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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.

COMMS AT 136KHZ
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.

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

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. Ian Hickman gives this neglected topic an airing.

SPEAKERS’ CORNER: HOW DOES IT RADIATE?
John Watkinson looks at how sound pressure waves propagate and interact.

DC-COUPLED VALVE POWER
Numerous single-ended audio power amplifiers with no coupling capacitors have appeared in the past, but there have been few DC-coupled push-pull designs like these, from Wim de Haan et al.

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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 subcontinent – 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 dilated 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 more worldly-wise than former generations.

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 provision of more teachers and a relaxation of the rules about secondments to other government departments. The professional development of teachers and other staff in this area has not been given the attention it deserves.

There have already been protests from 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 of State for Employment John Butcher.

The Committee’s first report looks at the shortage of skilled graduates at Alvey, and estimates that there is a current shortage of some 10,000 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 spokesmen 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 emerges with a first degree to become a fully-fledged IT professional. 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-government doing this. It is a Labour government.

However, the more cynical of you will see it as just another example of the return of the third world. As if, when we need 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.
Consumer Association questions mobile phone hands-free test methods

Which? magazine’s second set of mobile phone hands-free kit test results were 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.

It claims that, in normal use, in-brain signal levels are mainly determined by the length of cable between the phone antenna, 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

UK 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.

“At 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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**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 reliably 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.

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**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.

Consequently 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.

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**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-Vitec 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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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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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.," he said.

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'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.

Nanomat's guide to contrast ratios

<table>
<thead>
<tr>
<th>Angle</th>
<th>0°</th>
<th>30°</th>
<th>45°</th>
<th>60°</th>
</tr>
</thead>
<tbody>
<tr>
<td>Book</td>
<td>9:1</td>
<td>8:1</td>
<td>7:5:1</td>
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</tr>
<tr>
<td>Newspaper</td>
<td>6:1</td>
<td>5:1</td>
<td>4:1:5</td>
<td>4:5:1</td>
</tr>
<tr>
<td>CH-LCD</td>
<td>4:5:1</td>
<td>4:1</td>
<td>4:1</td>
<td>3:1</td>
</tr>
<tr>
<td>STN LCD</td>
<td>4:1</td>
<td>3:1</td>
<td>2:1</td>
<td>0</td>
</tr>
<tr>
<td>ECD</td>
<td>18:1</td>
<td>7:5:1</td>
<td>6:5:1</td>
<td>3:5:1</td>
</tr>
</tbody>
</table>

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

SETN LCD - LCD Newspaper 9:1 ECD 18:1

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, 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.

Steve Bush
TEKTRONIX 577 Curve tracer + adaptors -T900.
Johns Radio, Whitehall Works, 84 Whitehall Road East, Birkenshaw, Bradford BD11 2ER. Tel: (01274) 684007. Fax: 651160

TEK P6701 Optical Converter 700 MC/S-850 £250.

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Anritsu MN95B Variable Att. 1300 £100.

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HP3455/3456A Digital voltmeter - £

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HP86222A+B Sweep PI -01-2.4GHz + ATT £1000-11

HP3586A or C selective level meter - £500.

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HP432A-435A or B -436A -power meters + powerheads to HP180TR. HP1B1T, HP182T mainframes £300 - £500.

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Marconi mod meters

Racal/Dana 9300 RM


6650PI - 18-26.5 GHz or 6651 PI - 26.5-40GHz- £280 each.

L Bridge type TF2700 -1150.

M - 9009A £350.

8411a -8412 -8413 - 8418 -600. - 110Mc/s - £500 - £1000.

HP Sweep Oscillators type 8690 A+B + plug-ins £500. Both £1000.

Racal/Dana 9300 RM

Marconi TF2374 Zero Loss Probe - £200.

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HP3709B Constellation ANZ 11500.

Heads 11664 Extra - 1150 each.

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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.

A sked 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 telecoms 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 the denials 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 does what the official site advocates, i.e., connects to the Web site, but then, instead of being displayed on the screen, is surreptitiously by altering the user's dial-up number. The so-called viewer program does exploit weaknesses in the site's software; when a new user connects, the program changes the user's number to one of the fraudsters'.

Suchsites are more than the usual terms; in fact they are targeted to individuals (DUN) files.

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 going to change. It's up to us, the members of the telecommunications community, to represent the whole UK telecommunications industry, mobile, cable and fixed-network operators. TUFF now has 36 member organisations which represent over 15% of the UK's telecommunications customers."

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, the commission of an offence. As well as legal and technical solutions, it also gives law enforcers a much stronger weapon for punishing the perpetrators. The Act 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.

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

Legal issues

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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 subservive material can be found throughout the Web and in particular at http://www.members.tripod.com/~SeusslyOne/F/AQ.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 people being asked to present identity checks.

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. And the fraudsters are often multiple criminals, who will keep their victims on the phone as long as possible, at charges reported as up to $25 per minute. The scamsters then try to get the victims to call or fax a number in the 01932-251200. ‘Guidance and Best Practice for Avoidance of Dial-Through Fraud’, from industry body BABT. Phone 01932-251200.

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 $240000 in cash before the police discovered the front companies.

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 behaviour 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 $50000 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 £12000 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 $100000 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 calls charge 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.

Easy peasy

A highly reputable telecoms 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 consultant 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 Avoidance of Dial-Through Fraud', from industry body BABT.
- 'Avoiding PBX Fraud – working actively to safeguard your system', from phone company Telewest.
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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

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.

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.


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

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

Here, \( R_r \) is radiation resistance in ohms, \( h \) is antenna height in metres, and \( \lambda \) 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 \( \eta \) of this kind of arrangement is very poor and the resulting radiation resistance is only few milliohms,

\[
\eta = \frac{R_r}{R_L + R_G}
\]

considering that,

\[
R_L << R_r + R_G
\]

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,

\[
C_H = \frac{24l}{\log \left( \frac{d_H}{4l} \right)}
\]

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:

\[
l_t = \frac{C_H I_b}{C_H + C_v}
\]

Here, \( l_t \) 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,
Atmospheric conductivity for various terrains.

Fig. 3. Ground conductivity for various terrains.

dBuV/m 1W ep

Ground Propagation
Sky Wave Propagation
Ground Propagation
Ground Propagation

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

dBuV/m

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

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

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 10⁻¹⁰ to 10⁻⁹Ω/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 rod7.

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 1000Ω/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 0.016Ω/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 3mm² 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 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, a well known paper from J. A. Adcock 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.

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 are at 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 1.78dB - 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 almost always. This means that the gain is always 4.78dB, 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 Ga.

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 138.8kHz station with 40kW radiated power - are shown on the web site of DK8KW.

**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 tests 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 in Canada. These tests were carried out at 107kHz with a bandwidth of 4500Hz and are shown in Fig.

**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 30kHz.

---

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

<table>
<thead>
<tr>
<th>Height (m)</th>
<th>Degree</th>
<th>R1 (Ω)</th>
<th>R2 (Ω)</th>
<th>R3(*)</th>
<th>Po1 (mW)</th>
<th>Po2 (mW)</th>
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</thead>
<tbody>
<tr>
<td>5.0</td>
<td>197</td>
<td>0.0020</td>
<td>0.0081</td>
<td>0.0070</td>
<td>3.49</td>
<td>224</td>
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<tr>
<td>6.0</td>
<td>236</td>
<td>0.0029</td>
<td>0.0117</td>
<td>0.0101</td>
<td>5.04</td>
<td>323</td>
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<tr>
<td>7.0</td>
<td>276</td>
<td>0.0040</td>
<td>0.0159</td>
<td>0.0139</td>
<td>6.95</td>
<td>445</td>
</tr>
<tr>
<td>10.0</td>
<td>394</td>
<td>1.14</td>
<td>0.0081</td>
<td>0.0324</td>
<td>0.0282</td>
<td>14.10</td>
</tr>
<tr>
<td>14.0</td>
<td>551</td>
<td>2.28</td>
<td>0.0159</td>
<td>0.0636</td>
<td>0.0554</td>
<td>27.69</td>
</tr>
<tr>
<td>20.0</td>
<td>787</td>
<td>3.26</td>
<td>0.0324</td>
<td>0.1298</td>
<td>0.1129</td>
<td>56.46</td>
</tr>
<tr>
<td>28.0</td>
<td>1102</td>
<td>4.56</td>
<td>0.0636</td>
<td>0.2544</td>
<td>0.2216</td>
<td>110.80</td>
</tr>
</tbody>
</table>

| G4 = 3dB, comprising gain of short vertical plus ground reflection. |
|------------------------|------------------------|------------------------|------------------------|------------------------|------------------------|
| Ibw = 0.85             | Ibw = 0.85             | Ibw = 0.85             | Ibw = 0.85             | Ibw = 0.85             | Ibw = 0.85             |

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

<table>
<thead>
<tr>
<th>Conditions</th>
<th>BW = 100Hz</th>
<th>BW = 25Hz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Summer day</td>
<td>-15 dB</td>
<td>-21 dB</td>
</tr>
<tr>
<td>Summer night</td>
<td>-5 dB</td>
<td>-11 dB</td>
</tr>
<tr>
<td>Winter day</td>
<td>-30 dB</td>
<td>-36 dB</td>
</tr>
<tr>
<td>Winter night</td>
<td>-18 dB</td>
<td>-24 dB</td>
</tr>
</tbody>
</table>
5. Of course, man-made noise is also important, but it is unpredictable. 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 1202 line – i.e. the telephone's twisted pair – has an output of +20dBm.

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,

$$Q = \frac{2\pi f L}{R_{AC}}$$

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.

<table>
<thead>
<tr>
<th>Coil type</th>
<th>Wire dia (mm)</th>
<th>Turn length (m)</th>
<th>Wire length (m)</th>
<th>L (mH)</th>
<th>Q*</th>
<th>RAC (Ω)</th>
<th>136kHz</th>
<th>200kHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Grey PVC</td>
<td>0.90</td>
<td>175</td>
<td>87.6</td>
<td>2.43</td>
<td>2665</td>
<td>205</td>
<td>13.0</td>
<td>3919</td>
</tr>
<tr>
<td>Air + wood</td>
<td>0.18</td>
<td>120</td>
<td>76.0</td>
<td>1.38</td>
<td>1196</td>
<td>513</td>
<td>2.33</td>
<td>1758</td>
</tr>
<tr>
<td>Air + wood</td>
<td>0.18</td>
<td>157</td>
<td>72.2</td>
<td>1.63</td>
<td>2306</td>
<td>507</td>
<td>4.53</td>
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<tr>
<td>Air + wood</td>
<td>0.90</td>
<td>85</td>
<td>87.6</td>
<td>2.43</td>
<td>2618</td>
<td>237</td>
<td>11.9</td>
<td>4145</td>
</tr>
<tr>
<td>Air + wood</td>
<td>0.90</td>
<td>85</td>
<td>165</td>
<td>2.43</td>
<td>2135</td>
<td>318</td>
<td>6.71</td>
<td>3140</td>
</tr>
<tr>
<td>Air + wood</td>
<td>0.18</td>
<td>94</td>
<td>168</td>
<td>1.96</td>
<td>2562</td>
<td>597</td>
<td>4.28</td>
<td>3768</td>
</tr>
</tbody>
</table>

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,

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

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. \( D/Le \), 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,

\[ N = \frac{5.08 \times 10^5 L_p + \sqrt{(5.08 \times 10^5 L_p)^2 + 4.572 \times 10^5 LD^2}}{D^2} \]

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 LO6 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 resistance 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, \( R_{AC} \), 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.

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.

\*\( Q = \frac{X_L}{R_{AC}} \)
RF DESIGN

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.

Our first loop antenna, Fig. 12, has an area of 0.48m² and uses 18 turns, resulting in a Q of 50. Its equivalent height at 136kHz, where $\lambda$ is 2206m, is,

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

At 136kHz, the operational Q is always near 50. To have a working bandwidth of not less than 2kHz 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,

$$h_e = KxAN = 0.142xAN$$

For the new version of the loop, with an area of 2.0m², 44 turns and the same Q, the effective height will be,

$$h_e = KxAN = 0.142x44 = 0.142x44x2 = 12.5$$

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.

References

4. ON7YD, 'Antennas for 136kHz,' www.qsl.net/0n7ys/136ant.htm
14. DK8KW, 'LF field strength measurements,' http://home.t-online.de/home/dk8kw/index.html

Further reading

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### HARD DISK DRIVES

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High-frequency impedance 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:

- 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 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 known3.

To confirm these methods, I modelled a theoretical 200 µF 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 millivoltmeter. To phase meter reference channel

To phase meter reference channel

To phase meter measurement channel

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.
voltmeter. 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,

\[ |Z| = R_{\text{in}} \times \frac{V(1) - V(2)}{V(2)} \]

As already proved in Fig. 2,

Here, \( V(1) \) and \( V(2) \) are the complex difference voltages while \( V(2) \) is the voltage drop across the capacitor current sense \( 'R' \), Fig. 3.
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 tangent meter to give a direct display of measured impedance. The tangent 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 I\(Z\)I 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 effects 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 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 is a summary of Method 2, as detailed in an earlier article\(^2\). Combined impedance of the test capacitor and sense resistor is represented by \(|Z|_{local}\):

\[
|Z|_{local} = R_{sense} \times \frac{VM(1)}{VM(2)}
\]

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

\[
ESR = (\cos phase-angle \times |Z|_{local}) - R_{sense}
\]

where phase-angle is \(VP(1) - VP(2)\).

If phase-angle is negative, capacitance is,

\[
C = \frac{1}{2 \pi f X_c}
\]

If phase-angle is positive, inductance is,

\[
L = \frac{1}{2 \pi f X_c}
\]

Capacitor impedance is given by,

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

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_{15} and U_{16} 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_{15} 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_{3} and R_{53}. The IZI 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 IZI meter ground to the PM128 input ground.

To accommodate the maximum voltage from the IZI meter, R_{6}, originally 470kΩ, 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 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 IZI 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

---

**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 inductive 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 represent ESR (resistance) and reactance.

**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 300-350kHz from a signal generator at the jig input and with the IZI 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:

\[
\frac{V(1) - V(2)}{V_M(2)}
\]

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.
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 IZI 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.

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

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 U5.

Capacitance and inductance values
These values are calculated from measured R±jX for typical 220pF/10V capacitors. One random sample only of each style was measured.

<table>
<thead>
<tr>
<th>Unit/parameter</th>
<th>10 kHz</th>
<th>30 kHz</th>
<th>100 kHz</th>
<th>300 kHz</th>
<th>1 MHz</th>
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<tbody>
<tr>
<td>220μF/10V Philips 037</td>
<td>Aluminum</td>
<td></td>
<td></td>
<td></td>
<td></td>
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<tr>
<td>Impedance (Ω)</td>
<td>0.824</td>
<td>0.785</td>
<td>0.738</td>
<td>0.706</td>
<td>0.671</td>
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<tr>
<td>ESR (Ω)</td>
<td>0.818</td>
<td>0.783</td>
<td>0.74</td>
<td>0.71</td>
<td>0.67</td>
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<td>Capacitance</td>
<td>129μF</td>
<td>68.4μF</td>
<td>27.1μF</td>
<td>9.1μF</td>
<td>2.0μF</td>
</tr>
<tr>
<td>220μF/10V Rubycon YXF</td>
<td>Aluminum</td>
<td></td>
<td></td>
<td></td>
<td></td>
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<tr>
<td>Impedance (Ω)</td>
<td>0.404</td>
<td>0.372</td>
<td>0.341</td>
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<td>0.306</td>
</tr>
<tr>
<td>ESR (Ω)</td>
<td>0.394</td>
<td>0.372</td>
<td>0.34</td>
<td>0.32</td>
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<td>Capacitance</td>
<td>153μF</td>
<td>94μF</td>
<td>44μF</td>
<td>20.7μF</td>
<td>9.2μF</td>
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<tr>
<td>220μF/10V Rubycon ZL</td>
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<td></td>
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<td></td>
</tr>
<tr>
<td>Impedance (Ω)</td>
<td>0.122</td>
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<td>0.082</td>
<td>0.076</td>
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<tr>
<td>ESR (Ω)</td>
<td>0.10</td>
<td>0.09</td>
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<td>0.08</td>
<td>0.08</td>
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<td>Capacitance</td>
<td>187.5μF</td>
<td>69.5μF</td>
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<td>104.5μF</td>
<td>1.21nH</td>
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<tr>
<td>220μF/10V Rubycon ZA</td>
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<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Impedance (Ω)</td>
<td>0.087</td>
<td>0.041</td>
<td>0.030</td>
<td>0.028</td>
<td>0.035</td>
</tr>
<tr>
<td>ESR (Ω)</td>
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<td>0.04</td>
<td>0.03</td>
<td>0.03</td>
<td>0.03</td>
</tr>
<tr>
<td>Capacitance</td>
<td>189.8μF</td>
<td>185.4μF</td>
<td>200.9μF</td>
<td>0.69nH</td>
<td>2.39nH</td>
</tr>
<tr>
<td>220μF/10V Elna RSH</td>
<td>Aluminum</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Impedance (Ω)</td>
<td>0.313</td>
<td>0.290</td>
<td>0.270</td>
<td>0.254</td>
<td>0.242</td>
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<tr>
<td>ESR (Ω)</td>
<td>0.31</td>
<td>0.29</td>
<td>0.27</td>
<td>0.26</td>
<td>0.25</td>
</tr>
<tr>
<td>Capacitance</td>
<td>171.7μF</td>
<td>119.7μF</td>
<td>55.0μF</td>
<td>27.6μF</td>
<td>19.33μF</td>
</tr>
<tr>
<td>220μF/10V AVX TPS</td>
<td>Tantalum</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Impedance (Ω)</td>
<td>0.143</td>
<td>0.089</td>
<td>0.06</td>
<td>0.045</td>
<td>0.040</td>
</tr>
<tr>
<td>ESR (Ω)</td>
<td>0.10</td>
<td>0.08</td>
<td>0.06</td>
<td>0.05</td>
<td>0.05</td>
</tr>
<tr>
<td>Capacitance</td>
<td>149.2μF</td>
<td>100.9μF</td>
<td>62.3μF</td>
<td>45.7μF</td>
<td>0.83nH</td>
</tr>
<tr>
<td>220μF/10V Sanyo Oscon Oscon</td>
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<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Impedance (Ω)</td>
<td>0.077</td>
<td>0.027</td>
<td>0.01</td>
<td>0.01</td>
<td>0.029</td>
</tr>
<tr>
<td>ESR (Ω)</td>
<td>0.02</td>
<td>0.02</td>
<td>0.01</td>
<td>0.01</td>
<td>0.01</td>
</tr>
<tr>
<td>Capacitance</td>
<td>201.3μF</td>
<td>192.0μF</td>
<td>230.9μF</td>
<td>2.13nH</td>
<td>4.23nH</td>
</tr>
</tbody>
</table>

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.

Three-component modelling
As shown in Fig. 3, a capacitor is usually represented as a series combination of capacitance 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.

The software calculates parameters at frequencies where the measured impedance is a factor of \( \sqrt{2} \) smaller and larger than the maximum and minimum values measured.

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 plan 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 163.7uF. This represents excellent performance and is a much more realistic value than the device's apparent capacitance, which increased to 230.9uF, calculated from an uncorrected, single-frequency measurement.

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

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 1.52Ω 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 1Ω 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.
The definitive biography of the century’s godfather of invention—from the pre-eminent Edison scholar “Israel’s meticulous research and refusal to shy away from the dodgier aspects of Edison’s personality offers a fresh glimpse into the life of the inventor.” – New Scientist

“Remarkable.” – Nature

“An authoritative look into Edison’s working methods, here leavened by enough personal detail to give the achievements shape.” – Publishers Weekly

“Israel’s book should go a long way toward taking Edison out of the shadows and placing him in the proper light.” – Atlanta Journal-Constitution

“Exhaustively researched, with strong emphasis on Edison’s methods and achievements.” – Kirkus Reviews

The conventional story of Thomas Edison reads more like myth than history: With only three months of formal education, a hardworking young man overcomes the odds and becomes one of the greatest inventors in history. But the portrait that emerges from Edison: A Life of Invention reveals a man of genius and astonishing foresight whose career was actually a product of his fast-changing era. In this peerless biography, Paul Israel exposes for the first time the man behind the inventions, expertly situating his subject within a thoroughly realized portrait of a burgeoning country on the brink of massive change. Informed by Israel’s unprecedented access to workshop diaries, notebooks, letters, and more than five million pages of archives, this definitive biography brings fresh insights to a singularly influential and triumphant career in science.
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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

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 0.9xVcc. 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 1/6 and 2/3 Vcc.

Timer IC2A forms an astable multivibrator generating an exponential reference signal. This signal is compared in IC2B with the control voltage created with P1 (R3), forming a PWM signal.

Output is taken from the discharge pin of IC2B. This pin drives the voltage-regulator IC3 to create an output voltage that is regulated and has short circuit protection.

Amplitude of the output voltage is set with P2 (R6) between 3V and 15V. Pin 7 short circuits the resistance R5 when the control voltage is lower than the reference signal.

The reference voltage for P1 is taken from the voltage-control pin of IC2A, 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 IC2A is connected to the reset and trigger pin disabling the output of IC2B while capacitor C4 is discharging.

In the CMOS version of the NE556, the internal reference voltage chain uses higher value resistors, and it is necessary to scale R3 accordingly.

Bernard Van den Abeele
Evergem
Belgium
E24
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 \( T_{r1} \) (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 \( T_{r1} \) 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 \( 1.8/(R_3\times C_2\times V_{cc}) \). With the values shown in the diagram the charging current is about 50uA and the oscillator frequency is about 3kHz with supply of 5V.

M Hughes
York

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

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 5V 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.

<table>
<thead>
<tr>
<th>Power supply nominal o/p (V)</th>
<th>Trip voltage with germanium D2 (V)</th>
<th>Trip voltage with silicon D2 (V)</th>
</tr>
</thead>
<tbody>
<tr>
<td>5.0</td>
<td>5.9</td>
<td>7.2</td>
</tr>
<tr>
<td>10.0</td>
<td>11.0</td>
<td>12.3</td>
</tr>
<tr>
<td>15.0</td>
<td>15.9</td>
<td>17.3</td>
</tr>
<tr>
<td>20.0</td>
<td>21.0</td>
<td>22.3</td>
</tr>
<tr>
<td>25.0</td>
<td>25.8</td>
<td>27.2</td>
</tr>
<tr>
<td>30.0</td>
<td>30.8</td>
<td>32.3</td>
</tr>
</tbody>
</table>

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

Measure fuse blowing time at a given overcurrent level.
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 '13' 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 \( U_{1B} \) changes state and sets \( U_{2A} \). Transistor \( Q_3 \) conducts and trip circuit \( Q_1 \) is disabled. Counters \( U_3 \) is now forced into reset with all its outputs, \( Q_1-8; \) low and no voltage appears at the R-2R ladder network. Green LED 'D' 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. Bistable device \( U_{2B} \) changes state and sets \( U_{2A} \). Counter \( U_3 \) is now forced into reset with all its outputs, \( Q_1-8; \) low and no voltage appears at the R-2R ladder network. Green LED 'D' turns off to signal protection and the circuit is now off.

Crowbar protection circuit for a laboratory bench supply.
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- UL Listed
- 5½ Digit Multimeter for PCMCIA

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

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info.uk@ni.com www.ni.com.
Simple relay tester

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 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.

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36 ELECTRONICS WORLD January 2001
The HS801: the first 100 Mega samples per second measuring instrument that consists of a MOST (Multimeter, Oscilloscope, Spectrum analyzer and Transient recorder) and an AWG (abritary waveform generator). This new MOST portable and compact measuring instrument can solve almost every measurement problem. With the integrated AWG you can generate every signal you want.

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

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

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

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

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

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

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Track
2. Good Old Summertime, The American Quartet 1904
3. Marriage Bells, Bells & xylophone duet, Burckhardt & Daab with orchestra, 1913
4. The Volunteer Organist, Peter Dawson, 1913
5. Dialogue For Three, Flute, Oboe and Clarinet, 1913
6. The Toymaker’s Dream, Foxtrot, vocal, B.A. Rolfe and his orchestra, 1929
7. As I Sat Upon My Dear Old Mother’s Knee, Will Oakland, 1913
8. Light As A Feather, Bells solo, Charles Daab with orchestra, 1912
9. On Her Pic-Pic-Piccolo, Billy Williams, 1913
10. Polka Des English’s, Artist unknown, 1900
11. Somebody’s Coming To My House, Walter Van Brunt, 1913
12. Bonny Scotland Medley, Xylophone solo, Charles Daab with orchestra, 1914
13. Doin’ the Raccoon, Billy Murray, 1929
14. Luce Mial 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

Total playing time 72.09

All tracks on this CD were recorded on DAT from cylinders produced in the early 1900s. Considering the age of the cylinders, and the recording techniques available at the time, these tracks are of remarkable quality, having been carefully replayed using modern electronic technology by historian Joe Pengelly.

21 tracks – 72 minutes of music.
Published by Electronics World. All recordings reproduced by Joe Pengelly.
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.

**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 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.

**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 LED indicator light is available in red, yellow, green, blue or white.

**20V MOSFETs**
Fairchild has introduced 20V p-channel MOSFETs with 8 and 12V VGS ratings in surface mount packages. The devices have maximum RDS(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 VDS, 12V VGS MOSFET has a maximum RDS(ON) of 11mΩ at 4.5V VGS. The FDW254P single 20V MOSFET in a TSSOP-8 at 2.5V VGS has an RDS(ON) of 15mΩ.

**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.
Voltage reference

National Semiconductor has announced a sub-bandgap series voltage reference. With a voltage reference of 1.024V, the LM4140, it 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 1ppin/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 simple 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, 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 radiating 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 = dv/dr = d( rH)/dr = v = rH.

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 = GM/v2, where we predict G = 3H/4mp. 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 furthest 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 ASYMmetrics 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 = ma, where a = Hubble acceleration (because the dielectric motion in balances the Hubble acceleration). Thus R = dv/dt = d(RH)/dt = vH = RH2, 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 = 4πr2. Now, n2 = effective mass of the dielectric of free space (the 377 ohm dielectric or continuous fabric of space) acting on the area A; hence n2 = mass of matter in hollow volume of the universe with an inner radius of r and an outer radius of R, so n2 = (4/3)x 123p (R2 - r2)p, where r2 is insignificant compared to R2, so the formula reduces to simply n2 = (4/3)p R2p, 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 various objects, 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 = ma/A = (1/3)Rx H2p / r. 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 acceleration towards its target, represents the shielding: hence, net pressure, P = P(A,n2g, / (4E r2) = RH2 M / (471r2).

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.

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 recall). A shielding mass M, will therefore stop dielectric pressure in proportion to the ratio of M to the mass of the universe, (4/3)p R2. The effective shielding cross-sectional area is therefore Aeffective = [surface area of a sphere of radius r] xM[(4/3)p R2p] = 3M r2 / (R3p). The net pressure towards mass M is the unshielded pressure multiplied by Aeffective divided by the area of the sphere (4r2) of radius r which mass M is shielding; hence, net pressure, P = P(An2g, / (4E r2) = RH2 M / (471r2).

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 = PA/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] x(surface area of sphere with radius r) / [surface area of sphere with radius of universe] = (4/3)p R2p, and A is the surface area of a sphere of radius r, so A = 4πr2. Hence, a = (3/4) H2 M/(πr). Hence, we have theoretically derived the Newtonian expression a = GM/r2, where we find G = (3/4) H2 / (πp). This is a testable prediction, as H and p (including the invisible matter like black holes and neutrons) 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

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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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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, +2 or -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

**400/200MHz 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.

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...
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National says it expects the SCANSTA 400 chip to accelerate acceptance of the IEEE1149.4 mixed-signal test standard. Sample quantities will be available in Q4/00 from National as a technology test vehicle.
National Semiconductor
Tel: 0870 240 2171

Real-time emulator
A non-intrusive, real-time emulation hardware module from RF Solutions is for developing embedded systems based on Arizona Microchip's PIC16F87x eight-bit microcontrollers with on-board functionality including flash memory and 10-bit a-to-d conversion. The ICEPIC DB877 personality daughter board, with the company's ICEPIC and ICEPIC 2 non-intrusive in-circuit emulators, provides real-time source level debugging in assembler or C at maximum processor speeds for all 16F87xs. The user can set unlimited hardware trigger breakpoints on any address or range of addresses. Developers can identify and rectify software bugs by executing code in single-step and procedure-step modes. A hardware filter controls multi-cycle instruction capture. It runs under Windows 95, 98, 2000 and NT.
RF Solutions
Tel: 01273 488880

Low-Power CMOS VLSI Circuit Design
A comprehensive look at the rapidly growing field of low-power VLSI design

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Valve Radio and Audio Repair Handbook

A practical manual for collectors, owners, dealers and service engineers. Essential information for all radio and audio enthusiasts. Valve technology is a hot topic.

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Zero-drift op amp

Linear Technology has announced the LTC2051 and LTC2052 dual and quad zero-drift amplifiers. Housed in the tiny MSOP-8 and SSOP-16 packages, these DC accurate op amps feature a maximum input offset voltage of 3μV and an offset drift of 30nV/°C. 75μA input bias current, CMRR and PSRR in excess of 130dB and 140dB large signal gain. In addition to DC accuracy and a small footprint, the LTC2051/LTC2052 feature a wide gain bandwidth of 3MHz, a high slew rate of 2V/μs and low DC to 10Hz noise of 1.5pVpp. Optimised for use in portable battery-powered as well as industrial applications, these amplifiers are capable of operation from a single +2.7V supply to ±5V, drawing just 75μA per channel. Rail-to-rail output voltage swing and a wide input common mode range that includes ground make these op-amps ideal for use with a range of sensors from thermocouples to airflow meters, pressure sensors and any variety of wheatstone bridge sensors.

Linear Technology
Tel: 01276 677676

LEDs that emulate lamps

Litton VEAM/TEC’s range of fascia panels use a combination of white LED technology, optical filters and custom dimming circuitry to emulate the appearance of conventional incandescent lamps. Primarily designed for retrofit applications in defence and aerospace instrumentation, the new panels have the same colour and dimming characteristics as surrounding incandescent-based displays while still possessing the features of surface mount LED technology, including lower power consumption, lower operating temperature, longer lifetime and increased reliability.

Litton VEAM/TEC
Tel: 0208 8366 1291

Li-ion charger IC

Texas Instruments claims to have the industry’s first Li-ion battery charger management IC to have an integrated MOSFET and Schottky diode. The bq2404x family includes a 1.2A MOSFET pass transistor and a reverse-current blocking Schottky diode with a combined maximum dropout voltage of 0.7V. According to the supplier, this should make it possible to build a full battery charger circuit within a 44mm² footprint. In addition, the devices perform a programmable charge algorithm, including pre-conditioning for deeply discharged cells and voltage, temperature and time monitoring for safety. Battery makers recommend pre-conditioning for deeply discharged Li-ion and Li-Polymer cells. The process applies a reduced current to raise the battery voltage above 3V. The devices also include a safety timer so that if pre-conditioning fails to produce the desired result, the battery is presumed damaged and charging stops.

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

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 bandpass 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.

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 affects 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,

\[ F = 1 + \frac{(L-1)T}{290} \]  

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,

\[ F = F_1 + F_2 - \frac{1}{G_1} + \frac{1}{G_2} + \ldots + \frac{1}{G_N} \]  

Here, \( F \) is the equivalent noise factor, \( F_1, F_2, \ldots, F_N \) are the noise factors of stages 1, 2, and 3, \( G_N \) is the gain of the nth stage and \( G_1, G_2 \) 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,

\[ F = mF_R + nF_{LO} \]  

Here, \( F_R \) 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,
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,

\[ \text{Duplex image} = F_T - \Delta f \]

\[ \text{Half - duplex image} = F_T + \frac{\Delta f}{2} \]

Here, \( F_T \) is the transmitter frequency and \( \Delta f \) is the difference between the transmitter and receiver 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 \( (2FRF \pm 2Fw) \). The 1/2-IF rejection is given by,

\[ \text{IP}_{1/2} = S - C \]

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

\[ \frac{45\text{dBm} + 120\text{dBm} - 6\text{dBm}}{2} = 79.5\text{dB} \]

**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,

\[ \text{IM} = \frac{2IP_3 - 2S - C}{3} \]

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 \( 2F_1-F_i \) 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 nth order**

Once you know the input levels of signals applied to the receiver, you can calculate the nth order intercept points using,
Here, $IP_n$ is the $n$th order intercept point, $n$ is the order of the intercept point, $P_R$ is receiver input signal power level and $P_{IM2}$ 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.

- **First**, 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.

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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January 2001 ELECTRONICS WORLD
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.

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 ±Vp about 0V ground. Imagine a window comparator set at a level more negative than Vp, 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 +Vp, the probability becomes 0.5, assuming that the mark:space ratio is indeed 50:50. Beyond +Vp, clearly the probability of the voltage lying within the window is zero again. The probability is 0.5 while the level Vp remains within the window, regardless of the window’s width. In other words, the area of the shaded rectangle at Vp represents a probability of 0.5.

Now consider an infinitely narrow window, moving up from A = = = A, and encountering the -Vp edge of the squarewave. The long thin rectangle at -Vp, 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 -Vp and one at +Vp, 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 +Vp or -Vp. The probability of occurrence of any voltage between +Vp and -Vp or indeed outside that range, is zero.
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 IC3 integrates a voltage of \(+15\text{V}\) or \(-15\text{V}\), applied via \(R_5\) from the output of IC1d.

When the positive-going output of IC1c reaches \(+7.5\text{V}\), the output of IC1c falls from \(+15\text{V}\) to \(-15\text{V}\), while the output of IC1d rises from \(-15\text{V}\) to \(+15\text{V}\). The output of IC3 therefore reverses direction and moves linearly towards \(-7.5\text{V}\).

On reaching \(-7.5\text{V}\), the cycle repeats, and a \(\pm 7.5\text{V}\) triangular wave of about 0.5Hz is applied to the reference input, pin 2 of comparator IC2.

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 \(-15\text{V}\) or \(+15\text{V}\). Filter \(R_6/C_2\) 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 IC1a, and applied as Y deflection to a scope used in the XY mode. The triangular wave output of IC3 was applied as the X deflection.

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

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.

For this test, capacitor C2 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 C1b was ignored, and the filter output voltage across C2 monitored directly with a 10MΩ, x10 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 3½-digit, or better still 4½-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 a 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 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_c + f_m) where f_m is the deviation - a few megahertz upwards in this case - and f_c 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.

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. hfe 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.

Capacitance meter

There was an error in November’s Beginners’ Corner. On page 907, in the formula for C_p, 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.
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.

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.

---

**How does it radiate?**

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

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 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.
Diaphragm is considered as an infinite number of point radiators.

Effect here is obtained by integrating sound due to every point radiator.

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.
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 quali-
ty of an amplifier. The goal here was to pro-
duce 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 compo-
nent count.
We have found that single-ended designs
are affected by speaker choice so we pro-
duced 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 cho-

An ECC85 is used for its low impedance
in combination with a relative high ampli-
fication 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 cath-
ode voltage has to be something like +85V.
This is derived from,
\[ V_{\text{cath}} = V_{\text{ps}} - (V_{\text{plate}} + V_{\text{transformer}}) \]
Now that the cathode voltage of the EL84
is known, it is easy to determine the plate
voltage of the ECC85. As the cathode volt-
age 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 oper-
ating point.
The current source also makes the ampli-
fier more stable via DC feedback by means
of the two 120kΩ resistors. These resistors
apply some correction for current differ-
ences 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 cur-
rent source. We chose -6.2V for conve-
nience, 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 alter-
native, Fig. 1c), the DC feedback is again
superfluous.
Potentiometer P1 is 2200 and is added to
correct imbalance in the output stage. We
used a voltmeter made from a small 50µA-
0-50pA 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 cur-
rent feedback of the large cathode resistors.
Adjustment should be made 20 seconds
after power-up and again after 15 minutes

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

January 2001 ELECTRONICS WORLD
High voltage supply should have a delay of at least 30 seconds!

Gain approx 20 dB

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 EL34s 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 EL34s are fed to the screen grids of the EF86s of the long-tail triode with two EF86 pentodes and using the same pentode as the active device in the current source.
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,

\[
\frac{150}{100k} - \frac{15k}{180k}
\]

or about 3.6mA. This configuration should also work with ECC88s.

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.

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output power at 1% distortion</td>
<td>11.9W</td>
</tr>
<tr>
<td>Sensitivity for 10W output power</td>
<td>700mV</td>
</tr>
<tr>
<td>Noise/hum related to 1W</td>
<td>78dB</td>
</tr>
<tr>
<td>Frequency response @ 1W</td>
<td>10Hz - 72kHz</td>
</tr>
<tr>
<td>Distortion at 10W @ 1kHz</td>
<td>0.55%</td>
</tr>
<tr>
<td>Distortion at 10W @ 30kHz</td>
<td>2.35%</td>
</tr>
<tr>
<td>Distortion at 10W @ 45Hz</td>
<td>2.7%</td>
</tr>
<tr>
<td>Damping 4Ω @ 100Hz</td>
<td>4x</td>
</tr>
<tr>
<td>Damping 4Ω @ 10kHz</td>
<td>5.3x</td>
</tr>
</tbody>
</table>

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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.
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.
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.

---

**The irresistible force meets the immovable object**

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

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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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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, 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 \( \delta \) – 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-
The lettering in Fig. 4 denotes an ideal exponential diode, and the \( V_D \) characteristic. See also equation 1 in the main text.

\[
I_D = I_s \exp \left( \frac{V_D}{V_T} \right) - 1
\]

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,

\[
I_D = I_s \exp \left( \frac{V_D}{V_T} \right)
\]

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 \( I_s(\frac{V_D}{V_T}) \) but pass currents scaled in the ratio \( m:1 \), then,

\[
V_S = V_T \log_m m
\]

and,

\[
\frac{dV_S}{dT} = \frac{k}{q} \log_m m = 86.2 \log_m m \mu V/C
\]

The parameter \( m \) is a design choice.

**Ideal piecewise-linear BJT models**

An \( n-p-n \) BJT is 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 $\beta$ 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 $V_t$, of around 0.7V, in series with $DE$, 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 $DE$ 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 $V_{BE}$ is less than 0, i.e., $\beta_N$ is what was previously called $\beta$.

Diode $DC$ models the base-collector junction. Similar comments apply to the (reverse, $R$) current components,

$$I_{RR} = \frac{I_{CC}}{\beta_R}$$

and $I_{CR}$ associated with it. Normally, $\beta_R$ is less than $\beta_N$ 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 $R_R$, Fig. 11 shows a load line for an instantaneous value of $I_G$, in this case the maximum value $V_M$, plotted on the output characteristic for,

$$I_s = \frac{V_C}{R_s}$$

Switch $S_1$ in Fig. 9 models the action of a gating waveform. When this switch is at $X$, $Q$ is on and $V_0$ is zero, provided that,

$$(\beta_s + 1) \frac{V_C}{R_s} > \frac{V_B}{V_{BE}}$$

When the switch is at $Y$, $Q$ is off provided $V_M$ is less than $V_{KK}$. Then,

$$\frac{V_0}{V_G} = \frac{R_s}{V_{BE} + R_R}$$

Ideal exponential BJT models

Figure 12 illustrates an ideal exponential BJT model when $V_{BE}$ is less than zero. If $V_{BE}$ is more than 100mV,

$$I_c = I_s \exp \frac{V_{BE}}{V_T}$$

$I_s$ is a property of the base region. The change in base width with voltage is allowed

![Fig. 9. An analogue gate circuit.](image)

![Fig. 10. Input and output waveforms, $V_G$ and $V_0$ respectively, for Fig 9.](image)

![Fig. 11. Showing worst load-line position for satisfactory operation of the gate circuit.](image)

![Fig. 12. Model, a), and transfer and input characteristics of an ideal logarithmic BJT, b).](image)
for by multiplying $I_3$ by a factor,

$$I = \frac{V_{CE}}{V_A}$$

where $V_A$ is the magnitude of the Early voltage.

Equation 7, which emphasizes 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 $\beta$ is infinity, it can be shown that,

$$I_0 = \sqrt{I_1^2 + I_2^2}$$

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 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 $V_{BE}$ is shown in Fig. 15. It is a transport version of the classic Ebers-Moll injection model. For $V_{BE}, V_{BC} >> V_T$,

$$I_C = \frac{I_{CN} V_{BE}}{V_T}$$

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 of the BJT in saturation that is useful in interface-circuit design.

References
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Letters to the editor

Letters to ‘Electronics World’ Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS e-mail jackie.lowe@bli.co.uk using subject heading ‘Letters’.

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 contribution 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 ‘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 contrast 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 2X is 6.8 and 3.9 and so on. Now, of course, one might take 2X 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 Dennis is dubious to the point of opining 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 should have had a sine-wave 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 waveform 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, delay 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.

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 signals 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 the above 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?

I would like to build a homing device for a radio-controlled boat, where the boat would automatically ‘return to base’ by default. Can anyone help me find such a kit, or perhaps point me towards another solution to the problem, as designing and building such a system alone is a bit beyond me?

Paul Hampshire
Specialist Instrument Mechanic
Sasol (Secunda)
South Africa

January 2001 ELECTRONICS WORLD
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 Ericsson, published in IEE paper 2747 July 1966. And, 'Measurements of non-linearity in cracked carbon resistors', by G.H. Millward of the BBC, published in Electronics Northern Ireland in 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 Ericsson.

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. (CL11 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 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/hotone 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.
'Distortions inherent in PWM Class-D amplifiers...,' Dec. 1968, p.672.

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- Uses USB/Recharger supplied, or batteries for real portability
- Includes Chip testers for TECOMOS, DRAM and SRAM devices
- Optional EPROM/RAM emulators also available

<table>
<thead>
<tr>
<th>Model</th>
<th>Support</th>
<th>Price</th>
</tr>
</thead>
<tbody>
<tr>
<td>EPMaster LV48</td>
<td>48-pin support for EPROM/Flash, Pin &amp; Serial PROM</td>
<td>£995</td>
</tr>
<tr>
<td>Speedmaster LV48</td>
<td>As EPMaster LV48, plus SOP, PSOP, SSOP, PLCC and over 300 microcontrollers</td>
<td>£495</td>
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<tr>
<td>MicroMaster LV48</td>
<td>As Speedmaster LV48, plus over 300 microcontrollers including 87C451/51/61, PIC, ICS, 93CXX, 51XX, 68HC05/12, SAB-C000, TM83205/370, 28XX, COP etc...</td>
<td>£695</td>
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<tr>
<td>LVUSB Portable</td>
<td>40-pin version of MicroMaster LV48 + LCD &amp; Keypad</td>
<td>£995</td>
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