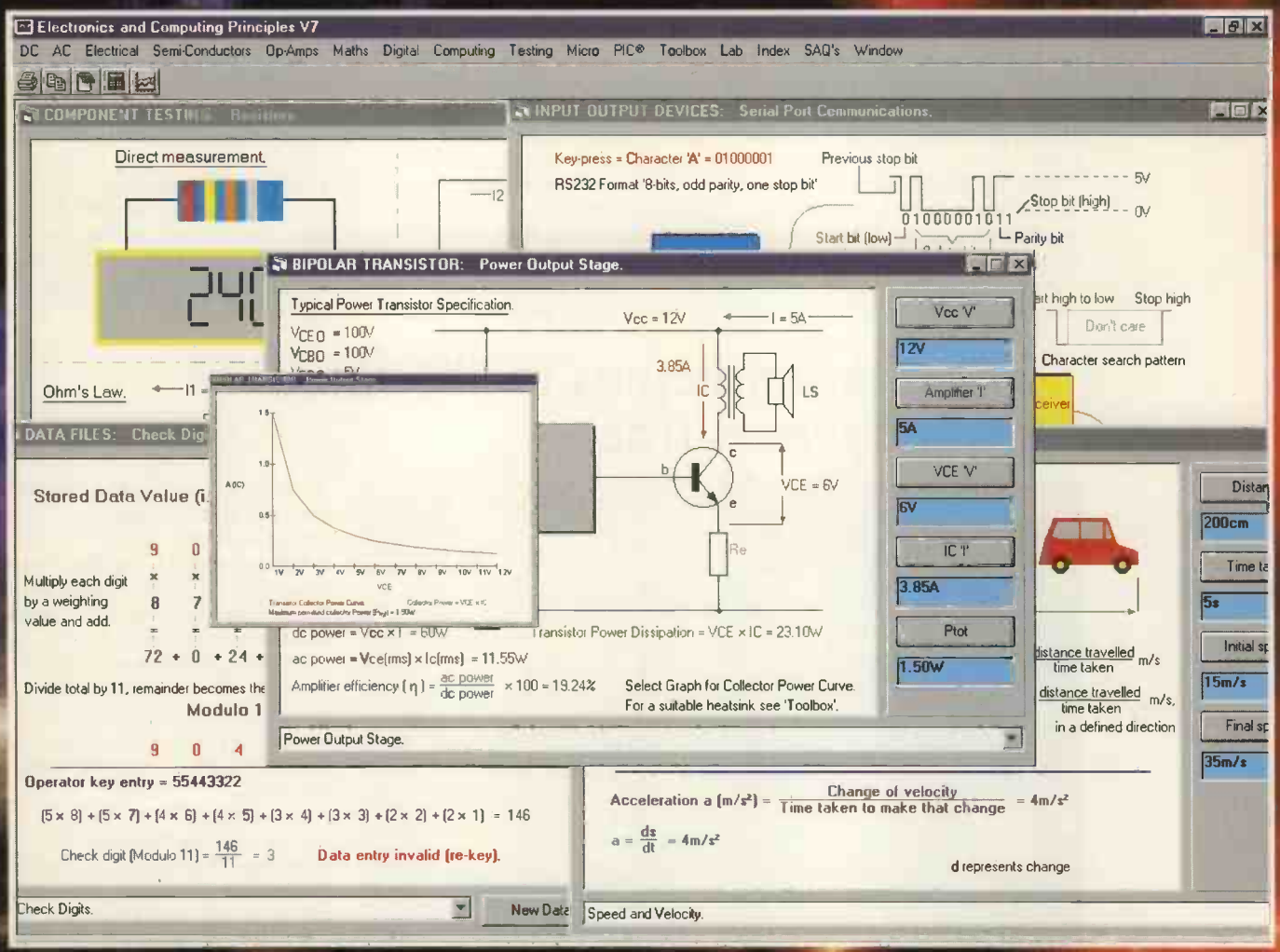

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CIRCLE NO. 109 ON REPLY CARD

All Fords will have digital radio by 2004

The fledgling digital audio tuner market has been given a shot in the arm with Ford's announcement that it is to fit digital radio receivers in all new cars sold in the UK by 2004.

The car market has been considered crucial for digital radio by analysts and the industry as a means of achieving volume production which will result in lower prices.

Ford has also become a shareholder in the MXR consortium, which is bidding for UK regional digital radio licences. "This represents the single biggest positive step for the uptake of digital radio in the UK," said Phil Riley, MXR consortium's chairman.

Also, in a move to promote the technology, an organisation named the Digital Radio Development Bureau, is being created.

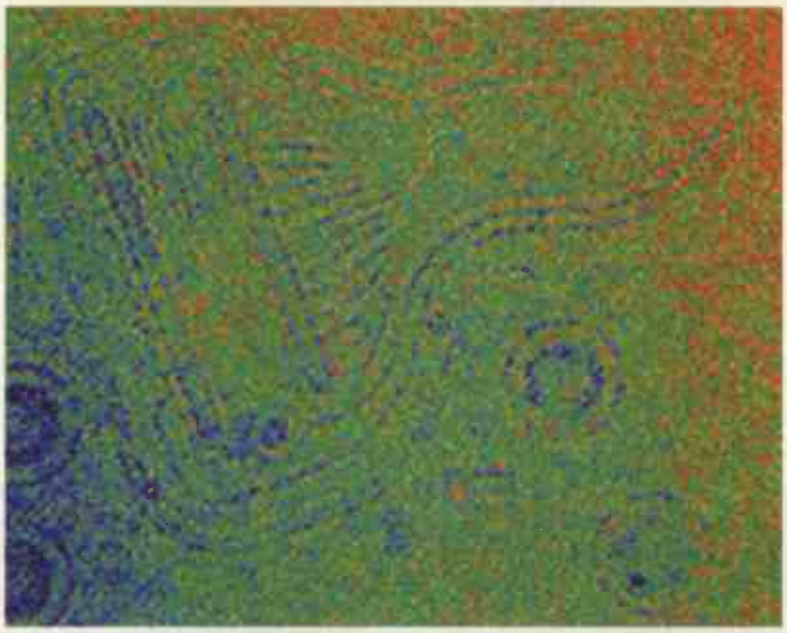
The Bureau will bring together broadcasters, retailers and manufacturers in a united effort to promote digital radio and to work together on developing sets.

"The task of the Bureau is to ensure that there are a wide variety and large number of sets in the market-place at competitive prices in as short a time as is reasonably possible," said Paul Brown, chief executive at the Commercial Radio Companies Association.

Brown played down a suggestion that the industry would follow the route taken by the digital TV industry and subsidise sets. "We're a very different medium; our income is derived entirely from sponsorship and advertising," said Brown, "We don't get anything back in the form of a subscription."

● A digital radio receiver for less than £300 is now available. The tuner, from VideoLogic Systems, connects to any hi-fi and will initially be available from selected Dixons stores.

Sensor film indicates tactile force between two surfaces



A sensor film that measures tactile force has been developed by US firm Sensor Products.

Called Pressurex, the film is used to measure the compression force between two surfaces. When placed between the surfaces and compressed, the film changes colour, reflecting the force between the surfaces. Comparing the film colour to a colour chart, or using an imaging system, allows pressure to be measured.

A typical application would be checking a laminating press for multi-layer PCB manufacture or checking the pressure between the rollers of a dry film laminator.

It could also reveal inconsistencies in resistor thickness across large substrates in hybrid chip manufacture. The image shows the lamination pressure of a multilayer PCB.

Internet buyers get goods return rights

A law protecting British consumers from dodgy Internet selling has come into force.

The Consumer Protection (Distance Selling) Regulations give consumers the right to obtain information about the sale of goods and the right to withdraw from a contract.

The law affects all forms of distance selling, such as mail order and catalogue shopping, but its most interesting impact is on the Web.

Whether buying a new PC mouse or a car, the buyer has the right to return the goods and get their money back for any reason.

Technology lawyer Dai Davis, a consultant with UK law firm Nabarro Nathanson, said sellers will have to accept some goods will be returned in less than perfect condition.

And if the seller fails to inform the buyer of their rights, then the right to withdraw from the contract extends to three months and seven days.

There are, as one would expect, exceptions, including sales of services such as accommodation, transport and catering. This also covers things like cinema tickets, personalised goods and software that is opened by the buyer.

200 pixels per inch liquid-crystal displays

IBM has started production of a 22-inch liquid-crystal display (LCD) that packs in 200 pixels per inch – a resolution around 12 times sharper than current monitors.

With nine million pixels in total, the display is said to be as clear and sharp as a photograph. Based on well known active matrix technology, the LCD uses aluminium instead of the more traditional molybdenum and tungsten.

Applications for the displays include medical imaging, weather forecasting, satellite mapping and publishing.

The first customer for the displays is the Lawrence Livermore National Laboratory in California, which will use it to display models of nuclear weapons ageing.

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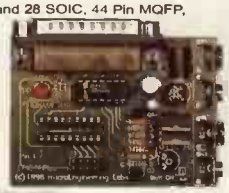

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CIRCLE NO. 110 ON REPLY CARD

New display technology

Dublin start-up Nanomat is looking at the advantages of using electrochromic chemicals to make displays

Nanomat, a spin-off from University College Dublin, is taking a close look at electrochromic chemicals with a view to making displays. These rely on chemicals which change colour or transmissivity when electrons are added or removed and have been used in electrically-dimmable car rear view mirrors.

In a paper to the SID conference, Nanomat's senior scientist Dr Diarmuid O'Brien outlined the advantages and disadvantages of electrochromic displays (ECDs) and showed how the company has jumped the first hurdle, of super-slow response time.

ECDs have been made with electrodes in a solution of electrochromic chemicals.

"When the materials are used in solution, the response time is around 15 seconds because the electrochromic molecules have to diffuse to the electrodes," said O'Brien.

To quicken the electron transfer, Nanomat is attaching the colour-change materials directly to the electrodes. "By injecting electrons directly into the dye [EC material] we get a 15ms response time, and a memory of up to five days."

This means that the display can be updated quickly. It also means that it will retain that display for almost a week without power.

Achieving a profound colour change means altering the state of a lot of molecules, all of which have to be connected to the electrodes.

O'Brien has created a large surface area electrode that can accommodate a three dimensional matrix of EC molecules by sintering conductive nano-crystals over a layer of indium tin oxide (ITO) on a glass substrate.

Two similar sheets of glass, positioned with conductive sides together, are used to make the display cell.

A potential applied between the two conductors transfers electrons to the colour-change material, called a viologen, which changes from transparent to coloured.

Currently, different viologen chromophores can be blended to get green, blue, violet and black.

Between the conductors is a charge-balancing lithium-based electrolyte. This is loaded with white particles to give the chromophores a white background.

Now that speed is not an issue, Nanomat is looking for a way to multiplex a two-dimensional array of EC pixels.

Like liquid crystals, the materials exhibit a knee in their voltage/optical characteristic that lends itself to passive-matrix addressing. Unfortunately, the knee is not so profound as with liquid crystals which could lead to contrast reduction through the partial turn-on of pixels in the row and column of an addressed pixel. Solutions to this problem are under investigation.

Steve Bush

Nanomat claims that its ECD exceeds the optical performance of other displays – including ink on paper.

Throw away your paperbacks?

Many attempts are being made to develop an effective display for electronic books.

What is needed is a device that has advantages over the original.

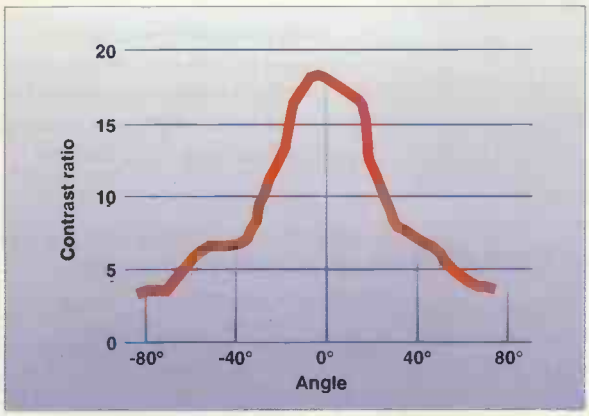
While it is unlikely to compete on cost for a long time, it must: be more compact, be at least as easy to read, and have negligible power consumption.

Assuming an electronic book only needs to compete with normal books, it will not need video-speed, may not need full colour and perhaps not even a grey-scale.

Taking this into account, several backwater display technologies begin to look attractive for electronic books.

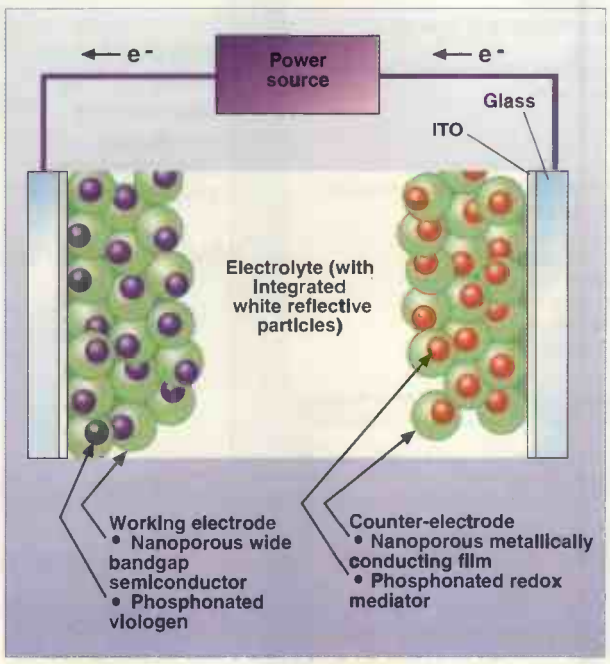
Among these are cholesteric LCDs (Ch-LCDs), ferroelectric LCDs, zenithal bistable LCDs, rotating micro-particle displays and electrophoretic displays. All of these can be used in non-emissive mode, so power is very low, and many of them are bi-stable, power consumption is zero between updates.

To this list can now be added electrochromic displays.



Nanomat's guide to contrast ratios

	@0°	@30°	@45°	@80°
Book	9:1	8:1	7.5:1	3.5:1
Newspaper	6:1	5.5:1	5:1	4.5:1
CH-LCD	4.5:1	4:1	4:1	3:1
STN LCD	4:1	3:1	2:1	0
ECD	18:1	7.5:1	6.5:1	3.5:1



Nanomat's electrochromic display cell. The viologen on the left changes from transparent to violet when it accepts an electron. When viewed from the left, the colour shows up against the background of white particles suspended in the electrolyte. The electron source is also a colour-change material so a red on white image can be seen from the back of the display. Response speed is 15ms at 3V for small pixels. Tests indicate life exceeds 5.5 million display cycles.

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HP 8410-A-B-C Network Analyser 110Mc/s to 12 Ghz or 18 GHz - plus most other units and displays used in this set-up - 8411a-8412-8413-8414-8418-8740-8741-8742-8743-8745-8650. From £1k.
Racal/Dana 9301A 9302 RF millivoltmeter - 1.5/2GHz - qty in stock £250-£400.
Racal/Dana Modulation Meter Type 9009-9008 - 8Mc/s - 1.5GHz - £150/£250 – 9009A £350.
Marconi RCL Bridge type TF2700 - £150.
Marconi Microwave 6600A 1 sweep osc., mainframe with 6650PI - 18-26.5 GHz or 6651 PI - 26.5-40GHz-£750pr - only £600. MF only £250.
Gould J3B test oscillator + manual - £150.
Barr & Stroud Variable filter EF3 0.1Hz-100Khz + high pass + low pass - £150, other makes in stock.
Racal/Dana 9300 RMS voltmeter - £250.
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Marconi mod meters type TF2304 - £250 – TF2305 - £1,000.
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HP180TR. HP181T, HP182T mainframes £300 - £500.
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HP8620C Mainframe - £250. IEEE £350.
HP8165A Programmable signal source - 1MHZ - 50Mc/s - £1k
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HP5370A Universal time interval counter - £
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TEKTRONIX 1502/1503 TDR cable test set - £400.
HP8699B Sweep PI YIG oscillator .01 - 4GHz - £300. 8690B MF-£250. Both £500.
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HP8405A Vector voltmeter - late colour - £400.
HP8508A Vector voltmeter - £2500

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Anritsu ML93B & Optical Lead Power Meter - £350.
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Anritsu MW97A Pulse Echo Tester.
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Anritsu MW98A Time Domain Reflector.
PI available - MH914C 1.3 - MH915B 1.3 - MH913B 0.85 - MH925A 1.3 - MH929A 1.55 - MH925A 1.3GI - MH914C 1.3SM - £500 + one P.I.
Anritsu MZ100A E/O Converter.
+ MG912B (LD 1.35) Light Source + MG92B (LD 0.85) Light Source £350.
Anritsu MZ118A O/E Converter.
+MH922A 0.8 O/E unit + MH923 A1.3 O/E unit £350.
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Anritsu MN95B Variable Att. 1300 £100.
Photo Dyne 1950 XR Continuous Att. 1300 - 1500 £100.
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HP84801A Fibre Power Sensor 600-1200 £250.
HP8158B ATT OPT 002+011 1300-1550 £300.
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STC OFR10 Reflectometer – £250.
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HP MICROWAVE TWT AMPLIFIER 489A11-2GHZ 30DB – £400
HP PREAMPLIFIER 8447A 0.1-400MC/S – £500. Dual – £3
HP PREAMPLIFIER 8447D 0.0-1.3GHZ – £400.
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TEK 2465 4ch-300MC/S – £1,150.
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TEK 2465ACT 4ch-350MC/S – £1,750.
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Racal/Dana Counter 1992-1.3GHz – £600.
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HP3312A Function Generator AM-FM 13MC/S-Dual – £300.
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Racal/Dana 9084 SYN S/G AM-FM-PH-.001-104MC/S – £300.
Racal/Dana 9087 SYN S/G AM-FM-PH-.001-1300MC/S – £1K.
Anritsu MG3601A SYN AM-FM 0.1-1040MC/S – £1.2K.

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CIRCLE NO. 111 ON REPLY CARD

Cybercrime goes big-time

Andrew Emmerson reports on a communications crime that's completely out of control – phone fraud. What's more, we're all paying the price.

Asked to name the most prevalent wire fraud you'd probably guess bogus credit card orders over the Internet. But you'd be wrong. Global losses to Internet-related fraud reported last year by MasterCard and Visa amounted to a 'mere' $10 million apiece, whereas phone calls charged to other users' accounts are over 2000 times higher.

In fact a new report issued by British consultancy Chorleywood Consulting conservatively estimates the impact of fraud in the telecomms industry world-wide to be over $22bn in 2000 – and rising. And that's without taking into account the large amount of fraud on cellular networks.

The ease with which these frauds are perpetrated is amazing. What's even more surprising is the indolence of users to protect themselves against their antagonists. The techniques employed by the phraudsters – and the straightforward remedies of defence – are publicised widely yet users continually fail to take the most basic precautions.

For instance, the last organisation you'd expect to suffer an attack from phone fraudsters would be the Metropolitan Police headquarters at New Scotland Yard. But you'd be wrong. Only a few years ago the law-enforcers were obliged to spend £15 million revamping their phone system after losing a sum claimed as close to £1 million to phone crooks.

Users aren't the only victims

Mobile-phone operators have been stung too. BT Cellnet had the misfortune to sell over a million pre-pay handsets that allowed deceitful users to modify them to give free phone calls for life. This was in addition to an earlier fraud in which prepaid cellular phone cards were being topped up with stolen credit cards.

What's even more amazing is that these frauds are allowed

to continue unabated. Eight years ago I wrote in this magazine,

"The fact these people get away with this so easily speaks volumes for the lack of security on modern user systems. Call it an allegory of modern times if you will, but it could not have occurred without the de-skilling of communications management in large firms and the general ignorance of telephone users. The net result is that many phraudsters know more about the programming of modern call-connect systems than their official custodians, which is how the former can manipulate them undetected to their own ends."

Since then fraud has mushroomed, proving that while the phraudsters have sharpened their skills, comms managers and security experts have not.

Not surprisingly the main targets are smaller firms and public authorities ill-qualified in detecting and preventing this kind of theft. Some victims though are high-profile organisations that should have known better. Part of the problem, according to BT, is that nobody expects to be a victim of phone fraud; it only happens to other firms. And while everyone has heard of computer hackers and recognises that danger, they have a mental block accepting that a processor-controlled switchboard is just another form of computer and just as vulnerable.

"This unawareness is what makes phone fraud so prevalent," declares communications consultant Richard Cox, who has acted as expert witness in several cases involving deception of this kind. "What's more, unlike computer hacking, telephone fraud has not attracted major media attention, leading most people to underestimate the scale of the problem or how easily it is perpetrated."

Some kinds of phone crime target phone companies; the mobile phone frauds already mentioned are an example, he continues. But in most cases the thieves' victims are other users, using three main techniques.

How is it done?

The crudest method of phone fraud is tapping into another user's line at an external junction box; this is blatant but effective and quite easy to conceal.

Far less detectable is the 'in and out' dodge; many organisations now have a facility for people working from home or in the field to dial into the company switchboard and make long-distance calls through the firm's own facilities. If not properly supervised, this provision can be exploited by unauthorised users as well.

The latest – and most cunning – ruse entraps people visiting certain sex sites on the Internet. They are told they must download a special viewer program and on installing this, click to agree to the program's 'conditions'.

It's not obvious among all the verbiage that these are more than the usual terms; in fact they include an agreement allowing the software to re-route all subsequent Internet calls via an extremely pricey premium rate or overseas number. The so-called viewer program does this surreptitiously by altering the user's dial-up network (DUN) file.

Dave Millett, security specialist with Lucent Technologies' new enterprise networks group, warns that the incidence of phone fraud is growing fast. He explains: "As new technologies are introduced, so fraud tends to migrate to them. However, 'social engineering' – tricking an insider into facilitating the fraud – is not dependent on technology and this is also on the increase. It's up to users to safeguard their own environment and to choose secure products. Hackers are here to stay, so never drop your guard."

All modern call-connect systems have call-logging facilities and users should exploit these capabilities to identify normal calling patterns and traffic levels. From this point on, detecting abnormal call activity is simple so long as a constant watch is maintained.

Specialist software to further simplify the task is widely available too. Employees have a part to play in fighting fraud too and must be educated to be vigilant and report abnormal telephone activity.

"It's this combination of speed, simplicity and stealth that makes phone fraud the perfect crime; no physical access is needed to the victim's premises. The perpetrators are also almost impossible to trace, too and while the police and phone companies all know where in east London the epicentre of this criminal activity is, that information alone is little help," states Cox.

Pennsylvania Telephone Association (PTA) was established in 1902 to devise and implement sharing of toll revenues. PTA represents the interests of its 38 local exchange carrier members in the legislative, regulatory and public arena of the Commonwealth. PTA is pleased to have over 90 associate/affiliate members who provide goods and services to the member companies and participate in the Annual Convention and Conferences. Our mission is to serve the regulated telecommunications local exchange companies in Pennsylvania and furnish leadership by promoting the interests of our members and their customers. PTA is actively preparing for the future, one which will bring about dramatic change. .

Telecommunications UK Fraud Forum (TUFF), was born in March of 1996. It was formed because it was recognized that preventing telecommunications fraud depends on the security satndards and practices of every licensed network operator in the UK. Where telecommunications fraud is concerned all operators are interdependent. It is therefore imperative that we, as operators, work together to ensure this new fragmented environment remains as secure as if it were a single network. TUFF is the only forum which attempts to represent the whole UK telecommunications industry, mobile, cable and fixed-network operators. TUFF now has 35 member organizations which represent over 95% of the UK's telecommunications customers.

PARK REGION Mutual Telephone Co.

Park Region is a local exchange cooperative headquartered in Underwood, MN, serves six exchanges in the Heart of the Lakes area of West central Minnesota and also has a competitive venture video conference center called Otter Tail Telcom in Fergus Falls, MN. In addition to state-of-the-art, digital, fiber optic local service, Park Region provides many wireless services such as cellular, television and alarm systems along with Internet access. **Federal Law Enforcement Training Center (FLETC)**

Clip from the 'Alliance to Outfox Phone Fraud' web site highlighting the UK fraud forum. The AOPF is a broad-based group of telecommunications industry and related companies whose goal is to reduce fraud by educating businesses and consumers about fraud prevention. http://www.telefraud.com/about.htm

The law offers little consolation to victims because the perpetrators can commit their crime from any location and cover their tracks well. Consequently they are seldom caught.

Legal issues

The Telecommunications (Fraud) Act 1997 has made it an offence to possess or supply equipment for use in the course of, or in connection with, obtaining telecommunication services dishonestly. It also give law enforcers a much stronger weapon for punishing the perpetrators they manage to catch. Previously, the only indictable offence was the 'unlawful abstraction of electricity' – a somewhat nebulous crime which, for many judges at least, understated the true gravity of the offence.

That said, having adequate penalties in place is one thing, actually catching the phraudsters is quite another matter. This confirms technology as the chief defence against fraud in business telephone systems.

Although the invisible enemy is capable and highly skilled, it can only succeed when security is lax, where system passwords have been left as the default code set at the factory – or

Learn the lingo

Phone pirates and phraudsters are motivated by money and are criminals who happen to have selected telephone companies or users as their victims.

Phone freaks or phreakers claim to be explorers, driven to find loopholes and work-arounds in technology for the sheer joy of discovery. Like genuine computer hackers, they wreck nothing, neither do they steal. Phrackers are people who inhabit both worlds of computing and communications, and h/p (hacker/phreaker) is another term with the same meaning.

The 2600 magazine, sold in Borders bookshops and Tower Records in London, is the unofficial organ of the h/p fraternity; it has a website at www.2600.com, while other subversive material can be found throughout the Web and in particular at http://www.members.tripod.com/~SeusslyOne/FAQ.html and http://www.phonelosers.org.uk/

Double trouble for Cellnet

BT's Cellnet mobile phone operation has been the victim of two kinds of fraud. Other networks have managed to avoid both of them through more stringent security measures.

Fraud number one occurs because, unlike other operators, the company was accepting anybody's debit or credit card numbers to top up a pre-pay phone, without positive authorisation or identity check.

Victims have not been compensated for their loss and inconvenience, since Cellnet claims to be a victim as well. A fascinating website, namely www.pardoes.com, sets out the claimants' case and runs a campaign on their behalf.

The second fraud is the 'bottomless top-up'. It enables people to make free calls indefinitely on Philips Diga phones. An additional chip is soldered to the main circuit board of the Diga, which is then programmed with software obtainable over the Internet.

Onboard routines are supposed to record the amount of allocated call credit and deduct this from the allowed call total, but switching the handset off and back on again restores the credit to the original amount. Although the rogue phones are no longer on sale, fraudulent use continues because prepay users are effectively impossible to trace.

But why is mobile phone fraud so rife? Avoiding payment is not the only pull; the ability to conceal calls is a boon to drug dealers and other criminals who suspect they are under surveillance. There's also a booming business in selling stolen calls to immigrants wanting to call home to their relatives; people on low wages are extremely willing to pay a modest sum to the 'brokers' who set up calls for them.

indeed where they cannot be changed – and where little or no examination is made of call logging records. A competent communications manager will be abreast of all these issues already; lack of experience in these matters may make others the next victim. You cannot say you have not been warned!

Networks can fight back

Cerebrus, widely considered to be the most effective telecommunication fraud detection system available, is now even better at detecting pre-pay mobile phone fraud, according to its developers, Fraud Solutions.

The Cerebrus fraud detection and management system monitors up to 200 million calls each day for signs of fraudulent behaviour. It's a hybrid, and combines basic 'rules and thresholds' techniques – setting limits on spend or usage – with neural-network based artificial intelligence. In so doing Cerebrus identifies behavioural patterns – the signatures of fraud – and adaptively learns to detect new, unpredictable forms of fraud.

Cerebrus targets pre-pay fraud specifically by enabling operators to compare account use to their respective credit and debit levels. For example, if balances don't reduce in proportion to account usage there's a strong chance that fraud is taking place.

Mike Waddell, Fraud Solutions' Business Development Manager says, "The pre-pay sector has been caught out by its own success. Phones are available to people that couldn't previously qualify for one, such as infrequent users, teenagers and credit risks. This factor, combined with the customer's anonymity, has proved a fertile ground for fraud. Cerebrus gives companies the very best defence against pre-pay fraud."

Fast work

Phone fraud is characterised by its speed, simplicity and stealth; no physical access is needed to the victim's premises either. A total of 9400 stolen phone calls made between December 5 1999 and February 8 this year represented a loss of $50 000 to the US federal government, according to phone company officials.

A US army sergeant was arrested the same month and charged with giving out White House telephone access-code information that allowed individuals to make calls at government expense.

A British company providing a faxback service neglected to bar return calls to premium numbers. The result was a £12 000 phone bill in just three weeks – the people running the premium rate numbers share half of the phone company's proceeds. Other phraudsters make faster work; in the USA a Seattle business was stung for $100 000 of fraudulent calls in just one weekend.

Playing on ignorance – and greed

Many Americans have responded to a message that they immediately call or fax a number in the 809 area code. The incentive is to receive information about winning a wonderful prize, or getting a job, or else to avoid litigation or receive information about someone who has been arrested or died.

The number is indistinguishable from any other North American number but because the 809 area code is in the Caribbean, international call charges apply. The scamsters then try to keep their victims on the phone as long as possible, at charges reported as up to $25 per minute.

An identical fraud is not possible in Britain but a number of press advertisements for 'Sexy Susie' chat services format their overseas numbers in a way that disguises that they are actually in Guyana. There, the telephone administration pays back commission to operators of telephone information services in recognition of the enlarged incoming call traffic these numbers generate.

Low-tech works too

Not all telephone-related frauds involve high technology. There are other 'social engineering' dodges that can also cost phone users dear. One of these, rampant in New York a while back was master-minded by an elusive Russian named 'Serge'. It milked a fortune worth millions from various corporations on Madison Avenue. Serge and his colleagues set up two premium-rate '540' telephone numbers of the kind used by phone-sex lines. These were named 'Get Rich Fast Inc.' and 'Work for Yourself Inc.'

The gang despatched a fake messenger to pick up packages from reception desks. When told there was no parcel waiting, the 'messenger' asked the receptionist for permission to "call the office to see what is wrong". He would then dial the premium-rate number and hold a long and involved conversation – in Russian – at a cost of $2.25 per minute. The charge was automatically transferred by the New York telephone company from the business victim to the crooks' accounts'. It was not known how much the con-men made altogether, but they withdrew $240 000 in cash before the police discovered the front companies.

Easy peasy

A highly reputable telecomms consultancy tells of a client company that has already been ripped off for tens of thousands. With phone company assistance, the calls and the PBX ports used were traced and the culprits identified. The victim company refused to take action in court, however, for fear of bad publicity.

Direct Inward System Access is the feature that made the fraud possible. According to this consultancy, DISA is nothing less than a time-bomb and should not be employed.

"It's just too open to fraudsters," says our informant. "Today's phrackers can erase their tracks so cleverly, you cannot even detect they have been through the switch." The consultancy has purchased service manuals on every PBX in the country and claims it can 'hack' the lot of them. And if it's that easy for them, it's equally easy for the phrackers.

The consultancy also institutes unannounced security checks on its customers – with the blessing of the comms managers concerned – and manages to fool employees into revealing passwords every time, even when warned to be on their guard. People are so trusting! ■

Preventing fraud

These free booklets provide useful advice on preventing fraud:

'Guidance and Best Practice for the Avoidance of Dial-Through Fraud', from industry body BABT, phone 01932-251200.

'Avoiding PBX Fraud – working actively to safeguard your system', from phone company Telewest, phone 0118-945 8522.

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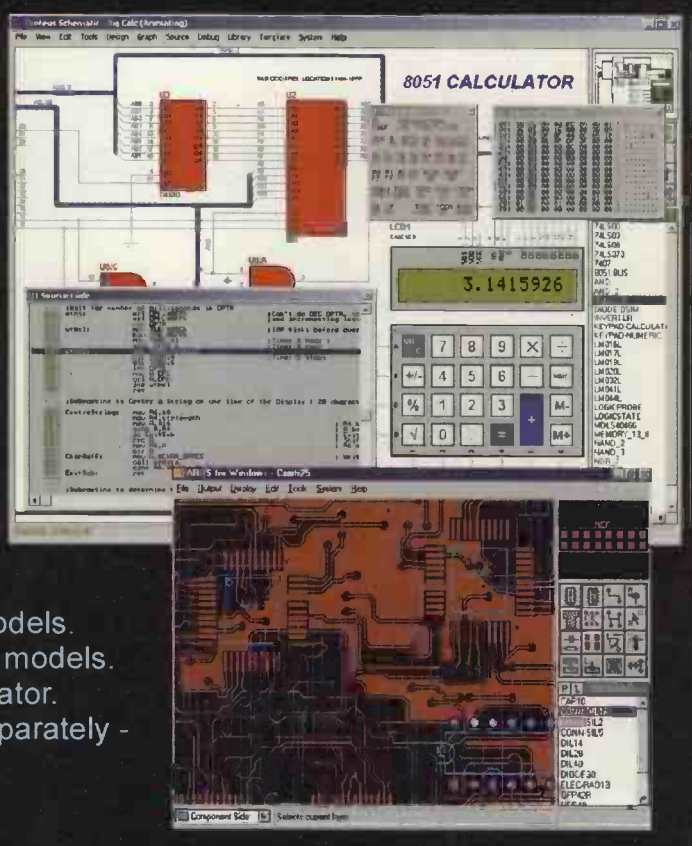

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Paolo Antoniazzi and Marco Arecco have been addressing the challenge of transmitting and receiving on the Europe-wide amateur-radio allocation at 136kHz – a frequency low enough to allow an audio power amplifier to be used as a transmitter.

Comms at 136kHz

Fig. 1. At the base of the Marconi vertical antenna for 136kHz.

Europe's new 135.7-137.8kHz LF amateur band was launched on 30 January 1998 in the UK. This low-frequency wireless band, and its 160-190kHz counterpart in the USA, presents unusual challenges. Working at a wavelength of 2206m for example, a hypothetical quarter-wave antenna would be a 550m high tower!

But this part of the RF spectrum is a wonderful place for experimenters. Only quasi-audio type instrumentation and lots of wire are needed to start.

Combining the modern PC's processing capability with old-style large coils and big vertical antennas with capacitive 'hats', you can obtain very interesting results[1].

To explain how, we will use a step by step method starting with transmission antennas. In the following, we analyse grounding, propagation, noise and load coils. To round the article off, we present a 136kHz mini-system suited for beginners and for preliminary tests in the field.

Very-short vertical antennas

It is very difficult to realise a quarter-wave vertical antenna because the wavelength corresponding to 136kHz is 2206m.

A practicable dimension would be in the 7 to 10m range, assuming that most people have a garden or some roof space. However, efficiency resulting from such a short antenna would be very small – less than 0.1%. For comparison, the height for a similar antenna used for the 14MHz band should be 7 to 10cm.

More complete technical documentation on the big aerials, are available from reference 2 – 'The monster antennas' – and reference 3.

Power radiated by a vertical antenna is related to radiation resistance R_r, effective value of antenna base current I_b and antenna directivity G_A.

The first way of improving radiation resistance is to make a taller antenna, but managing a 15 to 20m high tower is not easy. The radiation resistance of a short vertical antenna is given by the following equation,[4]

$$R_r = \frac{40\pi^2 h^2}{\lambda^2}$$

Here, R_r is radiation resistance in ohms, h is antenna height in metres, and λ is wavelength, which is 2206m at 136kHz.

This kind of antenna has a triangular current distribution: zero at the top and I_b at the base, i.e. the feed point. It is not difficult to show that the efficiency η of this kind of arrangement is very poor and the resulting radiation resistance is only few milliohms,[5]

$$\eta \approx \frac{R_r}{R_L + R_G}$$

considering that,

$$R_r << R_L + R_G$$

where R_L is coil resistance in ohms and R_G is ground resistance, also in ohms.

To improve the antenna efficiency it is necessary to increase the radiation resistance and antenna current distribution. This can be done by putting a capacitance at the top of the antenna, also known as an antenna's 'hat.' This kind of capacitance is obtained using one or more wires connected to the upper end of the vertical rod.

Another way of improving antenna efficiency is to minimise coil and ground losses R_L and R_G, but the reduction of the last one is not so easy.

The following relationship gives the distributed capacitance, of the vertical part of the antenna:

$$C_v = \frac{24h}{\log\left(\frac{1.15h}{d_v}\right)}$$

This will be around 10pF/m, considering the antenna dimensions involved here. In this equation, C_V is antenna vertical capacitance in picofarads while d_V is antenna vertical rod diameter in metres.

The following gives the horizontal distributed capacitance,

$$C_H = \frac{24l}{\log\frac{4h}{d_H}}$$

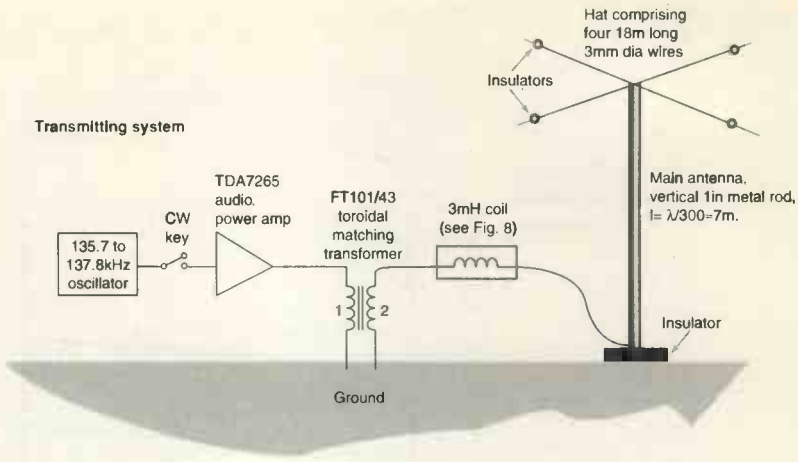
Here, C_H is antenna horizontal capacitance in picofarads, l is the antenna's horizontal wire, i.e. hat, length in metres, and d_H is horizontal wire diameter in metres. For the example involved here, this will be about 5pF/m.

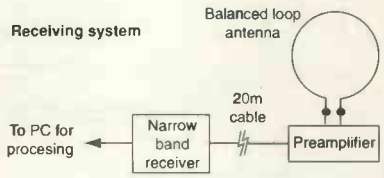
Current at the top of the vertical antenna, assumed to have the same current increase versus capacitance as the whole antenna, is given by the following:

$$I_t = \frac{C_H I_b}{C_H C_V}$$

Here, I_t is current at the top of the antenna in amps while I_b is current supplied by the transmitter at the antenna feed point, also in amps.

These considerations allow you to calculate the radiation resistance of a vertical antenna with hat using,





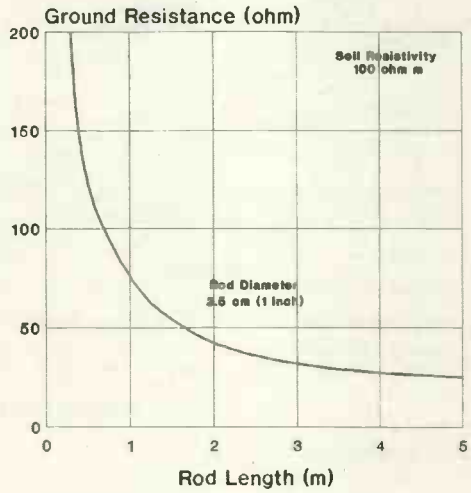


Fig. 2. Ground resistance versus rod length in a typical earth 100Ω/m.

$$R_r = 40\pi^2\left(\frac{2C_H + C_V}{C_H + C_V}\right)^2 \times \frac{h^2}{\lambda^2}$$

At this point, you can calculate the effective power radiated by the antenna discussed here, considering that G_A is 3dB – the gain of a short vertical antenna – and not the theoretical 4.78dB[6].

$$P_r = R_r I_b^2 G_A$$

Here, P_r is effective power radiated by the antenna in W_{erp}.

To check the complete system, a relatively simple antenna has been constructed with a vertical rod of 7m isolated from ground by a plastic plate, Fig. 1. This antenna was implemented using 1in (25.4mm) stainless-steel tubes loaded at the upper end by 70m of 3mm diameter copper wire.

This kind of arrangement has a total capacitance of 450pF. It allows trapezoidal current distribution on the antenna with a ratio between the top and the base currents of about 0.85. A consequence of the capacitive hat is a radiation resistance estimated at about 14mΩ. This corresponds to an effective radiated power increase of about 5dB relative to the same antenna without hat, in which case, R_r is 4mΩ.

Table 1 shows the radiation resistance calculated in three short vertical antennas: R_{r1} is without hat, R_{r2} is with a per-

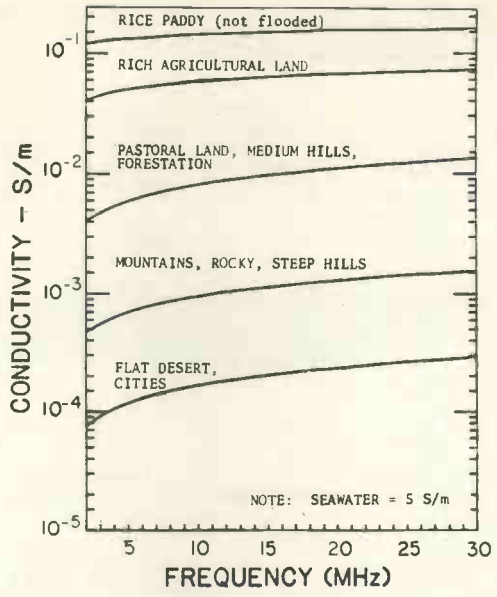


Fig. 3. Ground conductivity for various terrains.

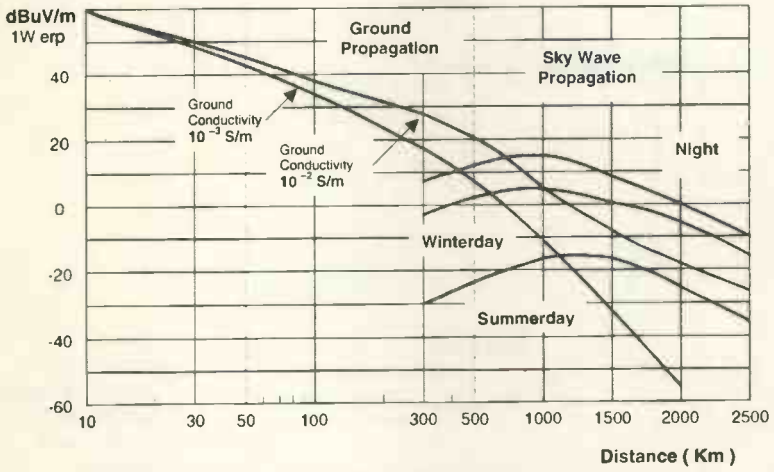


Fig. 4. CCIR ground and sky-wave propagation at 136kHz.

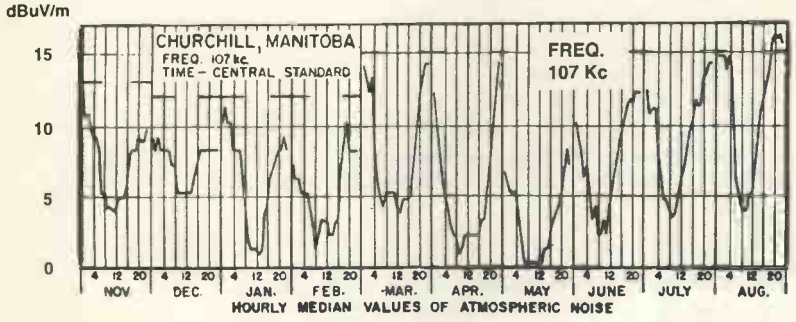


Fig. 5. Atmospheric noise measurements for F=107kHz and BW=4.5kHz.

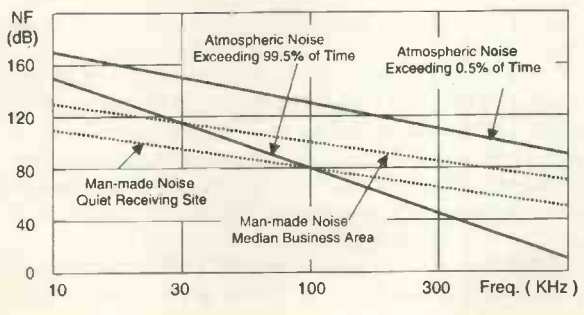


Fig. 6. Noise levels – atmospheric plus man-made.

fect hat, and R_{r3} involves a normal hat. Using radiation resistance R_{r3}, which is the most practicable, we computed the effective radiated power considering an effective base current of 0.5A (P_{O1}) or 4A (P_{O2}).

We tuned our antenna by making the loading coil together with the total antenna capacitance resonant at 136kHz. The antenna bandwidth, considering the total circuit resistance, R_L+R_G, is about 800Hz. To fine tune it, we used an additional inductor of 250µH selectable in 50µH steps.

The importance of a good ground

A big problem with LF is making an effective ground connection. If you have a country house with a big garden, you probably won't have grounding problems. As is well known, earth is inherently poor conductor. It normally has resistivities in the range of 10 to 1000Ω per metre, or 10^{-1} to 10^{-3}S/m, so the conductivity of the metal used for the earthing rod is not too important.

Ground resistance R_G can be viewed as the resistance resulting from a series of equally thick concentric shells of earth around the ground rod. With a typical 3m rod, half of the resistance is contained within a cylinder of 10-15cm radius around the rod[7].

The only way to reduce the ground resistance is to use multiple electrodes. Adding more ground rods reduces the earthing resistance, but the final resistance is higher than the value simply obtained by dividing the resistance of a single rod by the number of parallel rods.

A single 3m rod, 25mm in diameter, driven into soil with 100Ω/m average resistivity, **Fig. 2**, will have an earthing impedance of about 30Ω measured at 50-60Hz. Using four rods in parallel, placed at 10-15m in a square, will give a final LF resistance of 10-15Ω.

At 136kHz, the inductance of the connecting cable is not important, but a large wire is needed to avoid skin-effect resistances. Using for example the 42 by 0.18mm Litz wires that we used for the coils results in just 0.0164Ω/m DC resistance. This equates to 0.2Ω of RF resistance for a 10 metre run.

When using four or more parallel ground connections, the resistance of the wires is not too important. We used standard 3mm^2 flexible copper wire.

In our tests 2 by 4m deep rods and 2 by 2m deep rods were used at a distance from the common ground point – at the base of the Marconi vertical antenna – of 10-12m. The measured value of our ground resistance, R_G, at 136kHz was 11 to 14Ω.

For any type of electrode that may be used to connect the system to earth, its ground resistance is directly proportional to the resistivity of the soil. Knowing the earth resistance and impedance is very important since it governs the efficiency of the complete system, **Fig. 3**. See references 8 and 9 for more information on the electrical characteristics of the surface of the earth. The world atlas of ground conductivities is available from reference 10.

Distance covered

Propagation at 136kHz is mainly groundwave during the day with some skywave propagation at night, **Fig. 4**. The skywave behaviour is calculated using the CCIR simplified model[11] for typical ground conductivity. Achievable range depends very much on the transmitting station's capabilities. The better equipped stations now achieve several hundred kilometres in daytime and up to 1000 to 2000km at night.

Most of the available information on LF propagation originated from an old book by A. Watt[12], a well known paper from J. A. Adcock[6] and from practical field experiments.

Low-frequency propagation – at 10 to 200kHz – differs

from high-frequency propagation in a number of respects. Surface wave propagation is very strong. There is no skip zone, although low angle sky wave radiation is dominant. This is because of the low height of the reflecting plane and the long distance travelled by the surface wave.

Where the surface wave meets the sky wave, the two merge together, possibly with some cancelling or adding of the two where the strengths are equal.

From 200kHz downwards, both surface and sky wave improve. The result of this is that the zone where the surface wave equals the sky wave remains between 500 and 1000km. At 136 to 200kHz, propagation is poor in the day and reasonable at night. Between 10 and 30kHz, propagation is excellent both day and night.

Propagation issues

All LF transmissions are vertically polarised. In fact, it is almost impossible to radiate any horizontal component unless you are transmitting from an aircraft. From an LF perspective, the main limiting factor is the noise level.

The frequencies where propagation is best is also the place where noise is highest. This is the main reason for very high powers being used for very long distance communications by commercial and military stations. Another reason for high powers being used is poor antenna efficiency.

At LF, ground loss is very low. At all low frequencies, it only starts to become significant at distances of more than 500km, even over poor ground.

Although surface wave propagation depends on the fairly basic physical phenomenon of diffraction, calculating path loss depends upon a number of factors. Path loss graphs from 10kHz to 30MHz are given by CCIR/ITU[13].

At all radio frequencies, the surface wave is composed of several components. The most significant waves are direct, reflected and a diffracted wave derived from the edge of the wave shadow. At LF, the direct and reflected wave[6] at a low angle to the ground are opposite in phase and totally cancel each other. This leaves the diffracted wave as the dominant wave.

Gain of a short dipole is 1.78dBi – i.e. 1.78dB relative to an isotropic – at right angles to the wire. If the antenna is above a perfect ground, the lower half of the radiated energy will be added to the upper half and this adds another 3dB.

In normal operation on long wave, the antenna is always vertical and always short. This means that the gain is always 4.78dBi, or threefold. Due to losses of the signal to the ground and to supplying the surface wave by diffraction, the lower edge of the radiation pattern is depleted. In our calculations for Table 1 we used only 3dB for G_A.

The characteristic of loss of signal to ground is known as 'cut back'. At LF, cut back is less than at HF. Low-angle signals bend due to the diffraction that produces the surface wave, and the sky wave path just above is also curved. For an ionospheric reflecting layer 90km high, the expected maximum length of a single hop sky wave is about 2000km. In practice, the hop can be usefully extended to more than 2800km.

Very interesting field-strength measurements in the 136kHz band and on DCF39 – a German 138.830kHz station with 40kW radiated power – are shown on the web site of DK8KW[14].

Band limiting beats the noise

The final factor and the biggest problem in the band is noise. No matter how strong a signal is, the signal-to-noise ratio is always the limiting factor. This noise comes mostly from atmospheric electrical discharges throughout the whole world. It is applicable whatever real antenna is used for receiving.

The only real antidote to the noise is to work with an extremely low bandwidth – much lower than the 100-250Hz used by HF people.

For the tests, we used an interesting high-quality 10kHz IF filter designed for telecoms. It has a –3dB bandwidth of 20Hz and is still only ±50Hz wide at –50dB. It attenuates an interfering signal separated by 100Hz by more than 90dB.

Slow CW

To overcome the noise limits, many stations use the so-called 'slow-CW' to operate in the LF bands. Slow-CW allows you to detect signals that are far below the noise levels and can not be detected by ear. Using a personal computer, an audio card and specific spectrum analyser (FFT) software like Spectrogram 5 for example, signals 20dB below the noise floor can be detected! Typical dot lengths of 3 to 5 seconds have been established to do the job.

Atmospheric noise at LF is high and increases at a very high rate with decreasing frequency. Noise maps and curves are given in texts on the subject.

Typical noise figures are given in **Table 2**. These are intended as a guide to receiving noise level on 136-190kHz using CW reception with 100Hz and 25Hz bandwidth. The figures are relative to 1µV/m.

More scientific data on LF noise are available from tests made in 1957[15,16] in Canada. These tests were carried out at 107kHz with a bandwidth of 4500Hz and are shown in **Fig.**

Table 1. Radiation resistance and radiated power for Marconi antennas at 136kHz.

Height (m)	(inches)	Degree	R_{r1} (Ω)	R_{r2} (Ω)	R_3(*) (Ω)	P_{O1} (mW)	P_{O2} (mW)
5.0	197	0.82	0.0020	0.0081	0.0070	3.49	224
6.0	236	0.98	0.0029	0.0117	0.0101	5.04	323
7.0	276	1.14	0.0040	0.0159	0.0139	6.95	445
10.0	394	1.63	0.0081	0.0324	0.0282	14.10	902
14.0	551	2.28	0.0159	0.0636	0.0554	27.69	1773
20.0	787	3.26	0.0324	0.1298	0.1129	56.46	3613
28.0	1102	4.56	0.0636	0.2544	0.2216	110.80	7091

G_A = 3dB, comprising gain of short vertical plus ground reflection.

* $\frac{I_{top}}{I_{base}} = 0.85$

136kHz amateur band

In the UK, the introduction of the 'European Harmonised Amateur Low Frequency' spectrum allocation took place in January 1998. It spreads from 135.7 to 137.8kHz and replaces the old 73kHz allocation, which was withdrawn in June 2000.

Note that only amateurs with a Class-A licence are allowed to operate equipment at frequencies below 30MHz.

Table. 2. Typical LF noise levels at 136 to 190kHz – referred to 1µV/m.

Conditions	BW = 100Hz	BW = 25Hz
Summer day	–15 dB	–21 dB
Summer night	–5 dB	–11 dB
Winter day	–30 dB	–36 dB
Winter night	–18 dB	–24 dB

5. Of course, man-made noise is also important, but it is unpredictable,[13] **Fig. 6**.

Interference on the 136kHz band has been caused by high-speed data carried by telephone networks being radiated due to incorrect screening practices. ADSL carrying internet traffic at high speed over normal twisted pair telephone lines is likely to compound the problem. It uses 256 modulated carrier frequencies from 26kHz up to 1.5MHz. It is not yet clear whether there will be serious interference problems from ADSL though.

It is likely that noise from these sources will increase in time, as the commercial advantages of high speed internet and other data links seem to outweigh many other considerations. Bear in mind that the balanced driver that sends ADSL signals over the 120Ω line – i.e. the telephone's twisted pair – has an output of +20dBm.

Fig. 7. Minimum coil wire length versus coil diameter and D/Le – a starting point for realising high-current, high-quality inductors.

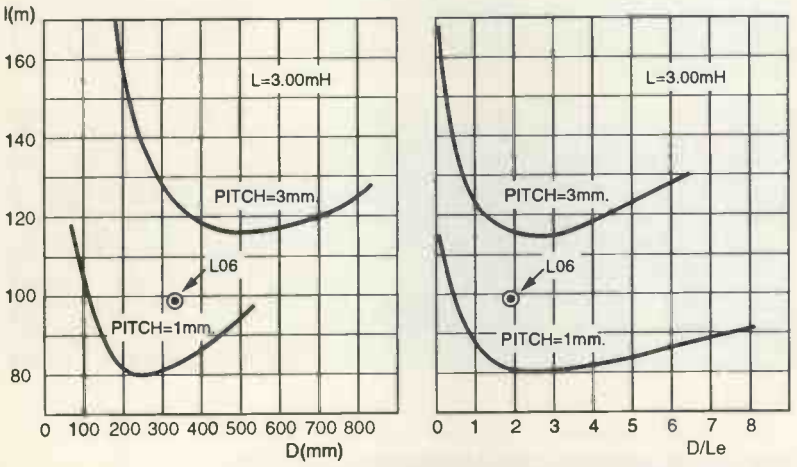


Fig. 8. Final load inductor – 3mH, with a Q of 600 at 136kHz.

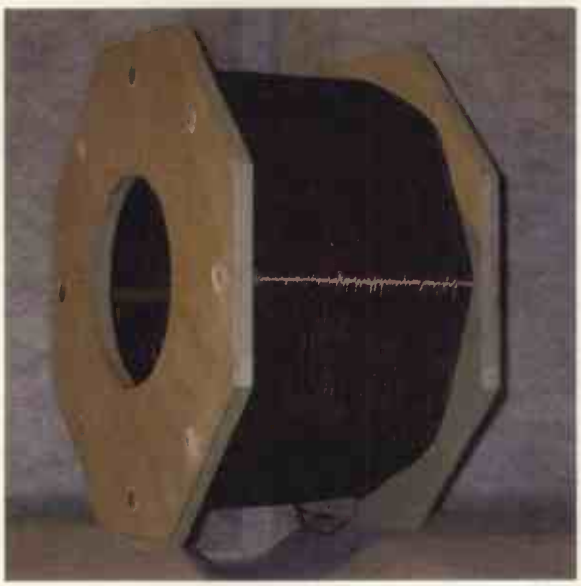
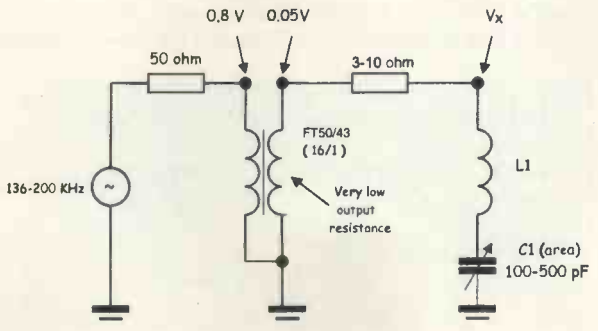


Fig. 9. Circuit for simple and accurate measurements of quality factor at 136 to 200kHz.



The art of high-Q coils

The simplest form of loading network is a coil in series with the vertical Marconi-type antenna. This coil tunes the capacitive reactance of the antenna and matches the feed point down to a reasonable level of 15 to 50Ω.

As already mentioned, a way to improve the antenna efficiency is to drastically reduce the series resistance of the coil – down to less than 5Ω for example – by designing an inductor having a very high quality factor 'Q'.

This equation defines coil quality factor,

$$Q = \frac{2\pi fL}{R_{AC}}$$

Here, f is frequency in kilohertz, L is inductance in millihenries and R_{AC} is the LF equivalent series resistance in ohms.

Losses that impact the quality factor of an LF coil are:

- skin effect of the wires
- proximity effect between contiguous winding turns
- dielectric properties of the distributed capacitance
- coil-support material electrical performance

The resistance that a copper wire presents against a direct current flowing is given by,

$$R_{DC} = \frac{2.23 \times 10^{-2}}{d^2}$$

Here, R_{DC} is DC resistance for unit of length in Ω/m and d is the wire diameter in millimetres.

If the current alternates, the inner part of the wire has a higher reactance than in the previous case and the wire resistance becomes,

$$R_{AC} = 8.374 \times 10^{-2} \frac{\sqrt{f}}{d}$$

where R_{AC} is the AC resistance for unit of length in Ω/m and f is frequency in megahertz. A more accurate computation of the resistance in alternating current, namely skin effect, is reported in reference 17.

To reduce the resistance generated by skin effect, Litz wire is used. It comprises a lot of thin insulated wires connected together at the ends.

When two or more adjacent wires are carrying current, the current distribution in every conductor is submitted to the magnetic field generated by the adjacent wires. This effect, namely proximity effect, significantly increases the value of R_{AC} calculated previously.

While experimenting, we verified the impact of this effect on the coil quality factor. Have a look at L04 and L05 data in **Table 3**. There, the 'Q' increase is related to the pitch increase from 1 to 2mm.

Dielectric losses occur due to the material used to insulate the winding's conductors, enamel for instance. Such losses are negligible though in relation to the total capacitance needed for resonance at LF.

During the experimental phase, we tried using common grey PVC tube. We found this to be the worst material, as you will see from L01 in Table 3. Compare that coil's quality factor with that of the coils wound on wood and air.

The best coil was one wound on a support comprising eight cylindrical pieces of wood connected together by two

Table 3. Specifications of high-Q coils we tested for use at LF.

Coil type	Coil dia	Wire dia	Litz	Turns	Coil L_e	D/L_e	Wire length	L	R_{DC}	136kHz			200kHz		
		(mm)	(mm)	(cores)	(mm)		(m)	(mH)	(Ω)	X_L (Ω)	Q (*)	R_{AC} (Ω)	X_L (Ω)	Q (*)	R_{AC} (Ω)
Grey PVC	160	0.90	–	175	176	0.91	88	3.12	2.43	2665	205	13.0	3919	187	21.0
Air+wood	160	0.18	30	120	210	0.76	60	1.40	1.38	1196	513	2.33	1758	534	3.29
Air +wood	200	0.18	42	157	272	0.73	99	2.70	1.63	2306	507	4.55	3391	435	7.80
Air +wood	330	0.90	–	85	87	3.80	88	3.30	2.43	2818	237	11.9	4145	230	18.0
Air +wood	330	0.90	–	85	165	2.00	88	2.50	2.43	2135	318	6.71	3140	309	10.16
Air +wood	330	0.18	42	94	168	1.96	97	3.00	1.59	2562	597	4.29	3768	541	6.97

$$*Q = \frac{X_L}{R_{AC}}$$

plates of wood with a hole in the centre. This kind of support minimises the mass of material – i.e. wood – within the inductor winding in order to reduce losses.

The inductance of an LF coil can be calculated using,

$$L = 2.53 \times \frac{10^7}{f^2 C_A}$$

where L is coil inductance in millihenries, f is frequency in kilohertz and C_A is total aerial capacitance comprising vertical + hat capacitances, in picofarads.

A further method of minimising the winding series resistance is to have the minimum wire length that corresponds to have a ratio of 2.2 between the coil diameter and its effective length, i.e. D/L_e, **Fig. 7**. You can vary this ratio from 1 to about 4.8 without a drastic worsening of the inductor performances.

At this point you can calculate the number of turns of the winding using,:

$$N = \frac{5.08 \times 10^5 Lp + \sqrt{(5.08 \times 10^5 Lp)^2 + 4.572 \times 10^5 LD^3}}{D^2}$$

Here, N is the number of turns of the coil, L is the value of loading inductance in millihenries, p is the pitch between two adjacent turns in millimetres and D is inductor diameter, also in millimetres. Pitch p is equal to, or more than, the wire diameter including its insulation.

The above formulas allow you to calculate the inductance of a single-layer coil with an accuracy of 1%, as confirmed by the experimental results.

Coil performance tests

Table 3 shows the results from all the inductors we tried during our evaluation. Coil L06 was the last one tested and it summarises the experience we achieved in this field.

This inductor, **Fig. 8**, has been used to load the 136kHz transmitter to the antenna and ground arrangements previously described.

After building the coils, we needed to perform the merit factor measurements in order to confirm the theoretical results of our study. To do this, we developed an original method for minimising losses caused by making measurements. The advantage of our novel test circuit, **Fig. 9**, is that it measures the series resonance instead of the parallel one.

When voltage V_X is one half of the voltage at the output of the transformer, the resistance is equal to R_{AC}. The relevant value, R_S, can be measured by a simple digital ohmmeter. Caution is needed while making the measurements to prevent errors. It is important that the coil under test and its magnetic field are far away from metallic surfaces and do not intercept external unwanted magnetic fields.

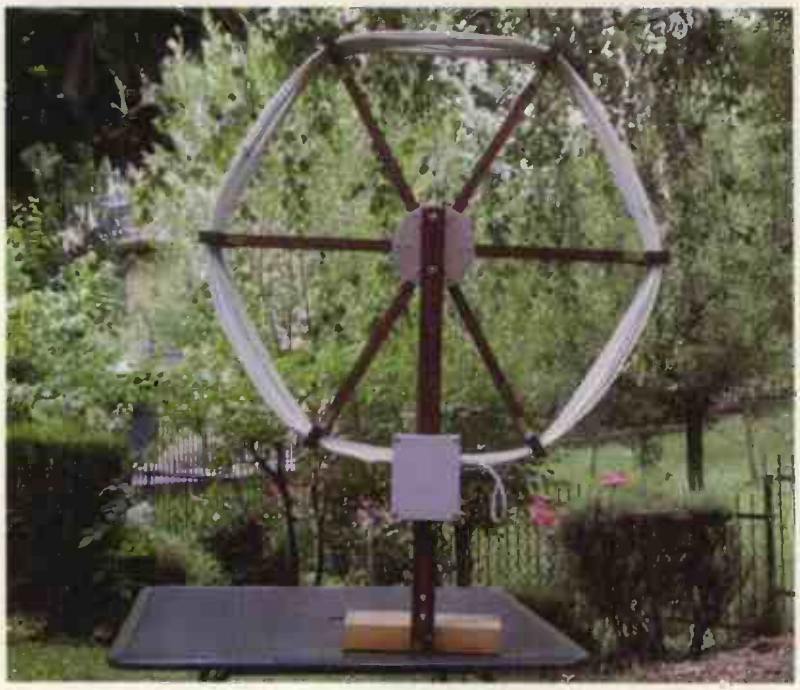


Fig. 10. A group of large coils that we used for the Q factor tests.

Such large coils become a loop receiving antenna so the measurements must be repeated with the winding in different positions and orientations to allow errors to be detected. Another thing worth considering when making measurements is whether the length of the antenna's connecting wire is affecting the test result.

Considering the number of possible error sources, we suggest that you use a reference inductor to check the quality of the measurements performed. For this purpose we used a Boonton shielded coil, with an inductance of 2.5mH and a Q of170. It is visible in **Fig. 10** at the top right, together with the group of inductors made.

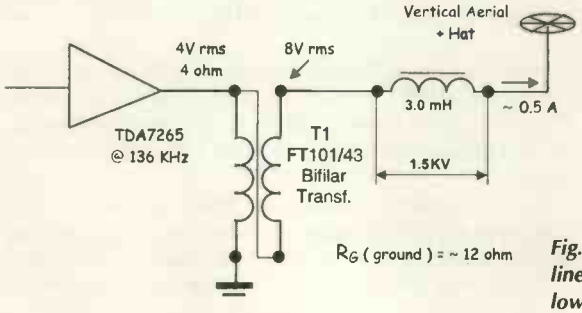


Fig. 11. Simple 136kHz linear amplifier using a low-cost monolithic IC.

Fig. 12. Shielded and tuned LF loop antenna comprising 18 turns of TV-satellite cable, together with a balanced amplifier.

A starting point for work at 136kHz

A very simple transmission circuit for use at 136kHz is shown in Fig. 11. It includes a common audio power amplifier used in high-quality TV sets, namely the TDA7265 from STMicroelectronics. A printed circuit board and more technical information on this device are available on the web site at www.st.com.

At audio frequencies, typical output power of this amplifier is 25+25W with 4Ω speakers. At 136kHz the maximum output power drops to about 4W. This output power is more than adequate for LF system tests, generating up to 0.5A of antenna current.

Transformer T_1 is implemented using a standard FT101-43 one inch toroidal core. It is used to match the required 4Ω load to the 16-18Ω total resistance of the system, i.e. the antenna plus load coil plus ground.

The coil selected for use as a load for the Marconi antenna is shown in Fig. 8. It has an inductance of 3mH and a Q of 600. Total loading seen by the output transformer, i.e. at the base of the 7m long vertical antenna, comprises 4.3Ω of the coil, R_L, and 12Ω for the ground, R_G, i.e. 16.3Ω. Loading on the amplifier is 4Ω.

You can see the base of the antenna and its load coil in the Fig. 1.

Receiving antenna

In the receiving section of the system, the shielded loop antenna, Fig. 12, performs best. A tuned-loop antenna provides the required front-end selectivity. A balanced input preamplifier using a TSH151 low-noise operational amplifier with power bandwidth up to 200kHz provides the required gain.

We use a gain of 26dB to provide a +6dB for the output matched to 50/75Ω. In effect the shield, made using satellite TV coaxial cable, is not so important. It is also a limiting factor as it reduces the number of turns, lowering the coil's Q factor.

The shielded-loop version in the photo was used for tests. Since then we have started to produce a new, larger loop that is well balanced and doesn't have any shield. A directional magnetic loop, sensitive only to the magnetic component of the field, will help to eliminate the effects of the various near-field sources of interferences – particularly man-made noise[1].

Our first loop antenna, Fig. 12, has an area of 0.48m^2 and uses 18 turns, resulting in a Q of 50. Its equivalent height at 136kHz, where λ is 2206m, is,

$$h_e = 2\pi AN\frac{Q}{\lambda} = 6.28\times 0.48\times 18\times\frac{50}{2206} = 1.23\text{m}$$

At 136kHz, the operational Q is always near 50. To have a working bandwidth of not less than 2.1kHz from 135.7 to 136.8kHz, all the parameters of the loop can be fixed, apart from A and N.

The simplified formula for the equivalent height of receiving loops at 136kHz is,

$$h_e = K\times AN = 0.142\times AN$$

For the new version of the loop, with an area of 2.0m^2, 44 turns and the same Q, the effective height will be,

$$h_e = K\times AN = 0.142\times AN = 0.142\times 44\times 2 = 12.5\text{m}$$

Note that this is a 20dB more gain! More turns is good, more diameter is even better.

W.E. Payne, N4YWK, suggests that the product of A and N is the only term that describes the characteristics of the loop itself. A figure of merit is the effective aperture, A_e, which is the physical area multiplied by the number of turns[18]. ■

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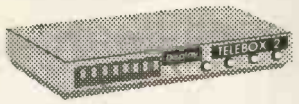

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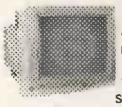

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High-frequency impedance meter

Measuring a capacitor's value at low frequencies tells you little about how it is likely to perform in, say, a switch-mode power supply, where the device's inductive components come into play. Here, Cyril Bateman describes an instrument for looking at a capacitor's impedance at anything from audio frequencies to several megahertz. It reads from a few milliohms to almost 2Ω on a low-cost digital panel meter.

The ever increasing demand for small, lightweight, efficient equipment, has resulted in an explosion in the use of minute surface mount packages, encouraging designers to use physically small capacitors.

For a given capacitance and working voltage, the smallest physical size usually results from using either high-K ceramic chips, or tantalum or aluminium electrolytic capacitors. These possess three undesirable attributes.[1]

- As frequency increases, their apparent, measurable capacitance reduces. It can be much smaller than the marked, low-frequency, nominal value.
- For a given *CV* product, case size reduction invariably increases the ESR of the capacitor.
- These styles exhibit measurable self-inductance.

Resonant capacitors

At some frequency, the capacitive and inductive reactances become equal and opposite and so cancel. The phase angle then measures zero. At this resonance frequency, and for one octave lower, the apparent or measured capacitance increases dramatically, tending ultimately towards infinity at resonance. Measured impedance is then equal to the capacitor's ESR.

Capacitance increases that are measured near the device's resonance are not real. They result from an *LCR* meter simply converting measured impedance and phase into a capacitance value.

With increasing frequency, the actual or true capacitance values reduce and the ESR of an electrolytic capacitor can exceed its capacitive reactance. When this happens, the capacitor's measured phase angle becomes small – only a few degrees. The measured impedance curve then appears flat bottomed over a wide frequency band.

Above resonance, the inductive reactance dominates over the capacitive reactance, so the capacitor's measured impedance increases with frequency, **Fig. 1**.

Some capacitor manufacturers provide nominal impedance or ESR values for their ranges. These are usually specified at 100kHz and room temperature.

High-frequency capacitance and inductance values are rarely stated. In any case it is better to perform your own comparative measurements. If you do, you can use frequencies and temperatures that are appropriate to your application, ensuring a more useful comparison between capacitor makes and types.

Measuring *C* at high frequencies

A suitable, variable-frequency *LCR* meter allows parameters that change with frequency to be determined accurately. But

such meters can be extremely expensive.

However, such measurements are also possible using simple methods and low cost laboratory instruments.

In my last article,[2] 'Method 2' described how accurate measurements of impedance and reactance are possible using only a suitable test jig with current sensing resistor, a signal generator, a phase meter and a high-impedance RF millivoltmeter. The current-sensing resistor used should be close in value to the mean impedances to be measured. It must be non-inductive and its true resistance value must be accurately known.

Unfortunately, using this equipment, impedance values are not directly measurable. Several calculations are needed to convert the measured parameters. Method 2 is outlined in a separate panel entitled 'Determining capacitor impedance'.

I also introduced my improved Method 3, which *does* produce a direct readout of impedance. It needs no calculations and is usable from audio frequency to several megahertz.

This method depends on accurately measuring the differential voltage drop across the capacitor using a wideband, differential input, millivoltmeter. Divided by the capacitor's through current, this voltage represents the device's impedance.

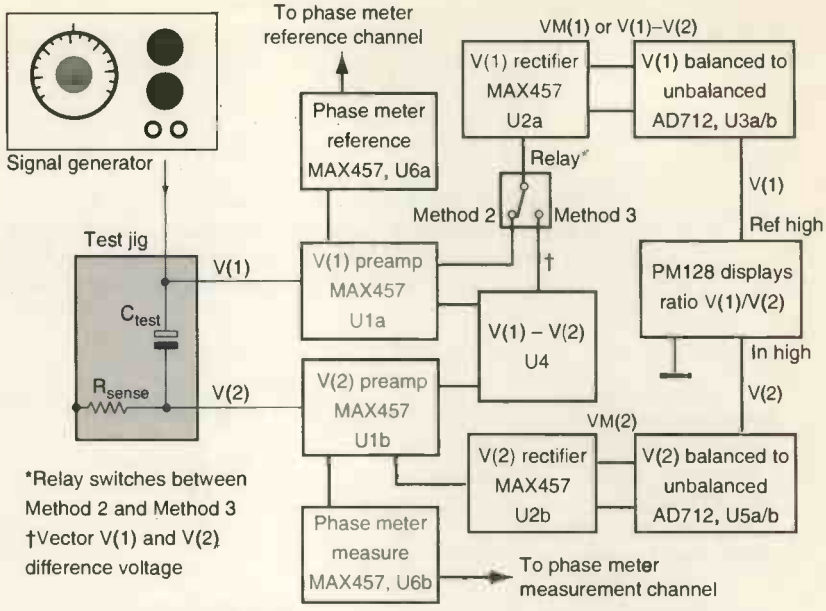
Capacitor through current is most easily determined by measuring the voltage drop across a small ground-return sense resistor whose value is known[3].

To confirm these methods, I modelled a theoretical 200µF capacitor having an ESR of 0.05Ω, measured using a 0.4995Ω sense resistor. The simulation shows this direct derivation of impedance with frequency, and confirms the accuracy of both techniques, **Fig. 2**.

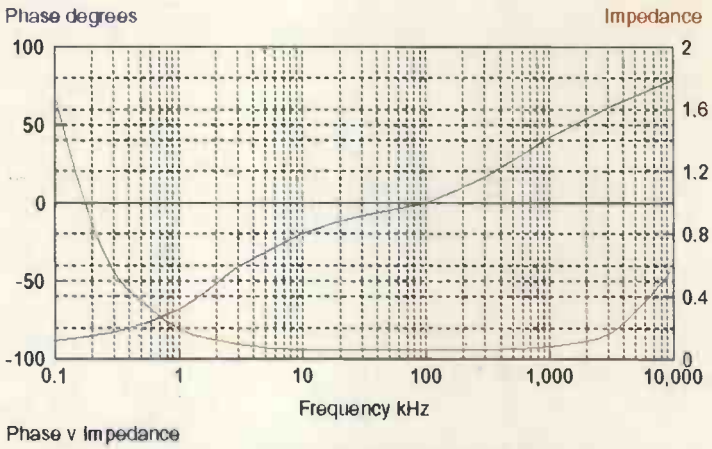
This article details the design of my dedicated impedance meter, which is usable from audio frequencies to several megahertz, measuring from a few milliohms to 1.999Ω.

Circuit concept

As a starting point I considered the circuit of my RF milli-



The meter is self-contained except for power supply, and needs only a signal generator to supply the test signal Use of a phase meter is optional.



1000 µF 25v Philips 135 Capacitor.

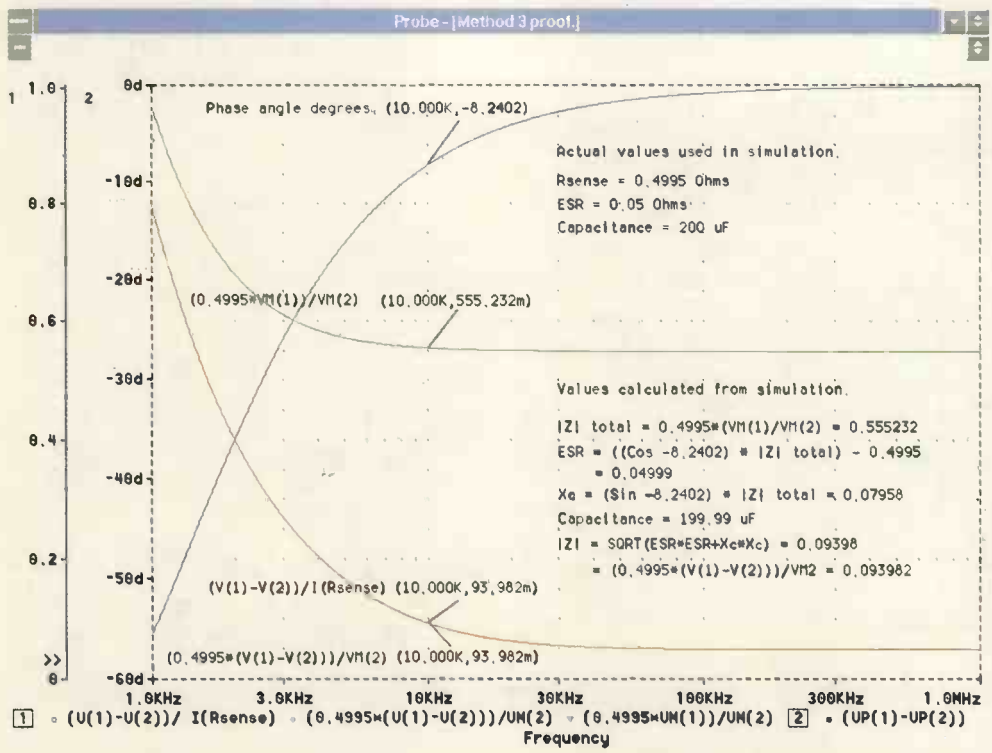


Fig. 1. Impedance measured using the |Z| meter with phase angles measured using the meter's phase buffer output. The capacitor under test was mounted on my test jig as in Fig. 7.

Fig. 2. Simulation used to verify my test methods. Impedance was plotted two ways, the classic 'voltage drop/current' and also 'sense-resistor-value×vector-difference-voltage/sense-resistor-voltage-magnitude'. Both curves overlay exactly in the plot. The method 2, 'sense resistor×VM1/VM2' ratio and capacitor phase angle, needed to calculate ESR and capacitance, is also plotted.

voltmeter[4]. This is based on a Maxim 457, a high-frequency, high input impedance, low input capacitance, dual op-amp in a standard 8-pin DIL package. It provides differential phase and gain of 0.2° and 0.5% respectively. The device slews at 150V/µs and can drive a 1V signal into 75Ω.

To measure this differential voltage, I needed at least two similar input circuits, followed by a differential amplifier and a high-frequency rectifier. But I also need to measure the voltage magnitude at the series resistor, and then perform a division of these voltages,

As already proved in Fig. 2,

$$|Z| = R_{sense} \times \frac{V(1) - V(2)}{VM(2)}$$

Here, $V(1)$ and $V(2)$ are the complex difference voltages while $V_M(2)$ is the voltage drop across the capacitor current sense '*R*', **Fig. 3**.

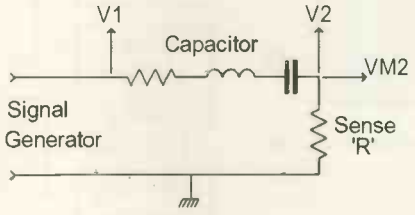


Fig. 3. Three-component equivalent circuit of a capacitor. Measurement points V1, V2 and VM2 are identified. The sense resistor should be non-inductive and comparable in resistance to the range of impedances to be measured.

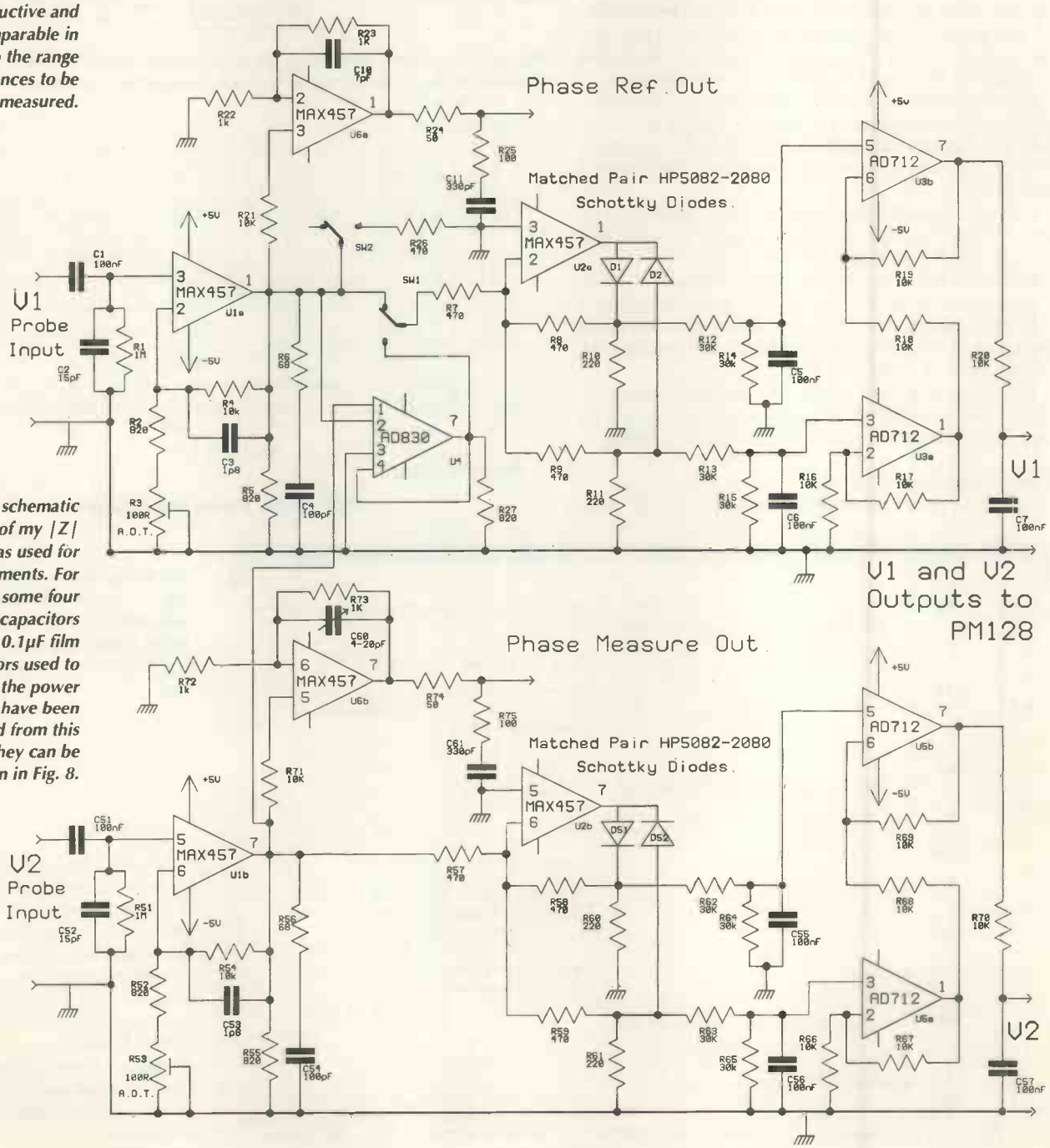


Fig. 4. Full schematic drawing of my |Z| meter as used for measurements. For clarity, some four electrolytic capacitors and ten 0.1µF film capacitors used to decouple the power supply rails have been omitted from this drawing. They can be seen in Fig. 8.

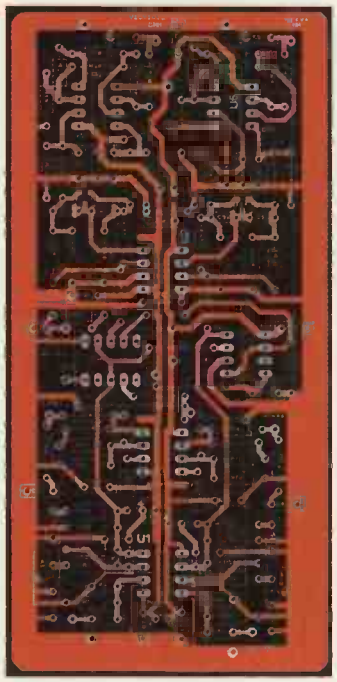


Fig. 5. The revised printed board pattern used for my prototype |Z| meter. Notice the additional dummy pads at U_2 and U_5 to equalise board capacitances between the two channels. Also note the guard tracks around sensitive pins of U_1. Together with mirror imaging of tracks, these measures overcame the phase difficulty of my original board.

I decided to investigate this approach using two matched high impedance input channels, similar to the one in my RF millivoltmeter[4]. These were followed by two of the rectifying stages used in the same millivoltmeter.

The rectified DC outputs were divided, using a PM128 panel meter, modified to ratio mode, as applied to my tanδ meter to give a direct display of measured impedance. The tanδ meter mentioned was described in the January 2000 issue.

A relay could select between measuring the two input channels, or the differential measurement could be taken, as required. This displays either measured impedance as the vector result of $(V(1)–V(2))/VM(2)$, or $VM(1)/VM(2)$ for Method 2.

By adding a dual-channel buffer immediately following the preamplifier circuits, I could supply an amplified, phase equalised output to my phase meter. Thus the |Z| meter could be used simply to measure impedance at a given frequency.

Alternatively, when combined with a phase meter, it could be used to calculate capacitance or inductance with frequency, using Method 2, answering many measurement needs at modest cost.

Following a few measurements and simulations to establish the voltage levels, I increased the preamplifier gain from 5 to 10, ensuring 1V drive to the rectifiers with a 100mV maximum input to the test jig.

To minimise the affects of circuit offsets when measuring the small differential $V(1)–V(2)$ voltage, I revised the rectified output attenuators, giving a maximum output of 2V to the modified PM128 display module, **Fig. 4**.

The differential amplifier

When measuring a low-impedance capacitor, the $V(1)–V(2)$ difference signal is some 35dB smaller than the common-mode voltages. Consequently the common-mode rejection of the difference amplifier must be significantly larger, and maintained to the highest measuring frequency.

I evaluated a number of instrument and differential amplifiers. Most worked well at lower frequencies but failed at

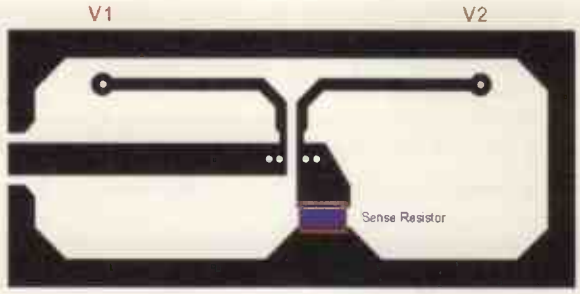


Fig. 6. Double-sided printed board, 54 by 25mm used for the test jig, shows mirror imaged measurement tracks to the V(1) and V(2) SMA connectors. The underside is completely covered by a ground plane except for a circular area at the test capacitor. Both top and bottom earths must be well connected together – especially at the earthy end of the sense resistor.

1MHz and above. I opted for the AD830, which claims differential gain and phase errors of 0.05% and 0.08° respectively. It also provides a 50dB common-mode rejection and better than 0.1dB gain flatness at 10MHz.

Phase-meter buffer

The phase-meter buffer circuit proved particularly difficult to design. My first prototype board worked well measuring impedance up to 1MHz, but performed poorly when measuring phase even as low as 100kHz.

I made many abortive attempts to trim the phase performances by adding small compensation capacitors. I then realised the circuit board had minor capacitance and track layout differences. $V(2)$ was not quite an exact mirror image of $V(1)$. Perhaps more important was the Maxim 457's use of an 8-pin DIL package.

This meant that $V(1)$'s non-inverting input, being next to the –5V supply pin and PCB tracks, exhibited slightly more capacitance to ground. Also, at high frequency, the output pin of $V(2)$ was similarly affected by being next to the +5V supply pin and its PCB tracks.

These differences were not significant when measuring

Determining capacitor impedance

This a summary of Method 2, as detailed in an earlier article[2]. Combined impedance of the test capacitor and sense resistor is represented by '$|Z|_{total}$'.

$$|Z|_{total} = R_{sense} \times \frac{VM(1)}{VM(2)}$$

where $VM(1)$ and $VM(2)$ are voltage magnitudes measured at $V(1)$ and $V(2)$ using a normal voltmeter.

$ESR = (\cos \textit{phase-angle} \times |Z|_{total}) - R_{sense}$.

where *phase-angle* is $VP(1)–VP(2)$.

$X_c = \sin \text{phase-angle} \times |Z|_{total}$.[6]

If *phase-angle* is negative, capacitance is,

$$\frac{1}{2 \times \pi \times f \times X_c}$$

If *phase-angle* is positive, inductance is,

$$\frac{X_c}{2 \times \pi \times f}$$

Capacitor impedance is given by,

$$|Z|_{capacitor} = \sqrt{ESR^2 + X_c}$$

Voltages $V(1)$ and $V(2)$ are complex voltages, having both magnitude and phase. But for this method you only need to measure voltage magnitudes.

gain by frequency, but contributed phase differences between the channels.

Revised layout

I carefully rebuilt the board using pads as for a 10-pin package, together with exact mirror imaging of tracks and with matched components for the preamp, phase buffer and rectifier stages. Adjusting one small trimmer capacitor, I achieved better than ±0.1° phase difference up to 1MHz between the two channels, measured at the phase meter buffer outputs.

I considered the preamplifier stage to be even more sensitive so I laid it out carefully using earthed guard tracks around the most sensitive pins and components, and carefully mirroring the channels.

This layout equalisation also improved the common-mode rejection at high frequency such that measuring a wire short circuit in place of a test capacitor in the jig, the short circuit now measured less than 4mΩ up to 10MHz.

Not being suited to an auto-routed layout, both boards were carefully laid out by hand. **Fig. 5**.

Relay switch

One final area that needed attention was the switching between two discrete measuring channels and the differential measuring circuit. It was important to equalise circuit loadings for both relay positions.

You will see two apparently unnecessary resistors R_{26} and R_{27} on the circuit diagram, **Fig. 4**. When switched as two discrete measuring channels, U_{1a} and U_{1b} both drive into 820Ω and 470Ω resistors in parallel. One end of each 470Ω is at virtual ground.

Switched to differential measurement and without R_{26}, U_{1a} would drive only into an 820Ω resistor and the high input impedance of the AD830.

Similarly without R_{27} on the output of the AD830, the two 470Ω resistors would see unequal impedances. These differences may seem unimportant, but without these arrangements I was not able to attain as good common-mode rejection at high frequency.

Display meter

I modified a PM128 display module to ratio mode by removing two resistors R_2 and R_3. The |Z| meter's $V(1)$ output connects to 'REF-HI' on the meter, the pad that originally linked both resistors. Output $V(2)$ connects to the normal 'IN-HI' terminals and the |Z| meter ground to the PM128 input ground.

To accommodate the maximum voltage from the |Z| meter, R_6, originally 47kΩ, was removed from the PM128, and replaced with 470kΩ.

Test jig requirements

To minimise line reflections to the signal generator, a 10dB attenuator should be inserted in the signal coaxial cable, immediately adjacent to the test jig.

If a 1Ω sense resistor is used, then the meter automatically reads impedance in ohms. Because I have standardised on a nominal 0.5Ω, the gain of the $V(2)$ output stage must be increased to compensate.

Using a sense resistor larger than 1Ω, the gain of the $V(1)$ output stage should be increased instead. If a sense resistor larger than 2Ω is used, you may need to alter both rectifier attenuators, to restrict output voltage into the PM128 meter.

The |Z| meter printed board provides spaces for additional gain-setting resistors in both channels. Not shown in the schematic, these connect between pin 2 and pin 6 of the AD712 amplifiers.

Setting up

Adjustment is simple. While measuring a known 1% resistor

Setting up

Solder a good short circuit to the jig in place of a test capacitor. A 3mm-wide piece of copper sheet will do for the short. The jig should be fitted with your chosen sense resistor.

Apply 100mV AC at 300-350Hz from a signal generator at the jig input and with the |Z| meter set to read $V(1)/V(2)$. Adjust R_3 then R_{53} to obtain 1V outputs at both $V(1)$ and $V(2)$. Set the meter to read,

$$\frac{V(1)-V(2)}{VM(2)}$$

and slightly adjust R_{53} to minimise the $V(1)$ output.

Remove the short circuit and replace with a known resistor, say 1Ω at 1%. Attach the PM128 panel meter.

Leaving R_3 and R_{53} as now set, add gain adjusting resistors to adjust the gain of the appropriate output stage until the PM128 display reads correctly for your known resistance.

When you are satisfied with the reading, apply higher test frequencies. The display reading should increase slightly according to the inductance of your known resistor.

Remove the known test resistor and replace with the short circuit.

Attach phase meter to the phase buffer outputs then apply 100kHz to 1MHz from the signal generator. Adjust trimmer C_{68} to obtain as close to 0° of phase as possible, over this frequency range.

Remove short circuit. |Z| meter is ready for use.

Impedance

When an alternating current passes through a perfect capacitor, the voltage waveform lags that of the current by 90°. An ideal capacitor has neither inductance nor resistance. The capacitor produces an impedance with a phase angle of –90°.

At any one frequency, a practical capacitor can be represented by a series combination of inductance, capacitance and resistance. These combine to produce an impedance with a much reduced phase angle. Depending on frequency, the measured phase angle can be either positive or negative.

An *LCR* meter converts this measured impedance and phase angle into two components only,

$$|Z|\angle\theta = R \pm jX$$

representing ESR (resistance) and reactance.

If the measured phase angle is negative, the meter calculates a capacitance value,. If positive it calculates an inductance value.

Because only a single frequency has been measured, the meter cannot segregate this reactance into its inductive and capacitive components.

$$|Z|\angle\theta = R \pm jX = \sqrt{R^2 + X^2}$$

Here, R is the capacitor's ESR at the measured frequency while X is the capacitor's reactance at that frequency.

With practical capacitors, this net reactance has both a capacitive and inductive component. In principle, provided impedance and phase angle are measured at a minimum of two frequencies, it is feasible to extract these components from the values of 'X' measured. There's more on this in the panel entitled, 'Three component modelling'.

mounted on the test jig, at 300-350Hz, adjust the gain of the relevant channel to display the correct value on the PM128 meter.

As for the main printed circuit board, the test jig was also carefully mirror imaged when laying out the tracks that connect the test capacitor to the |Z| meter. Ensure that track lengths and capacitances between both channels are similar, **Fig. 6**.

Sense resistors

The sense resistor used must be non-inductive. If it isn't, as frequency increases every capacitor measured will read less than its true value. Conventional 1Ω 1% resistors are readily available, but these usually have a spiral 'cut', used to trim to final value. Combined with the resistor's physical length, this results in sufficient self-inductance to increase the resistor's impedance measurably at 100kHz, degrading our measurement.

Fig. 7. Test jig specially developed for the |Z| meter differential voltage and phase measurements. It incorporates a non-inductive current sensing resistor and is usable to at least 10MHz. Three PRC201 1.5Ω 1218 chip resistors are used to make 0.4995Ω. The earthy end of the sense resistor is connected to the ground plane using vias and a brass U channel, well soldered along its length.

Fig. 8. Assembled |Z| meter printed board with two SMA connectors arranged to directly accept the test jig. By switching the central relay, the board can be made to operate in either measurement mode. The AD830 IC mounted near the relay, provides excellent common-mode rejection for differential measurements. To provide direct impedance measurement with 0.4995Ω sense 'R', gain increasing resistors for V(2) output are fitted next to U_5.

A spiralled reference resistor should not be used, except when measuring only very low frequencies. Some makers provide metal-film resistors with either a straight cut or no cut at all. These are considerably more expensive than conventional spiralled resistors.

If you are in doubt, carefully scrape away the resin coating from a resistor, to reveal whether it has been spiralled in

Capacitance and inductance values

These values are calculated from measured $R \pm jX$ for typical 220µF/10V capacitors. One random sample only of each style was measured.

Unit/parameter	10 kHz	30 kHz	100 kHz	300 kHz	1 MHz
220µF/10V Philips037	Aluminium				
Impedance (Ω)	0.824	0.785	0.738	0.706	0.671
ESR (Ω)	0.818	0.783	0.74	0.71	0.67
Capacitance	129µF	68.4µF	27.1µF	9.1µF	2.0µF
220µF/0V Rubycon YXF	Aluminium				
Impedance (Ω)	0.404	0.372	0.341	0.322	0.308
ESR (Ω)	0.394	0.372	0.34	0.32	0.32
Capacitance	153µF	94µF	44µF	20.7µF	9.2µF
220µF/10V Rubycon ZL	Aluminium				
Impedance (Ω)	0.122	0.091	0.082	0.076	0.073
ESR (Ω)	0.10	0.09	0.09	0.08	0.08
Capacitance	187.5µF	69.5µF	122.3µF	104.5µF	1.21nH
220µF/10V Rubycon ZA	Aluminium				
Impedance (Ω)	0.087	0.041	0.030	0.028	0.035
ESRΩ (Ω)	0.04	0.04	0.03	0.03	0.03
Capacitance	189.8µF	185.4µF	200.9µF	0.69nH	2.39 nH
220µF/10V Elna RSH	Aluminium				
Impedance (Ω)	0.313	0.290	0.270	0.254	0.242
ESR (Ω)	0.31	0.29	0.27	0.26	0.25
Capacitance	171.7µF	119.7µF	55.0µF	27.6µF	19.33µF
220µF/10V AVX TPS	Tantalum				
Impedance (Ω)	0.143	0.089	0.06	0.045	0.040
ESR (Ω)	0.10	0.08	0.06	0.05	0.05
Capacitance	149.2µF	100.9µF	62.3µF	45.7µF	0.83nH
220µF/10V Sanyo Oscon	Oscon				
Impedance (Ω)	0.077	0.027	0.01	0.01	0.029
ESR (Ω)	0.02	0.02	0.01	0.01	0.01
Capacitance	201.3µF	192.0µF	230.9µF	2.13nH	4.23 nH

I made these measurements using my dedicated meter, switched to Method 3 for impedance then Method 2 for ESR and capacitance. Many other suitable capacitors are available, from other stockists. ESR obviously cannot exceed |Z|. Where this occurs in the table, it is caused by insufficient resolution in my phase-angle measurement.

While the rapid increase in capacitance of the 220µF/10V Rubycon ZA and the 220µF/10V Sanyo Oscon looks odd, it is simply a reflection of the effect that self-inductance has on apparent capacitance, when approaching series resonance.

With electrolytic capacitors, series resistance usually dominates. So the notable impedance null, frequently found at resonance with low-loss capacitors, cannot be observed. See the panel entitled 'Three-component modelling' for more information.

manufacture. I have found some makers' low value straight cut 'Melf' resistors have extremely low inductance.

I recently bought some 1W low-value surface-mount resistors, listed by the distributor as non-inductive. I then wasted time getting very strange results.

On telephoning the manufacturer to complain, I was advised that this part should never have been listed as non-inductive because it was in fact a miniature, resin sealed wirewound!

Surface-mount 'thick-film' chip resistors ensure low inductance for two reasons. A straight 'L' cut is often used to trim such resistors, and their physical lengths can be very short.

Even better, certain types are available that are wider than they are long. Sometimes these are effectively three 1206 resistors in parallel. A typical 1206 chip has some 1 to 1.5nH inductance, so with three in parallel, this construction provides minimal self inductance[5].

Three-component modelling

As shown in Fig. 3, a capacitor is usually represented as a series combination of resistance as ESR, self inductance and capacitance.

When one frequency only is measured then the reactive component can only be interpreted as being either capacitive of inductive, according to the sign of the measured phase angle.

When measuring capacitors at several frequencies and especially if some of these measurements are taken at well above the capacitor's self resonance, its reactance can be segregated into capacitive and inductive parts. This results in a more realistic capacitance measurement at all frequencies.

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

where X_C is the capacitive reactance and X_L the inductive reactance at the measured frequency.

Solving for three unknowns requires a minimum of two measurements at differing frequencies. Ideally a swept measurement at several frequencies is used. Certain recent swept-frequency component analysers, like the HP4194 and HP4195, are provided with internal software routines. These routines automatically calculate the three-component model[7].

The software calculates parameters at frequencies where the measured impedance is a factor of √2 smaller and larger than the maximum and minimum values measured[8].

This method works well for many stable components but not with electrolytic capacitors. These have ESR and capacitance values that change with frequency. Self inductance for electrolytics, however, is relatively constant with frequency.

I prefer to estimate this inductive component by taking a series of impedance measurements at frequencies well above resonance. Taking the Oscon capacitor, which resonated at 190kHz as an example, I measured impedance at 1MHz intervals up to 10MHz. Above 2MHz its apparent measured inductance stabilised slightly below 5nH. Its impedance then increased linearly with frequency.

Correcting the measured results for 4.5nH self inductance reduces the Oscon's actual capacitance at 100kHz to 163.7µF. This represents excellent performance and is a much more realistic value than the device's apparent capacitance, which increased to 230.9µF, calculated from an uncorrected, single-frequency measurement.

Applying the same inductance correction to the 30 kHz measurement reduces its apparent 192µF to an excellent 186.5µF.

These figures clearly show the effect that a small amount of self inductance has on conventionally measured capacitance values – especially at frequencies less than a decade below resonance.

Having a reasonable estimate for self-inductance, using the above three-component equation can produce a far better estimate of the true capacitance value at all frequencies, both above and below resonance.

Using a magnifying glass, the 'L' trimming cut of a thick-film resistor can usually be seen through the coloured lacquer that coats the resistor element.

I bought a number of 1.5Ω resistors, Philips type PRC201. These are 1218 size, comprising three 1206 resistors in parallel. Measuring voltage drop while passing a 100mA DC, I was able to select a number of identical sets of three, effectively nine 1206 resistors in parallel. Each set makes a non-inductive, near 0.5Ω value, for my test jigs. These can be seen stacked and soldered on a test jig in **Fig. 7**.

Measurement range

Using the components and gains as shown, the |Z| meter measures from a few milliohms up to 1.999Ω, from low audio frequencies up to 10MHz. This range is ideal when measuring electrolytic capacitors. By scaling the sense resistor and output stage gain settings, other measurement ranges can be provided.

This system has been designed for use with a conventional 50Ω laboratory signal generator capable of supplying at least 7V into a 50Ω load. In use, the signal generator output is adjusted so as to provide 100mV input to V_1, **Fig. 8**.

Performance

Intended to measure from a few milliohms to 1.999Ω, this meter makes accurate impedance measurements on low-impedance capacitors ridiculously easy and is usable at frequencies up to 10MHz.

The preamplifier and buffer stages provide phase reference and phase measurement outputs at levels that simplify the measurement of phase angles. Phase measurement is needed to ascertain resonance frequencies or to calculate capacitance and inductance. It is not needed to measure impedance magnitude.

Of course if you have no need to measure phase, then the buffer components can be omitted, with significant saving on component costs.

The prototype meter was assembled using Augat low-profile turned pin sockets. These exhibit some 1pF of capacitance pin to pin. If other makes are used, or indeed no sockets at all, it would be prudent to adjust the values of C_3 and C_{53} to compensate for change in capacitance.

I plan to make PCBs together with full drawing sets and assembly notes available. For details and prices, send an SAE to me at Nimrod, New Road, Acle, Norfolk NR13 3BD. ■

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4. Bateman, C, 'Measure AC millivolts to 5MHz' *Electronics World*, Apr. 2000
5. Bateman, C, 'Understanding Capacitors' *Electronics World* Apr. 1998.
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7. Parametric Analysis for Electronic Components and Circuit Evaluation, AN339, Agilent Technologies, (H P) USA.
8. Practical Design & Evaluation of High Frequency Circuits, AN317, Agilent Technologies, (H P) USA.

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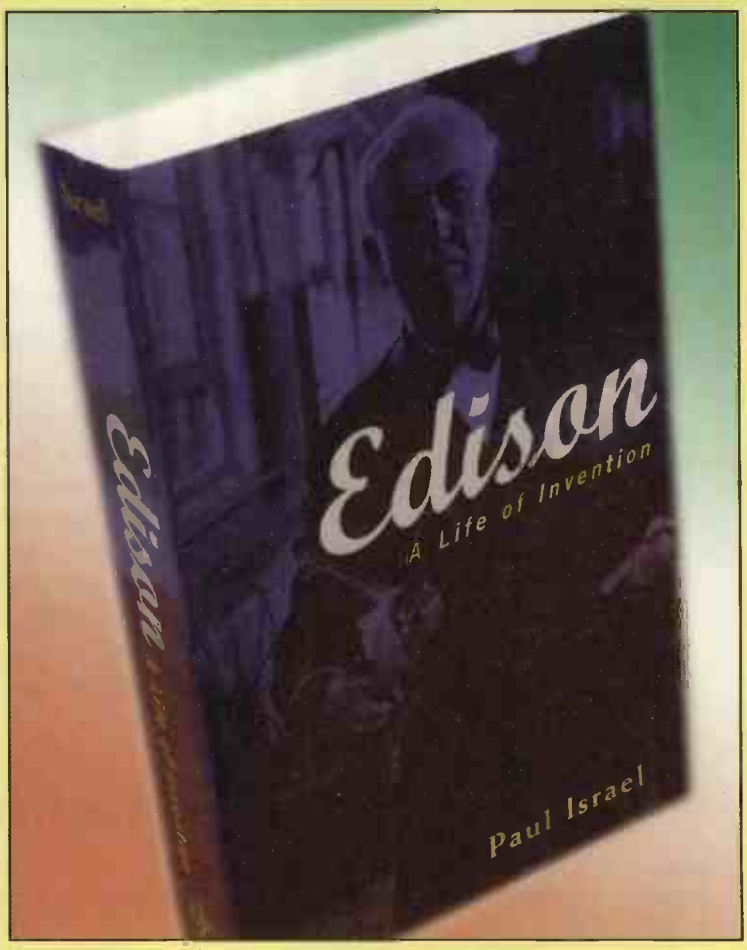

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CIRCUIT IDEAS

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

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

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

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

Pulse-width modulated power supply

£50 Winner!

This voltage-controlled PWM converter needs but a single NE556. Note that the input voltage on the voltage-control pin has to be between 0.45 and 0.9×V_{cc}. To allow for this, a resistor is inserted between the voltage control and the input pin.

The voltage divider thus created converts a 0-10V input voltage into the correct voltage range for the control pin. The conventional capacitor circuit is used, so the control circuit must create a control voltage between $^1/_3$ and $^2/_3$ V_{cc}.

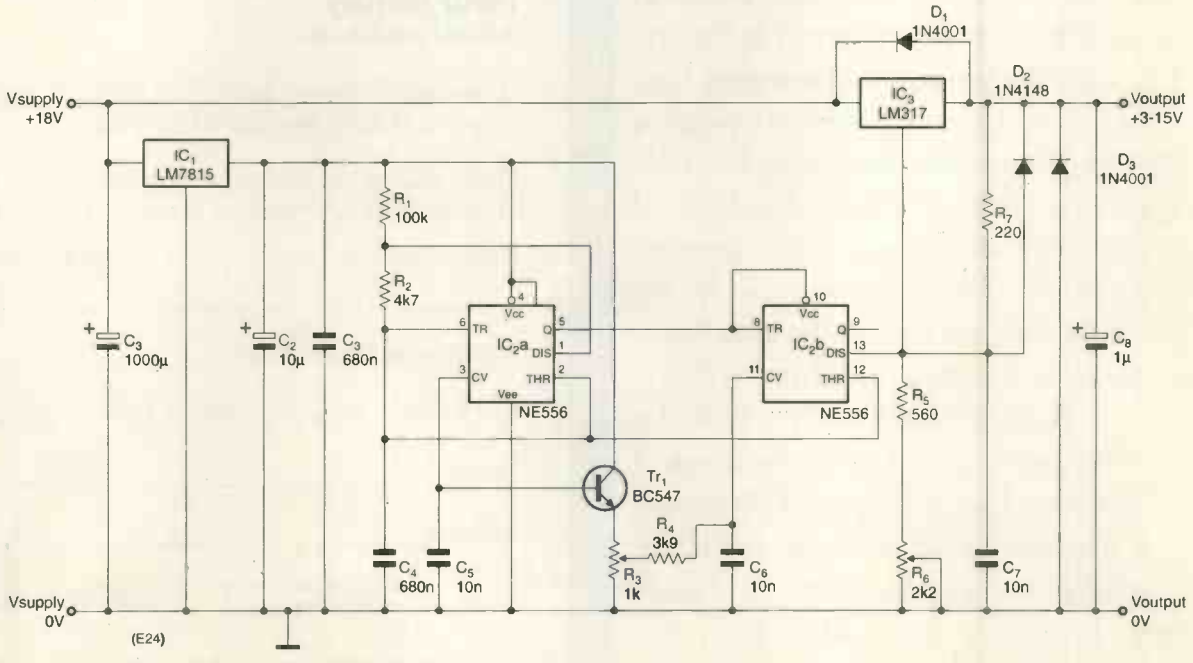
Timer IC_{2A} forms an astable multivibrator generating an exponential reference signal. This signal is compared in IC_{2B} with the control voltage created with P_1 (R_3), forming a PWM signal.

Output is taken from the discharge pin of IC_{2B}. This pin drives the voltage-regulator IC_3 to create an output voltage that is regulated and has short circuit protection. Amplitude of the output voltage is set with P_2 (R_6) between 3V and 15V. Pin 7 short circuits the resistance P_2 and R_5 when the control voltage is lower than the reference signal.

The reference voltage for P_1 is taken from the voltage-control pin of IC_{2A}, but any voltage between 0-10V DC can control the PWM setting. Range of the PWM is limited to the rising edge of the reference signal. To achieve this the output of IC_{2A} is connected to the reset and trigger pin disabling the output of IC_{2B} while capacitor C_4 is discharging.

In the CMOS version of the NE556, the internal reference voltage chain uses higher value resistors, and it is necessary to scale R_3 accordingly.

Bernard Van den Abeele
Evergem
Belgium
E24



PWM controlled 3-15V supply.

Linear sawtooth oscillator with buffered output

All that is needed to turn a 555 timer into a linear sawtooth oscillator with buffered output is a resistor, R_1, and p-n-p transistor Tr_1 (see diagram).

The transistor acts as an emitter follower with emitter bias resistor R_3. However, resistor R_1 across the base emitter junction of Tr_1 provides a near constant current of about $0.6/R_1$ (ignoring base current) to charge C_2. Resistor R_2 discharges C_2, as in a normal 555 astable circuit.

Charging current through R_1 changes according to the variation of the base emitter voltage. This change is largely due to the variation in the transistor's collector-emitter current, which is proportional to R_3 and any output load.

Provided that the collector-emitter current is much larger than the charging current, the variation in emitter current is approximately 2:1 so the base emitter voltage change is in the region of 20mV, i.e. about 3%.

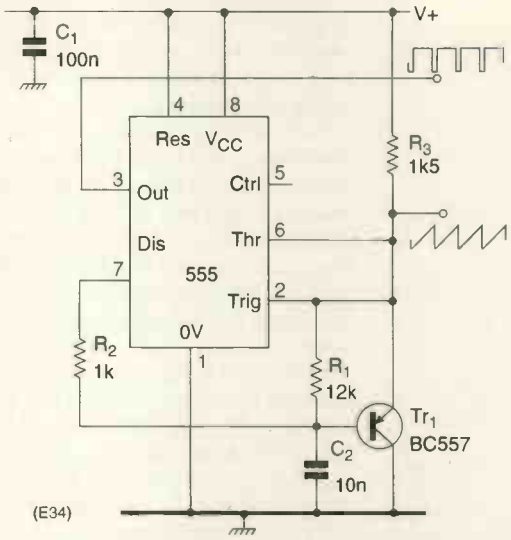
If the oscillator frequency is not more than a few kHz, the sawtooth amplitude is from $^1/_3$ to $^2/_3$ V_{cc} just like a normal 555 astable. At higher frequencies, the additional delay imposed by the transistor causes an increase in output amplitude. This effect can be reduced by increasing the transistor's bias current.

The discharge part of the sawtooth is still exponential in shape. If this part is very short the amplitude of the output will increase (see above).

Remember that the oscillation frequency is now no longer voltage independent since a constant current is involved and that there is a –0.3% per °C temperature dependence.

Operating frequency is approximately $1.8/(R_1 \times C_2 \times V_{cc})$. With the values shown in the diagram the charging current is about 50µA and the oscillator frequency is about 3kHz with supply of 5V.

M Hughes
York



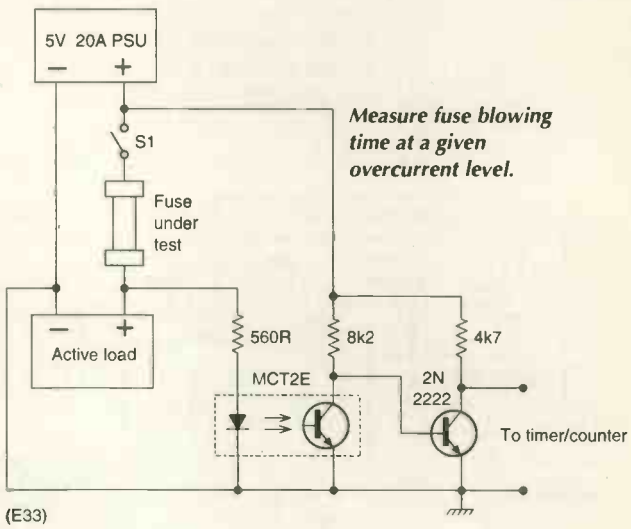
This 555 timer-based circuit provides a buffered sawtooth output.

Test fuse blowing time

It can be important when designing an equipment to choose the correct fuse rating. In the interests of reliability, it may be useful to measure the time it takes to blow a fuse at a given current, which exceeds the rating by a given amount.

This circuit provides the means for such a test. The required test current is set up on the active load. On closing the switch, the time to blow is recorded on the timer/counter. It is essential that the switch does not exhibit bounce; a knife switch is recommended.

J Kathe
Mumba
India
E33



Measure fuse blowing time at a given overcurrent level.

Crowbar protection for variable out-output supply

In a regulated power supply, the traditional crowbar circuit is designed for fixed voltage output only, as the trip point must be hard wired. The lack of over voltage protection on a variable-voltage output power supply is a hazard when connecting circuits to be tested or repaired to such power supply.

The following circuit was developed to provide crowbar protection on a regulated power supply with 0-30V output, using only two CMOS ICs and one voltage comparator IC. The circuit is powered at 15V via an LM317T three-terminal regulator connected to the main rectifier smoothing capacitor. The common 7815 is not applicable because of its low maximum input voltage.

The circuit consists of an oscillator, touch-operated switch, digital sample-and-hold circuit, trip voltage comparator and crowbar SCR. The touch point is made to the metal knob or the metal spindle of the variable resistor that the operator adjusts to change the regulated supply output voltage.

Table. Measured test results for the variable-voltage power supply over-voltage trip.

Power supply nominal o/p (V)	Trip voltage with germanium D2 (V)	Trip voltage with silicon D_2 (V)
5.0	5.9	7.2
10.0	11.0	12.3
15.0	15.9	17.3
20.0	21.0	22.3
25.0	25.8	27.2
30.0	30.8	32.3

If the variable resistor is mounted on a conductive panel, the resistor case and spindle must be insulated from the panel for the touch sensor to function properly.

Oscillator U_{1B} generates a 300Hz square wave that feeds the digital sample-and-hold circuit and the touch switch. Clock polarity is inverted via U_{1A} and applied to the U_{2B}'s clock input. The 4013 bistable multivibrator samples the data input 'D' during the rising edge of the clock.

If the metal knob is not being touched, because of the delay caused by U_{1A}, 'D' will always be low while clock rises. When the operator touches the metal knob to change the power supply output voltage, the human body introduces capacitance. This results in delay to the signal applied to 'D' and a change of state occurs at the 4013 outputs.

While the operator adjusts the supply output voltage, the touch switch operates. Bistable device U_{2B} changes state and sets U_{2A}. Transistor Q_3 conducts and trip circuit Q_1 is disabled.

Counter U_3 is now forced into reset with all its outputs, $Q_{1\text{-}8}$, low and no voltage appears at the R-$2R$ ladder network. Green LED D_3 turns off to signal protection and the circuit is now off.

After the correct output voltage is set, the operator releases the metal knob and the touch switch resets. Diode D_4 becomes inactive and the reset line of the 4040 returns to low.

The 4040 starts counting up. When the voltage output at the R-$2R$ network matches that of the regulated supply output voltage at point A, U_{1C} operates to enable U_{2A} to change state. This results in clock pulses from 4040 being bypassed, and turns on the LED.

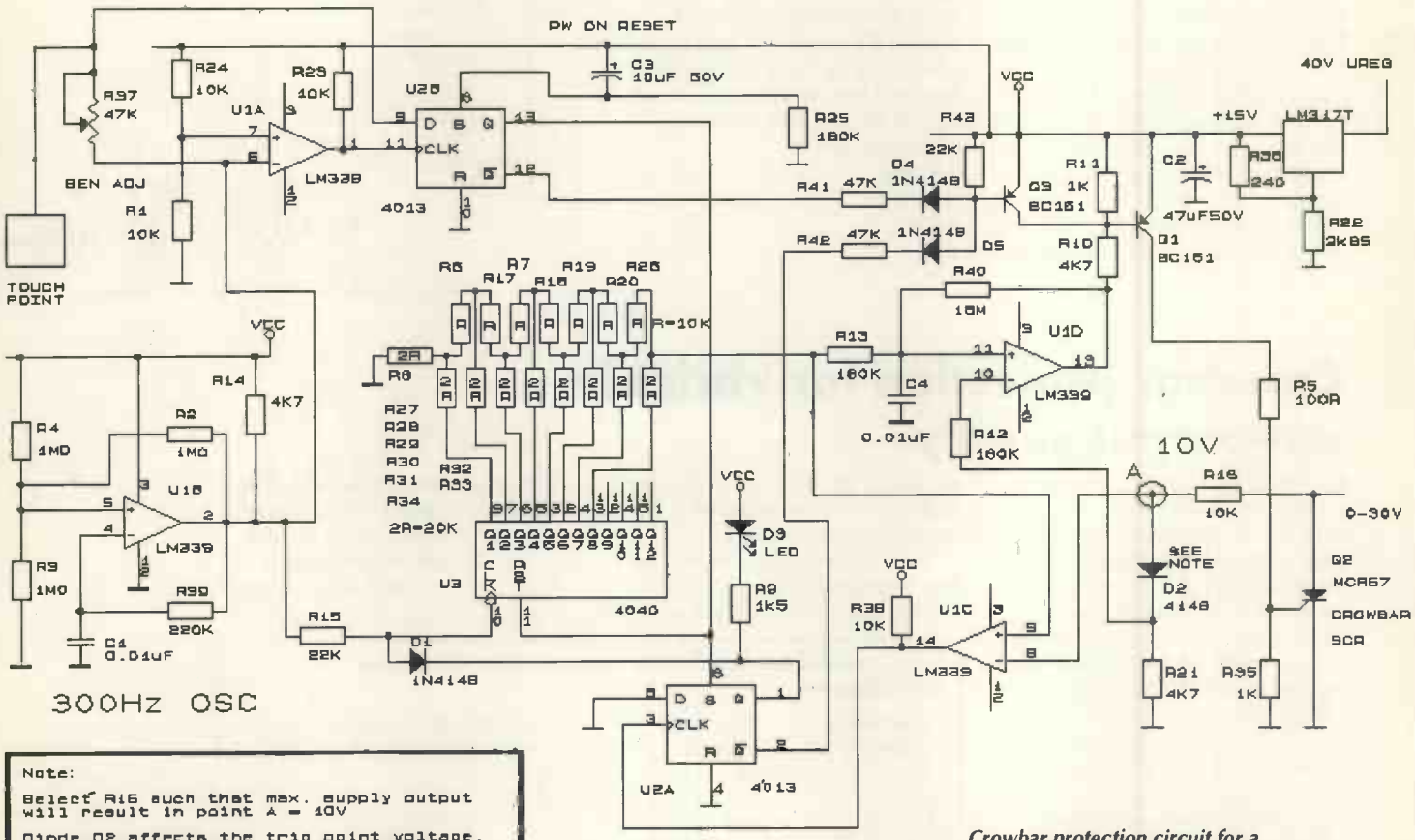
Simultaneously, the trip circuit is armed. However, as the sampling point of the trip comparator U_{1D} pin 10 is taken after diode D_2; a voltage offset exists between the R-$2R$ network voltage and the pin 10 voltage. It is this offset voltage that gives the ceiling of the over voltage protection.

As the input is sampled via R_{16} and R_{21} voltage divider, this translates to 300% of the diode voltage drop added to the external voltage that would result in a trip. Choice of D_2 between germanium or silicon types gives a different trip ceiling voltage.

Select R_{18} such that point A is +10V at maximum output. The test results of the actual tripping voltage on my prototype were measured as in the Table shown earlier.

Lim Chung
Haywards Heath
Sussex
E30

Winner of the National Instruments digital multimeter worth over £500



Note:
Select R16 such that max. supply output will result in point A = 10V
Diode D2 affects the trip point voltage. Use silicon diode for a 0.2V offset and germanium diode for a 0.9V offset.

Crowbar protection circuit for a laboratory bench supply.

National Instruments sponsors Circuit Ideas

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

Once every two months throughout 2000, National Instruments is awarding an NI4050 digital multimeter worth over £500 each for the best circuit idea published over each two-month period. At the end of the 12 months, National is awarding a LabVIEW package worth over £700 to the best circuit idea of the year.*

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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 $5^1/_2$ 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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Simple relay tester

£50 Winner

This simple circuit allows dynamic relay tests to be performed quickly and inexpensively on a large number of devices, as long as one normally open and normally closed pair of contacts is available. It is useful when checking a suspected sluggish relay, or seeking to qualify a new vendor.

Upon energising the circuit, the relay's normally closed contact applies a high logic level to a CMOS flip-flop's 'set' input. In turn, its Q output is now set high, driving the power MOSFET on, energising the relay's coil.

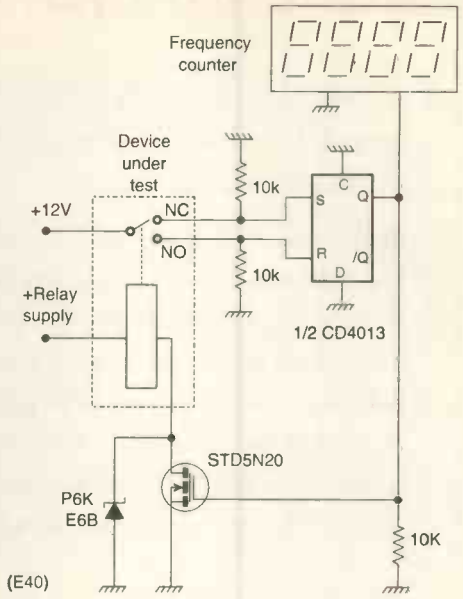
As the relay's wiper starts to move, the high level is removed from the 'set' input. This is of no consequence as the flip-flop won't change state until the wiper reaches the other, normally-open contact. At this time the flip-flop is reset and ceases to drive the FET so the relay is de-energised.

The wiper then goes back to the normally closed position, and the cycle repeats itself, creating an oscillation whose frequency is highly dependent on the relay's mechanical properties. This frequency may be read by a frequency counter or oscilloscope, and a rough estimate of the relay's response time may be obtained.

Any suspect relay may be sent to the lab for further testing in a controlled environment. The beauty of this circuit is that untrained personnel may quickly sort out potentially defective devices. In fact,with a little experience, the frequency counter is no longer required, as a trained ear may easily identify a sticky relay by listening to the cadence from the relay's buzzing sound.

Note that separate supplies are used to power the logic and the relay. Not only will this accommodate different coil voltages, but most importantly, it will prevent electrical noise – of which there will be plenty – coupling into the logic circuit, creating an erratic behaviour. For the same reason, I would recommend that you use CMOS flip-flops operating at least at 12V for improved noise immunity.

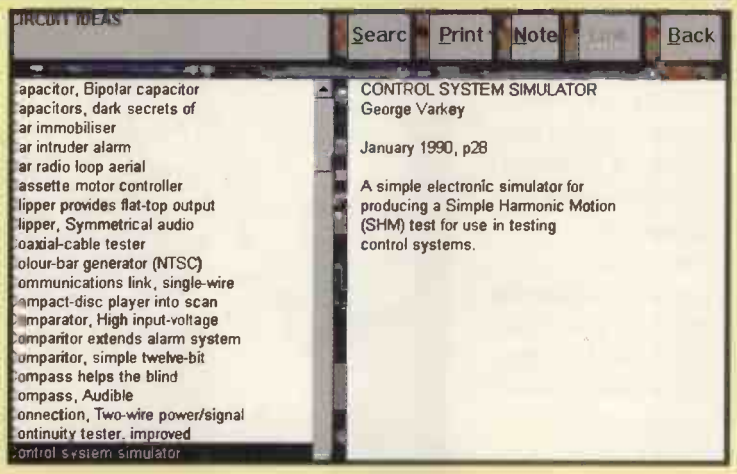
It is worth mentioning that the ubiquitous freewheeling diode normally placed across the relay coil should not be used. This dramatically increases the hold time after the relay is de-energised, and may even make it long enough to mask mechanical defects.



Rather, a 200V MOSFET, assisted by a relatively large-voltage transient suppressor, allows for dissipation of the magnetic energy with a minimal delay. ■

Fernando Garcia
Brownsville
Texas
USA
E40

Ten year index: new update



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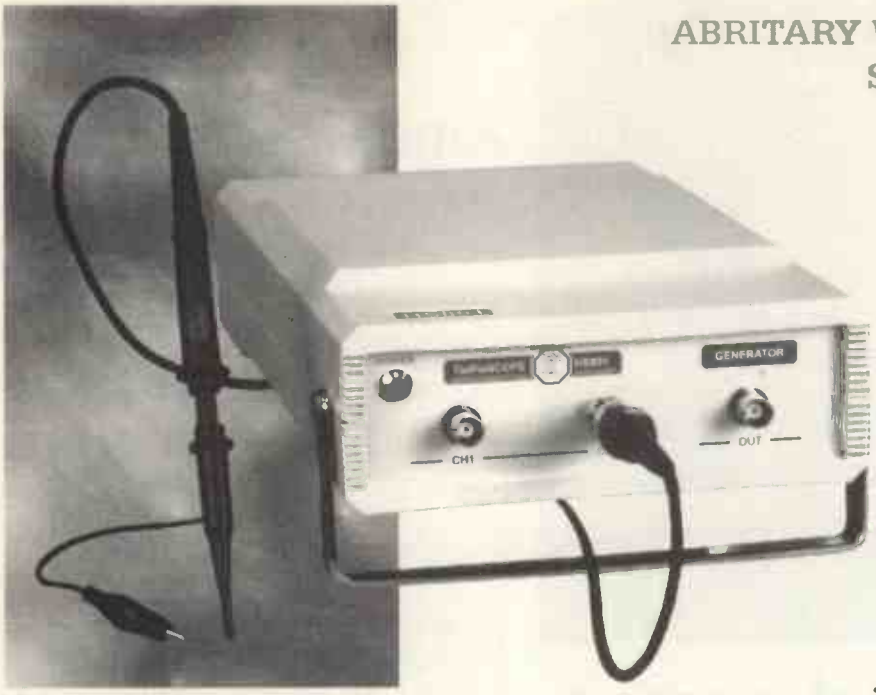

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- Measured signals and instrument settings can be saved on disk.This enables the creation of a library of measured signals. Text balloons can be added to a signal, for special comments. The (colour) print outs can be supplied with three common text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

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

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

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

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Track

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

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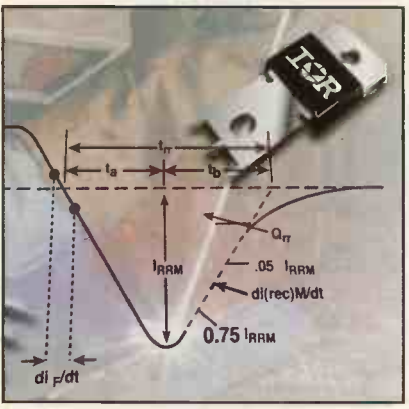
NEW PRODUCTS

Please quote *Electronics World* when seeking further information

Arc-welding diodes for 500A

International Rectifier has introduced five recovery diodes for inverter welder output stages. Typical voltage and current requirements for arc welders are 20 to 50V at 50 to 500A DC. Four of the five



diodes – the 80EBU02, 150EBU02, 80EBU04 and 150EBU04 – are available in the firm's PowIRtab package, and are rated at 80A. The fifth device (60EPU02) is rated at 60A and comes in a TO-247 package.

International Rectifier
Tel: 0208 645 8001

LED pilot lights

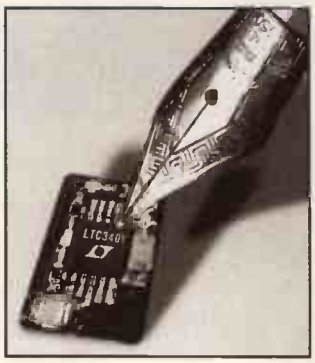
From EAO are series 04 pilot lights with direct feed 110 and 230V BA9 AC LEDs for 22.5 and 30.5mm mounting. The units can also be supplied with 24V LEDs, filament lamps and

transformers. The LEDs consume 3 to 5mA depending on voltage. The pilot lights are fitted using a bayonet fixing and two screws for assembly. The compact version is installed using a threaded fixing nut. The lights can be fitted into a 22.5mm hole with round or square fronts or flush mounted into a 30.5mm hole. Connection to the units is by screw terminal or 6.3mm fast-on terminals. Applications include control systems and industrial machinery. The lens cap or diffuser marking plate can be engraved for identification of the pilot light function. Illumination is in red, yellow, green, blue or white.

EAO
Tel: 01444 245021

Synchronous boost converter

Linear Technology has announced the LTC3401 synchronous step-up DC-to-DC converter in a ten-lead MSOP that operates from an input voltage down to 0.85V (single alkaline cell). It can deliver over



500mA output with up to 97 per cent efficiency and draws 38µA of supply current in burst mode or less than 1µA in shutdown. Switching frequencies from 300kHz to 3MHz may be programmed with an external timing resistor. Applications include pagers, cordless phones, GPS receivers and hand-held instruments. It incorporates a 0.16Ω n-channel MOSFET switch and 0.18Ω synchronous rectifier. Its current mode architecture has adaptive slope compensation to facilitate loop compensation. In user controlled burst mode, the internal power MOSFETs operate intermittently based on load demand. Its oscillator can be synchronised to an external clock to improve suppression of switching harmonics in sensitive audio and IF bands. Burst mode operation is inhibited during synchronisation or when the mode-sync pin is pulled low, which prevents low frequency ripple interference with audio circuitry.

Linear Technology
Tel: 01276 677676

20V MOSFETs

Fairchild has introduced 20V p-channel MOSFETs with 8 and 12V V_{GS} ratings in surface mount packages. The devices have maximum $R_{DS(on)}$ ratings at 1.8V gate voltage for switching low voltages, so there is no need for additional boost circuitry. The SuperSOT-8 FDR840P 20V V_{DS}, 12V V_{GS} MOSFET has a maximum $R_{DS(on)}$ of 11mΩ at 4.5V V_{GS}. The FDW254P single 20V MOSFET in a TSSOP-8 at 2.5V V_{GS} has an $R_{DS(on)}$ of 15mΩ.

Fairchild Semiconductor
Tel: 01793 856819

Risc in-circuit emulator

Available from Noral Micrologics, the Lauterbach TRACE32 FIRE-166 Risc in-circuit emulator provides real-time, non-intrusive debugging of embedded applications based on Infineon's 3.3 and 5V C166 16-bit microcontrollers. It can be used with all derivatives of the C166 family and features include 40MHz no-wait-state operation, hardware dequeuing, dual-ported emulation memory for on-chip flash and XRAM, and full tracing of information on bondout buses. A context tracking system reduces the time needed to identify and rectify software problems by letting developers recreate and analyse the embedded system's debug environment after the code has executed in real time.

Noral Micrologics
Tel: 01254 295800

Please quote *Electronics World* when seeking further information

Voltage reference

National Semiconductor has announced a sub-bandgap series voltage reference. With a voltage reference of 1.024V, the LM4140, is for battery-powered instruments and test equipment. Accuracy is 0.1 per cent and temperature coefficient 3ppm/°C. It comes in an SO-8 package. Noise is 2.2μV p-p from 0.1 to 10Hz and load regulation is 1ppm/mA. Stability is 60ppm over 1000 hours and thermal hysteresis 20ppm. Quiescent current is typically 230μA, with shut-off current less than 1μA. An enable pin helps prolong battery life. Because this is a series reference, it can also be used in low drop-out applications. The drop-out voltage is 20mV at 8mA and the device can source up to 8mA. There are five reference-voltage options – 1.024, 1.25, 2.048, 2.5 and 4.096V. All are specified from 0 to +70°C.

National Semiconductor
Tel: 0870 240 2171

Rabbit 2000 processor kit

2001 is stocking a module-based development kit for the Rabbit 2000 microprocessor. The kit provides a ready-built processor core containing processor, memory, I/O and control functions which simply plugs in to the prototyping

board or finished system. By using the kit designers can check out concepts and produce working module-based prototypes. In addition modules are available in production volumes. Three module variants are available – with 18MHz or 25MHz clock and 128k or 512k SRAM as required for the application. As well as the Core Module (RCM2020), the Module development kit includes: manual with schematics and documentation on CD-ROM, getting started page, prototyping motherboard board, programming cable, and a complete Dynamic C SE software development system.

2001
Tel: 01438 742001

3mm pitch connector

A 3mm pitch addition to its Mate-n-Lok range of connectors has been announced by AMP. With the smallest pitch in the MNL family, the series is suitable for wire-to-wire and wire-to-board applications, and can handle up to 5.0A per circuit. Features include a dual-beam contact design and PCB pegs for board polarisation. Offering 2-24 position vertical and right angle headers for surface mount and through-hole applications, the products are footprint compatible and intermateable with competitive equivalents. Featuring high-temperature header housings for IR processing, the connector series also features a pre-staged, two-piece housing assembly which will not lock unless the contacts are correctly positioned, and which then holds them securely in place with retention to withstand at least 20 pounds of force per contact. No special tools are required to remove a contact, so disconnection is quick and simple. They are available in 2-

24 position sizes with three row configurations available for 9,12 and 15 positions.

AMP
Tel: 0208 420 8072

Embedded Web browser

Amino Communications, the network appliance firm, has licensed the ANT Fresco Web browser from embedded communications appliance software developer, ANT, for use across all its information appliances and networked devices.

The ANT Fresco browser, ported onto Amino's IntAct modules adds another element to Amino's development 'toolkit', providing a browser suitable for a wide range of network devices. The browser has a small memory footprint, allowing it to be integrated into many designs for devices such as set-top boxes, seat back entertainment systems, in-vehicle terminals, kiosks and so on. The browser is processor and system software independent and so can be ported to a number of different platforms.

Amino Comms
Tel: 01954 784500

Programmable dual VCXO clock

American Microsystems has announced a programmable clock generator chip for video set-top boxes. The FS6219 is a 3.3V programmable, three PLL clock chip with two integrated VCXOs. The device meet worldwide set-top box signal frequency standards. The two VCXO circuits make it possible to receive simultaneous, but independent, signals for picture-in-picture. The VCXO and PLL circuits are user-programmable over an I²C bus. Each VCXO uses an external voltage to tune the crystal frequency for the phase locking of each reference to an independent source, such as the incoming MPEG clock. Features include programmable reference, feedback and post-dividers for generation of clock frequencies with no synthesis errors. Either VCXO can be used as the PLL reference. Output frequencies can be selected from one or more of the PLLs.

American Microsystems
Tel: 00 49 351 31 530 23

Advert. SCIENCE WORLD DISCOVERY BOOK. Further to Catt's work on the nature of the 377 ohm dielectric in electromagnetic theory, new research has established the cause of gravity. Volume = (volume of fundamental particles) + (volume of dielectric or fabric of space). This continuity equation implies that the motion of distant matter in the universe radially away from us, discovered by Hubble from the red-shifted spectra of clusters of galaxies, produces an equal and opposite effect on the dielectric of space, which is the cause of gravity.

SCIENTIFIC PROOF: THE MECHANISM OF GRAVITY

When a submarine moves underwater, or when you take clothes out of a suitcase (fixed volume), or when a person walks down a corridor, an equal volume of water or air moves in the opposite direction to fill the volume being vacated by the matter. In the universe, the motion of clusters of galaxies has a similar effect on the fabric of space. A unit volume is full of a mixture of electrons and nuclear matter and the fabric of space, like a suitcase containing some clothes and some air. The predicted effect of this from the Hubble motion of matter in the universe (the big bang) is therefore an equal and opposite motion of the fabric of space. Hence, the fabric of space moves towards us to fill in the volume being vacated by the outward moving matter as clusters of galaxies rush away from us. Furthermore, the fabric of space (or dielectric of space) permeates through atoms and planets.

The speed, v, of the clusters of galaxies increases with distance, r, according to $v = rH$. This variation in speed constitutes an acceleration since the distance is directly proportional to the time which has elapsed since the light was emitted (the space-time effect). Hence, by definition acceleration, $a_H = dv/dt = d(rH)/dt = vH = rH^2$.

Hence, the effect of the inward acceleration of the fabric of space, caused by the outward motion of matter in the universe (like the motion of air into the suitcase when you take clothes out of it) produces the acceleration due to gravity which keeps us on the earth's surface. The shielding effect of the Earth's mass on the all-round fabric of space creates the geometric effect of gravitation. Hence, $a = MG/r^2$, where we predict $G = 3H^2/(4\pi\rho)$. Thus, this theory actually predicts Newton's gravitational acceleration formula as well as the value of the constant, which can be verified experimentally by astronomy to prove or disprove the theory. Furthermore, this theory of gravity, the only ever proposed which accurately predicts the value of the constant G, also explains the recent experimental discovery that the furthermost galaxies in the universe are not "slowing down" due to gravitational retardation: gravity is not a mysterious "law" of nature but a mechanistic *effect* of the expansion itself.

MATHEMATICAL DERIVATION OF GRAVITY BY CONSIDERING IN DETAIL THE ASYMMETRIC DIELECTRIC PRESSURE FROM THE EXPANSION OF THE ENTIRE UNIVERSE

The shielding factor of matter (mainly nuclei in space) against the all-round accelerative pressure of the dielectric of space can be considered as follows. We consider a sphere around the observer with a radius equal to the distance of the mass from the observer. If a hypothetically massive shield *completely* stopped the dielectric pressure on one side of the observer, the maximum possible gravitational acceleration would be induced in the observer, forcing him towards the shield, like the bubbles pushed to the sides of a glass of beer or lemonade by the pressure of the liquid and the absence of pressure from the glass.

Dielectric pressure $P = F/A$. F = force $= m_z a_H$, where a_H = Hubble acceleration (because the dielectric motion in balances the Hubble mass motion outwards in the universe out to radius R) $= dv/dt = d(RH)/dt = vH = RH^2$, where R = effective radius of universe (radius at which the expanding mass produces the greatest contribution to the dielectric pressure where we are), and H = Hubble constant. A = area upon which dielectric force acts to produce the gravitational acceleration we wish to calculate = area of sphere with radius r, so $A = 4\pi r^2$. Now, m_z = effective mass of the dielectric of free space (the 377 ohm dielectric or continuous fabric of space) acting on the area A, hence m_z = mass of matter in hollow volume of the universe with an inner radius of r and an outer radius of R, so $m_z = (4/3)\pi (R^3 - r^3)\rho$, where r^3 is insignificant compared to R^3, so the formula reduces to simply $m_z = (4/3)\pi R^3\rho$, where (p = average density of universe (currently being carefully assessed by astronomers by working out the mass and distribution of galaxies in space, plus the gravitational effects of invisible matter, such as nuclear particles, on the shape of rotating visible galaxies), because the effective *dielectric* equivalent mass moving inwards is equal to the actual mass moving outwards.

Hence, $P = F/A = m_z a_H/A = (1/3)R^4H^2\rho / r^2$. This dielectric pressure acts equally on all sides of any sphere of radius r, thus preventing any net force or acceleration, but explaining inertia. Waves of dielectric flow, like Aristotle's arrow self-sustaining momentum mechanism in his book *Physics* (he was wrong only in mistaking the arrow for matter, when the actual matter is the sub-nuclear particles and electrons in the arrow; but we excuse Aristotle since he never had an X-ray or measured radiation penetration of atomic-composed matter), around moving fundamental particles from front to rear. The continuous, non-particulate, 377 ohm electromagnetic dielectric of space exhibits no drag because drag is only caused by particles colliding and thereby dissipating energy. [The continuous nature of the dielectric is illustrated by careful inspection of photos of starlight deflected by the sun's gravity during eclipses: the deflected starlight (photons) still produce a sharp image, indicating that the source of gravity (the dielectric) is continuous, rather than a diffuse image due to particle-particle scattering (which would occur between photons and "gravitons" if the cause of gravity was "graviton" particles.]

The shielding of this pressure P by a mass M at distance r creates an asymmetry (a net force upon the observer, due to a reduction in pressure coming specifically from the direction of M), hence the observer is accelerated towards M, thus producing gravity.

Since the equivalent mass of inward accelerating dielectric is equal to the mass travelling outward in the universe, a shield with the mass of the universe would be required to exactly and completely stop the inward dielectric (as elastic recoil). A shielding mass M, will therefore stop dielectric pressure in proportion to the ratio of M to the mass of the universe, $(4/3)\pi R^3$(. The effective shielding cross-sectional area is therefore $A_{shielding}$ = [surface area of a sphere of radius r].[$M/\{(4/3)\pi R^3\rho\}$] $= 3M r^2 / (R^3\rho)$. The net pressure towards mass M is the unshielded pressure multiplied by $A_{shielding}$, divided by the area of the sphere ($4\pi r^2$) of radius r which mass M is shielding: hence, net pressure, $\mathbf{P} = P(A_{shielding})/(4\pi r^2) = RH^2 M / (4\pi r^2)$.

To get the net acceleration of the dielectric towards mass M at distance r from it, we remember that pressure $P = F/A = ma/A$, so acceleration $a = \mathbf{P}A/m$, where **P** is the net pressure of the dielectric towards the mass M at distance r, m is the effective mass of the continuous dielectric accelerating inward, taking spherical divergence into account, m = [mass of universe].[surface area of sphere with radius r] / [surface area of sphere with radius of universe R] $= (4/3)\pi Rr^2\rho$, and A is the surface area of a sphere with radius r, so $A = 4\pi r^2$. Hence, $a = (3/4) H^2 M/(\pi r^2\rho)$. **Hence, we have theoretically derived the Newtonian expression $a = GM/r^2$, where we find $G = (3/4) H^2 /(\pi \rho)$.** This is a testable prediction, as H and ρ (including the invisible matter like black holes and neutrinos) become better known via further astronomy and nuclear physics research.

THE "SCIENCE WORLD DISCOVERY BOOK" SHOULD BE PUBLISHED IN 2001, PRICED £25. TO REQUEST AN ADVANCE ORDER, OR TO ENQUIRE ABOUT "SCIENCE WORLD" RESEARCH JOURNAL (ISSN 1367-6172), PLEASE WRITE TO NIGEL B COOK, 28 GATE COURT, WEYBRIDGE, SURREY, KT13 8NW, OR EMAIL: nigelbryancook@hotmail.com

CIRCLE NO. 115 ON REPLY CARD

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Electrolytic caps

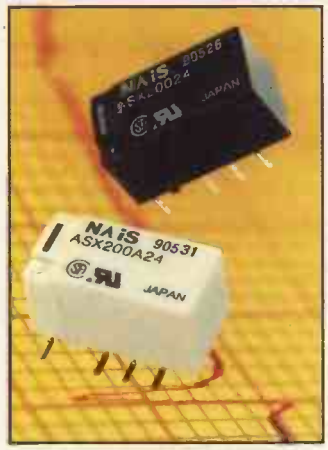
Components Bureau is stocking Jamicon's TXR series aluminium electrolytic capacitors. With radial leads on a 5 or 7.5mm pitch, the devices feature a high ripple current capability of up to 2000mA and low ESR of 0.25 to 4.47Ω, depending on value. Available in values from 3.3 to 330µF, and voltages of 160 to 450V DC, the low-cost capacitors are designed for operating at –40 to 105°C, or –25 to 105°C for the 450V devices.

Components Bureau
Tel: 01480 386565

Relays for small signals

Matsushita Electric Works has introduced its SX relay with dual changeover contacts that allow it to be used where only reed relays or other solid-state devices have previously been used. The combined effects of a guaranteed maximum initial contact resistance of 100mΩ, with the ability of being able to control loads as small as 10µA at 1mV DC, make the relay suitable for medical applications such as electrocardiogram equipment, where minute signals are passed by the relay without degradation. A minimal self-heating effect is generated through the SX's coil power pickup consumption of just 35mW, ensuring that thermoelectric voltages are reduced to a minimum 3mV. Measuring a subminiature 15 by 7.4 by 8.2mm, the SX is available in both through-hole and surface mount types, the contact arrangement is 2 form C, with latching configurations also available.

Matsushita
Tel: 01908 350700



Optical receiver

Anadigics has introduced an optical receiver that Stratos Lightwave has selected for use in its small form factor transceivers for storage area networks. The AMT8301T46F/L can operate at 1x and 2x Fibre Channel data rates.Sensitivity is better than -19dBm over voltage and temperature at 1.0625 and 2.125Gbit/s data rates at a supply voltage of 3.3V with less than 45mW power dissipation.

Anadigics
Tel: 001 908 668 5000

Microwave multi-chip module

Alpha Industries has introduced a proprietary multi-chip module packaging technology which it claims can reduce the cost of manufacturing high-speed and high-frequency datacomms equipment by replacing labour intensive wire-bonding with a surface-mounted package specifically designed for high frequency and high speed ICs. Called Alpha-2, it is claimed to be smaller and easier to assemble

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Wide XGA reduces optical components

Densitron's LightCaster WXGA display is a single panel projection system that is intended to reduce the number of optical components required. The system provides two display modes, which allows user selection between XGA (1024 by 768 pixels) and HDTV (720 P) with widescreen format (16:9 aspect ratio). Applications include television, data graphic, video-ready front projectors for business settings, desktop and presentation monitors; and videoconferencing.

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CALIBRATORS

Fluke 5220A Transconductance Amplifier	2350

COMPONENT ANALYSERS

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HP 4145B Component Analyser	11950
HP 4191A-002 1GHz Impedance Analyser	6500
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DATACOMMS

Microtest Penta Scanner Cat 5 Cable Tester	1650
Fluke DSP100 Cat 5 Cable Tester	1650
Fluke DSP4000 Cat 5/5e/6 Cable Tester	2650

ELECTRICAL NOISE

HP 346B 18GHz Noise Source	1150
HP 8970B Noise Figure Meter	8500

EMC

R&S EB 100 Miniport Receiver 20-1000MHz	3750
R&S EPZ 100 Mini Panoramic Display	1250
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Marconi CPM46 Counter Power Meter	5000
Philips PM6654C/526 1.5GHz/2ns GPIB Counter Timer	1650

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Various Marconi 69xx RF Power Sensors (from)	350
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Wandel & Goltermann OLP-2 Optical Power Meter	900

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Anritsu MS610B 2GHz Spectrum Analyser	2650
Anritsu MS612A 50Hz to 5.5GHz Spectrum Analyser	6500
HP 3561A 100KHz Dynamic Signal Analyser	3250
HP 3562A 2-Channel 100KHz Dynamic Signal Analyser	4500
HP 35660A 102.5KHz Dynamic Signal Analyser	2950
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HP 70000 2.9GHz Spectrum Analyser c/w Tracking Gen	13500
HP 8560A 2.9GHz Spectrum Analyser	10500
HP 8561A 1KHz-6.5GHz Spectrum Analyser	10850
HP 8562B 22GHz Spectrum Analyser	17500
HP 8591A / 021 1.8GHz Spectrum Analyser	4650
HP 8592A /021 Spectrum Analyser	9500
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HP 8903B 20Hz To 100KHz Audio Analyser	2750
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than traditional 'chip-and-wire' component packaging. Compatible with standard tape-and-reel manufacturing, the modules are suitable for use in high-speed data communications equipment, including LMDS wireless systems.
Alpha Industries
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Design for GPS infrastructure

Trimble has introduced a design for global positioning system (GPS) infrastructure, the Virtual Reference Station (VRS), which it claims supports centimetre-level positioning accuracy without the need to set up a local reference station. Developed in the US by Spectra Precision's subsidiary TerraSat, the VRS consists of software and a network of Trimble GPS receivers communicating with a control centre. The VRS software uses the data to calculate GPS error corrections that are applicable over a wide area. These error corrections are transmitted via wireless communication devices (radios or cellular phones) to users in the field within the network area. Additionally, users can retrieve stored GPS correction data from the control centre via the Internet for postprocessing. The system is claimed to reduce systematic errors in the reference station data to improve reliability and operating range. This allows a user to increase the distance at which the rover receiver is located from the physical reference stations while improving on-the-fly initialisation times.
Trimble Navigation
Tel: 01256 760150

400/200MHz amplifiers

Unique Memec has introduced two fixed gain amplifiers, configurable for gains of +1, –1 of +2 which feature identical bandwidths in both gain-of-1 and gain-of-2 configurations. With a bandwidth of 400MHz, the Elantec EL5196C offers a slew rate of 3000V/µs, while operating from just 9mA of supply current. The Elantec EL5197C offers a bandwidth of 200MHz with a slew rate of 2200V/µs and a supply current of 4mA. The EL5196C and E15197C are available in the industry standard 8-pin SOIC and SOT23-6 packages.
Unique Memec
Tel: 01296 397396

IEEE1532 standard PLDs

Lattice Semiconductor has announced that its latest generations of in-system programmable CPLDs will be compliant with the newly approved IEEE 1532 standard for programmable devices. Lattice plans to ship fully compliant ISP devices by the end of the year. Programming is accomplished through the IEEE 1149.1 boundary scan test access port (TAP).
Lattice Semiconductors
Tel: 01276 803223

Configurable supplies to 1kW

XP has introduced a 1000W unit in its RB series of configurable single and multi output modules which now ranges from 400 to 1000W. All units are approved, tested and guaranteed. The PFC units offer 1 to 10 outputs, have universal (85-264V) input and are CE marked. Output ripple and noise is 0.1 per cent RMS, 1 per cent peak-to-peak typical at 20MHz.

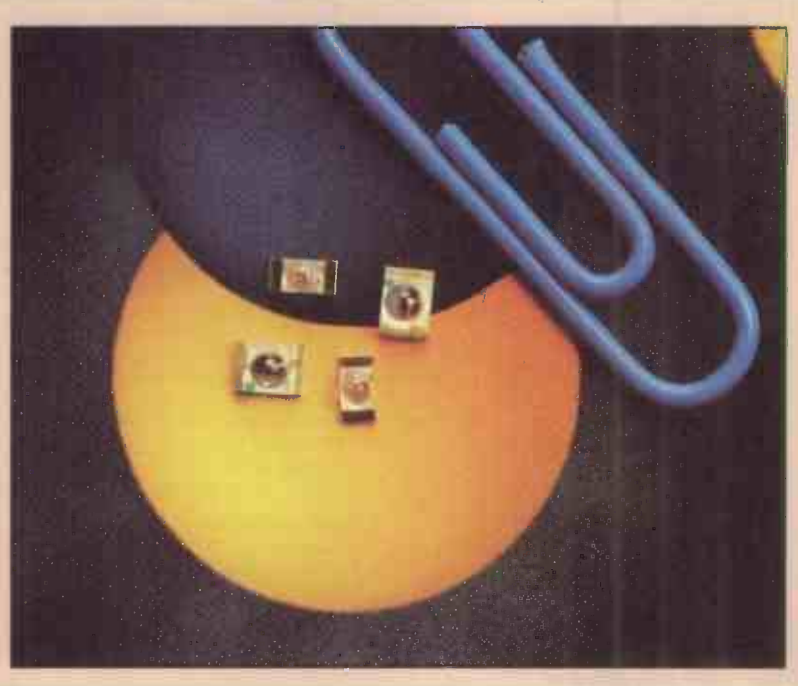
Features include overload protection, short circuit and overvoltage protection, remote sense and current sharing. Low earth leakage versions are available. Prototypes are available in three days. Every module is burned-in and tested prior to despatch, and covered by a comprehensive warranty and full technical support. Key markets include instrumentation and broadcast.
XP
Tel: 001189 845515

IEEE1149.4 test access IC

National Semiconductor has announced the development of a general-purpose SCAN IEEE 1 149.4-compliant IC. The Analog Test Access prototype uses embedded test circuitry from LogicVision to provide analogue access to board-level circuit nodes.
Digital board designs can make use of the IEEE1149.1 boundary scan standard for high fault coverage automated test generation but to access analogue test points an external In-Circuit-Tester (ICT) must be used. The IEEE1149.4 mixed-signal test standard is designed to reduce the need for ICT. The analogue test points can then be accessed according to the IEEE 1149.4 standard which defines

SM LEDs with lenses

Lumex has introduced two styles of surface mount light-emitting diodes (LEDs) that feature a built-in domed lens on top of a flat-bottomed carrier. According to the supplier, compared with conventional SM LEDs which emit light over an angle of 130° to 180°, these lensed devices focus the light to a narrower 50° angle. These LEDs are intended for applications such as displays or backlighting, as well as for status or circuit fault indicators. Two versions are available, the LXL1307 series is packaged on a chip carrier with a footprint that is 3.2mm x 1.6mm. The high point of the lens measures 1.8mm above the mounting plane. Colour choices include super intensity red, super orange, super red, super yellow, and ultra super blue. AlInGaP, GaAlAs, and InGaN/SiC chips are used, and provide on-axis light intensities as bright as 1500mcd. The LXL1209 series has a 3.2 x 2.4mm footprint and a 2.6mm high point above the board.
Lumex
Tel: 00 1 800 278 5666



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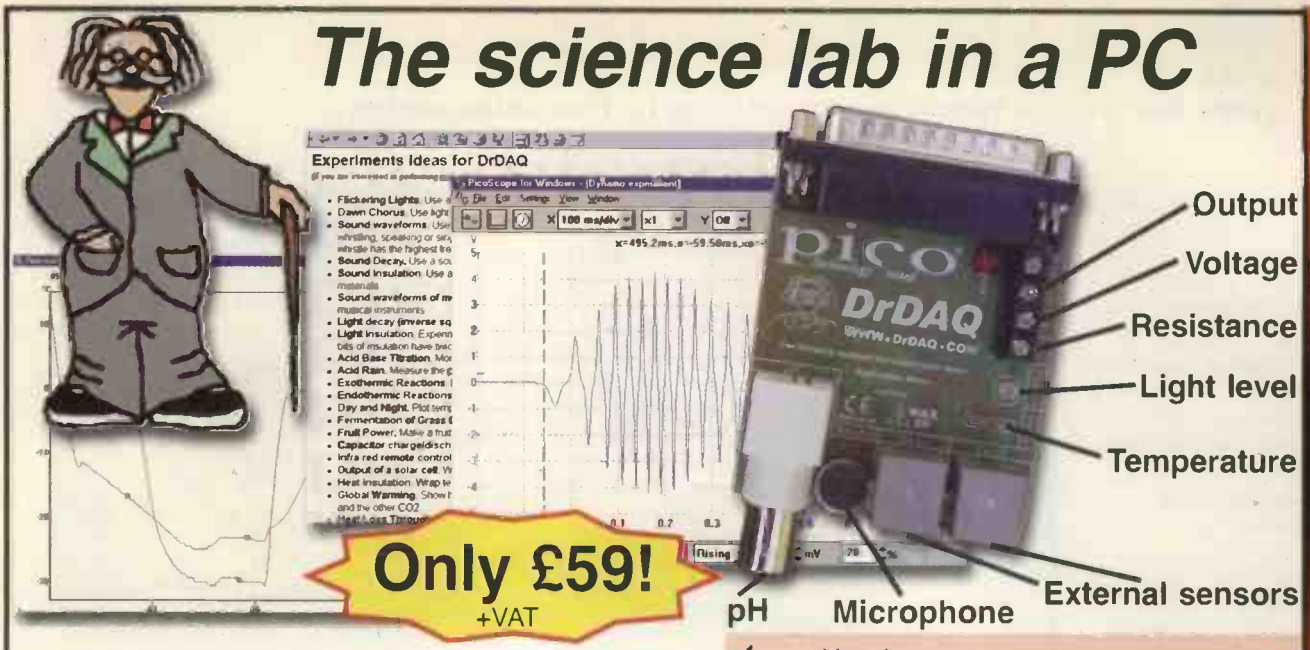

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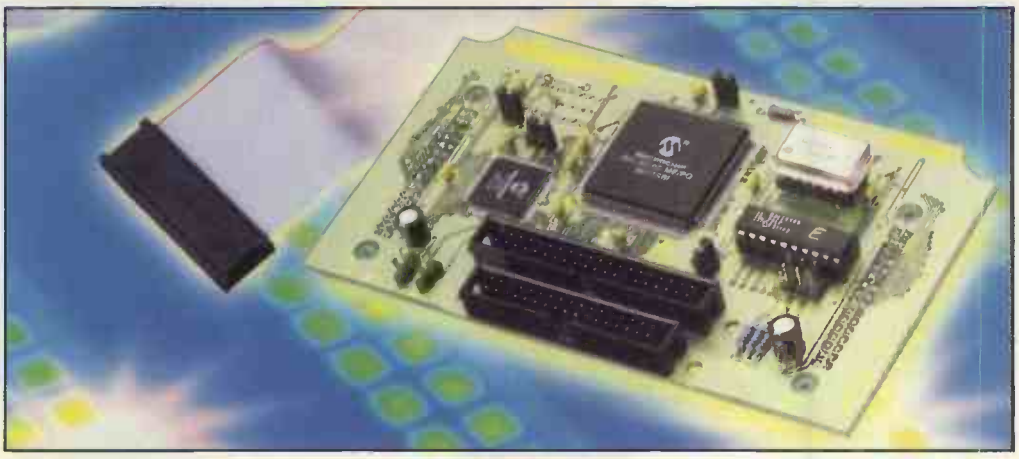
National Semiconductor
Tel: 0870 240 2171



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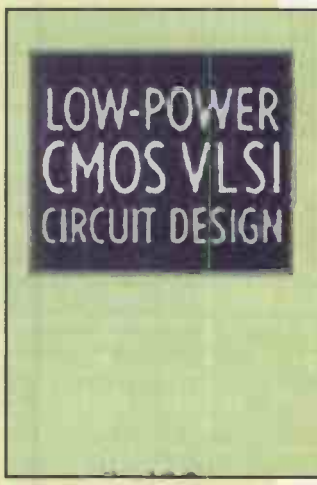

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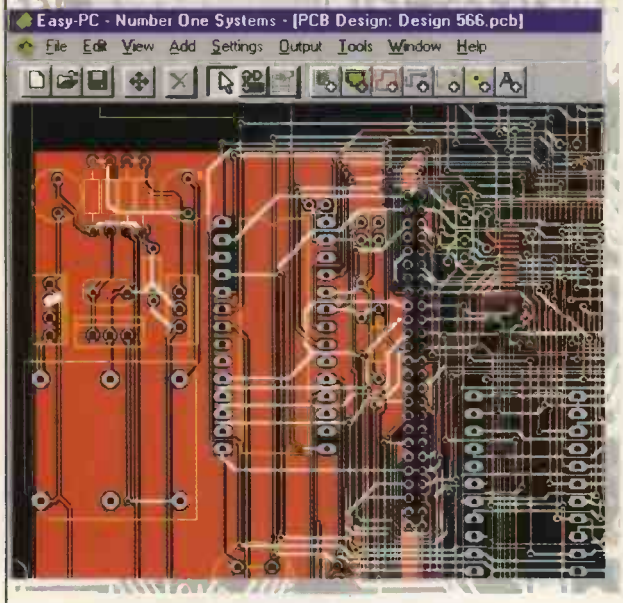

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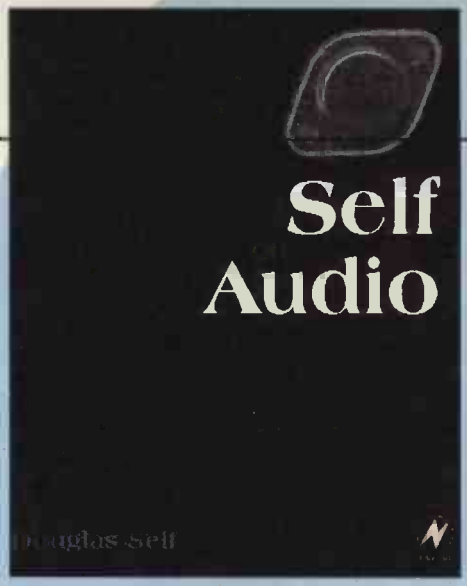

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Joe Carr explains how to get the most out of a radio receiver front-end design.

DESIGNING FRONT-ENDS

A receiver's front-end – that is the portion of the receiver prior to the IF amplifier – is the key to its dynamic performance. Areas such as dynamic range, intermodulation distortion, –1dB compression point and third-order intercept point are a function of how well the front end of the receiver performs.

Front-end architectures

Several different architectures are used in receiver front-end circuits. **Figure 1** shows the simplest form. It consists of a mixer stage and local oscillator preceded by a band-pass filter. Input to the band-pass filter comes from the antenna. The band-pass filter can be narrow or broad, depending on design.

There are two main issues regarding this type of architecture. First, there is cost. It costs less than the other architectures in some implementations. Secondly, the theory advanced by some authorities is, "why amplify noise prior to mixing?" The goal is to not use up the mixer's head room with processing unneeded energy. This theory has some merit, as was evident in the Squires-Sanders SS-1 receiver in the 1960s.

The main attributes of the bandpass filter are good forward performance – within the pass-band that is – and good reverse isolation. The second of these attributes is needed to prevent the local oscillator signal from reaching the antenna where it can be radiated. It has three important duties:

- It must limit the bandwidth of the input signal to minimise intermodulation distortion.
- It must attenuate spurious responses, mainly the image frequency and the $^1/_2$-IF frequency problems.
- Suppress local oscillator energy to prevent it from reaching the antenna.

Table 1. Affects of mixer attributes on receiver performance.

Third-order intercept point	Intermodulation distortion
Second-order intercept point	1/2-IF response
Noise balance	Sensitivity and AM noise rejection
LO to RF isolation	LO energy radiated by antenna
RF to IF isolation	Direct IF pick-up
Conversion loss	Sensitivity

A second version of the front-end architecture is shown in **Fig. 2**. This version uses an RF amplifier. The gain of the RF amplifier is low – certainly less than 20dB. Gains of more than 20dB may cause stability to be compromised, and the intercept point may not be achieved.

The purpose of the RF amplifier is to isolate the mixer as well as give the signal a small boost prior to mixing. This boost overcomes the losses in the mixer and the bandpass filter. The principal benefit of the RF amplifier is that it improves the isolation of the mixer/local-oscillator circuit from the antenna circuit.

A third version is shown in **Fig. 3**. Like the other two architectures, this one has a mixer and local-oscillator circuit – or a converter containing both mixer and local oscillator. The difference between this architecture and the previous one is the addition of a second band-pass filter.

This second band-pass filter may be the same frequency as the first, but that is not the only arrangement. It is often tuned to the image frequency. This frequency is the RF frequency plus or minus twice the IF, and is located on the other side of the local oscillator from the RF signal, **Fig. 4**. That way, the image frequency gets the same treatment in the mixer as the RF, so comes through the system as a valid signal.

Having a filter tuned to notch the image frequency, while passing the desired frequency, can limit this problem. Of course, the image filter must track the band-pass filter at the input if the receiver is multiple frequency.

The second band-pass filter may also attenuate the receiver's other spurious response and direct IF pick-up. In addition, it attenuates noise originating in the RF amplifier, preventing it from reaching the mixer.

Finally, it suppresses second-harmonic energy arising in the RF amplifier, thereby improving the receiver's second-order intercept point.

This filter should have no return responses at high frequencies. The reason is that the mixer has poor response for odd harmonics of the receive frequency, so they may ride through the system.

The RF band-pass filter's attributes will be determined by a combination of the first IF frequency and the injection side of the local oscillator signal. If low-side injection is selected, some of the spurious signal products may be on the low side of the RF signal.

On high-side injection just the opposite occurs: all of the spurious signals will be on the high side of the RF signal.

The trade-off between insertion loss and selectivity in the filter should usually be made in favour of insertion loss in band-pass filter number 1, but it can be sacrificed in band-pass filter number 2.

Mixer/local-oscillator performance

The performance of the first mixer is key to the performance of the receiver. It is a nonlinear device. Furthermore, it usually sees the highest level radio frequency signals in the system – the local oscillator, largely. So it needs to have a very high intercept point.

Single device active mixers are cheap, but they have the poorest performance of all the mixers. Generally speaking, the best performance comes from passive, double-balanced mixers. These generally have the highest intercept points, and better noise balance relative to most mixer designs. **Table 1** shows the mixer performance parameters and the things they affect.

Sometimes there is a third band-pass filter at the interface between the mixer and the local oscillator. This LO filter is used to attenuate wideband noise and its harmonics around the local-oscillator frequency, which could degrade the mixers second-order intercept point.

There is a trade-off in the type of mixer circuit used in a receiver. On the one hand, passive mixers have better intermodulation distortion performance than active mixers. However, they do not provide any conversion gain, and are in fact lossy devices.

Active mixers require less in the way of local-oscillator power, but their noise performance is worse than passive mixers. Furthermore, at high temperatures, the high third-order intercept point performance of the active mixer degrades.

A diplexer network is often placed between the mixer's IF output and the IF amplifier. The diplexer network absorbs some frequencies, while passing others. The diplexer network must be non-reflective up to several times the LO frequency. If not, those frequencies would be reflected back to the mixer, degrading its performance.

The single-sideband phase-noise performance of the local oscillator is important to the receiver's adjacent-channel selectivity. Wideband noise often afflicts the receiver sensitivity. Further, the LO signal must be as pure as possible to prevent spurious responses in the receiver.

It is not prudent to ignore the LO signal. It is a large signal that causes switching in the mixer, which generates its own harmonics. As a result, the local oscillator signal should be as pure as possible.

The local oscillator must be able to operate normally despite changes in temperature and power supply voltage. Its output should also remain stable if the receiver is subjected to mechanical vibration or impact.

Noise performance

All radio reception is a matter of manipulating the signal-to-noise ratio, i.e. SNR, of the system. Because of this problem, the noise generated by the mixer, local oscillator, band-pass filters and RF amplifier should be minimised.

For a passive, lossy device, such as the filter or some mixer stages, the noise figure is given by,

$$F = 1 + \frac{(L-1)\mathrm{T}}{290} \quad (1)$$

Here, F is the noise factor of the device, L is the loss of the device ($1/G$) and T is the temperature of the device in kelvin (K). Some double-balanced mixers can have slightly higher noise figures.

The Friis equation for noise governs the system,

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \ldots \frac{F_N - 1}{G_1 G_2 \ldots G_{N-1}} \quad (2)$$

Here, F is the equivalent noise factor, F_1, F_2, F_3 are the noise factors of stages 1, 2 and 3, F_N is the gain of the nth stage and G_1, G_0 and G_3 are the gains of stages 1, 2, and 3. Gain G_{N-1} is that of the stage before the Nth stage.

The overall noise factor of the receiver is determined by the noise performance of the stages within the system.

Spurious responses

A spurious response is a response that is not intended. On a superheterodyne receiver, these spurs can be created in the mixer stage, although they have their origin elsewhere. Most receiver spurs are a result of the heterodyning of the receiver, according to,

$$F_{IF} = mF_{RF} \pm nF_{LO} \quad (3)$$

Here, F_{IF} is the intermediate frequency, F_{RF} is the radio frequency, F_{LO} is the local-oscillator frequency and m and n are either integers or unity.

By solving equation 3 for F_{RF}, you get two possible RF frequencies at which spurs can occur. These are,

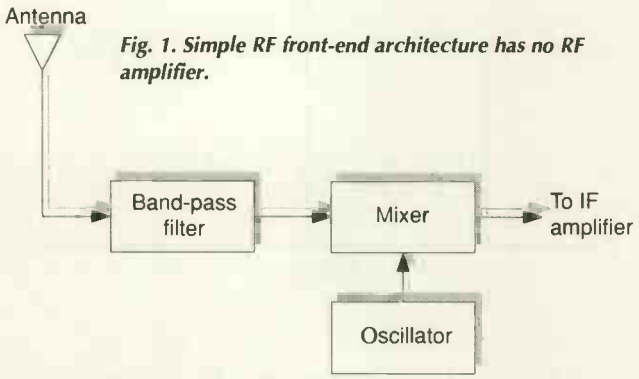


Fig. 1. Simple RF front-end architecture has no RF amplifier.

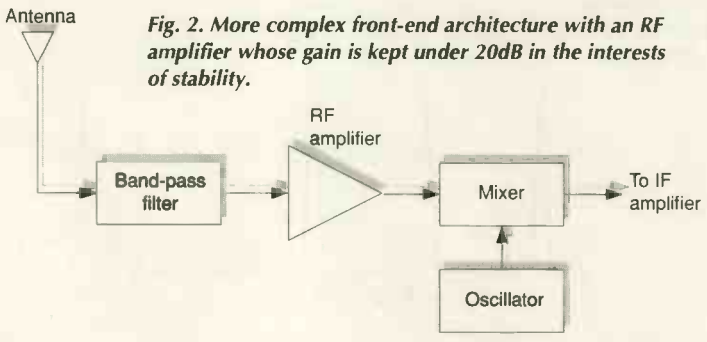


Fig. 2. More complex front-end architecture with an RF amplifier whose gain is kept under 20dB in the interests of stability.

$$F_{RF} = \frac{nF_{LO} - F_{IF}}{m} \quad (4)$$

and,

$$F_{RF} = \frac{nF_{LO} + F_{IF}}{m} \quad (5)$$

The most common spurs are,

- Image frequency (previously defined, see Fig. 4)
- $^1/_2$-IF (see Fig. 5)
- Direct IF pick-up
- n×LO frequency
- LO spurious frequencies
- Second mixer spurs (dual-conversion receivers only)

In full-duplex radio receivers, i.e. those that are used in conjunction with a transmitter at the same time, there are two additional responses that must be considered: full duplex image and half duplex image. These are defined as,

$$Duplex\ image = F_T - \Delta f \quad (6)$$

$$Half\text{-}duplex\ image = F_T + \frac{\Delta f}{2} \quad (7)$$

Here, F_T is the transmitter frequency and Δf is the difference between the transmitter and receiver frequencies.

Fig. 3. In the more complex form of front-end architecture, the second band-pass filter may have the same centre frequency as the first, or it may be tuned to the image frequency.

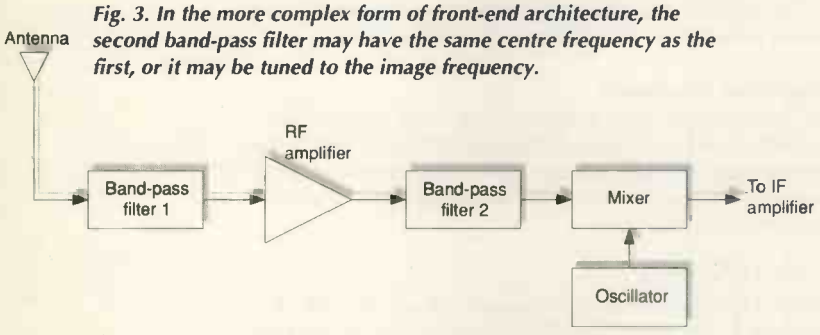


Fig. 4. RF, LO and image spectrum. The image frequency is the RF plus or minus twice the intermediate frequency.

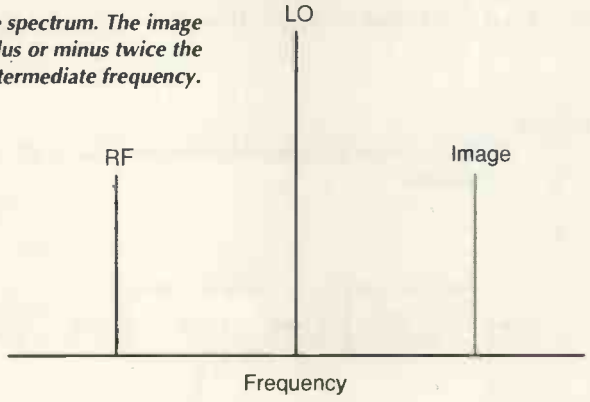
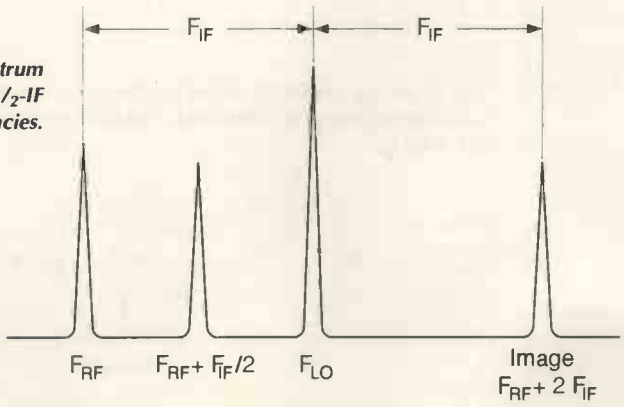


Fig. 5. IF spectrum showing $^1/_2$-IF frequencies.



Intercept points

An intercept point is a measure of circuit linearity. It allows you to calculate intermodulation distortion levels from the input signal levels. The intercept point represents an input amplitude, Fig. 6, at which the desired fundamental frequency is equal in amplitude to the undesired signal.

Second-order intercept point

The second-order intercept point, or SOIP, is due to the operation of the second-order products of a signal, and increases at a rate of 2dB for a 1dB increase in the fundamental level.

The $^1/_2$-IF response of the mixer can be predicted from the second-order intercept point. The $^1/_2$-IF point is due to the second harmonics of the RF signal and the LO signal, both of which are internally generated ($2F_{RF} \pm 2F_{LO}$). The $^1/_2$-IF rejection is given by,

$$\frac{IP_2 - S - C}{2} \quad (8)$$

Here, IP_2 is the second-order intercept point, S is the receiver sensitivity in dBm and C is the capture ratio or the cochannel rejection in decibels.

For example, suppose a receiver has a second-order intercept point of 45dBm, and a sensitivity of –120dBm. If the co-channel rejection is 6dB, the half-IF rejection is ,

$$\frac{45\text{dBm} + 120\text{dBm} - 6\text{dBm}}{2}$$

$$= \frac{159}{2} = 79.5\text{dB}$$

Third-order intercept point

The third-order intercept point, or TOIP, is the point at which the fundamental signal and its own third-order products are equal in amplitude. For each decibel increase in the fundamental signal, the TOIP increases 3dB.

The TOIP is predominantly responsible for the intermodulation distortion performance of the receiver. Intermodulation performance of the receiver can be defined as the difference, in decibels, between the receiver's sensitivity and the signal level that is sufficient to produce a specified level of interference. It can be calculated from,

$$IM = \frac{2IP_3 - 2S - C}{3} \quad (9)$$

where, IM is the intermodulation rejection ratio in decibels, IP_3 is the TOIP, S is the receiver sensitivity in dBm and C is the capture ratio or co-channel rejection in decibels.

Equation 9 covers the situation for one carrier. Unfortunately, real receivers see many carriers. The number of such products is $n(n-1)$, where n is the number of carriers present for both $2F_1-F_1$ and $2F_1+F_2$, and, for triple beats, $n(n-1)(n-2)/2$ for $F_1+F_2-F_3$ situations.

Intercept points of the *n*th order

Once you know the input levels of signals applied to the receiver, you can calculate the nth order intercept points using,

$$IP_N = \frac{nP_A - P_{IMN}}{n-1} \quad (10)$$

Here, IP_N is the nth order intercept point, n is the order of the intercept point, P_A is receiver input signal power level and P_{IMN} is the power level of the IMD signal

RF amplifier

The RF amplifier can have a deleterious effect on the performance of the mixer stage, hence the entire receiver. There's a number of methods that can be used to reduce the effect.

The first method is to use a high-power device operating well below its maximum range. There is a trade-off with noise performance, however, and that must be taken into consideration.

Second, reduce the signal level to the device. This can be done with attenuators, in some cases. Care must be taken though to balance the needs of sensitivity in this respect.

Third, reduce the stage gain. Again, noise and SNR considerations apply. Fourth, use negative feedback in the amplifier. And Fifth, increase the selectivity of the RF amplifier. A narrower bandpass produces less noise than wider bandwidths.

A sixth way is to use push-pull amplifiers because they tend to cancel even-order products – odd-order products are not affected – which tend to take up mixer head room.

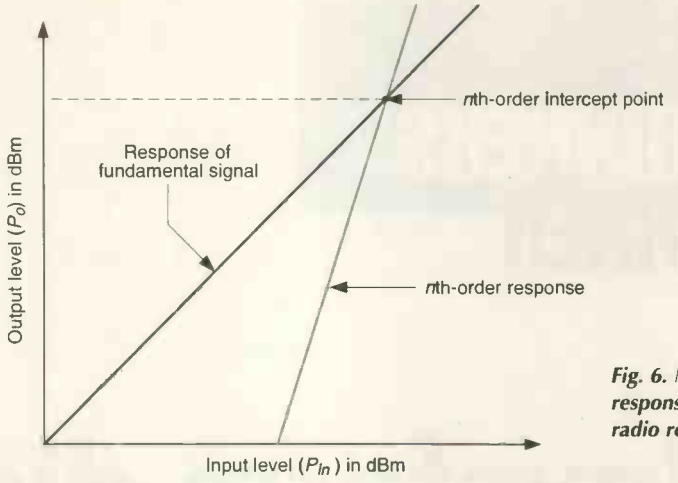


Fig. 6. Nth order response of a radio receiver.

In summary

A receiver's front-end dominates its dynamic performance far more than the IF or other sections of the receiver. The matters of sensitivity and selectivity are dominated by the IF performance characteristics, but the dynamic performance is influenced by the front-end of the receiver. ■

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BEGINNERS' CORNER

Waveform distributions

Probability distributions are an important topic in electronics, and in particular in communications, yet you rarely see an explanation of them. A spectrum analyser will tell you about the frequency distribution of a signal, but other techniques are needed to investigate the signal's voltage-probability distribution. Ian Hickman gives this neglected topic an airing.

I first came across voltage probability distribution diagrams many years ago, but they seem to be little discussed in the literature. Yet they aid our fundamental understanding of signals, in a way rather complementary to the voltage/frequency information provided by a spectrum analyser.

There are two sorts of probability distribution diagrams; those showing the cumulative probability of a voltage waveform, and those showing the probability density.

Voltage probability density distribution

Figure 1a) shows a squarewave of amplitude $\pm V_p$ about 0V ground. Imagine a window comparator set at a level more negative than V_p, indicated at A = = = A. If the window is moved up to the level indicated by the dotted lines, the probability that the voltage will lie within the window is zero – assuming that the squarewave is perfect, with infinitely fast rise and fall times.

As the window reaches and encloses $+V_p$, the probability becomes 0.5, assuming that the mark:space ratio is indeed 50:50. Beyond $+V_p$, clearly the probability of the voltage lying within the window is zero again. The probability is 0.5 while the level V_p remains within the window, regardless of the window's width. In other words, the area of the shaded rectangle at V_p represents a probability of 0.5.

Now consider an infinitely narrow window, moving up from A = = = A, and encountering the $-V_p$ edge of the squarewave. The long thin rectangle at $-V_p$, represented as a line with arrow head, has an area of 0.5, but its width is zero. So its length must be infinite. It is in fact an example of a 'delta function' – a function well known to mathematicians.

The usual zero-width infinite-amplitude delta function has an area of unity. But here there are two of them, one at $-V_p$ and one at $+V_p$, so each has an area of 0.5. The total area of unity indicates a probability of unity, or 100% if you prefer.

For 100% of the time, the voltage level is either $+V_p$ or $-V_p$. The probability of occurrence of any voltage between $+V_p$ and $-V_p$, or indeed outside that range, is zero.

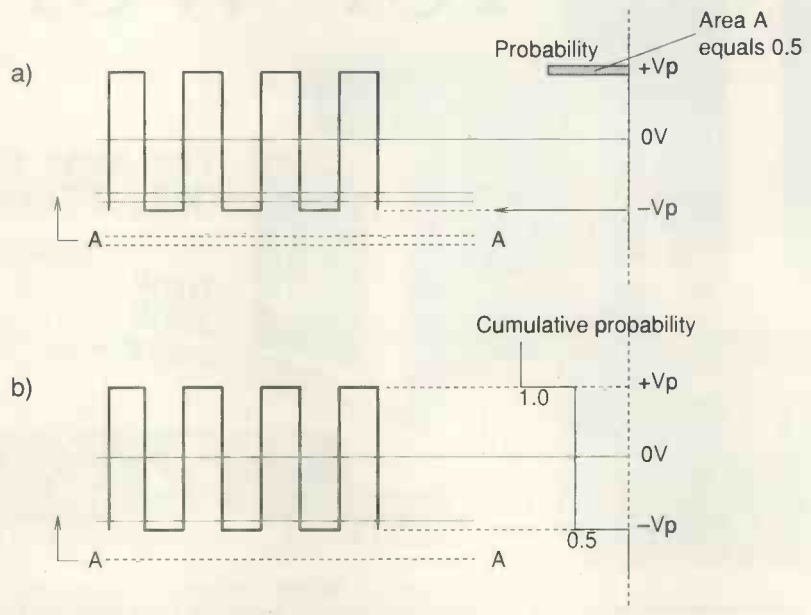

Fig. 1. Probability density of a squarewave a), and cumulative probability of the same waveform, b).

Determining probability density

While Fig. 1a) shows a probability density diagram, probability density is not the easiest thing to measure. You can use a window comparator as described, to obtain an approximation to it, but obviously a window of zero width is impracticable.

Much more manageable is the cumulative probability diagram. The cumulative probability diagram for a squarewave is given in Fig. 1b).

Imagine an ideal comparator whose output is zero when the signal input is above its reference input, and +1V when below. Then the cumulative probability will be as in Fig. 1b). Note that the cumulative probability is the integral of the probability density.

Other waveforms

Every waveform type has its own characteristic probability density diagram. Figure 3 shows that for a sinewave. The shape shown gives roughly the right idea, although it is not guaranteed accurately to scale.

As the infinitely-narrow window encounters the negative peak of the sinewave, suddenly there is a finite probability, since the sinewave has a 'stationary point' at $-V_p$. Above this point, the curved sides of the negative peak cross the window at an increasingly acute angle, so the probability decreases. It is least at zero volts, where the rate of change of voltage is greatest, and then increases again, only to cease abruptly at $+V_p$.

Note that Fig. 2 shows quite a different signal frequency from Fig. 1, to emphasise that the probability density depends only on the wave shape and is independent of its frequency.

It is not difficult to derive the probability density diagrams for other common waveforms. For example, looking at Fig. 2, it is clear that if the sinewave were replaced by a triangular wave with its constant slope sides, the dip in the curve around zero would disappear, the probability density being constant between the limits $-V_p$ and $+V_p$. And the result would be the same for a triangular wave with a non-50:50 ratio, or even for a sawtooth waveform.

Real probability distributions

Measuring real probability distributions is really quite straightforward, if you are content to measure the cumulative probability rather than the probability density.

The one can be derived from the other, for, as noted above in connection with Fig. 1, the cumulative probability is the integral of the probability density. So if you measure the cumulative probability, differentiating it will give the probability density.

I made up a circuit to measure the cumulative probability of various waveforms, and this is shown in Fig 3. Op-amp IC_3 integrates a voltage of +15V or –15V, applied via R_5 from the output of IC_{1d}.

When the positive-going output of IC_3 reaches +7.5V, the output of IC_{1c} falls from +15V to –15V, while the output of IC_{1d} rises from –15V to +15V. The output of IC_3 therefore reverses direction and moves linearly towards –7.5V.

On reaching –7.5V, the cycle repeats, and a ±7.5V triangular wave of about 0.5Hz is applied to the reference input, pin 2 of comparator IC_2.

Depending on whether the signal input at pin 3 of the comparator is higher or lower than the voltage at pin 2, the comparator's output will be –15V or +15V. Filter R_4/C_2 smooths out the waveform, providing a dc level indicative of the probability that the signal voltage is lower than the reference voltage.

As the filter is such a simple one, only signals of very much higher frequency than the 0.5Hz triangular wave can be used with the circuit as it stands.

The filtered voltage level across C_2 was buffered by IC_{1b} and applied as Y deflection to a scope used in the XY mode. The triangular wave output of IC_3 was applied as the X deflection.

The arrangement will trace out on the screen the cumulative probability of any waveform applied

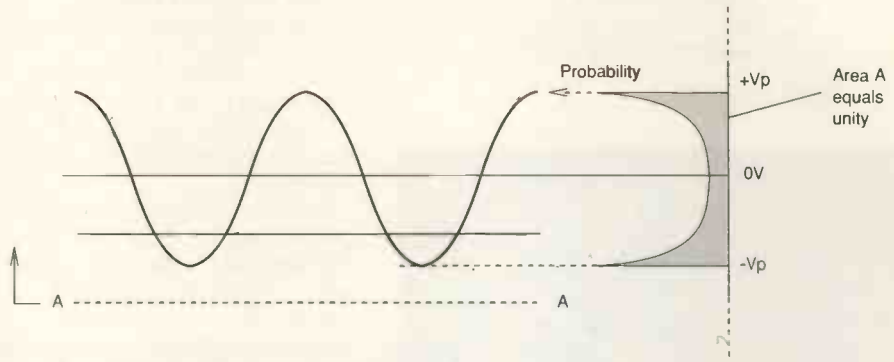

Fig. 2. Probability density diagram for a sinewave

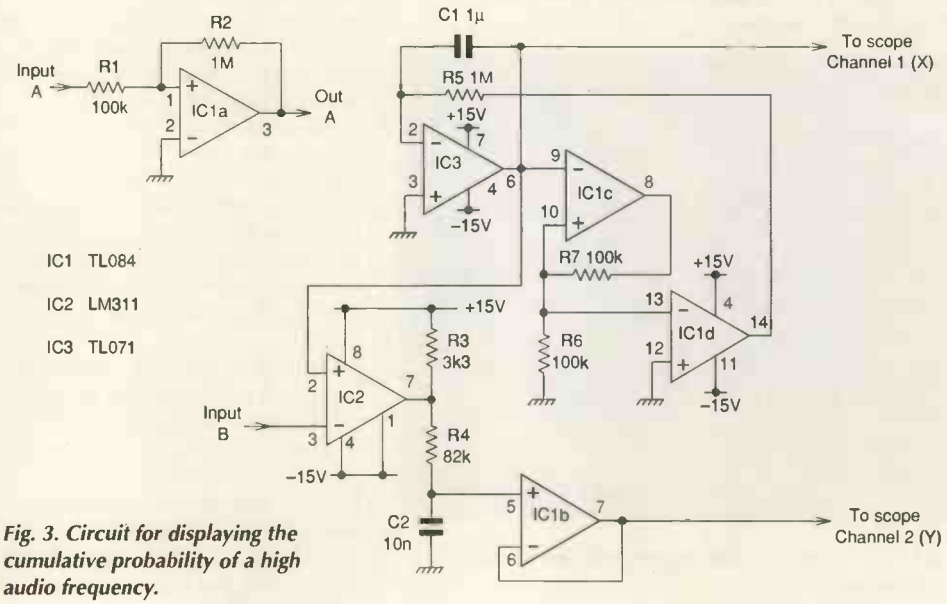

Fig. 3. Circuit for displaying the cumulative probability of a high audio frequency.

to the comparator, provided only that the peak-to-peak value of the waveform is less that ±7.5V.

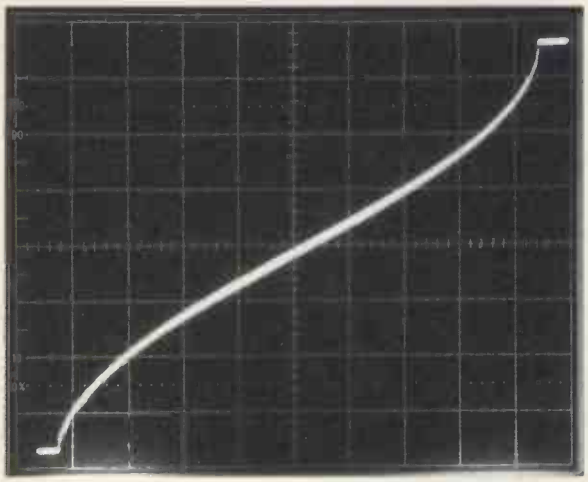

Fig. 4. Display of the cumulative probability distribution of a sinewave.

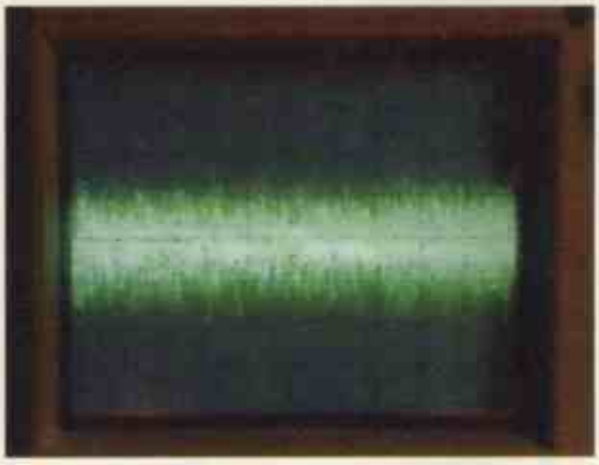

Fig. 5. High-pass filtered audio-frequency noise: 5V/div. vertical, 5ms/div. horizontal.

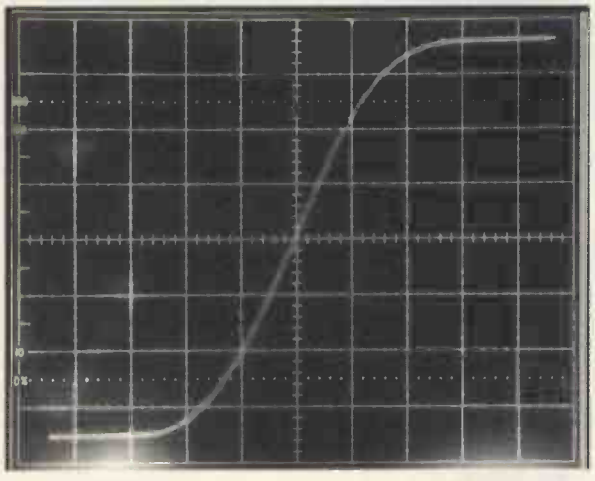

Fig. 6. Display of the cumulative probability distribution of Gaussian noise.

Some practical results

Figure 4 shows the result obtained for the cumulative probability of a sinewave. The horizontal deflection is provided by the ±7.5V triangular wave, and the deflection sensitivity has been adjusted to more or less fill the screen.

When the comparator output is permanently at logic zero or logic one, the limits of the filter output are ±15V, and the filter output supplies the vertical deflection. Again, the deflection sensitivity has been adjusted to more or less fill the screen.

Due to the 20 second exposure, Fig. 4 shows a complete sinuous trace, but during that time, the spot was repeatedly tracing out the curve, first from left to right and then back again, every two seconds. The fact that the go and return traces are identical shows that the filter time constant was not too long. If it had have been, the go and return traces would be slightly separated.

The amplitude of the applied ground-centred 20kHz sinewave was 13.1V peak-to-peak, and it was applied at input B. Thus the reference triangular wave recorded the cumulative probability from just below the negative peak to just above the positive.

When the reference is below the negative peak, at the extreme left of the trace, the probability of the comparator providing a logic 1 (+15V) output is zero, and the level out of the filter is constant at –15V.

When the reference encounters the negative peak, there is a discontinuity in the slope of the trace output by the filter, which starts to rise rapidly. The slope then reduces, being a minimum around the central portion of the sinewave, before increasing again only to revert abruptly to a horizontal straight line above $+V_p$.

If you were to plot the slope of the trace – i.e. differentiate it – it would be zero at the left, then rise abruptly before dipping in the middle, and being symmetrical about this point, just like that shown, on its side, in Fig. 2.

Figure 5 shows a very important waveform encountered in all branches of electronics and elsewhere, namely Gaussian noise. It is so called because the probability density distribution is the Gaussian or 'normal' curve.

The sample shown is band-limited white noise, that is to say it contains all frequencies between a lower and an upper cut-off point. The lower point was set at 10kHz, using the filters incorporated in the particular audio-frequency noise generator used.[1]

For this test, capacitor C_2 was increased to 47n, to avoid excessive thickening of the trace due to residual ripple from the filter. Also, in view of the limited output voltage available, it was applied to input A, output A being patched into input B.

In this case, the buffer stage IC_{1b} was ignored, and the filter output voltage across C_2 monitored directly with a 10MΩ, ×10 probe. This avoids the problem that the buffer cannot handle the full ±15V output of the filter. In fact, it would have been better not to use the buffer in the earlier tests; the discontinuities at each end of the trace in Fig. 5 should have looked even more pronounced.

Before recording the result, the amplitude of the noise was adjusted so that its peaks did not exceed ±7.5V. This proved difficult if not impossible to do by observing the waveform of Fig. 5 directly, so the output of the comparator at pin 7 was monitored instead. The amplitude of the noise input was reduced until momentarily the comparator output stopped changing state at each end of the ±7.5V reference excursion.

The resultant cumulative probability display is shown in **Fig. 6**. This looks very much like the textbook curve, and shows that the amplitude distribution of the noise-generator's output is indeed Gaussian.

A maximal length pseudo-random bit stream, or PRBS, circuit is used in the the noise generator to produce white noise. The PRBS is a series of logic noughts and ones, and this is converted to Gaussian by being low-pass filtered at well below the clock rate.

I had always assumed the distribution to be Gaussian normal, but it is nice to have confirmation. You can see that the slope of the curve increases, from left to right, smoothly to a maximum and then dies away again to zero. This corresponds exactly to the bell-shape of the normal or Gaussian distribution curve.

The degree of departure from the Gaussian normal characteristic could be further investigated in detail as follows. The triangular

waveform would be replaced by a dc level, accurately set by a potentiometer in steps of say 200mV from exactly –7.5V to +7.5V, with the aid of a DVM.

At each step, the dc level out of the filter should be recorded, also with the DVM, and the results tabulated. A $3^1/_2$-digit, or better still $4^1/_2$-digit, DVM will provide the resolution to record readings differing but little from the rail voltage, such as will be encountered at several standard deviations from the mean.

Filter output readings from –15V to +15V are then normalised to the range zero to unity by adding 15 and dividing by thirty. They can then be plotted on cumulative probability paper. If the distribution is Gaussian, the plot will be a straight line.

Cumulative probability paper has one linear axis, which would be used for the ±7.5V reference scale. The other is divided to nearly cover the range nought to one.

Probability of 0.5 occurs in the middle of the axis, increasing in cumulative normal fashion at one end to 0.9999 and decreasing to 0.0001 at the other. A suitable paper is 'Chartwell', Reference number D5572, though note that this gives probability in 'milliunits', so that the centre line is not 0.5 but 500.

Other applications

Probability density distribution is important in a number of electronics and communications applications. One such comes to mind from my days in defence electronics.

When trying to jam an adversary's communications, one will be modulating a transmitter to blanket the target bandwidth used by the enemy, without jamming one's own communications, quite probably in an adjacent part of the band.

The jammer could be frequency modulated with noise, but the resultant Gaussian distribution across the band would be unsuitable, due to the spill-over outside the target band. Wideband modulation with a waveform with precise limits would be better, perhaps wideband sinusoidal FM at a suitable rate of, say a few kilohertz.

The occupied bandwidth of a wideband sinewave modulated carrier is approximately given by $2(f_d+f_m)$ where f_d is the deviation – a few megahertz upwards in this case – and f_m is the modulating frequency – a few kilohertz, say. So the fall-off of power at the ends of the band is rapid and predictable.

Using sinewave frequency modulation though, the envelope of the discrete lines forming the jammer's output spectrum will have a dip in the middle, **Fig. 7a)**.

The similarity to Fig. 2 is clear, though the dip in the middle is less pronounced, due to the vertical scale being log – at 10dB/div – rather than linear.

On the other hand, wideband frequency modulation with a triangular waveform will provide a constant degree of jamming energy across the target band, though the roll-off beyond the band edges is not quite so compact, see **Fig. 7b)**. This is because the modulating triangular wave contains harmonics, unlike the sinewave modulation of Fig. 7a).

A somewhat better jammer may result from passing the spectrum of Fig. 7a) through a low-Q tuned circuit, partially flattening the power spectral density across the jammed band, and retaining or even slightly improving the rapid band-edge roll-off. ■

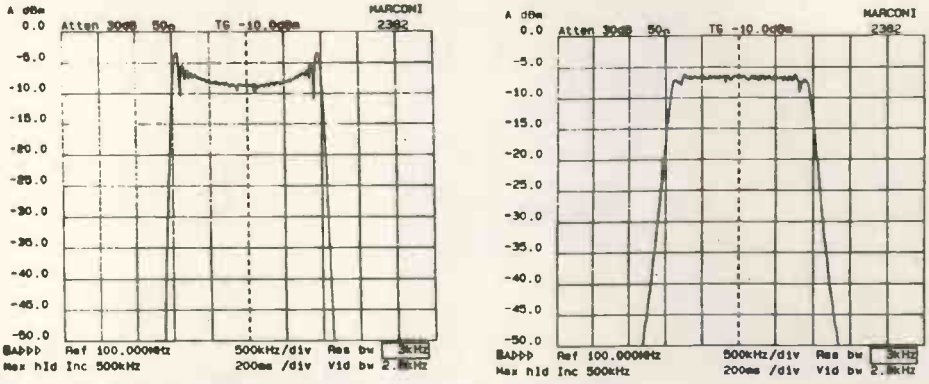

Fig. 7. Power spectral density of carrier, wideband modulated with a) a sinewave, b) a triangular wave

Forthcoming 'Beginners' corner' topics

This was to be the last article in the 'Beginners' Corner' series, but due to its popularity, Ian is currently working on further tutorials. This discussion on probability density distributions will be followed by articles on an audio PA stage using common-emitter devices driven from op-amp supply lines, transformer equivalent circuits, the superheterodyne receiver, balanced circuits, constant current generators, voltage references and linear ramp generators among others – not necessarily in that order.

Earlier in this series...

As explained in a preliminary article in the May 2000 issue, this series is intended to help students – and anyone interested in getting to grips with RF design – gain a background in practical electronic circuitry and troubleshooting.

Originally, the series was developed in response to the government's RF Engineering Education Initiative. Below is a list of the tutorials that have already appeared.

1 Timer circuit using the 555, June 2000 issue
2 Audio oscillator – Wien bridge based, July issue
3 h_{fe} tester, August.
4 Radio-frequency oscillator, Colpitts type, September.
5 Audio frequency filter/oscillator – state variable based, October.
6 Capacitance meter, November.
7 Radio-frequency oscillator involving negative resistance, December.

Reference

1. Hickman, Ian, Wide Range Noise Generator, *Wireless World*, July 1982, p 38.

Capacitance meter

There was an error in November's Beginners' Corner. On page 907, in the formula for C_o, the multiplication sign in the denominator on the right-hand side of the equation should be a minus sign. I missed it at the proof-reading stage; apologies.

SPEAKERS' CORNER

How does it radiate?

John Watkinson looks at how sound pressure waves propagate and interact.

The sensation of sound is a function of the velocity of the air. Displacement is the integral of the velocity. **Figure 1** shows that to obtain an identical velocity, or slope, the amplitude must increase as the inverse of the frequency. Consequently for a given sound-pressure level, low-frequency sounds result in much larger air movement than high frequencies.

Sound pressure is proportional to the volume velocity, U, of the source. Volume velocity is obtained by multiplying the vibrating area in square metres by the velocity in m/s. As sound-pressure level is proportional to volume velocity, as frequency falls the volume, or displacement, must rise.

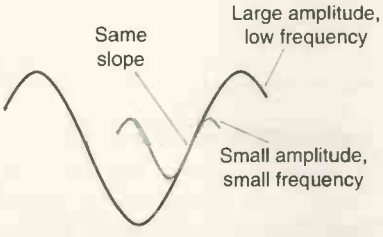

Fig. 1 Low frequencies must have larger air movements for the same sound-pressure levels as high frequencies.

This means that low-frequency sound can only be radiated effectively by large objects, hence all of the bass instruments in the orchestra are much larger than their treble equivalents. This is also why loudspeaker cone movement becomes great enough to see at low frequencies.

Volume velocity is measured in cubic metres per second and so sound is literally an alternating current of air. Pressure p is linked to the current by the impedance just as it is in electrical theory.

Electrical analogies

There are direct analogies between acoustic and electrical parameters and equations that are helpful. One small difficulty is that whereas alternating electrical parameters are measured in rms units, acoustic units are not. Thus when certain acoustic parameters are multiplied together, the product has to be divided by two. This happens automatically with rms units.

The intensity of a sound is the sound power passing through unit area. In the far field it is given by the product of the volume velocity and the pressure. In the near field, the relative phase angle will have to be considered.

Intensity is a vector quantity as it has direction, which is considered to be perpendicular to the area in question. Total sound power is obtained by multiplying the intensity by the cross sectional area through which it passes. Power is a scalar quantity because it can be radiated in all directions.

In a spherical sound wave, there is negligible loss as it advances outwards. Consequently the sound power passing through the surface of an imaginary sphere surrounding the source is independent of the radius of that sphere. As the area of a sphere is proportional to the square of the radius, clearly, the intensity falls according to an inverse square law.

Inverse square law?

Everyone has heard of the inverse square law, but unfortunately not everyone has heard that it does not always apply. It should be used only with caution because a number of exceptions to it exist.

Very close to a sound source, the proximity effect causes a deviation from the inverse square law for radiating objects that are small compared to the wavelength. The

area in which there is deviation from inverse square behaviour is called the near field.

In reverberant conditions, a sound field is set up by reflections. As the distance from the source increases at some point the level no longer falls.

It is also important to remember that the inverse square law only applies to near-point sources. A line source radiates cylindrically and intensity is then inversely proportional to radius. Noise from a busy road approximates to a cylindrical source. Flat panel loudspeakers that operate with chaotic bending waves are not point sources, and have an extended near field.

Wave theory

Wave acoustics is the sonic subset of wave theory is used in many different disciplines including radar, sonar, antenna design and optics. Consequently the designer of a loudspeaker may obtain inspiration from studying a radar antenna or a CD pickup, although most seem to get their inspiration from carpentry.

Figure 2 shows that when two sounds of equal amplitude and frequency add together, the result is completely dependent on the relative phase of the two. On the left of the diagram, when the phases are identical, the result is the arithmetic sum. On the right, where there is a 180° relationship, the result is complete cancellation. This is constructive and destructive interference. At any other phase and/or amplitude relationship, the result can only be obtained by vector addition.

The wave theory of propagation is based on interference and was originally proposed by Christiaan Huygens. The theory suggests that a wavefront advances because an infinite number of point sources can be considered to emit spherical waves which will only add when they are all in the same phase. This can only occur in the plane of the wavefront.

Figure 3a) shows that at all other angles, interference between spherical waves is destructive. For any radiating body, such as a vibrating object, it is easy to see from **Fig. 3b)** that when a radiating object is small with respect to the wavelength, only weak spherical radiation is possible.

Radiators large and small

On the other hand, when the radiator is large in comparison with the wavelength, a directional plane wave can be propagated, or 'beamed'. Consequently high-frequency sound behaves far more directionally than low frequency sound.

When a wavefront arrives at a solid body, it can be considered that the surface of the body acts as an infinite number of points that re-radiate the incident sound in all directions. When the body is large compared to the wavelength and the surface is flat, constructive

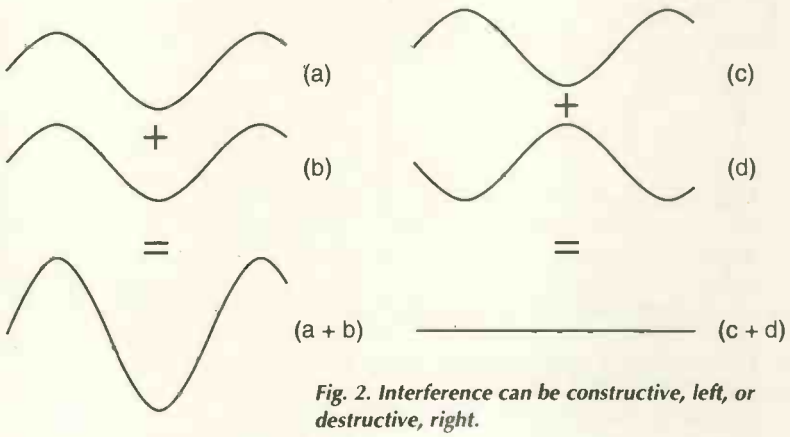

Fig. 2. Interference can be constructive, left, or destructive, right.

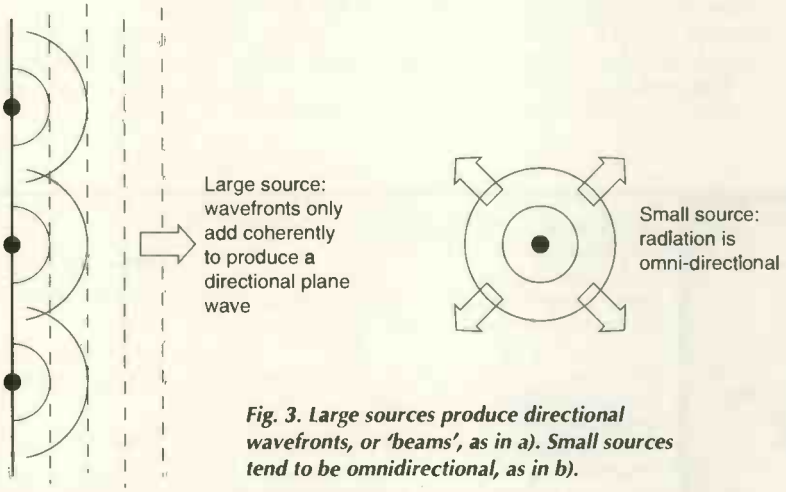

Fig. 3. Large sources produce directional wavefronts, or 'beams', as in a). Small sources tend to be omnidirectional, as in b).

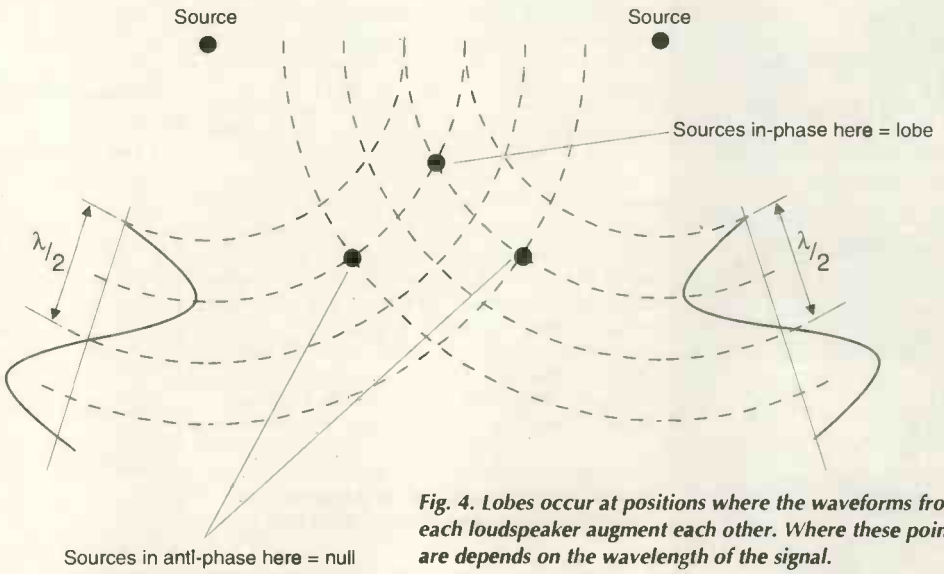

Fig. 4. Lobes occur at positions where the waveforms from each loudspeaker augment each other. Where these points are depends on the wavelength of the signal.

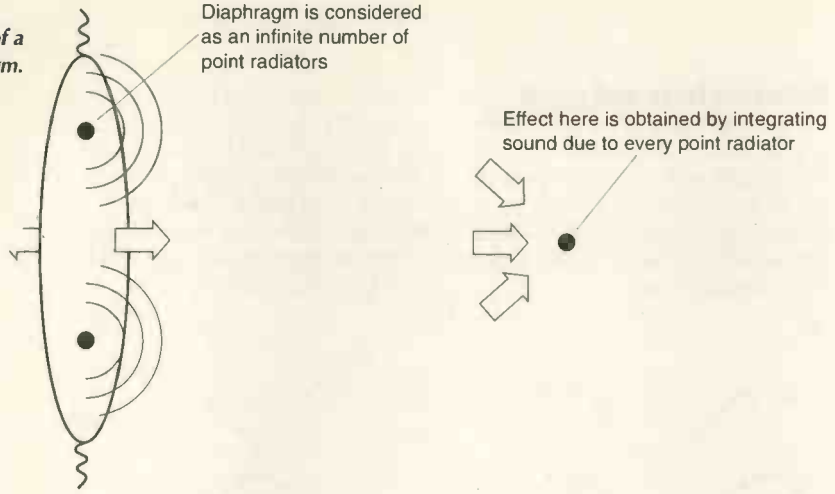

Fig. 5. Radiation of a real diaphragm.

interference only occurs when the wavefront is reflected such that the angle of reflection is the same as the angle of incidence.

If the body is small, the amount of re-radiation from the body compared to the radiation in the wavefront is very small. Constructive interference takes place beyond the body as if it were absent, thus it is correct to say that the sound diffracts around the body.

Figure 4 shows two identical sound sources that are spaced apart by a distance of several wavelengths, and which vibrate in-phase. At all points equidistant from the sources, the radiation adds constructively. The same is true where there are path length differences which are multiples of the wavelength.

Destroyed sound

However, in certain directions the path length difference results in relative phase reversal. Destructive interference means that sound cannot leave in those directions. The resultant diffraction pattern has a polar diagram consisting of repeating lobes with nulls between them.

The radiation of a pulsating sphere is interesting, but it does not model many real-life sound radiators. The situation in Fig. 4 can be extended to predict the results of vibrating bodies of arbitrary shape.

Figure 5 shows a hypothetical rigid circular piston vibrating in an opening in a plane surface. This is more like a real loudspeaker, or at least a woofer.

As the piston is rigid, all parts of it vibrate in the same phase. Following concepts advanced earlier, a rigid piston can be considered to be an infinite number of point sources. At an arbitrary point in space in front of the piston, the result is obtained by integrating the waveform from every point source.

Interestingly, the result of this integration is identical to the Fourier transform of the spatial disposition of the diaphragm. To see why, you will need to consider how transforms work. This will be the subject of a future article. ■

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DC-coupled valve power

Coupling capacitors and output transformer greatly simplify valve power amplifier design, but they both degrade linearity. Because of the inherently high impedance of valves, eliminating the output transformer is difficult. Numerous single-ended designs without coupling capacitors have appeared in the past, but there have been few DC-coupled push-pull amplifiers like this one from Kees Heuvelman and Wim de Haan.

This article was prepared by Wim de Haan, Kees Brakenhoff and Kees Heuvelman. The idea for the amplifier came from Kees Heuvelman.

Throughout the years I have seen many valve circuits. I have seen single-ended designs without coupling capacitors, but I have never seen a push-pull design that was DC coupled.

I believe that coupling capacitors have a major influence on the overall sound quality of an amplifier. The goal here was to produce a relatively simple push-pull valve amplifier using an output transformer, but without any coupling capacitors.

Valve amplifiers have appeared without an output transformer. But the benefits of transformerless designs are are outweighed by their high complexity and high component count.

We have found that single-ended designs are affected by speaker choice so we produced this DC-coupled amplifier using an ECC85 and EL84s in push-pull.

Design goals, amplifier No 1

The goal of this amplifier project was to design an amplifier with a low component count that was DC coupled throughout. It also had to be low cost and relatively easy to build.

Figure 1 shows the complete amplifier. Because one of the design goals was low component count, we had to come up with a simple input stage. A long-tail pair was chosen.

An ECC85 is used for its low impedance in combination with a relative high amplification factor. The low impedance is important because the valve has to work with a relatively low plate voltage. We have found the ECC85 to perform well in audio applications. It is also widely available and inexpensive.

For the output stage, we chose the EL84. It is also inexpensive and widely available. In push-pull mode, it can deliver over 10W.

Most valve amplifiers use a common-cathode circuit directly coupled to a long-tailed pair as driver stage. We thought, "why not skip the common cathode and find a way to couple the long-tail pair directly to the EL84 push-pull output circuit?" After all, the ECC85 can provide enough gain.

The EL84 needs –11V of bias for 40mA at 300V plate voltage. Thus, we had to find a way to couple the long-tail pair and the output stage in such a way that the 11V bias voltage remained.

Our power supply provides +390V. Knowing this allows the anode voltage of the ECC85 to be calculated. Maximum plate voltage of the EL84 is 300V, so cathode voltage has to be something like +85V. This is derived from,

$$V_{cath}=V_{psu}-(V_{plate}+V_{transformer}).$$

Now that the cathode voltage of the EL84 is known, it is easy to determine the plate voltage of the ECC85. As the cathode voltage is +85V, this has to be +74V in order to maintain the –11V bias voltage. The screen grid voltage should be 85V–11V=74V.

A transistor is used as a current source, setting current for the ECC85. Because of the high values of the cathode resistors, the EL84s automatically find the correct operating point.

The current source also makes the amplifier more stable via DC feedback by means of the two 120kΩ resistors. These resistors apply some correction for current differences in the EL84s by setting the voltage at the base of the transistor. This correction feature is not essential, but it works fine with the current source used.

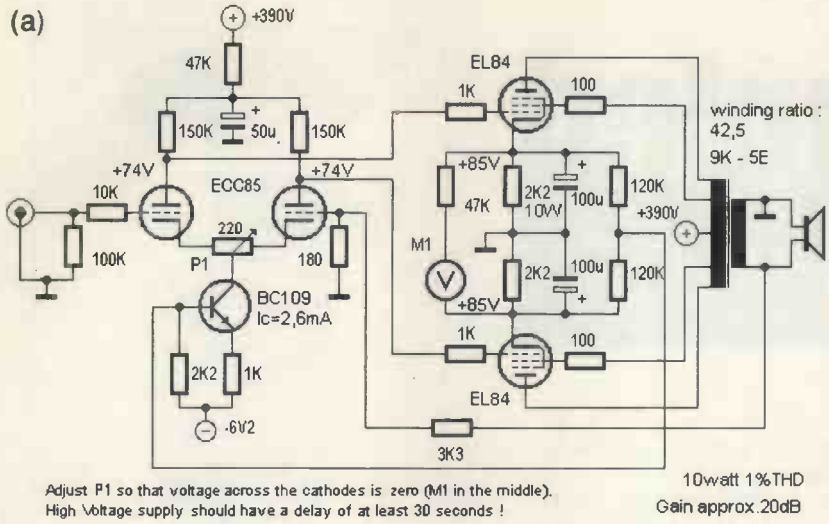
A negative supply is needed for the current source. We chose –6.2V for convenience, using the heater winding and zener diode to produce it. If you use 12V and modify the current source as shown in Fig. **1b)**, you can remove feedback and the amplifier will be very stable.

Using a cascode as a current source with a LED forming the reference should improve the current source even further. In this alternative, Fig. **1c)**, the DC feedback is again superfluous.

Potentiometer P_1 is 220Ω and is added to correct imbalance in the output stage. We used a voltmeter made from a small 50µA-0-50µA meter for adjustment. If the needle of this meter is in the middle, balance is good.

Adjustment for imbalance is not critical. A difference in the cathode voltages of 2.2V will result in a current imbalance of only 1mA. This is because of the large DC current feedback of the large cathode resistors.

Adjustment should be made 20 seconds after power-up and again after 15 minutes



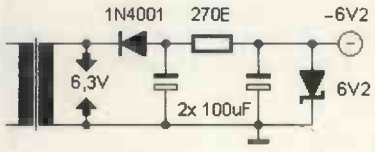
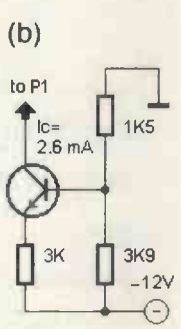
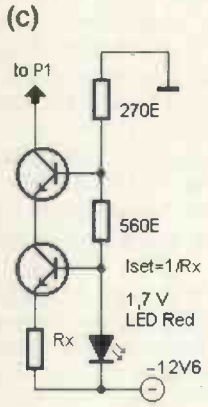
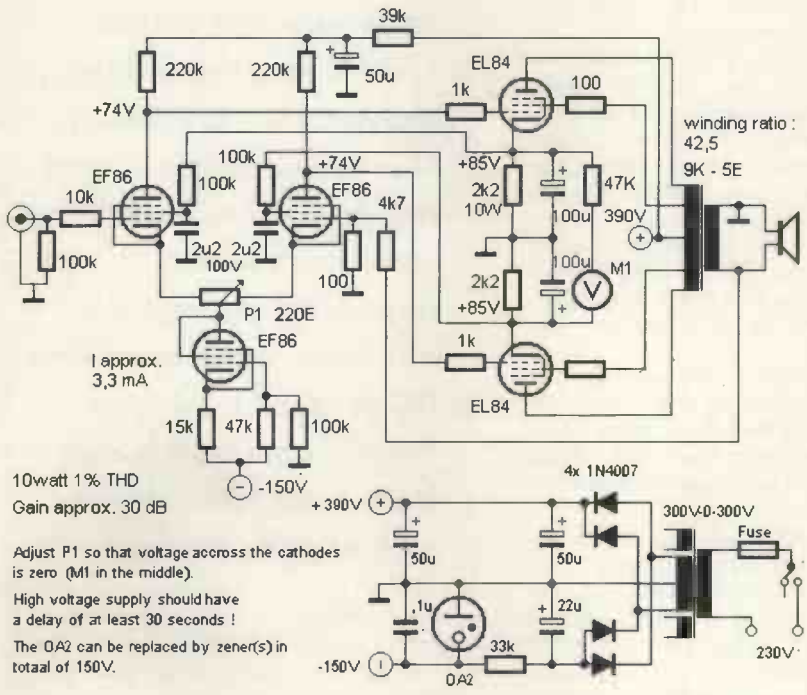


Fig. 1. Three-valve audio power amplifier using a transistor current source in the long-tail pair cathode. Insets b) and c) show alternative ways of forming the current source. In the original design, –6.2V was chosen as the source supply rail due to the convenience of the 6.3V heater winding.







or so. In this way the balance of the output stage remain stable over a very long period.

Overall feedback is accomplished by the 3.3kΩ resistor, the 5Ω transformer winding and the negative input of the long-tail pair. Total gain of the amplifier is approximately 20dB while overall feedback is approximately 6dB.

Transformer alternatives

The Amplimo VDV8020 toroidal output transformer would be the perfect choice as output transformer. The primary impedance of this toroidal transformer is 8kΩ and ultra-linear taps are provided. Frequency response is exceptional with –3dB points at 2Hz and 134kHz.

An extremely high coupling factor, very low leakage inductances and low internal capacitances have achieved this. There are no conflicting resonances below 100kHz. This performance is achieved through the use of multi-sectioned windings together with special combinations of series and parallel connections. The multi-segment secondary winding has a 5Ω impedance, contrived to be suitable for both 4Ω and 8Ω loudspeakers.

In 1993, Amplimo was the first company in the world to supply toroidal output transformers with a quality factor greater than 290 000. This factor is synonymous with frequency range.

Other suitable transformers are the Hammond 1608 or 1650F (USA), the BorderPatrol OTTP-610 (England), the Sowter U004, U064, UA23 or U082 (England).

A vintage Unitran 9U13 is used as output transformer throughout this project. This Dutch transformer is made back in the sixties. Unitran manufactured some fine transformers.

All transformers mentioned are still in production, except for the Unitran.

Amplifier 1 in summary

This amplifier circuit has no coupling capacitors, it works well and it is very stable.

As is common with DC-coupled valve amplifiers, the heater filaments must be up to temperature before the HT is applied to the circuit. This means a power-up delay of at least 30s.

Cathode capacitors on the output valves ensure that AC signals cannot feed back to the input stage through the current source transistor.

We designed this amplifier to give good performance yet remain simple. One area that could possibly benefit from a little extra complexity is the current source.

The suggested circuit works only with output valves biased fully in Class A. If for any reason the power supply voltage rises by, say 10%, then the output valves will be automatically overloaded.

In a recent design, we used a regulated DC power supply for the driver stage. This makes the circuit more flexible.

Amplifier design 2

We designed a second DC-coupled amplifier using three EF86s and EL84s in push-pull. Shown in **Fig. 2**, this circuit provides even better stability.

To enhance stability, the current source incorporates an EF86 pentode and the long-tail pair uses two more EF86s. Stability is improved because the DC differences in the output stage taken from the cathodes of the EL84s are fed to the screen grids of the EF86s of the long-tail

Fig. 2. Stability is improved by replacing the dual triode with two EF86 pentodes and using the same pentode as the active device in the current source.

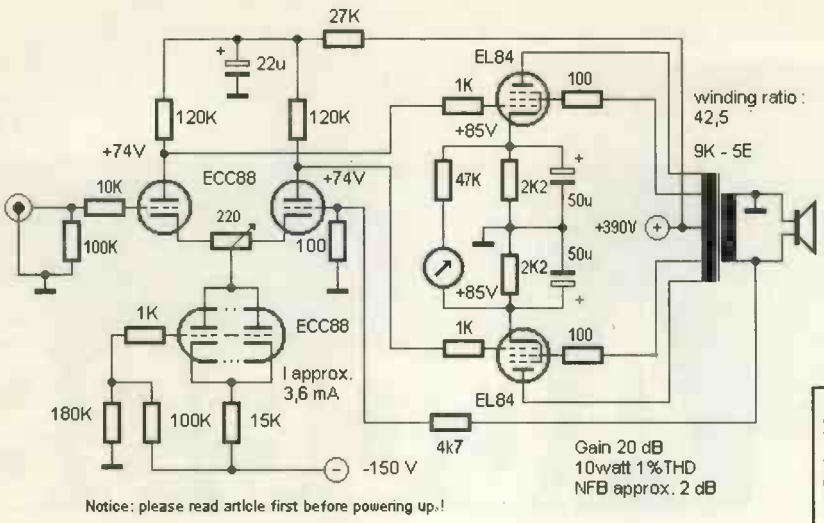


Fig. 3. Practical realisation of design 3, which provides good stability and an easy-to-implement circuit. Two ECC88 double triodes are used at the input.

pair. In this way each output valve has its own DC regulation, regulated by its associated EF86.

Using the EF86, the long-tail pair has a higher output impedance. This tends to cause a problem with high frequencies if no feedback is used. However, gain is high so more feedback can be applied to overcome this problem.

Square waves fed into this amplifier come out fine. Overall feedback is approximately 10dB.

Good stability, simpler circuit – design 3

We tried replacing the three EF86s with two ECC88s, **Fig. 3**. Stability was just as good. After 36 hours of continuous operation, the voltage across the cathode resistors of the EL84s changed only 0.3V.

Only 2dB feedback is used here. Current for the long-tail pair can be set by changing the 100kΩ resistor connected to the negative supply. It is approximately,

$$\left(150\frac{100k}{100k+180k}\right)\div 15k$$

or about 3.6mA. This configuration should also work with ECC85s.

Amplifier design 4

Figure 4 shows a practical DC-coupled design using an ECC85, a current source and two additional potentiometers. These pots are there to allow current and the high voltage of the input stage to be adjusted.

Current of the ECC85 is set to half that of ECC88. The other section of the ECC88 is used in the plus section of the power supply, in this way the voltage of the input stage is adjustable.

The grid of the ECC88 is controlled by the resistor divider between the negative rail and ground. Using the potentiometer, the high voltage can be easily set to approximate 206V.

The current source should be set to approximately 1.7mA. Bear in mind that the current source should be adjusted in so that the anode voltage of the ECC85 reads +74V. ■

Specifications, amplifier design 4

All measurements were made using a Hammond 1608 output transformer, a Tesla JJ ECC88, Tesla JJ EL84s, an RFT ECC85.

Output power at 1% distortion	11.9W
Sensitivity for 10W output power	700mV
Noise/hum related to 1W	78dB
Frequency response @ 1W –3dB	10Hz-72kHz,
Distortion at 10W @ 1kHz	0.55%
Distortion at 10 W @ 30kHz	2.35%
Distortion at 10 W @ 45Hz	2.7%
Damping 4Ω @ 100Hz	4x
Damping 4Ω @ 10kHz	5.3x

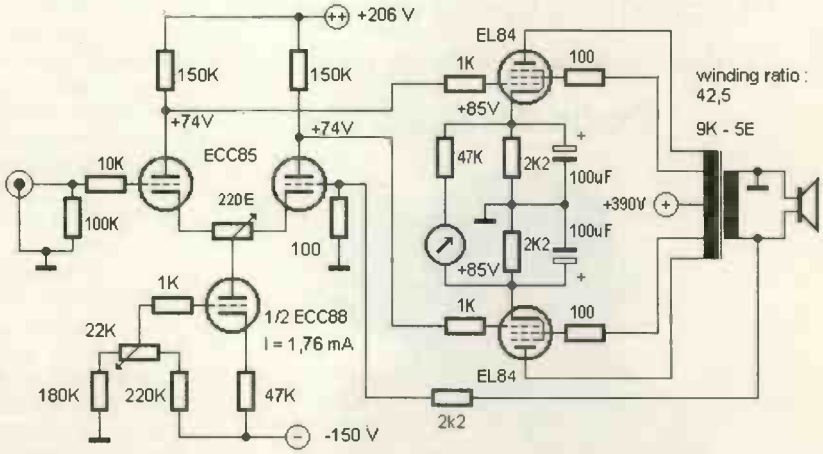
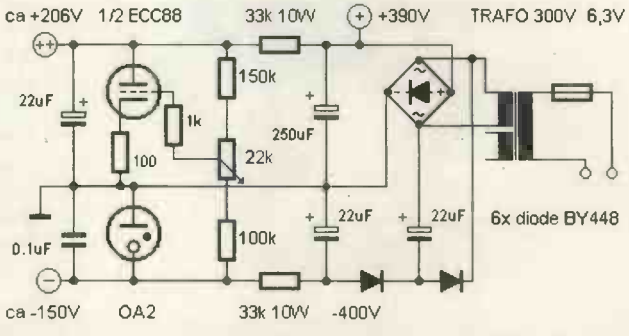


Fig.4. Of the four designs, this one is the most practical. It provides good stability and is easy to adjust.



Power supply for amplifier design 4 with a simple triode shunt regulator on the positive supply and a neon shunt stabiliser for the –150V rail.

Technical support

Amplimo toroidal transformers. Telephone: + 31 74 376 3765.
Website: http://www.amplimo.nl/
BorderPatrol transformers. Telephone: + 44 1273 276716.
Website: http://www.borderpatrol.net/
Hammond transformers: Telephone: + 1 716 651 0086.
Website: http://www.hammondmfg.com/
Sowter transformers. Telephone: + 44 1473 252794.
Website: http://www.sowter.co.uk/

Access

The UK is in danger of becoming the cave man of the telecoms industry if the market for broadband services is not opened up. Are the interests of operators being put before the needs of the economy? David Manners reports

Logical behaviour in the telecommunications industry has gone out of the window. The big US players, Lucent Technologies, Nortel Networks and Cisco Systems – all stock market darlings until recently – have seen their shares hit by reports that the telecommunications network operators are reducing equipment orders.

But in Europe, the biggest player of them all, Alcatel of France, reports sparkling business – sales up 50 per cent in the third quarter, data and optical networking up 88 per cent, and CEO Serge Tchuruk is very bullish about the state of the market.

"We're revising upwards our guidance from telecom revenue growth for the full year to the high thirties, with the increase from operations growing twice as fast," said Tchuruk, projecting that the demand for adding broadband technology to the world's telecommunications networks was robust.

On the other hand we see telecommunications network operators like BT digging their heels in on adopting broadband to the point where the UK regulator is hauled up before the House of Commons select committee, and a couple of the entrepreneurial network operators known as CLECs, which would like to sell broadband access to UK consumers, are reported to have pulled out of the market.

The UK is not alone in this. Across the world the anecdotal evidence is of telecommunications network operators doing what they can to slow down the adoption of broadband services. They create difficulties for customers who want them, delaying installation, making access to their premises difficult for new operators wishing to install their equipment, and charging both operators and customers too much to make the whole process viable.

This is why would-be operators, like the CLECs, are burning cash like dot.coms and not showing any returns. The amount of cash already used up varies from the $221m of Covad to the $871m of Rhythms. Most will have to return to the capital markets for more cash next year – which is reflected in 80 and even 90 per cent declines in the share prices of some CLECs.

Could the rollout of broadband-to-the-home technology, such as ADSL, be killed off in the same way as ISDN to the consumer was killed off in the 1990s? It seems an awful question to ask, with so much of the world's economic growth expected to come from e-commerce, which depends on upgrading the telephone network. But the question has to be asked.

denied

Could the interests of the national telecommunications network operators be put above the interests of the world economy?

Analysys, the Cambridge-based telecommunications consultants, says: "Competitively priced Internet access services are viewed as a critical component for EU global competitiveness". The political and economic stakes are too high for the telecommunications operators to stand in the way of the upgrading of the network.

Interestingly, the companies that are doing really well in providing the equipment to upgrade the network are finding contracts in countries where the national telecommunications operators have less political influence than they do in Europe and the US. For instance: Korea and Taiwan are big buyers of the equipment; last week Marconi announced a $550m order from Dubai.

The same irrationality, though not for the same reasons, seems to be occurring in the wireless market. Whereas Motorola and Ericsson announced weak results from handset sales in the third quarter citing declining markets – and Ericsson has long been rumoured to be considering exiting the handset market – No 1 handset player Nokia announced a fantastic third quarter, with sales up 59 per cent and revenues on track to top $30bn. What's more, it projected a record fourth quarter.

It must be awful for the executives at the handset divisions of Motorola and Ericsson to see Nokia's handset business storming ahead while they falter. Here the difference can only be in execution – Nokia actually came up with seven new handset models in the third quarter.

Something that Nokia does resonates with the public. Maybe it's the fact that its CEO Jorma Ollila gives priority to fashionability, ease-of-use and touchy-feely attributes like styling, over the engineering instinct to produce ever-more elaborate technological *tours de force*.

Latest models are said to allow you to customise your own ringing tone – like having your school song, and putting your own screensaver, say a photo of your partner, on the screen. Those ideas never came out of an engineering department.

So, in wireless communications, as in broadband wired communications, the market doesn't seem to be responding to trends, but to individual initiatives, varying abilities to execute, and opportunism. That's frustrating for analysts seeking trends and logical explanations, but its par for the course for an industry pioneering new frontiers. ■

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The irresistible force meets the immovable object

The irresistible force is the huge and growing human demand to communicate:

- We are sending 8bn e-mails a day – expected to rise to 26bn by 2005.
- We bought 400m mobile phone handsets this year; 60% more than last year. If that growth continues, 7bn hand-sets will be bought in 2006 by a world population of 6bn!
- Internet traffic doubles every 100 days.
- 350m people are on-line; 800m expected to be on-line in 2005. $650m in net commerce is transacted every year; $6.9trn is expected for 2004.
- The immovable object is the reluctance of telecoms network operators worldwide to upgrade the network...

The Big Question is: 'When does the network gridlock?"

Personal voice communications without a licence over a two mile radius for £75, exclusive.

Reader offer

A *pair* of two-way PMR radios for just £75*

RS 446 personal mobile radio...

To celebrate its launch, new test and instrumentation company Tecstar is offering *Electronics World* readers *two* RS446 personal mobile radios for just £75 excluding VAT and carriage.

Capable of transmitting and receiving voice over a distance of up to two miles, depending on terrain, the PMR 446 needs no licence. It offers eight channels, scanning – and with CTCSS up to 304 channel combinations.

A backlit liquid-crystal display shows volume level, channel number, sub-channel number, battery level and transmit/receive or channel busy. A unique call feature enables the user to alert the person they wish to contact.

Transmission distance is up to 2 miles. The radio has an accessory socket for an external headphone, earpiece or vox-microphone/headphone combination. A keypad lock and battery save feature are also standard.

The unit measures only 120 by 50 by 20mm and weighs less than 150 grams – including batteries. It is supplied complete with instructions and belt/mounting clip.

Compact, lightweight and low cost, the RS446 wireless personal-communications hand set has a wide range of applications. These include fetes, events and rallies. Builders on building sites could benefit from these radios, as could exhibitors at exhibitions and staff at warehouses, winter activities, sports events, maintenance departments, schools and care homes. Of course you can also use the RS446 just to keep contact with someone locally. The uses are almost limitless.

Send a cheques or postal orders to Tecstar Electronics Limited, 1 Nuffield Road, St Ives, Huntingdon Cambridgeshire PE27 3LX.
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RS446 key features...

- Up to 2 mile coverage radius
- Scans main and subchannels
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- Battery status indicator
- Auto battery save feature
- Keypad lock-out

What is CTCSS

CTCSS – or 'continuous-tone controlled squelch system' – allows sub channels of the main channels to be used. There are 38 sub channels to each main channel. Using subchannels decreases the likelihood that someone else will be using the same frequency.

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An 'ideal' transistor?

With the aim of helping you produce more effective analogue designs, Bryan Hart explains how it is possible to develop a set of 'ideal' transistor models that suit particular 'families' of applications. This first article deals with static models. A second will cover dynamic alternatives.

This article, and one to follow, emphasises the importance of the idea of a family of models for 'ideal' transistors. By 'ideal' transistor, I mean a device that is ideal for a given class of applications.

I describe circuit categories for which each family member may be regarded as ideal, either for a first-order understanding of circuit operation or for initial estimates prior to more detailed calculations and experimental work.

In this first article, static models are covered. Here, it is first necessary to consider, briefly, what is meant by an 'ideal diode' because that concept is central to the description of an ideal bipolar junction transistor, or BJT. In the second article, I deal with dynamic models.

Ideal-diode models

A p-n junction diode is shown schematically in **Fig. 1** and an 'ideal piecewise-linear' dc characteristic of it, together with its 'boxed' representation,[1] is shown in **Fig. 2**.

From a physical viewpoint, there is assumed to be an unlimited supply of charge carriers within the device to support the conduction process for one (forward) polarity of applied voltage but zero availability of carriers for the other (reverse) polarity. An analogue is a non-return valve. Such a valve permits the unhindered flow of fluid through a pipe in one direction but completely blocks the flow in the opposite direction.

The model in **Fig. 2** is useful in a first-order analysis of a wide range of applications that includes the following: rectification; clipping, limiting and clamping; and the diode pump circuit, used early in designs for pulse counters and frequency-to-voltage converters.

When the forward voltage drop of a diode is not negligible in comparison with signal and supply voltage levels, the model in **Fig. 3** can be used. The device is off when V_D is less than δ – typically 0.7V – but when I_D is greater than 0, it behaves like a battery being charged.

A limitation in the availability of suffi-

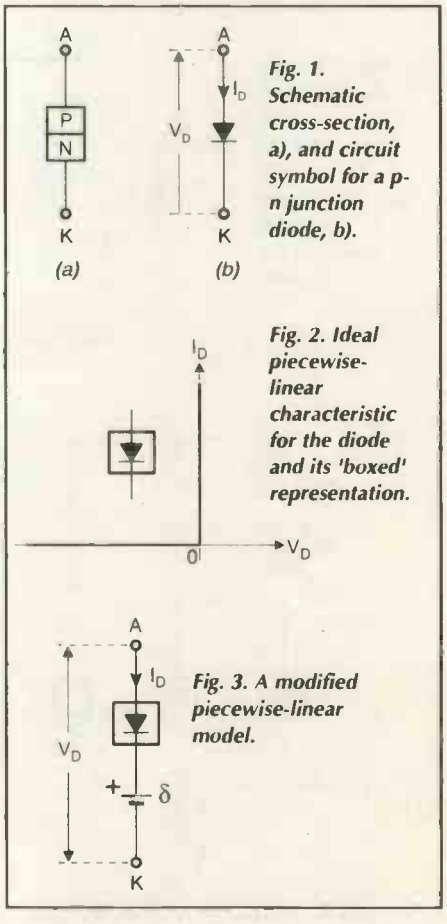


Fig. 1. Schematic cross-section, a), and circuit symbol for a p-n junction diode, b).

Fig. 2. Ideal piecewise-linear characteristic for the diode and its 'boxed' representation.

Fig. 3. A modified piecewise-linear model.

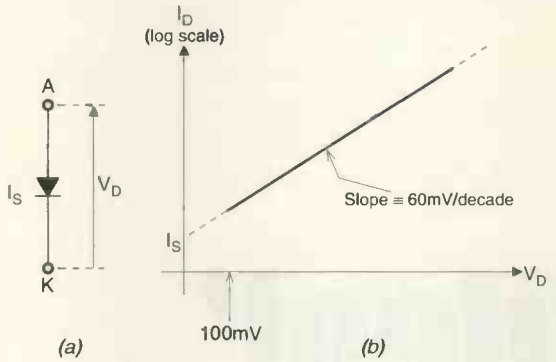


Fig. 4. The I_S-lettering denotes an ideal exponential diode, a), and the I-V characteristic, b). See also equation 1 in the main text.

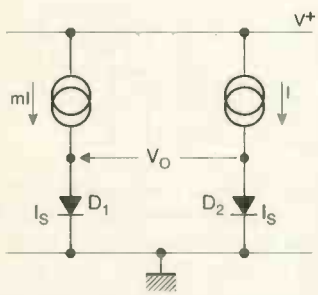


Fig. 5. An elementary p-n junction thermometer.

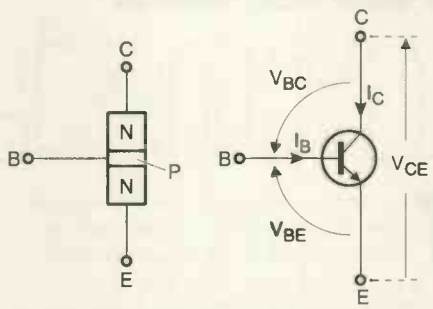


Fig. 6. Schematic cross-section, a), and annotated circuit symbol for an n-p-n-type bipolar transistor, b).

Fig. 7. Circuit model, a) and output characteristics, b), for an ideal piecewise-linear bipolar transistor for $V_{BC}<0$.

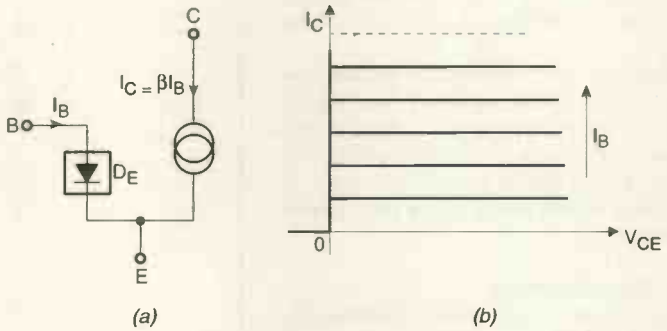
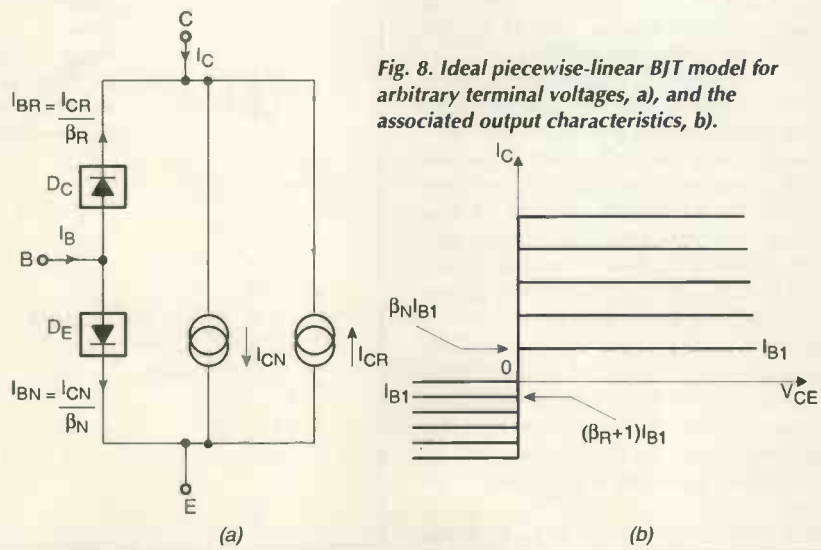


Fig. 8. Ideal piecewise-linear BJT model for arbitrary terminal voltages, a), and the associated output characteristics, b).



ciently energetic charge carriers in the p and n regions accounts for the 'ideal exponential' diode model. The model is conveniently represented **Fig. 4a)** by showing a general diode symbol with a current parameter, I_S, next to it. This is intended to indicate that,

$$I_D = I_S\left[\exp\frac{V_D}{V_T} - 1\right] \qquad (1a)$$

This equation typifies the *I-V* characteristic of a p-n junction. In it, the parameters have the following meanings: V_T, which is approximately 25mV at room temperature, is the product of Boltzmann's constant, k, with the absolute temperature, T, divided by the magnitude, q, of the charge on an electron; I_S, the reverse current when V_D is less than 0, is a parameter dependent on device geometry and doping levels.

Referring to equation 1a, the exponential Boltzmann factor is related to the fraction of two entities. The first is the number of holes approaching the edge of the depletion layer in the p region. The second is the number of electrons approaching the boundary of the depletion layer in the n region, that have sufficient thermal energy to surmount the junction potential barrier and enter the opposite region to become the minority carriers responsible for the conduction process within the device.

When V_D is greater than 100mV there is insignificant error in rewriting equation 1a) as,

$$I_D = I_s\exp\left(\frac{V_D}{V_T}\right) \qquad (1b)$$

A plot of I_D on a log scale against V_D on a linear scale is a straight line with extrapolated y axis intercept I_S, Fig. 4b).

A circuit for which the model is applicable is shown in **Fig. 5**. This forms the input stage of an elementary solid-state thermometer. If D_1 and D_2 have identical values of $I_S(<<I_D)$ but pass currents scaled in the ratio m:1, then,

$$V_O = V_T\log_e m \qquad (2)$$

and,

$$\left(\frac{dV_O}{dT}\right) = \left(\frac{k}{q}\right)\log_e m \approx 86.2\log_e m\ \frac{\mu V}{^\circ C} \qquad (3)$$

The parameter m is a design choice.

Ideal piecewise-linear BJT models

An n-p-n BJT is shown in shown in **Fig. 6** and its earliest, and simplest, model for $I_B>0$, $V_{BC}<0$ – the amplifying mode – in **Fig. 7**.

This is a current-controlled model. The ideal piecewise-linear diode, D_E, models the base-emitter p-n junction. Base current, I_B, in D_E controls the ideal collector current generator I_C. Here, I_C describes the transport of minority carriers across the base from the base-emitter junction, where they are inject-

ed, to the base-collector junction where they are extracted. As indicated, $I_C=\beta I_B$, where β is the well-known common-emitter direct current-gain factor.

As it stands, the model of Fig. 7a) has only limited practical use because of the neglect of a finite V_{BE}. However, by adding a voltage source, δ, of around 0.7V, in series with D_E, as in Fig. 3, you obtain a model that is useful in all but the most refined bias-circuit calculations in linear amplifier circuit applications.

An ideal piecewise-linear model of a BJT for switching DC and low-frequency signals is shown in **Fig. 8**, together with its associated output characteristics. In the common-emitter switching mode, both the emitter and the collector act simultaneously as sources and sinks for carriers crossing the base.

If you think of charge carriers as coins then a commercial bank is both a sink and source for them. Consequently, the BJT can be viewed as two transistors, each operating in the linear mode, connected in inverse parallel.

The component I_{BN} of base current in D_E that models the base-emitter junction controls the collector current component, I_{CN}, that flows in 'normal' operation. Hence the subscript N. This subscript is usually omitted when V_{BC} is less than 0, i.e., β_N is what was previously called β.

Diode D_C models the base-collector junction. Similar comments apply to the (reverse, R) current components,

$$I_{BR}=\frac{I_{CR}}{\beta_R} \qquad (4)$$

and I_{CR} associated with it. Normally, β_R is less than β_N because a BJT is intentionally designed to be asymmetrical.

This model is useful in demystifying the operation of the analogue-gate circuit in **Fig. 9**. I first used such a gate while designing an interscan target-marker symbol-generator for radar displays – long before the advent of cmos transmission gates.

Typical input and output waveforms are shown in **Fig. 10**. Assuming the impedance of C at the input signal frequency is much less than R_S, **Fig. 11** shows a load line for an instantaneous value of v_G, in this case the maximum value V_M, plotted on the output characteristic for,

$$I_B=\frac{V_{CC}}{R_B}$$

Switch S_1 in Fig. 9 models the action of a gating waveform. When this switch is at X, Q is on and v_O is zero, provided that,

$$(\beta_R+1)\frac{V_{CC}}{R_B}>\frac{V_M}{R_S} \qquad (5)$$

When the switch is at Y, Q is off provided V_M is less than V_{KK}. Then,

$$\frac{v_O}{v_G}=\frac{R_L}{R_L+R_S} \qquad (6)$$

Ideal exponential BJT models

Figure 12 illustrates an ideal exponential BJT model when V_{BC} is less than zero. If V_{BE} is more than 100mV,

$$I_C=I_S\exp\frac{V_{BE}}{V_T} \qquad (7)$$

I_S is a property of the base region. The change in base width with voltage is allowed

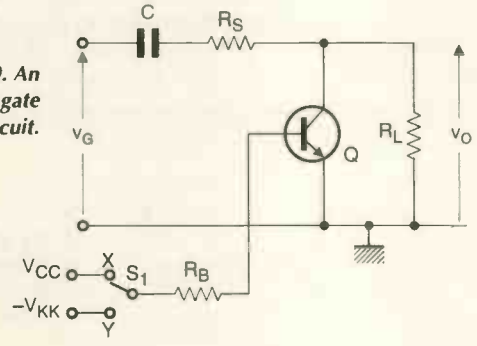


Fig. 9. An analogue gate circuit.

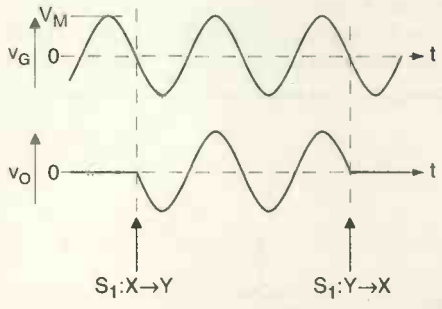


Fig. 10. Input and output waveforms, v_G and v_O respectively, for Fig 9.

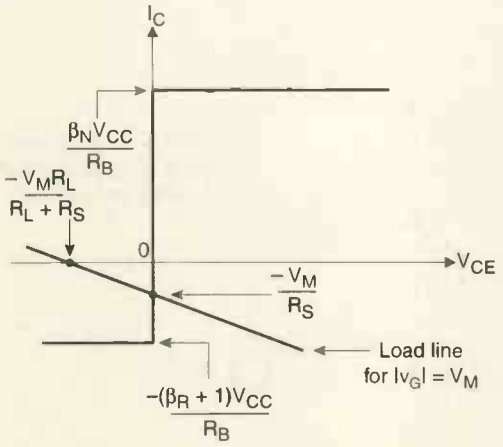


Fig. 11. Showing worst load-line position for satisfactory operation of the gate circuit.

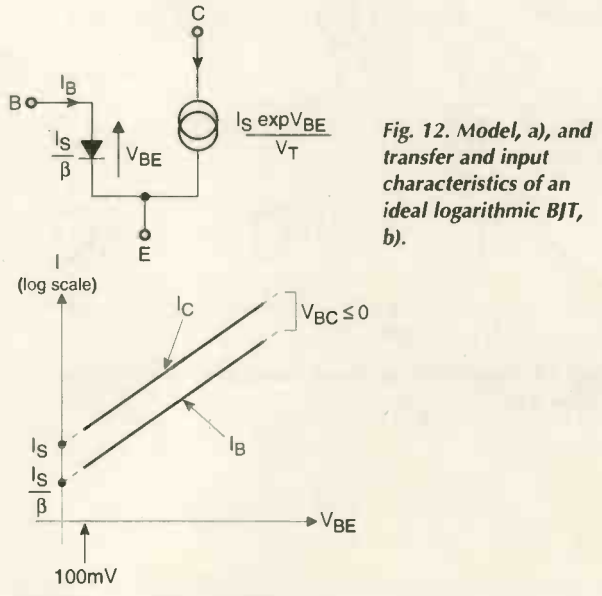


Fig. 12. Model, a), and transfer and input characteristics of an ideal logarithmic BJT, b).

for by multiplying I_S by a factor,

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

where V_A is the magnitude of the Early voltage.

Equation 7, which emphasises the voltage-control aspect of BJT operation, is the basis for the design of a wide range of bipolar analogue ICs. These include current mirrors, multipliers and 'root-law' circuits.

A simple root-law circuit using matched devices is shown in **Fig. 13**. By repeated application of equation 7, and the simplifying assumption that β is infinity, it can be shown that,

$$I_O=\sqrt{I_1^2+I_2^2} \qquad (8)$$

In practice, equation 7 is obeyed precisely over an I_C range exceeding seven decades for collector currents less than a milliamp.

However, the relationship between I_B and V_{BE} does not follow the ideal form of equation 7. Neither, incidentally, does the real relationship between diode current and voltage follow accurately that indicated in equation 1b. This is because of the existence of 'anomalous' current components such as that due to recombination of minority carriers in the base-emitter transition region.

The base current has been described graphically[2] as a 'sewer' for ill-behaved components of bipolar transistor current. That is why the function performed by the circuit of Fig. 5 can be more accurately achieved using BJTs instead of diodes. It is also why the logarithmic amplifier scheme in **Fig. 14** uses a BJT with its base earthed, rather than a diode for the feedback component.

An ideal exponential BJT model that applies for arbitrary polarities of V_{BE} and V_{BC} is shown in **Fig. 15**. It is a transport version[3] of the classic Ebers-Moll injection model[4]. For V_{BE}, $V_{BC}>>V_T$,

$$I_C=I_{CN}-I_{CR} \qquad (9)$$

$$I_{CN}=I_S\exp\frac{V_{BE}}{V_T} \qquad (10)$$

$$I_{CR}=I_S\exp\frac{V_{BC}}{V_T} \qquad (11)$$

$$I_{BN}=\frac{I_{CN}}{\beta_N} \qquad (12)$$

$$I_{BR}=\frac{I_{CR}}{\beta_R} \qquad (13)$$

In this saturated state, it turns out that if I_B and I_C change by the same numerical factor, then the change in V_{BE} is exactly matched by the change in V_{BC}, so V_{CE} remains constant. This fact is exploited in devising a simplified dc model[5] of the BJT in saturation that is useful in interface-circuit design. ■

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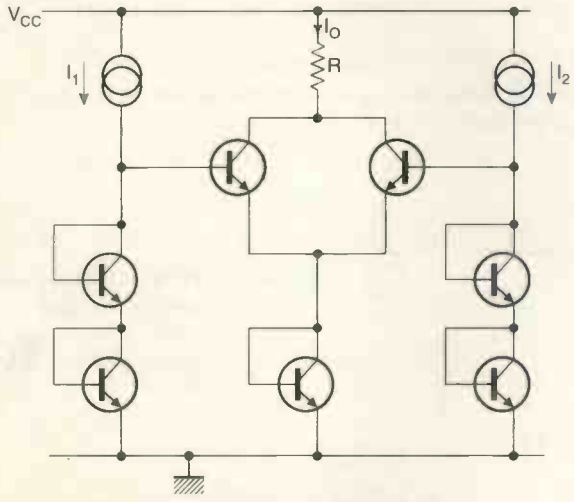

Fig. 13. Simple root-law circuit based on equation 7 of the main text: $I_O=\sqrt{I_1^2+I_2^2}$

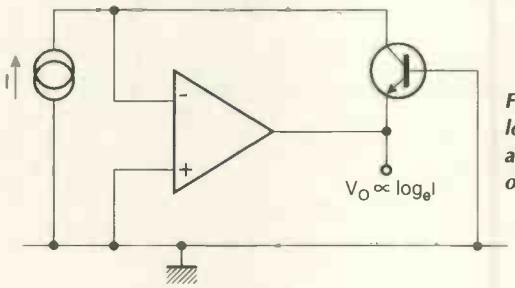

Fig. 14. A logarithmic amplifier based on equation 7.

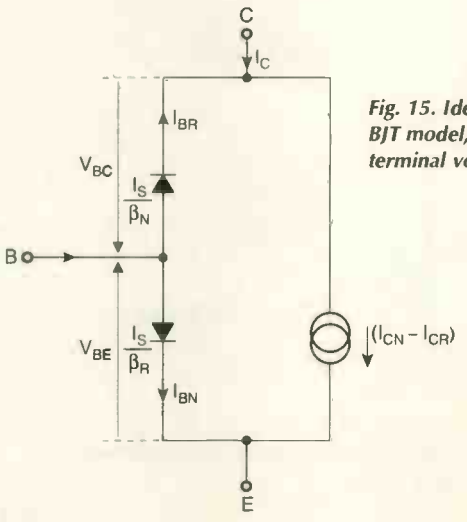

Fig. 15. Ideal exponential BJT model, for arbitrary terminal voltages.

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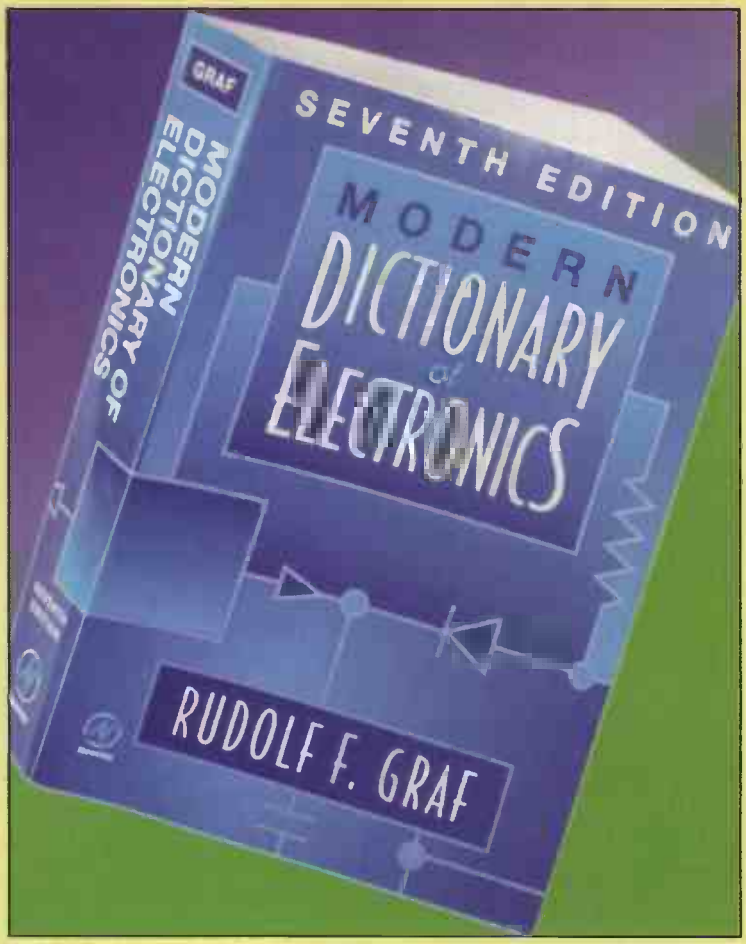

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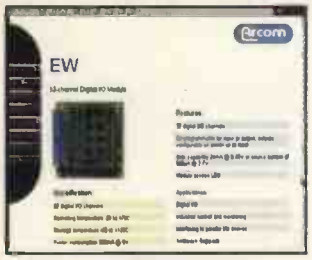

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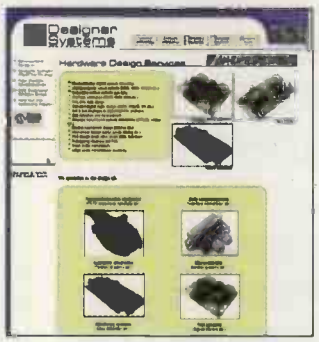

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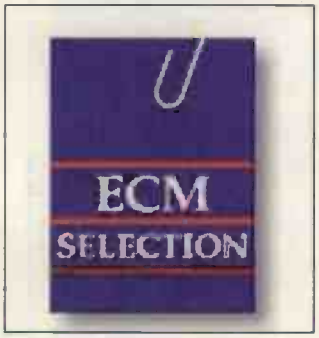

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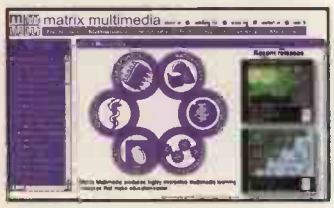

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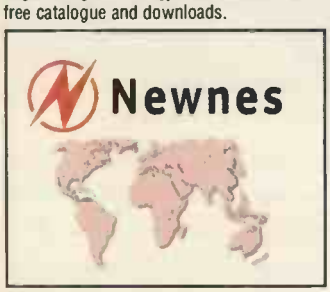

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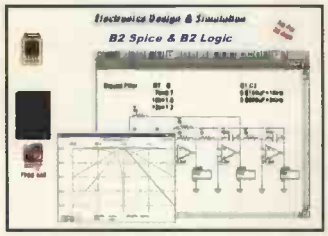

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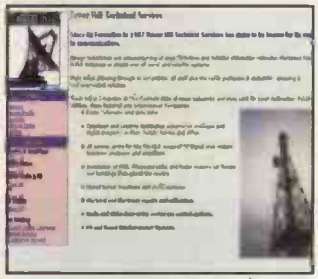

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The low-down on E numbers

With reference to the letter on 'E numbers and resistors' in the December 2000 issue, Mr Wells has clearly been dilatory in not learning the entire contents of his collection of *Electronics World* – formerly *Wireless World* – by heart.

A similar letter by one 'Scott' was published in the January 1984 issue, when Mr Wells' amplifier was approaching two years old. It received its retribution in a response from the *WW* stalwart 'Thomas Roddam'. I am fairly sure that this was one of the *WW* classic pseudonyms, published in the May 1984 issue.

The story goes back to the E6 series for ±20% tolerance resistors. The (British Standards?) committee dealing with the subject of preferred value components before the 1939-45 war deliberately deviated from the theoretically-correct values.

For practical reasons. 3.3 was chosen instead of 3.2 to get 'more colour contrast', presumably with 2.2. It hardly mattered technically, because '3.3' actually means '2.7 to 3.9' in the E6 series.

There was a choice to be made between 4.7/6.8 and 4.6/6.7, and again, the combination with the greater colour differences was chosen. Provided the manufacturer can sell all the resistors produced, as one preferred value or another, the values chosen as 'preferred' do not matter much.

Roddam does not discuss the 'modulation' of the E12 values in detail, but 2.7 is the geometric mean of 2.2 and 3.3, and 3.9 is the mean of 3.3 and 4.7. The 'modulation' of the E6 series has indeed influenced the E12 and E24 series, but the E96 series can use theoretical values because no colour-coding is involved.

Roddam's letter also discloses the way to design *RC* filters with E6 preferred values, by initially making the engineering approximation that 2π is 6.8 and using logs to base 6. Now, of course, one might take 2π as 6.2 and use logs to base 12 or 24.

John Woodgate
Via e-mail

Input-filter distortion

With reference to letters in the December 2000 issue, in his response to my own letter regarding the use of a series 10kΩ resistor plus 220pF shunt capacitance as an input filter, Phil Denniss is dubious to the point of opinionating that I fuss ridiculously. As all aspects must be addressed technically, please allow me to clarify, and hopefully promote a better understanding of the matters I raised.

First off, an apology is due. By the time I realised that I my computer had been set for a fast examination, and thus too coarse a series of examination subdivisions, the 0.3% distortion figure had already appeared in print. The distortion figure should have been only 0.1%, though the 15° phase shift at 20kHz was correct. I'm sorry.

Phil's understanding of linear fundamentals appears to be the same as mine. Of course a sine-wave that has passed through the filter will not show distortion on an audio test set, and the amplitude will be little altered. Nor will we be able to hear any distortion at 2kHz or any other test frequency.

What I relate to here is the way that we use signal generator inputs, and then examine the output wave-form *alone*. This is oh so wrong, because it leads to misleadingly good specification figures being produced in isolation. Just as we do with audio amplifiers!

We do not 'hear' musical tones in isolation; they are component parts of an intimately interrelated spectrum of recognisable fundamentals and harmonics.

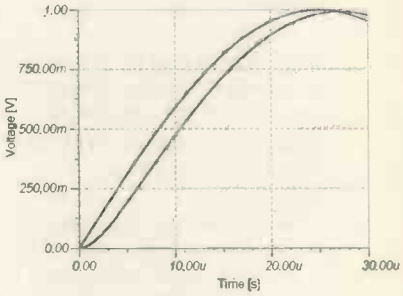
A 10kΩ resistor plus 220pF introduces a tiny delay, with a phase shift that increases sharply with frequency; this is separate to the individual, but series additional, delays and phase shifts that are due to our amplifiers themselves and the series output chokes they incorporate. Occasionally the latter two characteristics predominate and mask input filter effects.

This HF 'time' distortion – see the accompanying plot which shows in red the leading output edge of a 10kHz sine-wave that has passed through the filter – shows up as a loss of definition on transient and sibilant sounds, also on musical instruments that are rich in higher frequency harmonics.

A major point in my letter was to explain that these filter induced losses occur at all signal levels, and not just at ear damaging loudness when other amplifier problems might arise.

Clearly the 10kΩ+220pF filter causes a sine-wave lag that increases with frequency, as if the wave is replayed following a slight delay. Thus an initiated waveform becomes time lagged with respect to the input and all other frequencies that it had initially been specifically related to. The input signal did not stop to allow for this particular time delay to occur at just one frequency, and thus the filtered audio comprises a wide range of phase delays that relate to component frequencies at their time of passing.

This time I have illustrated and shown that there are measurable signal changes between a filter's source and its output – a difference that becomes superimposed upon the output with respect to the original input waveform. This distortion is equivalent to the original, plus an entire family of phase lagging harmonic components. These components are related to the instantaneous difference in amplitude between the input and output waveforms.



The right-hand curve is the leading output edge of a 10kHz sine-wave that has passed through the 10kΩ+220pF filter.

It is thus a fact that the suggested filter distorts my example of a 2kHz sine-wave by 0.1% with respect to input, and all other coexisting source signal there present.

The resulting reproduction might sound 'sweet' but it is not accurate, and our recognisable audible reference point becomes 'soft floating', constantly shifting as different sound frequencies dominate. I am not saying that anyone will notice this, but a system that has lesser frequency dependent propagation delays is instantly recognisable as being more transparent.

If anyone doubts my writing, will they please try switching this filter into and out of circuit with top flight audio gear using a microphone or CD source, because FM and vinyl simply might not be good enough. If you manage to hear the difference then you must admit that you are hearing signal distortion: You might find the sound acceptable, but I could never suggest that anyone must accept it.

There is nothing that is newly challenging or unconventional in this letter, yet it does challenge the conventionalised spot and swept frequency examination methods that we use to test our amplifiers.

It is realistic reproduction that I strive for. Over the decades I have come to understand that equipment does not always perform as its specification sheet implies; 'hi-fi' specifications tend to fall far short of what is necessary to regenerate realistic sound.

It was good to read John Well's letter on the same page. Hands-on

Will my ship ever come in?

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Paul Hampshire
Specialist Instrument Mechanic
Sasol (Secunda)
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experiences are always worth sharing, for not everyone can directly benefit from the unexpected empirical findings that challenge our theoretical bases.

Graham Maynard
Newtownabbey
Northern Ireland

In the response from Phil Dennis to the comment by Graham Maynard that a 10kΩ+220pF low-pass filter can introduce distortion, Phil states that this distortion 'defies understanding'.

Not so. Distortion from capacitors and resistors has been a known fact to component engineers for at least thirty years. See 'Harmonic testing pinpoints passive component flaws,' by V. Peterson and P. Harris of Ericcson, published in *Electronics* July 1966. And, 'Measurements of non-linearity in cracked carbon resistors', by G.H.Millward of the BBC, published in IEE paper 2747 Jan 1959.

Obviously if one used ideal components, such distortion is not possible. But in practice, depending on one's choice, it certainly can arise.

While the most likely culprit is the chosen capacitor, resistors are not blameless in this respect. Over the years, they have even been 100% production tested for this parameter.

In the heyday of the metal oxide resistor, many telecommunications makers specified a maximum permissible distortion for their resistors, initiated I believe by Ericcson.

Certainly my employers performed routine tests on their products using two specialist harmonic distortion testers and a 10kHz test frequency. These were high-speed commercially-available test equipments capable of testing ten components a second. I still have catalogue data listing some ten other references. (CLT1 Component Linearity Tester, March 1991)

As to the capacitor, the 220pF value quoted by Maynard raises some interesting choices. Almost all capacitor dielectrics exhibit two main deviations from the ideal. These are reduction in capacitance with frequency and dielectric absorption. Air performs best and electolytics – both aluminium and tantalum – worst.

Obviously for 220pF, you can forget electrolytic types. You are then left with a wide choice between plastic films, mica and ceramics. For cost and size reasons, the natural choice would be ceramic.

In the seventies, many reputable amplifier makers used tubular ceramic types in their preamplifier/tone controls. But these were usually restricted to low 'K' materials. This restriction resulted from comparative distortion measurements, performed I believe, by the Acoustical Company.

Today, my preferred choice would be to use only COG ceramic, polystyrene, polypropylene or polyphenylene sulphide. Of these the lowest cost would result from a COG ceramic-disc capacitor, the smallest size from a COG multilayer.

Using these capacitor materials with metal-film resistors should minimise distortion, but I suspect a measurable level would be found.

So in response to both correspondents, I suspect that Maynard may have chosen his capacitors unwisely, while Phil Dennis is assuming the use of near ideal components.

While I have no valid current comparative measurements to offer, I am willing to evaluate and investigate some modern components in order to prepare an article for *Electronics World* should there be sufficient interest.

Cyril Bateman
Acle
Norfolk

Has anyone got these back copies?

Can anyone help me with copies of the following please?

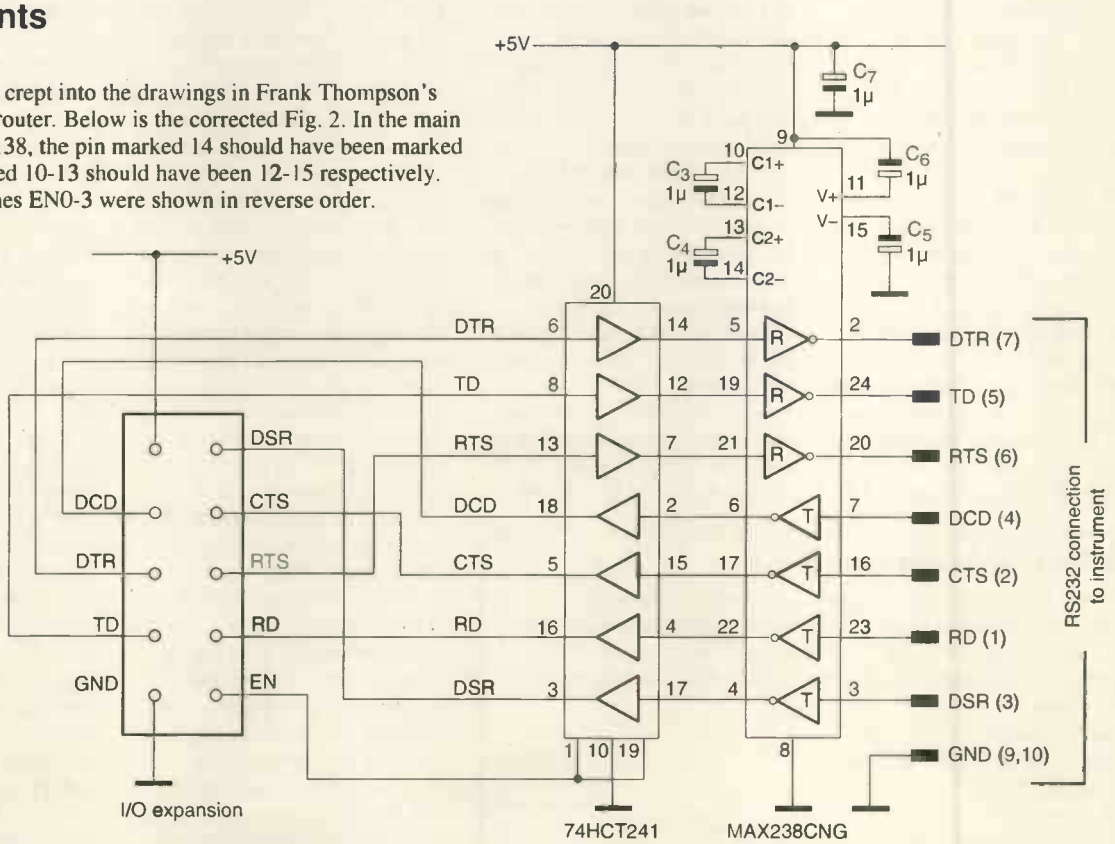
'Modulated pulse audio amplifiers,' Jan 1976, p. 76.
'Design of ceramic magnet loudspeakers,' Jan 1976, p.320.
'Class-D principles analysed,' Dec. 1967, p.576.
'A critique of Class-D amplifiers...' Dec. 1968, p.645.
'Distortions inherent in PWM Class-D amplifiers,' Dec. 1968, p.672.
'Compression chamber behind horn drivers,' Feb. 1971, p. 64.

My e-mail address is guidonoselli@outline.it. Fingers crossed.

RS232 port expander

Amendments

One or two errors crept into the drawings in Frank Thompson's four-way RS232 router. Below is the corrected Fig. 2. In the main circuit, on the 74138, the pin marked 14 should have been marked 16 and pins marked 10-13 should have been 12-15 respectively. Finally, enable lines EN0-3 were shown in reverse order. Apologies.



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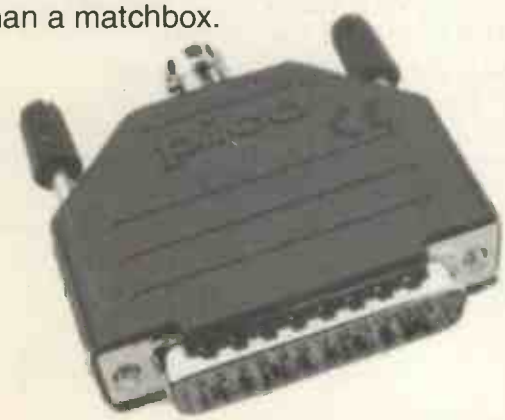

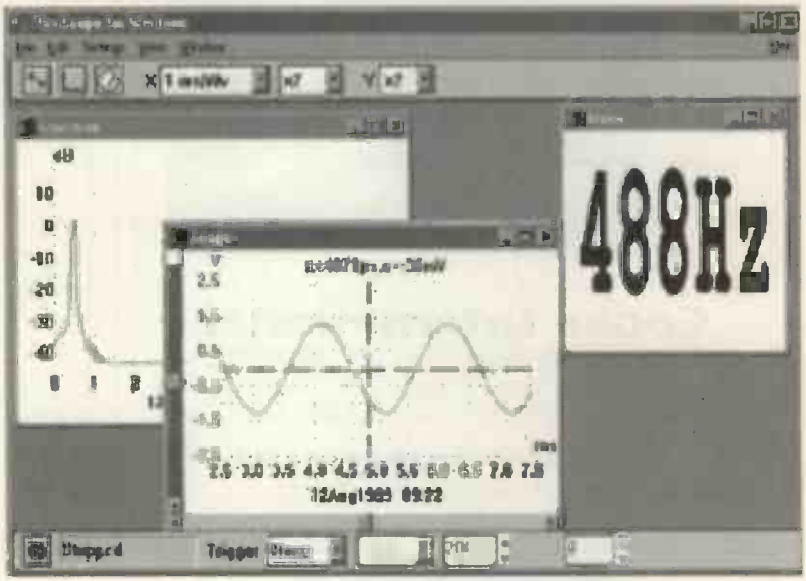

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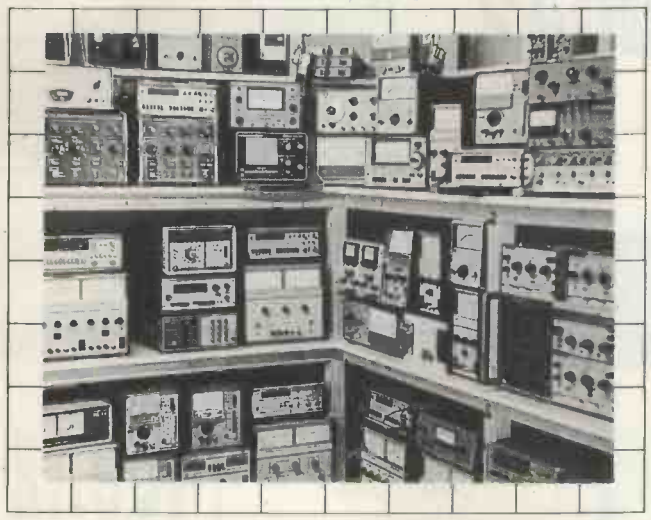

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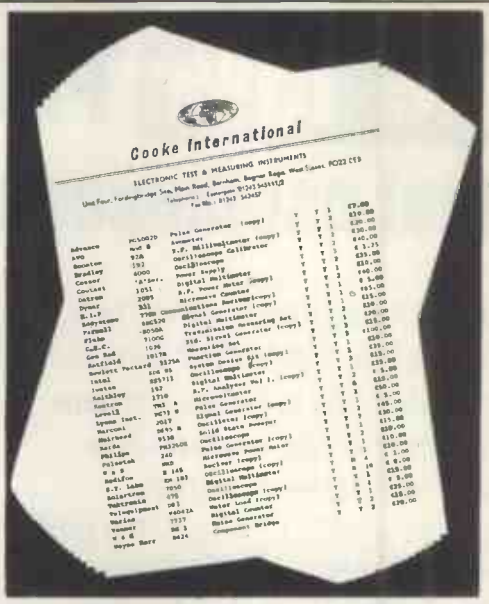

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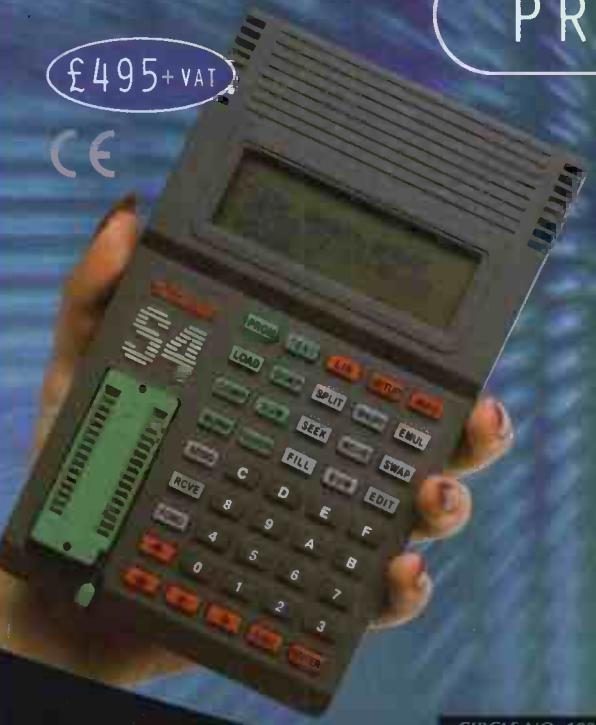

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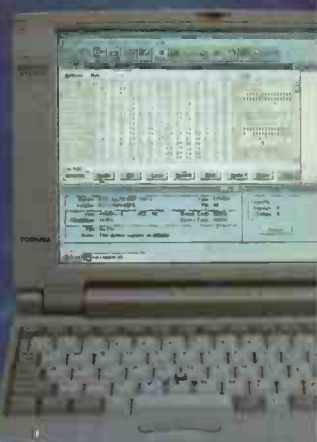

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