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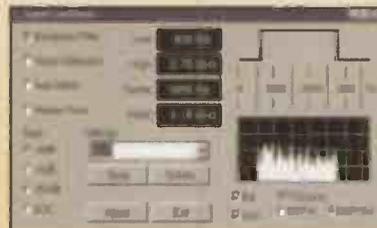
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IF bandwidths	6 kHz (AM/SSB), 17 kHz (FM-N), 230 kHz (W)	2.5 kHz(SSB/CW), 9 kHz (AM) 17 kHz (FM-N), 230 kHz (W)	2.5 kHz(SSB/CW), 9 kHz (AM) 17 kHz (FM-N), 230 kHz (W)
Receiver type	PLL-based triple-conv. superhet		
Scanning speed	10 ch/sec (AM), 50 ch/sec (FM)		
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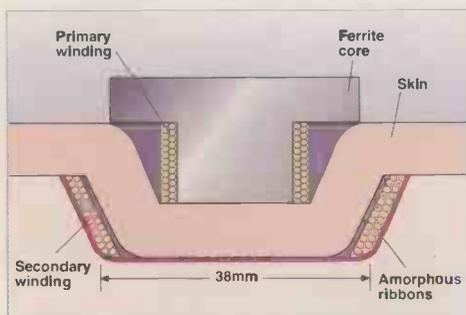
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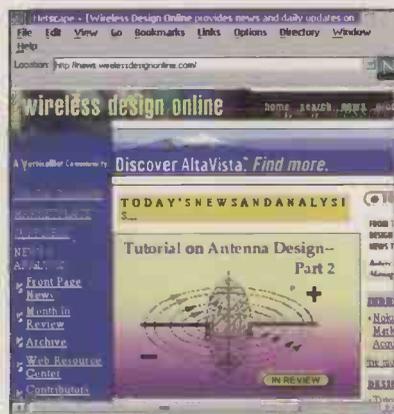
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Cover illustration: Jamel Akib



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INFORMATION

Cache 22

The European Parliament has rapidly become the alternative comedian's equivalent of 'My mother-in-law' as a sure-fire way of setting-up an audience for a good belly-laugh at the presumptions and follies of politicians – and of course largely foreign politicians at that.

You can hear the sniggers breaking out all over the room almost before the words are properly out. What is it this time, they're wondering. Another straight banana edict?

This time it's the vote to ban Internet service providers from cacheing. Not quite as rib-tickling as the straight banana, but just as absurd. And, for service providers and all of us who have come to rely on the Internet, just as serious.

Service providers routinely set up caches – simply temporary files – of websites in order to make it easier for subscribers to access them at busy times. They enable subscribers to access the content locally, speeding access and reducing traffic – and hence congestion problems – on the Internet.

Cacheing is pretty much like distributing films to local multiscreens, rather than making everyone drive halfway round the M25 to watch them, but with one important difference. The process does not increase revenue for the ISPs. Indeed, where access is metered, it would actually mean less revenue.

Nor does it deprive rights owners of revenue – a vital point, given the context of the decision. The vote was taken as part of the first reading of a new draft directive on electronic copyright, and its specific intention was to prevent the theft of recorded music by illicitly downloading it off the Internet.

It's hard to see how such a measure would make it harder to steal pirated CDs over the net, except insofar as it would make it harder for everyone to access anything at all.

The inspiration for the move appears to be persistent lobbying by the record and film industry itself, which has waged a long-running campaign against 'theft' of electronically transmitted material – a crime which it defines to include time-shifting of tv programmes.

The same draft directive would ban all forms of private copying of tv and radio programmes, effectively outlawing the domestic video recorder.

The industry does indeed have a serious problem

with criminal interests making illegal and unauthorised copies of discs or tapes on a large scale for profit. It is a problem which could admittedly increase with the Internet, and which needs addressing.

However, most of the measures it has sought to introduce over the years have not been to address this problem – which is covered by criminal law in any case. Rather they have been implemented to increase its profits by taxing what to most reasonable people looks like an aspect of the legitimate enjoyment of the material in question, which has been bought and paid-for. Time-shifting a tv show is one example, taping a CD for playing in the car is another.

It has succeeded to the extent that some European states now impose a levy on blank tapes and recording equipment.

That the British government has so far robustly resisted such moves is probably a tribute to the visibility of our democratic processes and our ability to bring the cause of public interest and common sense to bear against the pressures of vested interest groups and lobbyists. That the

European Parliament has yielded to those pressures demonstrates the exact opposite.

One of the problems with the European Parliament is its remoteness from each of the national communities it represents. If our own MPs at Westminster get a silly or inequitable idea into their heads, we soon get to hear of it, and bring our displeasure, anger and ridicule, if needed, to bear on them. Particularly if we think they've been stitched up by some special interest group.

In Brussels, or Strasbourg, it's different. We don't know what they're up to half the time, but the lobbyists sure as hell do.

The result is that an opportunity to align copyright legislation with the technology of the 21st century has been hijacked and distorted by this interest group.

Curiously, at much the same time, another EU report, prepared by British MEP Christine Oddy for a draft law on e-commerce, was published, arguing that caching was essential if the Internet was to run smoothly.

What a tangled Web may be woven if we leave it to the MEPs.

Peter Willis



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CIRCLE NO. 107 ON REPLY CARD

UPDATE

New mobile phones hampered

UK third generation mobile phone operators will be handicapped by the inadequate spectrum allocation being proposed by the government for its new UMTS licences.

That is the claim of the UMTS Forum, a multinational group of manufacturers and mobile phone operators promoting the UMTS third generation standard.

The UK government is considering two spectrum proposals where either three or four of the five UK third generation operators get a 2x10MHz band pair and single 5MHz band.

"This is fine for just voice but it is not enough for multimedia services," said Dr Chris Wildey, a UMTS Forum vice chairman. "The minimum for UMTS just to get started is 2x15MHz plus 5MHz."

If not enough spectrum is made available, said Wildey, it will adversely affect the services operators will be able to offer.

The Radiocommunications Agency (RA), charged with spectrum

ALLOCATION AT A GLANCE

First option

3 licensees with 2 x 10MHz pair and 1 x 5MHz single
2 licensees with 2 x 15MHz pair

Second option

4 licensees with 2 x 10MHz pair and 1 x 5MHz single
1 licensee with 2 x 20MHz pair
Total band is 155MHz with 10 to 15MHz for licence-free or "lightly licensed" private operation.

allocation proposals in the UK, says the situation is still under review.

"Our proposals went out for comment about four weeks ago and we are just receiving the last of the responses from interested parties, one of which is the UMTS Forum," said Jeremy Clayton, head of the RA spectrum auction team.

This month has been allocated to review comments and make firm proposals to ministers. "We should be finding a way forward by April-

May time," said Clayton.

The spectrum claims were backed by UMTS Forum chairman, Dr Bernd Eylert, who is also at Deutsche Telekom's mobile subsidiary DeTeMobil: "Our 2x12MHz spectrum band is not enough just for GSM [services]."

The RA's Clayton said UMTS will use the available spectrum better than GSM.

Roy Rubenstein and Steve Bush
Electronics Weekly

A loudspeaker you can see the tv screen through?

NXT, the flat-panel loudspeaker company, has developed a thin, transparent speaker that can be fitted over a display and produce stereo sound.

In the first public demonstration, one was fitted to a slim-line notebook computer, although the technology is said not to be limited to small screens. "It could be applied to a 60 inch television," said company spokesman John Watson.

The company claims that the technology, which it is calling SoundVu, is ready for use. "If someone takes it up and runs with it, it could see production before the end of the year," said Watson.

For NXT, it is too early to talk about licensees because last week's demonstration was the first time anyone outside the company had experienced SoundVu in action. "It is an in-house development, we did not have a development partner," said Watson.

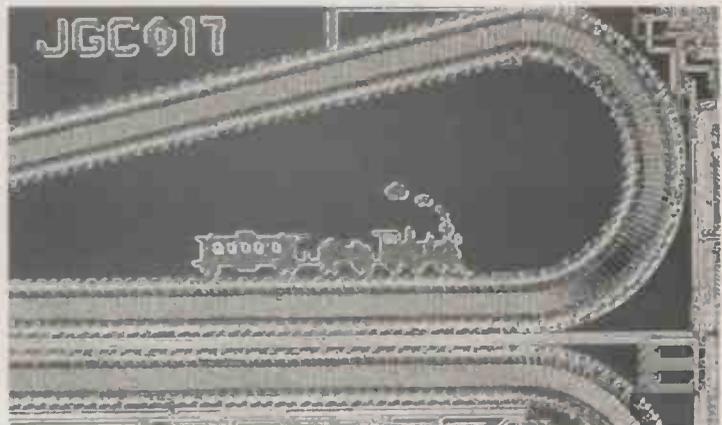
The display itself takes no part in the sound production, although Watson will not rule out using the display structure to make sound in the future. Instead the speaker sits in front of the display with an air gap. The speaker adds "a couple of millimetres, if that" to the display

thickness, said Watson.

Driving the speaker from the top or bottom allows stereo sound to be produced, although Watson thinks that it would also be possible by driving from one side only "but you would need a DSP".

Steve Bush

Track signalling... The microscopy department of the Florida State University has made a collection of the doodles chip designers have put onto the chip metallisation and has set up a Web Site called Silicon Zoo to show them to the world. The train found rolling on the tracks here is on a LeCroy MVV 200 chip. The tracks are a charge-coupled device analogue shift register



New Euro maths processor uses logs to speed up calculations

A European project co-ordinated by Newcastle upon Tyne University aims to produce an arithmetic unit that is more accurate and twice as fast as existing designs.

The concept is stunningly simple. As every school child knows, multiplying and dividing numbers can be easily achieved by adding or subtracting their logarithms.

Expressing a 32-bit floating-point value as its logarithm turns it into a 32-bit fixed point number and makes the maths easy. It reduces the computation time by a factor of five for multiplies and 15 for divides.

So why aren't all arithmetic units, microprocessors and DSPs organised this way? Unfortunately, while logarithms make multiplying and

What about accuracy?

When arithmetic is performed on floating point numbers, each operation, such as the common multiply accumulate, can cause a half bit rounding error. Over the course of a long computation, a Fourier transform for instance, these errors can reach several bits.

Fixed-point numbers on the other hand do not suffer from such errors. And, hey presto, the log of a 32-bit floating point number is fixed point. Newcastle claims the improvements in accuracy can be seen during simulations when 3D graphics data is being computed.

Why Bother?

What's so great about a mere doubling of arithmetic speed? The inherent growth rate of the industry says performance will double within 18 months anyway.

Well, looking at it from another angle, for a specific application the clock speed can be halved. This has a corresponding effect on the power consumption, and improves electromagnetic compatibility to boot.

Offer any mobile phone manufacturer a painless way of doubling talk time and they'll snap your hand off. At the shoulder.

dividing a breeze, addition and subtraction are very tricky when the original numbers are expressed as logs.

The obvious answer is to do adds and subtracts in floating point and multiply and divides in logs. Indeed, if the data could be converted to and from logarithms on the fly, as and when needed, the problem would be solved.

But this would take far too much time. Therefore when data for the processor comes into the system, from an a-to-d converter for example, it has to be stored as a logarithm, never as a floating point number. And so additions and subtractions must be done in logarithms.

The sum of two numbers, x and y , when they are expressed in logs – $i=\log(x)$ and $j=\log(y)$ – is $\log(2j+2i)$, which can't be calculated on a simple processor.

The project team has solved this problem by recourse to some clever maths techniques.

$\log(2j+2i)$ can be expanded to equal $i+\log(1+2j-i)$. This is a non-linear function which can be evaluated by storing the solutions in a look up table.

However, for 32-bit numbers, the table would be enormous, so a small one is used with interpolation using the Taylor approximation. This adds errors, so the team has developed a brand new algorithm that runs in parallel with the look up table and calculates the error in the Taylor series which is subtracted out.

This process of addition or subtraction while in logs takes the same amount of time as it would in floating point. In typical algorithms, the acceleration of the multiplies and divides results in an overall doubling of speed.

Novel power supply gets to the heart

Increasingly, electrical devices are being implanted into people. Some of the more ambitious in-body systems, including artificial hearts, are too power-hungry to be

powered from batteries and have to be supplied from outside the body.

One option for getting the power in is to run a cable right through the skin. This is possible, but infection is a constant threat.

Transformers, with one coil outside the body and one coil inside, avoid the infection problem. Unfortunately, coupling efficiency is compromised because the skin is thick, between 5 and 15mm.

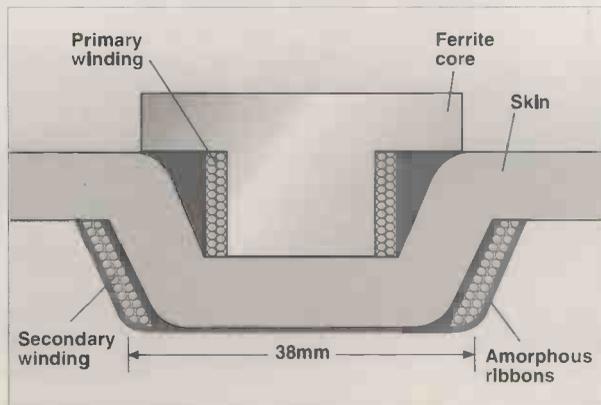
In a paper submitted to the IEE's *Electronic Letters* (volume 35,

number 2), a team from the Nanyang Technological University in Singapore has described a transcutaneous transformer which can transmit 20W of power at 90 per cent efficiency.

The transformer employs an unusual geometry (see diagram). The primary core is ferrite around which a 21 turn coil is wound. The secondary core is made from amorphous metal ribbon and has 27 turns.

Litz wire, made from many thin insulated strands, is used for both windings to reduce losses due to the skin effect.

An external 100kHz oscillator provides power which is rectified inside the body. The coupling coefficient between the coils is 0.65 with a 5mm skin thickness.



Body talk... Research team's transcutaneous transformer can transmit 20W of power at 90 per cent efficiency.

Micromachining puts RF tuning and switching components on chip

Micromachining is coming to RF chips. The only important question remaining is when?

"It has been 20 years since a technology has come around that will give us such an improvement," said Professor Elliot Brown, a micromachine specialist at the University of California, Los Angeles.

Silicon semiconductors, so useful in digital and low-speed analogue circuits, are more limited in RF applications.

True, there are silicon processes that will work at 2GHz, but any tuning components or signal path switches have to be off-chip. And the tuning problem extends all the way down the RF spectrum.

Only semiconductors with high bulk impedances, such as expensive GaAs, can be used to make filters with any appreciable selectivity.

Making micro-electromechanical systems (MEMS) by micromachining looks set to side-step the switching and tuning problems by allowing the construction of low-loss relays, mechanically resonant filters, high-Q inductors and variable high-Q capacitors.

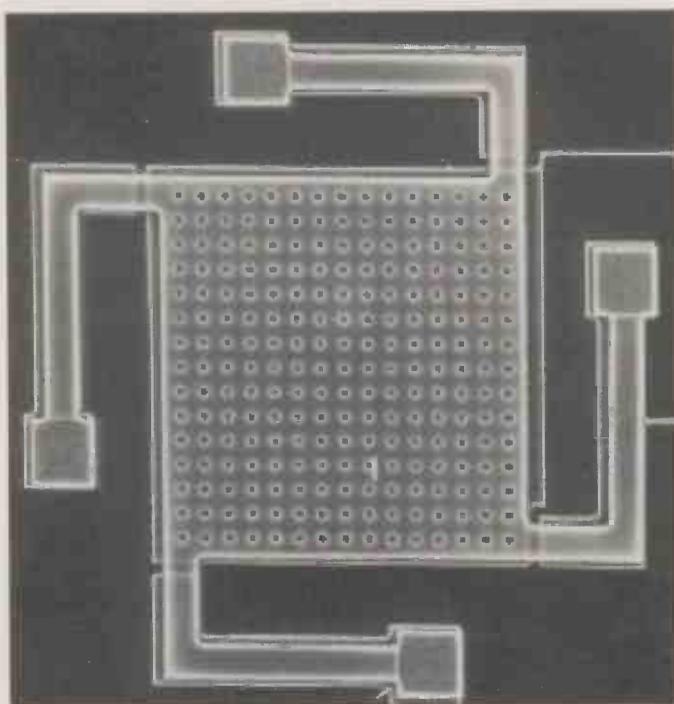
All of these have been constructed on-chip using modifications of standard semiconductor processes and are small enough to lay along-side other circuitry. The University of Michigan, for instance, has constructed a 92MHz bandpass filter with a Q of 8000 which is only 13.1µm long.

Soon, claimed UCLA's Brown, RF MEMS will be commonplace on chips. "You will start to think of them like inverters," he said.

First will come on-chip RF switches in the form of electrostatically deflected relays.

The standard solution to implementing a compact gigahertz-range RF switch is to use a PIN diode where more than 1dB switching loss is inevitable. A 3-6GHz PIN diode 8-by-8 crossbar switch has an insertion loss of around 21dB. Individual MEMS switches have already been constructed with losses of under 0.1dB and isolation over 35dB.

And they are fast, bounce-free switching in 4µs has been recorded, 20µs is more typical. Although 5V is the target, it takes 40V to operate a MEMS switch this quickly at the moment.



Tuning in... On-chip electronic filter components, inductors and capacitors, are getting the MEMS treatment at the University of California, Berkeley. The 200 x 200µm capacitor is an aluminum plate suspended in air above an aluminum layer by four springs. A 3V DC bias electrostatically causes the plate to pull down, increasing capacitance from 2.04 pF to 2.35 pF. Q is 60 at 1GHz, ten times better than silicon on-chip capacitors. The inductor is made around a 650nm by 500nm former. Inductance is 4.8nH with a Q of 30 at 1GHz. VCO tuning range is from 855MHz to 863MHz.

Other drawbacks are low manufacturing yield and limited life. Pellon said: "We need to improve the single switch yield to greater than 95 per cent and want a 30 billion operation life, enough for five years of continuous operation."

MEMS frequency-control components for filters and oscillators fall into two types. One uses micromachining to make conventional capacitors and inductors without the limited Q of traditional on-chip reactive components. This is being championed by the University of California, Berkeley – see photographs on this page. These components require special packaging, but operate at atmospheric pressure.

The other approach is to micromachine mechanically resonant structures that are capacitively coupled to the signal path and profoundly affect the signal. These are being made at the University of Michigan by professor Clarke Nguyen. So far his devices, which use combinations of inter-linked

masses and cantilevers, have all operated at VHF frequencies.

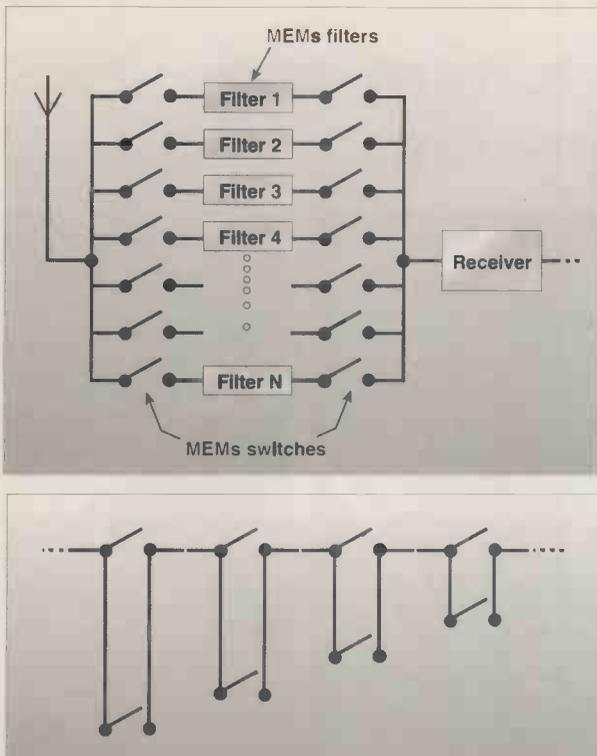
Some suggest that this is the upper frequency limit of the technology, but he is strongly dismissive: "You could easily make 300MHz without sub-micron lithography. Gigahertz frequencies are reasonable."

His work has shown that not only can filtering devices be made, but that

Vibration problems?

"To make a sensitive accelerometer you need a large mass and a very compliant spring. Resonator devices operate at VHF, and in future, UHF. Mass has to be small and stiffness high. These resonators are very much 'anti-accelerometers', less sensitive than crystals and surface acoustic wave devices," said Professor Clarke Nguyen of the University of Michigan.

This said, variable capacitors made by micromachining are exactly masses on compliant springs and are therefore somewhat vulnerable to microphony under vibration. In addition, variable capacitors also suffer from Brownian motion noise, although this is not thought to be a big problem at this stage.



New horizons... Architectures that are unthinkable using off-chip components become viable with a MEMS approach. For instance, rather than implement a variable filter at the front of a radio receiver, the diagram above shows one mechanically-resonant filter per reception band. Each performs all out-of-band rejection needed for the application and they are switched in and out using micromachined electrostatic relays. The whole thing could be built in the corner of a chip. The diagram below shows a phase-shifter, constructed simply from lengths of track and electrostatic micro-relays.

they can be as time-stable as quartz if they are fabricated from mono-crystalline silicon. There are several factors holding back the introduction of RF MEMS. Technology stability is one. Micromachined devices are made

using many types of wafer processing. What is needed is a standard process, even if it does have some limitations, so that fabs can offer it for production and IP developers can start making libraries. Michigan's Nguyen favours a four-mask addition to the usual CMOS process as a basic MEMS process. Designers are another problem. Few of them think in terms of MEMS solutions yet.

Lastly, design tools are needed to develop micromachined products. These have to be able to handle the physics of the mass-spring structures used as well as the fluidics of surrounding atmospheres. When will RF MEMS be qualified for production? Professor Bernhard Boser of the University of California, Berkeley draws an analogy: "Micromachined accelerometers were five years from invention to qualification for use in airbags. RF MEMS have been around for two years so far."

Steve Bush

Companies advised to act now on electronics waste recycling

Companies need to start acting now to ensure they can comply with the electronic waste recycling directive even though its legal enforcement is several years away.

"One of the best ways to prepare is to start now and learn," said Claire Snow, director at ICER during a meeting on the subject at Sutton Borough Council. "It's no good putting our heads in the sand."

ICER is a cross-industry organisation that is seeking to influence the directive by giving feedback on the drafts. "We're trying to achieve sensible legislation and promote electronics recycling," said Snow.

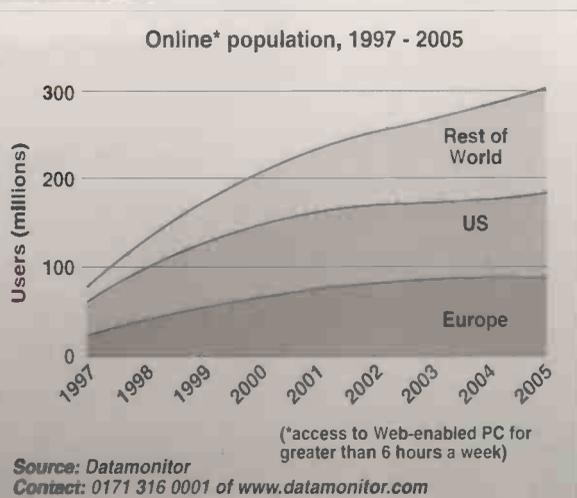
But designers currently have little awareness of environmental issues and their knowledge needs to be increased. "Eco-design is relatively advanced thinking so you cannot expect companies to get to that level without building up to it," said Martin Charter a co-ordinator at The Centre for Sustainable Design. One of the Centre's aims is to help designers gain this knowledge.

The EC's directive on recycling waste from electrical and electronic equipment (WEEE) is expected to be implemented nationally between 2002 and 2004. It will cover the recycling of 11 categories of

electrical and electronic equipment from large household appliances to telecoms and hi-fi equipment.

At a recent ERA conference on the subject the high cost of compliance was highlighted. Speaking at the conference, ICL's head of environmental affairs Joy Boyce said that it will be producers that have to face the cost, estimated to be between one and three per cent of turnover. But it will be an opportunity for producers to forge new links between themselves, their suppliers and their customers as an essential part of the recycler relationship.

DATA WEEK



400 new jobs in mobile-phone sector

Motorola is to create 400 manufacturing jobs at its Easter Inch mobile phone plant in Silicon Glen, Scotland.

"We will be recruiting over the next few months," said Motorola spokesman Derek Milne.

The jobs will be in assembly, inspection and testing of the handsets. Many of the positions will be filled from staff brought in for the pre-Christmas rush. "Current temporary employees will be given the opportunity of applying for the new permanent positions," said Milne.

Motorola put its expansion down to the popularity of its existing handsets and its new product line-up. "This will be the only Motorola site worldwide that makes the GSM version of the

new V-series," said Milne. "Although the analogue version is made in the US," he added.

Manufacturing decline may be bottoming out

Manufacturing decline may be bottoming out, according to the latest monthly order figures from the Chartered Institute of Purchasing and Supply. Manufacturing orders continued to decline in February but more slowly than in previous months.

The survey indicated the first signs of stronger demand in some sectors. New orders were down, but the monthly fall in February was far less than at the height of order book losses last October.

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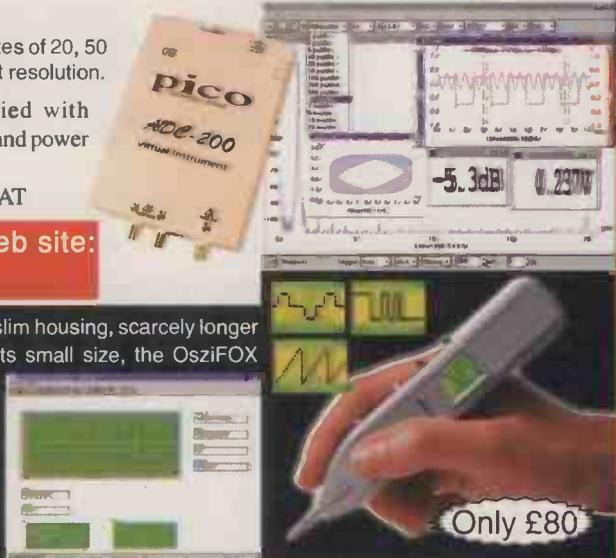
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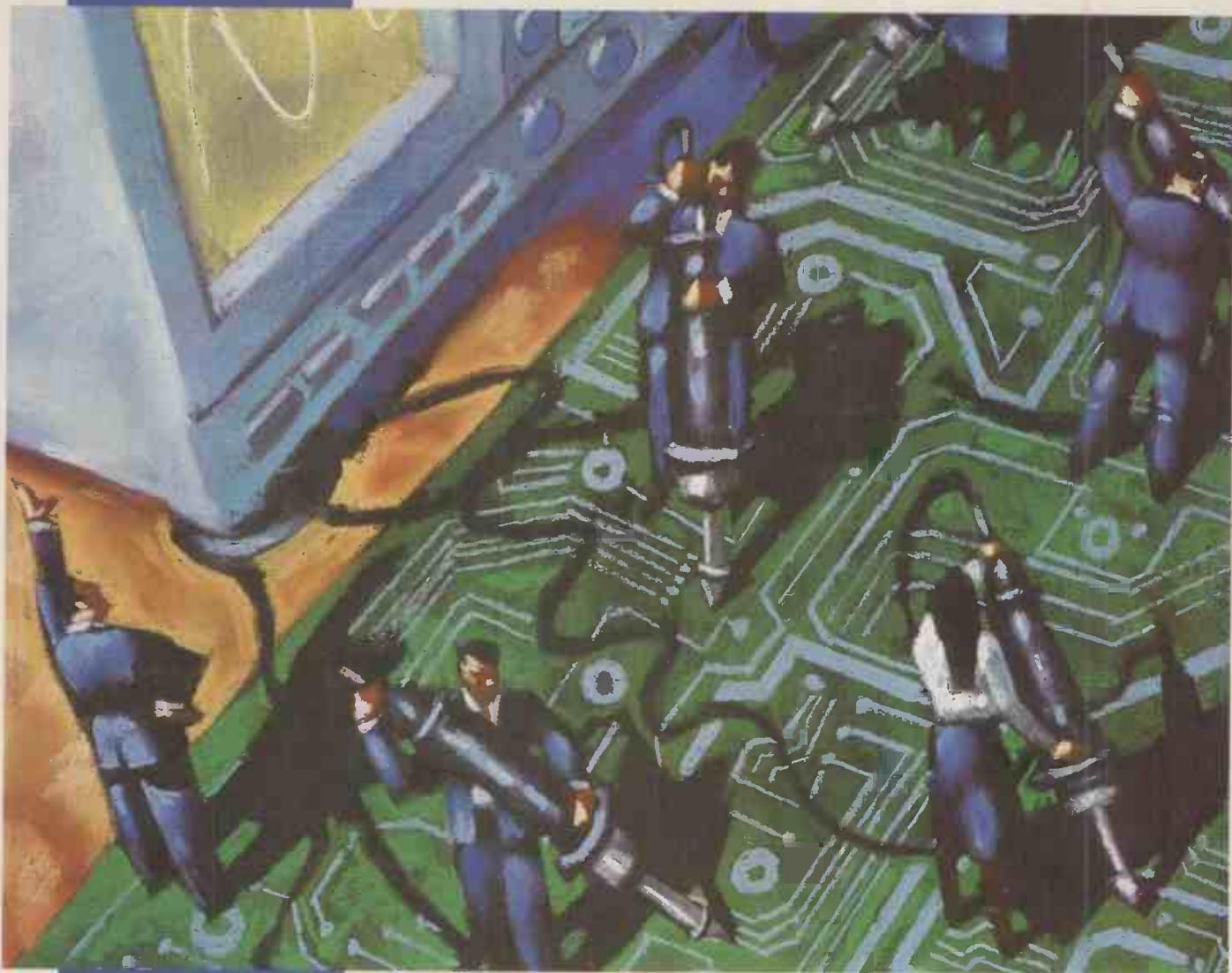


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Check Cs *in situ*

If you think that the best way to check an electrolytic capacitor is to measure its capacitance, think again. Cyril Bateman explains why it isn't, and starts a discussion on how to identify failed electrolytic capacitors without removing them from the board. This discussion ends next month with a design for a radically different in-circuit electrolytic tester.

Any capacitor that fails open or short circuit can be quickly identified without removing it from its circuit board. Most non-electrolytic capacitor failures result in a permanently short-circuited component.¹

The traditional aluminium electrolytic capacitor though is different. While it is self-repairing, it also has a built-in ageing mechanism.²

Such a capacitor's useful life ends when the oxygen needed to maintain or repair its dielectric oxide film can no longer be provided by the electrolyte. At this point, the conductivity of the electrolyte is much reduced, increasing the capacitor's equivalent series resistance, or ESR, at all frequencies.

Failed aluminium electrolytic capacitors exhibit four common symptoms. The first two are visual and hence easily identifiable.

- Leakage of electrolyte, usually around one termination. This can result from internal gas pressures, caused by internal reverse cathode bias voltage.
- Discoloured insulating sleeve caused by excess heat. This may result from adjacent components raising the local ambient temperature or from the capacitor self-heating.
- Electrolyte exhaustion, commonly called 'drying up'. The measured capacitance may be little changed and still within tolerance. The electrolyte may still be liquid, but have insufficient conductivity for the capacitor to function properly.
- Cathode-foil oxide growth, called 'forming up', results when the cathode foil's normal voltage, 'reverses' to some 1.5V positive, relative to the electrolyte. This internal 'reverse bias' induced cathode oxide growth can provoke the other symptoms.²

Aluminium electrolytic capacitors used correctly within their ratings usually fail due to the electrolyte being exhausted. Such capacitors are frequently described as 'dried up'.

Electrolyte exhaustion has two causes. These are consumption of the available oxygen by the capacitor leakage currents, and permeation of electrolyte through the capacitor seals.

Electrolyte exhaustion results in increased ESR, increase of $\tan\delta$ and a measurable increase of impedance at higher frequencies. Aluminium electrolytic failures that result in a short circuit are rare. I have never seen one.

Understanding electrolytics

To identify a failed aluminium capacitor, you need to understand how the capacitor works. Capacitor ESR is a combination of three frequency-dependent mechanisms. These are the electrolyte resistance, the electrode foils with their connection resistances, and the anode and cathode foils' dielectric loss factors expressed as a series resistive loss, Fig. 1.

There's more on this in the panel entitled 'Equivalent series resistance.'

Figure 1 illustrates a conventional 'polar' electrolytic capacitor.² Both anode and cathode foils possess capacitance. The capacitance of the cathode is very much larger than that of the anode. The measured value of the capacitor is the series sum of both values.

A 'bi-polar' or reversible electrolytic capacitor differs by having two anode foils formed to the same voltage. The diodes then have the same breakdown voltage. Both foils have similar capacitance values, each double the capacitor's marked value.

Leakage currents

This equivalent circuit of an electrolytic capacitor merits exploration. Aluminium is a 'valve' metal, so called because in the 1850s, researchers first noted that insulating oxide films, grown on the anode metal in a bath of suitable electrolyte, exhibited a rectifying action. While this metal is connected to a positive voltage, the oxide film is an excellent insulator.

Immersed in electrolyte and connected to a few negative volts, this oxide film becomes a conductor.² Current flow releases oxygen which travels to this negative voltage on the oxide film. If prolonged, hydrogen chemically 'reduces' some oxide back to aluminium, degrading its insulating properties.

Oxides of many other materials including niobium, zirconium, hafnium, uranium and silicon exhibit similar characteristics. Traditionally only aluminium and tantalum have been used to manufacture capacitors.³

Capacitor diodes?

Following my last series of articles, one reader asked how capacitor dielectric oxide could rectify. Alumina substrate used to manufacture thick film hybrid circuits is an excellent insulator. He posed the question, "Could this rectifying behaviour result from the electrolyte?" An explanation appears in the panel entitled, 'Substrates versus dielectric.'

Aluminium electrolytic capacitors do exhibit this rectifying behaviour. It easily demonstrated. Simply measure a polar capacitor's leakage current with forward and reverse DC bias applied. Under DC bias, if the dielectric film were non-rectifying, then regardless of polarity, similar leakage currents would flow. Any rectifying behaviour must result from the oxide films.

Take a new polar aluminium electrolytic capacitor and subject it to a low direct voltage of the correct polarity. To prevent any re-forming of the oxide films from influencing the results, charging currents should be limited. Supply this voltage via a 1k Ω current limiting resistor in series with the test capacitor.

Wait two minutes for the capacitor to charge and for the leakage current to stabilise. Increment this voltage and again allow the current to stabilise. Then plot leakage current from zero to the capacitor's rated or surge voltage.

Now reverse the polarity of the DC supply and measure and plot leakage current from zero to, say, negative 5V in 1V steps. Leakage current for negative voltages is considerably higher than for the same positive voltage.

To avoid any possibility of oxide formation, as well as for safety, the stabilised capacitor leakage currents should not be

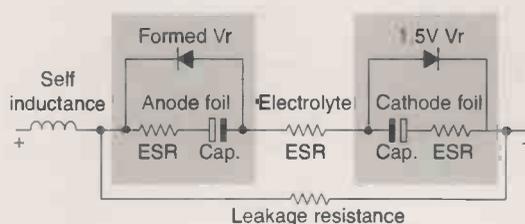


Fig. 1. Equivalent circuit of a polarised aluminium electrolytic capacitor. The cathode foil is etched, not formed. Its natural atmospheric oxide roughly equates to 1.5V electrical formation. Using similar foil thickness and etch ratio as the anode foil, a 6V rated capacitor's cathode capacitance will exceed that of the anode by a good margin. CV products on the other hand will be similar.

Alumina substrate versus dielectric oxide

The oxide dielectric of an aluminium electrolytic capacitor is grown on the surface of the anode foil by connecting to a positive voltage while it is immersed in a bath of suitable electrolyte.²

Oxygen, freely available from the electrolyte, combines with aluminium from the surface of the foil to form an aluminium oxide, namely Al_2O_3 . This oxide is extremely thin, attaining some 14 ångströms thickness for each volt applied.³

The best purity materials are used both for the aluminium anode foil and for the forming electrolyte, so the resulting oxide is extremely pure. It has a dielectric strength approaching the theoretical strength predicted by the ionic theory of crystals.

Growing anodically under this DC voltage stress ensures consistent alignment of the extremely small aluminium oxide particles as they form. This oxide growth is self-limiting. As thickness approaches 14Å for each volt applied, anode current falls. Oxide growth slows and almost stops.

A typical capacitor dielectric has an oxide thickness less than 1µm. For example, a 10V capacitor has a dielectric oxide thickness of some 0.02µm.²

A thick-film hybrid substrate is made in similar fashion to the ceramic capacitor dielectric, described in an earlier article.⁷ The Erie company, for which I then worked, was almost certainly the first major UK maker of thick-film circuits. Having considerable ceramic development and production resources, the company used these facilities to make alumina substrates for hybrid production.

The various processes of milling, spray drying, pressing or casting, followed by sintering, ensure a random alignment of alumina particles. These particles are much larger than those grown as the electrolytic capacitor dielectric.

The spray drying process typically produces dried powders in the form of minute hollow spheres. Subsequent processes of pressing or casting flatten these spheres, ensuring the deposited grains have a completely random alignment. Most substrates incorporate a significant glass or frit content. This acts as a flux during the sintering process, when these grains clump together and grow in size.

Typically manufactured as a 0.5mm thick ceramic plate, alumina substrates comprise a large number of randomly-aligned grains, electrically in series with each other, to build up the required thickness of substrate. This makes them a good insulator.

allowed to exceed 1mA. The resulting graph displays typical diode behaviour, Fig. 2.

Should a suitable microammeter not be available, the voltage drop across this resistor can be used. A 200mV DMM measures 1µA as 1mV.

Component quality

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

The reciprocal of Q, called $\tan\delta$, is defined as the capacitor's ESR/reactance ratio. $\tan\delta$ is used to describe the quality of almost all general-purpose capacitors.

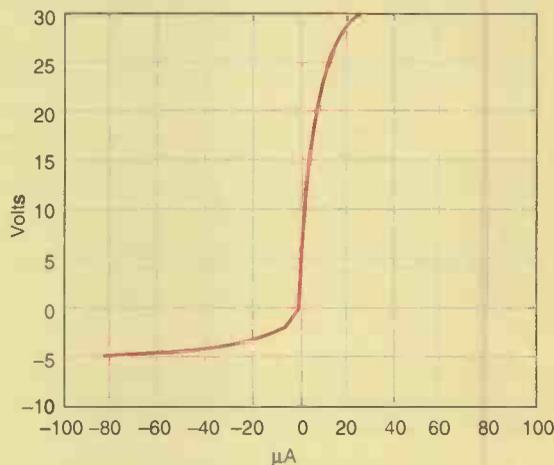
$$\tan\delta = \text{abs} \frac{ESR}{X_C}$$

$$X_C = \frac{1}{2\pi fC}$$

Conversely, $ESR = X_C \tan\delta$.

The capacitor's reactance reduces in proportion to its

Philips 135-4700µF/25 volt
Forward/reverse 'diode' voltage characteristic
100kHz impedance – red 25mΩ



Value measured with 1kΩ limiting resistor
Unused capacitor.
Capacitance 4 743µF, $\tan\delta$ 0.0927

Fig. 2. Typical 'diode' forward/reverse characteristic common to all polar aluminium electrolytic capacitors. This capacitance value and voltage rating was plotted to answer the question Chris Green posed in the letters pages of the October 1998 issue. It was selected to provide easily measurable leakage currents for forward and reverse bias voltages.

The rectification effect permitted by the extremely thin, electrolytically grown capacitor dielectric oxide is easily measured, Fig. 2.

If you are interested in this topic, I recommend that you try this. The rate that leakage current increases when the capacitor's voltage is reached and exceeded is a good indication of the 'formation' voltage margin used, and hence the potential life time of the capacitor. Traditionally, long life or professional capacitors are built using foils with higher formation voltages than those used for miniature and commodity capacitors.

capacitance value and frequency. Being a combination of fixed and variable losses, ESR also reduces with frequency but to a lesser extent.² Having reached its minimum value, ESR then increases with frequency.

The measured $\tan\delta$ of a capacitor therefore must always increase with frequency. From the first equation, $\tan\delta$ has no upper limit. It can exceed unity.

Equivalent series resistance is related to the construction of the capacitor, its voltage rating and capacitance value. A capacitor's ESR varies with frequency. At any frequency, ESR and impedance of each capacitor value and voltage rating varies widely. No single global good or bad figure can thus be assigned. ESR or impedance can be used to identify a worn capacitor, but only by comparison against known good identical capacitors.

At 100 or 120Hz depending on supply frequency, $\tan\delta$ is used by capacitor makers to indicate aluminium electrolytic capacitor quality. Every capacitor is tested for $\tan\delta$ in production. While many makers also table a 10kHz or 100kHz impedance or ESR value by capacitor, these parameters are not tested.

Variation of $\tan\delta$ for capacitor size and voltage rating is small compared to the wide range of ESR and impedance.

ESR is not a fixed value

One extremely common mistake is to consider that a capacitor's ESR has a fixed value.

Electrolytically-formed aluminium oxide is a low-loss dielectric, changing little with frequency.³ Expressed as the equivalent series loss, its contribution to ESR reduces with increasing frequency. The resistances of the metal foils and the electrolyte/paper combination, tend to increase but more slowly, with frequency.

Dominated at low frequencies by the $\tan\delta$ of the oxide system, the capacitor's measured ESR initially reduces with frequency. With further increase of frequency, resistances of the metal foils and the electrolyte/paper combination, dominate. The capacitor's ESR and impedance become almost constant, finally rising with frequency, Fig. 3.

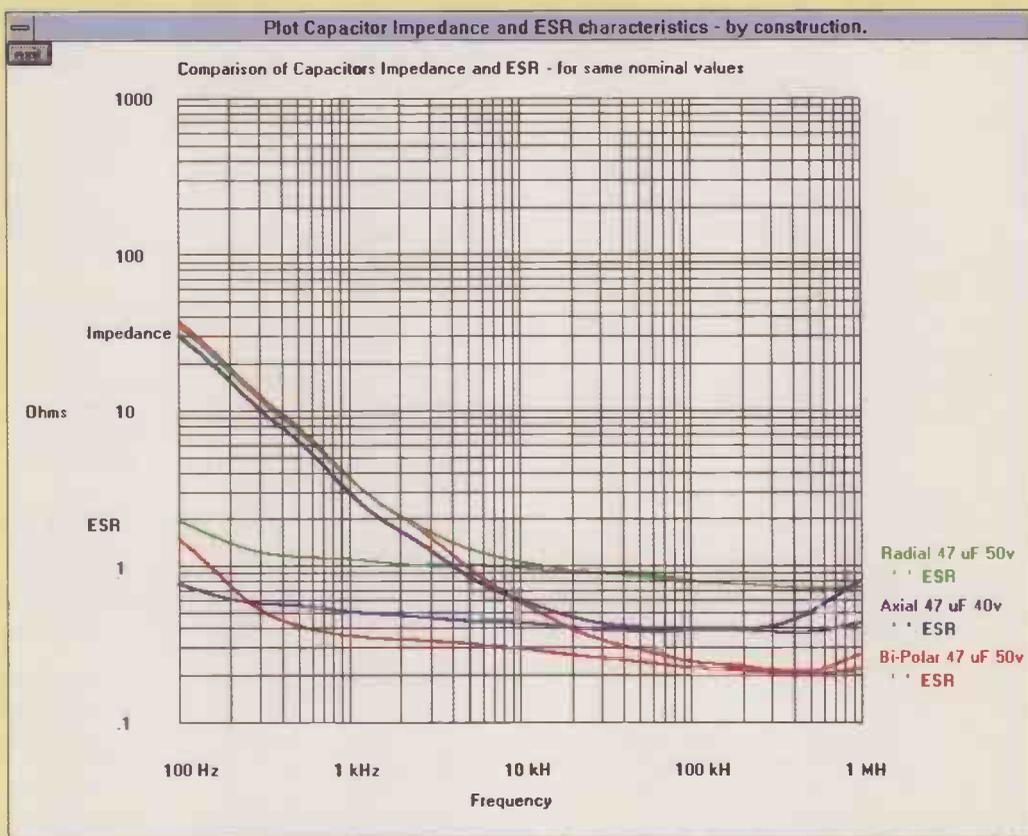


Fig. 3. Plots of impedance and ESR against frequency for three 47µF capacitors, typical of the range of impedances and ESR found at 100kHz with medium voltage electrolytic devices. Lower and higher voltage capacitors will further increase this spread.

$\tan\delta$ provides a good figure of merit for all capacitors.

ESR is of interest to a capacitor designer to determine ripple current or power rating.⁴ But where capacitor quality issues are concerned, 100Hz $\tan\delta$ is the criterion used – not ESR.

From equation 1, $\tan\delta$ responds to change of capacitance or ESR. A bad or failing aluminium electrolytic's capacitance usually reduces slightly, while ESR increases significantly. $\tan\delta$ reflects both changes.

Temperature effects

Capacitor leakage currents are temperature dependent, roughly doubling for each 10°C increase, according to Arrhenius' law. Since the consumption of free oxygen from the electrolyte determines a capacitor's service life, so does its working temperature.

Subject to an AC ripple current, the capacitor's ESR results in the dissipation of real power as the product of I^2ESR . This raises the capacitor's internal temperature. Each aluminium electrolytic capacitor has a sinusoidal ripple-current rating, usually based on a frequency of 100 or 120Hz. Correlation factors for other frequencies and change of ambient temperature are provided.

Assuming the circuit that the capacitor is used in applies a sinusoidal or other similar easily defined waveform, then compliance with the capacitor makers ratings is easily confirmed.

These ideal waveforms rarely occur in practical circuits though. Given a repetitive voltage or current waveform, power dissipation can be calculated, but this can be difficult.

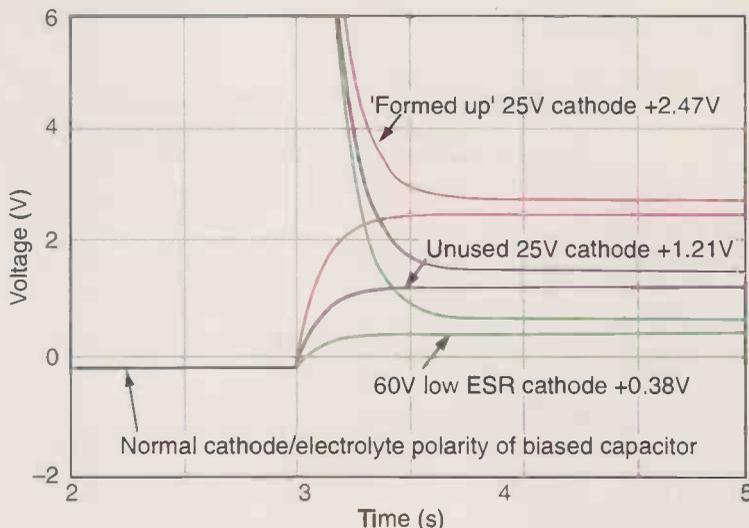


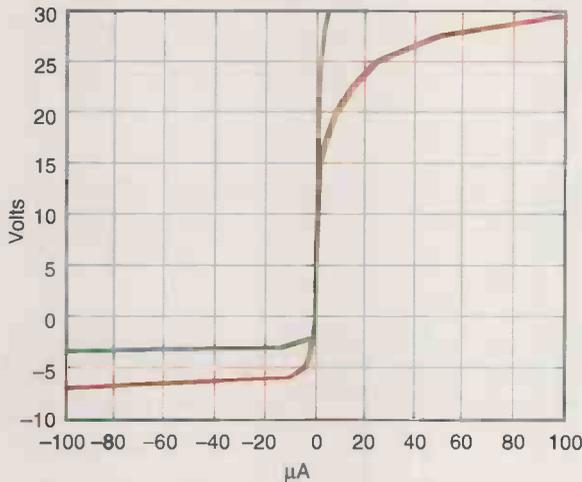
Fig. 4. Simulations of capacitors charged to 25V then discharged via a 10Ω resistor demonstrate how the cathode foil's voltage becomes positive, with respect to the electrolyte. With positive anode bias voltage, the cathode voltage stabilises slightly below that of the electrolyte. With similar anode and cathode foil CV product, the cathode foil charges to a positive voltage, compared to the electrolyte as in the green trace. When the cathode foil CV exceeds the anode's, red trace, a much lower voltage develops. In the blue trace, the cathode has 'formed up' and only half of original capacitance value remains.

In many instances – especially with non-repetitive waveforms – the only practicable method is measurement of the working capacitor's case temperature rise.

A full treatment of a mathematical method, applicable to any repetitive waveform, has already been published in *Electronics World*.⁴

220µF/25V new/used comparison

Forward/reverse 'diode' voltage characteristic
100kHz impedances – red 371mΩ, green 243mΩ

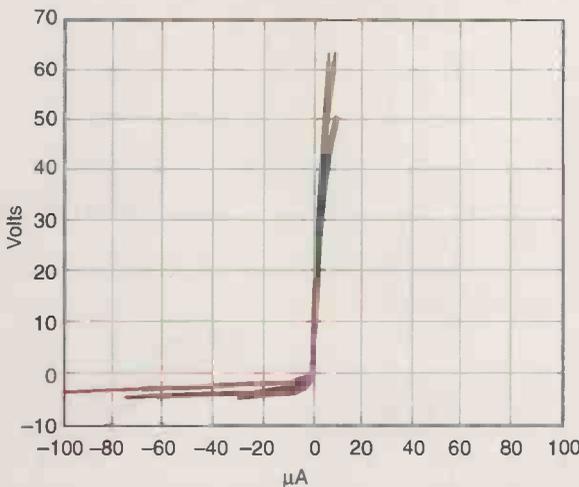


Value measured with 1kΩ limiting resistor
Red, the used capacitor's cathode has 'formed up'
Capacitance now 199µF, tandelta 0.107

Fig. 5. The red curve shows the degraded forward leakage and increased reverse voltage sustain of a capacitor that has suffered from internal reversed cathode bias. The cathode foil has clearly 'formed up'. Total capacitance at 100Hz has reduced to 199µF, but is still within tolerance. The cathode capacitance of this worn out capacitor has reduced to 30% of its initial value while tandelta has increased to 0.107. The green curve provides comparison with a similar unused capacitor, having 218µF and a tandelta of only 0.044, a 2.4:1 ratio of tandelta. The 100kHz impedances were 371 and 243mΩ, a ratio of only 1.5:1. Clearly, tandelta is the much more sensitive test.

220µF/63V and 50V new/used comparison

Forward/reverse 'diode' voltage characteristic
100kHz impedances – red 77.4mΩ, blue 119mΩ green 30mΩ



Value measured with 1kΩ limiting resistor
Red, the used capacitor, has not degraded
Capacitance now 220µF, tandelta 0.016

Fig. 6. The red curve for the used low ESR 220µF compared with two unused general-purpose capacitors. This shows that after four years service, this higher voltage, low ESR replacement has not degraded. Measured values, red 220µF, 0.016 tandelta; blue 213µF 0.024 tandelta; green 214µF, 0.028 tandelta.

Used correctly within its ratings, an aluminium electrolytic capacitor's leakage current will slowly consume the oxygen available from its electrolyte. Its capacitance value will change and ESR will increase. A capacitance change of ±10% or tandelta or impedance increasing by 1.2 times the rated limits indicates that the capacitor's useful life has ended.

A capacitor used within its published ratings should provide many years of service though. Recently, I refurbished an elderly Hewlett Packard test instrument. Because of its age, as is good practice, I replaced all of its 43 aluminium electrolytic capacitors. These were all dated 1974 or 1975. Only two failed to meet their specifications when tested on a bridge.

Why then do some capacitor applications cause such a drastic reduction in a capacitor's life? Ignoring those obvious reasons – excess ambient temperature, voltage, ripple current or reversed polarity – most premature aluminium electrolytic capacitors failures occur because their cathode foil has become reverse biased internally. The foil has become what is known as 'formed-up'.

Internal cathode reversed bias

Forming up cannot easily be observed using an oscilloscope, except by manufacturing special capacitors to allow measurement of the electrolyte's voltage relative to the anode and cathode foils. Using the circuit of Fig. 1, though, it can be simulated.

The anode and cathode foil diodes can be represented by the 'default' diode in PSpice. I used 'BV=1.5' to represent the unformed cathode foil diode and the capacitor's rated voltage for the anode foil diode. For accuracy, I also selected 'RS' values, which replicate the measured DC leakage current/voltage plots.

If prolonged, internal reverse bias can cause the cathode foil to form up above its natural atmospheric oxide's equivalent electrical value. This reduces its capacitance.²

Both anode and cathode foils exhibit capacitance. They are connected in series internally, back to back, by the electrolyte. When a capacitor is charged or discharged, charge transfers between anode and cathode foils. This develops voltages between the anode foil and electrolyte and between electrolyte and cathode foil, according to each foil's capacitance value, Fig. 4.

Under normal positive anode bias voltages, the cathode voltage is slightly negative with respect to the electrolyte voltage. When the capacitor is discharged, the charge transfer develops a cathode voltage which is positive compared to the electrolyte. The cathode has become internally reverse biased, relative to normal operation.

This transfer of charge may cause the cathode foil voltage to approach or exceed that sustainable by the cathode foil's naturally occurring oxide film. Considerable leakage current then flows between cathode foil and electrolyte, causing oxide growth on the cathode.

Once initiated, this oxide growth enters a runaway condition. With repeated capacitor discharge, the reducing cathode capacitance develops an ever higher voltage. Cathode foil oxide growth, cathode voltage, ESR, tandelta and leakage currents all increase until the capacitor fails.

Heavy-duty capacitors

Repetitive crash discharging, as needed for photoflash, requires a specially constructed capacitor. While photoflash presents an extreme case, many circuit applications use electrolytic capacitors to couple irregular waveforms but without ensuring adequate bias voltage.²

Depending on the applied waveform, these uses are considered as repetitive charge/discharge or AC applications. Commercial polarised capacitors are not suitable for either application so special capacitors are needed for each.

Frequently, the base drive waveforms of switching transistors include a small capacitor in parallel with a resistor to shape and couple the drive current into the transistor base. My satellite receiver's power supply originally had a $1\mu\text{F}$ electrolytic paralleled with a $1\text{k}\Omega$ resistor to drive the base of a *BUT11A* transistor. This capacitor failed very quickly, destroying the power supply.

This is a well known design fault. Replacing the component with a $1\mu\text{F}$ polyester film capacitor provided a permanent remedy.

Similarly, my TV set had a $220\mu\text{F}$, 25V capacitor paralleled with a 12Ω resistor to drive the base of its *BU508* line-scan transistor. After only four years service, this TV exhibited an underscanning display area.

Causes of rapid failure

Capacitor anodes are made using super-pure aluminium. Cathodes are deliberately made using lower purity foils. This is done to discourage cathode formation under the normal charge/discharge cycles found during equipment switch-on or off.

Capacitors described as charge/discharge proof in compliance with CECC 30 300 are tested to survive a million switch-on or off cycles using a time constant of 0.1s.

When a capacitor is internally or externally reverse biased, abnormal amounts of oxygen are consumed from the electrolyte. The consequent excessive free hydrogen can force electrolyte past the capacitor seals.

Hydrogen collecting at the anode foil degrades its dielectric oxide, increasing leakage current when re-biased correctly. The cathode foil can form up, reducing its capacitance.

Capacitance values drop while leakage current, ESR and $\tan\delta$ all increase, until the capacitor or the equipment fails. For example, see the red plot of the $220\mu\text{F}$, 25V capacitor removed from my TV in 1995. While its measured capacitance was still in tolerance, the display no longer filled the screen.

This damaged capacitor has changed dramatically. Its cathode foil has formed up and the cathode's capacitance has reduced to 30% of its original value, Fig. 5.

For this article I removed and measured the low-ESR 63V capacitor that replaced the damaged original one. Although it had seen a similar length of service, as can be seen from the leakage current plots, the replacement has survived undamaged, Fig. 6.

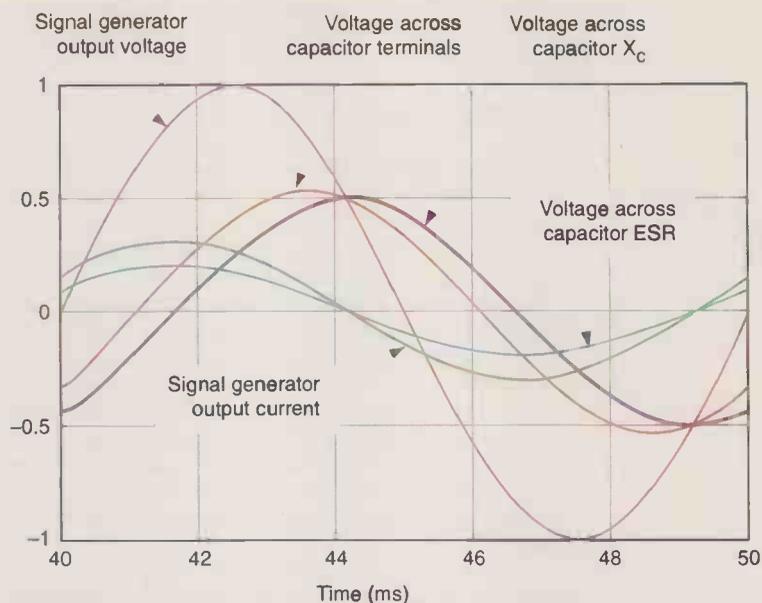


Fig. 7. PSpice simulation showing voltages and currents of a $1000\mu\text{F}$, $0.4 \tan\delta$ capacitor. Note how the peaks of the voltage waveform for the capacitor's ESR coincide exactly with the generator's current waveform. The peak of the capacitor X_C voltage waveform is separated by exactly 90° phase. Similar simulations confirmed that these relationships hold over the range of $\tan\delta$. The generator voltage/current phases vary widely though, depending on the capacitor's ESR and the generator's source resistance.

Anode foil is considerably more expensive than cathode foil. To ensure meeting the charge/discharge requirements and reduce costs, aluminium electrolytic capacitor designers ensure the CV product of the cathode exceeds that of the anode. This maximises the available capacitance at minimum cost.

To avoid cathode formation in capacitors that could be subjected to internal cathode reverse bias, the CV product of the cathode foil should be increased. The best solution is to use a bi-polar capacitor. Alternatively, a low ESR capacitor of higher voltage than required for the circuit voltages may be the answer.

Low-ESR capacitors are made using high-gain cathode foils with more conductive paper and electrolyte relative to standard capacitors. This cathode foil provides more capac-

Electrolytic capacitor ESR

Capacitor electrolytes are conducting solutions, usually a neutralised weak acid in a solvent. This electrolyte must not freeze or boil at the extremes of the capacitor's working temperature range, nor attack pure aluminium at any temperature. Very pure ethylene glycol was for many years the standard solvent used in capacitors.²

Most modern electrolytes are made without added water. Using only dry ingredients, such as super purity ammonium borate crystals, some water appearing as water of crystallisation is inevitable.

Low-voltage electrolytes have a low resistivity – typically a few millisiemens – and are not polarity sensitive. In a capacitor, some electrolyte is contained within the minute voids and channels in the anode and cathode foils' oxide coatings. These can be tenuous and very long relative to their cross section,² so the effective electrolyte resistance in them increases quickly with frequency.

Most of the electrolyte however will be absorbed in the separator. This is usually a paper tissue, interwound with the anode and cathode foils during assembly. Resistivity of the

electrolyte/paper separating tissue is increased compared to the bulk electrolyte, according to paper type and thickness used.

Low-voltage capacitors use very low resistivity paper and electrolyte. To help lower ESR, a single open weave 'rag' tissue is often used. Higher voltage capacitors incorporate a higher resistivity paper/electrolyte combination, and more than one tissue thickness.

As the capacitor's rated voltage increases, electrolyte, paper and cathode-foil thickness are varied to minimise $\tan\delta$ losses, capacitor case size and production costs while ensuring a satisfactory capacitor service life and performance.

Low-ESR capacitors use thicker, higher surface gain cathode foils together with more conductive paper/electrolyte combinations than standard capacitors of the same voltage. Naturally, these carry a size and cost penalty.

Electrolyte/paper resistivity varies with temperature. Above room temperature, this resistivity reduces, but due to the paper's influence, it perhaps only halves by 85°C . Below 0°C though, resistivity increases rapidly as the solvent approaches its freezing point.

itance, and hence CV product, for a similar foil area. Increasing capacitor voltage increases foil area. These contribute a much improved service life, Fig. 6.

In-circuit capacitor diagnosis

Having removed a suspect capacitor from a circuit board, you can confirm whether it has failed by measuring its $\tan\delta$ using a bridge.⁵ But it is usually easier, quicker and much cheaper simply to remove and replace all suspect capacitors.

If you do this though, you will no doubt be throwing away a lot of good capacitors. Is there a better way?

Measuring a capacitor *in situ* poses three main problems. To avoid false readings, the test voltage must be sufficiently low that adjacent semiconductor junctions are not turned on. Measurement times must be much less than needed to remove and replace a suspect capacitor. The test result should be unambiguous, needing little or no interpretation.

Measurement of capacitance is of little help. As with my TV, most failed aluminium electrolytic capacitors still have a measurable capacitance and can appear well within tolerance.

At present, a few commercial in-circuit impedance testers, incorrectly called ESR meters, are available. These measure the capacitor's impedance at high frequency, which is a much easier measurement – and not its ESR.

A capacitor's impedance, $|Z|$, comprises the vector sum of its ESR and its capacitive/inductive reactances. At low frequency, capacitor self-inductance matters little. But as frequency increases, every capacitor ultimately becomes a DC-blocking inductive reactance. Its reactance then increases rapidly with frequency.

$$|Z| = \sqrt{ESR^2 + (X_{C(s)} - X_{L(s)})^2}$$

$$X_{L(s)} = 2\pi fL(s) \text{ and } X_{C(s)} = \frac{1}{2\pi fC(s)}$$

At any one frequency, the term $X_{C(s)}^2 - X_{L(s)}^2$ can be simplified to $\pm jX_s$, giving the fundamental vector capacitor equation for impedance,

$$|Z| = ESR_s \pm jX_s$$

Many writers assume that capacitive reactance at 100kHz is so small that impedance is the same as ESR. This is not correct. Many larger electrolytics – especially axial types – may be above resonance, hence inductive. At 100kHz, a the-

oretical 1 μ F capacitor has a reactance of around 1.6 Ω . A practical electrolytic capacitor's reactance will be considerably above this theoretical value, Fig. 3.

Most so-called ESR meters with a pre-set good/bad buzzer cause confusion when used to measure smaller value electrolytic capacitors.¹ But smaller capacitors are the ones most frequently measured. Often the value and voltage of the capacitor being measured is not visible, making judgement difficult.

Because of the range of impedances involved, an 'ESR' meter's results require considerable interpretation and comparison against known good capacitors. Since 100kHz impedance values for known good electrolytic capacitors range from 0.01 Ω to 24 Ω , it is clearly not possible to define an impedance value representing 'good' or 'bad', Table 1.

Compared to these impedance variations, the $\tan\delta$ of a good capacitor is a reasonably constant number. For a typical commercial aluminium electrolytic capacitor, $\tan\delta$ ranges from a low of 0.02 to a high of 0.3 for very large low voltage parts.

Philips' capacitor data handbook requires general-purpose capacitors subjected to endurance testing to have a $\tan\delta$ of ± 1.5 times the catalogue limits or 0.4, whichever is larger.⁶

As a general guide, the $\tan\delta$ for typical good board mounted capacitors should be less than 0.1. Medium sized capacitors having a $\tan\delta$ of 0.2 or more should be replaced for reliability, Table 2.

Advantages of looking at $\tan\delta$

Does in-circuit $\tan\delta$ measurement have any particular disadvantages? Using the internet and various trade publications, I have been unable to locate a suitable low cost in-circuit $\tan\delta$ meter, so I decided to build one to supplement my conventional capacitor bridges.

Using PSpice I plotted the waveforms simulating an electrolytic capacitor with 0.4 $\tan\delta$, representing a failed capacitor. These demonstrate the difficulty measuring $\tan\delta$ or true ESR, compared to the easy measurement of impedance.

While the vectors representing the capacitor's ESR and reactance remain separated by 90°, their phase relationship to the signal generator's voltage varies according to source impedance and test capacitor $\tan\delta$, Fig. 7.

This PSpice simulation confirmed that the signal generator's current and the current, and hence voltage waveforms, of the capacitor's ESR, do remain in phase. Simulation also confirmed that the voltage waveform for the capacitor's reactance is delayed by exactly 90° relative to that of the capacitor's through current.

Encouraged by these PSpice analyses, I set out to investigate the design of a suitable in-circuit $\tan\delta$ tester. Its design proved rather more difficult than I at first expected. But I have managed to design and develop an easy to use $\tan\delta$ in-circuit tester. It will be described in my next article. ■

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Table 1a). Typical impedances measured at 100kHz – low capacitance values.

Capacitor	1 μ F	2.2 μ F	4.7 μ F	10 μ F	22 μ F	47 μ F	100 μ F
50V bipolar Al.	4.0 Ω	3.2 Ω	1.4 Ω	0.9 Ω	0.35 Ω	0.3 Ω	0.22 Ω
63V polar Al.	4.3 Ω	3.5 Ω	1.8 Ω	1.4 Ω	0.5 Ω	0.4 Ω	0.28 Ω
450V polar Al.	24 Ω	11 Ω	5 Ω	3.8 Ω	1.5 Ω	1.0 Ω	

Table 1b). Typical impedances measured at 100kHz – high capacitance values.

Capacitor	1000 μ F	2200 μ F	4700 μ F	10 000 μ F
25V polar Al.	0.090 Ω	0.07 Ω	0.045 Ω	0.022 Ω
63V polar Al.	0.050 Ω	0.025 Ω	0.015 Ω	0.010 Ω

Table 2a). Typical $\tan\delta$ values of new stock capacitors measured at 100Hz – low capacitance values.

Capacitor	1 μ F	2.2 μ F	4.7 μ F	10 μ F	22 μ F	47 μ F	100 μ F
50V bipolar Al.	0.05	0.05	0.05	0.05	0.05	0.05	0.06
63V polar Al.	0.04	0.04	0.035	0.035	0.035	0.045	0.04
450V polar Al.	0.1	0.1	0.08	0.05	0.05	0.05	

Table 2b). Typical $\tan\delta$ values of new stock capacitors measured at 100Hz – high capacitance values.

Capacitor	1000 μ F	2200 μ F	4700 μ F	10 000 μ F
25V Polar Al.	0.06	0.075	0.09	0.1
63V Polar Al.	0.03	0.05	0.06	0.07

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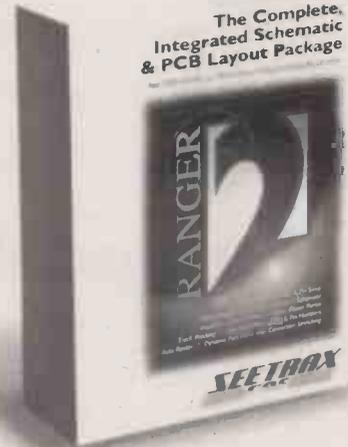
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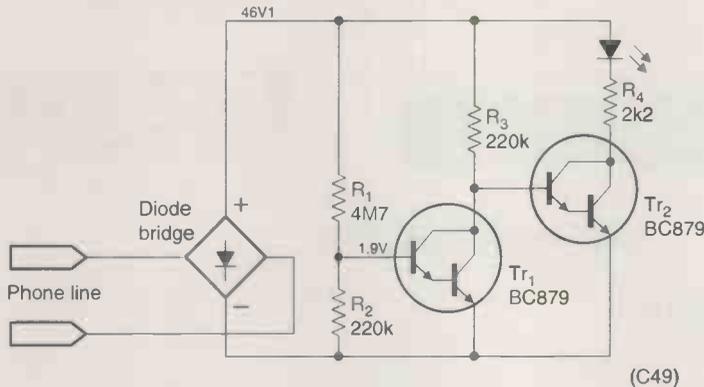
Since the state of an internal modem is sometimes a little indeterminate, this was intended to indicate whether the modem is on-line or off; it also

shows when the telephone is in use. It needs no batteries and may be fixed to the computer, possibly with Velcro.

At the base of Tr_1 there should be a

voltage of 1.9V, the transistor being a 100V, 1A BC879 darlington, which therefore conducts when the modem is off-line. When on-line, the line voltage falls, Tr_1 turns off and Tr_2 conducts, activating the led, a low-current, high-brightness type.

Robert L A Trost
Duiven
Holland
C49



(C49)

Line monitor indicates whether a modem is on-line or off-line and will tell you when your three-year-old has dialled the fire brigade from the bedroom.

Video muting circuit

While maintaining sync. and colour burst signals so that the picture may be switched on or off as required with no effect on other equipment, this circuit switches

composite video to black, an advantage in video multiplexing.

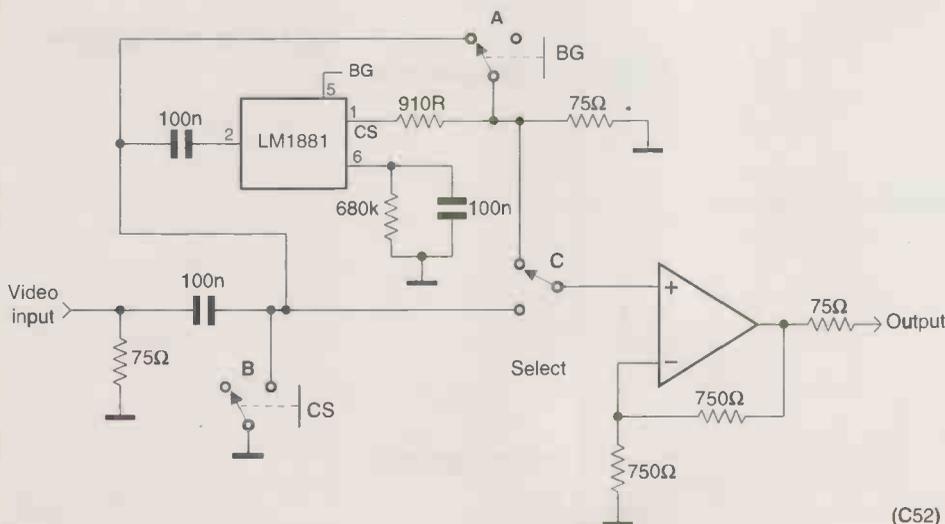
The LM1881 sync. separator provides composite sync and burst gate signals used to control 74HCT4053 cmos

switches A, B and C, shown here as mechanical types for simplicity.

Composite sync. is potted down and may receive colour burst at switch A to produce a video signal synchronised to the video input, albeit without picture content. Video input is clamped to the same level at switch B and both signals are selected by switch C.

Switching is best carried out in the vertical interval of the first field, which can be done by a little logic controlled by the odd/even flag on the sync. separator.

Jim Irlam
Wraysbury
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C52



(C52)

For use during operations such as video multiplexing, this simple circuit allows the picture to be turned on or off while keeping sync. and colour burst intact.

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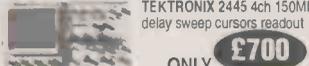
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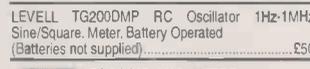
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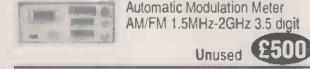
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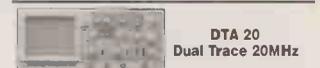
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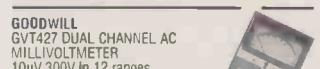
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Semiconductor 'barretter'

To limit an inrush of current in, for instance, lamps, devices whose resistance varies with current are often used. Normally though, these so-called 'barretters' are not protected against overload and their limiting current cannot be set.

This circuit avoids these problems. When connected to a short circuit, current through Tr_1 flows through R_3 and potential divider $R_{4,5,6}$. Voltage on the potentiometer output, which is proportional to the load current, controls Tr_3 and consequently the base current of Tr_2 , which in turn controls the power transistor Tr_1 .

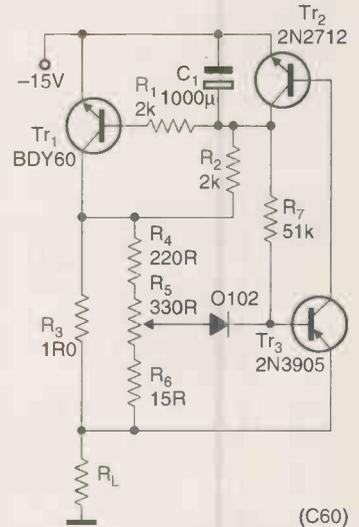
The result is that current increases smoothly to the point set by the potentiometer, where it stays with no

further increase. The diode and resistor R_7 set initial bias on Tr_1 to 500mV. Resistor R_6 sets the maximum load current at 100-600mA and R_4 determines the minimum current of 10-100mA.

Time taken for current to increase to the steady state may be varied by C_1 and $R_{1,2}$. Transistor Tr_1 should be on a heat sink.

Michael A Shustov
Tomska
Russia
C60

Transistor version of a current-limiting barretter, this one with overcurrent protection and variable current limiting.



Ring of leds for first-year students

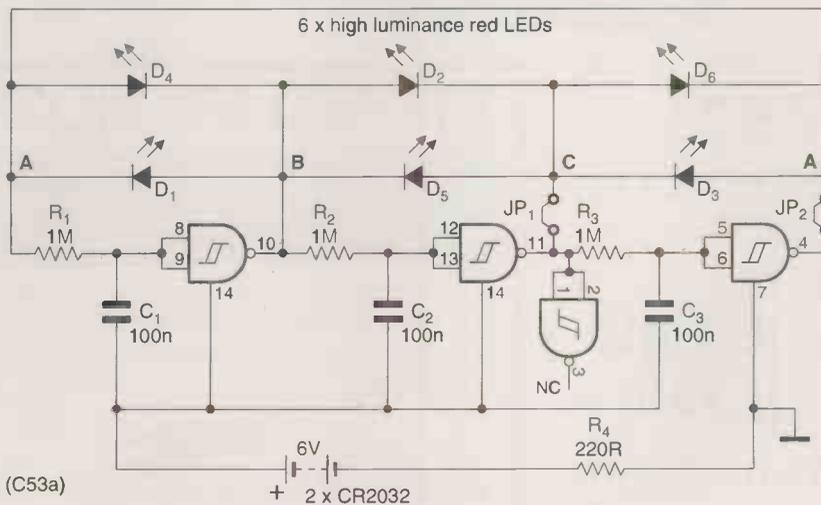


Fig.1. As a project for students, this illustration of capacitor charging is also entertaining.

Capacitor charging seemed a good subject for study in an electronics first year, so this collection of leds came about.

Capacitors $C_{1,3}$ in Fig. 1 charge in sequence with time constants R_1C_1 , etc., with sufficient overlap to produce the timing diagram in Fig. 2. The result is that the leds come on in the sequence $D_6D_1, D_1D_2, D_2D_3, D_3D_4, D_4D_5$ and back to D_6D_1 . Reversing the connections of JP1 and JP2 reverses the sequence.

The leds are high-brightness types such as TLLR 4400 and the 220Ω resistor limits current in the leds and logic. The circuit will work at voltages down to 3V, a current draw of 5mA giving 50h from a couple of CR2032 lithium cells.

J M Terrade
Clermont-Ferrand
France
C53

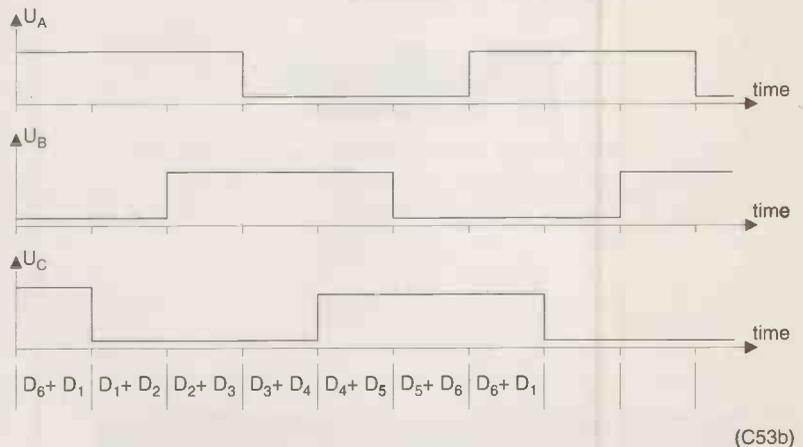
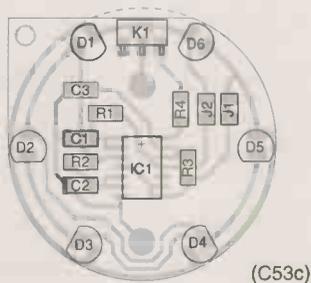


Fig.2. Waveforms at points A-C and the sequence of led illumination



Buffer connects RGB monitor to a pc

Five-wire VGA video produced by a standard pc is incompatible with the three-wire standard used by high-quality Unix workstation monitors. This circuit allows such monitors to be used with pcs, while also driving VGA displays.

Two three-way video buffers, IC_{1,2}, take the same video input from VGA in. Output to the VGA display is direct and includes the sync., ID lines and power management signals. The MAX499CWG is a 135MHz full-power-bandwidth triple video switch, giving buffered outputs to the

workstation monitor.

Since, in Unix monitors, composite sync. is superimposed on the green signal, the buffer's 3ns switching time is necessary here to switch between sync. and colour signal. Composite sync. is obtained by an Ex-Or operation between horizontal and vertical sync., R_{10,11} setting composite sync. level.

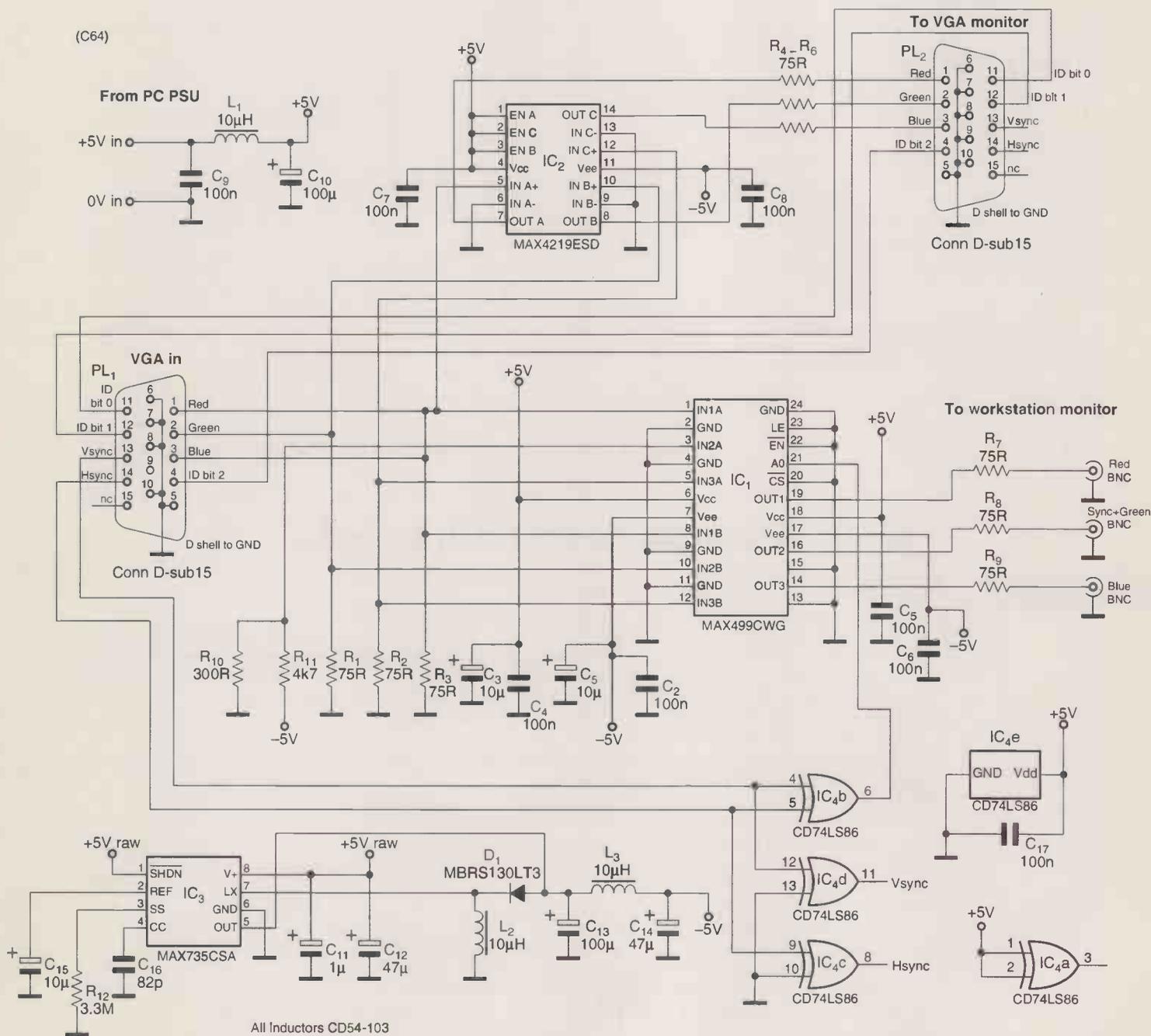
The 200MHz MAX4219 buffer drives a VGA monitor from the original input, so that a workstation monitor may be used for the audience in a presentation, for example, while the

presenter sees a VGA display.

A convenient power source for the circuit is the standard four-wire lead in the pc, which gives +5V and +12V for peripherals. The 5V is used and is filtered by L₁C₁₀. An inverter, the MAX735, provides the -5V.

The graphics card must be carefully chosen to ensure that the refresh timing is adjustable to suit the workstation.

Tim Herklots
Maxim Integrated Products
Theale
Berkshire
C64



Oscillator runs at 0.24µA

A CD4007 contains three pairs of n-channel and p-channel fets and is used here to form an oscillator drawing 0.24µA at an output frequency of 0.87Hz and 9.1µA at 32Hz.

Two of the n-channel fets form a Schmitt trigger, the output of which on pin 5 drives an inverter input on pin 10. Capacitor C₁ charges towards the supply rail through R₁ and when its voltage reaches the threshold of the Schmitt, pin 5 goes high and the inverter output on pin 12 goes low, C₁ discharging through a diode.

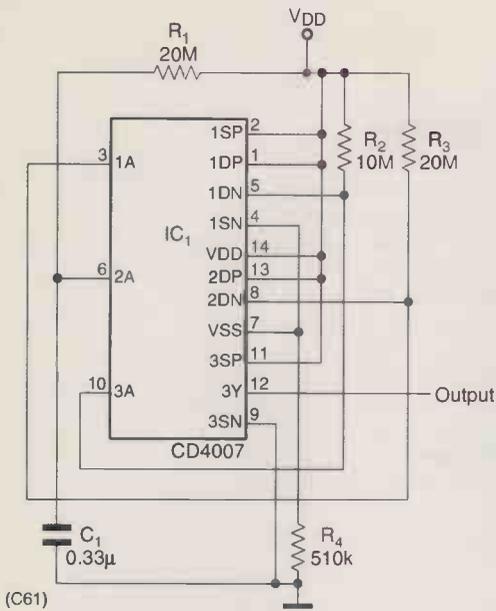
As the C₁ voltage reaches the lower switching threshold of the Schmitt, pin 5 goes low, pin 12 goes high, blocking the diode. The cycle then repeats.

Output pulse width is 0.35-0.6ms with a supply voltage of 5-2.5V.

Yongping Xia
Torrance
California
C61

V _{DD}	I _{DD} (µA)	f _{out} (Hz)
2.5	0.24	0.87
3.0	0.37	0.71
3.5	0.51	0.93
4.0	0.88	2.2
4.5	2.4	7.1
5.0	9.1	32

This oscillator takes only 0.24µA at a frequency of under 1Hz.



(C61)

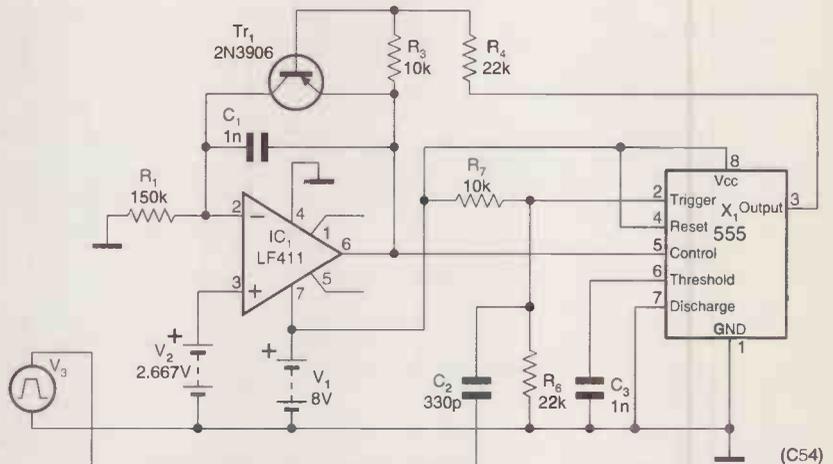
Ramp timer

This is a voltage-variable timer delivering a ramp waveform, controllable by a voltage.

Capacitor C₁ charges at a constant rate in Miller integrator fashion. When the op-amp output reaches 67% of the supply voltage, the 555 output goes low, turning on the p-n-p transistor to discharge C₁. The transistor stays on while the 555 output is low, so the op-amp output and input are at the same voltage. A negative pulse from the trigger source drives the 555 output high to restart the cycle.

A voltage on the 555 control pin varying between 33% and 67% of the supply voltage will produce an output pulse varying in width between zero and R₁C₁.

Dinarte Santos
Brazil
C54



Triggered timer produces an output pulse controllable in width by voltage input.

(C54)

4-bit flash makes 16-bit flash a-to-d

A high-resolution a-to-d converter can be made using two or more low-resolution types, as has been described¹. This note illustrates the use of a single 4-bit converter to obtain a 16-bit circuit.

There are four phases of operation, in each of which an analogue 'nibble' is generated. Every such nibble output from the 4-bit converter is latched by 4-bit latches, which in turn drive d-to-a converters.

To take the operation in sequence, analogue input v_i goes to the 4-bit converter to produce the 4-bit most significant output D₁₅-D₁₂. In the second phase of operation, the analogue of the most significant nibble of the d-to-a digital output is inverted (I₁), subtracted from v_i in S₁ and amplified by 16 times, the result being the analogue of the most

significant nibble. Four such phases are carried out.

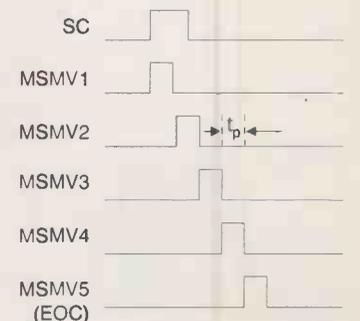
Five monostable flip-flops generate control pulses for the process of enabling the analogue input to the 4-bit converter and for latching its output, the chain of flip-flops being activated by the start conversion pulse.

K Balasubramanian
B Gayathri
Huseyin Camur
European University of Lefke
Turkish Republic of Northern
Cyprus
C50

Reference

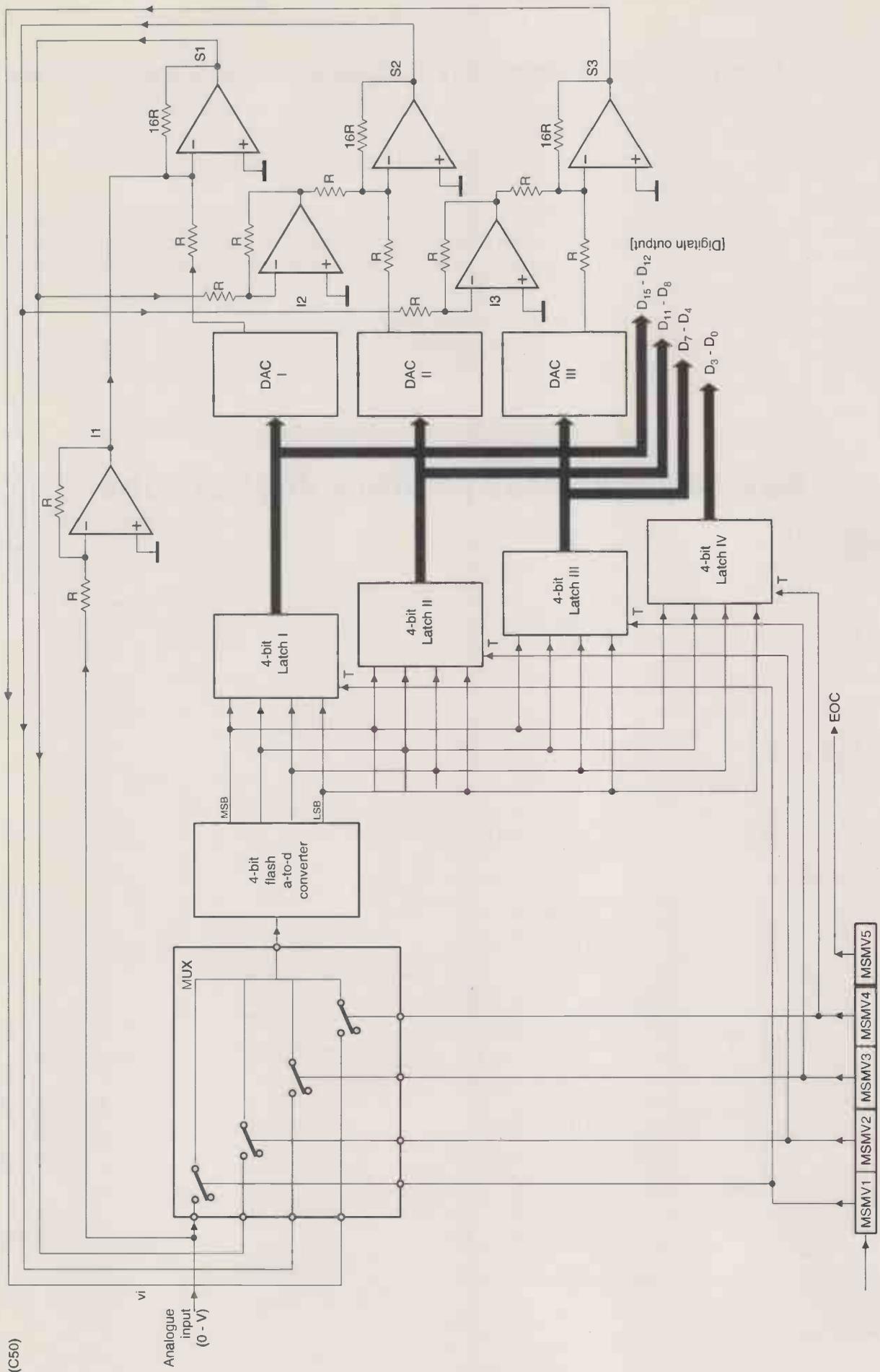
1. Balasubramanian, K. High resolution a-to-d using low-resolution converters. *Electronics World*, December 1991, p.1052.

Four-bit analogue to digital flash converter used sequentially to process nibbles of the input gives a 16-bit digital output.



(C50)

£100 WINNER



(C50)

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

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

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

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

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

Power supply for photomultiplier tube

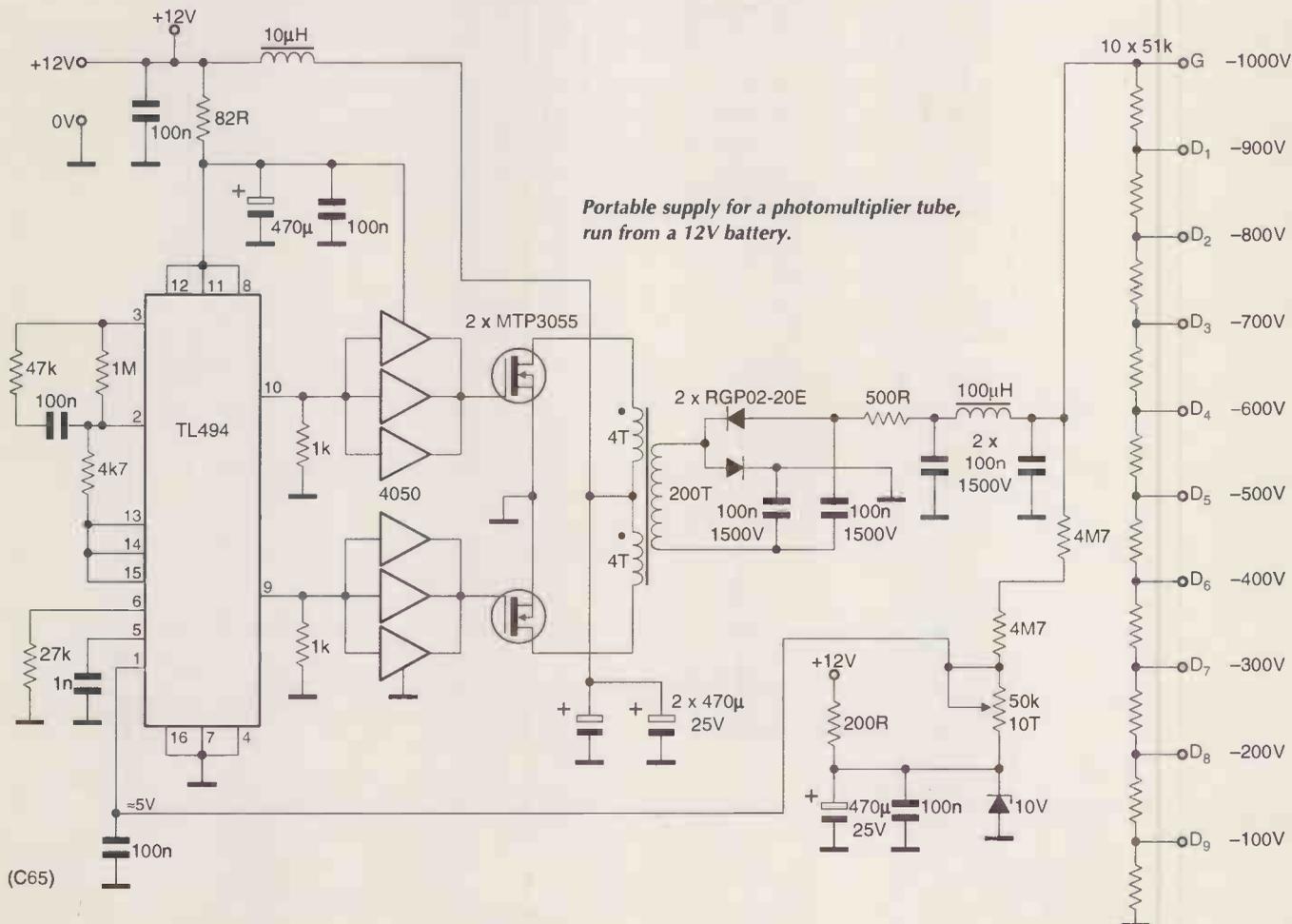
£100 WINNER

This circuit arose from the need to power a photomultiplier tube from a portable 12V source. It drives a 931A tube, but may be used with other types, providing a voltage of -1kV divided in a resistive chain to drive the tube dynodes. The use of a

zener diode to provide an error signal to the TL494 pulse-width modulation controller allows a negative output; improved stability can be obtained by the use of a voltage reference diode. The pot core for the transformer is an FX2242.

When setting the circuit up, set the potentiometer to maximum resistance before adjusting for the required output.

Matthew A Curlis
Melbourne
Australia



Telnet

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RF mixers

Joe Carr explains the importance of choosing the right rf mixer in all applications up to microwave.

My last two articles have developed the basic theory behind RF mixers, and introduced a simple mixer circuits. This month I take a look at higher performance mixer circuitry.

A mosfet double-balanced active mixer¹ is shown in Fig. 1. The dual-gate mixer is ideal for this type of application, but the fets need careful selection.

Instead of the 3N211 mosfets shown, 40673s could be used. These have the advantage of being relatively insensitive to casual electrostatic damage during handling, and are low cost. In addition, in a circuit such as this, the 40673 can provide between 15 and 20dB of conversion gain.

However, this gain is not without cost. The devices are relatively easily overdriven in the presence of large RF input signals. When this occurs the advantages of the device evaporate in increased intermodulation products and degraded noise performance.

The same circuit using 3N211 devices, or equivalent, will produce less conversion gain, of about -5 dB, but better overall performance. With local-oscillator injection of 8V peak-to-peak, and a 10dBm RF input signal, this circuit will exhibit a respectable third-order intercept point of +17dBm.

Bipolar alternative useful to 500MHz

An active double-balanced mixer based on n-p-n bipolar transistors² is shown in Fig. 2. This circuit is usable to frequencies around 500MHz.

Normally, the use of non-IC transistors in a circuit such as this normally requires matching of the transistors for best performance. That need is overcome by using a bit of degenerative feedback for $Tr_{1,2}$ in the form of un-bypassed emitter resistors $R_{3,4}$.

The base circuits are driven with the

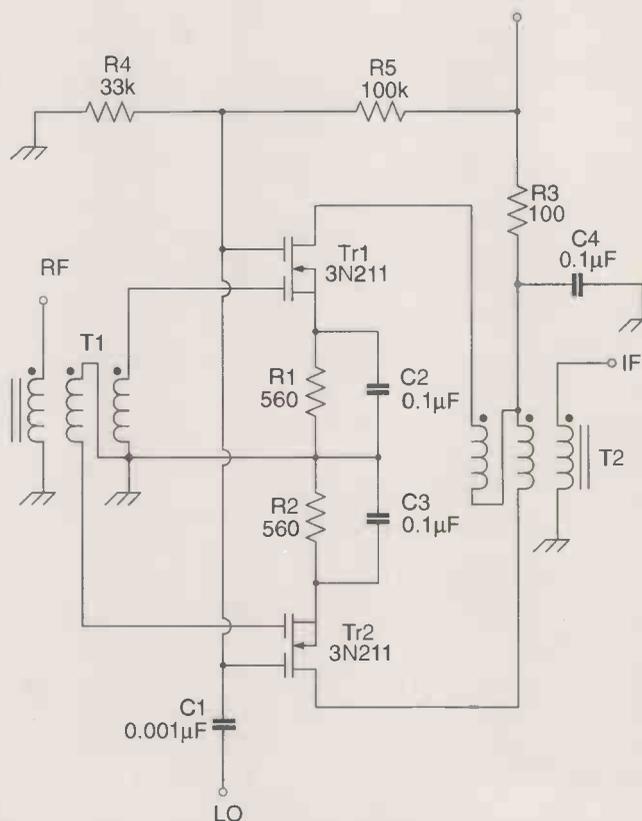


Fig. 1. The dual-gate mixer is useful, but the fets need to be chosen carefully for best performance.

LO signal from a balun transformer, T_1 in a manner similar to the earlier junction-fet circuit. The output transformer, however, is rather interesting.

It consists of four windings, correctly phased, with the IF being taken from the junction of two of the windings. The RF signal is applied to the remaining two windings of the transformer.

This mixer exhibits a third-order intercept of +33dBm, with a conversion loss of 6dB, and only 15 to 17dBm of LO drive power.

Benefits of quad mos and junction fets

Although using bipolar transistors can result in an active double-balanced

mixer with a high third-order intercept point, or TOIP, there is a trend towards using junction and mos fets.

Typical designs use four active devices. This approach is made easier by the fact that many IC makers are producing RF mos and junction fet products that include four matched devices in the same package.

Figure 3 shows a mixer circuit based on the use of four junction-fet devices (Tr_{1-4}). These transistors are arranged such that the source terminals of $Tr_{1,2}$ are tied together, as are the source terminals of $Tr_{3,4}$.

The source-pair terminals receive the RF input signal from transformer T_2 . The gates of these transistors are con-

Joseph J. Carr,
MSEE

Fig. 2. Bipolar mixers such as this are useful to 500MHz. Normally, discrete transistors in a circuit like this would need matching for best performance. Adding a little degenerative feedback gets round this problem.

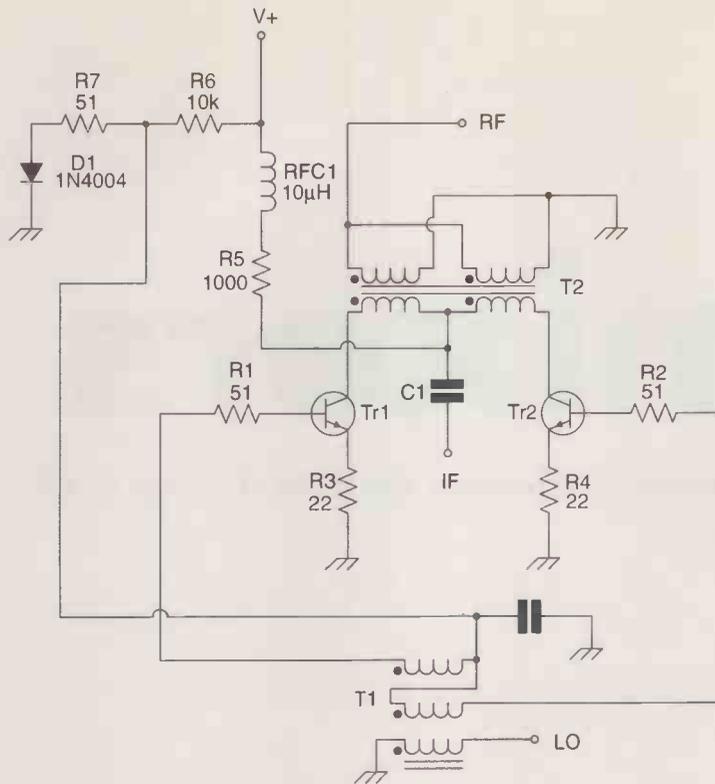


Fig. 3. Quad junction-fet balanced mixer. The fets need to be matched, but packages with four matched fets designed for such applications are readily available.

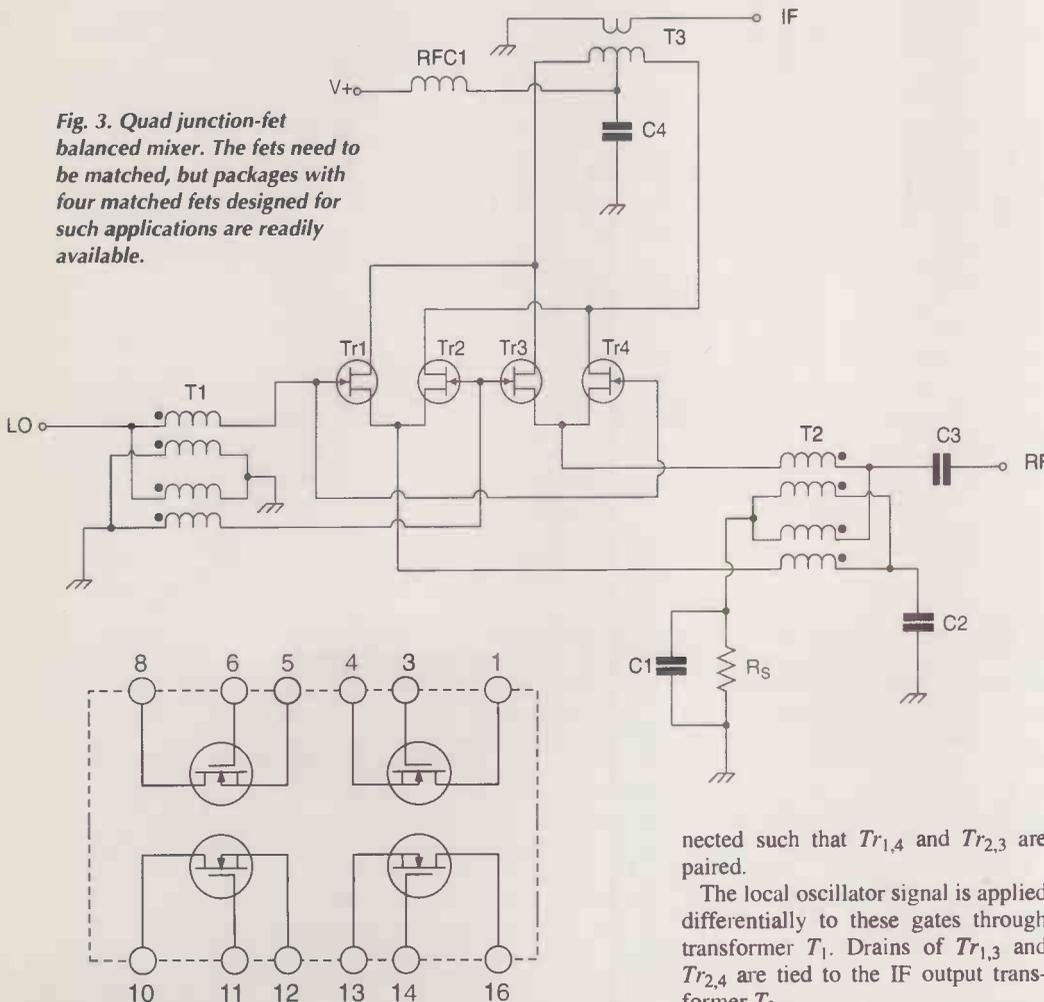


Fig. 4. Siliconix quad MOSFET device contains four fets that can be used independently. When connected as a ring, they form a mixer.

SD5000 quad DMOS fet is shown in Fig. 4.

This device contains four fets that can be used independently. When connected as a ring, they form a mixer.

Calogic carries the theme a little further in its SD8901 DMOS quad fet mixer IC, Fig. 5. The fets, $Tr_{1,4}$, are connected in a ring such that opposite gates are connected together to form two LO ports, LO_1 and LO_2 .

The RF signals are applied differentially across drain-source nodes $Tr_{1,2}$ and $Tr_{3,4}$. Similarly, the IF output is taken from the opposite pair of nodes: $Tr_{1,4}$ and $Tr_{2,3}$. The SD8901 comes in an eight-pin metal can package.

The circuit for using the SD8901, Fig. 6, is representative of this class of mixers. The RF and IF output terminals are through transformers T_1 and T_2 , respectively.

The local-oscillator signal is applied directly to the LO_1 and LO_2 ports, but requires a divide-by-two circuit comprising a J-K flip-flop.

Note that this divider makes the LO signal a square wave rather than a sine wave. An implication of this circuit is that the LO injection frequency must be twice the expected LO frequency.

Gilbert-cell mixers

The Gilbert transconductance cell, Fig. 7, is the basis for a number of IC mixer – for example the NE602 shown in Fig. 8, and analog multiplier devices like the LM1496.

The Gilbert mixer consists of two cross-connected n-p-n pairs fed from a common current source. The RF signal is differentially applied to the transistors that control the apportioning of the current source between the two differential pairs. The local oscillator signal drives the base connections of the differential pairs.

Integrated Gilbert-cell devices such as the Philips/Signetics NE602 in Fig. 8 are used extensively in low-cost radio receivers. The Gilbert cell is capable of operating to 500MHz. An on-board oscillator can be used to 200MHz.

One problem seen on such devices is that they often trade off dynamic range for higher sensitivity.

Passive double-balanced mixers

The diode double-balanced mixer, Fig. 9, is one of the more popular double-balanced mixer approaches. It has the obvious advantage over active mixers of not requiring a DC power source.

The circuit uses a diode ring, $D_{1,4}$, to perform the switching action. In Fig. 9, only one diode per arm is shown, but some commercial double-balanced mixers use two or more diodes per arm.

This configuration is capable of 30 to

nected such that $Tr_{1,4}$ and $Tr_{2,3}$ are paired.

The local oscillator signal is applied differentially to these gates through transformer T_1 . Drains of $Tr_{1,3}$ and $Tr_{2,4}$ are tied to the IF output transformer T_3 .

There are several quad FET ICs on the market that have found favour as mixers in radio receivers. Siliconix's

60dB of port-to-port isolation, and is not difficult to apply.

With proper design, it is easy to build passive diode double-balanced mixers with frequency responses from 1 to 500MHz.

Commercial models that work into the microwave region are not hard to find. Intermediate-frequency outputs of typical diode double-balanced mixers can be DC to about 500MHz.

The diodes used in the ring can be ordinary silicon small-signal diodes such as 1N914 or 1N4148, but these are not as good as hot-carrier Schottky diodes like the 1N5820, 1N5821 and 1N5822.

Whichever diodes you choose though, they should be matched for use in the circuit. Diode differences can degrade a mixer's performance.

The usual approach is to match the diode forward voltage drop at some specified standard current, such as 5 to 10mA, depending on the normal forward current rating of the diode. It is also important to match the junction capacitance of the diodes.

The mixer in Fig. 9 uses two balun transformers, T_1 and T_2 , to couple to the diode ring. The double-balanced nature of this circuit depends on these transformers, and as a result the LO and RF components are suppressed in the IF output.

Diode double-balanced mixers are characterised according to their drive level requirement, which is a function of the number of diodes in each arm of the ring. Typical values of drive required for proper mixing action are

0dBm, +3dBm, +7dBm, +10dBm, +13dBm, +17dBm, +23dBm and +27dBm.

Figure 10 shows the internals of a commercially available passive double-balanced mixer made by Mini-Circuits (PO Box 166, Brooklyn, NY, 11235, USA: Phone 714-934-4500; Web site <http://www.minicircuits.com>). These type no. $SBLx$ and $SRAx$ devices are available in a number of different characteristics, Table 1.

The standard package for SRA/SBL devices is shown in Fig. 11. Non-insulated pins are grounded to the case. Pin-outs for common SRA/SBL devices are shown in Table 2.

The regular SRA/SBL devices use a local-oscillator drive level of +7dBm, and can accommodate RF input levels up to +1dBm. The devices will work at lower drive levels, but performance deteriorates rapidly, so it is not recommended.

Note that the IF output port is split into pins 3 and 4. Some models tie the ports together, but for others an external connection must be provided for the device to work.

The nice thing about this type of commercially available mixer is that the system impedances are already set to 50Ω. Otherwise, impedance matching would be necessary for them to be used in typical RF circuits.

Note, however, that if a circuit or system impedance is other than 50Ω, then a mismatch loss will be seen unless steps are taken to effect an impedance match.

The problem becomes considerably

greater when a mismatch occurs at the IF port of the mixer. The mixer works properly only when it is connected to a matched resistive load. Reactive loads, and mismatched resistive loads deteriorate performance.

Figure 12 shows a circuit using a passive diode double-balanced mixer. The diplexer is a critical component to this type of circuit.

Proper loading

The diplexer is a passive RF circuit that provides frequency selectivity at the

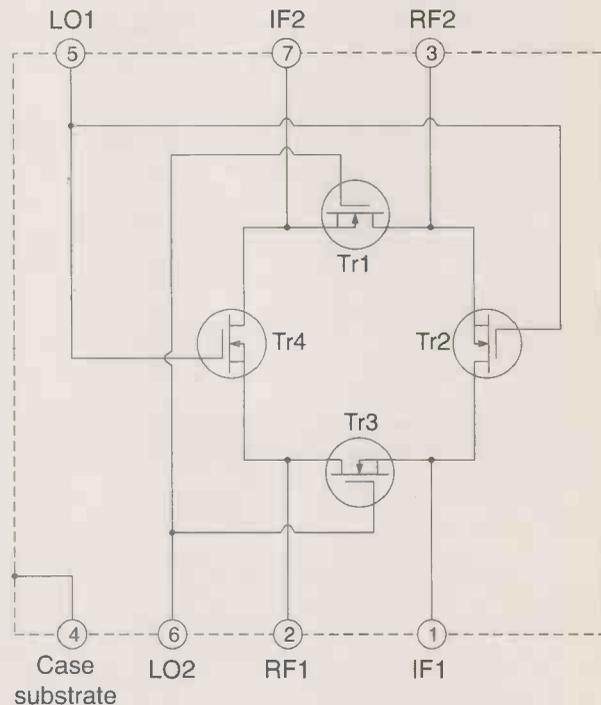


Fig. 5. In Calogic's SD8901 mixer, fets Tr_{1-4} are connected in a ring such that opposite gates are connected together to form two LO ports, LO₁ and LO₂. RF signals are applied differentially.

Fig. 6. Typical mixer application circuit using the SD8901. The local-oscillator signal is applied directly to the LO₁ and LO₂ ports on pins 5 and 6, but a divide-by-two circuit is needed so the LO signal is a square-wave – not a sine wave.

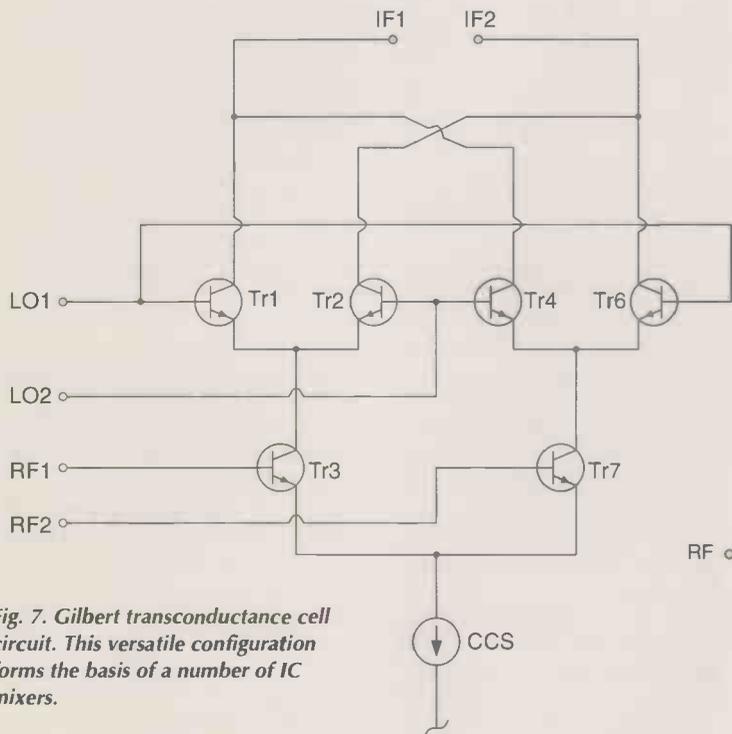
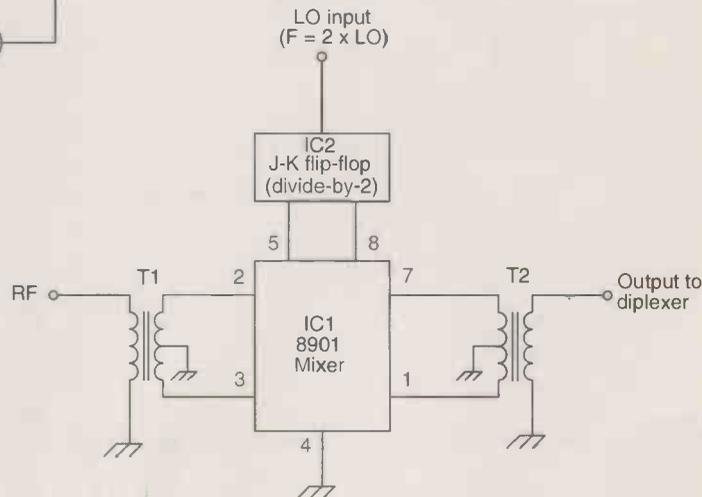


Fig. 7. Gilbert transconductance cell circuit. This versatile configuration forms the basis of a number of IC mixers.

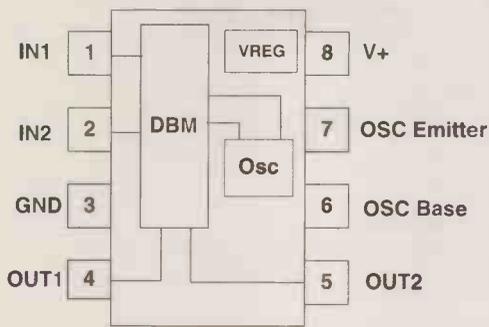


Fig. 8. One IC based on the Gilbert-cell principle is the NE602.

Table 1. Key performance figures for Mini-Circuits SBLx and SRAx passive double-balanced mixers, Fig. 11.

Type No	LO range (MHz)	IF range (MHz)	Mid-band loss (dB)
SRA-1	0.5-500	DC-500	5.5-7.0
SRA-1TX	0.5-500	DC-500	5.5-7.0
SRA-1W	1-750	DC-750	5.5-7.5
SRA-1-1	0.1-500	DC-500	5.5-7.5
SRA-2	1-1000	DC-500	5.5-7.5
SBL-1	1-500	DC-500	5.5-7.0
SBL-1X	10-1000	5-500	5.5-7.5
SBL-1Z	10-1000	DC-500	5.5-7.5
SBL-1-1	0.1-400	DC-400	5.5-7.0
SBL-3	0.025-200	DC-200	5.5-7.5

Table 2. Pin designations of SBLx and SRAx passive double-balanced mixers.

Type No	LO	RF	IF	GND	Case
SRA-1	8	1	3,4	2,5,6,7	2'
SRA-1TX	8	1	3,4	2,5,6,7	2
SRA-1W	8	1	3,4	2,5,6,7	2,5,6,7
SRA-1-1	8	1	3,4	2,5,6,7	2
SRA-2	8	3,4	1	2,5,6,7	2,5,6,7
SBL-1	8	1	3,4	2,5,6,7	2
SBL-1X	8	1	3,4	2,5,6,7	2
SBL-1Z	8	1	3,4	2,5,6,7	XX
SBL-1-1	8	1	3,4	2,5,6,7	XX
SBL-3	8	1	3,4	2,5,6,7	XX

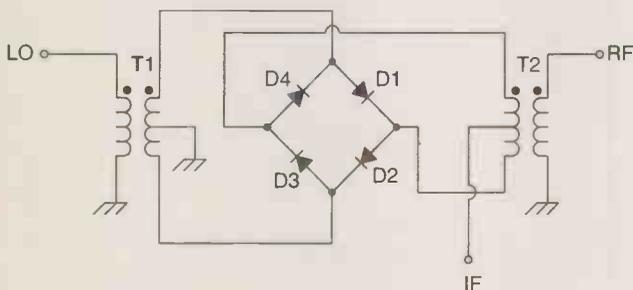


Fig. 9. Involving diodes instead of active switches, the passive double-balanced mixer has the benefit of not needing a DC power source.

binates two frequencies, F_1 and F_2 , to produce an output spectrum of $mF_1 \pm nF_2$, where m and n are integers representing the fundamental and harmonics of the two frequencies.

In some cases, you will only be interested in the difference frequency, so you will want to use the low-pass output of the diplexer, Fig. 14a). The high-pass output is terminated in a matched load so that signal transmitted through the high-pass filter is fully absorbed in the load.

The exact opposite situation is shown in Fig. 14b). Here we are interested in the sum frequency, so the high-pass output port of the diplexer is used, and low-pass output port is terminated in a resistive load. In this case, the signal passed through the low-pass filter section will be absorbed by the load.

output, while looking like a constant resistive impedance at its input terminal.

Figure 13 shows a generalisation of the diplexer. It consists of a high-pass filter and a low-pass filter that share a common input line, and are balanced to present a constant input impedance.

With appropriate design, the diplexer will not exhibit any reactance reflected back to the input terminal – eliminating the reflections and voltage/standing-wave ratio problem. Yet, at the same time it separates the high and low frequency components into two separate signal channels.

The idea is to forward the desired frequency to the output and absorb the unwanted frequency in a dummy load.

Figure 14 shows the two cases. In each case, a mixer nonlinearly com-

Bandpass diplexers

Figures 15 and 16 show two different bandpass diplexer circuits commonly used at the outputs of mixers. These circuits use a bandpass filter approach, rather than two separate filters.

Figure 15 represents a π -network approach, while the version in Fig. 16 is an L-network. In both cases,

$$Q = \frac{f_o}{BW_{3dB}}$$

and,

$$\omega = 2\pi f_o$$

Fig. 10. Mini-Circuits SRA/SBL passive double-balanced mixer. Its characteristics are shown in Table 1.

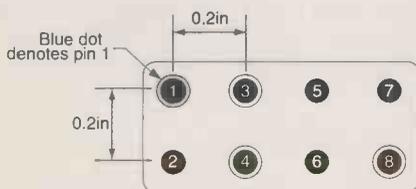
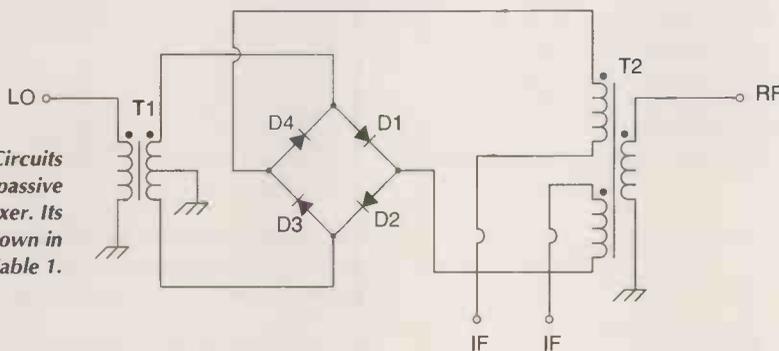


Fig. 11. Package details of the double-balanced mixer in Fig. 10.

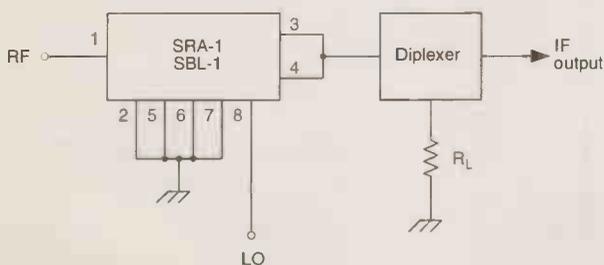


Fig. 12. Terminating a passive double-balanced mixer with a diplexer.

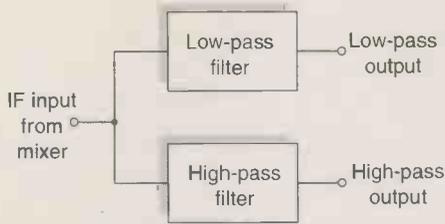


Fig. 13. A passive diplexer following the mixer provides selectivity while looking like a constant resistive impedance at its input terminals.

where f_0 is the centre frequency of the passband in hertz, BW_{3dB} is the desired bandwidth in hertz and Q is the relative bandwidth.

For the circuit of Fig. 15,

$$L_2 = \frac{R_0 Q}{\omega}$$

$$L_1 = \frac{R_0}{\omega Q}$$

$$C_2 = \frac{1}{R_0 Q \omega}$$

$$C_1 = \frac{Q}{\omega R_0}$$

And for the circuit of Fig. 16,

$$L_2 = \frac{R_0 Q}{\omega}$$

$$L_1 = \frac{R_0}{\omega Q}$$

$$C_1 = \frac{1}{L_1 \omega^2}$$

$$C_2 = \frac{1}{L_2 \omega^2}$$

A dual double-balanced mixer

The normal passive diode mixer provides relatively high third-order intercept and -1dB compression points. It also provides a high degree of port-to-port isolation. Because of the switching action of the diodes in the ring, they are shut off at the instances where they would feed through the other ports.

But where an even higher degree of performance is needed, designers sometimes opt for the dual double-balanced mixer as shown in Fig. 17. The -1dB compression point is usually ≤ 4 dB.

Image-reject mixers

In cases where very good image rejection performance is needed in a receiver, a circuit such as Fig. 18 can be used. This circuit uses a pair of passive double-balanced mixers, a 0° power splitter and two 90° power splitters to form an image reject mixer.

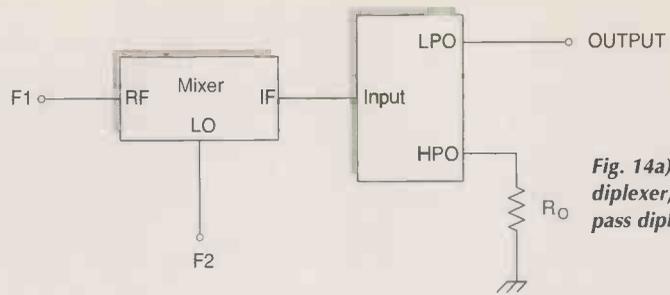
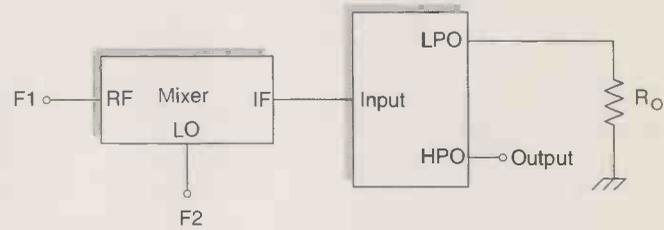


Fig. 14a) Low-pass diplexer; b) high-pass diplexer.



The local oscillator ports of mixer-1 and mixer-2 are driven in-phase from a master local-oscillator source. The RF input, however, is divided into quadrature signals and applied to the respective RF inputs of the two mixers.

The IF outputs of the mixers are then recombined in another quadrature splitter, to form separate USB and LSB IF outputs.

VHF/UHF microwave mixers

When the frequencies used for LO, RF and IF begin to reach into the VHF and above region, the design approaches change a bit.

Figure 19 shows a simple single diode unbalanced mixer. Variants of this circuit have been used in UHF television and other types of receivers. The circuit is enclosed in a shielded space in which a strip inductor L_1 and variable capacitor C_1 form a resonant circuit.

The LO and RF signals are applied to the mixer through coupling loops to L_1 . A UHF signal diode is connected to L_1 at a point that matches its impedance.

External to the mixer chamber, an IF filter is used to select the mixer prod-

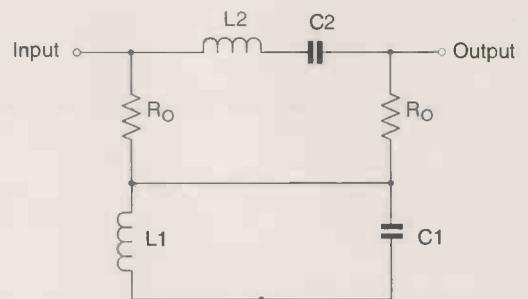


Fig. 15. Pi-filter bandpass diplexer.

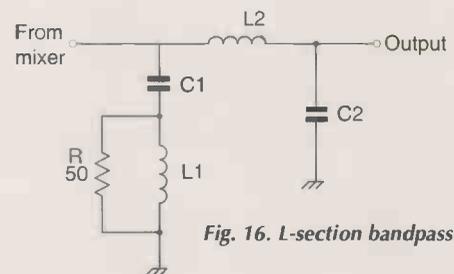


Fig. 16. L-section bandpass diplexer.

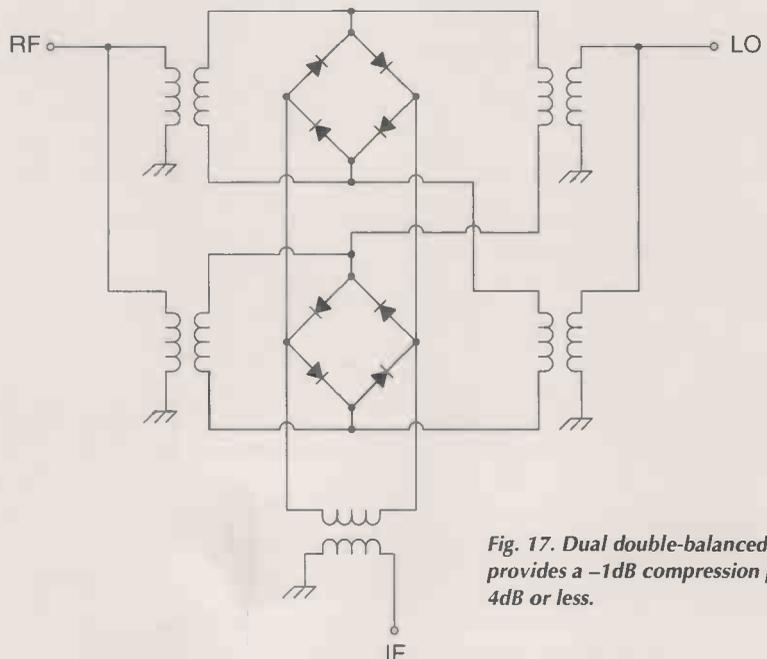


Fig. 17. Dual double-balanced mixer provides a -1dB compression point of 4dB or less.

uct desired for the IF section of the particular receiver.

A single-balanced mixer is shown in Fig. 20. This mixer circuit uses two diodes, D_1 and D_2 , connected together, and also to the ends of a half-wavelength 100Ω transmission line. The LO signal is applied to D_2 and one end of the transmission line. IF and RF filters

are used to couple to the RF and IF ports.

These mixers suffer from RF and LO components appearing in the output. Figure 21 shows an improved version that will solve the problem. It is used in the UHF and microwave regions.

The LO and RF input signals are applied to two separate ports of a

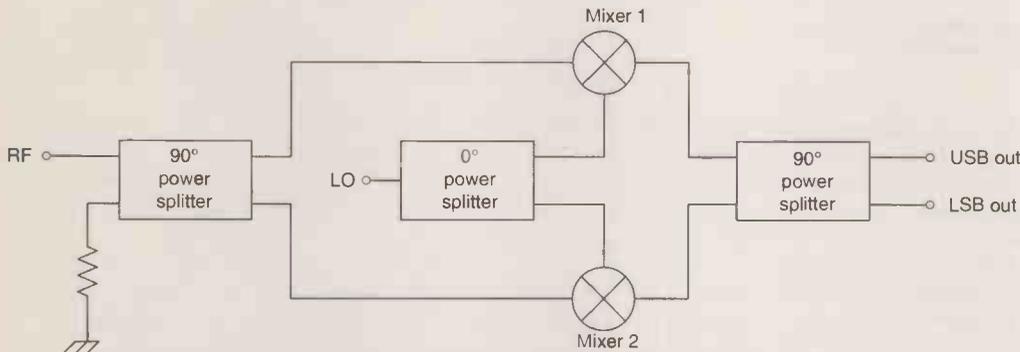


Fig. 18. Image-rejection is improved by using two passive double-balanced mixers together with power splitters.

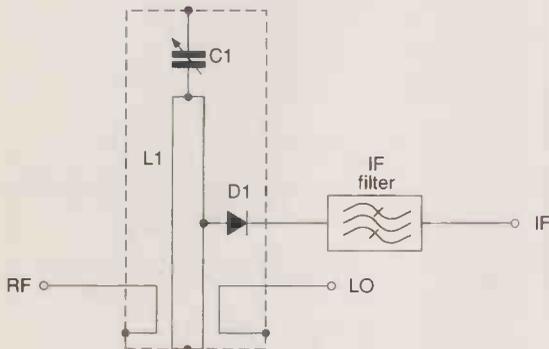


Fig. 19. Simple single-ended diode mixer for VHF/UHF.

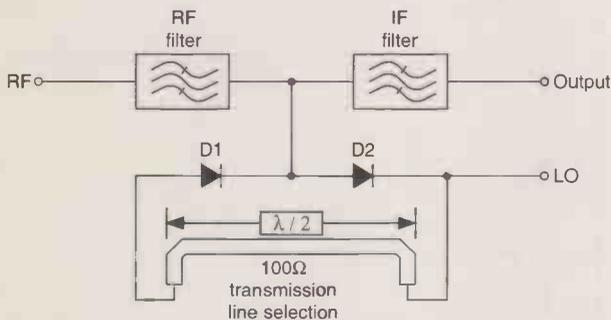


Fig. 20. Such UHF single-balanced mixers suffer from RF and LO components appearing in the output.

quadrature hybrid coupler. The input and output filtering are made using printed-circuit-board transmission lines.

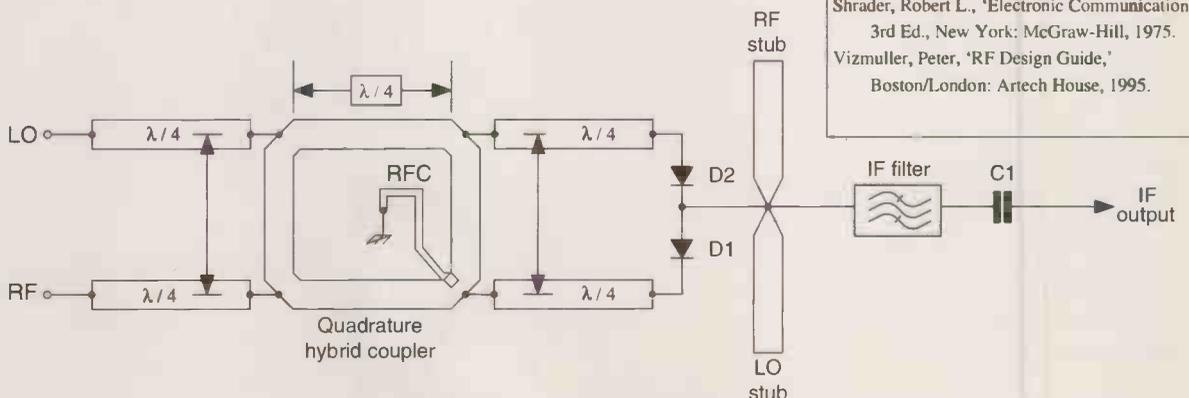
Each of the lines is a quarter wavelength, although the actual physical lengths must be shortened by the velocity factor of the printed circuit board being used. A printed circuit RF choke is used to provide a return connection for the diodes.

Note the RF and LO stubs at the output of the mixer, prior to the input of the IF filter. These stubs are used to suppress RL and LO components that pass through the mixer.

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Fig. 21. Microwave mixer is an improvement over the single-balanced mixer of Fig. 20, significantly reducing RF and LO components in the output.



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Power amplifier

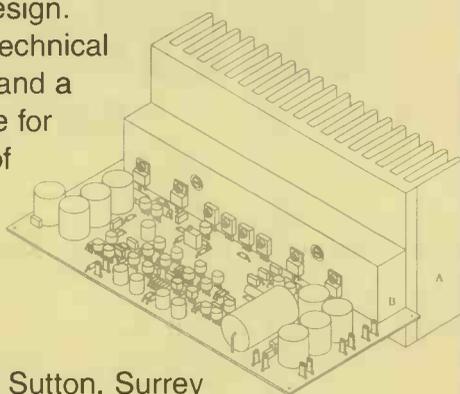
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Positive slew-rate	+320V/µs
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How

Inductance is the essential principle in applications as diverse as 50Hz power transformers, metal detectors and tuned circuits in mobile phones. But how many electronics engineers are truly conversant with it? Ian Hickman looks at inductance from the viewpoint of the practising rf engineer.

long is L?

Over the years, the experienced design engineer develops a feel for the right value of a component in any circuit — at least to a first-order guess. In the case of inductors and capacitors, this guess is based on knowledge of the reactance of any particular component value at any given frequency.

My introduction to this practical aspect of electronic engineering was as a sandwich student in GEC's Central Research Labs at Wembley, which has long since disappeared. Someone was developing a rather sensitive circuit, and was troubled by 100Hz hum. Supply-line decoupling with a reasonably large electrolytic capacitor reduced it, but not enough.

A colleague jumped to his aid, brandishing a 1000 μ F electrolytic, exclaiming that if this looked like 160 Ω at one cycle per second it should be pretty effective at 100 cycles. Hertz hadn't been invented yet.

The figure stuck in my then impressionable memory, and was subsequently embroidered upon. So: it follows that 1 μ F has a reactance of 160 Ω or so at 1kHz, and a CR lowpass or high-pass combination of 1M Ω and 1 μ F has a time constant of 1s. Hence it

will be 3dB down at 0.159Hz, while 1nF and 1k Ω will be 3dB down at about 160kHz.

And any budding rf engineer working with mobile phone technology would do well to remember that at 1GHz, a 1pF capacitor has a reactance of just 160 Ω .

About inductance...

It is also handy to have a feel for the reactance of inductors of course — even though they are not so readily available in a wide range of close tolerance values, unlike resistors and capacitors.

Some years after graduation, I was working at a different firm, small but well known in the rf field. It produced, among other things, hybrids, baluns, attenuators and rf bridges. There, I soon developed a feel for inductance, similar to the feel I already had for capacitance. Just as the odd picofarad of strays can be disastrous at vhf and even more so at uhf, likewise too many nanohenrys in the wrong place can be a real headache.

To dimension the problem in practical terms, I experimented with an rf bridge, to get a feel for the inductance of an inch or so of wire.

Finding myself now, thirty years

later, working on L Band equipment, the odd bit of inductance here or there is of real consequence. So, having access to a network analyser the likes of which could only grace my home laboratory in my dreams, I decided to check up on my earlier measurements. These have led to a convenient *aide-mémoire*, similar to the 'reactance of a capacitance at 1Hz' figure mentioned earlier.

Fixed inductance?

Capacitors come ready made. You can't add some more dielectric to increase the capacitance for example, even if you would want to.

Inductors are different. Frequently, you have to make your own inductor, either as a printed spiral on a pcb, or as a 'curly' — a coil which is either a self-supporting air-cored type, or wound on a former.

Using a former, the inductance can be increased by a ferromagnetic core or 'slug'. This increases the flux density within the coil. But the return path for the flux is still in air, so the resulting increase in inductance may only be in the region of 10% to 40% — even with a slug of high permeability material.

The important thing though is that the value of inductance is now readily adjustable, e.g. for tuning purposes, where the inductor forms part of a tuned circuit. It may also mean that the required inductance can be achieved with a fewer turns, reducing the coil's 'copper loss'. If the reduction of copper loss outweighs the 'iron loss' due to the slug, the Q of the coil may even be increased – an additional benefit over and above tunability.

Where a larger inductance is needed in the available space, and/or screening of the inductor is required, a coil former complete with a ferrite or dust iron sleeve or pot can be used, in conjunction with the slug. Now, most of the flux path is in a material with a relative permeability μ_r of anything in the range 2 to 100 or more, so the inductance per turn will be greatly increased: or is it per turn squared?

Most of the flux path is within the core, yes, but not all. The design of the former and core will be such that, even with the slug set for maximum inductance, there is still an appreciable air gap in the flux path.

The air gap stabilises the inductance, at the expense of reducing it from the fully closed path condition. For the permeability of ferromagnetic materials is likely to vary somewhat with selection, temperature and life, especially if the winding of the inductor is carrying any dc.

A_L and all that

The case with hybrids, baluns and untuned rf transformers of all sorts is different. If there is no dc in the windings, then a fully closed flux path can usefully be used, and is always preferred. Thus, the necessary magnetising inductance for the primary can be achieved with the minimum number of turns, keeping down the copper loss.

With a fully-closed flux path in high permeability material, virtually all the flux will be contained within the core, with very little 'leakage inductance'. Apart from the question of screening, leakage inductance, which is associated with flux in air rather than in the core, is of no consequence as long as the flux links both primary and secondary.

The term leakage inductance is therefore usually reserved for inductance due to flux that links one wind-

ing but not the other. This adds a reactive component in the primary circuit, in series with the transformed secondary load, reducing the effectiveness of the transformer, especially at higher frequencies.

The inductance of two 'close-coupled' coils of inductance L_1 and L_2 connected in series and wound in the same sense is L_1+L_2+2M . Here, the mutual inductance $M=\sqrt{L_1L_2}$ and the assumption is made that there is no leakage, all the flux in each coil linking with all the turns of the other.

The reason for the $2M$ is that M is the inductance due to the flux of coil L_1 which links L_2 , to which must be added as much again due to the flux of coil L_2 which links L_1 . So if the two 'coils' consists of two identical turns on the same core, then if the inductance of each is L nanohenries while that of the two together is $4L$ nanohenries. This is because the coils are closely coupled, so in this instance, $L_1=L_2=M$.

If there are three turns, the inductance will be $3L$ plus $6M$ or $9L$ all told, since the flux of each turn links with both of the others. So extending the argument, the inductance of N turns on the said core will be N^2L nanohenries.

I have illustrated this graphically in Fig. 1 for an eight-turn winding. Each dot represents an inductance L – or equally well, M . Because of this result, you will find the inductance A_L of, for example, a two-hole balun core quoted as the inductance of a single turn, i.e. as so many nH/turns².

A challenge

Prove that $n^2=n+nP_2$. This sounds like

one of those arid academic exercises from Chapter N of an old maths textbook. But there is an application, right here. In Fig. 1, the left-hand column represents $8L$, due to the eight individual turns.

The seven dots in the left-hand column of the lower triangle represent the inductance due to the flux from turn 1, linking with turn 2, with turn 3, and so on to turn 8.

Likewise, the next column represents $6L$ – or strictly speaking $6M$, which is the same thing here. This is due to the flux in turn 2 linking with turns 3 to 8, and so on, totalling in all $28L$. But the total due to mutual inductance is twice this, since turn 2 links with turn 1, as well as turn 1 with turn 2.

This is represented by the upper inverted triangle. At a glance, you can see from the diagram that the grand total is $64L$.

It is also clear from Fig. 1 that $2(7+6+\dots+1)=56$ – the block representing the two triangles – is equal to $7 \times 8 = T$ where T is the total. Now $T=(8 \times 7 \times 6 \times 5 \dots \times 1)/(6 \times 5 \dots \times 1)$, which is conventionally denoted by the shorthand ${}_n P_2$, where P stands for permutation. Add the column on the left, and QED.

There's more on permutation in the separate panel.

How long is a piece of inductance?

So what has all this to do with the inductance of a length of wire? Well, it turns out that the inductance of a length of wire depends on just how you measure it. The simplest way is as a single air-cored turn.

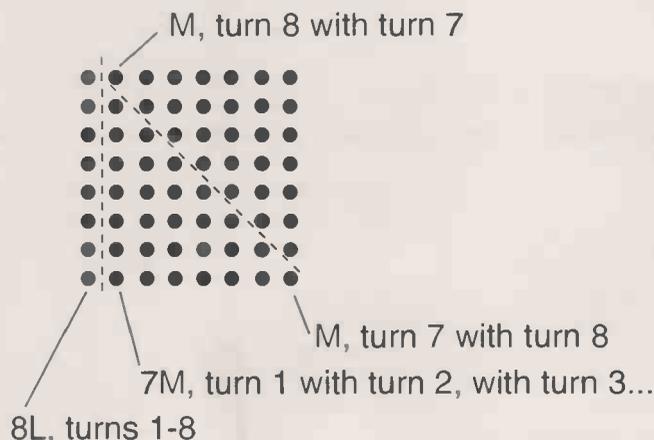


Fig. 1. Illustrating how the inductance of an eight-turn winding on a high permeability core is 64 times that of a single turn.

The fact is, you can't measure the inductance of an isolated straight piece of wire. For the phenomenon of

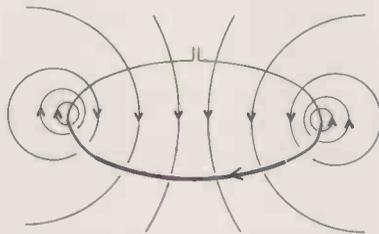


Fig. 2. Showing the lines of flux surrounding a current carrying wire.

inductance to manifest itself, a current must flow. The current can only flow in a complete circuit, so even if you make a jig to hold a length of straight wire under test, what you actually measure is the inductance of the wire plus that of the jig.

Figure 2 shows a length of wire W centimetres carrying a current, and the resultant lines of magnetic flux. Some of these are only shown in part, as the ones passing through the turn near the centre are very long. The flux is naturally more concentrated near the wire, where the flux paths are shorter.

Perms and coms

Permutations and combinations are conveniently expressed by means of factorials. Factorial N , written $N!$, means $1 \times 2 \times 3 \dots \times (N-1) \times N$. So factorials 4! is $1 \times 2 \times 3 \times 4 = 24$.

The permutation ${}_n P_r$ represents the number of ways of selecting r things from a group of n , the order being significant. So imagine for instance selecting four ornaments to range on the mantel-piece, from a choice of nine. For the item on the left there are nine possible choices. For the next, you have eight possibilities, seven for the third and six for the fourth.

The total number of ways therefore is 3024. This is equal to $n!/(n-r)!$, where here, $n=9$ and $r=4$.

The combination ${}_n C_r$ represents the number of ways of selecting r things from a group of n , the order not being important. So for instance the number of ways of selecting four apples to put in your shopping bag from nine on the greengrocer's shelf is smaller than ${}_9 P_4$.

As there are 4! ways of arranging the 4 apples, ${}_n P_r$ is simply too large an answer, by a factor of 4! So ${}_n C_r = {}_n P_r / r! = n! / \{(n-r)! \times r!\}$

I'm told that a perm on the football pools is not a perm, it's actually a combination, but never having done the pools, I wouldn't know.

Measuring a length of wire

What should the inductance be? In Fig. 2, if the length of the wire forming the loop were halved, the area of the loop through which all the flux must pass is reduced to a quarter of the previous figure. So it looks as though the reluctance S of the magnetic circuit – the equivalent of resistance R in an electric circuit – must be quadrupled.

Now inductance equals N^2/S , or here, with just a single turn, equals $1/S$. So the inductance of a length of wire, as a single-turn loop, would appear to be proportional the square of the length. This is a somewhat surprising statement.

On the other hand, if a single loop of inductance L nanohenrys were formed into two tight-coupled turns of half the circumference, S being quadrupled, each would then have an inductance of $L/4$.

By the earlier reasoning, the two together would have an inductance of $4(L/4)$. But would the inductance really be the same for one large or two small turns? Some careful mathematical analysis apart, there is only one way to find out.

Figure 3 shows an SMA connector terminating in an open spill, fitted

Puzzle corner

If you followed the argument about $n^2 = n + {}_n P_2$ in the main article, you might like to try proving that $n^3 = 3n^2 + {}_n P_3 - 2n$. If you need a hint, imagine layers like Fig. 1 stacked up on top of each other.

Is there a recurrence formula that will enable the value of n^{m+1} in terms of perms to be deduced from that of n^m ?

with a solder tag, connected to the coil. The SMA plug was connected to a 'between-series-adaptor' which picked up on the APC7 connector on the network analyser's front panel, as indicated.

Before connecting the coil, the test frequency of the HP8753D network analyser was set to 5MHz, and the measurement plane calibrated as the spill and solder tag used as the test terminals.

A 20cm length of wire was formed into a circle and connected to the terminals. The wire was 0.23mm diameter, measured over the usual self-fluxing insulation; this means it was probably 36SWG.

The wire was actually bifilar, with two parallel untwisted strands glued together side by side along their length. Each strand has a different colour enamel. This type of wire is available from the better stockists, and is very convenient when winding baluns, Ruthroff type line transformers and the like.

Various connections were tried, the results being recorded in Table 1. The final measurement was with the test terminals shorted.

The outcome

The short is close to perfect – at least as far as the reactive term goes. It actually looks like a very large capacitance, of such low impedance as to be negligible relative to the measured inductance of the loop.

However, it measured as having 100mΩ in series with the small reactive impedance – although in principle the earlier measurement-plane calibration supposedly factored both terms out.

Consequently, in addition to the value of Q calculated for the coil from the measured R and X figures, a

Table 1. Frequency 5MHz. Bifilar wire. 36SWG. Length 10cm – plus small allowance for soldered connections.

Meas.	Conditions	Impedance (Ω)	Q	Q'	Inductance (nH)
1	Both strands in parallel	0.195+j 2.88	14.8	30	91.5
2	As 1, but one turn disconnected	0.265+j3.19	12.0	19	101.3
3	The two turns in series	0.455+j11.4	25.1	32	363.0
4	As 1, but connections reversed for one turn	0.185+j0.33	1.8	4	10.7
5	As 2, but disconnected turn removed	0.249+j3.26	13.1	22	103.6
6	2t in series as 1t of 2x dia. (20cm circ.)	0.385+j7.41	19.2	26	235.5
7	Terminals short circuited	0.100-j0.09	-	-	

corrected value is given in column Q' , with 0.1Ω subtracted from the measured value of R . Which figure is the more nearly correct is a moot point; it is probably somewhere in between.

Comparing row 5 with row 3 shows that two turns do not give four times the inductance of a single turn. This is because the shortest lines of flux surrounding each wire manage to sneak through the two thicknesses of insulation separating the turns.

This also happens when a high permeability core is present, as in a transformer. But the effect is then barely noticed, due to the much-enhanced flux linking both turns, resulting from the much greater flux density in the core. In fact, the leakage inductance is indicated by row 4, where the two turns are in parallel, but with the connections to one reversed.

With perfect coupling, the inductance would of course have been zero.

Was it the right figure?

The results agree reasonably well with my previous idea that an inch of wire looks like $25nH$; those of you brought up metricated might find $1nH/mm$ an easier figure to bear in mind.

At $0.915nH/mm$, the figure in the first row of Table 1 for two strands of 36SWG in parallel is about 10% lower. But the figures in the second and fifth rows are nearer. Comparing these results with row 3 shows that for an air-cored coil, lacking the flux concentration provided by a high permeability core, the inductance falls short of being proportional to N^2 .

As a check, I repeated the experiment with 28SWG wire. Firstly, I measured a loop of 20cm of 28 gauge EnCu wire, result $201nH$. This was then formed into two turns of exactly half the diameter, a couple of millimetres of extra wire at each end of the 20cm having been allowed for the soldered connections: result, $306nH$.

One of the two turns was then removed, the measurement returning $86nH$. Finally, as a sanity check, the terminals were measured shorted, this

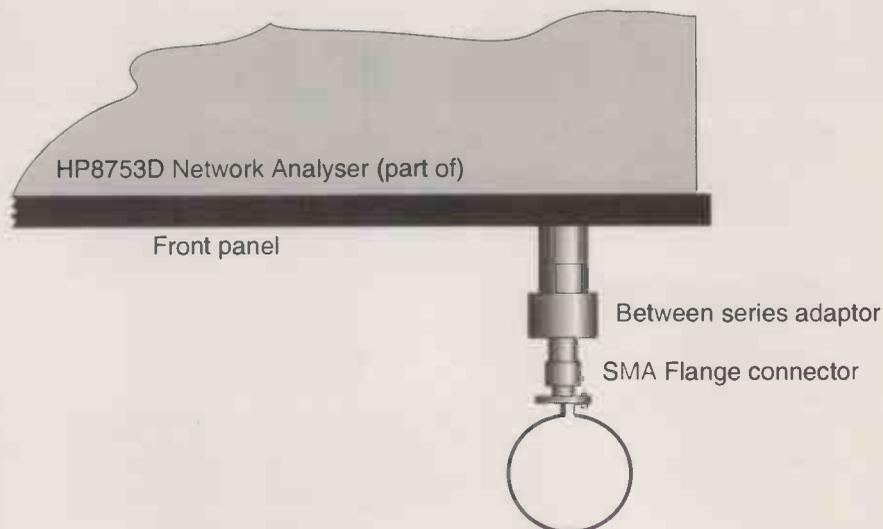


Fig. 3. The test set-up used to record the results in Tables 1 and 2.

time the answer being $2.5nH$. And for good measure, the inductance of 1 turn of 20cm circumference 16SWG wire was also determined. The results for the three gauges are compared in Table 2.

The results confirm the story told by Table 1. In particular, rows 5 and 6, like the figures $201nH$ and $86nH$ above, show that the inductance of a length of wire is certainly not proportional to the square of its length. Nor is it simply proportional to the length either, but somewhere in between.

In fact, the figures show that a single turn air-cored coil of twice the circumference will have about 2.33 times the inductance.

The final question

The other question I set out to answer was what happens to the inductance if the wire of a single turn coil is refashioned into a two-turn coil of half the diameter. Why did the inductance go up from $201nH$ to $306nH$ when the single turn was fashioned into two smaller ones – or $235.5nH$ to $363nH$ in the case of Table 1?

Well, it is true that the reluctance S of that part of the magnetic path where it passes through the loop is quadrupled due to the smaller area, but that does not apply to the path as

a whole. So the overall increase in reluctance is less than four times. The figures show that when a single turn is rewound as two tightly-coupled smaller turns, the inductance increases by 53%.

This is the result of two separate effects; the increase in reluctance of the magnetic circuit, and the leakage resulting in less than perfect coupling. The effect of the leakage is dramatically shown by the measured results for 1 and 2 turns of $86nH$ and $306nH$ (28SWG wire, above) or $91nH$ and $308nH$, Table 1.

So, if you are talking about 28SWG or thereabouts, my earlier mental note that an inch of wire equals $25nH$ equals 16Ω at 100MHz was not far out. They are of course reactive or $+j\Omega$, assuming the inch of wire has a reasonable Q , as it usually will.

Now 16Ω is only a rough estimate; it depends on the gauge of the wire, a thin wire having greater inductance per unit length than a thick one. In the mean time, I still reckon $25nH/inch$ is a good figure to bear in mind.

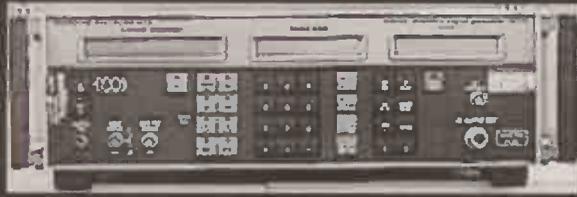
Although I do think that having a feel for the reactance per unit length at any given frequency is more useful in practical terms than the inductance per unit length. ■

Table 2. Frequency 5MHz, length 20cm, again plus small allowance for soldered connections.

SWG	36	28	16
Inductance, 1 turn	235.5nH	201nH	142nH

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Microstripline update

Using a UHF amplifier as a design example, Nick Wheeler explains how easy it is to apply microstripline.

My first article on this topic, 'Microstrip made easy,' described how most microstripline calculations refer to traces of width W on an infinite layer of dielectric whose dielectric constant is ϵ_r and whose thickness is h .¹ The dielectric is backed by an infinite ground plane.

In practice, the electric field tends to concentrate in the volume of dielectric lying just under the track. This tendency increases with increasing frequency.

Some of the field is in air though. This leads to the concept of the effective dielectric, ϵ_{eff} , which is lower than the that of the dielectric alone.

The greater the width of the track, the more ϵ_{eff} tends towards ϵ_r . While trace width is a strong determinant of impedance Z_0 , it also has to be taken into consideration where the desired track length is frequency dependent, as in $\lambda/4$ transformers.

I sacrificed a sheet of double-sided photo-etch circuit board by laying down a set of traces ranging in width from 0.007in (0.5pt) to 0.5in (36pt). For such traces, Z_0 can be determined from Fig. A, duplicated from my earlier article.

Resistive pads were constructed to match each trace to a 50 Ω source, the other ends being left open-circuited. λ_g was then determined for each line by probing for nodes, Fig. B.

Findings

The 'velocity ratio', V , is λ_g divided by the free-space wavelength, and for each trace ϵ_{eff} is determined by squaring $1/V$.

There are simple theoretical limits for ϵ_{eff} . When the line is infinitely narrow, ϵ_{eff} tends to $1/2(\epsilon_r+1)$, where ϵ_r is the dielectric constant of the substrate. In this case 4.5 is assumed.

When the line is infinitely wide ϵ_{eff} tends asymptotically to ϵ_r , since the

electric field is entirely within the substrate. The results obtained with the practical range of trace widths noted above are shown in Fig. 1.

Values taken from Fig. 1 are good enough for the production of artwork. In general, end effects tend to result in lines that are too long. They may need to be shortened by trial and error for best results. End effects include the parts attached to the lines at each end.

A UHF amplifier

Figure 2 is the circuit board artwork and circuit for the UHF amplifier. This is a broadband amplifier centred on 527MHz – roughly the centre frequency of my local television transmissions from Crystal Palace.

I have disregarded the Channel 5 transmission, since I can only receive it at a very low level. Other signals here are so strong that I do not need a preamplifier, but a television set is a good way of detecting impedance mismatch, which produces multiple reflections in the antenna lead. The circuit includes two microstriplines, the dimensioning of which I shall now describe.

The input circuit, L_1 , is a conventional $\lambda/4$ strip earthed at one end and tapped to match the low impedance of the input from the antenna. With a dual-gate FET such as the CF739,

which has a very high input impedance at G_1 , it can be fed directly from the 'high' end of the line. Its input capacitance of just over 1pF will slightly reduce the optimum line length, as will other stray capacitances.

A wide, rather than a narrow, trace is chosen as a resonator. A wide trace has a lower Q than a narrow one, making the input circuit broadband.

The free-space wavelength of 527MHz is 56.7cm. From Fig. 1 you

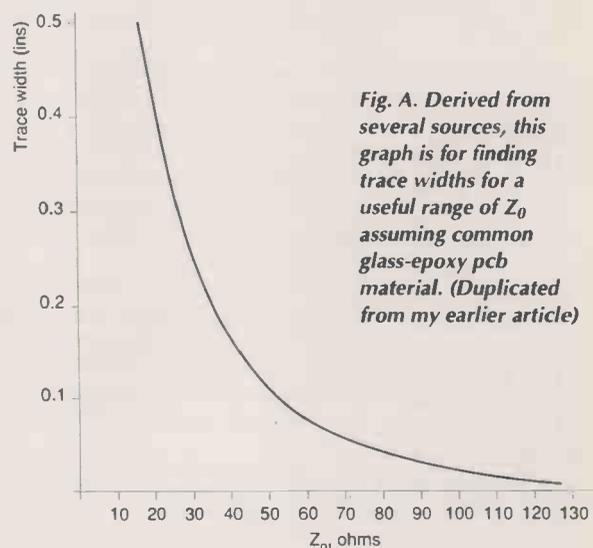


Fig. A. Derived from several sources, this graph is for finding trace widths for a useful range of Z_0 assuming common glass-epoxy pcb material. (Duplicated from my earlier article)

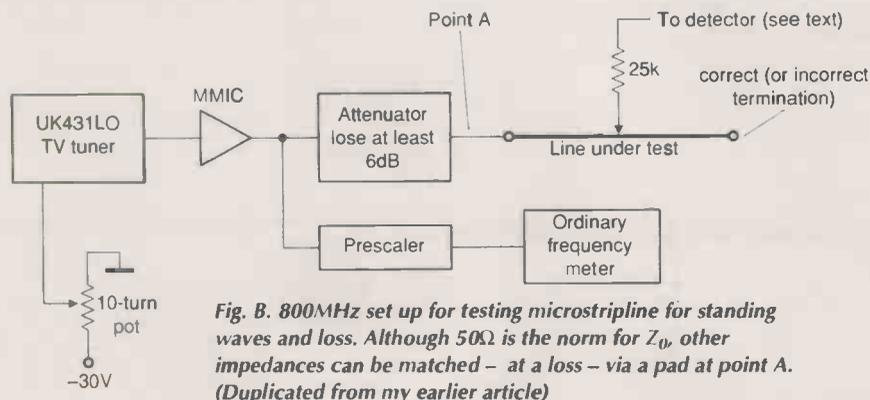


Fig. B. 800MHz set up for testing microstripline for standing waves and loss. Although 50 Ω is the norm for Z_0 , other impedances can be matched – at a loss – via a pad at point A. (Duplicated from my earlier article)

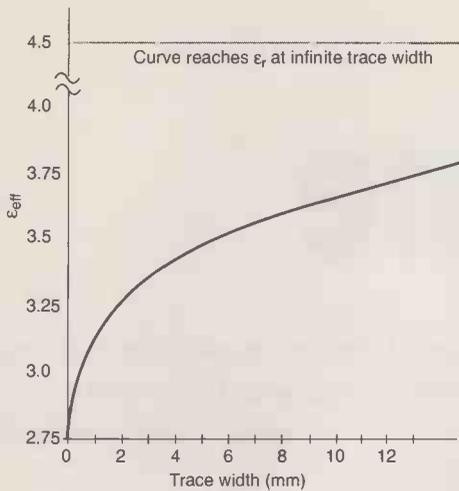


Fig. 1. A graph of effective dielectric value ϵ_{eff} versus trace width shows that the real dielectric value is different from that of the pcb substrate material ϵ_r . Here, ϵ_r is 4.5 and the substrate thickness is 1.6mm.

will see that for the chosen trace width of 2.8mm, ϵ_{eff} is 3.35. Hence λ_g is $56.7/(1/\sqrt{\epsilon_{eff}})$, or 30.98cm, making $\lambda_g/4=7.74$ cm.

As I pointed out earlier, the actual length required will be less than this. A relatively wide trace such as this can be terminated at the earthy end by drilling a 1mm hole through the centre line and earthing to the ground plane on the other side of the circuit board. Ordinary 'track pins' are suitable for this.

Start with the track too long and progressively shorten it, moving the input tap to be arbitrarily 1cm from the earth point.

Positioning is not too critical. I ended up with a track 70mm long. Resonance can be determined by the same means as was used for finding the nodes, though in this case we are looking for a voltage maximum, which cannot be done as accurately as finding a minimum.

Inductor L_2 is a $\lambda/4$ transformer, which matches the drain load of the the CF739 to the 75 Ω of the output to the TV set. Because the output impedance of a dual-gate FET is very high, the termination seen by signals reflected

back from the TV set is equal to the drain load, as transformed, for all practical purposes.

In this case the dominant consideration is the Z_0 of L_2 , since this equals $\sqrt{(R_{drain} \times 75)}$.

Using a graphic program such as *Serif Draw* it is convenient to choose trace widths offered in the program menus – particularly if there are to be bends. If a straight trace is acceptable, almost any trace width can be closely approached.

In this case, I chose a trace width of 2 points, which is around 0.7mm. Such a trace has Z_0 approximately equal to 92 Ω , whence R_{drain} has to be 113 Ω .

I used chip resistors of 150 Ω and 470 Ω in parallel.

With this trace width, ϵ_{eff} is close to the limiting minimum value of 2.75. From Fig. 1, I estimated it at 3.07, so $\lambda_g/4=80.9$ mm. Note that this is appreciably more than the equivalent at a track width of 2.8mm.

I used 8cm, and this proved to be satisfactory. The performance at the edges of the band, BBC2 and ITV was good, but not quite as good as in the middle. The Channel 5 signal was not improved.

This amplifier has a gain of 7-10dB over the local TV band, but it is not intended as a serious contender in this field. It is described to illustrate the practical application of microstripline. The design also illustrates how microstripline complements SMD technology.

Experiments

I also tried to use a $\lambda/4$ transformer with a 0.35mm track. This worked well for one TV channel, but at the expense of the others.

The higher transformation ratio permits a larger drain load of 220 Ω in parallel with 1.5k Ω , giving more gain. This is what you would expect. Broadbanding is always at the expense of gain.

The tracks were made straight for ease of experimentation, but provided they are curved on an adequate radius can be folded to give a more compact board layout

In summary

To get good results from a microstripline circuit board, practice and experimentation are necessary.

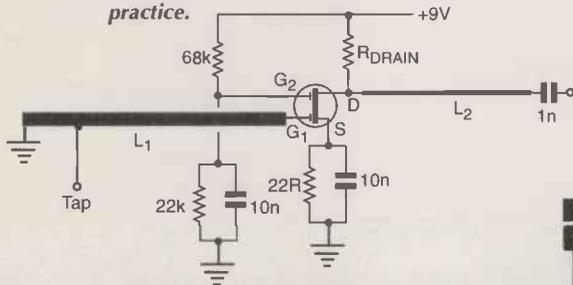
This brief article, read in conjunction my earlier one, sets out most of what you need to know. But because of the variables involved – not the least of which is the fact that circuit board obtained from ordinary sources is not tightly specified as regards ϵ_r – your first attempt may not work as well as you might expect.

I have rather taken to adding a few test tracks to each artwork print so that I can evaluate ϵ_{eff} for the material being used. ■

Reference

- 1. Wheeler, N., 'Microstrip made easy,' *Electronics World*, December 1997.

Fig. 2. UHF amplifier using a CF739 dual-gate mosfet and its circuit board artwork. Note that all component grounds shown on the circuit diagram go through to the ground plane on the reverse side of the pcb in practice.



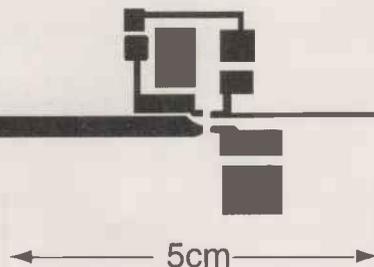
Drawing tracks to 0.001in accuracy with Serif

I use a general-purpose drawing package called *Serif Draw* for laying out circuit boards. Its menus allow a useful variety of line weights to be selected in point steps. A point is a typesetting measure equal to 1/72in by the way.

With patience though, it is possible to draw lines of any width to an accuracy of ± 0.001 in. At a magnification of 1000%, the smallest divisions on the vertical and horizontal ruler scales are a sixteenth of 0.1in, or 0.00625in.

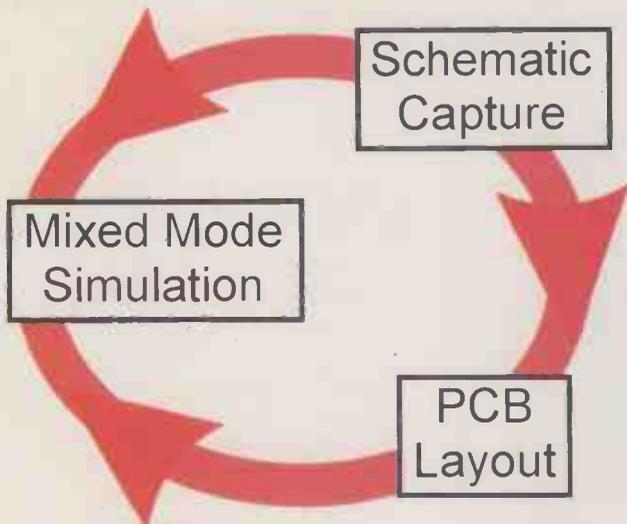
You should find it possible to draw 13 'hairline' traces, separately identifiable, in the space of two of these small divisions. Abutting a hairline trace to another trace which is just too narrow allows the narrow to be widened in steps of 0.001in. This enables artwork to be created for microstripline to greater accuracy than is ever likely to be needed.

These lines can only be vertical or horizontal, since sloping or curved lines consist of a series of steps.



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Spice-derived distortion figures for mosfet power audio output stages are notoriously inaccurate. They can be as much as 63% out. Ian Hegglun's breakthrough involves four new mosfet models that allow you to produce realistic Spice distortion figures and data sheet curves.

Hotter Spice

Spice – an acronym for 'simulator program with IC emphasis' – was written in the seventies before power mosfets were developed. The Spice level 1 mosfet model, also known as the Schickman and Hodges model,¹ is present in all simulators but the more accurate level 3 mosfet model² is not.

The level 1 model is a simple model that runs fast and presents few problems with numerical convergence. It was originally intended for analogue simulation of switching circuits. The level 3 model is better suited for modelling linear circuits, although it is slower and has some annoying convergence problems in class AB circuits when several feedback loops are used.

Better mosfet models have been developed since then, but most of these are for ICs (BSIM models) or for microwave jfet devices³ but are only found in later and more expensive Spice 3 versions.

Spice's macromodel, or 'subcircuit,' technique, allows users to add new models to the standard library. Because macro-

models are made from predefined components it can be difficult to model some functions — particularly if your simulator lacks a multiplier. Consequently, I have developed several maths macromodels for use in power-mosfet macromodels.

A useful historical perspective over the last three decades of Spice can be found on the Web⁴ or from ref. 5.

These mosfet models are accurate enough for distortion simulations. In addition, they can be adapted to run on almost any version of Spice because they use only standard Spice components.

Simulating distortion

The Macquarie University team at Sydney showed class AB distortion using the Spice level 3 model gave a distortion result that was a third of the measured thd – or an error of 63%, see ref. 6. When a more accurate model was used, the distortion was within 20% of measured values.

For class AB, the level of crossover distort-

tion generated is sensitive to the bias setting. Subthreshold conduction is the conduction region where significant conduction begins and therefore plays a part in crossover distortion generation.⁷

Subthreshold conduction is not modelled in the level 1 mosfet. Since 1990, the level 3 parameter NFS has been included for subthreshold conduction, but not all simulators since then include the updated model.

In the following macromodels, several missing effects are added; namely subthreshold conduction, velocity saturation, a smooth pinchoff and electrothermal effects. Electrothermal feedback is present in practical power amplifiers where junction temperature change during a cycle affects linearity and hence distortion.

You may find that some of these effects are not essential for your application. But to discover whether they are needed or not, a simulation with and without these effects is usually necessary. The following four macromodels

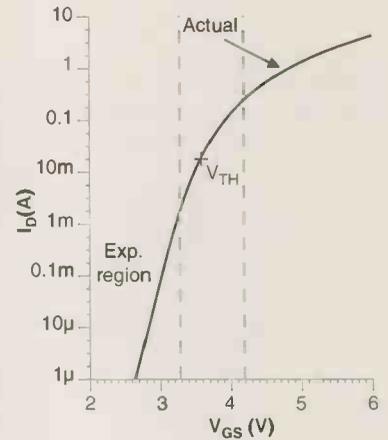
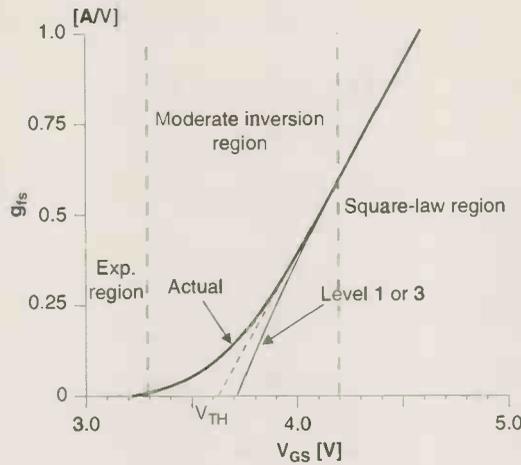


Fig. 1a) A g_{fs} transconductance plot for a power mosfet and Spice models without subthreshold conduction.

Fig. 1b) Logarithmic current plot for the same mosfet.

also show various techniques for those who would like to develop their own macromodels.

Subthreshold conduction

On a transconductance plot with gate-to-

Simple curve tracer

A simple curve tracer can be made with a scope, signal generator, power supply and a few other discrete components. We have found this useful for demonstrating the various mosfet operating regions. Approximate values for V_{TH} and β_{eff} can be obtained from the g_{fs}/V_{GS} curve.

The circuit enables the I_D/V_{GS} curve and the g_{fs}/V_{GS} curve to be generated together. The value of V_{TH} can be

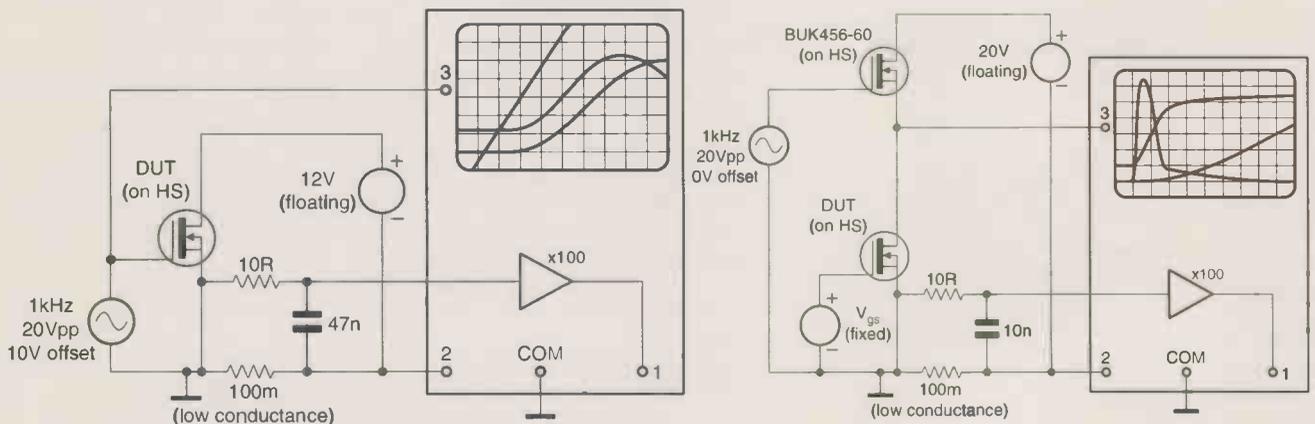
found from the g_{fs} curve x-intercept and β from the slope.

To reduce self-heating, the sweep duty cycle is kept low, at around 0.15, by offsetting the signal generator. An 18V gate zener with a series resistor is a good idea to prevent the signal generator plus offset value exceeding the gate breakdown voltage.

The peak seen in the g_{fs} curve is caused by the mosfet entering the resistance region. Here, V_{DS} is only 5V to minimise heating effects. When V_{DS}

is much higher and the pulse time short the g_{fs} curve levels off giving a linear I_D/V_{GS} curve.

The second circuit gives the I_D/V_{DS} curve and g_{DS}/V_{DS} curve. The value of λ can be calculated from g_{DS}/I_D where g_{DS} is the drain conductance beyond the knee region. In this rig the drain conductance incorrectly peaks at low V_{DS} because the V_{DS} ramp is not maintained below 1V. A power amplifier drive for V_{DS} would overcome this problem.



The left-hand circuit allows an I_D/V_{GS} curve and g_{fs}/V_{GS} to be generated simultaneously. The curves shown are for an ECF16N20 lateral power mosfet for V_{DS} of only 5V indicating $V_{TH}=0.3V$ and $\beta=2A/V^2$. Scope x-axis represents V_{GS} with 0.5V/div. For the y-axis $I_D \sim 2A/div$ and $g_{fs} \sim 1.3S/div$.

The right-hand circuit generates an I_D/V_{DS} curve and g_{DS}/V_{DS} curve. For the ECF16N20 the value of λ is around $0.01V^{-1}$. The knee in g_{DS} at pinch-off continues to fall until $V_{DS} \sim 7V$ at which point g_{DS} is around 20mS. Here $V_{GS}=2.3V$, $R_s=0.1\Omega$, $R_{diff}=100\Omega$, $C=10nF$, $f=1kHz$, time base 50µs/div giving x-axis $V_{DS} \sim 2V/div$, and y-axis $I_D \sim 0.5A/div$ and $g_{DS} \sim 75mS/div$.

Fig. 2a). A macromodel for adding subthreshold conduction and additional circuitry for analysing the macromodel.

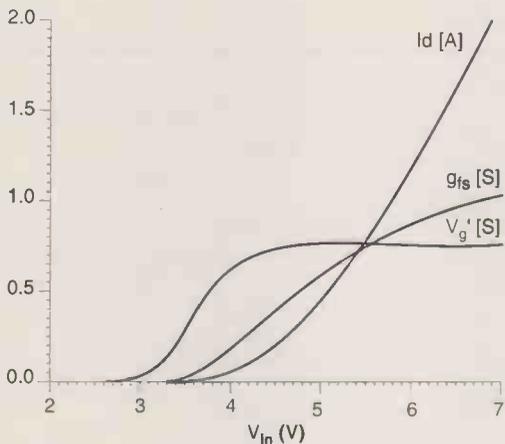
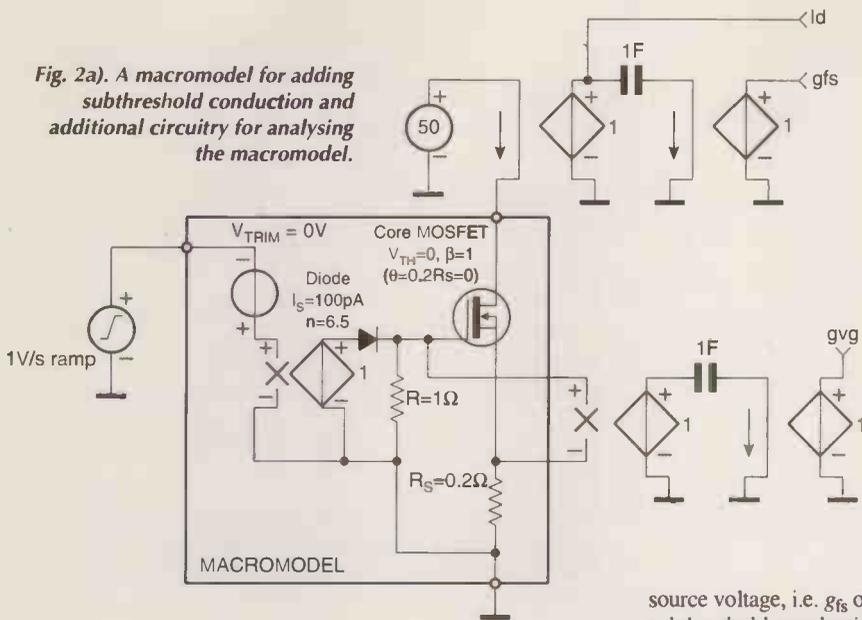
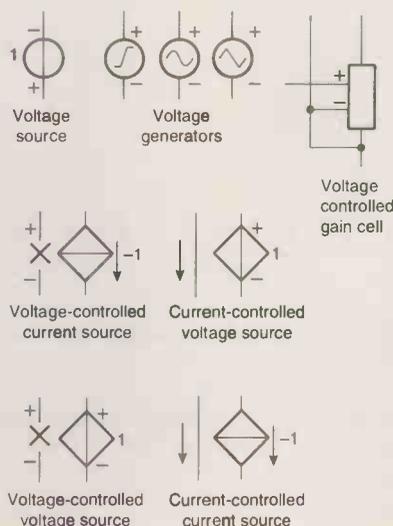


Fig. 2b). The resulting I_D curve, g_{fs} curve and the diode network transfer function V_g' .

Model symbols used in the Windows-based circuit simulation tool, Tina, distributed in the UK by Quickroute (does valves too, Ed).



source voltage, i.e. g_{fs} or g_m versus V_{GS} , subthreshold conduction is seen as a knee region near the V_{TH} threshold voltage Fig. 1a).

On a log-linear plot, Fig. 1b), the lower end of the knee region appears as a straight line indicating an exponential relationship with V_{GS} . Approaching V_{TH} , the relationship changes to a square law for a limited range of gate voltage. There's more on this in the panel entitled 'Simple curve tracer.'

In the mosfet model world, the threshold voltage is defined as the x-intercept of a projection back from the square law region, Fig. 1a).

The initial slope of the g_{fs} curve in the square law region is set by the value of β or K_p for the level 1. For the level 3, $\beta = K_p W/L$ and $K_p = \mu C_{OX}$ where μ is carrier mobility and C_{OX} is the dielectric capacitance per square metre of gate insulation. These process parameters may be ignored by setting $W=L=2\mu m$ so $\beta = K_p$.

The transition region between the exponential and square law regions is the moderate inversion region. At higher currents, g_{fs} curve levels off giving a reasonably linear region with V_{GS} as in Fig. 1b) once the gate voltage rises some 2V more than the threshold voltage.⁸

Throughout this article upper case subscripts denote the voltages measured at the mosfet's terminals and lower case subscripts are for voltages at the mosfet channel. It is necessary to make this distinction because the effect of ohmic volt drops at the drain and source are usually significant in power mosfets.

Adding subthreshold conduction Figure 2a) shows a subcircuit or

macromodel that can add subthreshold conduction to the core mosfet. It uses the exponential nature of a diode for the region below the threshold voltage. With a series resistor the moderate inversion knee region in Fig. 1a) can be produced, closely resembling the behaviour of an actual power mosfet.

Series resistor R determines the threshold voltage while the shape of the knee region is determined by the diode emission coefficient parameter, n . To help use this macromodel an empirical method for determining parameter values has been developed.

Parameter setting

Several simulations are performed in transient mode to generate the g_{fs} curve, Fig. 2. An input voltage source for V_{in} provides a voltage ramp that effectively converts the x-axis to voltage as seen in Fig. 2b).

The diode circuit is terminated at the macromodel's source terminal for convenience. Unexpectedly, it can be alternatively terminated at the core mosfet's source terminal with little difference in the shape of the V_g' transfer function.

Subthreshold conduction reduces the effective transconductance in the square law region, hence the effective values for V_{th} , β and θ are reduced, requiring higher values in the core mosfet. Typically β must be set about 50% higher in this macromodel. The trial and error method for parameter setting can be used but β and θ parameters interact making this difficult. An empirical method based on several simulations is much easier.

Before simulations can be performed, a value for diode emission coefficient, n in the above macromodel, needs to be determined. It can be found either from data sheets or via simple measurement of drain current below the threshold region. The slope of the $(\log I_D)/V_{GS}$ graph gives n using,

$$n = \frac{2}{2.3 \times \text{slope} \times V_T}$$

where $V_T = 25.86mV$ at $27^\circ C$ and slope is the slope of the $\log I_{DS}$ graph expressed in decades/volt. The factor 2.3 comes from the \log_{10} of e to convert decades to base e and the factor of two allows for the squaring effect of the mosfet.

A typical value of n for a vertical power mosfet such as the 5A/100V IRF510 is five decades/volt giving $n=6.5$. For a lateral audio power mosfet such as a 7A/140V 2SK134 it is about

15 decades/volt or $n=2.3V^{-1}$.

The effective or target velocity saturation parameter θ_{eff} can be determined from measured I_{Dsat}/V_{GS} points.⁸ The value for θ for direct entry into the level 3 mosfet can be calculated using

$$\theta \cong \theta_{eff}(A+B\theta_{eff})$$

where constants A and B are found from simulated points, Table 1. Constants A and B can then be found by plotting the ratio θ/θ_{eff} (y-axis) and θ_{eff} (x-axis), or use linear regression. The value for A is the y-intercept and B is the gradient.

The value of β for the core mosfet can be calculated from

$$\beta \cong C\beta_{eff}(1+D\theta_{eff})^2$$

Constants C and D are found from points in Table 1 by plotting $\sqrt{\beta/\beta_{eff}}$ with θ_{eff} . The resulting slope gives D, and C is found by squaring the y-intercept.

Table 2 shows the results for the 2SK134 lateral audio mosfet and the IRF510 vertical mosfet. The resulting effective values are within 10% of the target values. This accuracy should not be confused with modelling accuracy

since it is only the accuracy for converting the β and θ values from raw extracted values to values need in this macromodel.

The final threshold voltage can be set by finding V_{TH} from a simulation with $R=1\Omega$ and $V_{trim}=0$. Then set V_{trim} to the difference between the simulated and required values.

The value of R_s can be varied while θ is reduced according to $\theta_{new}=\theta-\beta R_s$. This allows the high current end to be altered while not affecting lower currents.

Since the value of n is reasonably independent of die size, constants A-D

Table 1. Simulations used to find a relationship between model settings and results.

θ	$n = 2.3$ (2SK134)				$n = 6$ (IRF510)			
	V_{TH}	β_{eff}	g_{fslim}	θ_{eff}	V_{TH}	β_{eff}	g_{fslim}	θ_{eff}
0	1.50	0.998	none	0	4.20	1.0	none	0
0.1	1.29	0.770	4.84	0.080	3.74	0.650	4.82	0.067
0.2	1.26	0.694	2.47	0.140	3.66	0.552	2.47	0.112
0.3	1.25	0.643	1.75	0.184	3.64	0.489	1.65	0.148
0.4	1.23	0.603	1.24	0.243	3.60	0.443	1.24	0.179
0.5	1.23	0.568	1.00	0.284				
0.6	1.23	0.540	0.83	0.325				

Note: β is fixed at unity, $n=6$, $I_s=100pA$ and R is 1Ω . The term $\theta_{eff}=\beta/2g_{sim}$ is derived from the level 3 (Crawford) equation, where g_{sim} is the simulated gain limit with large V_{GS} such as 30V — provided V_{DS} is kept larger than V_{GS} to ensure current saturation. The value of β can be found accurately by differentiating g_s and reading the peak value.

Table 2. Values for the 2SK134 lateral audio power mosfet and the IRF510 vertical mosfet.

	n	A	B	C	D	β_{target}	β_{eff}	θ_{target}	θ_{eff}	β	θ
2SK134	2.3	1.0	3.3	1.08	1.0	0.636	0.628	300m	332m	1.15	595m
IRF510	6.5	1.1	6.0	1.15	2.15	1.30	1.26	90m	91m	2.13	151m

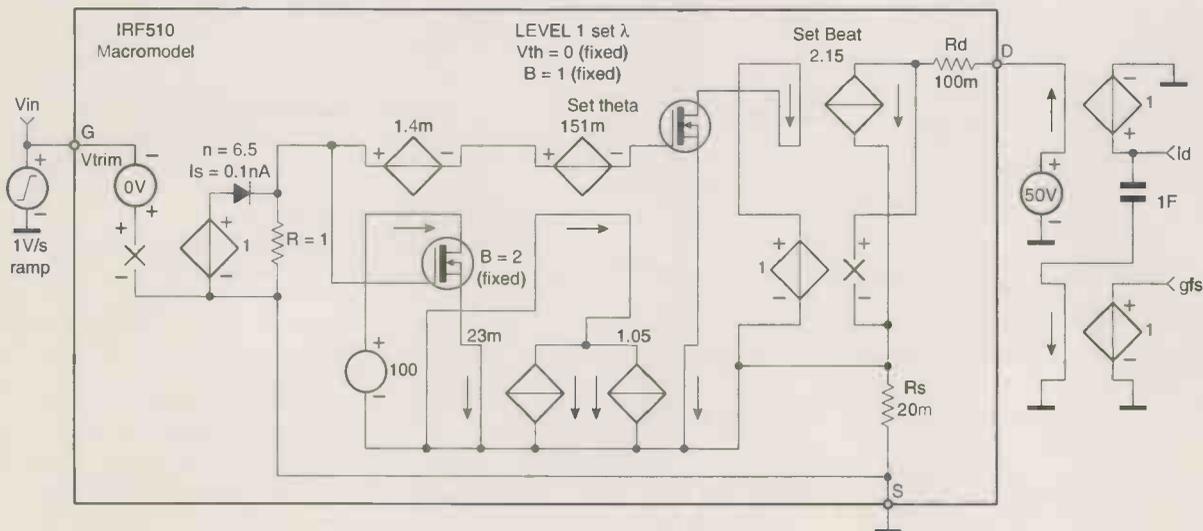


Fig. 3. A macromodel for adding velocity saturation and subthreshold conduction. Only level 1 mosfets are used here to ensure convergence in difficult circuits, such as symmetric class AB.

above can be applied to other mosfets of the same family even where n differs by up to 20%.

Adding velocity saturation

Although the level 3 model includes the velocity saturation parameter θ , it has some unfortunate convergence problems. Also, for Spice 2 simulators, it is reported that all mosfets should be of the same level.¹

The level 1 model is quite reliable but does not include the velocity saturation parameter θ . Instead additional source resistance is usually used to emulate the effect of velocity saturation. The total value for R_s can be calculated using $R_s \approx 0.8 (\theta_{eff}/\beta_{eff})$.

The factor 0.8 minimises modeling error up to the devices current rating, giving an averaged error of around $\pm 5\%$. The error increases rapidly at high currents and makes distortion prediction less realistic for large signal levels, both for class A and class B amplifiers.

Figure 3 shows a macromodel that emulates the level 3 model by adding the parameter θ . In the level 1 the model error increases with the square of the gate voltage, $V_{GS}-V_{TH}$. Generating $(V_{GS}-V_{TH})^2$ with a second mosfet and subtracting it from the input voltage allows this error to be almost completely removed.

In Fig. 3, sensing the source current

with a controlled source generates a virtual source resistance. This allows the true physical source resistance to be used for a more realistic model particularly in the controlled resistance region.

The value of θ can be entered at the controlled source marked. The core mosfet's β setting is fixed at 1 because of interaction between β and θ . Instead, β is set using a controlled current source. The resulting curves are within 1% of the level 3 model over the normally used range of current.

Adding smooth pinch-off

An enhancement mosfet is said to 'pinch-off' when the value of $V_{GS}-V_{th}$ is equal to the drain-source voltage. At pinch-off the mosfet changes from a voltage controlled resistance to a voltage controlled current source.

In practice the transition is gradual, creating a knee region on a g_{DS}/V_{DS} plot, as discussed in the panel 'Simple curve tracer.'

A g_{DS} knee is not modeled by the level 1 or level 3 mosfets. Both produce a discontinuity in the slope of the drain g_{DS} conductance curve at pinch-off which prevents accurate modeling of the controlled resistance region and slows simulation as Spice steps through pinch-off.

It is possible to generate a knee in the g_{DS}/V_{DS} curve near pinch-off by adding a voltage controlled drain series resistance. Figure 4a) shows a macromodel that adds a voltage controlled resistance where the conductance is proportional to the internal gate-drain voltage, V_{gd} , of the core mosfet. This causes the core mosfet to approach pinch-off asymptotically and creates a knee in the drain conductance g_{DS} curve, operating effectively as a cascode circuit.

The core mosfet's drain conductance parameter λ must be set to zero because the voltage controlled series conductance prevents the core mosfet entering saturation. Instead, the drain conductance parameter λ can be generated externally using a second controlled conductance cell. This is in turn controlled by the core mosfet's drain current.

Figure 4b) shows simulations for a 2SK134 macromodel. Several key data sheet points are shown for comparison. Dotted lines show the effect of reducing β_0 from $4A/V^2$ to $2A/V^2$. Varying the shape of the g_{DS} knee with β_0 allows the current in the saturation region to be varied to fit the I_D/V_{DS} curves much more closely. This is a vast improvement over the level 1 and level 3 models.

Fig. 4a). Macromodel using a voltage-controlled conductance V_{CG} cell to generate a knee in the drain conductance curve.

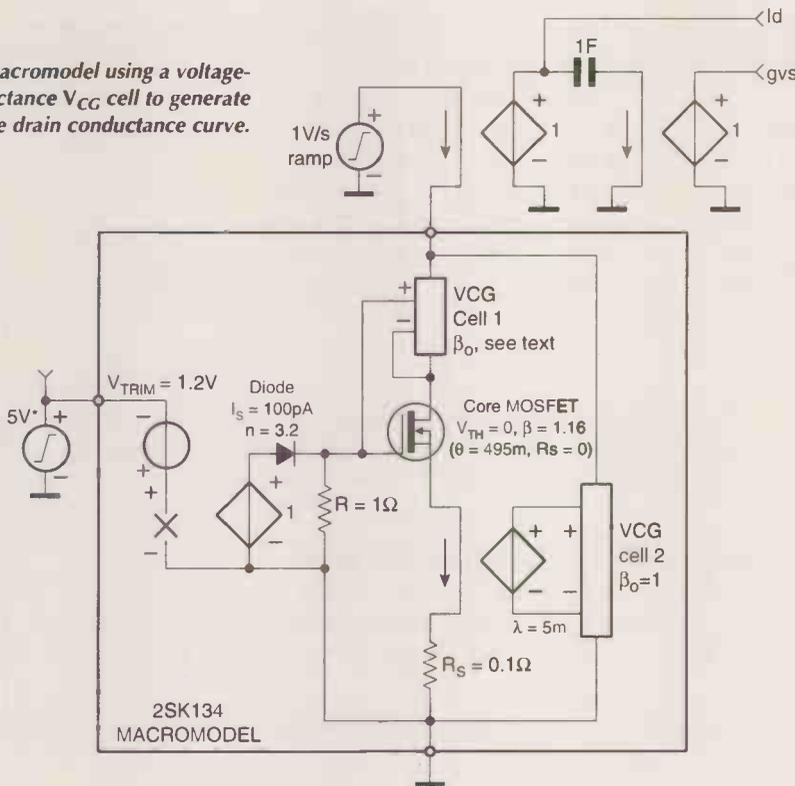
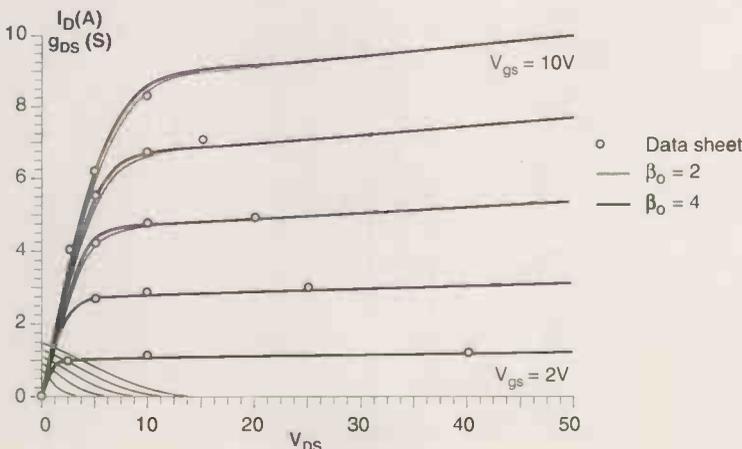


Fig. 4b). Simulated and actual results for the 2SK134 with the dotted line showing the effect of reducing β_0 . The lower family of curves shows drain conductance.

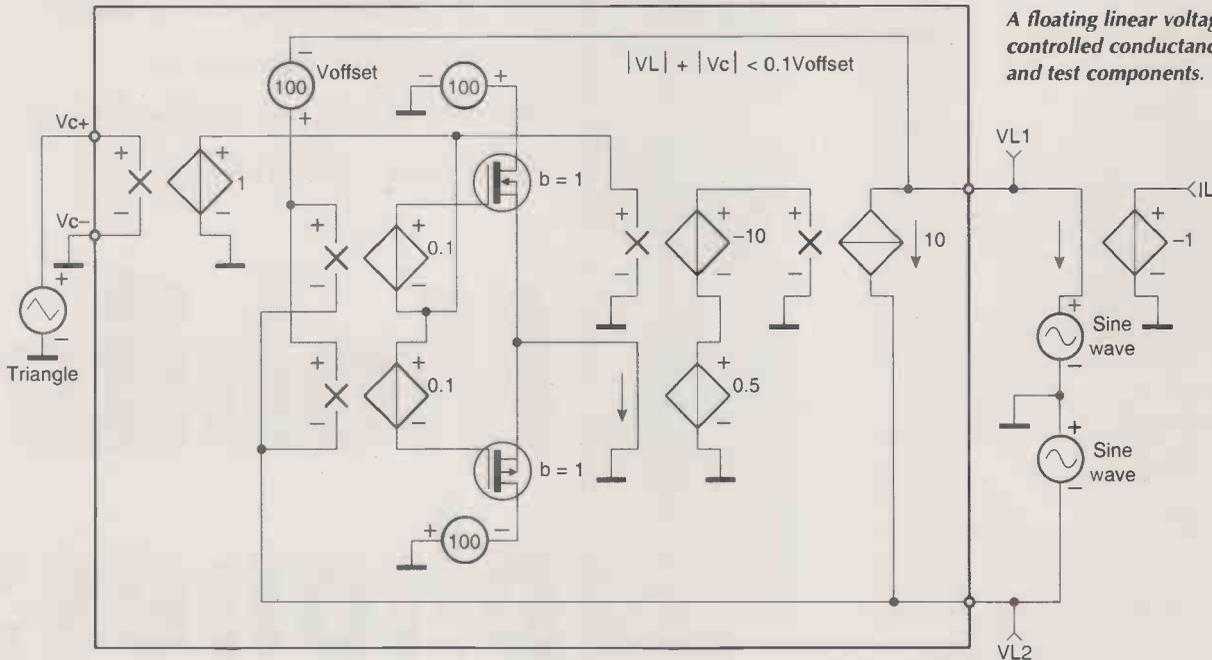


Sub circuit for a voltage-controlled conductance

Spice does not include a voltage-controlled conductance. This diagram shows one constructed using the square law multiplier and a voltage-

controlled current source. The output current is given by $I_{out} = \beta_0 V_1 V_2$. Making V_2 equal to V_{out} gives $I_{out}/V_{out} = \beta_0 V_1$

This is a linear voltage-controlled conductance. V_{out} should be kept within the range of voltage allowable. This is increased by scaling V_{out} by 0.1 in this case to increase the output voltage range.



A floating linear voltage-controlled conductance cell and test components.

Note that β_0 determines the shape of the transition region and does not significantly affect the drain conductance for V_{DS} is close to zero and hence $R_{DS(on)}$. This makes set up easier.

More parameter setting

A starting value for β_0 for lateral devices is around 5β of the core mosfet. Parameters β , θ , V_{th} , and n are found first from measurements in the saturation region. For vertical power mosfets R_s can be set at a typical value of $20m\Omega$ for TO220 devices or 5% to 10% of their $R_{DS(on)}$.

The usual method for finding R_d – a fixed series resistance – is as follows. First set R_d to zero and β_0 at around 5β . Next find the simulated $R_{DS(on)}$ at the gate voltage specified for the actual $R_{DS(on)}$. Then set R_d to the difference then vary β_0 to fit the I_D/V_{DS} curve at pinch-off.

Lateral devices have similar R_s and R_d values due to their geometry. The above method can be used for these devices if R_s is initially set to a likely value of R_d , such as 10% of $R_{DS(on)}$. If, after determining all the parameters, you find R_d is significantly different from R_s , then adjustment can be made to theta using $\theta_{new} = \theta - \beta \Delta R_s$ where ΔR_s is $R_{S(new)} - R_{S(old)}$.

In a second article, Ian looks at maths functions for Spice, modelling crossover distortion and thermal considerations. He also outlines how to put the models into practice.

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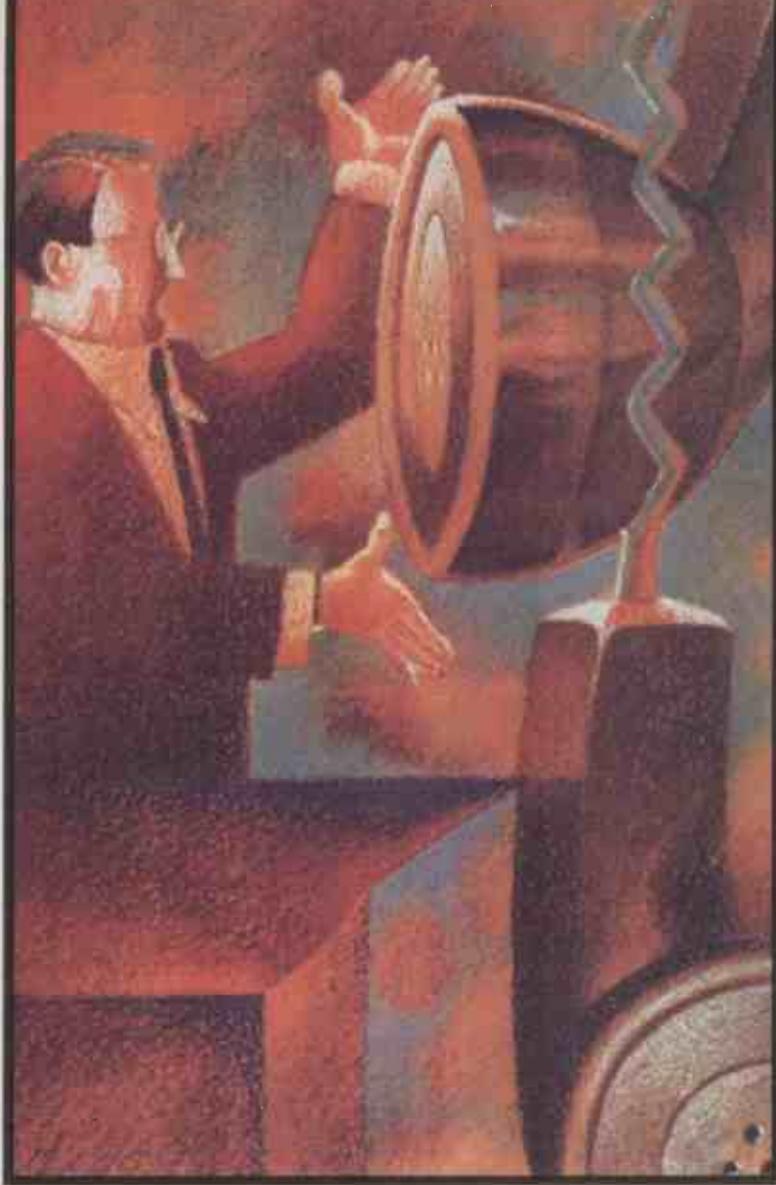
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In this second article looking into the pending third generation of mobile phones **Tim Forrester** details the new technology involved – CDMA.

Mobile phones: the next generation

Last month I looked at the politics surrounding the next generation of mobile phones. Here, I present an overview of the technology that will be used to provide the new service. This technology is code-division multiple access, or CDMA for short.

Forward and reverse links

Before you can appreciate CDMA, it is necessary for you to understand forward and reverse links and the interaction

between the base-station, the base-station controller and the public switched telephone network, or PSTN.

The forward link is the transmission from the base station to the mobile and the reverse link is the transmission from the mobile to the base station.

As you can see from Fig. 1, the base station is controlled by a base station controller, which may also be responsible for controlling a number of other base stations in other cells.

A base station is sometimes called a BTS, which is short for base station transceiver sub-system. From here though, I will use the term base station.

The base station controller provides a number of functions, such as switching a signal to the appropriate base station and billing information. It also acts as a link to the PSTN.

Returning to the radio links, while both of them use spread spectrum modulation, there are several important differences between them.

The forward link

The base station is continually transmitting a spread spectrum signal which has a number of important signals modulated onto its carrier.

Forward traffic transmission is generated as shown in Fig. 2. Incoming information from the vocoder or other data source first passes through a convolutional encoder. The pur-

What will CDMA mobile phone technology offer?

Mobile phones based on the new technology are expected in the shops in the UK early 2000. Below is a brief outline of new features and performance benefits that the new CDMA technology will make possible.

- Better hand-over between cells – fewer dropped calls
- Fewer holes in coverage
- Possibility of global coverage via low-orbit satellites – with suitable multi-mode phones
- Billions of telephone number options
- Longer battery life, handset transmitter power managed by the base station
- Less interference, better performance

pose of this encoder is to provide a forward error correcting function, shortened to FEC. The convolutional coding process provides data redundancy, which the receiver uses to correct errors.

Symbol repetition reduces the amount of energy needed for each symbol, lowering the transmitted power requirement. This in turn reduces the interference levels to other users.

Subsequent interleaving is performed on 20ms blocks of symbols. 20ms is also the vocoder frame rate.

Noise in a radio channel tends to induce 'burst errors', but FEC coding copes better with randomly spaced errors. So by interleaving the symbols prior to transmission, any burst errors are effectively randomly scattered when the symbols are de-interleaved at the receiver.

Scrambling traffic

The next stage is to scramble the traffic channel with a pseudo-random noise sequence. Each symbol from the block interleaver is added modulo-2 with one chip of the 19200 chips per second scrambling sequence.

A device known as a long-code pseudo noise, or PN, code generator supplies the scrambling sequence. The mask for the PN generator is 42 bits long of which 32 are based on the mobile's electronic serial number, or ESN.

The long-code generator is a linear feedback shift register with pre-determined taps to generate the pseudo random scrambling sequence. It produces a signal at 1.2288Mchars/s, which is then divided by 64 in the 'decimator' for the purposes of scrambling.

Mobile transmit power control bits are added to the data stream at a rate of 800Hz. These bits are essential for keeping the RF power output of the mobile as low as possible.

The power control bits actually overwrite data thus introducing errors into the data stream. However due to the heavy data protection provided by the FEC system, the mobile receiver is able to correct these errors.

In this manner the mobile transmitter power is adjusted 800 times per second.

The signal is now spread orthogonally using a Walsh code. The form of Walsh code and the meaning of orthogonal are explained later.

Each traffic channel in the forward direction involves a unique Walsh code. However certain Walsh codes are reserved for system purposes.

The pilot channel employs Walsh code 0. Paging channels employ Walsh codes 1 to 7 and the sync channel is always code 32. The purposes of these dedicated Walsh codes are described later.

After the signal has been spread by the appropriate Walsh code, it is spread over both quadrants by means of a universal short code - actually the pilot channel.

Each base station can have a unique identifier since the same short code is fed to both quadrature modulators but with a specific offset between I and Q channels. This works because all base stations are synchronised to an accurate frequency reference, namely the Global Positioning System time reference.

The base-band signal is now ready for frequency conversion to the transmitters' intermediate frequency before final translation to the transmission frequency. However, it first needs to be band limited to meet FCC specification of -3dB at 615kHz. It is then applied to a quadrature phase-shift keyed modulator ready for frequency translation.

The panel entitled 'Functions of dedicated forward channels' contains extra information.

The reverse link

At mobile switch-on the initial function of the reverse link is to announce the presence of the mobile to the selected base station. Thus the system knows where the mobile is

located so it can direct paging and traffic accordingly.

However, as you can be seen from Fig. 3, the generation of the reverse link signal is substantially different to that of the forward link.

As with the forward link, the data to be transmitted by the reverse link first passes through a convolutional coder and interleaver to provide an FEC capability. The inter-

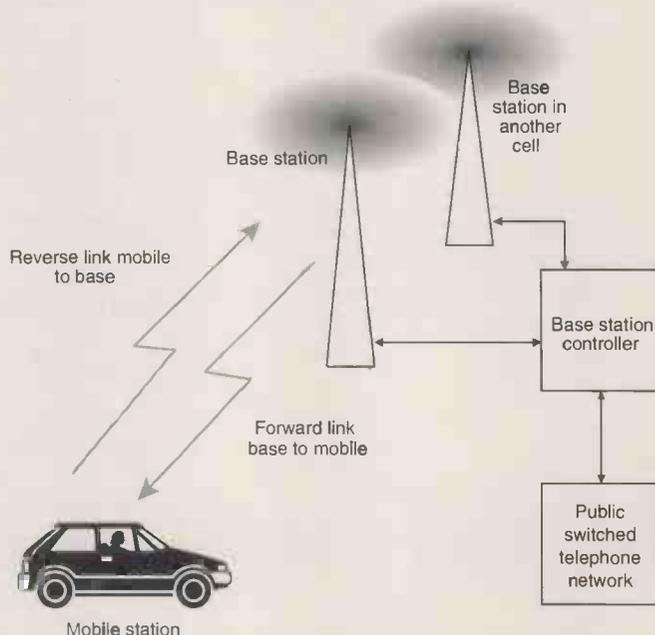


Fig. 1. The base station is controlled by the base station controller, which may also be responsible for controlling other base stations.

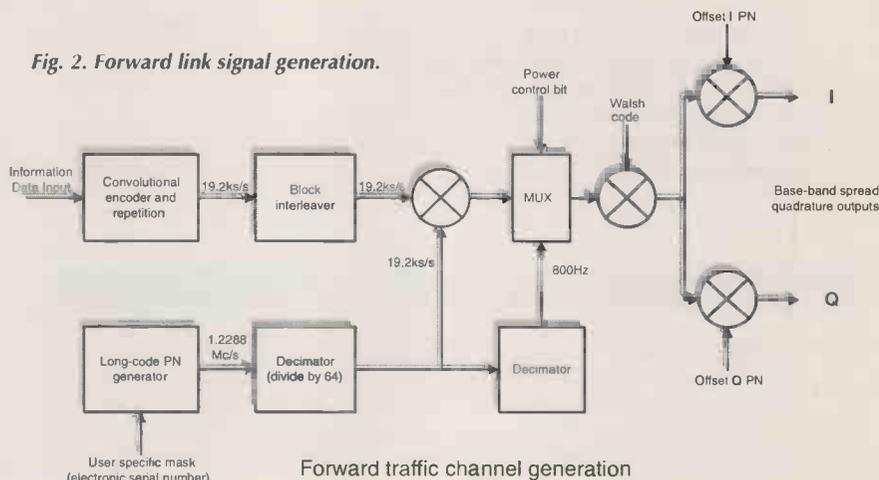
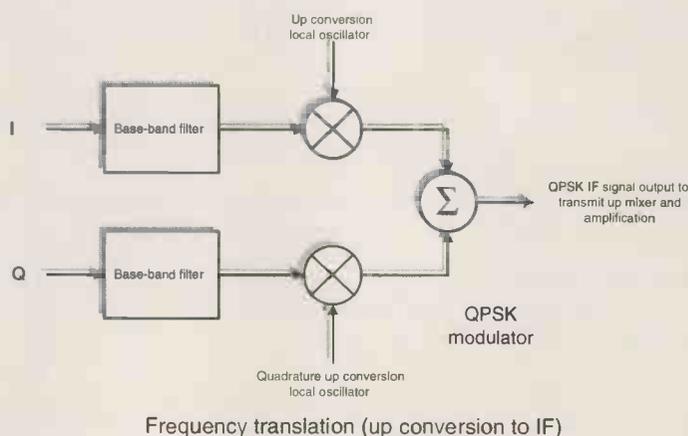


Fig. 2. Forward link signal generation.



Frequency translation (up conversion to IF)

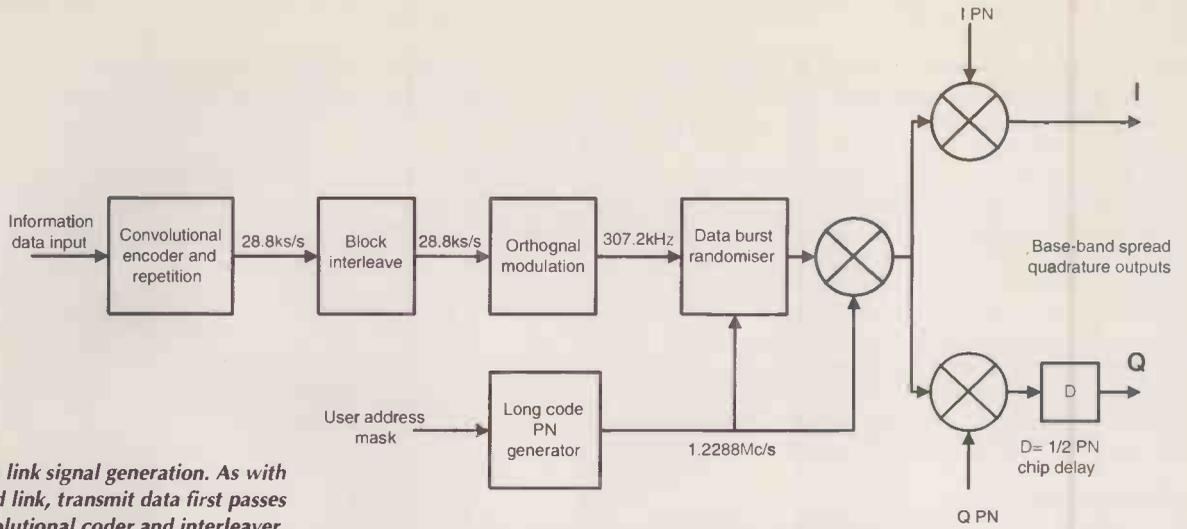
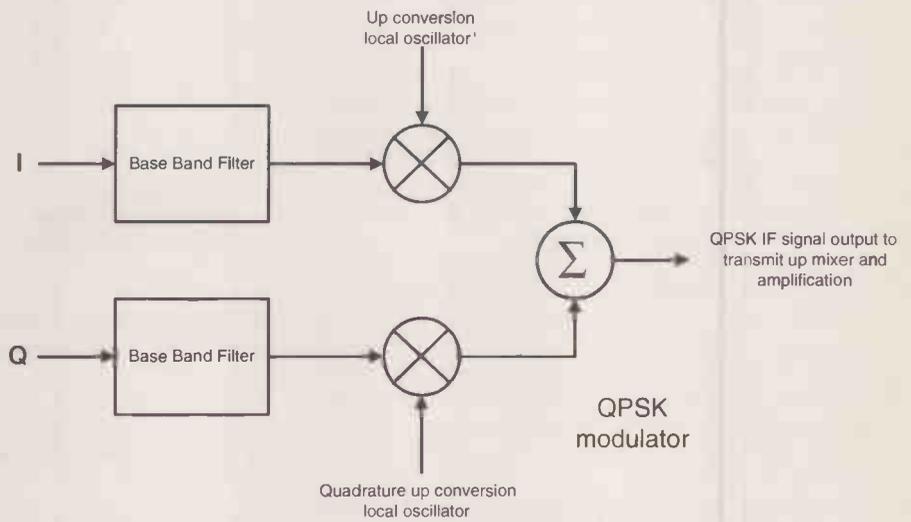


Fig. 3. Reverse link signal generation. As with the forward link, transmit data first passes through a convolutional coder and interleaver, but in many other respects, it is quite different.

Reverse traffic channel generation



Frequency translation (up conversion to IF)

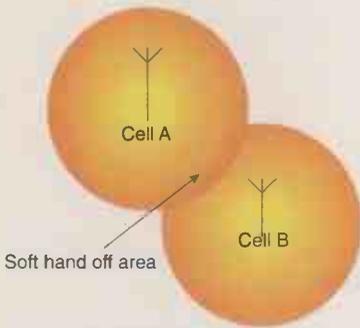


Fig. 4. Soft hand-off area. Unlike its predecessors, CDMA does not switch immediately from one transmitter to another as the cell boundary is crossed. Instead it makes use of both transmitters while in the soft hand-off area, resulting in better performance and fewer dropped lines.

Functions of dedicated forward channels

- **Pilot channel.** The mobile demodulates the pilot channel and so receives a reference for time, phase and signal strength. In this manner the mobile can determine if the base station it is currently decoding is the best candidate in terms of signal strength. The use of the pilot channel as a phase reference enables the rake receiver to coherently demodulate the incoming signals. Furthermore as different base stations have different short-code offsets, the mobile can determine which base station is the strongest even if they occupy the same channel. This is important for a mobile-assisted 'soft hand-off'.
- **Traffic channel.** As the name implies, this channel conveys the information to be transmitted between the mobile and base station.
- **Sync channel.** The sync channel transmits a repeating message that identifies the station and the absolute phase of the pilot sequence.
- **Paging channel.** The paging channel operates in a similar manner to the forward traffic channel, but as its name implies is used to 'page' the mobile and establish the call.

mediate symbol repetition is for system convenience only. The repeated symbols are not transmitted. Instead, the data-burst randomiser will select only one copy of each symbol for transmission. To conserve power and minimise interference, the transmitter is turned off for the redundant symbols.

The purpose of the orthogonal modulation scheme – again using Walsh codes – is to provide easily identifiable isolation between symbols.

The chosen Walsh codes contain 64 possible combinations. The information to be transmitted is broken up into groups of 6 symbols. These 6 symbols then correspond to a value between 0 and 63. This value is used to select a Walsh code for transmission.

As speech has a typical 'duty cycle' of 40%, during periods of reduced speech activity the vocoder will reduce its data rate – thus the overall transmitted power will also be reduced. This means that system capacity is freed up as the shared resource is rf power.

Furthermore, the data-burst randomiser will turn off the transmitter when the symbol repetition module produces redundant information.

The signal from the data randomiser occupies a bandwidth of 307.2kHz and is passed onto the long code spreader.

All reverse channels are isolated from one another by using the same long code but with different offsets. Thus

each mobile is allocated a specific offset from the start of the long code. As the long code takes 41 days to repeat at a clock rate of 1.2288Mchars/s there are billions of offsets available to address individual mobiles.

Finally the signal is spread in quadrature with the short PN codes. No offset is employed, but the quadrature branch is delayed by 1/2 of a PN chip to produce offset QPSK rather than QPSK.

This OQPSK relaxes the linearity requirement on the mobile transmitter, as there are no zero-amplitude phase transitions.

At the base station each mobile is separated from others by correlating the incoming signal containing a version of the long code with the appropriate offset. This offset is known to the base station.

Soft hand-off

One of the major benefits of CDMA is the 'soft hand-off' feature. What this means is that there is no fixed hand-off boundary. Instead, there is a hand-off region where the mobile may be in simultaneous communication with a number of base stations.

This is effectively a 'make before break' hand off. In this way there is inherent diversity, which helps to fill in any coverage 'holes' and reduce the probability of dropped calls.

Figure 4 shows the 'soft hand-off area' where a mobile might be in contact with a number of base stations. A soft hand-off area is typically on the fringe of a cell. It is exactly where coverage is normally patchy and might otherwise lead to dropped calls if a hard hand-off was employed – as in a typical TDMA system.

Therefore a soft hand-off not only minimises dropped calls, it also helps to improve signal integrity in locations which would otherwise have poor coverage from an individual base station.

Indeed if the mobile is being received by a number of base stations, the system will normally command the mobile to reduce power until it is only being received by one base station. This base station will then control the mobile transmit power as required to sustain the link.

Eventually, as the mobile moves away from its selected base station, the mobile transmit power will be increased to a point where another base station will detect the signal and the process of 'soft hand-off' starts again.

Walsh codes

In this particular CDMA application there are 64 Walsh codes, each 64 chips long. They are produced at a rate of 1.2288Mchars/s to spread the information signal. Each Walsh code is orthogonal.

Two binary sequence codes are called orthogonal if the process of carrying out an exclusive-or function on them results in an equal number of ones and zeros, i.e. the cross correlation is zero. An example is shown below.

```

0110
1100
-----
1010
    
```

The Walsh codes are a particular set of orthogonal codes, which are easily generated as follows. Start with a 'seed' of 0. Repeat to the right and below, then invert diagonally.

Seed

0	

Repeat horizontally and vertically,

0	0
0	

Invert diagonally,

0	0
0	1

The same rules are then applied to the whole pattern.

0	0	0	0
0	1	0	1
0	0	1	1
0	1	1	0

The above procedure is repeated until an array of 64 by 64 is generated.

When applied to the forward link, each information bit to be transmitted is spread, i.e. exclusive-ored, by a particular 64-chip sequence, or Walsh code, allocated to that traffic channel.

To retrieve the information bit at the receiver, the incoming 64-chip sequence is 'de-spread' – i.e. exclusive-ored again – with the same Walsh function.

If the wrong Walsh code is used for de-spreading, the resulting correlation produces an average of zero. Such is the power of orthogonal Walsh codes.

CDMA reception

So far this article has concentrated on the signal format of the forward and reverse links in a typical CDMA system. In many ways, the reception, demodulation and decoding of a spread spectrum signal is more challenging – and interesting.

However, due to space constraints it is not possible to cover all of the issues involved with the design of the rf receiver and subsequent digital signal processing, but I hope that the following paragraphs will give you an insight into the problems and solutions associated with CDMA reception.

The radio section

A typical CDMA receiver employs a conventional super-heterodyne architecture but with special attention paid to certain aspects. A few of the main areas of concern are:

- Blocking performance
- IF group delay
- Transmitter broad band noise.

Blocking performance. The receiver de-sensitises when receiving a weak CDMA signal in the presence of a strong

Fig. 5. Multi-channel base-station receiver. A base station with a single rf front end might have a thirty-fingered rake receiver capable of serving ten mobiles with three fingers each. This give a significant reduction in rf hardware complexity and cost.

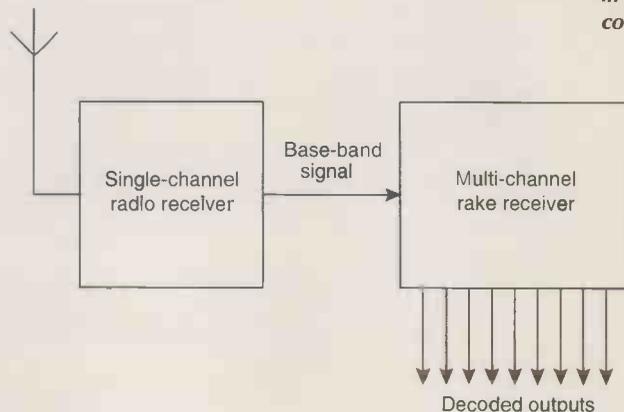
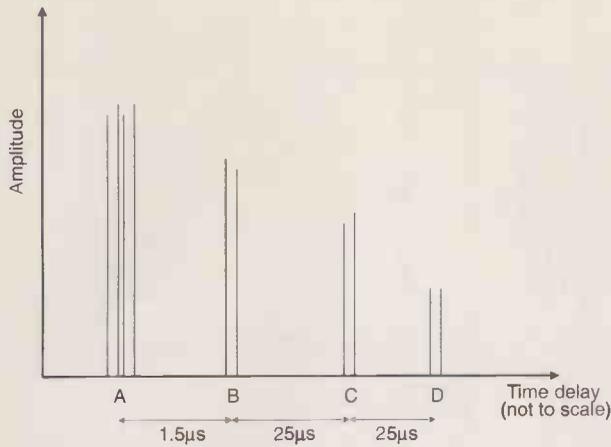


Fig. 6. Typical delays that a radio signal might experience in an urban environment.



blocking signal. This is especially critical when a CDMA system has to co-exist with a conventional cellular system.

Intermediate-frequency group delay. As the digital signal processing attempts to correlate multipath – time delayed – signals, it is essential that any additional signal delay distortion within the receive path is kept to a minimum.

Transmitter broad-band noise. As the receiver may be operating in full duplex with the transmitter, any broad-band noise from the transmitter must be managed in such a way that it does not adversely de-sensitise the receiver. This requirement often imposes severe filtering and screening requirements on both the transmitter and receiver.

For these reasons, the design of a CDMA receiver is considered more of a challenge than that of an equivalent TDMA receiver.

Base-band dsp and the 'rake' receiver

Base band digital signal processing is at the heart of a CDMA system. Typically, companies will have spent many man-years in the design and execution of an optimum processing system. For this reason the detail of the algorithms, ASIC hardware and digital signal processing code of a CDMA system are normally proprietary and closely guarded.

Other companies may have to make licence fee or royalty payments of tens of millions of dollars for access to the intellectual property rights for the sole purpose of being able to produce a CDMA phone. Therefore this article can only describe how a generic CDMA base-band section might operate.

One of the main advantages of CDMA is the ability of the receiver to constructively add multipath signals, which would otherwise be destructive in a narrow-band system.

A mobile receiver typically employs a three-channel rake receiver. Two of the channels are used to correlate incoming signals, while the third is used as a search receiver to detect the presence of pilot channels from other base stations which may be coming into range.

The name rake for the receiver is an apt description as each prong of the rake can be thought as a finger searching for a particular code signal. The greater the number of fingers on the rake the greater the number channels the receiver can handle.

Thus a base station equipped with single rf receiver but with a thirty-finger rake receiver might employ three fingers per mobile, resulting in a channel capacity of ten. This results in a great reduction – and hence cost – in rf hardware complexity, Fig. 5.

The ability of a rake receiver to correlate signals dispersed in time is determined by the chip rate and the number of fingers. The faster the chip rates the finer the resolution in time and hence distances for multipath.

There are however good reasons for not having an

excessively high chip and thus signal bandwidth. Presently, domestic CDMA systems use a chip rate of 1.2288Mchars/s, which equates to a time resolution of 0.8µs.

Likewise, due to power and size constraints, the mobile would normally have only three fingers used as described above. It will not be possible to correlate multipath signals arriving at the receiver at a time interval of less than approximately 1µs with the main signal. However a delay of 1µs equates to approximately 1000 feet. Provided the multipath signal path is greater than 1000 feet the rake receiver will either be able to correlate or ignore it.

Figure 6 shows the typical delays that a radio signal might incur in an urban environment. Clusters A and B are the main signals separated by a short delay. These signals are capable of being correlated if they have a sufficiently high chip rate and signal bandwidth.

Other signals, such as C and D, have a much longer delay and lower amplitude are effectively ignored. However if Fig. 6 were to represent a narrow band signal, it would be practically impossible to identify the individual components. There could be severe interference and probable signal destruction.

From the above descriptions it should be apparent that the combination of a multi-channel rake receiver and an adequately high signal chip rate offers excellent protection against multi-path interference. Thus the ability of a rake receiver to recover a signal under multi-path conditions combined with the power of Walsh codes and FEC coding makes for a robust data link.

Base-band signal decoding

Consider the mobile phone decoding the forward link. The mobile makes use of the base station's embedded pilot signal for a phase reference, enabling the use of coherent demodulation.

For the mobile to decode the base station transmission, it merely performs the reverse of the signal encoding, but with the appropriate Walsh code for channel selection.

Now consider the base station decoding the reverse link from the mobile phone. The base station cannot use coherent demodulation as there is no pilot. In any event, there is no practical means of keeping all mobiles synchronised at the base station.

However, demodulation of the signal is the reverse of the encoding, but with the appropriate offset for the long code for channel selection. Much of the above decoding is performed in hardware, such as ASICs, which have been developed to perform FEC decoding.

In summary

The technology of mobile phones has advanced at a stupendous rate over the past twenty years – driven by the ever-growing need to communicate.

As the popularity of the mobile phone has increased, so has the demand on that most limited resource called 'bandwidth'. In turn this has driven the development of ever more efficient coding and modulation schemes.

As the new millennium approaches, so does the prospect of a whole new mobile phone standard – not just 3G but the advent of a satellite-based mobile phone system. When such a system is fully operational mobile phone users will at last be freed from dependence on a ground based infra structure – typically at its best in areas of high population and alongside major highways.

Thus not too far in the future – if all goes according to plan – mobile phone users will be able to roam the planet in sure knowledge that a low Earth-orbiting satellite will be able to provide a stable communication link in even the most remote areas.

When this happens the mobile phone will truly have become of age. ■

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 TEKTRONIX 577 Curve tracer + adaptors - £300.
 TEKTRONIX 1502/1503 TDR cable test set - £400.
 HP8699B Sweep PI YIG oscillator .01 - 4GHz - £300. 8690B MF-
 £250. Both £500.
 Dummy Loads & Power att up to 2.5 kilowatts FX up to 18GHz -
 microwave parts new and ex equip - relays - attenuators -
 switches - waveguides - Yigs - SMA - APC7 plugs - adaptors etc.
 qty. in stock.
 B&K items in stock - ask for list.
 Power Supplies Heavy duty + bench in stock - Farnell - HP -
 Weir - Thurlby - Rascal etc. Ask for list. Large quantity in stock,
 all types to 400 amp - 100Kv.
 HP8405A Vector voltmeter - late colour - £400.
 HP8508A Vector voltmeter - £2500.
 HP8505A Network Anz 500KHz-1.3GHz - £1000.
 HP8505A + 8502A or 8503A test sets - £1200 - £1500.
 HP8505A + 8502A or 8503A + 8501A normalizer - £1750-£2000.
 Phillips 3217 50Mc/s oscilloscopes - £150-£250.
 Phillips 3296 350Mc/s IR remote oscilloscope - £500.
 R&S APN 62 LF S/G 0.1Hz - 260KHz with book - £500.
 Wavetek-Schlumberger 4031 Radio communication test set

LIGHT AND OPTICAL EQUIPMENT

Anritsu ML93A & Optical Lead Power Meter - £250.
 Anritsu ML93B & Optical Lead Power Meter - £350.
 Power Sensors for above MA96A - MA98A - MA913A - Battery
 Pack MZ95A.
 Anritsu MW97A Pulse Echo Tester.
 PI available - MH914C 1.3 - MH915B 1.3 - MH913B 0.85 -
 MH925A 1.3 - MH929A 1.55 - MH925A 1.3GI - MH914C 1.3SM -
 £500 + one PI.
 Anritsu MW98A Time Domain Reflector.
 PI available - MH914C 1.3 - MH915B 1.3 - MH913B 0.85 -
 MH925A 1.3 - MH929A 1.55 - MH925A 1.3GI - MH914C 1.3SM -
 £500 + one PI.
 Anritsu MZ100A E/O Converter.
 + MG912B (LD 1.35) Light Source + MG92B (LD 0.85)
 Light Source £350.
 Anritsu MZ118A O/E Converter.
 +MH922A 0.8 O/E unit + MH923 A1.3 O/E unit £350.
 Anritsu ML96B Power Meter & Charger £450.
 Anritsu MN95B Variable Att. 1300 £100.
 Photo Dyne 1950 XR Continuous Att. 1300 - 1500 £100.
 Photo Dyne 1800 FA. Att £100.
 Cossor-Raytheon 108L Optical Cable Fault Locator
 0-1000M 0-10km £200.
 TEK P6701 Optical Converter 700 Mc/S-850 £250.
 TEK OF150 Fibre Optic TDR - £750.
 HP81512A Head 150Mc/S 950-1700 £250.
 HP84801A Fibre Power Sensor: 600-1200 £250.

HP8158B ATT OPT 002+011 1300-1550 £300.
 HP81519A RX DC-400Mc/S 550-950 £250.
 STC OFR10 Reflectorometer - £250.
 STC OFSK15 Machine jointing + eye magnifier - £250.

COMMUNICATION EQUIPMENT

Anritsu ME453L RX Microwave ANZ - £350.
 Anritsu ME453L TX Microwave ANZ - £350.
 Anritsu MH370A Jitter Mod Oscillator - £350.
 Anritsu MG62A Pulse Patt Gen. £350.
 System MS02A Timer & Digital Printer - £500.
 Complete MS65A Error Detector.
 Anritsu ML612A Sel Level Meter - £400.
 Anritsu ML244A Sel Level Meter - £300.
 W&G PCM3 Auto Measuring Set - £300.
 W&G SPM14 Sel Level Meter - £300.
 W&G SPM15 Sel Level Meter - £350.
 W&G SPM16 Sel Level Meter - £400.
 W&G PS19 Level Gen - £500.
 W&G DA20+DA1 Data Anz £400.
 W&G PMG3 Transmission Measuring Set - £300.
 W&G PSS16 Generator - £300.
 W&G PS14 Level Generator - £350.
 W&G EPM-1 Plus Head Millivolt Power Meter - £450.
 W&G DLM3 Phase Jitter & Noise - £350
 W&G DLM4 Data Line Test Set - £400.
 W&G PS10 & PM10 Level Gen - £250.

MISCELLANEOUS ITEMS

HP 3852A Data Acquisition Control Unit + 44721A 16ch input
 £1,000.
 HP 4261 LCR meter - £650.
 HP 4274 FX LCR meter - £1,500.
 HP 4951 Protocol Anz - £500.
 HP 3488 Switch Control Unit + PI Boards - £500.
 HP 7500 VXI Bus Controllers + I3236B-DVM-quantity.
 HP 83220A GSM DCS/PCS 1805-1990Mc/S converter for use
 with 8922A - £2,000.
 HP 1630-1631-1650 Logic ANZ's in stock.
 HP 8754A Network ANZ 4-1300Mc/S + 8502A + cables - £1,500.
 HP 8754A Network ANZ H26 4-2600Mc/S + 8502A + Cables -
 £2,000.
 HP 8350A Sweeper MF + 83540A PI 2-8.4GHz + 83545A PI 5.9-
 12.4GHz all 3 - £3,500.
 HP MICROWAVE TWT AMPLIFIER 489A 1-2GHz-30DB - £400.
 HP PREAMPLIFIER 8447A 0.1-400Mc/S - £200. Dual - £300.
 HP PREAMPLIFIER 8447D 0.01-1.3GHz - £400.
 HP POWER AMPLIFIER 8447E 0.01-1.3GHz - £400.
 HP PRE + POWER AMPLIFIER 8447F 0.01-1.3GHz - £500.
 HP 3574 Gain-Phase Meter 1Hz-13Mc/S OPT 001 Dual - £400.
 MARCONI 2305 Modulation Meter-50KHz-2.3 GHz - £1,000.
 MARCONI 2610 True RMS Meter - £450.
 MARCONI 893B AF Power Meter (opt Sinad filter) - £250-£350.
 MARCONI 6950-6960B Power Meters + Heads - £400-£900.
 MARCONI SIGNAL SOURCE-6055-6056-6057-6058-6059 - FX
 Range 4-18GHz - £250-£400.
 RACAL 1792 COMMUNICATION RX - £500 early - £1,000 - late
 model with back lighting and byte test.
 RACAL 1772 COMMUNICATION RX - £400-£500.
 PLESSEY PR225 COMMUNICATION RX - £500-£900.
 TDC MODULE MAINFRAMES - TM501-502-503-504-506-
 TM5003-5006.
 TEK PI 5010-M1 - Prog Multi Interface - £250. FG Prog 20Mc/S
 Function Gen - £400 - S1 Prog Scanner - £250 - DM Prog DMM
 - £400.
 TEK 7000 OSCILLOSCOPE MAINFRAMES - 7603-7623-7633-
 7834-7854-7904-7904A-7104 - £150-£1,000.
 TEK 7000 PI's - 7A11-7A12-7A13-7A18-7A19-7A22-7A24-7A26-
 7A29-7A42-7B10-7B15-7B53A-7B80-7B85-7B92A-7D15-7D20.
 TEK 7000 - 7S11-7S12-7S14-7M11-S1-S2-S3A-S4-S5-S6-S51-
 S53-S54.

RADIO COMMUNICATION TEST SETS

BULK PURCHASE ONLY FROM JOHNS RADIO
 HP 8920 RF Communication Test Set - Opts 003-004-007-011
 unit contains Syn Signal Gen-Distortion Meter-Mod Meter-
 Digital Oscilloscope etc. 1000Mc/S - £1,500 each.
 MARCONI 2955 RF Test Sets-1000Mc/S - £1,200 each.
 MARCONI 2958 RF Test Sets-1000Mc/S - £1,300 each.
 MARCONI 2960 RF Test Sets-1000Mc/S - £1,400 each.
 MARCONI 2955A RF Test Sets-1000Mc/S - £2,000 each.
 MARCONI 2960A RF Test Sets-1000Mc/S - £2,500 each.
 ANRITSU MSS55A2 Radio Comm Anz-1000Mc/S - £1,200 each.
 MARCONI 2019A SYNTHESIZED SIGNAL GENERATORS -
 80Kc/S-1040Mc/S - AM-FM all functions tested off the pile as
 received from Gov - in average used condition - £650 each or
 in original Gov cartons. 1st class condition each fitted with IEEE
 plus added protection front cover lid containing RF-IEEE-mains
 cables + N to BNC adaptor - Attenuator etc. + Instruction Book
 - fully checked to high standards in our own workshop - £1k.
 MARCONI 2022E SYNTHESIZED SIGNAL GENERATOR -
 10Kc/S-1.01GHz AM-FM - made small and light for portability
 being the naval version - all functions tested off the pile as
 received from Gov - in average used condition - £1,000 each or
 in original Gov cartons as new condition - each fitted with IEEE
 + added protection front cover lid containing RF-IEEE - mains
 cables-N to BNC Adaptor - Attenuator-50-75OHM adaptor etc. +
 Instruction Book - fully checked to high standards in our own
 workshop - £1,250 each.
 WE KEEP IN STOCK HP and other makes of RF Frequency
 doublers which when fitted to the RF output socket of a
 S/Generator doubles the output frequency EG.50-1300Mc/S to
 50-2600Mc/S price from £250 - £450 each.

SPECTRUM ANALYZERS

HP 3580A 5Hz-50KHz - £750.
 HP 3582A Dual 0.2Hz-25.5KHz - £1,500.
 HP 3585A 20Hz-40Mc/S - £3,500.
 HP 3588A 10Hz-150Mc/S - £7,500.
 HP 8568A 100Hz-1.5GHz - £3,500.
 HP 8568B 100Hz-1.5GHz - £4,500.
 HP 8590B 9Kc/S-1.8GHz - £4,500.
 HP 8569B 10Mc/S (0.01-22GHz) - £3,500.
 HP 3581A Signal Analyser 15Hz-50KHz - £400.
 TEK491 10Mc/S-12.4GHz + 12.4-40GHz - £500.
 TEK492 50KHz-21GHz OPT 2 - £2,500.
 TEK493 50KHz-21GHz OPT 1-1.3 - £3,500.

TEK492AP 50KHz-21GHz OPT 1-2-3 - £4,000.
 TEK495 100KHz-1.8GHz - £2,000.
 ANRITSU MS710F 100Kc/S-23GHz - £4,000.
 HP 8557A 0.01Mc/S-350Mc/S - £500 + MF180T or 180C - £150 -
 182T - £500.
 HP 8558B 0.01-1500Mc/S - £750 - MF180T or 180C - £150 -
 182T - £500.
 HP 8559A 0.01-21GHz - £1,000 - MF180T or 180C - £150 - 182T
 - £500.
 HP 8901A AM FM Modulation ANZ Meter - £800.
 HP 8901B AM FM Modulation ANZ Meter - £1,750.
 HP 8903A Audio Analyzer - £1,000.
 HP 8903B Audio Analyzer - £1,500.
 MARCONI 2370 SPECTRUM ANALYZERS - HIGH QUALITY -
 DIGITAL STORAGE - 30Hz-110Mc/S Large qty to clear as
 received from Gov - all sold as is from pile complete or add
 £100 for basic testing and adjustment - callers preferred - pick
 your own from over sixty units - discount on qty's of five or
 more.
 A EARLY MODEL GREY - horizontal alloy cooling fins - £200.
 B LATE MODEL GREY - vertical alloy cooling fins - £300.
 C LATE MODEL BROWN - as above (few only) - £500.

OSCILLOSCOPES

TEK 465-465B 100Mc/S + 2 probes - £250-£300.
 TEK 466 100Mc/S storage + 2 probes - £200.
 TEK 475-475A 200Mc/S-250Mc/S + 2 probes - £300-£350.
 TEK 2213-2213A-2215-2215A-2224-2225-2235-2236-2245-60-
 100Mc/S - £250-£400.
 TEK 2245 4ch 150Mc/S + 2 probes - £450.
 TEK 2245A 4ch 150Mc/S + 2 probes - £600.
 TEK 2245B 4ch 150Mc/S + 2 probes - £750.
 TEK 468 D.S.O. 100Mc/S + 2 probes - £500.
 TEK 485 350Mc/S + 2 probes - £550.
 TEK 2465 4ch-300Mc/S - £1,150.
 TEK 2465A 4ch-350Mc/S - £1,550.
 TEK 2465ACT 4ch-350Mc/S - £1,750.
 TEK 2465B 4ch-400Mc/S - £2,000.
 TEK 2467 4ch-350Mc/S - £2,000.
 TEK D.S.O. 2230 -100Mc/S + 2 probes - £1,000.
 TEK D.S.O. 2430 -150Mc/S + 2 probes - £1,250.
 TEK D.S.O. 2430A -150Mc/S + 2 probes - £1,750.
 TEK D.S.O. 2440 -300Mc/S + 2 probes - £2,000.
 TEK TAS 475-485 -100Mc/S-20Mc/S-4ch + 2 probes - £900-
 £1,100.
 HP1740A - 100Mc/S + 2 probes - £250.
 HP1741A - 100Mc/S storage + 2 probes - £200.
 HP1720A - 1722A - 1725A - 275Mc/S + 2 probes - £300-£400.
 HP1744A - 100Mc/S storage - large screen - £250.
 HP1745A - 1746A - 100Mc/S - large screen - £350.
 HP54100A - 1GHz digitizing - £500.
 HP54200A - 100Mc/S digitizing - £500.
 HP54501A - 100Mc/S digitizing - £500.
 HP54100D - 1GHz digitizing - £1,000.

MICROWAVE COUNTERS - ALL LED READOUT

EIP 351D Autohot 20Hz-18GHz - £750.
 EIP 371 Micro Source Locking - 20Hz-18GHz - £850.
 EIP 451 Micro Pulse Counter - 300Mc/S-18GHz - £700.
 EIP 545 Microwave Frequency Counter - 10Hz-18GHz - £1K.
 EIP 548A Microwave Frequency Counter - 10Hz-26.5GHz - £1.5k.
 EIP 575 Microwave Source Locking - 10Hz-18GHz - £1.2k.
 EIP 588 Microwave Pulse Counter - 300Mc/S-26.5GHz - £1.4k.
 SD 6054B Micro Counter 20Hz-24GHz - SMA Socket - £800.
 SD 6054B Micro Counter 20Hz-18GHz - N Socket - £700.
 SD 6054D Micro Counter 800Mc/S-18GHz - £600.
 SD 6246A Micro Counter 20Hz-26GHz - £1.2k.
 SD 6244A Micro Counter 20Hz-4.5GHz - £400.
 HP5352B Micro Counter OPT 010-005-46GHz - new in box - £5k.
 HP5340A Micro Counter 10Hz-18GHz - Nixey - £500.
 HP5342A Micro Counter 10Hz-18-24GHz - £800-£1K - OPTS
 001-002-003-005-011 available.
 HP5342A + 5344S Source Synchronizer - £1.5k.
 HP5345A 500Mc/S 11 Digit LED Readout - £400.
 HP5345A + 5354A Plugin - 4GHz - £700.
 HP5345A + 5355A Plugin with 356A 18GHz Head - £1k.
 HP5385A 1GHz 5386A-5386A 3GHz Counter - £1K-£2K.
 Rascal/Dana Counter 1991-160Mc/S - £200.
 Rascal/Dana Counter 1992-1.3GHz - £600.
 Rascal/Dana Counter 9921-3GHz - £350.

SIGNAL GENERATORS

HP8640A - AM-FM 0.5-512-1024Mc/S - £200-£400.
 HP8640B - Phase locked - AM-FM-0.5-512-1024Mc/S - £500-
 £1.2k. Opts 1-2-3 available.
 HP8654A - B-AM-FM 10Mc/S-520Mc/S - £300.
 HP8656A SYN AM-FM 0.1-990Mc/S - £900.
 HP8656B SYN AM-FM 0.1-990Mc/S - £1.5k.
 HP8657A SYN AM-FM 0.1-1040Mc/S - £2k.
 HP8657B SYN AM-FM 0.1-2060Mc/S - £3k.
 HP8660C SYN AM-FM-PM-0.01-1300Mc/S-2600Mc/S - £2k.
 HP8660D SYN AM-FM-PM-0.01-1300Mc/S-2600Mc/S - £3k.
 HP8673D SYN AM-FM-PM-0.01-26.5 GHz - £12k.
 HP3312A Function Generator AM-FM 13Mc/S-Dual - £300.
 HP3314A Function Generator AM-FM-VCO-20Mc/S - £600.
 HP3325A SYN Function Generator 21Mc/S - £800.
 HP3325B SYN Function Generator 21Mc/S - £2k.
 HP8673B SYN AM-FM-PM-2-26.5 GHz - £5k.
 HP3326A SYN ZCH Function Generator 13Mc/S-IEEE - £1.4k.
 HP3336A-B-C SYN Func/Level Gen 21Mc/S - £400-£300-£500.
 Rascal/Dana 9081 SYN S/G AM-FM-PH-5-520Mc/S - £300.
 Rascal/Dana 9082 SYN S/G AM-FM-PH-1.5-520Mc/S - £400.
 Rascal/Dana 9084 SYN S/G AM-FM-PH-001-1040Mc/S - £300.
 Rascal/Dana 9087 SYN S/G AM-FM-PH-001-1300Mc/S - £1k.
 Marconi TF2008 AM-FM-Sweep 10Kc/S-510Mc/S - £200 Fully
 Tested to £300, as new + book + probe kit in wooden box.
 Marconi TF2015 AM-FM-10-520Mc/S - £100.
 Marconi TF2016A AM-FM 10Kc/S-120Mc/S - £100.
 Marconi TF2171/3 Digital Synthesizer for 2015/2016A - £50.
 Marconi TF2018A AM-FM SYN 80Kc/S-520Mc/S - £500.
 Marconi TF2019A AM-FM SYN 80Kc/S-1040Mc/S - £650-£1k.
 Marconi TF2022E AM-FM SYN 10Kc/S-1.01GHz - £1K-£1.2k.
 Farnell ES41000 AM-FM SYN 10Hz-1GHz - £500.
 R & S SMPD AM-FM-PH 5KHz-2720Mc/S - £3k.
 Anritsu MG3601A SYN AM-FM 0.1-1040Mc/S - £1.2k.

ITEMS BOUGHT FROM HM GOVERNMENT BEING SURPLUS. PRICE IS EX WORKS. SAE FOR ENQUIRIES. PHONE FOR APPOINTMENT OR FOR DEMONSTRATION OF ANY ITEMS, AVAILABILITY OR PRICE CHANGE.
 VAT AND CARRIAGE EXTRA. ITEMS MARKED TESTED HAVE 30 DAY WARRANTY. WANTED: TEST EQUIPMENT-VALVES-PLUGS AND SOCKETS-SYNCRONS-TRANSMITTING AND RECEIVING EQUIPMENT ETC.

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Converts your colour monitor into a QUALITY COLOUR TV!!



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CABLE COMPATIBLE*

The TELEBOX is an attractive fully cased mains powered unit, containing all electronics ready to plug into a host of video monitors made by makers such as MICROVITEC, ATARI, SANYO, SONY, COMMODORE, PHILIPS, TATUNG, AMSTRAD etc. The composite video output will also plug directly into most video recorders, allowing reception of TV channels not normally receivable on most television receivers* (TELEBOX MB). Push button controls on the front panel allow reception of 8 fully tuneable 'off air' UHF colour television channels. TELEBOX MB covers virtually all television frequencies VHF and UHF including the HYPERBAND as used by most cable TV operators. A composite video output is located on the rear panel for direct connection to most makes of monitor or desktop computer video systems. For complete compatibility - even for monitors without sound - an integral 4 watt audio amplifier and low level Hi Fi audio output are provided as standard.

TELEBOX ST for composite video input type monitors £36.95
TELEBOX STL as ST but fitted with integral speaker £39.50
TELEBOX MB Multiband VHF/UHF/Cable/Hyperband tuner £69.95
For overseas PAL versions state 5.5 or 6 mhz sound specification.
*For cable / hyperband signal reception Telebox MB should be connected to a cable type service. Shipping on all Teleboxes's, code (B)

NEW State of the art PAL (UK spec) UHF TV tuner module with composite 1V pp video & NICAM hi fi stereo sound outputs. Micro electronics all on one small PCB only 73 x 160 x 52 mm enable full software control via a simple 2 wire link to any IBM type computer. Supplied complete with simple working program and documentation. Requires +12V & +5V DC to operate.
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- 3 1/2" Mitsubishi MF355C-D 1.4 Meg. Non laptop £18.95 (B)
- 5 1/4" Teac FD-55GFR 1.2 Meg (for IBM pcs) RFE £19.95 (B)
- 5 1/4" Teac FD-55F-03-U 720K 40/80 (for BBC's etc) RFE £29.95 (B)
- 5 1/4" BRAND NEW Mitsubishi MF501B 360K £22.95 (B)
- Table top case with integral PSU for HH 5 1/4" Floppy or HD £29.95 (B)
- 8" Shugart 80/801 8" SS refurbished & tested £210.00 (E)
- 8" Shugart 810 8" SS HH Brand New £195.00 (E)
- 8" Shugart 851 8" double sided refurbished & tested £260.00 (E)
- 8" Mitsubishi M2894-63 double sided NEW £295.00 (E)
- 8" Mitsubishi M2896-63-02U DS slimline NEW £295.00 (E)
- Dual 8" cased drives with integral power supply 2 Mb £499.00 (E)

HARD DISK DRIVES

- 2 1/2" TOSHIBA (19 mm H) MK2101MAN 2.16 Gb. New £199.00
 - 2 1/2" TOSH (12.5 mm H) MK1002MAV 1.1 Gb laptop. New £115.00
 - 2 1/2" to 3 1/2" conversion kit for PC's, complete with connectors £12.95
 - 3 1/2" FUJI FK-309-26 20mb MFM I/F RFE £59.95
 - 3 1/2" CONNER CP3024 20 mb IDE I/F (or equiv.) RFE £59.95
 - 3 1/2" CONNER CP3044 40mb IDE I/F (or equiv.) RFE £69.00
 - 3 1/2" RODIME RQ3057S 45mb SCSI I/F (Mac & Acom) £69.00
 - 3 1/2" QUANTUM 40S Prodrive 42mb SCSI I/F, New RFE £49.00
 - 3 1/2" WESTERN DIGITAL 850mb IDE I/F New £185.00
 - 5 1/4" MINISCRIBE 3425 20mb MFM I/F (or equiv.) RFE £49.95
 - 5 1/4" SEAGATE ST-238R 30 mb RLL I/F Refurb £69.95
 - 5 1/4" CDC 94205-51 40mb HH MFM I/F RFE tested £69.95
 - 5 1/4" HP 97548 850 Mb SCSI RFE tested £99.00
 - 5 1/4" HP C3010 2 Gbyte SCSI differential RFE tested £195.00
 - 8" NEC D2246 85 Mb SMD interface. New £199.00
 - 8" FUJITSU M2322K 160Mb SMD I/F RFE tested £195.00
 - 8" FUJITSU M2392K 2 Gb SMD I/F RFE tested £345.00
- Many other drives in stock - Shipping on all drives is code (D)

TEST EQUIPMENT & SPECIAL INTEREST ITEMS

- mits. FA3445ETKL 14" Industrial spec SVGA monitors 1kW to 400 kW - 400 Hz 3 phase power sources - ex stock £245
- IBM 8230 Type 1, Token ring base unit driver £760
- Wayne Kerr RA200 Audio frequency response analyser £2500
- IBM 53F5501 Token Ring ICS 20 port lobe modules £750
- IBM MAU Token ring distribution panel 8228-23-5050N £95
- IBM 501 Low distortion Oscillator 9Hz to 330KHz, IEEE £550
- ALLGON 8360.11805-1880 MHz hybrid power combiners £250
- Trend DSA 274 Data Analyser with G703(2M) 64 I/O £90A
- Marconi 6310 Programmable 2 to 22 GHz sweep generator £6500
- Marconi 2022C 10KHz-1GHz RF signal generator £1550
- Marconi 2030 opt 03 10KHz-1.3 GHz signal generator, New £1510
- HP1650B Logic Analyser £3750
- HP3781A Pattern generator & HP3782A Error Detector £90A
- HP6621A Dual Programmable GPIB PSU 0-7 V 160 watts £1800
- HP6264 Rack mount variable 0-20V @ 20A metered PSU £675
- HP54121A DC to 22 GHz four channel test set £90A
- HP8130A opt 020 300 MHz pulse generator, GPIB etc £8500
- HP A1, A0 8 pen HPGL high speed drum plotters - from EG+G Brookdale 95035CS Precision lock in amp £650
- View Eng. Mod 1200 computerised inspection system £850A
- Sony DXC-3000A High quality CCD colour TV camera £1100
- Kelthley 590 CV capacitor / voltage analyser £90A
- Racal ICR40 dual 40 channel voice recorder system £3750
- Fiskers 45KVA 3 ph On Line UPS - New batteries £9500
- ICI R5030UV34 Cleanline ultrasonic cleaning system £90A
- Mann Tally MT645 High speed line printer £2200
- Intel SBC 486/133SE Multibus 486 system. 8Mb Ram £945
- Siemans K4400 64Kb to 140Mb demux analyser £2950

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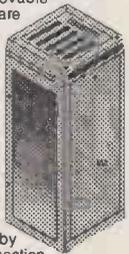
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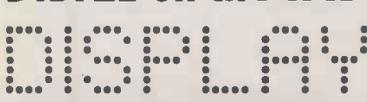
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CIRCLE NO. 125 ON REPLY CARD



SPEAKERS' CORNER



In his previous corner, John Watkinson looked at how loudspeakers can be modelled as electrical equivalents. In this one, he analyses the behaviour of real speakers based on the model.

The moving-coil loudspeaker is mechanically a simple mass, spring and damper system. But these characteristics are reflected into the load seen by the amplifier, making it complex.

Figure 1a) shows the equivalent circuit derived in my last article. It consists of a damped tuned circuit fed by a series resistance which is actually the coil resistance.

In most cases the output impedance of the amplifier will be zero because the amplifier uses heavy negative feedback which makes it a voltage source. As a result the coil resistance is essentially in parallel with the damping resistance.

In practice the damping effect of the coil resistance may be somewhat greater than the mechanical damping and so the equivalent circuit can be simplified to that of Fig. 1b).

It is important that the resistance of the cables between the amplifier and the speaker is low enough that the effective damping resistance isn't significantly changed. This is the only attribute a loudspeaker cable needs.

Boxing in

When a drive unit is mounted in a sealed enclosure, the enclosed air acts as a further stiffness in parallel with the stiffness of the drive unit. In other words the compliance seen by the moving cone goes down. This is modelled by Fig. 1c).

The result of the enclosure is that the fundamental resonant frequency of the cone goes up. Figure 2a) shows how this can be measured. A signal generator and a 1kΩ resistor provide a nearly constant current source and the impedance can be plotted by using the expression shown. Of course modern computerised speaker testers can do this with a few instructions and plot the result automatically.

Figure 2b) shows the result for a medium sized woofer. The impedance peaks sharply indicating the value of the resonant frequency. This is as

might be expected from a tuned circuit.

Above the resonant frequency, the mass component in Fig. 1c) dominates. Clearly, the impedance of the capacitor falls at 6dB/octave. As there is a series resistor R_e , the voltage across the capacitor, representing the cone velocity, also falls at 6dB/octave. Because radiation is proportional to frequency, the result is that the two effects cancel and the frequency response under mass control is uniform.

Below the resonant frequency, compliance dominates. As the frequency falls the inductors in Fig. 1c) progressively shunt away cone velocity so that it goes down below resonance at 6dB/octave. Now the radiation characteristic compounds the effect so that the frequency response falls at 12dB/octave.

Phase considerations

Under mass control, the velocity experiences a 90° phase lag whereas under compliance control there is a 90° phase lead. At resonance, the phase angle is zero. Around the resonant frequency the speaker undergoes a phase reversal. The sharpness of this reversal is a function of the Q-factor of the resonance.

Clearly, if this phase reversal is within the audio band it will do nothing for the time response of percussive transients. In other words a loudspeaker displaying resonant behaviour is not reproducing the input waveform, but is instead stamping a footprint of its own on the reproduced sound.

There are three approaches to the problem. One is to pretend it doesn't matter. Another is to make the fundamental resonant frequency so low that it is below the audio band. This is entirely feasible, but it does require physically large enclosures to avoid the compliance of the air spring becoming too low.

The third approach is to build an active speaker in which the resonant behaviour of the drive unit is cancelled

in amplitude and phase by an equal and opposite electronic transfer function. The low frequency roll-off of the speaker can then be set electronically below the audio band.

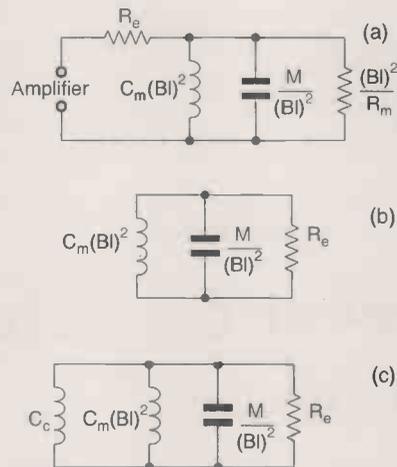


Fig. 1. Equivalent circuit of a moving-coil loudspeaker, a), equivalent where coil damping dominates, b), and drive unit in enclosure, compliance of enclosure increases resonant frequency, c).

Fig. 2. Measuring resonant frequency, a).

Cone velocity is proportional to back emf, e , and

$$e = V_s - iR_e = V_s - \frac{V_R}{1000} R_e$$

Graph b) shows the peak in response at resonance.

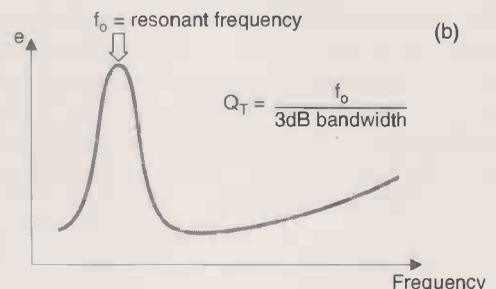
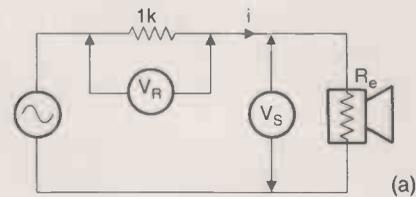


Fig. 3. Equations, a), showing that Q factor of speaker driven by amplifier is different from measured Q, Fig. 2.

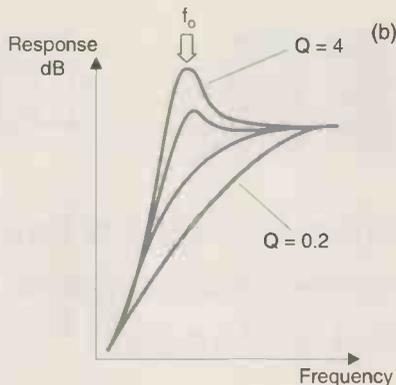
$$\frac{1}{Q} = \frac{1}{Q_T} + \frac{1}{Q_E}$$

In the above, $1/Q$ is total Q , $1/Q_T$ is mechanical Q from Fig. 2 and $1/Q_E$ is electrical Q . From Fig. 1, electrical Q is,

$$\frac{2\pi f_o MR_e}{(Bl)^2}$$

$$\therefore \frac{1}{Q} = \frac{1}{Q_T} + \frac{(Bl)^2}{2\pi f_o MR_e}$$

Graph b) shows how Q factor affects low-frequency response.



Several circuits have appeared in this magazine from time to time that do just that and the results are well worth the effort.

The arrangement of Fig. 2 is designed to show the resonance clearly. Because a current source is used, the Q factor measured there is not the Q factor which will result when driven by an amplifier. The electrical Q due to the damping of the coil resistance appears in parallel with the Q of Fig. 2.

Figure 3a) shows how the electrical Q factor

is calculated once the resonant frequency is known. Figure 3b) shows the effect of different Q factors on the frequency response.

No more juggling Bl

In a traditional passive speaker, the designer had to juggle the Bl product to avoid an obviously 'honky' high Q response, but also to avoid a premature roll-off of response due to a low Q factor.

The efficiency of a loudspeaker goes as the

square of its Bl product. Nowadays, using rare-earth magnets, some pretty impressive Bl products are now possible as a matter of straightforward design, resulting in highly efficient drive units.

Unfortunately the designer of the passive speaker can't use them. Such speakers have very low Q and suffer premature roll-off. However, this is not a problem in the active speaker, which simply equalises. There is thus a case for designing drive units specifically for active speakers. In my experience, commercially available drive units designed for passive use are seldom optimal for active applications.

Fortunately, designing a drive unit for an active speaker is easy, because the exact value of the parameters is not particularly important. When the parameters are being electronically equalised, what matters instead is consistency from one unit to the next.

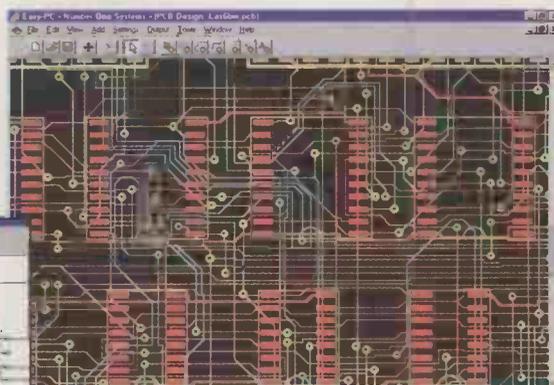
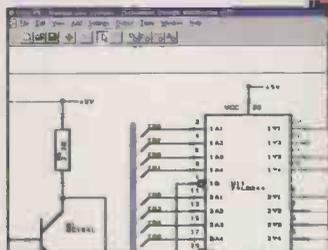
By using a small enclosure, the compliance of the air spring dominates and this swamps variations in drive unit compliance. Moving mass and cone area will not change, nor will Bl or R_e and so an equalised speaker should not go out of adjustment as it ages.

Time saved designing the drive unit can be used to optimise other parts of the system. ■

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The instrument has a dynamic range of 70dB with a typical accuracy of 1.5dB. Its frequency range is from 400kHz to over 1GHz, and its bandwidth – i.e. selectivity – is 250kHz.

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Centre adjust	0MHz to 1000MHz
Bandwidth	250kHz (-6dB typical)
Meter accuracy	1% of reading +1MHz
Calibration marker	50MHz fundamental, harmonics to 1GHz
Scan width	1MHz to 100MHz/div
Scan speed	0.5ms to 35ms/div
Amplitude	
Input impedance	50Ω
Amplitude range	-70dBm to 0dBm nominal
Amplitude scale	Logarithmic, 10dB/div
Amplitude linearity	Typically ±2dB
Amplitude flatness	Typically ±1.5dB
Max. input level	+10dBm
Calibration marker	-30dBm ±1dB at 50MHz

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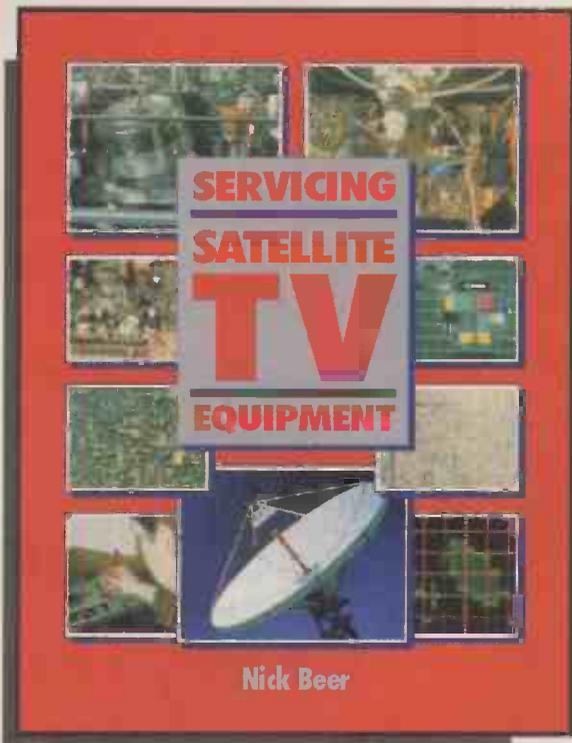
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CIRCLE NO.127 ON REPLY CARD

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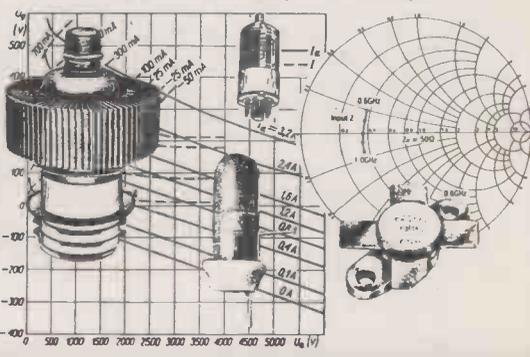
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Marble FX trackball

Mouse holes

As I pointed out in last month's article, there are several advantages to staying with a mouse-type device. The *Digital Pro* mouse uses a dedicated mat with a closely-spaced line grid to generate the digital signal. The grid is scanned with two fibre-optic coupled sensors, one at the top right and one bottom centre of the underside of the mouse.

Half of the grid on the mat has vertical lines and the other half has horizontal lines. One sensor only generates a signal from any side-to-side movement and the other only generates a signal from up-and-down movement. With this method a resolution of 525dot/in can be achieved – an accuracy significantly better than most mechanical-optical mechanisms, and with impeccable consistency and reliability.

The digital system continues to func-

Like many engineers using CAD on a daily basis, Rod Cooper soon became frustrated with the conventional mouse. So he set out to find a better solution. In this second article, Rod looks at a novel trackball and a high-performance graphics tablet. But first, the mouse with no moving parts.

tion normally up to 45° either side of the vertical axis of the mat, which accommodates most likely mouse movements. However, if your natural stance is with the mouse skewed on the mat, there is nothing to stop you rotating the mat to bring it back to normal. This gives you maximum mouse travel.

You might think that with two separate sensors, the digital mouse functions differently from its conventional cousin. It had a slightly different feel, but the handling of the digital mouse seemed surprisingly similar. I found I did not have to alter my style.

One digital mat is supplied, and consists of a plastic laminate line pattern on a foam rubber base, measuring 155 mm by 175 mm. These dimensions are smaller than an ordinary mat, which needs to allow for slippage of the rotating ball mechanism on the mat surface. There is no slippage with the digital

system; returning the digital mouse to given position on its mat brings the screen pointer to exactly the same position time and time again – provided you don't lift the mouse off the surface. If you take the digital mouse off its mat, you will find that its motion is relative, just like a conventional mouse.

One benefit of a small but fully utilised mat is that it can be placed in front of the keyboard without taking up much room. Also, I tried it with one of the lap-mounted mouse-mat pads designed to relieve muscle stress and found the small size a distinct advantage. There is more on mouse ergonomics in last month's article.

Details are given of how to produce a new mat should the one supplied wear out. This overcomes one of the snags of previous digital mouse designs, which involved the purchase of a new mat.

In use, the screen pointer should be roughly centred, then the mouse should be moved to the centre of the mat to start work. The maximum vertical movement of the device is determined by the distance apart of the sensors, and is about 70mm total.

The maximum horizontal movement is determined by the dimensions of the pad, about 100mm. The overall movement can be configured by the driver software settings, and some experimentation is needed to get the configuration right for your own personal preference, as with any mouse.

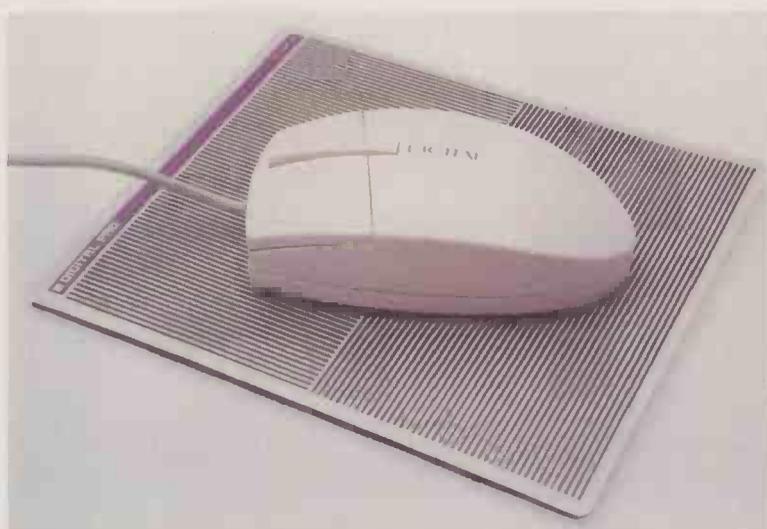
Click-selectable modes

Although it uses the standard Microsoft mouse driver, this mouse can be operated in three modes. It also has a third, middle button mounted just right of centre.

In standard mode, the speed or gearing between movement of the mouse and pointer is the same as for a conventional mouse. This is useful for general use on word processors, spreadsheets, etc.

The second mode, which is accessed by pressing the middle and the right buttons together for one second, is intended for more detailed work. In this mode, the mouse needed to move more for the same pointer movement. I found this useful for some aspects of pcb-CAD, like wiring up the schematic symbols.

To get out of this mode, you simply press the middle button for more than one second. In practical terms, it is very easy to enter and leave these two modes to do different types of work inside one program. The main mouse



The Digital Pro mouse looks similar to an ordinary mouse but note the third middle button, which provides a double-click, and the special gridded pad.



Look underneath and you will see no rubber ball. The two fibre optic sensor points can be seen at the bottom and top.

control is still held by the Windows driver menu so you can still adjust overall 'speed' of the pointer, etc. to your liking.

The third mode, accessed by pressing the middle and left buttons for a second, emulates a joystick – insofar as this is possible with a mouse. The literature claims that joystick mode is good for playing games like Doom, Quake, etc.

The middle button performs another function, that of a generator of double-clicks. If you have a program that requires many double clicks as part of its normal operation, like the wiring-up and manual tracing in *Easy-PC*, then this feature is a practical addition.

In summary

This digital mouse has no intrinsic wear mechanism and performs consistently without cleaning. Where CAD and graphics are concerned, being able to switch rapidly from standard to graphics mode and back again is useful, as is the double-click feature. Although the price is more than an

Review subjects

Digital Pro mouse. Supplier of the Digital Pro ball-less mouse is Verkonix Ltd, tel/fax 0121 354 5569. Price £25.

Easypen. Produced by Genius-Kye, the *Easypen* graphics tablet was reviewed last month and is supplied by Laton Technology, phone 01424 422562, fax 01424 423460. Price £29.50.

PenPartner. Made by Wacom, the *PenPartner* graphics tablet is supplied by Computers Unlimited, tel 0181 358 5857, fax 0181 200 3788. Price £74.95.

Tabby 2. Maker and supplier of the *Tabby 2* graphics tablet reviewed last month is Micrograf International Ltd, tel 0181 838 3750, fax 0181 838 3650. Price £49.95.

Trackman Marble FX. Logitech manufactures the *Trackman Marble FX* trackball, which is available from PC World. Price £49.



Increased productivity. As mentioned in last month's article, I experimented by replacing the two push buttons on a graphics tablet pen with a pistol grip designed to be used in the left hand, as shown. This made using the tablet much easier and faster.

ordinary mouse, its longer life should offset this.

Supplied by Verkonix Ltd, tel/fax 0121 354 5569 at £25 in the UK or by Good Systems Inc in the US, tel. 1 408 739 4713, fax 1 408 739 4702.

Trackman Marble FX

Logitech has devised an interesting solution to the problems that beset the conventional trackball with a technique it calls *Marble*.

Shown on the first page of this article, the *Marble FX* trackball replaces the revolving spindles and interrupters found on conventional trackballs with an all-electronic system. The only moving part is the ball itself, which rests by gravity on a three-point suspension.

The 52mm diameter ball is covered in numerous randomly-spaced black dots. These dots and the background are designed to give maximum contrast to infra-red light. The trackball is illuminated from within the trackball body

by two infra-red leds, via a window, and the reflected light focussed by a mirror/lens system onto a sensor. This sensor consists of 93 small photocells, and the output is passed to a microprocessor.

The cells work independently in a neural network and the average displacement of the dots from all these cells, caused by rotation of the trackball, is computed.

The image is sampled at 1kHz and any shift between successive samples reported to the microprocessor. If there is any dirt or scratches on the trackball, or on the window to the optical system, the collective calculation from this system is only marginally affected. This will be below the level of perception of the user – i.e. you will notice no error in operation. At this rate of sampling the response to movement is as good as any mouse or tablet.

To protect the dot array, the trackball is given an infra-red transparent coating. This gives the trackball a smooth glossy finish that is easily cleaned. The

ball is readily removed from the body of the device by gently pressing it out. This also exposes the optical window for cleaning. Dirt is simply wiped off, which, compared to a conventional trackball, is very simple.

The trackball can be operated by lightly gripping the ball between thumb and forefinger – or between thumb and the first two fingers. A glance at the photo will show this design is intended for right-hand use.

The ball is so freely suspended that it can also be operated by just the thumb or one finger. Very little effort or grip is needed so operation is very relaxed, and knowing that there is not going to be any jerkiness or other malfunction helps to maintain this. Weighing 260g, the base of the trackball rests solidly on the desk.

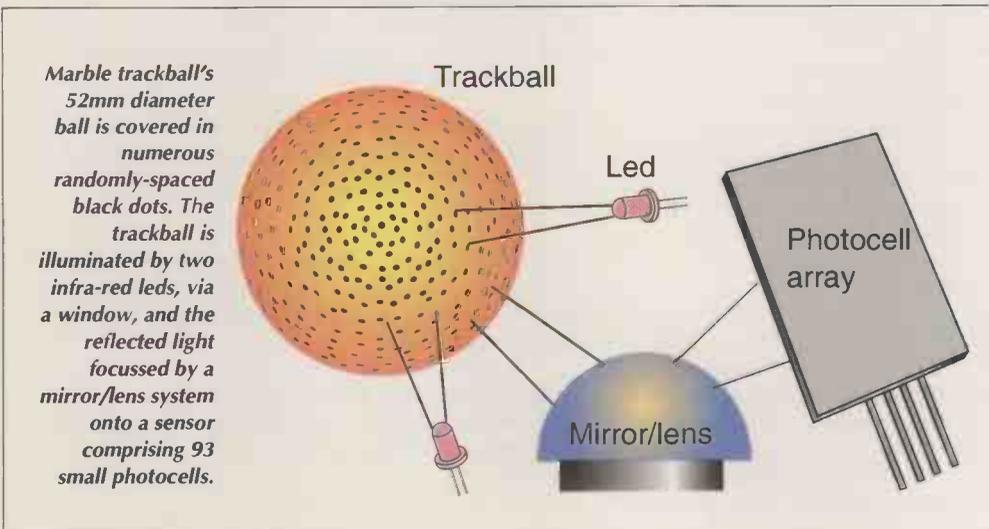
Marble FX can be connected as a direct replacement for a mouse on a serial port or a PS/2 socket, so the existing Windows mouse driver may be used. You will have normal pointer control and left and right clicks covered by the trackball's lower left and right-side switches respectively. But in this mode you will not take full advantage of the extra features provided by Logitech.

The device has four buttons in all, and they have an extensive range of configurations if you use the software provided. This comes on two 3.5in disks and is easily installed. Although advertised as a Windows 3.1 and 95 product, there is a readme file in the driver software describing Windows 98 and NT operation.

With this software, the buttons can be configured for such things as scrolling, escape, drag-lock, the F1 to F12 functions, and many more, as well as controlling the pointer style, speed and acceleration.

Also included are two aids to quicker operation, namely 'HyperJump' and 'CyberJump.' Basically, both of these bring together eight of the most commonly used functions on the screen in a small window. These functions include for example closing and scrolling. In this way, you don't have to travel to various parts of the screen to execute the commands. CyberJump is used for working the Internet, and HyperJump for Windows operations.

I checked out the effects of heavy grease and dirt contamination from the hand and the system did not falter. This bears out Logitech's claim to have overcome the dirt problem.



In summary

Logitech's *Marble FX* trackball is a well-constructed device. A lot of effort has clearly gone into the ergonomics of this design, successfully in my view. Operation is smooth and precise and noticeably consistent when compared to a conventional trackball.

If you currently use a trackball, this is the natural up-grade. Note that other design styles are available from Logitech using the Marble technology.

Price; £49, supplier; PC World.

PenPartner

Made by Wacom, *PenPartner* measures 182 by 205mm and has an active area of 96 by 128mm.

The tablet part is noticeably slimmer than the other tablets reviewed, and heavier than comparably-sized products, weighing in at 450g. The weight gave it noticeably more stay-put stability when used with a desk-top set-up.

Driver software for Windows 3.1, 95 and NT comes on a cd and installation was easy. The pen is electromagnetically coupled and has a range of about 10mm. The resolution is 1000line/in.

The product comes with a very brief installation guide and a small booklet describing the tablet's capabilities. There's also a spare nib for the pen.

Pressure-sensitive and cable-free pen

This digitising tablet represents the next level of sophistication in the review, the pen being both pressure-sensitive and cable-free. The pen is self-powered and does not require batteries.

The body of the pen is shaped to give a waist where the fingers go, with extra thickness further up the pen, where it is about 11mm diameter. I found this design very comfortable to work with. A rocker-type switch on the barrel can be configured to give right, middle, left single or double-click, or keyboard controls. To place the pen out-of-range, a free standing inkpot-type pen holder is provided.

The combined left switch and pressure sensor is in the tip of the pen, so pressing on the tip gives the left-click, and pressing harder gives a thicker line, more intense colour, or both, provided the application software will support this mode.

The reverse end of the pen has an eraser built in, so you do not need to select the erase function from the pro-

gram, you simply turn the pen round. With pressure-sensitive erase, it is possible to create shading effects. The eraser and pressure tip each provide 256 levels of pressure, and both features can be independently programmed to give the pen a soft or hard feel.

For the pressure features to be of use, the application program has to have this control built in, so this tablet is available either on its own or bundled with special editions of a photo-editing program and a paint-type program at nominal extra cost.

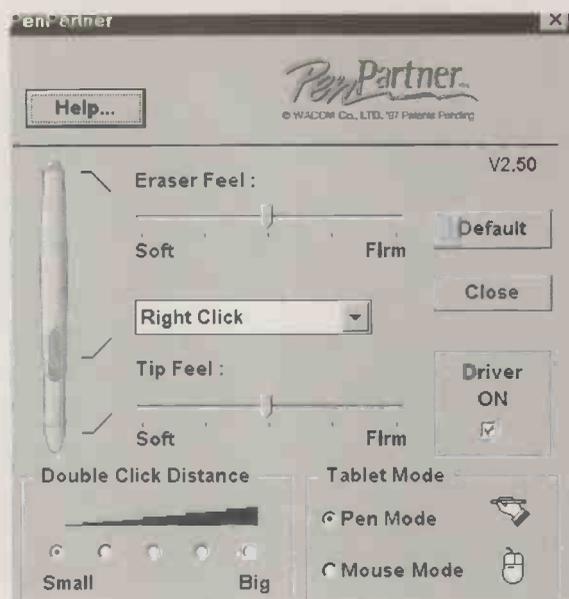
I tried out the programs provided and the system worked well. A further touch of realism is provided on the tablet surface, which comes close to resembling the feel of pen on paper. Of the tablets reviewed, this felt the most natural when used for graphics.

With a typical pcb-CAD program, the tablet performed very much like its non-pressure sensitive cousins, with good positive action of the left and right clicks from tip and barrel switches.

The pc connection is different from that on the other tablets reviewed. On an AT machine, an in-line adaptor fits between the keyboard socket and the keyboard, and an additional lead is plugged into the 9-pin serial port, i.e. two connections to the pc.

On a machine with a PS/2 port, the adaptor fits in between the PS/2 port and the mouse, the extension going again to the 9-pin serial port. A single lead goes to the tablet and is permanently cabled. A led on the tablet shows that this set-up is working.

The tablet co-existed with the mouse,



but if you wish, you can configure it to replace the mouse as the pointing device.

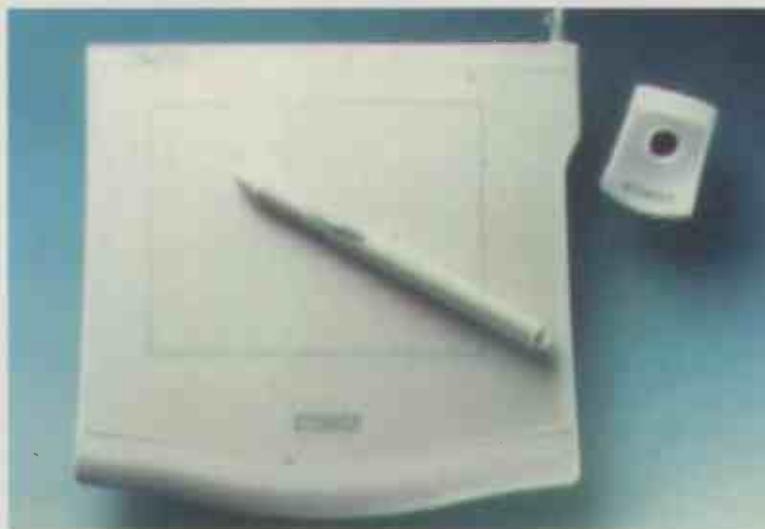
In summary

PenPartner performed well with pcb-CAD. No pcb-CAD program that I am aware of is pressure-sensitive, but as I pointed out in the introduction, if you use your pc for graphics as well as pcb-CAD, or if the pc is shared with another person in your company or household using graphics applications, it would make good sense to opt for this type of tablet.

A lot of design effort has clearly been expended to make this tablet act and feel like pen and paper and in this respect it is successful.

Supplier; Computers Unlimited, tel. 0181 358 5857 fax 0181 200 3788, e-mail general@unlimited.com. Prices are £74.95 for *PenPartner*, or £84.95 for *PenPartner* bundled with pressure-sensitive versions of Kai's Photo Soap 1.0SE and Dabblor 2.0SE. ■

PenPartner's write and erase functions are both pressure sensitive and adjustable via software.



PenPartner's cable-free and pressure-sensitive pen gives it more sophistication.



Hands-on Internet

Cyril Bateman presents a variety of sine-wave generation alternatives found on the net - including digital options.

I am currently designing a piece of equipment that needs a compact and stable sine-wave generator operating at low audio frequencies with less than 1%V distortion. For the development work, I have been using a separate generator, but eventually the equipment will need its own oscillator.

This generator has to drive a load with a very high capacitance, hence it needs a low output impedance. Since the instrument is portable, battery consumption is important too.

I had some ideas how I could achieve these targets, but I decided to see if there were any interesting alternatives on the Internet. Since my workshop's elderly, audio generator also needs updating, I decided to broaden my search to include very-low-distortion sine-wave oscillators.

Sine wave generators

Perhaps the most difficult part of any sine-wave generator is ensuring a constant amplitude output. Often this is

provided by a feedback AGC loop, which might be based on thermistors, lamps with low filament mass, variable resistance FET circuits or clipping diodes.

To stabilise amplitudes at the lowest

frequency needed, these control methods invariably use long time constants. Amplitude bounce on changing frequency is then almost inevitable. If the circuit has insufficient positive feedback, oscillation start-up

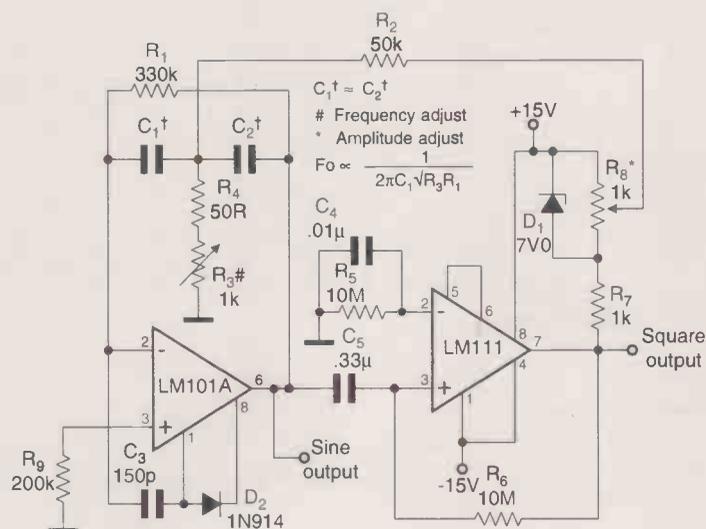


Fig. 1. Described as easily tuned, this filtered square-wave approach to sine-wave generation can be tuned over a four to one range of frequencies by adjusting a single resistor. Distortion is less than 1%.

Fig. 2. Eight times over-clocking, this simple circuit produces a very low distortion sine wave at any one of 128 discrete frequency steps between 1kHz and 25kHz. All harmonics are more than 80dB down.

can be erratic. Too much feedback on the other hand increases distortion.

A popular alternative way of obtaining a nominal 1% distortion, constant amplitude and stable sine wave is to remove harmonics from a square wave using either low-pass or band-pass active filter circuits. A constant amplitude, stable frequency square wave is extremely easy to generate without using AGC loops.

A Burr-Brown application note¹ describes in detail how this filtering can easily be provided using their dedicated UAF42 integrated-filter circuit. An

easily obtained Q of 10 filter attenuates the third harmonic by 28dB. This ensures that the filtered amplitude of the third harmonic is some 40dB below the fundamental, resulting in a sine-wave output having a nominal 1% distortion. Higher filter circuit Qs can provide much lower distortions.

Filter42 is part of the Burr Brown 'FilterPro' software package, which can be quickly downloaded. Dedicated to designing filters using the UAF42 integrated circuit, Filter42 simplifies this design task.

A second filter design software

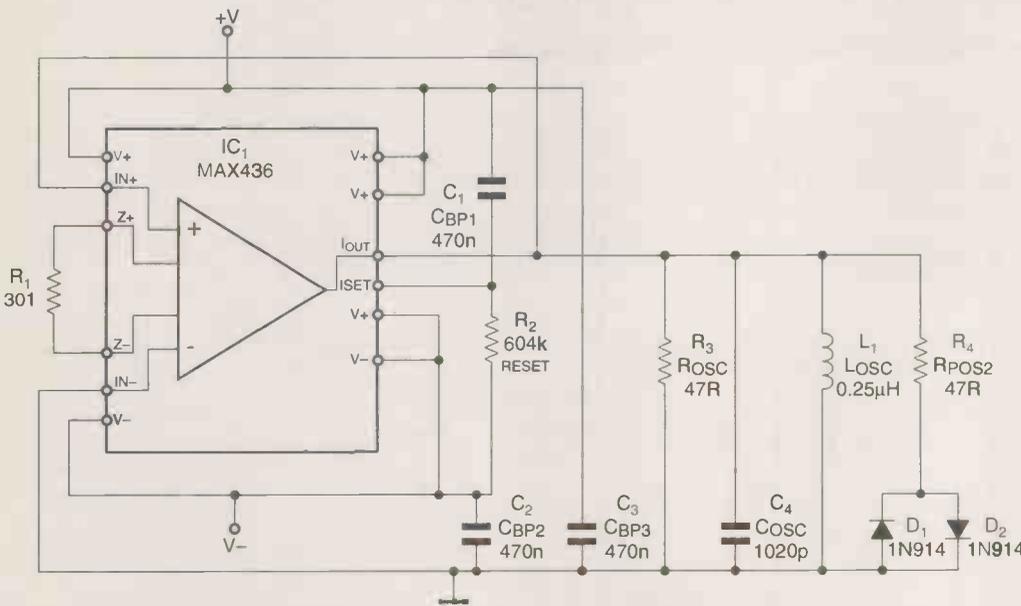
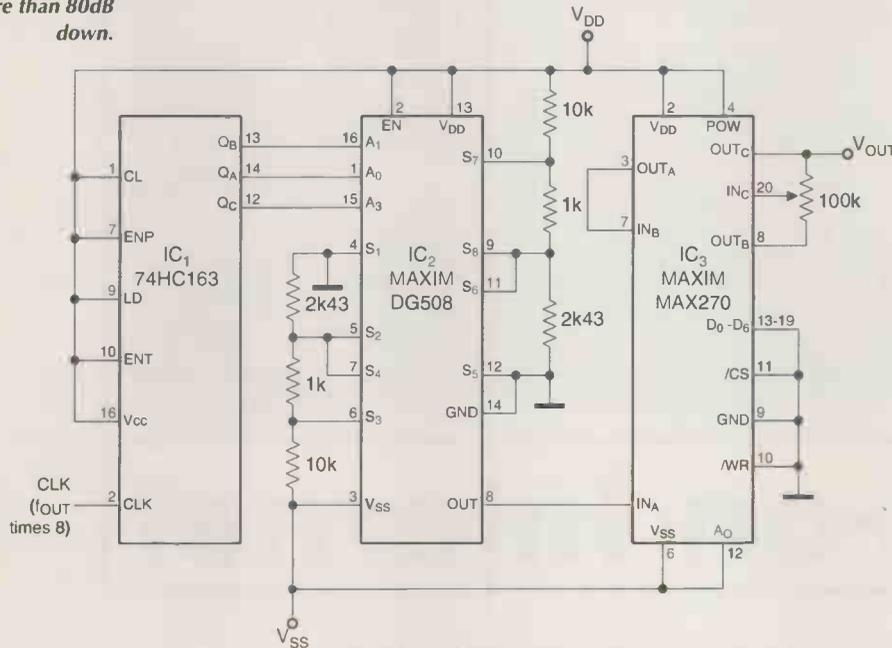
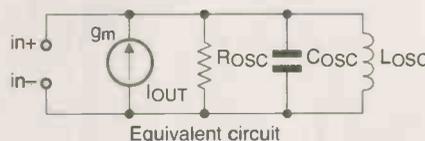


Fig. 3. This unusual 1% distortion LC oscillator performs up to radio frequencies. Losses in the L₁/C₄ tank circuit are cancelled using the Max436's ability to generate a controlled negative resistance.

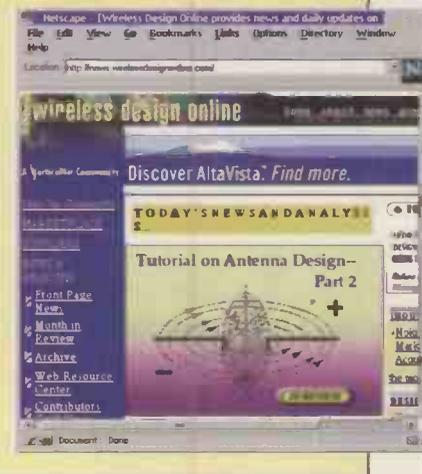


Wireless design

Wireless design online is a meeting place for readers interested or involved in the design of wireless systems. It is strong on tutorial articles, and has recently covered antenna designs in some depth. Perhaps of most interest to new visitors, is the easy access to its archives of past topics this site provides.

I 'signed up' with this community some months ago. Since then I have received regular e-mails with a synopsis of its latest news and topics, together with useful references to other sites.

Wireless Design Online is at <http://news.wirelessdesignonline.com>



program called Filter2, completes the FilterPro package. Filter2 is used when designing Butterworth, Chebyshev or Bessel filters in the Sallen & Key and MFB filter styles.

This filtered square-wave approach is excellent when a fixed frequency sine wave is needed. Change of oscillator frequency requires a corresponding change of filter frequency, which can be difficult.

Filtered square-wave oscillators

One of the first and oldest application notes to catch my attention provided a simple method of varying the filter's frequency.

Dating from March 1971, Linear Brief 16 from National Semiconductor² describes two very simple 1% distortion circuits. Both can be continuously tuned over a 4:1 frequency range using only a single variable resistor. They provide stable amplitude outputs with change of frequency, avoiding the bounce often associated with slow AGC loops and common in many designs.

These circuits use a variable-

resistance *RC* active filter circuit to remove harmonics from a square wave or clipped sine wave, leaving only the fundamental with noise and distortion. Both designs have feedforward compensation to speed up the relatively slow op-amps then available. This compensation method could be dispensed with however using modern op-amps.

The circuit shown, Fig. 1, differs from the usual arrangement used with filtered square waves, where the square wave oscillator determines the frequency. This circuit uses one op-amp as the active frequency determining filter. The second forms a comparator to square up the sine-wave output from the first stage.

Changing the filter frequency changes the frequency of the sine wave and the square wave in the circuit shown. Both waveforms are available for output.

Low THD to 25kHz in 128 steps

Another way that filter frequency can be changed is described in the 'Three ICs produce pure sine waves' application note from Maxim.³ The circuit described produces very low distortion, variable frequency sine waves.

Although the output frequency is not continuously variable, it does provide 128 discrete frequency steps, from 1 to 25kHz, and at much lower distortion than the circuit of Fig. 1.

An eight-channel *DG508* analogue multiplexer is driven from the *Q_{a-c}* outputs of a *74HC163* counter. The *DG508* generates an eight-times oversampled staircase approximation to a sine wave. This is filtered and smoothed by the *MAX270* circuit of two second order, continuous-time Sallen & Key filters, Fig. 2.

The *MAX270* filter cut-off frequency depends on the logic levels at its *D₀₋₆* inputs. Tying these all to ground as in the drawing sets the filter cut off to 1kHz. This makes the input clock required 8kHz.

With eight times oversampling, the third harmonic is 50dB below the fundamental, reducing the degree of filtering needed. The first significant harmonic is now at seven times the fundamental frequency. Having been filtered, all harmonics in the output stage are more than 80dB down, resulting in a low distortion, low noise, sine wave.

These two quite different examples illustrate how filtering a square wave can be extremely useful at audio frequencies. When higher frequencies are required, resonant inductor/capacitor circuits provide a practicable alternative.

At the heart of many high-frequency sine-wave oscillators is a parallel-

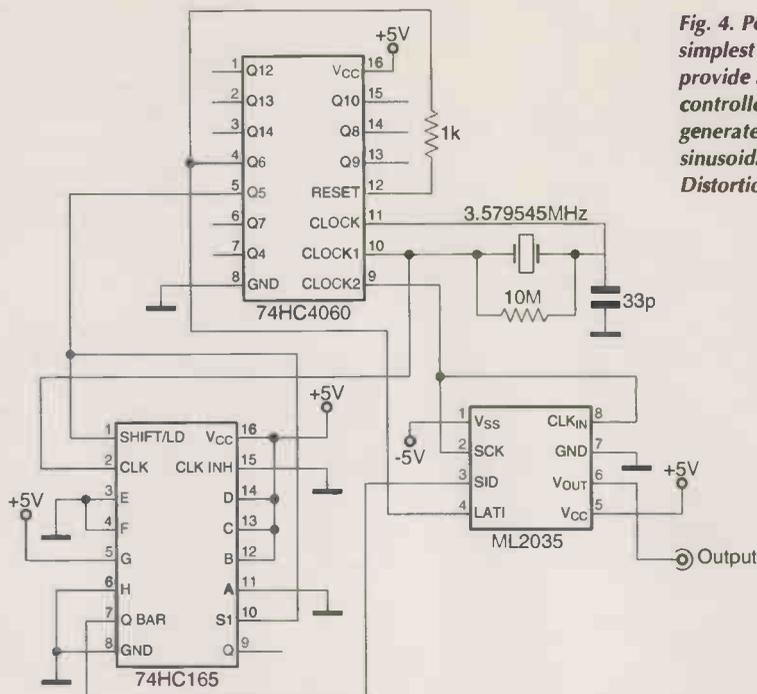
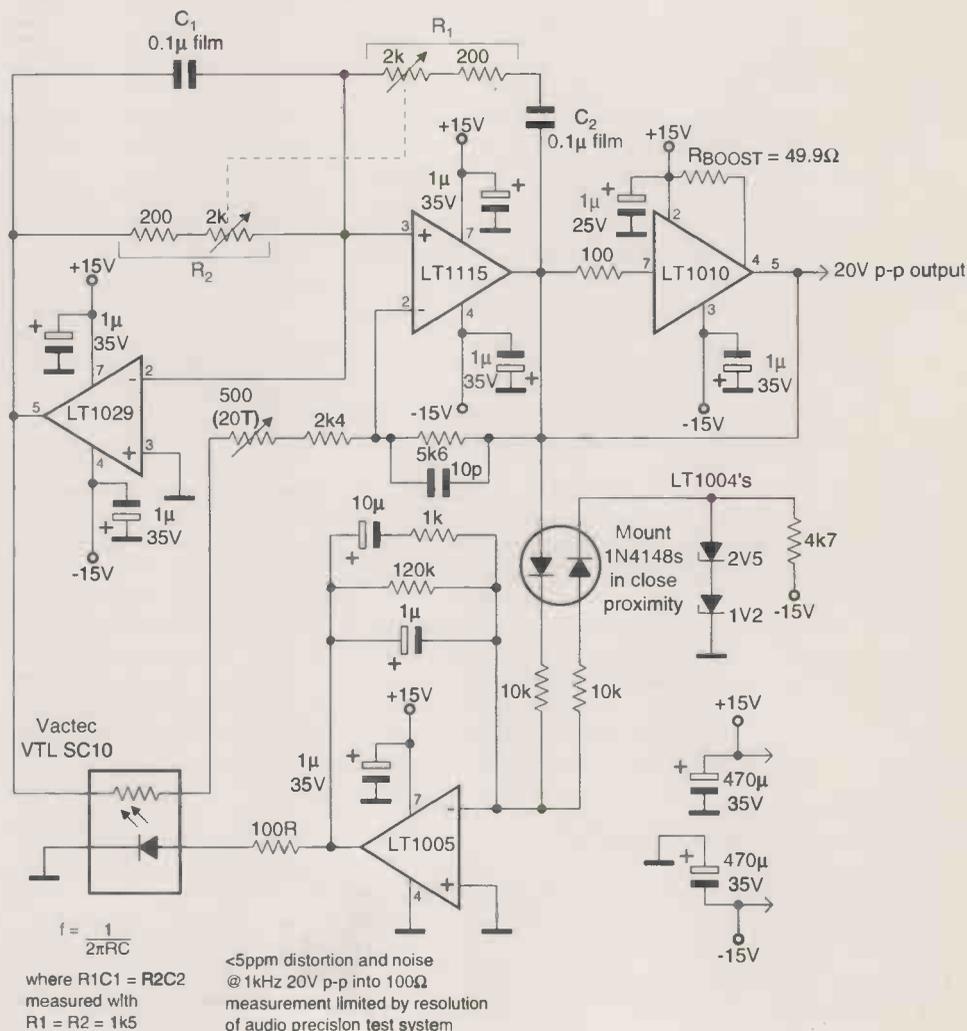


Fig. 4. Perhaps the simplest possible route to provide a stable, crystal controlled, digitally generated and accurate sinusoidal waveform. Distortion is typically 1%.

Fig. 5. Described as an ultra-low-THD sine-wave oscillator, this version of the Wien bridge claims to produce less than 5ppm - i.e. 0.0005% - distortion and noise.



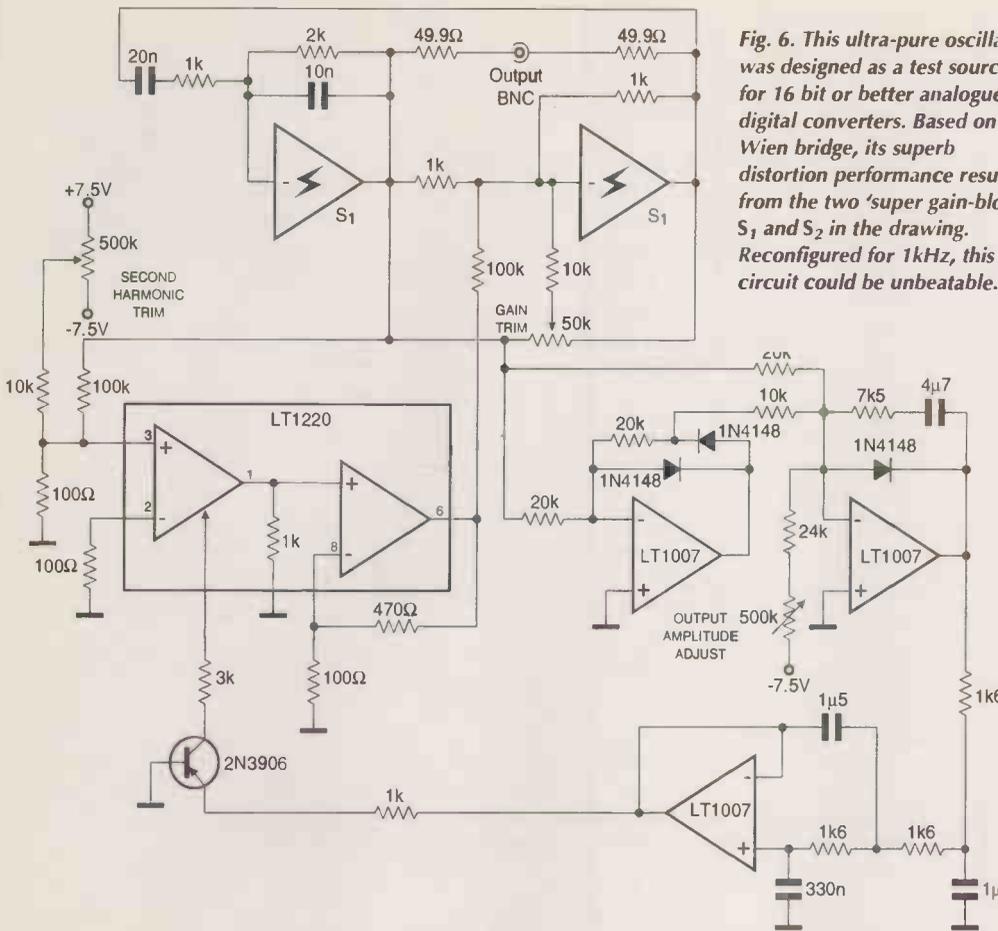


Fig. 6. This ultra-pure oscillator was designed as a test source for 16 bit or better analogue-to-digital converters. Based on a Wien bridge, its superb distortion performance results from the two 'super gain-blocks' S₁ and S₂ in the drawing. Reconfigured for 1kHz, this circuit could be unbeatable.

resonant LC tank circuit. A gain stage controlled by a feedback AGC loop, is used to overcome losses in the LC tank circuit. But there is another way.

Due to resistive losses in the capacitor and inductor, unaided the oscillations in the LC tank circuit would die away. Application of a negative resistance, equivalent in value to these resistive losses, should allow oscillation to continue indefinitely. Negative resistance can be synthesised quite easily using a transconductance amplifier.

The MAX436 and its companion MAX435, are high speed, wideband transconductance amplifiers having true differential, high impedance inputs and a current output.⁴ Their unique architecture provides accurate gain control without using overall feedback, eliminating closed-loop phase shifts.

A single gain-control resistor across the two secondary Z inputs controls the

circuit's gain. The MAX435 integrated circuit has a 275MHz bandwidth and 53dB CMRR at 10MHz. While it has a low noise figure of 7nV/√Hz at 1kHz it also provides an 850V/μs slew rate.

The application example shows the circuit arranged as a 9.3MHz oscillator with a 50Ω output, producing less than 1% distortion – mainly third harmonic. Higher oscillation frequencies are possible, provided a ground plane layout with short component connections is used, Fig. 3.

Generating waveforms digitally

When your circuit already incorporates a control processor, digital waveform generation becomes most attractive. In addition to being cheap, digital methods can generate accurate sine waves – or any other desired shape of wave.

Digital wave-shaping generators access a look-up table of X, Y co-ordinates held either in volatile or non-

volatile memory. These X, Y co-ordinate values are passed in sequence to a digital to analogue converter. Stepping through this co-ordinate table step by step can generate any desired waveform.

Microlinear provides two programmable digital generator integrated circuits, namely the ML2035 and ML2036. These have built-in 512-point sine look-up tables, dedicated to producing an accurate sine waveform. Both use a 16-bit serial data-word input.

The chips are intended to generate low-cost but accurate tones for telecommunications use and modem applications. Housed in an eight pin package, the ML2035 includes this look up table, an eight-bit d-to-a converter and a smoothing filter, Fig. 4.

Originally intended for use under microprocessor control to provide multiple frequencies, they can be used without a processor to generate a single frequency. An application note⁵ gives details how to generate accurate 50, 60, 400Hz and 1kHz sine waves, at around 1% distortion, using only simple logic circuits. This distortion could be further reduced using external filter stages.

Modern digital waveform generators can produce the widest variety of wave shapes. When the lowest possible distortion sine wave at audio frequencies is needed however, many designers still favour the Wien bridge circuit.

And back to Wien

I was particularly intrigued by two Wien bridge low-distortion circuits. The first a low-distortion, variable frequency generator, using only four integrated circuits, claimed less than 5ppm, or 0.0005% distortion at 1kHz. The second circuit, designed for 10kHz, was much more complicated, needing ten integrated circuits. Its distortion though is claimed to be 'unmeasurably low.'

An 'ultra-low THD oscillator' is described in Linear Technology's data sheet for the LT1115 op-amp.⁶ Comprising LT1022, LT1115, LT1010 and LT1006 amplifiers, this circuit used a Vactec 5C10 pre-packaged LED/LDR combination in its amplitude control circuitry.

While claiming less than 5ppm distortion, apparently the circuit's actual distortion measurement in practice was limited by the resolution of the Audio Precision test set used, Fig. 5.

99.9999% pure

Application note 67 also from Linear Technology⁷ details 'An ultra-pure oscillator' by Dale Eager. This 10kHz circuit was designed as a test source for use when calibrating 16-bit analogue-to-digital converters. Its distortion is stated

Site references

1. Application bulletin AB058.PDF
2. Linear Brief 16
3. Three ICs Produce Pure Sine waves, A0415.PDF
4. LC oscillator has 1% THD, A1616.PDF
5. Micro Linear Application note 14
6. Linear Technology Corp., LT1115
7. Application Note 67, AN67.PDF

- <http://www.burr-brown.com>
- <http://www.national.com>
- <http://www.maxim-ic.com>
- <http://www.maxim-ic.com>
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- <http://www.linear-tech.com>
- <http://www.linear-tech.com>

as beyond available measurement capability, probably into the parts per billion range, or much less than 0.0001%.

The key to this exceptional performance is the 'super gain block', identified as S_1 and S_2 in the diagram. These gain blocks comprise an LT1007 with two LT1230 separated by a passive CR bridged-T phase-correction network, Fig. 6.

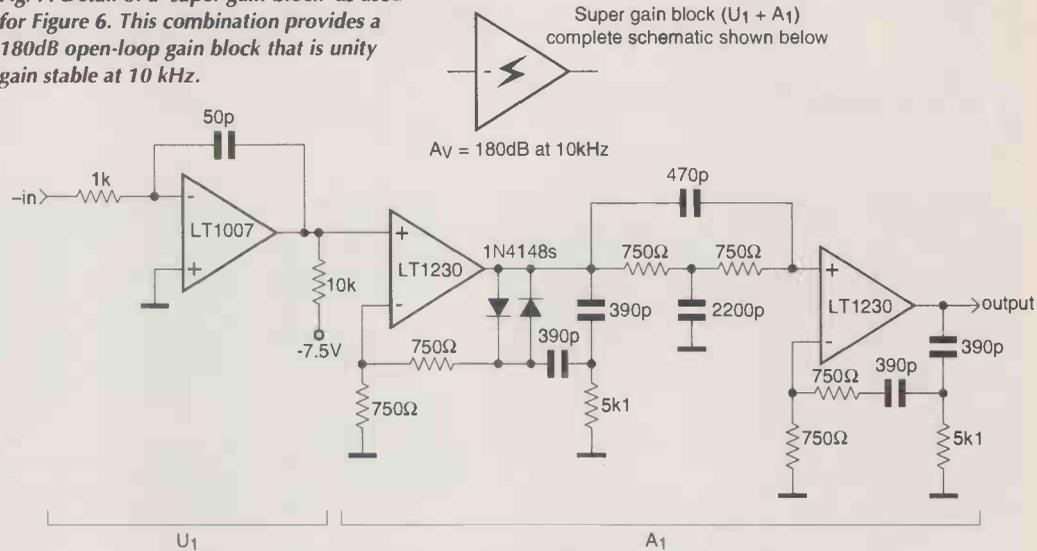
Unity gain stable, this composite super gain block has an open-loop gain of 180dB at 10kHz. This suggests that the Wien bridge distortion should be in the parts per billion range, Fig. 7.

Readers interested in following up this approach should download a copy of application note AN67. This 88-page 1.8Mbyte download covers many topics. The oscillator circuit will be found on pages 62 to 65. Readers not having Internet access can request the printed copy from Linear Technology.

Which one did I choose?

Having explored these options, just what method did I finally choose for my design ?

Fig. 7. Detail of a 'super gain-block' as used for Figure 6. This combination provides a 180dB open-loop gain block that is unity gain stable at 10 kHz.



For ease of design and low battery consumption, I stuck with my original plan. This is a CMOS square-wave oscillator driving a band-pass filter to generate the sine wave. The output stage is a dual op-amp driving a miniature push-pull

output transformer to provide the required high current, low impedance output.

However my searches did uncover some thought provoking ideas for updating my old laboratory signal generator. ■

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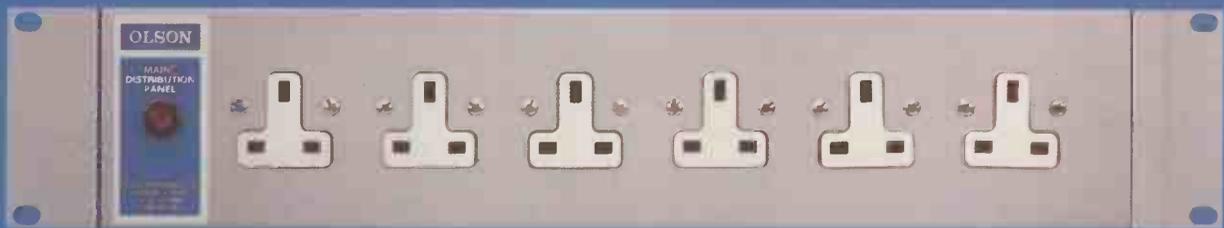
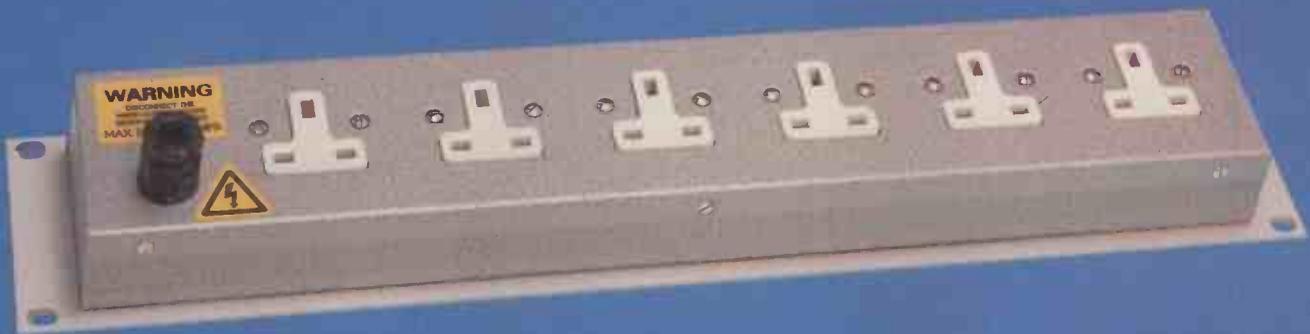
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Connectors and cabling

Fast connectors. SpeedPac from Siemens is a connector system for use with high speed data transfer up to 2.5Gb/s. This system has been achieved without the use of extra pins, which have in the past required greater insertion forces. A clamp design opens the connector to allow insertion of a daughter board, a lever then closing the connector. There are no male and female contacts; the design is that of a beam-on-pad type using only one spring-loaded contact pressing on the daughter card contact pads and those of the backplane.

Quiller Switches Ltd. Tel., 01202 436777; fax, 01202 421255; e-mail, sales@quiller.com; web, www.quiller.com.

Enq no 510

Data converters

16-bit a-to-ds. Crystal analogue-to-digital converters from Cirrus Logic, namely CS5521/23, minimise the need for external buffers and registers. Each has a low-drift, chopper-stabilised amplifier, a multi-channel multiplexer and a charge-pump drive with a programmable gain stage; they are resistant to 50/60Hz±3Hz noise and cope with the full industrial temperature range. These devices are capable of precise voltage measurement with full-scale ranges from 25mV to 5V. The voltage reference is 5V straight from the supply. Data settles in a single conversion, so that the converter automatically samples multiple channels and stores the data in an on-chip buffer, the microcontroller not being involved.

Cirrus Logic Inc. Tel., 01727 872424; fax, 01727 875919.

Enq no 502

Displays

Industrial lcds. NEC has new flat-panel displays for industrial use that are bright and have wide viewing angles. NL8060AC31-12 is a 12.1in type with 800 by 600 pixels, taking 6-bit digital RGB input. Brightness is 250cd/m². A 10.4in model, the NL6448AC33-24, has 640 by 480 pixels and a viewing angle of 80° away from the normal in any direction, with no preferred viewing angle. Brightness is 190cd/m² and the contrast ratio is 150:1.

Sunrise Electronics Ltd. Tel., 01908 263999; fax, 01908 263003; web, www.sunrise.co.uk

Enq no 503

Digital video interface. As lcds are beginning to rival crts as regards the cost and performance in pcs, the need for a digital interface between digital video on a pc and the digital lcd has been addressed by REP Design. The result is *PanelLink*, which has been adopted by VESA and some leading makers of pcs. The interface is also in use in epos terminals, embedded graphics boards, in-car displays and the like. Transmitter and receiver chips are linked by tmds to allow a scaleable, high-speed digital interface over a twisted pair up to about 5m or to 15m with twin-ax shielded cable. There are three pairs of Tx/Rx chips for data rates of 650Mb/s to 1.12Gb/s per channel, all compatible. A developers' kit, the *CuPipe*, is available.

REP Design (UK) Ltd. Tel., 01462 670770; fax, 01462 671670; e-mail, RepDesignUK@compuserve.com.

Enq no 504

Bright lcd. Mercator offers the LTE052T five-inch colour tft display handling composite video for Pal and NTSC or separate RGB. Resolution is 320 by 234 pixels and there is a replaceable cold-cathode backlight. Brightness is 300cd/m², controlled externally, and a reverse scan allows a 6 o'clock or 12 o'clock viewing position, the same facility being provided in the horizontal scan. An on-screen display switch allows the overlay of an RGB image on a composite video one.

Mercator. Tel., 01493 334000; fax, 01493 334050.

Enq no 505

Hardware

Compact heatsinks. Thermalloy has announced a family of bonded-fin heatsinks for use when space is limited, the fins being either for forced ventilation or are made high and thin for natural convection. *Powerfin* sinks will fit into spaces where extruded types would be too large and may therefore be used in motor drives, ups and many other high-power applications. The bonded-fin technique enables the fin height to be over 100mm, base and fins being made in copper or aluminium to choice.

Redpoint Thermalloy Ltd. Tel., 01793 537861; fax, 01793 615396.

Enq no 507

Linear integrated circuits

Alternative graphic equaliser. Audio semiconductor specialist Profusion is supplying Sanyo's 7-band graphic equaliser. Only two capacitors and one potentiometer are needed on each band to provide an adjustable 12dB of boost or cut over the audio

range. Typical distortion at 1kHz and 1V output with no boost or cut is 0.02%. With the inputs shorted and controls set for flat response, typical output noise is 7µV, provided that a 10Hz to 30kHz band-pass filter is used. Devices may be used in series to increase the number of equaliser channels.

Profusion plc, Tel. 01702 543500, fax, 01702 543700, e-mail sales@profusion.co.uk.

Enq no 508

Materials

Anti-static matting. TBA can supply conductive neoprene and static dissipation matting. The materials resist chemical attack and all soldering temperatures, are flexible and durable, lay flat and comply with BS EN100015 Pt 1:1991, section 4.3.3. The matting comes in roll form or may be cut to any size in 1.3mm and 2.3mm thickness for floor or bench use.

TBA Industrial Products Ltd. Tel., 01706 47718; fax, 01706 46170; e-mail, info@tbaecp.co.uk; web, www.tbaecp.co.uk.

Enq no 509

Thermal insulators. High conductivity, low-cost thermal insulator pads by Warth use boron nitride to give lower thermal resistance than is usual, but higher voltage breakdown. They are for use with hot-running power transistors, giving a thermal resistance of 0.28°C/W and a voltage breakdown of 1kV. *Kool-Pads K200* are made in silicone rubber compound on a layer of woven glass fibre and are flexible, clean and do not crack or age. Temperature range is -60°C to 180°C and the pads are available with adhesive or non-adhesive coatings.

Warth International Ltd. Tel., 01342 315044; fax, 01342 312969; web, www.warth.co.uk.

Enq no 510

Microprocessors and controllers

Otp microcontrollers. Siemens has produced a new family of one-time-programmable microcontrollers with six members, all with on-chip memory. There is the *C505L*, which has an on-chip led controller and driver; the *C504* an 8-bit device with 16K of otp memory and a pwm unit; and the *C505CA* with 32K of otp memory and CAN 2.0B. The family consists of otp versions of the *C501G*, *C504*, *C505CA*, *C513* and *C515C*. Starter kits are available.

Siemens plc. Tel., 0990 550500; fax, 01344 396721.

Enq no 511



'Thinnest' fan. Sunon's GB0535ACB blower measures 35 by 35 by 4.8mm. It is claimed to be the world's thinnest, yet produces a static pressure of 0.28in of water. Its aluminium housing allows it to be attached directly to the hot component without a baffle and also acts as a heat sink itself. The fan uses only 0.5W and, at an airflow of 0.67cf/min and 0.91cf/min, there is little noise.

Thermaco Ltd. Tel., 01684 566163; fax, 01684 892356; e-mail,

thermaco@compuserve.com

Enq no 506

Mixed-signal ics

1200V gate drivers. International Rectifier has a range of half-bridge and three-phase gate driver ics to handle 600V and 1200V, while being compatible with existing 600V parts in pin layout and function. This means that the same circuit design and layout may be used in different systems by choosing the relevant ic, in 115V, 230V and 460V designs. The devices simplify the drive to fets and igbts, reducing component count by over 50% and are immune to voltage spikes. For use in ac drives to 20hp and power supplies to 10kW, the gate drive current is up to 2A.

International Rectifier. Tel., 01883 732020; fax, 01883 733410; web, www.irf.com.

Enq no 512

Microwave components

Ku-band mmics. Toshiba's TMD1414-X series is a range of microwave ics working in the ku band of 13.75GHz to 14.5GHz. These are wideband devices to eliminate the need for multiple units on a board to cover the band. TDM1414-02/1/2 provide output power of 23dBm, 31.5dBm and 34.5dBm with power gains of 37dB for the 02 and 26 for the others. Toshiba Electronics UK Ltd. Tel., 01276 694730; fax, 01276 694800.

Enq no 513

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Protection devices

Small ptc fuse. Schurter offers the *PFMC*, a smaller, resettable, positive temperature coefficient, polymer fuse, which trips fast enough for the protection of pc motherboards, for example. Ratings are 0.2A to 1.1A with interrupt ratings of 40A and 6-15V. Packaging is 4.5 by 3.2mm. Schurter AG. Tel., 0041 413693111; fax., 0041 413693333; e-mail, contact@schurter.ch; web, www.schurter.ch. Enq no 517

Switches and relays

Optocoupled mosfet. *PS7241-1A* by NEC is an optocoupled mosfet, forming a solid-state relay with a drive current down to 2mA. It is in a 4-pin SOP and 2.1mm high, 4.5mm

Emi suppression chokes.

Siemens has cut the cost of its emi suppression chokes by about 30% by the simple expedient of leaving the lids off. This procedure is made possible by a new approach to surface-mounting component design; in the new *B82793* range of ring-core double chokes, full encapsulation is unnecessary. These chokes are meant for use in emi suppression on data and signal lines and are a lower-cost and more compact alternative to rfi shielded cables. They come in two varieties: *B82793-C* has bifilar windings and suppresses asymmetrical noise, passing symmetrical data signal at frequencies of several megahertz; *B82793-S* is a sector-wound type and suppresses both types of interference, reducing hf noise on data lines. The range of inductances is 5µH to 4.7mH at currents in the range 0.4A to 1.2A, voltage ratings being 40V ac or 80V dc on all components. Siemens plc. Tel., 0990 550500; fax, 01344 396721. Enq no 514



square. Isolation is provided by a GaAs led at the input and a photovoltaic array driving a thyristor, a diode, and a pair of normally-open-contact mosfets to form the single-channel output. Breakdown voltage is 400V and leakage current 1µA. The device is relatively insensitive to rapid voltage changes, which also makes it insensitive to noise. IMO Precision Controls Ltd. Tel., 0181 452 6444; fax, 0181 450 2274; e-mail, imo@imopc.com; web, www.imopc.com. Enq no 518

Shallow switches. *EAO Series 71* panel-mounted switches only stick out from the front panel by 2mm and 42.5mm at the rear. All components are on a single pcb, so that a range of functions may be panel-mounted, including full-travel push-buttons, emergency stops and keylock types. The contact block may be removed for maintenance or replacement. Switches are sealed to IP65 and may be backlit by lamp or led. EAO Ltd. Tel., 01444 236000; fax, 01444 236641; e-mail, uksales@eao.com; web, www.eao-group.com. Enq no 51

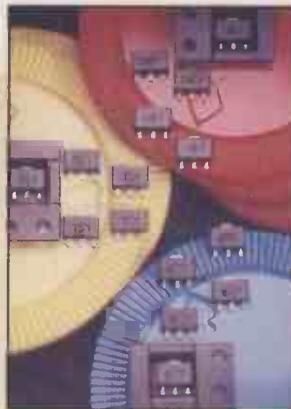
EQUIPMENT

Cameras

Shielded cameras. *Scout* closed-circuit television cameras by Pulse Power & Measurement are hardened against high-level electromagnetic fields. They are optically coupled and provided with remote control of focus, zoom, pan and tilt. A protective enclosure confers immunity to rf fields of more than 200V/m at frequencies over 1GHz with no effect on the picture and the cameras may be used in fields of more than 1kV/m at 18GHz. The cameras have two 75Ω video outputs. Pulse Power & Measurement Ltd. Tel., 01793 784389; fax, 01793 784391; e-mail, sales@ppm.co.uk. Enq no 522

Production equipment

Transformer testing. Voltech has introduced new software upgrades for its *AT1600* and *AT3600* transformer test platforms to provide fast, integrated transformer testing for frequencies in the 0-1MHz range at a rate of 20 tests per second. Basic accuracies are 0.05% and the tests are concerned in the main with the characteristics needed by transformers in modems, telephone systems and the like. Tests include common-mode rejection ratio, insertion loss, return loss and impedance matching. All the new tests are set up and made using the company's pc-based test editor, programs being saved to a server disk for later downloading. Voltech Instruments Ltd. Tel., 01235 861173; fax, 01235 861174; e-mail, sales@voltech.co.uk; web, www.voltech.com. Enq no 523



Power semiconductors

Power transistors. Zetex has introduced SOT89 versions of its low saturation voltage bipolar power devices, taken from the SuperSOT range; at present, there are four complementary pairs. *FCX617/8* are rated at 300% of the power of the smaller devices, handling 2W at 12A collector current. Saturation voltages for n-p-n and p-n-p types are 8mV and 10mV. Space saving is the outcome of a further variation in ranges: there are three SOT89 transistor pairs formerly only made in SOT223 and TO92 E-line form, reducing space taken up by about 50%. These are *FCX688/789*, *FCX1047/1147*, and *FCX1051/1151* pairs. Zetex plc. Tel., 0161 622 4422; fax, 0161 622 4420; web, www.zetex.com. Enq no 516

Passive components

Class X2 capacitors. Arcotronics' *R.46* range of metallised polypropylene film capacitors is suitable for interference suppression and in particular where failure could lead to injury. The range comes in values between 0.01µF and 2.2µF in 10% and 20% tolerances at 1kHz. Voltage rating is 275V ac. Windings are non-inductive, leads are tinned copper and the plastic casings are filled with polyurethane resin, the box material being resistant to solvents and flame retardant to UL94-V0. Easby Electronics Ltd. Tel., 01748 850555; fax, 01748 850556; web, sales@easby.co.uk. Enq no 515



Power supplies

Plug-top battery charger. Arlec Power announces a new version of its *BCN* charger, which offers a choice of British or European connection. The new range includes types using a negative delta-peak charge control where the unit detects a slight drop in voltage at full charge, switching at that point to a trickle charge to avoid damage to the battery. There are also models fitted with timers that charge for a set time. All models are protected against shorts, overload, and incorrect connection. In addition they all satisfy European and other safety and emc standards. Arlec Power UK Ltd. Tel., 01582 544520; fax, 01582 544521. Enq no 524

High-current dc. Kikusui *PAN-A* high-current, regulated supplies come in four powers: 175W, 350W, 700W and 1kW at voltages to 250V and currents up to 50A. A preregulator uses fets and the series regulator is a power transistor type, temperature drift is reduced by forced-air cooling and fast transient response is achieved by a wide-band error amplifier. There is provision for external control by analogue voltage and current signals or by GPIB with an optional unit. All units in the series may be connected in series or parallel with master/slave control. Telonic Instruments Ltd. Tel., 01734 786911; fax, 01734 792338. Enq no 525

1U, 1500W. Vicor's new range of ac/dc power supplies in the *Westcor PFC Mini* family has increased available power from 600W to 1500W without derating for 24V, 28V and 48V outputs. Units have 1 to 6 floating, independently regulated outputs providing 1-95V at up to 600W per output. Input is power-factor-corrected to handle 85-254V ac at 47-500Hz or 100-380V dc, and complies with the usual emc requirements. Case size is 304.8 by 152.4mm in a 1U height and a control interface provides for output sequencing and shutdown. Vicor UK. Tel., 01276 678222; fax, 01276 681269; e-mail, vicor@vicr.com; web, www.vicr.com. Enq no 526

Test and measurement

Digital signal generator. Kenwood's *DG 2432* digital signal generator produces all the test signals needed in the measurement of performance in digital audio equipment, d-to-a converters, filters and other dsp instruments, all in the EIAJ CP-1201 digital audio interface form. The instrument holds in memory 100 preset test patterns, which may be modified to suit a particular application remotely by way of an RS232 interface. Output is provided via optical connectors at either -19dBm or -27dBm, or through coaxial connectors. Operation is by key entry on the front panel, annunciators providing the relevant information on set up and functions,

Please quote "Electronics World" when seeking further information

and a digital display gives a readout of pattern information.

Kenwood UK Ltd. Tel., 01923 655291; fax, 01923 655297.
Enq no 527

Digital wattmeters. *WT1000* wattmeters from Yokogawa are accurate to within 0.1% at a bandwidth of 300kHz, being made in single-phase and three-phase versions and in a special version for the evaluation of motors. Input range is 15-1000V rms and a filter with a selectable cut-off frequency facilitates the examination of inverter waveforms; built-in pll circuitry allows harmonic analysis to be done on fundamentals from 10Hz to 440Hz up to the 50th. The three-phase model measures phase difference between phases, active, reactive and apparent power of the fundamental. The *WT1030M* works with a torque meter and computes torque, speed, mechanical power, synchronous speed, slip, motor efficiency, and total efficiency of the motor under test.
Yokogawa Martron Ltd. Tel., 01494 459200; fax, 01494 535002; e-mail, info@martron.co.uk; web, www.martron.co.uk.
Enq no 528

Temperature recorder. *DS1615* from Dallas is a single-chip diil instrument to record temperature integrity and variation over time, incorporating a Year 2000-ready clock, a digital thermometer, non-volatile memory, control logic, and a serial interface. Data is stored in a conventional data log, and in the form of a histogram for later analysis. The instrument may be used on its own as a complete data logger or as an embedded unit. As an embedded unit, it uses no system resources to monitor and record temperature 2048 times at preset intervals, continuing to operate during power-down.
Dallas Semiconductor Corporation. Tel., 0121 782 2959; fax, 0121 782 2156.
Enq no 529

Burst generator. Schaffner announces a full-function burst generator for emc testing that is housed in a 342mm by 134mm bench-top case, the smallest available. *NSG 3025* functions exceed the requirements of the relevant standards and is CE marked. Most IEC standards are programmed for fast recall and all pulse parameters are variable. Pulse amplitude is 200-4800V, burst frequencies 0.1kHz-1MHz with 1-255 spikes per burst. A coupling network is provided to allow the pulses to be superimposed on the supply lines of the equipment under test and the pulses are also available at the front panel for data and line testing. Control is by RS232 link using the company's WIN 3105 software for Windows.
Schaffner EMC Ltd. Tel., 0118 9770070; fax, 0118 9792969.

Instrument stack. *QuanteC* by ENG-INN is a single tower case combining

the functions of signal generator, pulse generator, dual-channel, true-rms 2MHz voltmeter, dual-channel digital storage oscilloscope, gain/phase analyser, dual-channel frequency meter, white noise generator, power meter, arbitrary function generator, phase meter, Icr meter and spectrum and harmonic analysers. All functions are available as hardware and/or software upgrades and each has a single set of connections and an RS232 port for control and configuration by software. The display is an electroluminescent type and there is a printer.
ENG-INN (Electronics) Ltd. Tel., 0116 2376467; fax, 0116 2376167; e-mail, sales@eng-inn.co.uk; web, www.eng-inn.co.uk.
Enq no 530

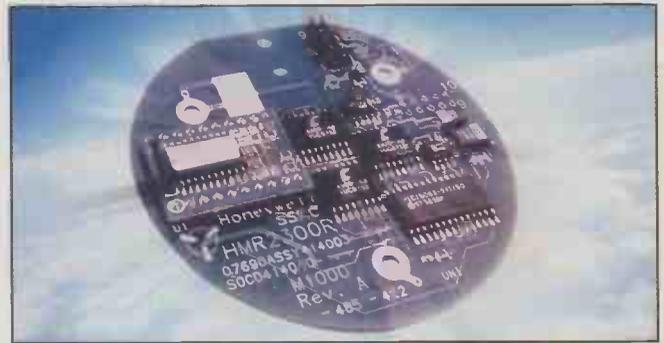
Frequency-stable function generator. TTI offers the *TG550* 5MHz function generator. It has digital frequency locking to a crystal reference; after setting the required frequency, one presses the lock switch to activate a measuring and correction circuit to obtain an accuracy of one digit on the 4-digit display. A second display gives output level in pk-pk, rms or dc offset. For low frequencies, an auto-ranging reciprocal measurement is made to avoid long gate times. Output is sine, square, triangle, pulse and ramp in the frequency range 0.005Hz to 5MHz; a sweep generator provides linear or log. frequency sweeps at periods of 20ms-20s and there is also amplitude modulation from zero to 100%. The *TG550* will also function as a seven-digit frequency counter to 20MHz.
Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409; e-mail, sales@ttinst.co.uk.
Enq no 531

COMPUTER AND DATA HANDLING

Computers

Hardy computer. Panasonic invites its customers for the *ToughBook 27* notebook computer to drop it onto concrete, attempt to drown it and generally treat it in a less than careful manner, since it has satisfied the IEC tests along these lines. The computer incorporates a 266MHz Pentium with MMX, a shock-protected, removable hard disk drive, 32MB of EDO ram, a 512 cache, a 12.1-in tft display, CD-rom, and two card slots. Its display panel is fitted with a reflection filter, to make the picture viewable in daylight.
Panasonic UK Ltd. Tel., 0500 404041; web, www.panasonic.co.uk.
Enq no 532

Compact sbc. The *PCM-3345* In Advantech's PC/104 series is a self-contained processor board that is only 90 by 96mm and yet offers all the capability of a much larger design. It



uses a 66MHz 486DX cpu that needs no frequency or voltage configuration, the board also being provided with 16Mb of memory (32Mb as an option). It supports ATX power to allow remote power-on and watchdog timing; there are one parallel and two serial ports, an FDD connector and one IDE connector. Flash card options are available up to 220Mb for use as a simulated hard disk. Support for crts is provided, and for lcds as an upgrade, now available.
Semicom UK Ltd. Tel., 01279 422224; fax, 01279 433339; e-mail, sales@semicom.demon.co.uk
Enq no 533

COMPUTER BOARD-LEVEL PRODUCTS

Quad dsp board. Blue Wave Systems produces the *VME/C6420* digital signal processor board, which is intended for applications such as radar and signal intelligence in which large amounts of data must be processed quickly. It is available with a choice of four *C6201* fixed-point dsps or four *C6701* floating-point processors and a fast multi-port crossbar controls data flow, enabling up to four 200Mbyte/s simultaneous, on-board, point-to-point links to be set up dynamically. All on-board housekeeping is the function of a *PowerQUICC* processor, which accesses 2Mbyte of flash, a total of 16Mbyte of sdram being shared between the four dsps., control processor and i/o.
Blue Wave Systems Ltd. Tel., 01509 634300; fax, 01509 634333; web, www.bluews.com.
Enq no 534

Transducers and sensors

Three-axis magnetometer. Honeywell has a new three-axis, strap-down magnetometer, the *HMR2300R*, intended to sense the strength and direction of the earth's magnetic field and send the x, y and z components to a navigation system by way of a serial port. The sensor is around 72mm in diameter and fits an ML-1 type of enclosure. The sensor is microprocessor-based with a range of ± 2 gauss to a resolution of ± 70 µgauss, heading accuracy being 0.02°. The device is meant to replace flux valve sensors and may be used in conjunction with systems such as GPS and remote vehicle monitoring.
Inertial Aerosystems Ltd. Tel., 01252 782442; fax, 01252 783749; e-mail, sales@inertialaero.demon.co.uk.
Enq no 520

Tilt sensor. The *Model AB* absolute tilt sensor from Control Transducers connects directly to any analogue data system. It is based on an optical absolute encoder with a pendulum referred to gravity and works from a single six-wire telephone connector and cable. Power supply is 9-36V dc and its output is in the form of 1-4096mV in 1mV steps for a 360° rotation while turning continuously. The unit comes complete with cables, plugs, software, power supply, control cards and can be supplied with matched, direct-reading indicators.
Control Transducers. Tel., 01234 217704; fax, 01234 217083.
Enq no 521



Please quote "Electronics World" when seeking further information

Dsp coprocessor. Datel announces the *PC-430M*, a high-speed coprocessor board for use in recovering signal from noise, in distortion analysis and in the examination of vibration and resonance filtering, mapping and imaging, etc. It is an ISA analogue-to-digital dsp coprocessor data acquisition board with four simultaneous sampling a-to-d converters and is suitable for continuous FFT processing, saving comms signal to disk or the simultaneous graphics display of spectra. With all four a-to-ds sampling in parallel, conversion rate is 200kHz/channel. An on-board crystal oscillator and timer/counter tracks phase precisely and up to 4Mb of memory stores data and programs.

Datel (UK) Ltd. Tel., 01256 880444; fax, 01256 880706; e-mail, datel.ltd@ge.geis.com; web, www.datel.com.
Enq no 535

Data acquisition

PCI data acquisition. From Datel, the *PCI-416M*, a four-channel, analogue input board having 32-bit architecture for PCI computers. There are four 16-bit inputs with $\pm 10V$ range single-ended, sampling at up to 200kHz per channel, the front ends having low-noise, wide-bandwidth a-to-d converters. The design allows the movement of two a-to-d words in each 32-bit PCI transfer, the fifo memory used for this purpose acting as a filter to dissociate the a-to-d timing from the block bursts of the PCI bus. The device is a temporary bus master to burst a-to-d blocks to memory, freeing the cpu for other tasks. Gapless sampling is provided with no data loss; a pretrigger collects data continuously to host circular ring memory of several megabytes. When an external trigger arrives, the *PCI-416M* counts down the number of preloaded samples after the trigger and stops when all are collected. No programming is needed.

Datel (UK) Ltd. Tel., 01256 880444; fax, 01256 880706; e-mail, datel.ltd@ge.geis.com; web, www.datel.com.
Enq no 536

Data communications

Single-chip modem. TDK announces the *73M2921* single-chip modem ic, which has all necessary to implement a V.23/V.22bis/V.29 modem, plus handshaking for all data modes. It operates at data and fax speeds up to 9600bit/s in V.29 and has microcontroller, data and analogue interfaces and an oscillator. The device consists of a dsp core with integrated ram and rom and codec; it is optimised for low power use and operates from 3.3V and 5V lines.

TDK UK Semiconductor Corp. Tel., 0181 4437061; fax, 0181 4437022; e-mail, europe.sales@tsc.tdk.com; web, www.tdksemi.demon.co.uk.
Enq no 537

Software

Wiring harness design. In partnership with Volvo, Analogy Inc. has produced a set of software for the design of wiring harnesses, including electrical analysis and the detection of possible hot spots in the harness. The software consists of the *SaberHarness* wiring editor, a dc simulator, an optional mixed-signal transient simulator, the *SaberBundle* editor to produce drawings for manufacture and the interfaces to mechanical 3D cad software. The physical wire model allows geometric and material wire characteristics to be simulated. Since *SaberHarness* is integrated with the 3D cad software, data may be passed to the mechanical cad program, manipulated and passed back to *SaberHarness* with changes affecting the electrical design shown. Many more features are provided to simulate the design and to predict possible electrical and mechanical failure.

Analogy Europe. Tel., 0049 811 60093-23; fax, 0049 811 60093-11; e-mail, acirkel@analogy.com.
Enq no 538

Algebra. *Maple V Release 5.1* is the latest edition of *Waterloo*, Maple's computer algebra system. It allows electronic publishing of mathematical information using IBM's Hypermedia Browser plug-in technique for Web browsing in LaTeX - a standard typesetting language for maths, using the Maple language. Output is of publication quality while producing smaller files than has been possible. Maple's newsletter describing the system is available free from Adept.

Adept Scientific Micro Systems Ltd. Tel., 01462 480055; fax, 01462 480213; e-mail, info@adeptscience.co.uk; web, www.adeptscience.co.uk.
Enq no 539

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Web shopping. Adept Scientific's *WebStore* at <http://directory.adeptscience.co.uk> is linked to the company's main site. It provides prices and will print quotations and customers may order hardware and software by credit card. The main site gives full information about the company's products, including those from *Mathcad*, *ComputerBoards*, *Electronics Workbench*, *NWA Quality Analyst* quality control software and the *Maple V* computer algebra software.

Adept Scientific Micro Systems Ltd. Tel., 01462 480055; fax, 01462 480213; e-mail, info@adeptscience.co.uk; web, www.adeptscience.co.uk.
Enq no 540

Murata. Information on Murata's whole range of products is now contained on CD rom, the 1999 edition now being available, complete with a new search engine. You can search by product type, emi problem or part number and, having found the component, there is a worldwide contact list with addresses, telephone and fax numbers. *Murata Electronics (UK) Ltd.* Tel., 01252 811666; fax, 01252 811777.

Enq no 541

Guide to ics. Farnell's *IC Master*, the integrated circuit reference, is

in its 25th year and is a three-volume set of books or on CD rom. In all, there are data sheets on 106 000 ics, 8000 of them being new listings from 350 makers, the CD also giving data on obsolete devices to take its total to 152 000 devices. The guide is easy to navigate, allowing selection by function, type number and keyword or package style. Links to makers' home pages are provided for on-line users. *Farnell Components Ltd.* Tel., 0113 263 6311; fax, 0113 263 3411, web, www.farnell.com. ■
Enq no 542

BACK ISSUES

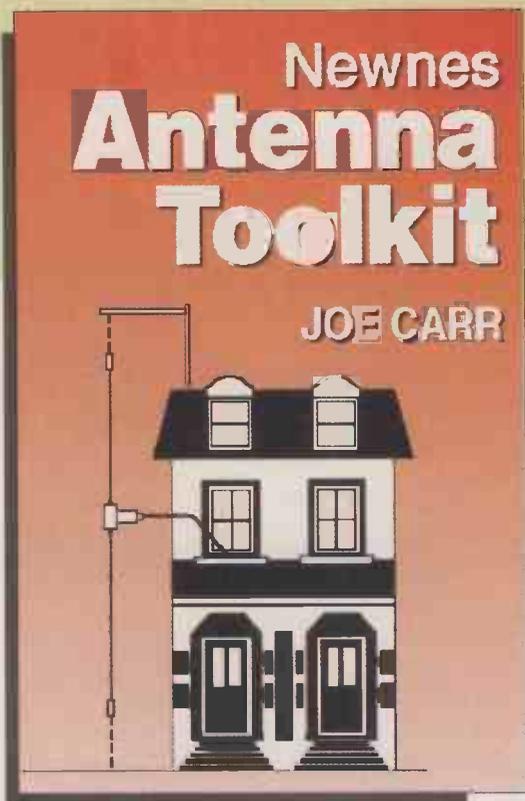
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Antenna Toolkit

by Joe Carr

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

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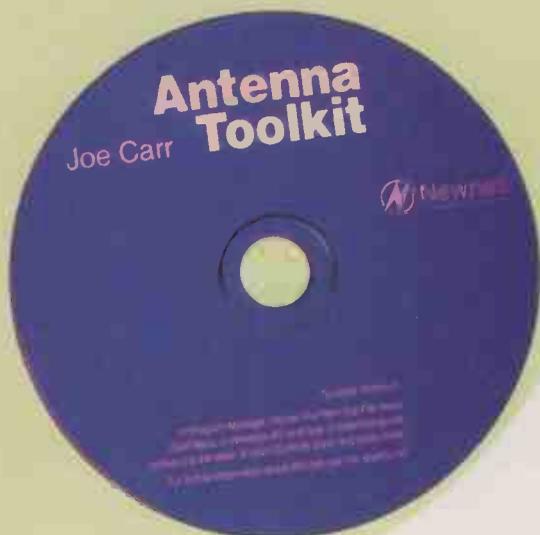
Also included is a Windows freeware package, from the Voice of America organization, called VOACAP. This is an hf propagation predictor which some commercial sources have offered unmodified for hundreds of dollars.

UK Price: £27.50 Europe £30.00 ROW £32.50

What's in the book?

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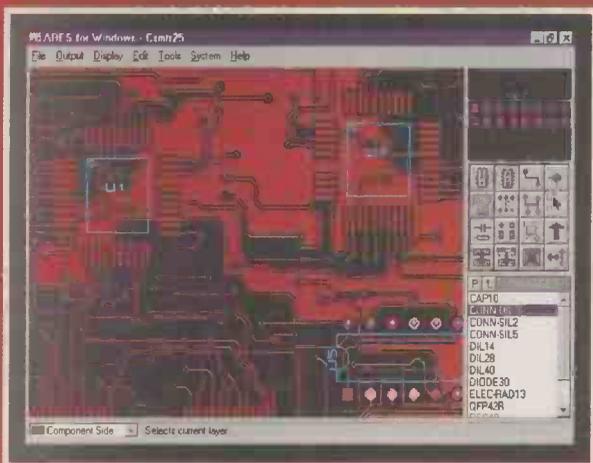
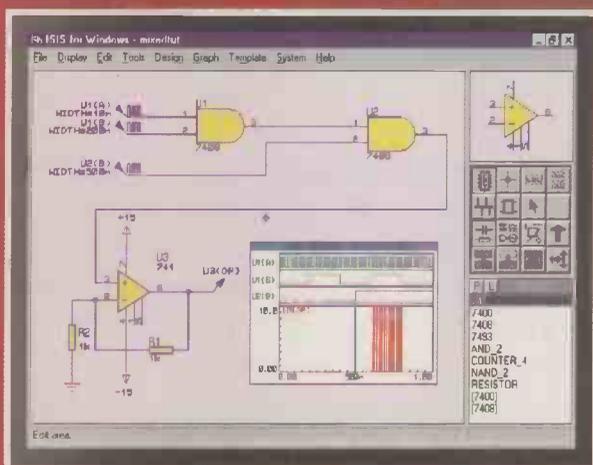
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Here are details of the special offers contained in this month's cover-mounted voucher booklet.*

*UK issues only. Details for overseas readers vary so please contact companies concerned directly, quoting Electronics World special offers.

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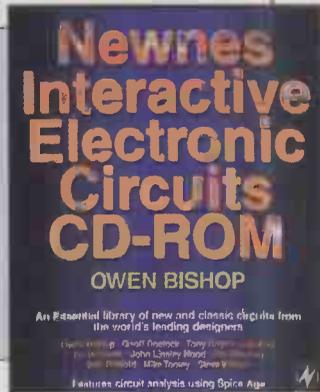
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Newnes, Linacre House, Jordan Hill, Oxford OX2 8DP, <http://www.newnespress.com>

A 30W hi-fi audio power amplifier for just £9.99

ILP's HY2000 power amplifier is a fully-encapsulated, high-quality power amplifier with integral heatsink.

By selecting the appropriate mains transformer and PCB programming link, the amplifier can be used with 4Ω or 8Ω loads. Input sensitivity is adjusted automatically.

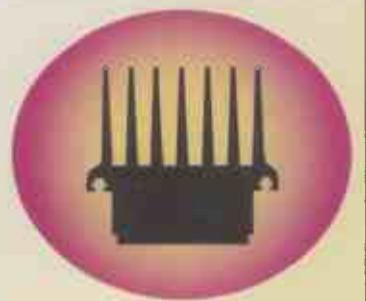
T-slots in the heatsink and nuts and bolts provided facilitate mounting.

Typical specifications

Parameter	Value
Maximum output power	30W rms
-3dB frequency response	15Hz-50kHz
Total harmonic distortion @ 1kHz	0.005%
Signal-to-noise ratio (DIN Audio)	100dB
Slew-rate, typical	10V/μs
Rise time	5μs
Input sensitivity	500mV rms.
Input impedance	100kΩ
Damping factor, 8Ω @ 100Hz	>400
Load impedance (programmable)	4 or 8Ω
Maximum DC rails, 8Ω load	+30V
Size, width by height by extrusion cut	76x75x40mm
Weight	240g

Features

- 4 or 8Ω operation
- Enhanced specifications
- Anti-thump circuitry
- Integral heatsink
- Thermal protection
- Encapsulated
- Fully assembled
- Easy to mount



Power supply requirements

ILP has developed a range of hi-fi quality low noise audio power toroidal transformers for the HY2000 they are low profile with resin filled centres for maximum noise absorption and ease of mounting. Finished in black to match the amplifiers these transformers are the ideal choice to power the HY2000.

Mode	Transformer	RMS supply
4ohm	AT0304	16-0-16V
8ohm	AT0308	20-0-20V



ILP, Spong Lane, Elmsted, Ashford, Kent TN25 5JU, phone +44 (0)1233 750481, fax +44 (0)1233 750578.

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TiePie introduces the HANDYSCOPE 2 A powerful 12 bit virtual measuring instrument for the PC

The HANDYSCOPE 2, connected to the parallel printer port of the PC and controlled by very user friendly software under Windows or DOS, gives everybody the possibility to measure within a few minutes. The philosophy of the HANDYSCOPE 2 is:

"PLUG IN AND MEASURE".

Because of the good hardware specs (two channels, 12 bit, 200 kHz sampling on both channels simultaneously, 32 KWord memory, 0,1 to 80 volt full scale, 0,2% absolute accuracy, software controlled AC/DC switch) and the very complete software (oscilloscope, voltmeter, transient recorder and spectrum analyzer) the HANDYSCOPE 2 is the best PC controlled measuring instrument in its category.

The four integrated virtual instruments give lots of possibilities for performing good measurements and making clear documentation. The software for the HANDYSCOPE 2 is suitable for Windows 3.1 and Windows 95. There is also software available for DOS 3.1 and higher.

A key point of the Windows software is the quick and easy control of the instruments. This is done by using:

- the speed button bar. Gives direct access to most settings.
- the mouse. Place the cursor on an object and press the right mouse button for the corresponding settings menu.

- menus. All settings can be changed using the menus.

Some quick examples:

The voltage axis can be set using a drag and drop principle. Both the gain and the position can be changed in an easy way. The time axis is controlled using a scalable scroll bar. With this scroll bar the measured signal (10 to 32K samples) can be zoomed live in and out.

The pre and post trigger moment is displayed graphically and can be adjusted by means of the mouse. For triggering a graphical WYSIWYG trigger symbol is available. This symbol indicates the trigger mode, slope and level. These can be adjusted with the mouse.

The oscilloscope has an AUTO DISK function with which unexpected disturbances can be captured. When the instrument is set up for the disturbance, the AUTO DISK function can be started. Each time the disturbance occurs, it is measured and the measured data is stored on disk. When pre samples are selected, both samples before and after the moment of disturbance are stored.

The spectrum analyzer is capable to calculate an 8K spectrum and disposes of 6 window functions. Because of this higher harmonics can be measured well (e.g. for power line analysis and audio analysis).

The voltmeter has 6 fully configurable displays. 11 different values can be measured and these values can be displayed in 16 different ways. This results in an easy way of reading the requested values. Besides this, for each display a bar graph is available.

When slowly changing events (like temperature or pressure) have to be measured, the transient recorder is the solution. The time between two samples can be set from 0.01 sec to 500 sec, so it is easy to measure events that last up to almost 200 days.

The extensive possibilities of the cursors in the oscilloscope, the transient recorder and the spectrum analyzer can be used to analyze the measured signal. Besides the standard measurements, also True RMS, Peak-Peak, Mean, Max and Min values of the measured signal are available.

To document the measured signal three features is provided for. For common documentation three lines of text are available. These lines are printed on every print out. They can be used e.g. for the company name and address. For measurement specific documentation 240 characters text can be added to the measurement. Also "text balloons" are available, which can be placed within the measurement. These balloons can be configured to your own demands.

For printing both black and white printers and color printers are supported. Exporting data can be done in ASCII (SCV) so the data can be read in a

spreadsheet program. All instrument settings are stored in a SET file. By reading a SET file, the instrument is configured completely and measuring can start at once. Each data file is accompanied by a settings file. The data file contains the measured values (ASCII or binary) and the settings file contains the settings of the instrument. The settings file is in ASCII and can be read easily by other programs.

Other TiePie measuring instruments are: HS508 (50MHz-8bit), TP112 (1MHz-12bit), TP208 (20MHz-8bit) and TP508 (50MHz-8bit).

Convince yourself and download the demo software from our web page: <http://www.tiepie.nl>. When you have questions and / or remarks, contact us via e-mail: support@tiepie.nl

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delivers raw MSF data from Rugby but there's an add-on controller for producing serial data.

Detailed on page 748 of the September 1997 issue of *Electronics World*, the decoder and antenna module are remarkably easy to assemble and feature an enhanced level matching circuit. The module has two pins for the antenna, ground and V+ connections and an output.

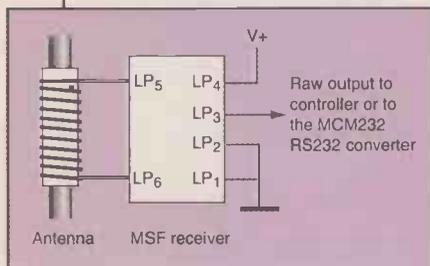
The time-code signal from the decoder is exactly as transmitted – i.e. raw data, whose format is described in the article just mentioned. If you want RS232 for feeding

into a pc or controller system, simply feed the raw data into the MCM232 microcontroller.

Connection details are supplied with the parts.

For price information, see the coupon in the booklet on the cover of this issue, or contact Galleon – details below.

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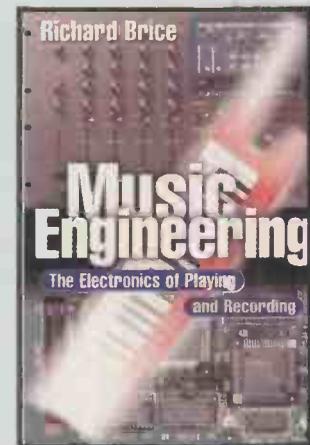
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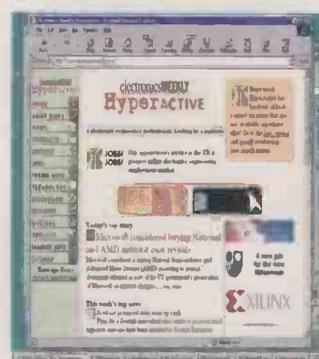
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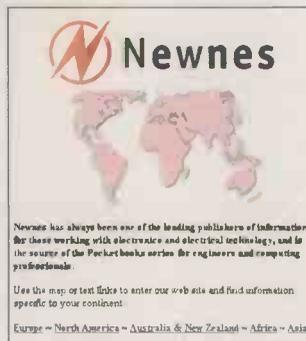
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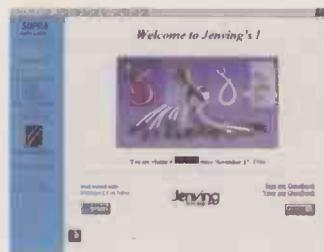
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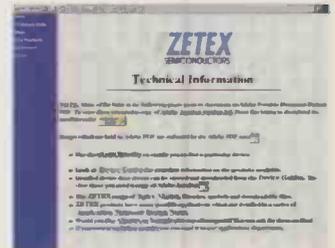
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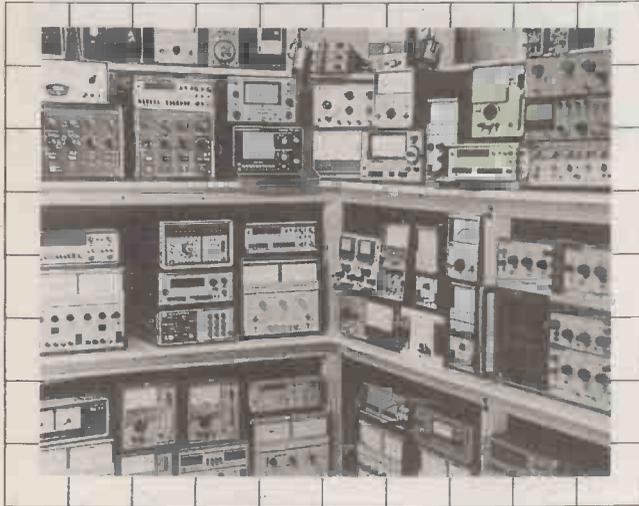
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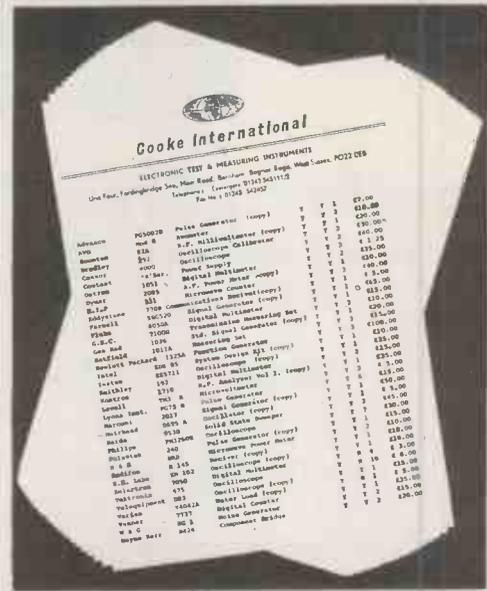
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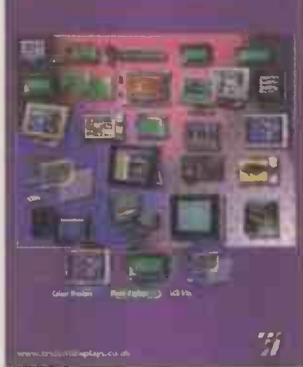
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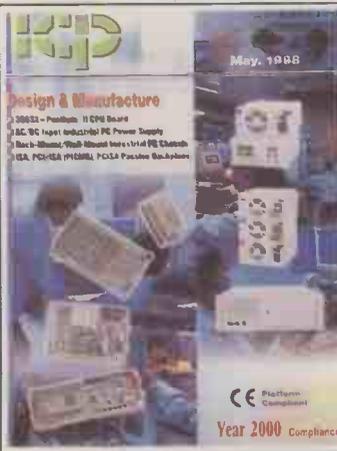
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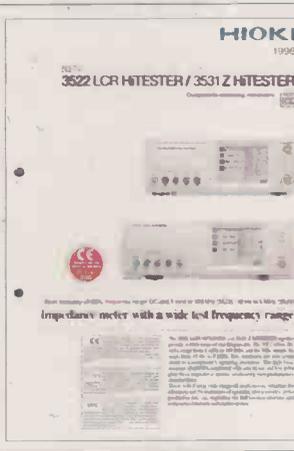


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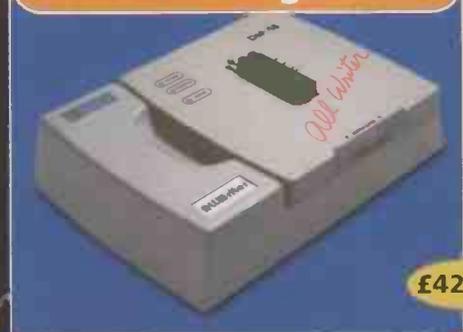


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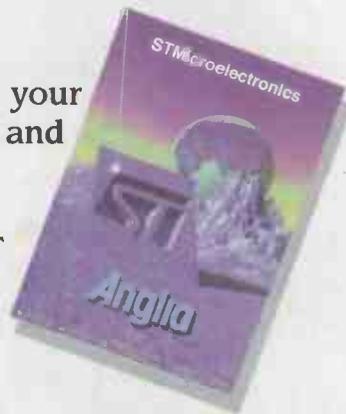
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