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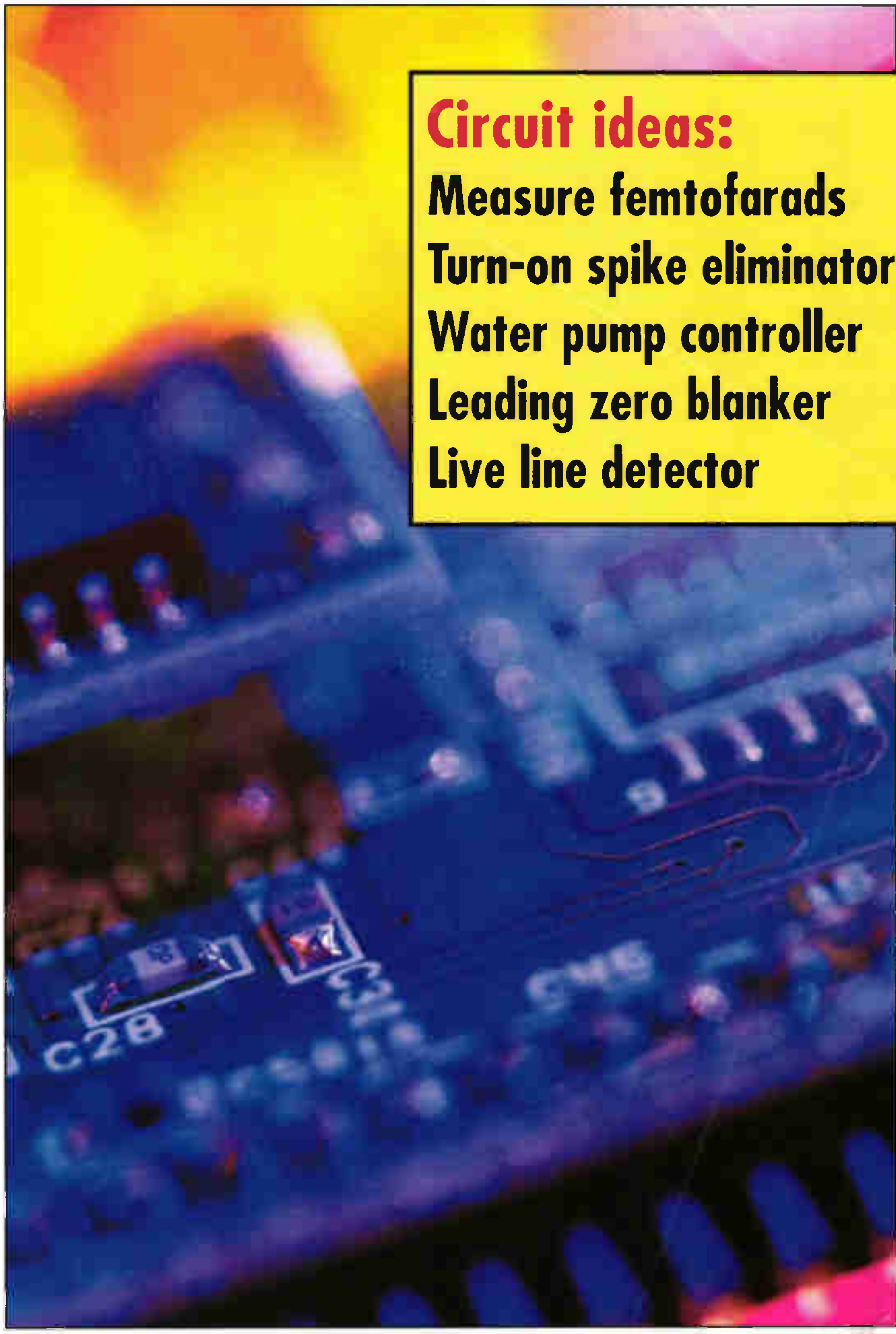

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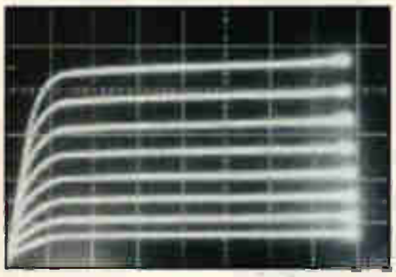

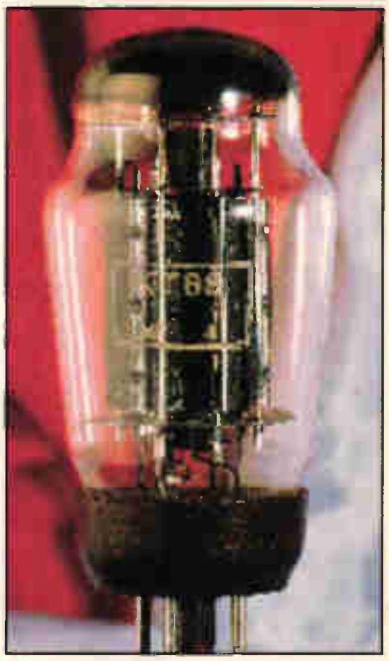

A valve's grid determines a surprising number of its parameters. Patrick Mitchell looks at the grid's evolution, starting on page 622.

Making a wireless network operating between 20 and 40GHz poses problems as each transmitter needs a line-of-sight link with the receivers. Melanie Reynolds looks at a new solution – page 600.

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REED BUSINESS INFORMATION

Piracy and power

A feeling of relief all round should have greeted Judge Jackson's long-awaited decision on Microsoft, that at last this long-winded business was coming to an end. But no – the Microsoft management have vowed to fight on, not conceding anything. So this saga will continue, to the disgust of the many Windows users who must now realise they have been taken for ride.

One of the depressing facets of this case has been the refusal of the Microsoft bosses even to acknowledge that they have done anything wrong, despite all the evidence to the contrary. They are clearly out of touch with reality. But if you are that big, and powerful, with vast quantities of money to hire the best lawyers, it is easy to fall into the trap of thinking that might is right, and believe that by spending enough money, it is possible stand the law on its head and prove that wrong is right.

And mighty they are. Which is why, even with all the bad publicity that they have recently accrued, Microsoft's management have picked this moment to launch a crusade against software piracy. You may have seen their campaign. They are comparing software piracy to knowingly owning a fake copy of the Mona Lisa – an approach that would be laughable if it were not true.

This is a truly bizarre situation. A company shown to have acted illegally, and that is about to be punished for it, has chosen this moment to lecture the rest of us about infringements of copyright. And the time they have picked to do this is when the free alternative, Linux, has taken off in a big way. What more evidence is needed that overwhelming power has affected their ability to size a situation up correctly?

However, Microsoft has shot itself in the foot, and something good could come out of this. For most users who just want to use their computer for basic tasks like browsing the net, using a word processor, a database and spreadsheet, and who don't want to be tainted by piracy, the complete answer is to drop Windows altogether and use Linux and StarOffice.

Both of these software packages are free and readily available. Why pay £200 for Office 2000 when the excellent Staroffice is free?

StarOffice can be downloaded from the Web, but it is rather large. A better method is to ring Sun, and you will receive a free CD with the program StarOffice not just for Linux, but for Windows and OS/2 as well, ensuring a degree of compatibility unobtainable with Microsoft's offerings.

And having entered the world of Linux you will discover that a vast amount of alternative software is available, some of it free, some of it modestly priced.

Since the main driving force behind piracy has been the high cost of Windows products, this mischievous activity should become a thing of the past under Linux.

Unfortunately, there are several factors working against Linux. Certain professionals in the computer world, who see their intellectual or financial investment in Windows being eroded as Linux rises, have been sniping away with a disinformation campaign. Most of what they say is plain old propaganda, with a few half-truths thrown in for good measure. Fortunately, it is possible to spot these people with a vested interest a mile off if you have been made aware of this disinformation.

The very fact that Linux is free poses a problem to some people, due to the quirk of human nature which equates being free with being without much worth. They say to themselves, "how can a free operating system put together by amateurs working part-time possibly challenge a multi-billion dollar OS like Windows?"

The reality is that some of the developers of Linux – far from being amateurs – are some of the most skilled exponents of their art on the planet. Thankfully Microsoft has no monopoly here. They provide the benefit of their intellectual effort on behalf of Linux for anyone who cares to make use of it, and encourage others to do the same. Eventually everyone gains, and not just a few ultra-fat-cats, as in Microsoft's case.

Readers of *EW* should have no trouble with this concept, because Contributors to the magazine, in the various features, and particularly in 'Circuit ideas', provide novel ideas and leading-edge designs for anyone to use, despite the fact that they may have some commercial value. Contributors can likewise, if they wish, make reciprocal use of others' ideas through the same channels. The buzz word sometimes used in the Linux community for this practice of openly offering intellectual property is 'copyleft' – the opposite of copyright.

If not all of your software is pirated, and you have legitimate Windows applications which you don't want to see abandoned by converting to Linux as your sole operating system, there are great possibilities in the 'Wine' program for Linux. This enables Windows applications to run on Linux without emulation, almost like a native application.

At present, Wine is usable but still under development, and shows considerable promise. Naturally, Wine is free and both source code and binary files are provided.

If you use illicit software, you may think you are safe from detection by virtue of being a small fry in the world of pirates. But judging from the tone of Microsoft's latest ploy you may be less safe than you think – particularly if you use the Internet and have unauthorised copies of software on your hard disk drive. Much better by far to side-step the issue altogether and adopt Linux. ■

Simon Wright

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UPDATE

UK teams feature in US technology display showcase

Display technology specialists from the UK made a significant contribution at the world's largest display technologies show, the Society for Information Display (SID) conference in California recently.

DERA in Malvern presented a paper on its zenithal bistable device – or ZBD, pronounced zebedee. This is an LCD that keeps its image after power is removed.

Even with no power applied, the image is robust, making it suitable for smart cards and shelf-edges in supermarkets. "If you press it, the image remains. ZBD is the only example of a bistable LCD which does this. All the rest can have the 'bistable' image erased by pressing the display," said DERA's Dr Emma Wood.

Modelled on a mobile phone display, DERA's demonstrator is 2 by 2cm with 90 by 83 pixels.

Latching voltages can be as low as 10V and contrast ratio is 40:1 per pixel or 15:1 including inter-pixel gaps. Displays for e-books and electronic paper using ZBD are proposed.

Another reflective technology is being shown by IBM. "It is as close as we can get to print on paper," said independent consultant Tony Lowe who worked on the display at IBM, "It reflects 60 per cent of light and has a 12:1 contrast ratio."

By suspending a 1μm polyester sheet through the middle of its display, IBM has shared a single-pixel electric field between two stacked cells.

Each cell contains a guest-host system. These are orientated at 90° from each other. In this way the cells are either transparent, or cross polarised – and therefore opaque.

The company showed a 320dpi 100mm diagonal TFT demonstrator.

Phosphor experts from the University of Greenwich presented their work on ultra-fine phosphor particles for the next generation of field emission displays.

Pixels are shrinking as display resolutions climb and pixels as narrow as 0.3μm can be found. At 2-5μm, common phosphor particles are too large to work well.

"We can make regularly shaped particles as small as 0.1μm," said Professor Jack Silver from the Greenwich team.

These experimental particles still work well as phosphors, having 60 per cent of the light output of more typical phosphor particles.

The University of Cambridge has been working on displays that give true 3-D images without the use of special eye-wear for many years now – so called autostereoscopic displays. Using a spinning polygon inside a mirrored cavity it can now produce a 3D image with a viewing angle of 90°.

Boinnnng... DERA in Malvern is presenting a paper on its zenithal bistable device – ZBD- pronounced zebedee – an LCD which keeps its image after power is removed.

Robot with the brain of a fish

A real android, living brain and mechanical body, is taking shape in a joint US-Italian project.

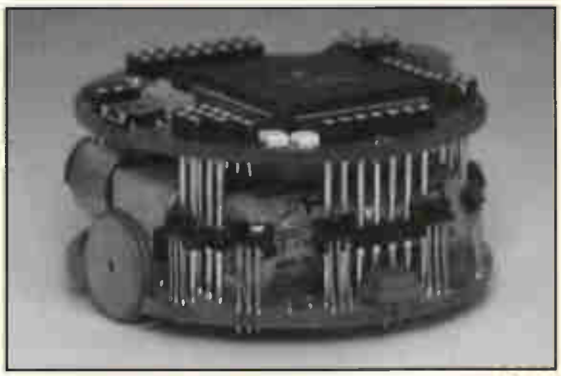
Not quite Robocop, its body is a commercial Swiss-made Khepera robot and its brain is a handful of cells taken from the brain stem of a sea lamprey – an eel-like fish.

According to project scientist Simon Alford of the University of Illinois in Chicago, using even a few cells for control gives the robot a higher level of sophistication than you might expect. "A neuron is no logic gate," he said, "There is a lot more to a neuron than that. They have an enormous capacity for integrating information – even on their own."

According to Alford, the robot changes behaviour as the cells, normally responsible for swimming, learn. "They have to adapt [in life], they have a self-teaching mechanism," he said.

The robot has light sensors on board. Signals from these are fed to the remotely-mounted cells through an Italian-programmed signal-conditioning computer. The cells produce signals that then control the robot motors through the computer.

The work is to be presented at Artificial Life 7 in Oregon in August.



Murphy Jnr... having cells from a fish brain, this robot's behaviour changes as the cells learn.

Turbo-coding for set-top box chips

STMicroelectronics is to use Turbo-coding, the error correction scheme for third generation mobile phones, in its set-top box chips. This allows 50% more channels in the same bandwidth, the firm claims.

"ST has entered into an agreement with France Telecom to exploit the technology," said Philippe Geyres, vice president of ST's consumer and microcontroller group. "We can increase the number of channels in a given satellite by 50%."

Alternatively, set-top boxes could receive Web pages over satellite, or the antenna could be reduced in size by 30%, Geyres said.

France Telecom received patents relating to Turbo-coding in 1991. It developed its first chip in 1995 and worked with ST to produce more advanced devices in 1998.

ST now plans to deploy Turbo-coding in high end set-top boxes starting early next year. It has already demonstrated the technology to broadcasters and set-top box makers.

Using Turbo-codes increases the spectral efficiency of the channel by around 2dB, said ST, which results in 50% more data being sent.

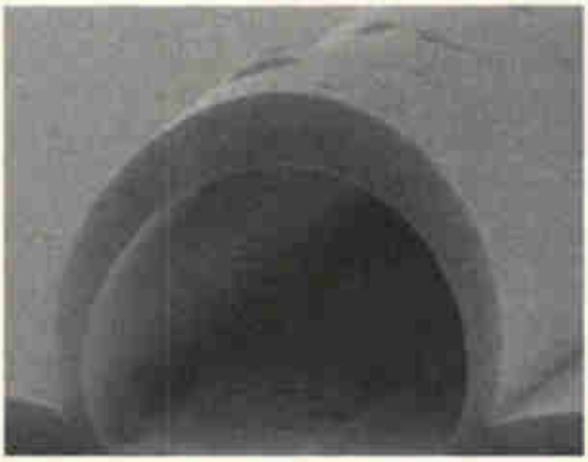
Turbo-codes approach the theoretical limit for error-correction limits set by Shannon and Hartley in 1948. With this form of coding, bit error rates reach 10^{-12} when transmitting 3bits per symbol.

UMTS – the third generation mobile standard – is based on the same Turbo-codes. ST believes that any chip manufacturer supplying silicon for mobile phones will have to take a licence from France Telecom.

Researchers create on-chip micro-canals

Researchers at the US's Sandia National Laboratories have created a technique that creates raised, microscopic canals on chips, through which liquids or gases can flow from one chip feature to another.

Such canals are useful for emerging 'microfluidic' devices that make use of the chemical properties of liquids or gases and the electrical properties of semiconductors on a single chip or among nearby chips.



Better detectors for airborne toxins, rapid DNA analysers for crime-scene investigators, and new pharmaceutical testers for drug development are among the possible future uses for inexpensive microfluidic devices.

The researchers have created raised, hemispherical canals on silicon, glass, and quartz surfaces that are 8 to 100μm in diameter. They've also made canals with tight turns, creating hairpin curvatures with radii as small as 8μm.

The traditional method for creating canals on chips, according to Sandia, is called 'trench and seal' and involves bonding two pre-trenched wafers at up to 1000°C – potentially damaging for other chip features.

To make the Sandia canals, the researchers pattern a thin layer of photoresist on the wafer's surface using a conventional photo mask and light, then develop away areas of the photoresist exposed to the light, leaving a network of photoresist ridges on the wafer's surface that eventually becomes the canals' interiors.

Next they heat the wafer to a relatively low 100°C for about 20 seconds, which causes the square-edged ridges to slump into a hemispherical shape. A 2μm-thick film of silicon oxynitride is deposited over the rounded photoresist, and the entire wafer is soaked in an acetone bath until the remaining photoresist is dissolved, leaving hollow tunnels on the wafer's surface.

The technique is 10 to 100 times faster than trench-and-seal techniques, says co-developer Carol Ashby, and the resulting tunnels are virtually indestructible.

Future micro-fluidic channels might be embedded with electronic controls, micromachines, or novel materials to create an electrokinetic pump or tiny valves or filters. They could deliver cooling fluids to hot spots on today's tightly packed microchips. Or they might help create on-chip insulators that reduce electrical cross-talk.

Academics told to be more commercial

The academic science base needs to be more aware of the commercial application and money-making opportunities of the research taking place in universities, according to the government.

The government's statement came in response to a report by the House of Commons Science and Technology Committee on engineering and science-based innovation. "The government agrees with the committee's conclusion that there is plenty of good research being produced in the UK, but our weakness lies in applying and exploiting that research," said the government.

While the committee blamed industry for not taking up innovative ideas that come from universities, the government said it "does not agree that the responsibility for exploiting the results of research lies solely outside the science base".

Elsewhere in its response to the committee's report, the government admitted that UK spending on R&D was not good enough. According to the committee, R&D as a percentage of gross domestic product has seen a greater drop since 1993 than any other G7 country. The government claims to already be tackling the decline in R&D with an increase in science and engineering spending.

There was also a rejection by the government of the committee's suggestion to expand the R&D tax credit scheme to include large companies. "To provide all firms with a volume tax credit based on a firm's total R&D expenditure would be very expensive, involve considerable dead weight and would not offer good value for money," the government added.

The future's video... This is the Orange Videophone, or at least the phone that Cambridge Consultants have designed for Orange. "This may not be the final form that Orange uses," said a CCL spokesman. It is said by CCL to combine the function of videophone, PDA and Web browser. Orange will not say when it is planning to introduce it commercially, except to say, "Some time in the summer, when we add high speed data capability to our network."

UK's Internet shame

The UK could be the last country in Europe to have high-speed access to the Internet via technology, according to one US telecoms operator.

"It is paradoxical and depressing that Britain, after being the first European country to de-regulate on voice, will be one of the last to de-regulate on broadband." Rupert Baines, European product manager for First Telecom.

Those countries that are already allowing free competition on broadband access are Germany, Finland, Austria, Denmark and the Netherlands. The EU has recommended that all European countries should allow open competition on the local loop by the end of the year. But Britain is not allowing it until July 2001.

The US allowed open competition in 1996, and Germany has had it for six months.

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

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

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

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

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

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

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

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

Web: http://www.tiepie.nl

FEI welcomes mobile health risk report

The mobile phone industry is putting on a brave face and welcoming the rather inconclusive results of the government sponsored report into mobile phone health issues.

The report in summary...

- No evidence to date that emissions from mobile phones and base stations put health at risk.
- Some scientific evidence shows RF exposure may cause subtle effects on biological functions although it is not known if this is detrimental.
- A precautionary approach to the use of mobile phones should be adopted until more information is available.
- The widespread use of mobiles by children for non-essential calls should be discouraged.
- Use of a mobile while driving should be discouraged.
- International standards for maximum exposure levels should be adopted. These are five times lower than current UK levels.
- The specific energy absorption rate of a phone should be made available.
- Base stations should be subject to planning permission.
- Shielding devices and hands-free kits should be independently tested.
- A leaflet should be circulated to every UK household giving information on mobile phone technology and related health aspects.
- A new research programme should be established.

"We believe the report is comprehensive and balanced and we welcome it," said a spokesman for the Federation of the Electronics Industry (FEI) which represents the industry on health issues.

Network operator Orange acknowledged there was public concern on the issue but said it was encouraged by the report's conclusion that "the balance of evidence does not suggest mobile phone technologies put the health of the general population of the UK at risk."

Despite this conclusion the Independent Expert Group on Mobile Phones and Health, originator of the report, has chosen to adopt a precautionary approach to the use of mobile phones.

The Group has been working on the report since September last year.

Lightweight circuit board material takes on FR4

Electromechanical component supplier Radiatron is adding PCBs to its portfolio, with the addition of a lightweight board that the firm claims will compete with traditional FR4 material.

The material, which is made by Dupont, is called Thermount, and boards are produced to military and commercial specifications by the Danish PCB manufacturer, Printca.

A spokesman for Printca said that over 60 million pieces have been shipped for four Japanese companies already for use in mobile telephones.

Thermount-based laminate weighs 25 per cent less than FR-4 and is highly stable. It is seen as having particular applications in high reliability areas. It has a low coefficient of thermal expansion – 8-12ppm per °C – and produces a 'cleaner' hole when drilled using a laser. The dimensional stability – of 0.02 per cent – helps improve registration. But there is a price premium to be paid of between ten and 20 per cent compared with FR-4 laminate.

Commenting on the deal with the Danish board maker, Ian Groves, divisional manager for interconnect and systems at Radiatron, said: "We are now able to offer our customers an extremely competitive PCB service, including lead-free boards."

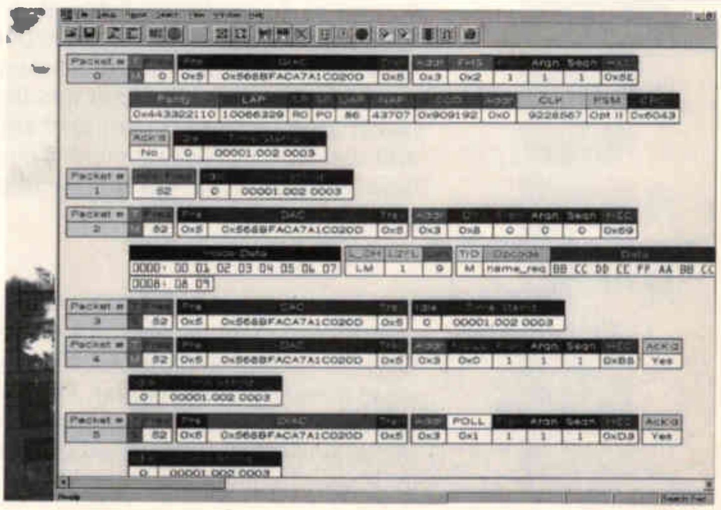
The reduced weight circuit boards will find applications in aerospace. In satellites for example, every kilogram saved in the construction is worth $50 000 in money terms. With each satellite carrying around 12kg of boards, this saving is significant.

"The military specification market in the US is worth some $140m, and given our product portfolio and printed-circuit board pricing policy, we aim to tap into that market," said Groves.

Radiatron's board service includes multi-layer boards up to 26 layers, blind and buried vias, flexible, rigid-flex as well as single and double-sided boards. ■

Free CD for Bluetooth developers

Nohau has introduced a development CD aimed at supporting Bluetooth developers. The Merlin Bluetooth Starter CD is free and contains a trace viewer from CATC that enables users to display, analyse and print Bluetooth protocol traffic files. Also included are copies of the wireless standard's core specification and white papers on the architecture, security and protocols. Nohau is also selling a protocol analyser dedicated to Bluetooth. Expected in products this year, Bluetooth is the new standard for wireless linking of products such as laptop PCs, mobile phones and other handheld equipment. Using a low power 2.4GHz radio link, Bluetooth can connect devices at data rates approaching 1Mbit/s with a range of 10m.



A new wireless network

Cambridge-based Radiant Networks has created clever nodes that can guarantee radio wave coverage in the 20 to 40GHz spectrum. Melanie Reynolds explains how it is done.

Cambridge based Radiant Networks has been working on the problem of radio waves in the 20 to 40GHz region of the spectrum being unable to whizz round, or go through, obstructions.

This fact causes problems in broadband fixed wireless networks because it means line of sight is required between the transmitter and receiver in order for the system to work.

Radiant's intelligent node can connect to four other nodes in a Web-like wireless network.

But Radiant says the technology it is developing can overcome this shortcoming. "It doesn't alter the basic laws of physics, propagation is still line of sight," explains Radiant's technical director, Tim Jackson. "The whole point here is that what it does is change the problem."

Typical broadband fixed wireless networks use a point-to-multipoint (PMP) network.

In this network a base station transmits to a number of subscriber units based at individual customer premises. Because of the frequencies

The intelligent node: what it can do

Radiant's mesh architecture consists mainly of customer nodes, but the same basic product is used for several different purposes.

- **Customer node** – provides network services to the customer's terminating equipment.
- **TNCP (Trunk Network Connection Point)** – is used to make the connection between the mesh and the trunk network. Typical connection bandwidth is 155Mbits/s.
- **MIP (Mesh Insertion Point)** – provides the means by which bandwidth is introduced to the mesh and consists of at least two connected nodes. One of these acts as a high speed interface to the TNCP while the others link to other nodes in the mesh.
- **Seed node** – these are added to the network to provide additional coverage. This could be when the network is initially deployed and there are few customers or if a customer is isolated. As more customers are added to the mesh the seed nodes become redundant.

used, the subscriber unit must be able to see the base station in order to receive signals from it. If there is a building or tree in the way the signal is blocked.

The appearance of obstacles in the path is obviously a common occurrence and the standard way round this is to overlap the base station coverage areas. This means customers can be reached from a variety of different directions but this approach has disadvantages. Base stations are expensive and because multiple frequencies have to be used to cover a particular area it, "knackers the spectral efficiency," according to Jackson.

Radiant uses an Internet style mesh configuration. The mesh system does not use a network of base stations and subscriber units. Instead it uses intelligent nodes equipped with four small directional antennas. These nodes combine the functions of the subscriber unit and base station equipment into one product.

Four rotatable beams

The antennas generate four separate microwave beams at frequencies between 28 and 40GHz, delivering up to 25Mbit/s in full duplex connections.

Under the control of the network, the antennas are able to rotate through 360 degrees. This allows each node to connect to up to four other nodes thus creating the mesh configuration. As the technology is adopted, this will mean a large number of other nodes being in the line-of-sight. "Instead of one chance to see a base station I could have tens, hundreds, maybe even thousands of chances and I really only need one of them," exclaims Jackson.

Each node transmits and receives from other customer nodes so instead of a base station providing the coverage, the customers are actually providing the coverage.

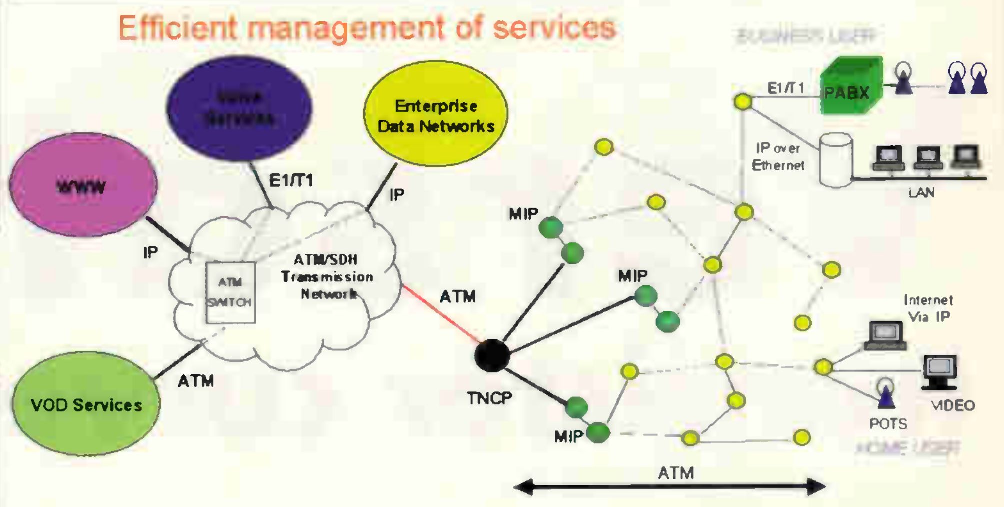
Radiant claims the system capacity is 50 times higher than PMP systems and it provides near 100 per cent coverage. Because narrow radio beams are used for the links between nodes, this allows frequencies to be re-used more frequently.

Protocols that can run over the link include asynchronous transfer mode, or ATM, and Internet protocol, IP.

Domestic services too

Initially, the system is aimed at business but it is also designed with the residential market in mind. Services that could be offered using the system include video-on-demand, video conferencing, interactive television, high-speed internet access and telephony.

Despite combining the base station and subscriber functions in one product, and it having four antennas, Radiant is happy its mesh network is competitive. "The cost of the node is very similar to the cost of subscriber equipment in point-to-multipoint but we don't have the base stations," emphasises Jackson. "You're getting a lot more performance and coverage but not actually paying significantly more per



Full service network with intelligent bandwidth management

The antennas generate four separate microwave beams at frequencies between 28 and 40GHz delivering up to 25Mbit/s full duplex connections

Radiant Networks: a brief history

Radiant Networks is based in Cambridge and was originally spun out of communication specialist Plextek in October 1997.

The company now employs 27 full time staff, but is currently recruiting and expects to have over 100 by the end of this year.

Venture capital funds of $20m were secured in April this year. The funding, from a number of venture capital firms including Advent Venture Partners and Intel Capital, will allow the firm to start trials and manufacture a product.

During the past three years the company has raised around $10m of funding.

Versatile transistor curve tracer

Useful for selecting matched pairs, this transistor curve tracer tests bipolar devices, junction field-effect transistors and metal-oxide-silicon field-effect transistors – both p and n types in each case. Ian Hickman describes its development, and illustrates its operation with typical traces produced by each type of device.

A transistor curve tracer provides an instant overview of the operation of a transistor, be it a bipolar type or an FET, covering a range of combinations of voltage and current. As such, it is more revealing than a spot h_{FE} measurement on a transistor tester – even though such a test may be quantitatively more precise.

A curve tracer is particularly useful for selecting matched pairs of devices, either of the same polarity, or complementary devices, for use in audio power amplifiers.

In days gone by, transistor curve tracer plug-ins were available for a well known make of oscilloscope. Also, now, as then, oscilloscopes designed specifically as transistor curve tracers are available. However, I had long intended to make myself such an instrument, using my existing TEK485A as the display.

The instrument described below is the result. It fits my limited requirements admirably, but those of you with more ambitious ideas will find the circuit readily modifiable, for extension to testing at higher voltages and currents.

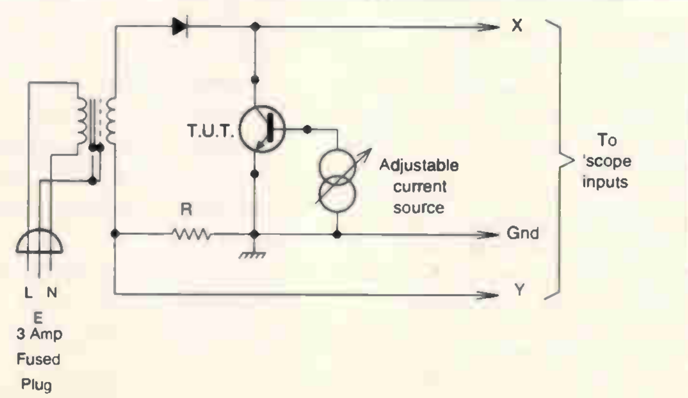


Fig. 1. Basic transistor curve tracer, plotting collector-emitter voltage against base current.

Curve-tracing principle

A transistor curve tracer should display the collector current of the transistor under test as the ordinate – or Y co-ordinate – of a plot. Varying collector/emitter voltage V_{ce} is the X co-ordinate, or abscissa, with several different fixed values of base current as parameter.

Figure 1 shows a convenient arrangement using a floating ac supply as the V_{ce} supply for an n-p-n device. Low value series resistor *R* is added to monitor the collector current. The fact that some of the supply voltage is dropped across *R* is of no significance. The arrangement displays the true characteristic provided only that the base current remains constant.

There is however a trade-off in exchange for for the convenience of Fig. 1: the voltage drop across *R* is due to the emitter current rather than solely to the collector current. Except for very low gain transistors, the difference is small, and the arrangement of Fig. 1 is commonly employed. The curve tracer described here adopts the same scheme.

The circuit

Figures 2 and 3 show the circuit. Supplies for the op-amps are derived by half-wave rectification from a tap on the mains transformer T_1. This was an obsolete RS rectifier transformer, rated for use at 0.7A dc output.

Although a high-gain transistor under test will draw more than 0.7A, that only happens on one step out of eight, and then only on alternate half cycles. As a result, the transformer actually leads an easy life – even when testing at high-currents.

The arrangement used permits testing at values of 12, 24 and 30V nominal V_{ce}. Other mains transformers, preferably with two or more secondary sections, may obviously be employed.

In Fig. 2. the transistor being tested is shown as an n-p-n device, fitted on one of three test sockets, all in parallel. The varying collector voltage is supplied by the secondary windings of T_1. via S_{1A} and D_1 in this case, since the transistor being tested is n-p-n.

Base current supplied via R_2 is constant over the whole of the positive half cycle of voltage applied to the collector of the transistor under test. The '*R*' of Fig. 1 is here represented by R_{12} plus R_{10}. or R_{12} alone for higher current tests. Op-amp IC_{1B} amplifies this volt drop by a factor of 20, giving an output scaling of 1V/100mA or 1V/1mA, depending on the setting of S_5.

Switch S_5 is arranged so that on the high current ranges, the contact resistance of the switch shorting out R_{10} does not add to its resistance and form part of the current sensing resistance.

Output at Y connects to the Y input of an oscilloscope. while the collector voltage at X_1 connects to the X input. the scope being set to operate in X/Y mode of course.

By using the Y setting at 1V/div. or 100mV/div. in conjunction with S_5. current ranges of 100μA. 1mA. 10mA and 100mA per division are available. Other Y 'scope sensitivities provide yet more ranges. such as 50mA. 20mA per division etc. Setting the Y input to 2V/div. provides a 200mA/div. range. up to the limit of the available swing from IC_{1B}. This is equivalent to. say. 1.4A or more.

Originally. I had intended to supply the base current. which must be held constant at each of the various steps of the I_c parameter, from a Howland current pump[1]. However, initial design work suggested that it might be difficult for a Howland-type circuit to accommodate the large range of base currents that I had in mind.

So another scheme was used, borrowed from an article that appeared recently in these pages[2]. This used the near infinite input resistance of a FET-input op-amp to sense the base-emitter voltage of a transistor. without affecting the base current.

The op-amp in question. IC_{1A}, provides a base reference voltage for the base-current generator section, which is shown in **Fig.** 3. This reference voltage equals the V_{be} of the transistor under test. or nearly so, allowing for the small attenuation provided by R_{52} and R_6.

Op-amp IC_{2A} produces a mains-derived 50Hz clock waveform. This signal clocks 12 stage ripple counter IC_3, of which only the first three stages are used. The seven-stage CD4024 would perhaps have been a more logical choice, but I had a 4040 to hand. Either will do – but not the 14-stage CD4020. as in this device, Q2 and Q3 are not available.

Resistors R_{44-46} produce a ramp of eight 25μA steps, resulting in a staircase of 1V steps at the output of IC_{1D}. At the output of IC_{1C}. the staircase is positive going for n-p-n transistors under test. or negative for p-n-p. as appropriately selected by S_3.

By means of S_2, base current steps in a 1, 2, 5 sequence from 1μA up to 1mA. can be applied to the transistor being tested. As the voltage steps applied to S_1 are produced relative to

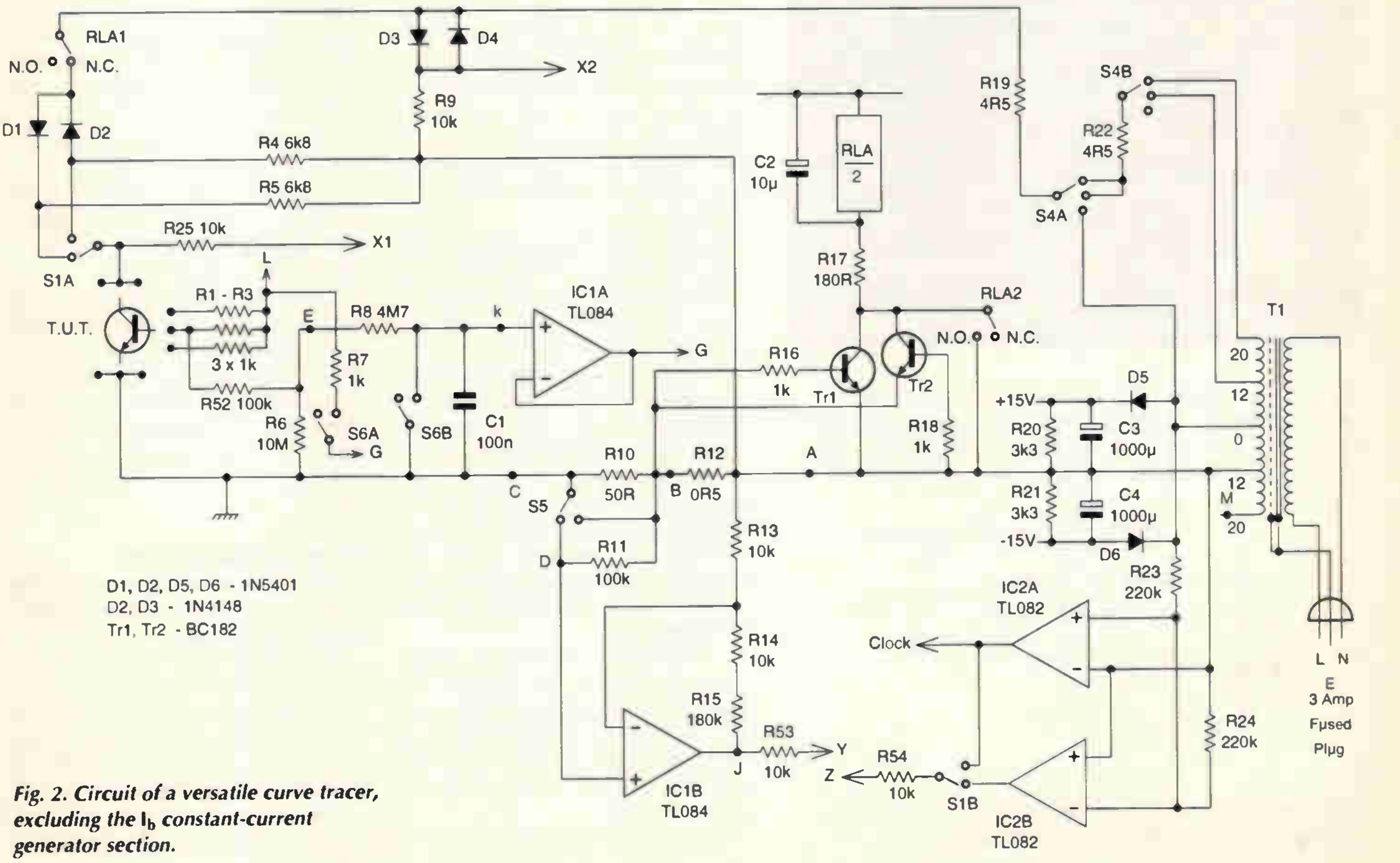


Fig. 2. Circuit of a versatile curve tracer, excluding the I_b constant-current generator section.

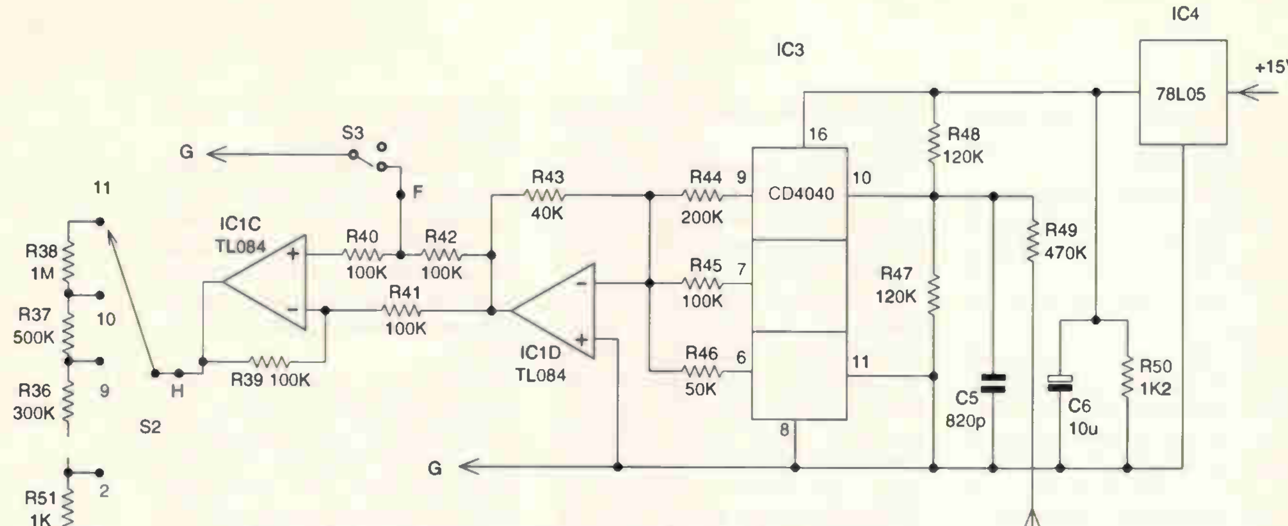


Fig. 3. The I_b constant current staircase generator section. A 4024 is a more elegant choice of counter, but I had a 4040 to hand, and it works.

point G, i.e. relative to the V_{be} of the transistor under test, both silicon and germanium devices can be tested, without errors due to V_{be} variations.

Managing the trace

To display a set of eight I_b/V_c curves, the curve tracer uses half-cycles of the secondary voltage, eight consecutive positive-going half-cycles for n-p-n devices and eight negative for p-n-p.

At the peak of each V_c trace, the spot on the screen of the oscilloscope is comparatively slow moving, compared to the rest of the trace. So that part of the trace appears brighter, as you can see in the screen shots illustrated.

However, each is at a different point on the screen. But with the oscilloscope's X input taken from X_1 of Fig. 2, during each and every alternate unused half-cycle, the spot remains at the origin of the parametric set of curves. The result is a bright spot, clearly visible in **Fig. 4**. This carries with it the danger of a burn mark, due to local overheating of the phosphor.

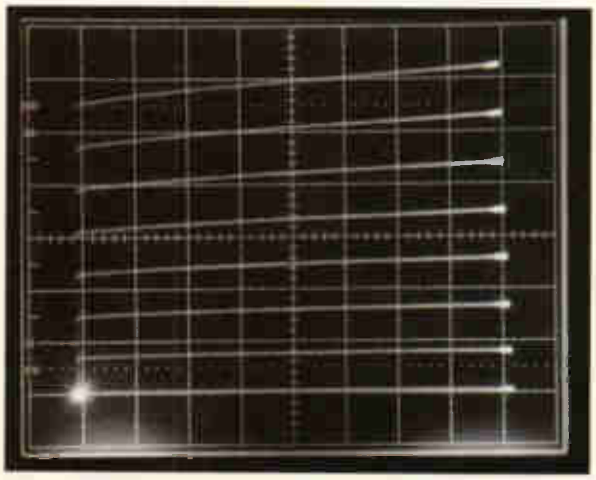


Fig. 4. Display of the I_b/V_{ce} characteristics of a BC108B, 1mA/div. vertical, 2V/div. horizontal. The excessively bright spot at the origin, where the beam rests for 50% of the time – with danger of spot-burn – is only too obvious.

To avoid this, IC_{2B} produces a 30V peak squarewave, for application to the Z modulation input of an oscilloscope. Unfortunately, it is not suitable for use with my TEK475A, as the input resistance of the Z modulation input, which operates up to 50MHz, is 15KΩ. The problem is that there can only be one 0V chassis connection between the oscilloscope and the curve tracer, which means that the signal current drawn by the Z modulation input must return via $R_{10}+R_{12}$. It is then indistinguishable from the current through the transistor being tested.

If your oscilloscope has a 1MΩ Z modulation input, and you are not interested in testing at very low levels of I_c, then the Z output from IC_{2B} will work fine. In my case, and doubtless others, another solution must be sought.

One solution is shown in **Fig. 5**.

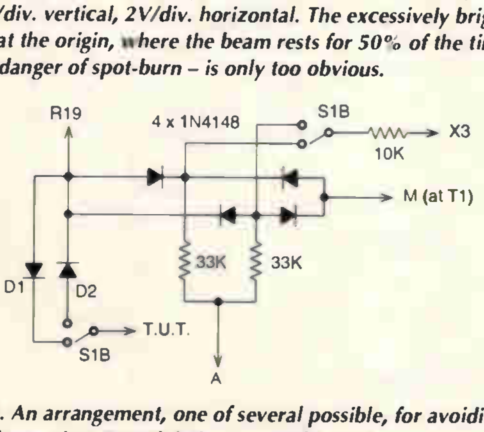


Fig. 5. An arrangement, one of several possible, for avoiding spot-burn when Z modulation cannot be used.

Here, the collector of the transistor being tested receives half cycles of the voltage selected at S_4 via S_{1A}. These are positive for n-p-n, or negative for p-n-p.

A similar half cycle of voltage appears at one input of S_{1B}, but the other half cycle also appears due to the diode connected to output M of the transformer.

Switch S_{1B} selects X_3, the version appropriate for n-p-n or p-n-p testing, and the result is as shown in **Fig. 6**. You can see from this diagram that instead of dwelling at the origin for 50% of the time, in the unused half cycles, it draws an 'underline' to the I_b=0 curve.

However, if like my oscilloscope, the amplifiers of yours will accept a 100% overscan beyond the graticule without complaint, the simplest arrangement is the one I have adopted, shown as X_2 in Fig. 2.

On n-p-n testing, the drop in D_3 closely matches that in D_1, while D_4 serves the same purpose in relation to D_2 for p-n-p working. Either way, on the unused half-cycles, instead of remaining stationary, the spot is deflected way off-screen, in the opposite direction to the set of curves.

Overload protection

With an instrument such as this, protection is an important consideration at the design stage.

My initial idea was to detect excess emitter current via R_{12}, and use this to force a high level on the reset line, pin 11 of IC_3, removing the base drive to the transistor being tested. But this would not allow for the connection of a device with a collector-emitter short circuit, or worse still, a device which develops a collector-emitter short circuit during testing.

Instead, Tr_1 – or Tr_2 in the case of a p-n-p fault – detects when the current through R_{12} exceed about 1.2A and operates relay RL_A. Contact RL_{A1} thus clears the short or overload, while contact RLA_2 provides a self-hold function.

To reset the circuit after removing the faulty device – or setting a lower base current – the mains supply to the unit must be switched off. Resistors R_{19} and R_{22} have no effect on small signal devices, but reduce the collector dissipation of the transistor under test on the highest-current steps. Their effect can be seen in some of the screen shots of devices under test, shown in the following illustrations.

this reason, the four highest current I_b ranges of S_2 should only be used in conjunction with the centre socket. Plug-in adapters were made for this socket, to take TO3, 'plastic TO3' and other large devices.

In addition to a choice of I_b step sizes, with the particular mains transformer used, three different collector voltage ranges are available.

...and other devices

The task I set myself was to design a curve tracer that would also characterise field-effect transistors – both MOSFETs and JFETs – in addition to bipolar transistors. For this purpose, constant gate/source voltage V_{gs} steps are required, rather than constant I_b steps.

An obvious way to achieve this is

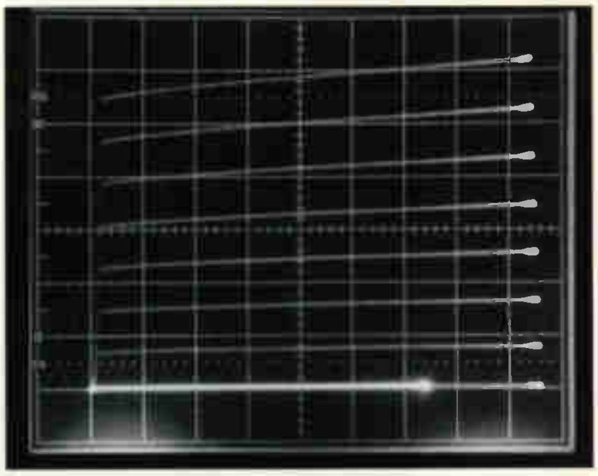


Fig. 6. The same device and characteristic as Fig. 4, but using the arrangement of Fig. 5 to avoid spot-burn.

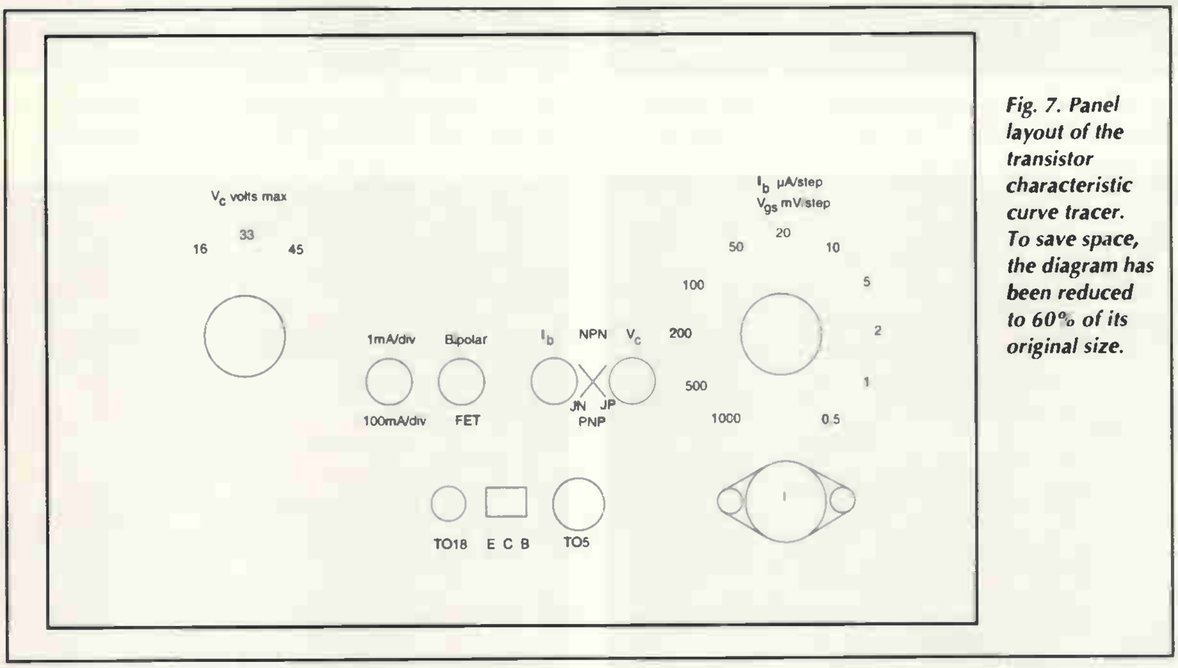


Fig. 7. Panel layout of the transistor characteristic curve tracer. To save space, the diagram has been reduced to 60% of its original size.

Checking bipolar transistors...

A wide choice of base currents is available, so that both small signal transistors and power transistors can be accommodated. Small-signal devices can be plugged directly into one of the three sockets provided, one for TO18 types, one for TO5 and a third between them, with three holes in a row, for TO92. The TO92 option has the usual centre-pin-collector layout, **Fig. 7**.

To minimise the possibility of parasitic oscillations during test, a 1kΩ 'base-stopper resistor' is mounted directly at the base pin of each socket. This resistor also doubles as the base-current defining resistor for the 1mA/step range, position 1 of S_2.

Op-amp IC_{1A} monitors the voltage at the base connection of the centre socket, and hence via R_2 when one of the other two sockets are in use. For

simply to dump the I_b steps to 0V, via a 1kΩ resistor, giving steps of 1V, 0.5V etc., depending on the setting of S_2. However, this would result in I_b flowing through R_{10}+R_{12}, and so being indistinguishable from the drain current I_d.

The solution is provided by S_6, which is shown in Fig. 2 in the 'bipolar' position. In the FET position, S_{6B} shorts the non-inverting input of IC_{1A} to 0V, point C, providing a low impedance pseudo 0V ground at IC_{1A}'s output point G.

Switch S_{6A} connects a 1kΩ resistor R_7 to this, giving increments of V_{gs} relative to 0V without passing any current through R_{10}+R_{12}. Op-amp IC_{1A} sources or sinks the current, swallowing it and returning it to the plus or minus 15V rail as appropriate.

A further complication is that while most MOSFETs require a V_{gs} of the

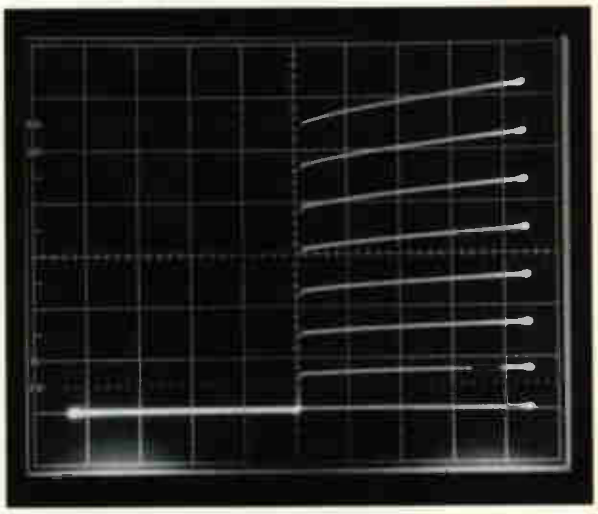


Fig. 8. The same device and characteristic as Fig. 4 again, but this time using the X_2 output from fig. 2. to drive the X sweep.

same polarity as V_{ds}. JFETs require a V_{gs} of the opposite polarity – as do n-channel MOSFETs of the depletion variety. This is the reason for making S_1 and S_3 separate switches.

For n-p-n bipolars and enhancement n-channel MOSFETs. S_1 and S_3 are used in the positions shown in Fig. 1, while for n-channel JFETs, S_3 only is set to the other position, as indicated by the 'JN' line in Fig. 7.

Some typical results

Figure 8 shows the same device and settings as in Fig. 4. but using the X2

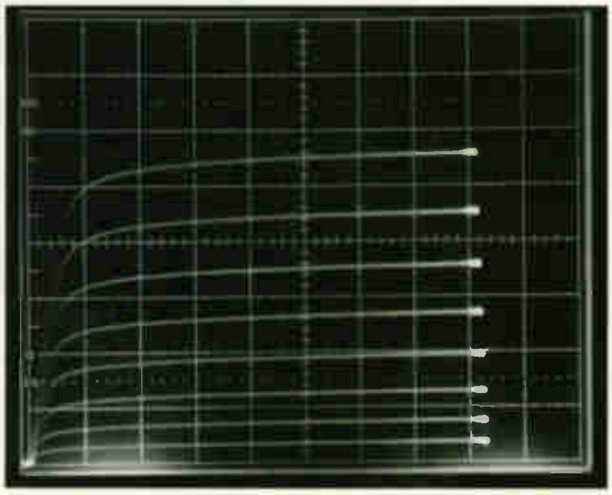


Fig. 9. Characteristics of an n-channel JFET: BF244A, nominal 12V V_{ds}, 1mA/div vertical, 2V/div horizontal, 0.2V/step V_{gs}.

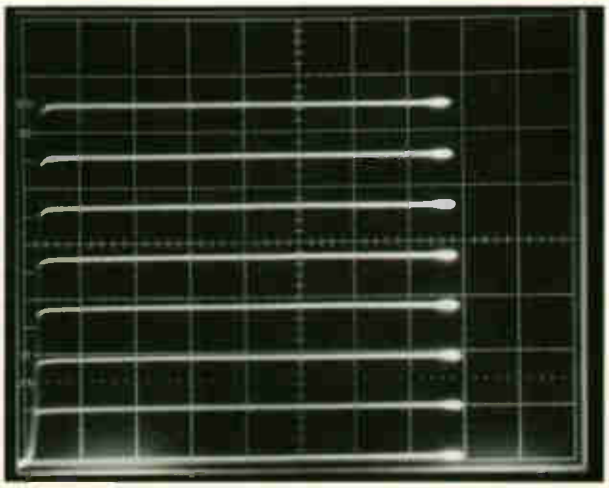


Fig. 12. Characteristics of a VN46AF N channel MOSFET, 1mA/div vertical, 2V/div horizontal, 1V/step V_{gs}, 12V nominal V_{ds}.

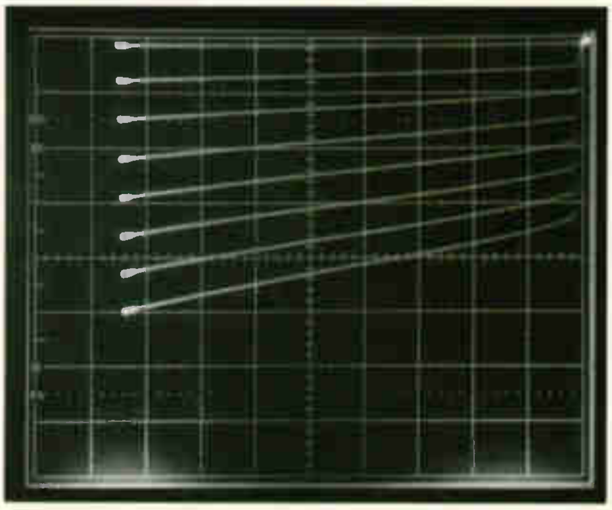


Fig. 10. Characteristics of a BC214L; nominal 12V Vds, 1mA/div vertical, 2V/div horizontal, 2μA/step I_b.

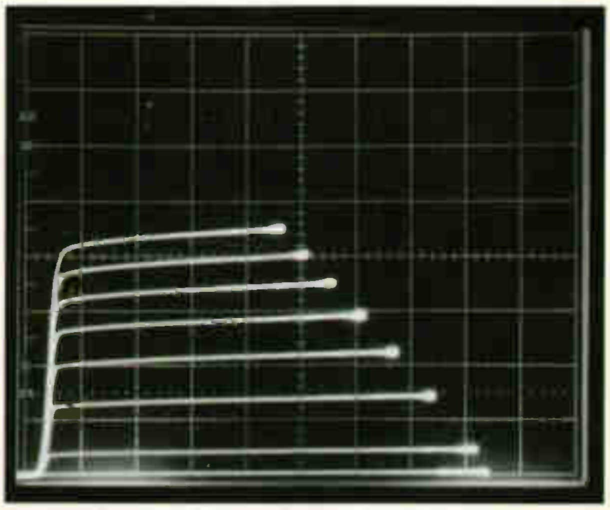


Fig. 13. Characteristics of a TIP41A n-p-n silicon power/switching transistor, 200mA/div vertical, 2V/div horizontal, 1mA/step I_b, nominal 12V for V_{ce}.

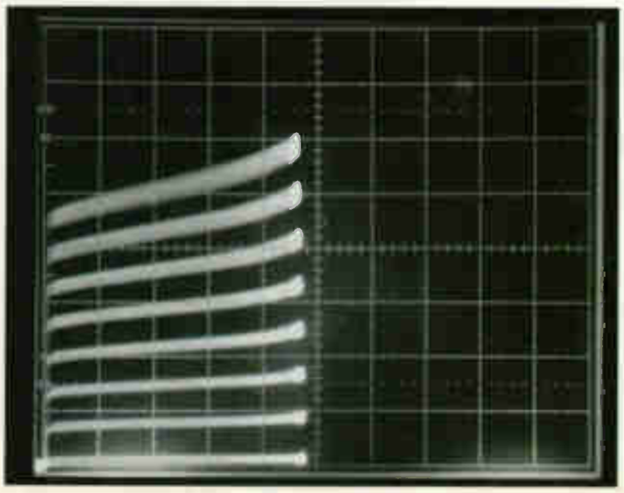


Fig. 11. Characteristics of a BC184B, 10mA/div vertical, 10V/div horizontal, 20μA/step Ib, 32V Vce nominal.

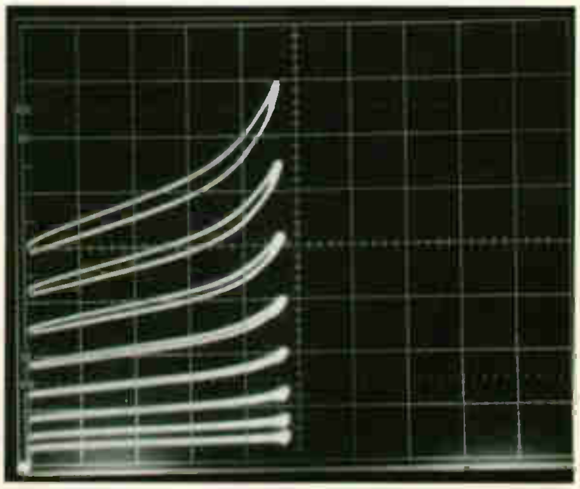


Fig. 14. Characteristics of a BC182, 2mA/div vertical, 10V/div horizontal, 20μA/step I_b, 32V nominal for V_{ce}, see text.

output of Fig. 2 to prevent burn spots. Normally, the trace would be set one major graticule division lower and five to the left. In this way, the origin of the parametric set of curves would be at the extreme bottom left of the graticule.

Figure 9 shows the characteristics of an n-channel JFET, namely a BF244A, at a nominal 12V V_{ds}, 1mA/div vertical, 2V/div horizontal, 0.2V/step V_{gs}. Note that the top curve corresponds to V_{gs}=0V, the I_{dss} curve, and that the device is largely cut off by a V_{gs} of −1.4V.

Figure 10 shows the characteristics of a p-n-p bipolar device, a BC214L. For a p-n-p device, of course, the origin is at the top right. The characteristics are shown at a nominal 12V V_{ds}, 1mA/div vertical, 2V/div horizontal, 2μA/step I_b.

Note that comparing these curves with those for the BC108B in Fig. 4, the BC214L has a distinctly lower collector slope resistance.

Figure 11 shows a BC184B with settings of 10mA/div vertical, 10V/div horizontal, 20μA/step I_b, 32V V_{ce} nominal. With the oscilloscope Y input set to 1V/div, S_5 gives a choice of 1mA/div or 100mA/div vertical. The 10mA/div setting was obtained by using 100mA/div at S_5, but increasing the Y sensitivity to 100mV/div.

Power issues

You can see that at the maximum voltage excursion on the highest current step, the instantaneous dissipation in the TO92 style plastic device was close on 3W. Of course, the average was less than this, but the gradual heating up of the device, over the 20 second exposure required by my home-made oscilloscope camera, is clearly visible; if left on test much longer, the device might have expired.

Figure 12 shows a VN46AF n-channel MOSFET on test, with settings of 1mA/div vertical, 2V/div horizontal, 1V/step V_{gs}, 12V nominal V_{ds}.

This device is remarkable for two of its characteristics. Firstly, the drain slope resistance is very high, regardless of the actual value of drain current. Secondly, the mutual conductance, g_m, is very nearly constant, from 0V V_{gs} upwards, as evidenced by the nearly equal steps of I_d. This is clearly a very linear device for small/medium signal amplification.

Figure 13 is the result of testing a TIP41A n-p-n silicon power/switching transistor. The settings were 200mA/div vertical, 2V/div horizontal, 1mA/step I_b, and 12V for V_{ce}

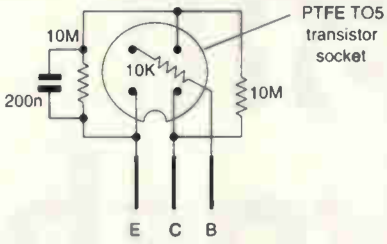


Fig. 15. Circuit diagram of an appliqué unit, used to permit the curve tracer to be used for evaluation dual-gate MOSFET devices.

nominal. Peak collector dissipation of over 8W would have been even greater but for the effect of R_{19}.

The reduction of peak collector voltage on the higher current traces is evident from the photo. On the 24V and 32V nominal V_{ce} settings, R_{22} is brought into circuit, further limiting the peak current. These resistors also help to limit the peak prospective fault current that contact RLA_1 may be called upon to clear.

Figure 14 shows the results for a BC182, at 2mA/div vertical, 10V/div horizontal, 20μA/step I_b, and 32V nominal for V_{ce}. Dissipation was much less than in the case of the BC184B, so there was none of the gradual heating up of the device over the period of the exposure visible in Fig. 11.

Nevertheless, the heating and cooling of the silicon pellet at each individual I_b step – especially the higher current steps – is clearly in evidence.

Modifications for dual-gate FETs

While the instrument will display the characteristics of both JFETs and MOSFETS, both depletion and enhancement, as it stands it will not display those of dual-gate MOSFETs. These devices are often used as AGC-able RF amplifiers at VHF and UHF. But a simple appliqué unit will remedy this lack.

Figure 15 shows the circuit of such an add-on, designed to plug into the central one of the three test sockets. With the component values shown, gate 2 of a dual-gate n-channel MOSFET is held at about +3V – a fairly typical sort of value. Any desired value could be arranged, by changing one of the two 10MΩ resistors; you could even substitute a 10MΩ preset potentiometer for one of them, to permit evaluation of the characteristics of the device under test at different values of V_{g2}.

I used the circuit of Fig. 15, constructed on a scrap of copper-strip matrix board, to record the collector

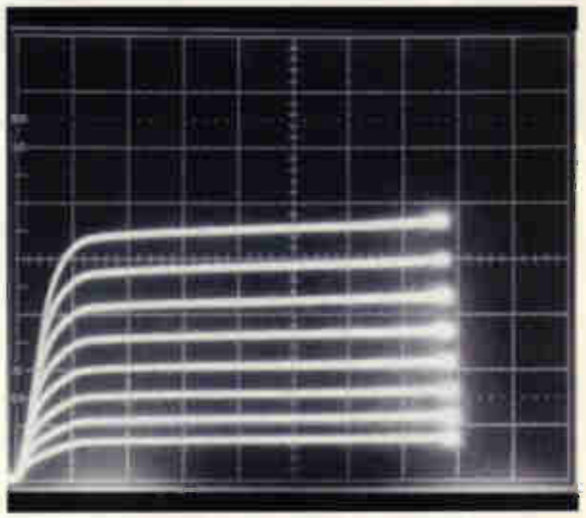


Fig. 16. Characteristics of an MEM614 dual-gate MOSFET, 2mA/div horizontal, 100mv/step V_{g1}, 16V nominal V_{ce}, see text.

voltage/drain current characteristics of a General Instrument Corporation dual-gate MOSFET, type MEM614, **Fig. 16**.

In summary

The circuit presented is a useful addition to any laboratory's instrumentation – especially if you need to select matched devices for use in power amplifiers. Incorporating 1kΩ stoppers has resulted in the absence of parasitic oscillations, which can be a problem especially when testing RF transistors.

If the unit will frequently be used for testing enhancement-mode MOSFETs, then a useful modification might concern the biasing arrangements. The linear results shown by the VN46AF of Fig. 12 are by no means typical of MOSFETs. Many of these do not conduct appreciably at all until the V_{gs} has risen to several volts, whereafter another volt or so switches them on fully.

A means of raising the potential of point G by a pedestal of a few volts, so that V_{gs} steps of 1V or even less can be applied, relative to the pedestal, would be useful for many power MOSFETs – especially those specifically rated for use in switching circuits. ■

References

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2. March, Ian, 'h_{FE} tester uses no meter', Electronics World, April 2000, p296.

Bluetooth whitewash

As life styles become more mobile, people are looking to wire-free technology to put critical information at their fingertips. It's a wireless world, and a Bluetooth one if you believe all you're told. Not necessarily though; other technologies may be more appropriate. Andrew Emmerson compares the contenders.

It's funny how some technologies catch the mood of the moment. Last year you couldn't open the trade press without reading about MP3, while this year if they're not wearying you with WAP, then they're sure as heck boring you with Bluetooth.

It's a triple tragedy really: as a technology Bluetooth is in grave danger of being over-hyped and misapplied to the extent that it will never achieve all these expectations. What's more, it's by no means the only technique for short-range wireless communication – whatever you might be led to believe – and it certainly shouldn't be applied to the exclusion of other, more appropriate solutions.

Most ironic of all, the current hoopla and tedium being invoked in the name of Bluetooth is poor tribute to the memory of a Viking ruler whose historical claim to fame was for spreading sweetness and harmony. Still, who said life was fair?

In the air and everywhere

'Wireless' – a word that until recently sounded as archaic as horse-drawn milk floats – is suddenly sexy again, having been appropriated by marketeers who find it more meaningful than, say, wire-free. Be that as it may, the umbilical cords that used to tie people to a fixed location in order to work are vanishing.

Within the workplace or out and about, people expect to have access to phone and data communication and now to powerful multimedia applications as well – regardless of location. It's lucky then that this dream of 'everywhere, all the time' wireless access to information is finally becoming a reality.

Just as people's perceptions of reality vary, however, so do the means of achieving the wireless reality. However much users might wish, providing universal access by means of a single technology is remarkably difficult – as any one who has tried to use a mobile phone deep inside a building will know.

Indoors and outdoors are totally different environments to radio system designers. Providing wireless connectivity within offices and homes calls for techniques developed specifically for those situations.

Three technologies are in contention, named HiperLAN/2, HomeRF and Bluetooth, but in fact contention is the wrong word. Although the platforms may appear as competing solutions, they serve radically different purposes.

Fortunately the task of choosing the right wireless product for users' applications and working style will be simplified by 'oven-ready' solutions offered by manufacturers. End users will have little or no part in any decision-making process. At least that's what one hopes, although there are already signs that Bluetooth may be employed as a marketing gimmick and persuader.

So what are Bluetooth, HomeRF and HiperLAN/2, and how do they differ? All three are essentially short-range radio systems using ultra-low power microwave radio frequencies and both also replace cable connections. But beyond that they differ in their concept and function.

Contenders contrasted

Bluetooth was described in the article 'What Is Bluetooth?' on page 421 of the May 2000 issue. It is a replacement for the leads that currently tie a printer to a PC or a video recorder to a TV screen. In effect, Bluetooth connects objects that have a logical affinity and never move far apart.

HiperLAN/2, on the other hand, is a local-area network or LAN product in which the fixed data cabling of conventional networks is replaced by a radio link that floods your complete premises. It does so in the same way as a domestic cordless telephone does at home.

You can then use laptop computers and other portable devices anywhere you like and link them to shared file servers or the Internet. Cross-industry collaboration is working to ensure standardisation for both techniques and heavyweight support from the major manufacturers will ensure their products are fully conformant and thus interoperable.

HiperLAN/2 can carry voice, data

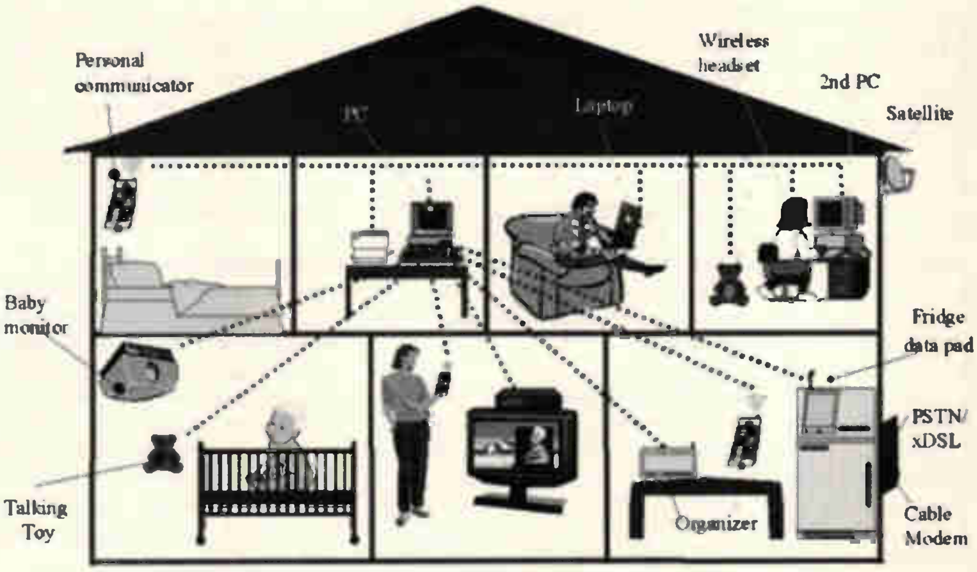
Want to find out more?

Entering HomeRF, HiperLAN/2 or Bluetooth into any Internet search engine will produce more hits than you'll have time to read. However, the official Bluetooth website can be found at *http://www.bluetooth.com*
and others at
http://www.bluetooth.com/bluetoothguide/intro/intro2.html
http://www.anywhereyougo.com/ayg/ayg/bluetooth/Index.po
HiperLAN/2 sites are at,
http://www.hiperlan2.com/site/home.htm
http://www.etsi.org/technicalactiv/hiperlan2.htm
...and HomeRF can be found at
http://www.homerf.org/

and video, which will appeal to facilities managers and office administrators looking for a simple, reliable, invisible and unified replacement for the spaghetti junction of cabling that leads to every desk and hinders unrestricted movement of desks and departments.

HomeRF – the third contender – is an intermediate solution. Like HiperLAN/2, it is an integrated, full-function wire-free replacement for structured cabling but without the capacity for intensive business use.

As its name implies, HomeRF is designed primarily to serve the domestic and small office environment, supporting data and telephony in a robust fashion. It does not supplant Bluetooth as such and users may well require both technologies together.



HomeRF is a wireless local-area network with a range of 150 feet, designed to provide communication throughout a single household. Its 'shared wireless access protocol' operates in the 2.4GHz band and uses digital frequency-hopping spread spectrum radio.

Horses for courses

Xircom Europe, a leading supplier of mobile information access solutions, has experience of Bluetooth and HiperLAN/2 and intends to exploit both of them on a 'horses for courses' basis.

Earlier this year at Cebit, Xircom announced plans to integrate a whole array of wireless technologies, including Bluetooth and HiperLAN/2 into offerings that could be customised easily by the corporate enterprise, small and mid-size business, home-office worker and consumer.

Bluetooth is a kind of 'wireless wire' used for what Xircom conveniently terms 'personal-area networks'. Dispensing entirely with cables and plugs, Bluetooth technology transfers data via short-range, low power – typically one milliwatt – 2.4GHz radio signals.

The data transfer rates of 1Mbit/s – 20000 times faster than a normal telephone modem – create a connection environment with no wires for devices such as laptop computers, mobile phones, handheld PCs and wearable information accessories. This allows all these devices to exchange and synchronise up-to-date information, and to connect to the Internet at high speed – all without the need for wires or cables.

When Bluetooth-enabled devices are within a 10-metre range of each other, users can transparently transfer schedule and address book information between their accessory device and notebook PC or automatically dial through their mobile phone to access the Internet. Intelligent handsets will even reduce the volume of their ringing signal when they detect the presence of other Bluetooth signals.

Untethered access

At the same time, changing workgroup needs and the increasing mobility of employees, combined with the higher-speed connectivity and continually decreasing prices, are making wireless LAN solutions extremely attractive to businesses large and small.

HiperLAN/2 meets this need by providing 'untethered' network access for computer users across an office, factory, campus or similar local area. It is also touted as a link mechanism for PCs, VCRs, cameras and printers in the home. HiperLAN/2 operates in the 5.8GHz microwave band with data transfer rates of up to 54Mbit/s. It looks set to trigger the mainstream adoption of wireless LAN systems.

Home networking

Serving less intensive demands, the 'shared wireless access protocol', or SWAP, of HomeRF operates in the 2.4GHz band and uses a digital frequency-hopping spread spectrum radio. Its principles are derived from existing DECT cordless and wireless LAN technology. SWAP supports both a time-division multiple access, or TDMA, service to provide delivery of interactive voice and other time-critical services, and a carrier sense multiple access/collision avoidance, or CSMAICA, service for delivering high-speed packet data.

A single network supports up to 127 data devices and six full-duplex voice channels. More than one network can be operated concurrently in the same location.

Range is stated as 150 feet – adequate for most homes and gardens. The next generation of HomeRF technology will provide enhanced wireless multimedia services in the home, it is claimed.

Imperfect solution?

While all three technologies will revolutionise in-house connectivity, they are still localised answers. As such, they will not solve the problems of nomadic workers who expect seamless telephone and Internet access to home, work and on the move.

Seamless remote access demands a wide-area network – a WAN as opposed to a LAN – and manufacturers are working on their wireless equivalent. As it happens, HiperLAN/2 would be well suited to these applications also. Planners are already considering deploying open-access HiperLAN/2 networks in hot spot areas such as airports and hotels.

Users of laptop and notebook PCs

Why 2.4 and 5.8GHz?

Radio frequencies are a precious and scarce commodity – something not normally surrendered to Johnny-come-lately users or those unprepared to pay a licence fee.

Charging for the use of the airwaves, however, would render wireless networking less attractive – and for some users, entirely unaffordable. Governments in most countries have therefore permitted wireless networking in frequency bands set aside many years ago for industrial, scientific and medical (ISM) use.

The upside for users is unrestricted access to the airwaves at no charge – subject, of course, to observing prescribed power levels. The downside is that these bands are already occupied by other users, some powerful and 'dirty', such as microwave ovens. This means that users must tolerate a level of mutual interference. There is no protection against this on these crowded frequencies and it is up to users to employ frequency hopping and/or other protocols that minimise the effects of interference.

Terming these frequencies 'unlicensed' or 'unrestricted' is mistaken by the way; a licence is required but in this case it held on behalf of all users (in Britain) by the Secretary of State for Trade & Industry. Restrictions do apply to use but end users need not worry about these so long as they use approved apparatus.

Give us a clue!

It takes little imagination to decode HiperLAN/2 as a second-generation high performance LAN but what about Bluetooth? Easy! King Harald Bluetooth was a Viking king who brought about peace and harmony, which is also the aim of the new Bluetooth standard.

can already buy wireless data connection kits that connect to GSM mobile telephone handsets. Solution providers are now developing means of connecting them up to the 'third-generation' mobile radio systems currently under construction. These developments are entirely separate from 'indoor' solutions such as Bluetooth and HiperLAN/2 though, and it remains to be seen how wide-area and local-area access can be made compatible with one another.

Technical troubles

Other potential concerns – among them multi-vendor compatibility and user density – are also brushed aside frequently. But they need to be addressed sooner rather than later. Patrick Rada of US retailer Hello Direct sums it up neatly: "How can you be sure your IBM Bluetooth notebook will be able to speak with your Toshiba Bluetooth PDA and your Hello Direct Bluetooth headset? And what if dozens of Bluetooth users use their Bluetooth PDA, headset, or device, simultaneously in the shopping mall?"

Because Bluetooth sends very short packets and changes frequency rapidly, packet collisions will occur increasingly as communication density increases. This will lead to decreased data throughput and noisy audio. To address these issues, says Rada, it is likely that the Bluetooth standard will need refining several times, involving major effort and investment.

The unregulated frequency band that Bluetooth shares with HomeRF and many other radio devices may create problems for Bluetooth. Not all devices on this band operate at equal power levels. Wireless LANs for instance use the same 2.4GHz band as Bluetooth and domestic applications may be stymied by interference from microwave ovens, also on 2.4GHz. Mobile phones with Bluetooth interfaces may find the Bluetooth receiver desensitised when they transmit.

Gargantuan growth...

None of the drawbacks mentioned is likely to cause long-term problems, although some redefinition of standards may be required. The prospects for all three technologies look bright, with heavyweight support from the major manufacturers. But no-one can predict exactly how the wireless LAN market will split between HomeRF and HiperLAN/2 – and any other proprietary contenders.

Bluetooth introduces entirely new functionality. With the prospects of a mass market, the concept should take off rapidly if the analysts and forecasters are to be trusted.

Research firm International Data Corporation, also known as IDC, predicts a compound annual growth rate of over 80 per cent for Bluetooth shipments world-wide between 2000 and 2003.

Another consultancy, Visiongain, forecasts that by 2003 some 70 per cent of mobile phones sold in Europe and North America will have Bluetooth capability, with 400 million Bluetooth-enabled devices in service by 2005. In turn, Frost & Sullivan predicts "gargantuan growth" for Bluetooth-based devices and technology, forecasting sales of $36.7 million in 2000 and $699.2 million by 2006.

Where HiperLAN/2 and HomeRF are concerned, a lot will depend on the result of the cost-versus-capability equation. In performance terms, conventional structured cable systems may still have the edge but there will be many situations where users are prepared to compromise on data rates if offered the opportunity to cut the cord. Signs are that they will.

Statistics from Cahners In-Stat propose growth from $771 million in 1999 to nearly $2.2 billion in 2004. This reflects changing workgroup needs, increased employee mobility and the trend towards working at home or in converted small office environments where fixed cabling would be unacceptable or unfeasible.

Tumbling prices and growing functionality are another factor rendering WLAN solutions increasingly attractive to businesses of all sizes.

...with SoHo setting the pace

These last two factors will propel the small office and home – often shortened to SoHo – market into the lead according to well-informed observers. Forrester Research estimates the market for home and small-office networks will blossom to the tune of $1 billion in sales by 2003, noting that wireless systems and other technologies that have struggled for acceptance in large networks are welcomed by small users, who value flexibility and simplicity over sheer speed.

The watchword in the small-office and home market is 'no new wires': homeowners are reluctant to drill holes through walls and landlords probably forbid it. Bright-blue data network cable is easily concealed below the floors of modern office blocks but it's far less acceptable snaking around skirting boards and doorways.

Ericsson's Bluetooth evaluation kit.

Money talks

The final decision comes down to cash, however. Wireless connectivity will catch on only if the price is right and for this the cost of the necessary semiconductors will be the key determinant.

At last year's Wireless Symposium in San Jose, California, the manufacturer Harris Corporation claimed its Prism II chip set would bring down the materials cost of a 2.4GHz network card to $70, while Benno Ritter, communications product marketing manager with Philips Semiconductors, stated that to be successful, HomeRF radio ICs would need to be priced under $10.

With games manufacturers clamouring for a $3 'HomeRF-Lite' chip to incorporate in smart toys, the pressure is now definitely on to make wireless networking affordable. ■

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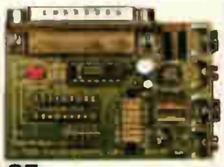

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Beginners' corner

Ian Hickman presents a further simple circuit for undergraduates – or anyone – to build, trouble shoot and test. In addition to being educational, the finished item is useful piece of test gear for measuring the small-signal AC current gain of n-p-n or p-n-p transistors.

This month's project depends on a useful and important technique, which may not have been specifically pointed out on many electronics-based courses. This is the separation of DC and AC characteristics.

The two can be separated not only conceptually, but also in practice – to the point where the circuit operation at DC is completely the reverse of the AC behaviour, as will appear in what follows.

The circuit

The diagram shows the complete circuit of the h_{fe} tester, built around a single op-amp. Like all practical electronic circuits, its design revolves around a number of approximations, which will be duly pointed out.

Originally, the circuit appeared in Chapter 11, 'Tricks of the trade, of Analog Electronics'[1], although it – and indeed the whole chapter – are omitted from the current edition.

In the original, I used a TL071 op-amp, although a TL081, described in last month's 'Beginners corner', would be equally suitable. The op-amp, a ceramic sounder, a couple of switches plus a few passive components are all that is required – apart of course from a couple of batteries to power the circuit, and a box to hold it all together.

How it works

The circuit implements two feedback loops, one positive and one negative. There is feedback from the output of the op-amp back to its non-inverting input, via the transistor under test. As the transistor is connected in the inverting common-emitter configuration, this actually constitutes a negative-feedback loop.

Resistor chain R_6, R_7, R_8 establishes a reference voltage at the op-amp's inverting input. This is pin 2, and is marked 'I'.

The effect of the negative feedback is to force just enough base current into the transistor under test to cause it to pass 1mA of collector current. This causes a 1V drop across R_2.

If the collector supply voltage were +10V – it isn't, but just imagine, to make the sums easier – then due to R_3 and R_4, the voltage at the op-amp's non-inverting input would be +4.5V, equal to the reference voltage at its inverting 'I' input: this is the first of many approximations involved in the design.

Ideally, R_8 pulls the reference voltage down from a value of one half of the supply voltage, by just the right amount to allow for the 1V drop across R_2. But resistor tolerances mean that the current forced through R_2 may have to depart slightly from 1mA to equalise the voltages at the op-amp's inverting and non-inverting inputs. In fact, those voltages will not be exactly equal, since a tiny difference is required to drive the op-amp's output to the appropriate voltage to close the loop.

A sea of approximations

Furthermore, this small differential input voltage may itself be small compared with the op-amp's input offset voltage – a sea of approximations, all of which are insignificant in practical engineering terms.

Note that the negative-feedback loop includes the whole open-loop gain of the op-amp, plus half – due to R_3 and R_4 – of the voltage gain of the transistor being tested. Loop stability is ensured by rolling off the loop gain at a low frequency, by the presence of C_2.

Output of op-amp will sit at the DC voltage needed to force the required base current into the transistor being tested, about +5.6V in the case of a device with a DC gain h_{FE} of 20 – less for higher gain devices.

Having dealt with the DC conditions, it is time to look at the AC conditions. There is feedback from the output of the op-amp, via the transistor being

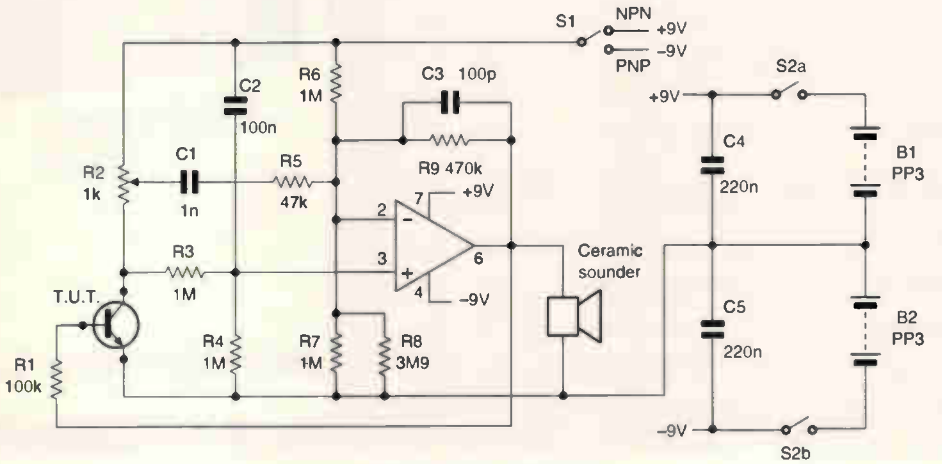
The benefits of double checking

Benefits of checking everything twice are many, including the avoidance of a red face – from which I am currently suffering. You may have noticed an error in Fig. 2 of last month's 'Beginners' corner', for which I humbly apologise.

The text correctly describes pin 2 as connected to pin 6, whereas the diagram shows it connected to pin 7 instead. I drew Fig. 2 last year and it was duly used at the time by the student members of Portsmouth University's RF Club. They all got the project up and running. So without thinking, I reused the drawing in 'Beginners' corner'.

Thinking back though, I now dimly remember having to write up a correction on the whiteboard: fortunately I had a copy of the LM555 data sheet with me.

The moral: double checking is always advisable – you can't expect the editor to act as longstop for your own gaffs!



This h_{fe} tester indicates the AC current gain of a transistor at the point where a tone is just audible.

tested, to the inverting input via C_1 and R_5. Around this loop, both the op-amp and the transistor being tested are inverting, so the feedback is in fact positive.

If C_1 were short circuit and C_2 open, this would upset the DC conditions. But imagine for the moment that it doesn't. Then the voltage gain from the wiper of R_2 to the op-amp output would be ×10. Now reinstate C_1 and consider the frequency fHz – at around 3.3kHz – where its reactance equals 47kΩ.

The combined impedance of C_1 and R_5 is now greater by a factor of $\sqrt{2}$, so the gain would now be ×7.07. Add C_3 back in, and at fHz, the feedback impedance falls by another factor of $\sqrt{2}$. At this point, the gain at fHz is just ×5.

Small changes

Now another approximation. Small changes in base current will produce negligible changes in the base-emitter voltage, or V_{be}, of the transistor being tested, so a voltage change δV of 1V at the op-amp's output will produce a change of base current δI_b of 10μA.

If the alternating current gain h_{fe} of the transistor being tested is 20, there will be a δV_c across R_2 of 200mV. Given the gain of the op-amp at fHz of ×5, this will produce a voltage change at the op-amp's output of 1V, if the wiper of R_2 is set fully clockwise, to the collector end of its travel. Thus the circuit will just oscillate – as will be apparent from the tone emitted by the ceramic sounder.

The sounder I used was salvaged from a defunct smoke alarm by the way. Scaling the value of C_1 and C_3, keeping the same ratio, will change the tone frequency, if the 3.2kHz given by the values shown is not optimum for the sounder used.

If the h_{fe} of the transistor under test is 40, then the circuit will just oscillate when the wiper is set half way, and so on. A pointer knob fitted to R_2 can thus indicate the tested transistor's h_{fe} directly, against a scale marked on the panel of the instrument.

Now make it...

The instrument can conveniently be built up in a small plastic instrument case. The front panel will carry the knob for R_2, with its associated scale, together with switches S_1 and S_2, and a TO5 or TO18 transistor socket. The sounder can be mounted on the front panel, or one of the sides, as most convenient.

Alternatively, with goods nowadays so commonly supplied in disposable plastic containers, one of these can be pressed into service. If a two-pole biased toggle switch or push button is used for S_2, there is no danger of the batteries becoming exhausted due to the instrument accidentally being left switched on.

Note that while the transistor being tested shown in the diagram is an n-p-n device, the instrument is equally at home measuring the h_{fe} of p-n-p devices, with S_1 appropriately set, of course.

...and get it working

Commissioning the circuit should not be difficult, though C_1 can be omitted initially. The first step is to calibrate R_2. This can be done with an ohm meter, with the test transistor absent and before fitting R_3.

The setting where the pointer indicates maximum resistance should be marked '20'. The 50% resistance setting corresponds to an h_{fe} of 40, 10% to 200 and so on.

With the circuitry completed, a known good n-p-n transistor fitted and S_1 in the n-p-n position, check that the op-amp output is between about +0.7V and +5.6V.

If all is well, C_1 can be added, and on adjusting R_2, the result should be silence fully anticlockwise, with tone starting to appear at some point as the setting is advanced.

The correct indication is the setting at which tone first starts to appear. This corresponds to the *small signal* current gain. At larger amplitudes, the transistor's AC gain falls slightly, so a further advance of R_2 is necessary if the amplitude is to increase.

Points worth noting

As mentioned earlier, this circuit – like any other – will be affected to some extent by component tolerances. Metal-oxide or metal-film resistors with 2% tolerance or better still 1% are preferable.

The exact values of the capacitors are unimportant, but the value of C_1 should be close to exactly 10 times that of C_3.

The tolerance on R_2 is likely to be worse than on the other resistors, but the circuit should still be accurate, if the value of R_1 is adjusted to be 100 times that of R_2. ■

Reference

1. Hickman, I. 'Analog Electronics, Heinemann-Newnes, 1st Ed. 1990.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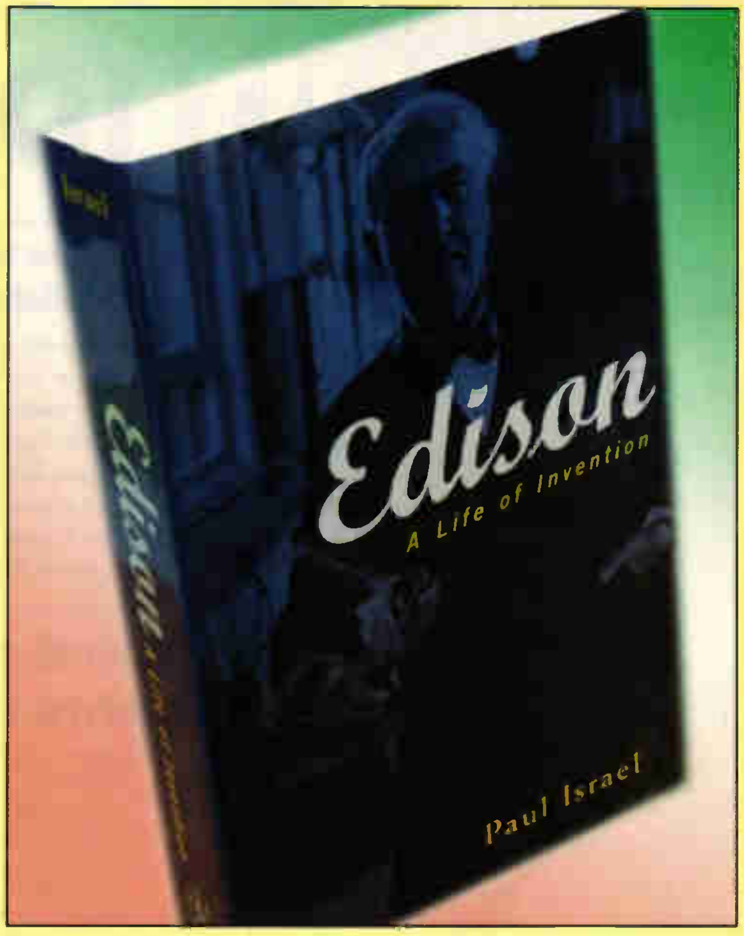

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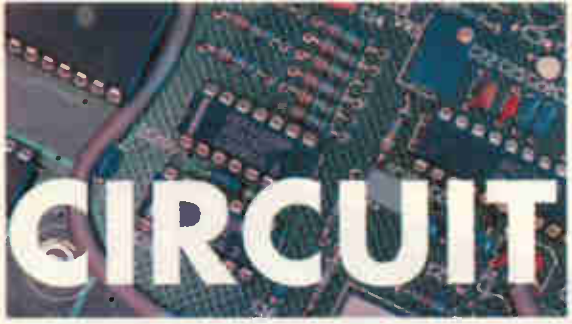


CIRCUIT IDEAS

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

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

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

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

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

Protection against turn-on voltage transients in 48V systems

In telecommunication systems, a battery with a nominal value of 48V is used to generate lower voltages such as the 5V, 12V and –12V required for electronic circuits in electronic switching systems using dc-to-dc converters.

The output from such a dc-to-dc converter generally has very short duration voltage spikes at the instant of turn-on. Filters cannot kill these spikes, which may cause sensitive electronic components to malfunction – or even fail. This circuit provides the necessary protection to sensitive electronic components.

The design consists of a soft-start circuit that allows the +5V DC voltage to be applied to the load circuit after a delay of about 20ms so that turn-on transients decay before the DC supply voltage reaches the load circuit.

You need to connect the protection circuit at the output of the +5V supply, where it is fed to the load circuits. The circuit shown is for +5V logic supply but similar circuits can be designed for + 12V, –12V, or any other voltage within the range.

The circuit consists of a *RC* network chosen to give a delay time of about 20ms, which is sufficient time to allow voltage transients to decay. The 470nF capacitor is charged through resistor *R* from the regulated 5V output from the dc-to-dc converter.

Initially, the capacitor is uncharged and the input of inverter 7404 is low, making its output high. It also makes the output of the transmitter of the inverting RS232 transceiver IC_2 low. This keeps diode *D* off and the MOSFET switch off too.

When the voltage across the capacitor reaches the high-level input threshold level of the inverter IC_1, a 7404 logic gate, the output of the inverter goes low. This makes the output of the RS232 transmitter high, turning diode *D* on and enabling conduction of the MOSFET.

The MAX233's high level output is about 9V, which is enough to turn the MOSFET fully on to conduct the load current. Diode *D* protects the MOSFET gate from the negative voltage output of –9V of MAX233 when a logic high appears at its input.

In this circuit, IC_2 is used as a voltage converter to step up logic output to about 9V required to drive the MOSFET into full conduction. Application of +5V to the logic circuits is thus delayed by about 20ms during turn-on, which allows time for the voltage level to settle down and transients to decay.

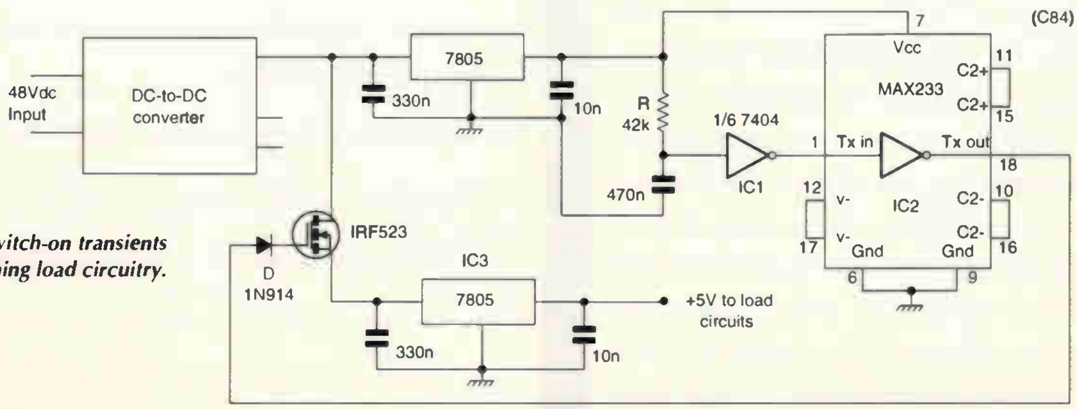
The MOSFET's drain connects to the unregulated voltage output of the dc-to-dc converter, from which the +5V is derived. Its source terminal of the connects to the input of the +5V voltage regulator.

Unregulated voltage appears at the input of the voltage regulator only when the MOSFET is on. The output of the regulator supplies +5V to the load circuits.

Using such an RS232 transceiver to drive the MOSFET avoids the need for a custom MOSFET driver or a step-up converter circuit to boost the +5V voltage to a higher voltage needed to turn the MOSFET fully on.

V Lakshminarayanan
Bangalore
India C84

The winner of the fourth National Instruments prize, to be selected from the July and August issues, will be announced next month.



Circuit to prevent switch-on transients reaching load circuitry.

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

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

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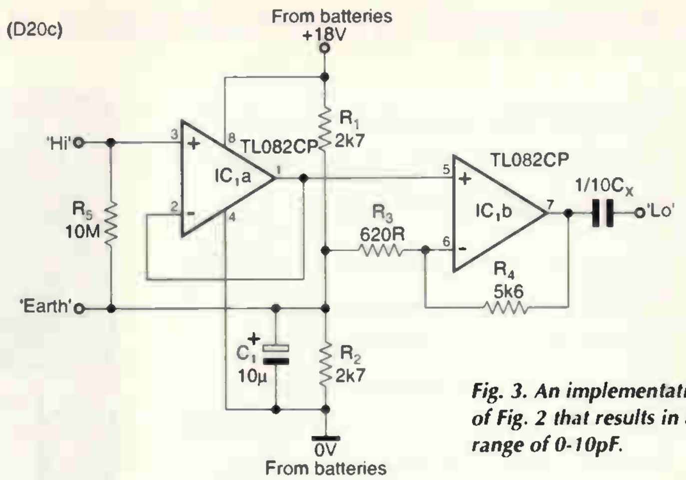

Fig. 3. An implementation of Fig. 2 that results in a range of 0-10pF.

Measuring femtofarads on traditional AC bridges

I needed to check the matching of two nominally 1pF capacitors, but all I had was a Marconi Instruments TF868B universal bridge having a lowest capacitance range of 0-100pF.

I could have made a transformer bridge, but this was not justifiable for the job in hand, given the extensive screening and constructional work required. The possibility of using the AC source and tuned detector in the TF868B was then considered; ideally without any alterations. Bridge resolution would have been adequate at 0.1 % at the top end of a range.

Figure 1 shows the basic bridge circuit for capacitance measurements; the earthing of one side of the detector allows three-terminal capacitance measurement to exclude strays. These are either across the detector or across C_2+R_1; so of insignificant effect at null.

If the voltage across C_2+R_1 were multiplied by a factor of, say 10, then a reduction in the unknown capacitor by a factor of 10 is needed to preserve bridge balance.

Amplification only need be applied to the voltage across the standard capacitor, since the magnified voltage is in phase with the original voltage across the capacitor, preserving phase balance to the detector. **Fig. 2**. No modifications are needed inside the TF868B.

Figure 3 shows the first circuit tried; to give a range of 0-10pF. The signal on 'Hi' never exceeded 1.2V pk-pk, so an op-amp with a gain of 10 could be used when driven from two 9V batteries in series; batteries were used to prevent earth loops and null offsets.

Op-amp IC_{1A} is a buffer – not strictly necessary – and R_5 protects the TL082 FET inputs against static build-up when the circuit is disconnected.

Components R_1, R_2 and C_1 gave a decoupled reference for the signal to swing about, R_4 and R_3 defining the gain. It worked as expected, giving a sharp null with a test 2.2pF capacitor at the 22pF dial reading.

Figure 4 showed the circuit tried to give 0-1pF range – with a resolution of 1fF at the top end; stray capacitances C_{S1} and C_{S2} are shown. I considered a transformer to be the easiest method for giving the voltage step-up needed. A tightly-coupled toroid transformer was found with the characteristics shown – and a self-resonant frequency well above the 1kHz of the bridge source.

Two op-amps – used as unity gain buffers – were connected in parallel to give the peak drive current of around 13mA. Individual and differential DC offsets, giving inter-amplifier and primary DC currents, were <1mV on the device used; should this be thought a problem capacitive coupling from the outputs to the transformer could be used at the risk of some phase shift; adding series resistance would also give significant phase shifts.

Supply current was around 11mA and the circuit worked well. Nulled 'tan-delta' setting was non-zero, due to phase shift from op-amp output impedance and primary inductance, but a sharp capacitive null was obtained.

Output voltage from the transformer was ≤120V pk-pk and at high impedance, but take care to avoid electric shock. Hand-capacitance effects were very significant and the high voltage output of the transformer secondary was placed at least 2 inches away from the 'Lo' terminal of the bridge. Even so, null readings of over 80fF were obtained, but use of a metallic screen between the high-voltage terminal and the 'Lo' terminal reduced this further. Low battery voltage is shown by a drifting null.

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Surbiton
Surrey

D20

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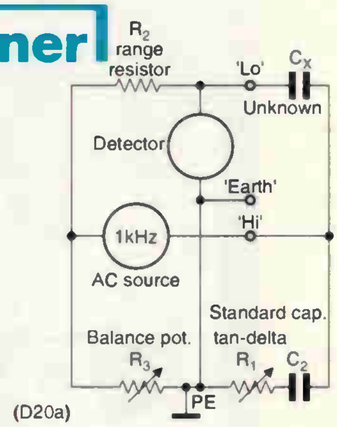


Fig. 1. Showing the basic bridge circuit for capacitance measurements.

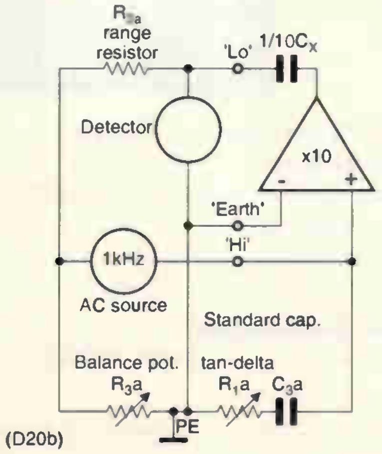


Fig. 2. Outline arrangement to extend the smallest range from 100pF maximum to 10pF maximum.

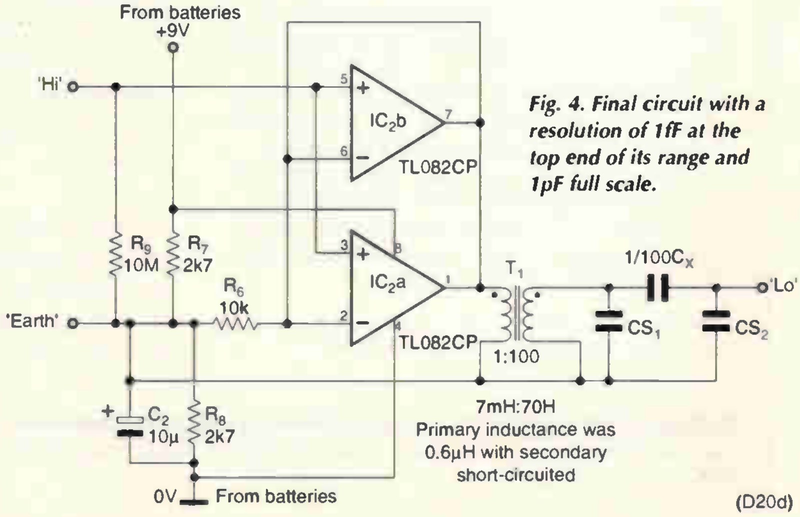


Fig. 4. Final circuit with a resolution of 1fF at the top end of its range and 1pF full scale.

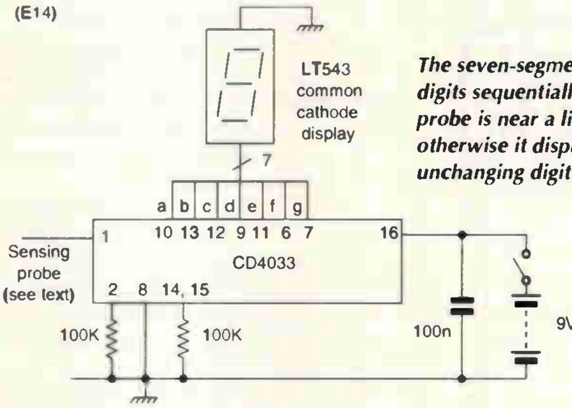
Live-line detector

This is a CMOS-based circuit, using a CD4033, that detects the presence of a live mains conductor. The count input of the CD4033 is used floating, connected to a 5 to 10cm length of insulated wire, which acts as the detector probe.

If the unit is brought close to a live conductor – insulated, and even buried in plaster – capacitive coupling between the live conductor and the probe clocks the counter, and causes the displayed digits to increment rapidly.

When remote from a live line, the unit stops advancing and displays an unchanging digit at random. The 'low' input, represented on the circuit by a chassis earth symbol, is adequately provided by the user grasping the case of the instrument in use.

Raj Gorkhali
E14

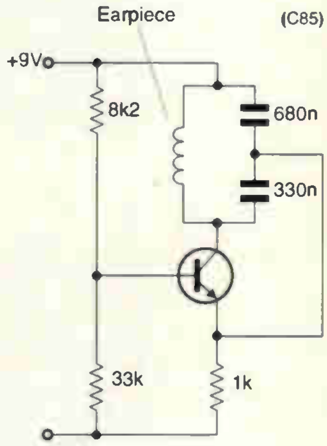


The seven-segment display flashes digits sequentially when the probe is near a live conductor, otherwise it displays a single unchanging digit.

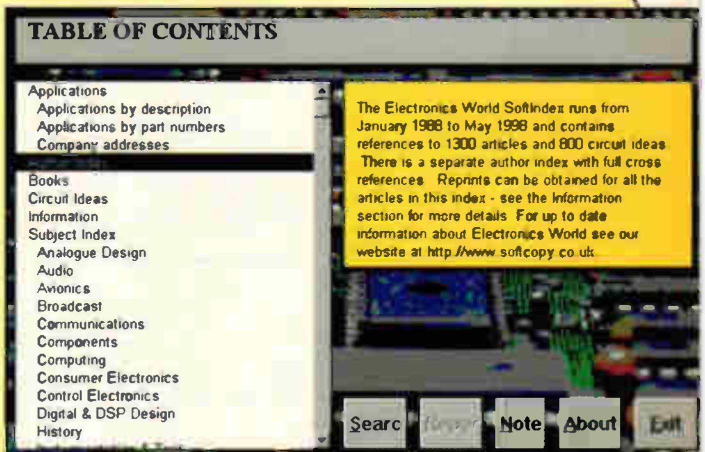
Sounder uses telephone earpiece

The tone sounder shown uses the inductance of an old telephone earpiece in the tank circuit of a Colpitts audio oscillator. The audio tone is radiated from the earpiece, and may be used as an audible alarm, etc.

D M Bridgen
Camberley
C85



An earpiece is used both as the sounder and the tank inductance in this audible tone generator.



74C926 leading-zero blanking

The 74C926 four-digit frequency counter is versatile, but lacks a leading-zero blanking facility.

Figure 1 shows the frequency counter driving common cathode LEDs. An additional logic circuit was included to suppress leading zeros, as shown in **Fig. 2**, using simple, cheap, readily available CMOS logic ICs.

The status of zeros and non-zeros from the detector output of the 74C926 were monitored by the CD 4068. Output at pin 13 of the latter is at logic 1 or 0 for non-zero and zero digits respectively. The CD4081 performs the AND operation between the output of the CD4068 and the 74C926's A, B, C and D multiplexing (MUX) outputs sequentially.

If the output of the CD4081 happens to be at zero for A, then the MSB will blank. The CD4013 D-type bistable device stores this event. On MUX output B, the CD4071 OR gate selects blanking or not blanking, depending on the pin 1 output of the CD4013.

Irrespective of the B output from the CD4081, the second digit is displayed if the output pin 1 output of the 4013 is at logic 1. The second digit is only blanked if the 4013 pin 1 output and CD4081 are both zero. Here, the CD4013 acts a simple set-reset bistable device.

The same logic is also applied to MUX output C. At the end of the cycle, i.e. after MUX output D, a narrow pulse of a few microseconds is generated by the CD4528. This pulse is applied to the reset pins of the CD4013.

It was necessary to use the CD4066 for the correct display of the digits. Leading zero suppression was not applied to MUX output D.

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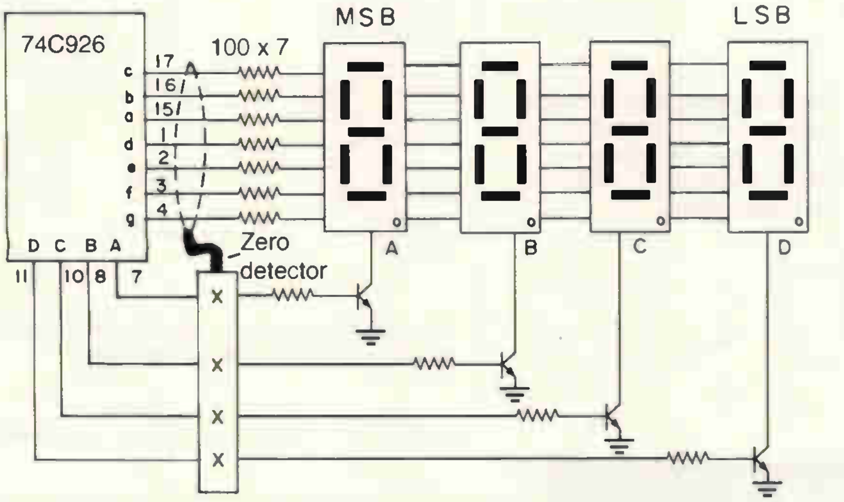


Fig. 1. In order to blank the leading zero of a four-digit frequency counter based on the 74C926, extra circuitry needs to be inserted to modify the drive signals.

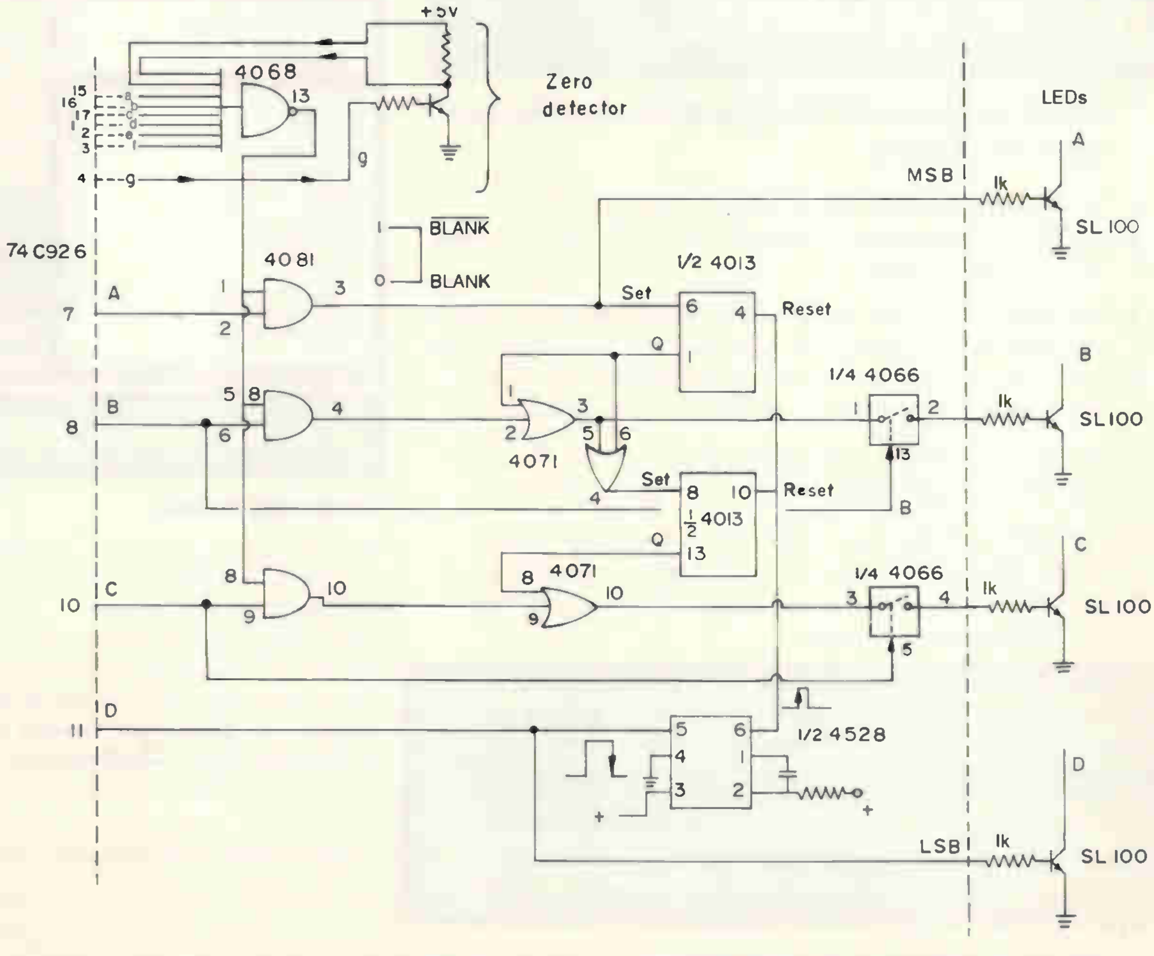
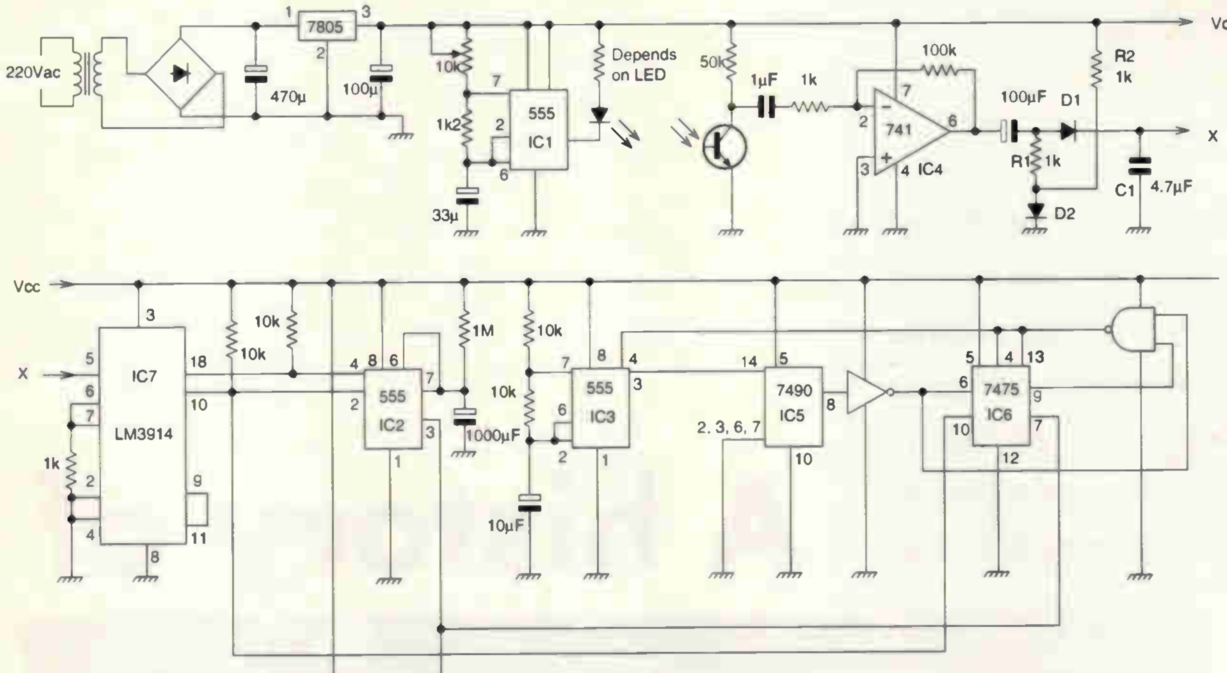
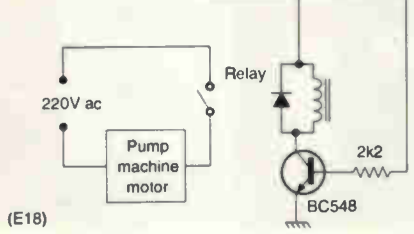


Fig. 2. Four-digit frequency counter with leading zero suppression, circuit diagram.





This circuit maintains the water level between fixed limits and safely auto-restarts after a mains outage.

Water pump control

We used an infra-red sensor to monitor water level to prevent wear and tear of mechanical parts, and to avoid electrolysis due to electrodes, which could affect the purity of the water in the tank.

Amplitude of the detected signal varies with the transceiver path length. The signal is amplified and peak detected by D_1 and C_1. Diode D_2, with R_1 and R_2, compensate for the forward diode drop of D_1.

The LM3914 dot/bar display driver, IC_7, decodes the amplitude of the signal into one of ten digital outputs. Pins 10 and 18 were chosen as the 'on' and 'off' points respectively, but in other installations, different settings may be required, as determined experimentally.

At the low level, pin 18 goes low, triggering 'on' the monostable, IC_2. At the full level, pin 10 goes low, switching off the monostable, which thus works like a bistable.

We added other circuitry due to frequent power cuts in Nigeria. Timer IC_3 and counter IC_5 form a delay circuit, which holds the pump off for about five seconds following restoration of power. This prevents transient voltage variations damaging the motor.

The delay circuit switches on the motor after the time-out period, whatever the water level in the tank, unless it is already full. Output from decade counter IC_5 passes through a transparent latch IC_6 and triggers the monostable.

The high level from the monostable feeds IC_6. From there, the signal goes to one input of the NAND gate, whose output goes low immediately after the next clock pulse into IC_5.

At this stage, IC_6 is latched and its outputs frozen. This disables the clock IC_3 and prevents the state of the NAND gate from changing. This condition persists until another mains outage occurs.

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Nigeria

E18

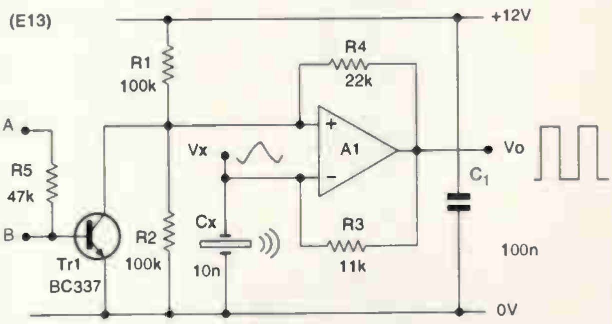
Ceramic sounder oscillator

The common ceramic sounder has a dielectric layer between the copper backing disc and the top metallisation. Its capacitance C_x is high – typically 10nF – and this property is used to advantage in the oscillator shown.

Op-amp A_1 acts as a triangle generator, with R_3 providing the path to charge and discharge C_x. Positive feedback via R_4 ensures that V_o is a clean square-wave. With the values shown, C_x displays a rounded tri-wave at f=2.5kHz and with peak-to-peak amplitude of 7.5V when the supply is 12V. This frequency can be altered by changing R_3, up to $f_{max}\approx$10kHz at resonance. Resistor R_3 at 7.5kΩ gives a nice 3.2kHz.

A simple but effective sounding-board for the transducer can be made by super-gluing it, off-centre, to a thin-walled plastic lid, say 8 by 5 by 1cm. Power consumption is very low, at less than 1mA at 12V. The circuit works down to about 3V, although the frequency falls, and is safe up to 30V.

By adding the n-p-n transistor Tr_1, the tone is easily modulated. A high signal on R_5 quells the oscillation, but a series of pulses can 'usurp' the tone. For example, point A was linked to +12V and point B connected to the 0.5V pk-pk 1kHz 'cal' signal from an oscilloscope.



The piezo ceramic sounder itself acts as the frequency-determining capacitor in this sounder circuit.

The result was a mixed 1kHz+2.5kHz sound.

For general enable/disable purposes, Tr_1 may be part of an open-collector TTL gate.

C J D Catto
Cambridge

E13

Patrick Mitchell looks at how the means of controlling a thermionic valve's emission have evolved and explains what the pros and cons of having multiple grids are.

A history of the grid

While the power efficiency of valves is largely determined by the performance of the cathode. other important properties depend on the nature and geometry of the valve's grid, or grids. These properties are gain, linearity, distortion, band width. and to some extent noise.

The origin of thermionic valves is usually traced to Edison's 1883 patent for what he called an 'electrical indicator'. This was the first device to use current flowing through the space between a filament and another electrode. Edison recognised that this current would flow in one direction only, but he put this rectifying effect to no use.

J. A. Fleming had been appointed as scientific advisor to the Edison Electric Light Company of London in 1882. He became involved in research into the blackening of light bulbs with vapourised filament carbon – the same area that had led Edison to his 1883 patent.

Some examples of Edison's 'indicator' bulbs were brought to Fleming in the course of this work. He noted their rectifying property, but had no immediate application for it. The bulbs were to remain forgotten for several years.

In 1899 Fleming was appointed technical advisor to Marconi's wireless company and began work on radio transmission and detection. At the time the weakest link in radio communication was the detector. The devices in use were the rather unsatisfactory coherers. as described in a separate panel entitled 'Coherers'.

Fleming realised that considerable improvements in detector performance could be achieved if a high-frequency rectifier could be devised. Edison's indicator bulbs were taken out of storage and tried as RF detectors. The experiments were successful. After making various changes to optimise RF rectification Fleming filed for patents in 1904 for his 'oscillation valve'.

Although in time thermionic devices would be used universally as detectors, they were not an immediate success. Other types of detector such as Marconi's magnetic detector patented in 1902 remained in use alongside the diode valve until the end of World War I. Rectifying RF signals was one part of the problem of detection. The very low signal power available from an antenna was the another. Truly satisfactory radio detection had to await the development of electronic amplification.

It was the development of a means of controlling the magnitude of the 'space current' that enabled valves to amplify.

This control mechanism was responsible for their transition from esoteric scientific instruments to valuable engineering components.

The well known grid was one of several means of achieving control that were experimented with. All methods shared the same principle: to control the field strength in the immediate vicinity of the valve cathode. If this could be achieved without introducing any additional currents then so much the better, but in reality it never was and never has been.

Lee de Forest is remembered as the man who put the grid into the diode valve. After receiving a PhD from Yale in 1899 on the reflection of electromagnetic waves, de Forest – like Fleming on the other side of the Atlantic – began work on radio. He had noticed that gas lights in his laboratory flickered in response to electromagnetic waves generated by sparks.

Having observed the flickering gas lights, de Forest spent several years experimenting with hybrid devices involving Bunsen burner flames and thermionic tubes in an attempt to find a better radio detector. This was to be a blind alley, but his involvement with research into detectors led him to acquire, and experiment with, several Fleming valves.

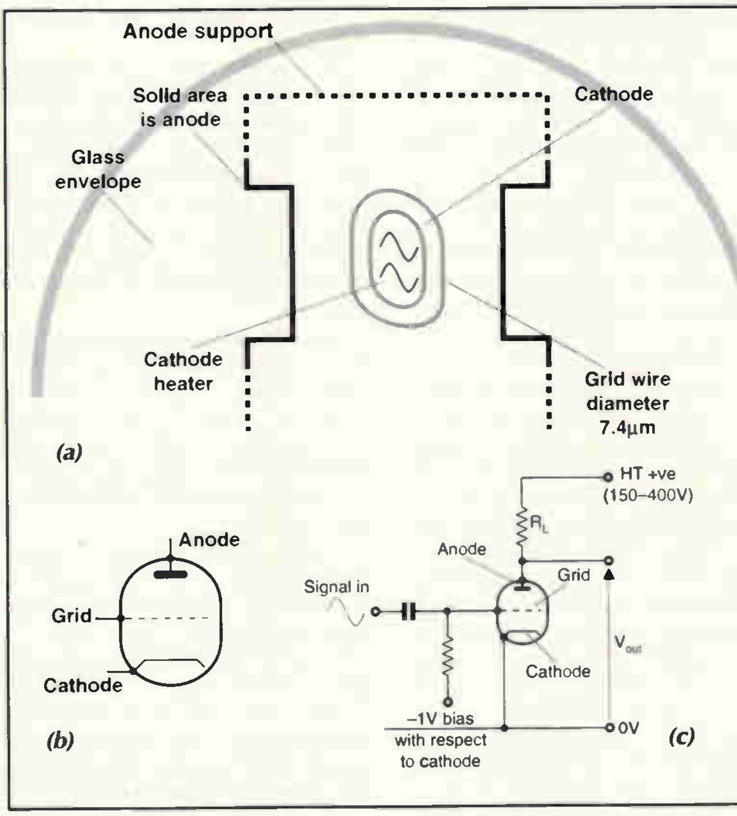


Fig. 1. The triode. Diagram a) is a cross section of the electrode assembly of a small-signal triode. Diagram b) is a circuit symbol commonly used for the triode and in c), a triode wired as a basic common cathode amplifier.

Early grids

In connection with this work de Forest had a New York lamp manufacturer called McCandless make copies of Fleming valves. Some had extra electrodes of various configurations.

Several US patents followed from this work, one of which, in October 1906, was for his first three electrode 'Audion'. This involved a type of Fleming diode with parallel plates on either side of the filament, connected together as the anode to maximise 'space-current capture'.

Wires to these two plates were brought out separately. One was connected to an antenna and the other to a detector. De Forest claimed that this arrangement was capable of amplification, which is doubtful, but he recognised that the current to one plate could be controlled by the voltage on the other.

De Forest reasoned that the performance of the device might be improved if the third electrode was placed between the filament and the anode. He realised that if a solid sheet were used it would largely block the current flow to the anode, so instead he specified a piece of wire bent in the shape of a grid iron.

The name grid stuck. He ordered some bulbs to this specification from McCandless which were delivered on 25 November 1906. At the time, de Forest was beset with legal and financial troubles stemming from patent law. Because of this, he did not test the new bulbs until 31 December. The tests showed that definite amplification was possible.

A patent followed on 29 January, 1907. De Forest's 'Audion' was not, however, the first amplifying valve. European workers had been experimenting with the cathode-ray tube, described in a separate panel entitled 'Cathode-ray amplifiers'.

Temperamental triodes

Early triodes were temperamental, varied widely in characteristics from one example to another and had limited life. These problems were investigated and gradually solved by engineers and manufacturers.

The main difficulties were the generation and maintenance of a high vacuum, and the production of durable filament cathodes. Being a fairly simple structure operating at lower temperatures than the cathode, the grid posed fewer problems and so changed little in the early years.

Mechanical strength, ease of manufacture and reproducibility led to the

Coherers

Coherers are temperamental electromechanical devices that work by arranging for a conduction path to be established by an RF signal which is then used to conduct DC.

Marconi's version consisted of fine nickel and silver filings contained in a partial vacuum between two electrodes. The resistance between the electrodes is largely determined by the number and nature of inter-filing contact points and drops markedly if an RF signal is applied between the ends.

The coherer does not reset itself and needs to be struck to shake the filings loose again. A relay could be connected and would close in the presence of a signal allowing for the reception of telegraphy.

Others coherers were used for telephony. One example is the Lodge Muirhead type which consisted of a bowl of mercury covered in a thin film of oil. A rotating sharp-edged iron disc was lowered into the mercury to almost – but not quite – break the oil layer. The RF signal would break the layer and allow conduction. Unlike the filings-based coherer, this device was self resetting.

abandonment of the unsupported bent wire arrangement and the adoption of ladder type grids, but it was not until other problems had been ameliorated – if not solved – that the electrical properties of the grid began to limit valve performance.

In the years leading up to World War 1, the development of the valve became mired in patent law. This was responsible for stagnation of the technology and financial troubles for the innovators and producers. As usual, the only winners were the lawyers.

Patents were suspended to allow development of military communications equipment to go ahead unhindered during the war. But the problem recurred thereafter and the need to avoid patent infringement prompted many novel arrangements – though few technical improvements.

The development of the grid and its geometry was dictated by three principle technical limitations: secondary emissions, microphony, and capacitive feedback – the Miller effect. Once the problems of consistency, reliability and life span had been addressed, the impetus was to increase the gain and raise the maximum operating frequency of valves.

General Electric appointed Dr Irving Langmuir as an engineer in 1909. He was to prove a valuable asset. GE had developed a high power 200kHz alternator for radio transmission but had no means of modulating its output. Langmuir thought that the triode might be adapted to make such a modulator and began a painstaking study of its electrical characteristics.

Several key improvements were to emerge from his work. An increase in gain by moving the grid closer to the cathode was the first improvement. Cathode current depends on the field strength immediately adjacent to the cathode. The closer the grid is to the cathode the lower the grid voltages needed to produce the necessary fields.

This modification had the happy side effect of increasing the distance between the anode and grid and hence reducing the anode-to-grid capacitance, which when amplified by the stage gain appears as Miller capacitance. The down side was increased grid heating and contamination with cathode atoms, leading to thermionic emission.

Suppressing emission current was to be a central theme in grid development. This is the exact opposite of what is required for cathodes. Incidentally, Langmuir was also responsible for noticing the potential of, and perfecting, the thoriated tungsten cathode.

Several factors conspire to cause grid emissions. The control grid is located near to the hot cathode to improve gain. This means that it is inclined to absorb vapourised constituents of the cathode – particularly as they are liable to be positively charged heavy ions and the grid is usually more negative than the cathode. Deposits of Ba, BaO_2, and Th on grids caused by cathode evaporation increase their thermionic emissions.

Another effect of the close proximity to the cathode is that the grid gets hot.

Unwanted emissions

Secondary emission is the emission of electrons from a surface induced by high energy electrons striking it. The secondarily emitted current may exceed the incident current – an effect that is exploited in photomultipliers.

Electrons emitted from the cathode have a range of energies, a few of them having enough to cause secondary emission when they strike the grid. The bombardment of the grid by high energy electrons also heats it, exacerbating thermionic emission.

As the grid is usually connected in a high impedance circuit, a small current of electrons away from it can cause a significant potential rise, attracting more electron bombardment and further heating and thermionic emission leading to thermal runaway.

Suppression of grid emissions depends on choice of surface coating, geometry and voltage. It is eased by increasing the grid-to-cathode spacing and by using finer grid wires. Holding the grid at a negative potential relative to the cathode also helps. In most valves, grid current approximates to zero when the grid voltage is more than 1V below cathode voltage.

Grid coatings

Numerous coatings were tried to increase the grid's effectiveness and defeat the thermionic effect of condensed cathode atoms. Graphite was one of the first coatings to be used. Although effective, it is fragile and deteriorates with time.

In 1930, it was discovered that gold plating was particularly effective at suppressing grid emission in valves using an oxide coated cathode. This came to be the commonest method used. A layer 1μm thick is needed for effective suppression.

Gold coating is very effective at stopping thermionic emission but places a considerable constraint on grid temperature as gold evaporation from the grid will poison the cathode. Thoriated tungsten cathodes operate and higher temperatures than oxide coated so platinum – also an effective emission suppressor but without the tendency to evaporate and poison the cathode – is used.

Grid geometry

For the grid's field to be fairly uniform at the cathode surface, the grid wires should be numerous and fairly distant from the cathode. Considerations of gain and emissions dictate otherwise.

In the event, many small-signal valves, such as the triode Western Electric 417A in **Fig. 1**, use uneven grid fields at the cathode. In this case the wires of the grid are 7.4μm in diameter, pitched at 65μm and placed 45μm from the cathode.

Valves of this type do not have a clean cut-off grid voltage as areas of the cathode close to a grid wire are cut off at smaller grid voltages than areas between wires. This led to a solution of the problem of automatic gain control. By altering the bias on the grid, a greater or lesser proportion of the

Cathode-ray amplifiers

In 1905, von Lieben constructed a cathode-ray tube with an electrode as the target. The cathode ray was steered by a magnetic field induced by coils.

A greater or lesser amount of the beam struck the target and thus the target current could be varied by changing the coil current. Based on his 1906 patent, the two German companies AEG and Siemens founded Telefunken AG to build Electron Tubes.

On 10 October 1906, Dieckmann and Clag of Strasbourg filed a patent for a similar device. In this alternative, variations in an electric field deflected the cathode ray such that a greater or lesser amount of it fell on a target electrode. Amplification was thus achieved by altering the proportion of cathode current striking an anode. This contrasted with the triode, in which amplification was achieved by varying the magnitude of the cathode current.

cathode surface can be 'cut off' thus altering the gain. Beginning in the mid 1920s, this effect was augmented in some types intended for use as automatic gain-control components in radio receivers. They had grids with wires more widely spaced in the centre of the grid than at the ends and were known as 'vari-mu' valves.

Microphony in valve amplifiers is the appearance of a component of noise produced by mechanical vibration. The main cause for it is vibration of the grid wires. Microphony was of little relevance in the early days of valve production. As demand increased for radio equipment that would perform satisfactorily in aircraft in the late 1920s though, it became an important issue and research intensified.

The requirements of the grid structure preclude any effective damping scheme, but as most mechanical vibrations are low in frequency, the grid is constructed with a high resonant frequency. This is done by placing the grid wires under tension. They are generally made of tungsten, which has a high tensile strength. They are pulled to about half the breaking strain of tungsten and brazed with gold onto a molybdenum frame. The whole is then coated with gold.

The tetrode

Capacitance between the grid and the anode is the main factor limiting the operating frequency range of triode valves. Modifications to improve this were to increase the grid-to-anode spacing and to use fewer, narrower grid wires, but the effects of this strategy were limited.

In 1915 Langmuir at General Electric, and simultaneously Schottky at Siemens, addressed the problem by placing a 'screen grid' between the grid – now 'control grid' – and the anode, nearer to the control grid. The new valve was called a tetrode, **Fig. 2**.

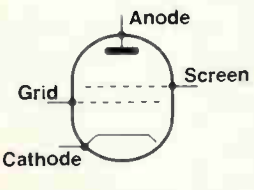


Fig. 2. Circuit symbol of a tetrode.

A screen grid is connected to a low-impedance circuit so its voltage varies little with variations in the anode voltage. The control grid is effectively screened from the anode. The screen grid's DC voltage must be higher than that of the control grid and cathode, or

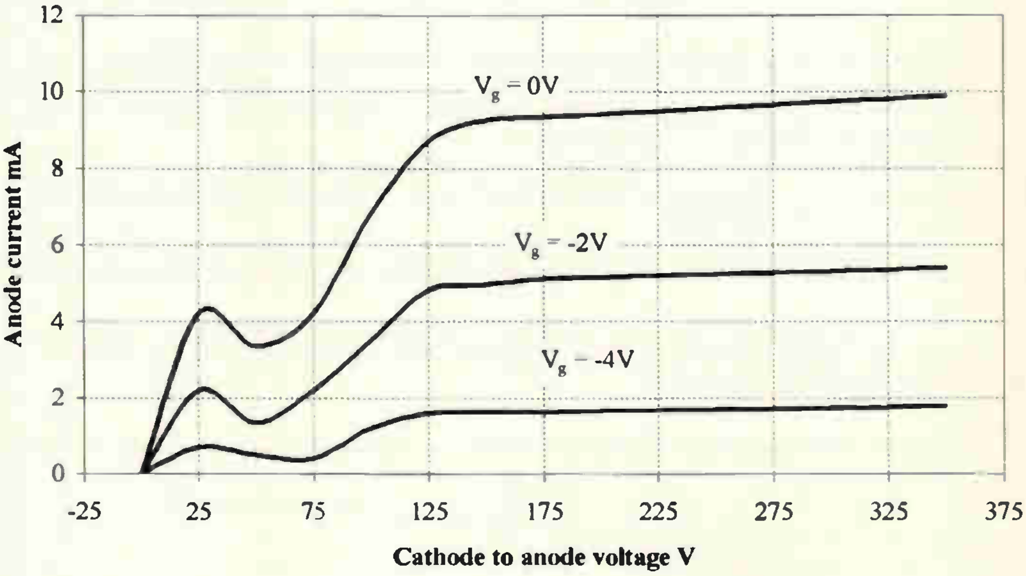


Fig. 3. V-I characteristic curves for a small-signal tetrode. The curves represent the relationship between anode current and anode voltage for various fixed control grid to cathode voltages (V_g). The kink between 25 and 75V on the anode is due to the exchange of secondarily emitted electrons between the anode and screen. Notice the negative impedance between 25 and 50V.

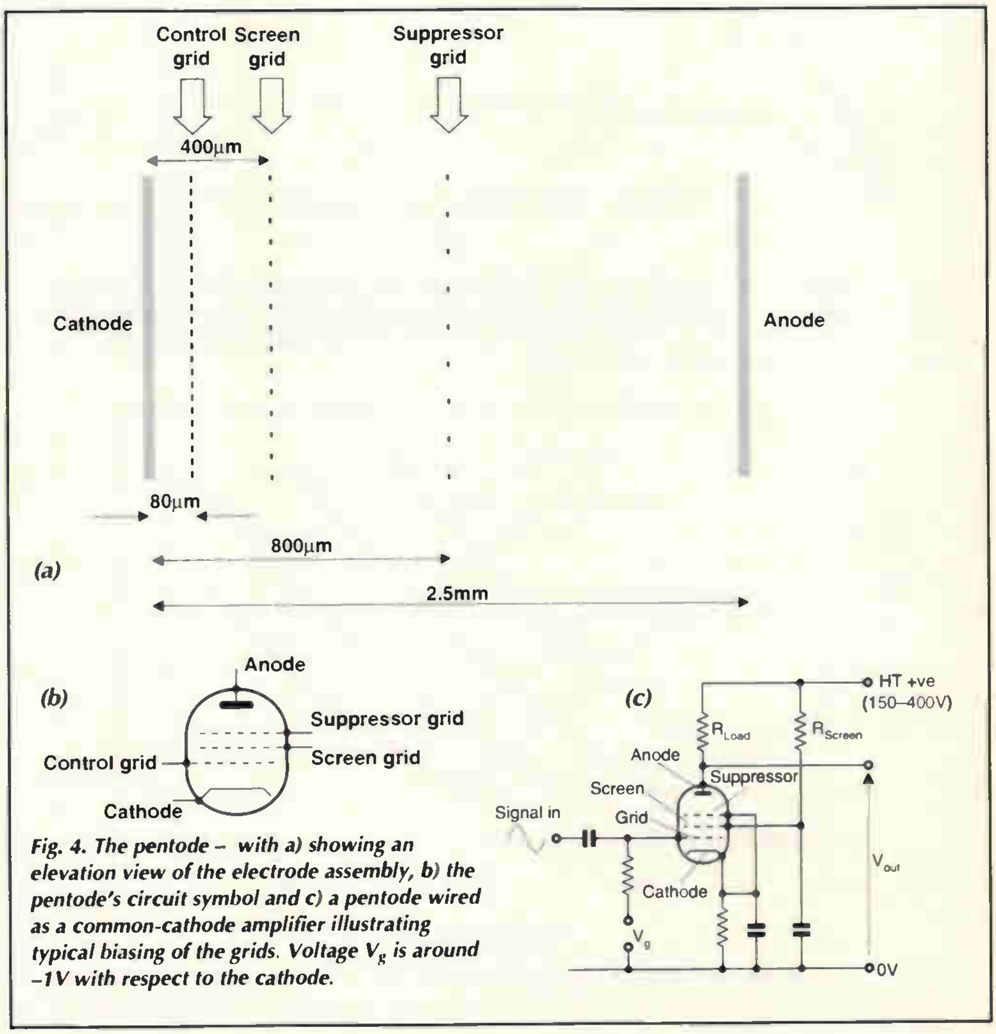


Fig. 4. The pentode – with a) showing an elevation view of the electrode assembly, b) the pentode's circuit symbol and c) a pentode wired as a common-cathode amplifier illustrating typical biasing of the grids. Voltage V_g is around –1V with respect to the cathode.

Characteristics and distortion

Graphs of the anode current versus anode voltage for various grid voltages for a small signal pentode are shown in **Fig. A**. These 'characteristics' are very similar to those of bipolar transistors and approximate to $I_a \propto (1-e^{kV_a})$. This leads to distortion by the introduction of odd harmonics spanning from low to high order.

On the other hand, the triode's anode *IV* characteristics, **Fig. B**, approximate to $I_a \propto V_a^{3/2}$. This produces more distortion but it is mainly by introducing lower harmonics, principally 2 and 3 times the fundamental.

In musical terms, these notes are the octave and dominant (fifth) notes of the fundamental key – the note being distorted. These notes are parts of the basic chord and sound musically harmonious.

Triode distortion sounds as though it belongs to the music and is difficult for the ear to detect. Even when it can be heard, it is not irksome and makes the sound 'richer'. Many audiophiles see this as desirable.

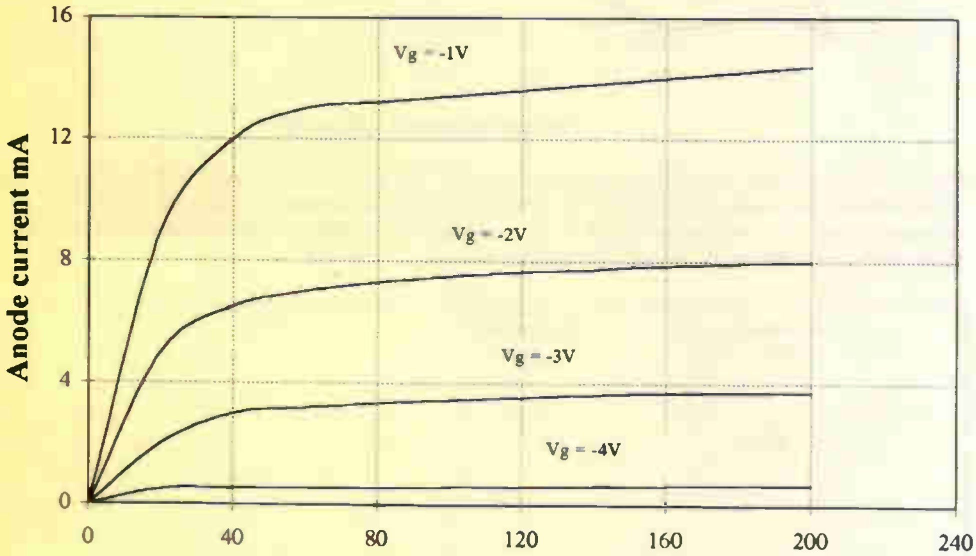


Fig. A. V-I characteristic curves for a small-signal pentode. The curves represent the relationship between anode current and anode voltage for various fixed control grid to cathode voltages (V_g). The curves have similar shape to those of bipolar transistors and consequently pentodes have similar distortion characteristics to transistors.

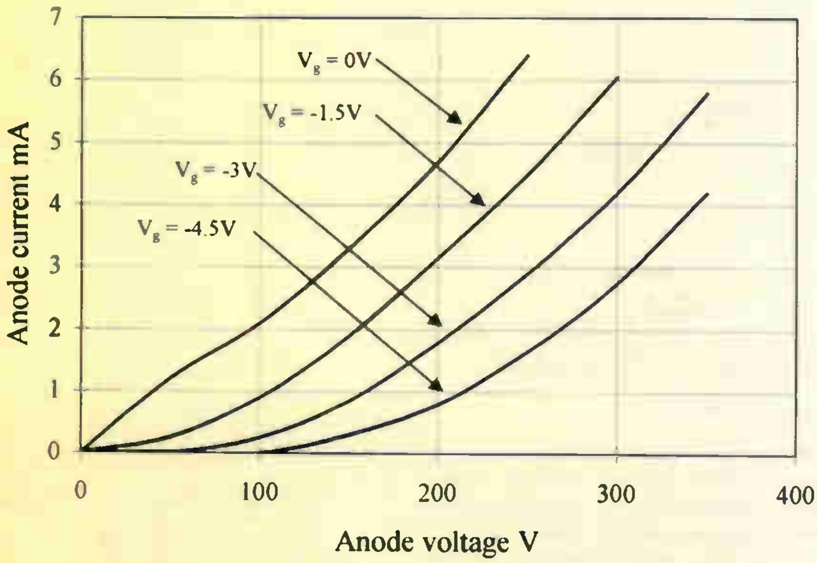


Fig. B. V-I characteristic curves for a small signal triode. The curves represent the relationship between anode current and anode voltage for various fixed control grid to cathode voltages, V_g.

electrons will not pass it. In practice it is held at or near V_{aa}.

The frequency range of the tetrode was a substantial improvement over that of the triode but at a cost: distortion. **Figure 3** shows a family of characteristic curves of a small signal tetrode.

Secondary emissions from both the anode and the screen are the problem. As a first approximation, the cathode current is shared between the screen grid and the anode in proportion to the relative areas they present to the oncoming space charge.

In practice, both the screen and the anode display a considerable degree of secondary emission due to high-energy electrons hitting them. Thus if the screen is more positive than the anode, current will flow from the anode to the screen and vice versa. This effect spoils the linearity of the relationship between anode current and control grid voltage by adding an undesirable dependence of anode current on anode voltage, seen as a kink in the *I/V* characteristic.

This nasty *I/V* characteristic was improved by increasing the grid-to-anode spacing and by coating the anode with an emission suppressing coating such as graphite. The basic problem remained however, and tetrodes – with the exception of beam power tetrodes – largely fell out of use as audio amplifier components by the 1940s.

Pentodes

Two Dutch engineers working for Philips, Gilles Holst and Bernard Tellegen, solved the problem of the tetrode by placing another grid between the screen grid and the anode. The result was the pentode and the new grid came to be called the suppressor grid, **Fig. 4**.

Philips was granted Dutch patents on the pentode in 1926 and patents around the world soon after that. This was one of the key developments that led to Philips establishing the dominant position it still enjoys today.

The suppressor grid is usually held at cathode potential. It thus repels electrons approaching it from either side and suppresses the exchange of secondarily emitted electrons between the anode and the screen grid. It does not intercept electrons passing from the cathode to the anode.

A suppressor grid has widely-spaced wires to minimise its influence on primary anode current, yet it still suppresses virtually all exchange of secondary electronic emission. The pentode offers several other advantages as well: effective control grid-anode screening, high anode dynamic impedance and high gain per stage.

The pentode became the most popular valve type by far.

Beam power tetrodes

The beam power tetrode was developed after the pentode to avoid infringing the pentode patents. It overcomes the problem of secondary emission currents by exploiting 'space charge' geometry. The control and screen grids are aligned so that electrons not impacting on the control grid also bypass the screen wires.

Two beam-forming electrodes, as shown in **Fig. 5**, further control the space charge. These electrodes are connected together and form a fifth external electrode connection. This is why the valve is often called a beam power pentode.

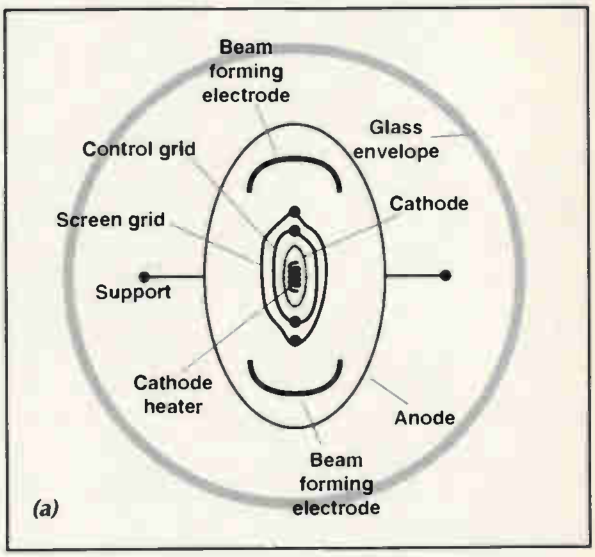
The space charge is channelled between the grid wires and beam forming electrodes. In the area beyond the grids, the charge spreads out to reoccupy the gaps left by the grid wires.

At any point in space within the valve, the potential depends on the voltages on the various electrodes and the space-charge distribution. The upshot is that the space-charge distribution is arranged to form a potential minimum down stream of the screen. This potential minimum acts as a virtual suppressor grid and repels electrons from both the screen and the anode.

The scheme works best when substantial space currents are present. This is why it is mainly used in high power valves – hence the term beam *power* valves. Beam power valves have very similar characteristics to pentodes, though their biasing is different so they cannot be used as plug-in replacements.

If a large number of electrons are emitted locally then the depth of the minimum is increased and fewer electrons have the energy to pass it. This smoothes out the random fluctuations in cathode current and can reduce noise power by as much as fifty times.

Space charge smoothing has little effect on the random distribution of electron velocity components parallel to the cathode. Because of the finite thickness of the screen and suppressor-grid wires, the amount of current collected by them depends on the transverse velocity of passing electrons, so a random noise is reintroduced to the space charge smoothed current. This is known as partition noise.



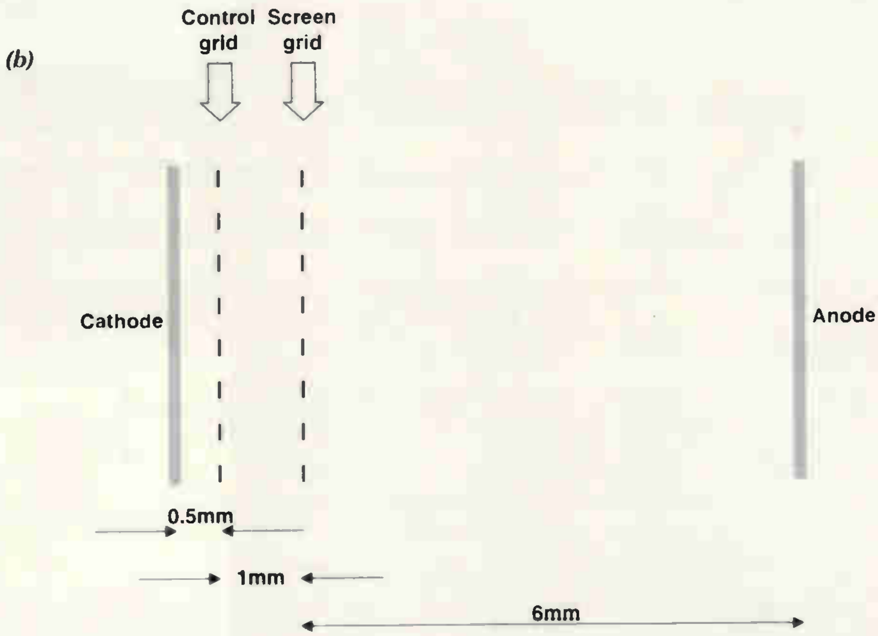


Fig. 5. The beam power tetrode. Diagram a) is cross section of the electrode assembly while b) is an elevation view of the electrodes illustrating the alignment of the grid wires.

Semiconductors and the return of the triode

Triodes are less noisy than tetrodes or pentodes. Most of the noise in tetrodes and pentodes is partition noise.

To explain partition noise it is necessary first to digress to describe shot noise in valves. Electrons are emitted from the surface of a hot cathode according to a Poisson type process. The process may be regarded as purely random and thus has a noise component $i^2=2e\Delta f$. This is a close approximation to the noise observed when the cathode current is saturated.

At currents below saturation, an effect called 'space-charge smoothing' markedly reduces this noise. In essence, what happens is that very close to the cathode there is a cloud of emitted electrons producing a potential minimum. Not all electrons will pass it.

Tetrodes and pentodes are affected by partition noise – but not triodes because they are usually operated with virtually zero grid current. Consequently, pentodes produce around ten times the noise power relative to triodes.

For audio applications, triodes also have better anode *I/V* characteristics than pentodes. There's more on this in a separate panel entitled 'Characteristics and distortion'.

Pentodes are stable, predictable, have a wide frequency range, and high output impedance. For these reasons they came to dominate the mass electronics market. Semiconductors though did these things better and have entirely replaced valves in that market, save for CRTs. But the particular properties of the triode have ensured it a niche in the growing high-quality audio sector.

Finally, I am indebted to Professor G. M. Sacchetti of Unison Research, Italy – makers of valve audio equipment – for his advice in the preparation of this article. ■

Further reading

Saga of the vacuum tube, Gerald F. J. Tyne. ISBN 0-672-21470-9.

70 Years of Radio Tubes and Valves, John W. Stokes. ISBN 1-886606-11-0.

Principles of electron tubes, J. W. Gewartowski and H. A. Watson, 1965, Van Nostrand, Princeton, New Jersey

Yet more grids

Valves have been produced with four or more grids. The extra grids are usually extra control grids to allow mixing of several input signals.

NEW PRODUCTS

Please quote *Electronics World* when seeking further information

RF chip set

Agilent Technologies has announced an RF chip set for CDMA 800 and 1900MHz handsets. Called CDMAdvantage, it will include all active RF functions required in a CDMA handset. The receive side has a switched LNA that can be used in an adaptive mode to reduce current draw. The combination of the MGA-72543 bypassed-switched LNA with the HPMX-7102 dual-band, dual-mode down converter can provide –106.5dBm sensitivity at less than 1 per cent frame error rate. Also part of the chipset are three power amplifier modules – the QCPM-9801 dual-band, dual-mode amplifier, QCPM-9804 dual-band, tri-mode amplifier and QCPM-9805 single-band, dual-mode module. The QCPM-9804 has the functionality of three amplifiers (Amps, CDMA800 and CDMA1900) in one 11.5 by 11.5mm package.

Agilent Technologies
Tel: 00 49 6441 9246 46

Dual-redundant fibre channel PMC

Radstone has introduced the FA2 dual-redundant fibre channel PCI mezzanine card adapter. It uses fibre channel technology developed by Fairchild for the B-1B stealth bomber. The dual redundancy provides system level fault tolerance by creating parallel paths through the transceivers and cable plant. This feature can automatically switch to the alternative, available channel if there is a failure. It provides dual-attach fibre channel fabric connectivity in air-cooled and conduction-cooled environments, and is capable of fibre channel data transfer rates up to 200Mbyte/s. It supports class two and three communications over all fibre channel topologies. Orbital's fibre channel technology uses active transceivers that have passed military EMC tests while transmitting at 1GHz over 21m of cable with standard aerospace connectors in line.

Radstone Technology
Tel: 01327 359444

Dual codec chip

From DT Electronics is the MSM7731 dual codec chip that integrates acoustic and line-side analogue interfaces with echo and noise cancellation. For hands-free communications systems, it is equipped with gain and mute controls for data transmission and reception, a μ-law PCM 16-bit linear digital interface for memo recording and message output, and a transfer clock generator for digital communications. An integral analogue input amplifier has a maximum gain of 30dB and the analogue output is a push-pull configuration to drive a 1.2kΩ load. There is a built-in transmit slope filter. The chip operates from a 3V power supply. Echo attenuation is typically 35dB for white noise and signal noise attenuation 17dB for white noise or 40dB for single tone. The built-in codec provides synchronous transmission and reception, permitting full duplex operation. Typical operating current is 35mA at a V_{DD} of 3.0V falling to 0.02mA in standby mode. The device can be controlled via serial or parallel ports.

DT Electronics
Tel: 024 7643 7437

Tactile switches

Thomas & Betts has broadened its Alcoswitch range with three tactile switches – the FSM CT, CTA and CTL. They can be positioned 5 to 6mm beneath switch foils or flat keypads. With a height of 1.5mm, the switches are for use under thin foils on the front panels of electrical and electronic equipment. They are aimed at the white goods market including supermarket cash registers and washing machine controls. All three operate from –25 to +70°C. The CTA and CTL are quoted at 100 000 cycles and the CT at 500 000 cycles. There are three

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operating strengths, that is the degree of force to cause actuation, of 120, 180 and 260gf. Surface area is 4.9 by 4.9mm.
Thomas & Betts
Tel: 00 32 2 359 8300

Scope weighs 3kg

Lecroy has developed the Literunner LP142 digital oscilloscope that weighs about 3kg. There is a built-in floppy drive, graphics printer and internal five-digit hardware frequency counter. It operates at 500MS/s and 100MHz bandwidth. Memory is 100kbyte/ch and it has three-way data storage capability. Horizontal accuracy is 50ppm. Basic functions required for common measurements are provided automatically. For example, it can automatically measure the time differential between signals. The oscilloscope automatically sets up vertical and horizontal range and triggers for measuring repetitive signals. A function knob is used to select a menu item and a press of a button to set it.
Lecroy
Tel: 01235 536973

Thermistors

Murata has produced the NTH5G10P chip NTC thermistors for temperature compensation in portable electronic equipment, including mobile phones, PDAs and audio equipment, and contrast compensa-

tion for LCDs. The 0402, 0603 and 0805 sized products measure from 1.0 by 0.5mm and have been made using multilayer forming technology. The surface mount 0402 thermistors are suitable for reflow soldering and the 0603 and 0805 can be flow and reflow soldered. Also available are the NTH5G16P and NTH5G20P measuring 1.6 by 0.8mm and 2.0 by 1.25mm respectively.
Murata Electronics
Tel: 01252 811666

Octal Ethernet switch

An integrated single-chip octal switch for 10 and 100baseTX desktop applications, the AC108SU from Solid State Supplies is based on the Nucleus switching architecture. The single chip, two layer octal switch can cascade up to 24 ports at full wire speed. Internal RAM supports 4k Mac addresses, and the 4.8Gbit Nucleus backplane enables full wire speed in eight, 16 and 24-port configurations. One SGRAM or SDRAM chip provides 4Mbyte of external packet memory. Each port uses physical layer mixed-signal core architecture and an 802.3 compliant Mac that provides half and full duplex with flow control. The switching fabric provides full wire speed store and forward with auto learning and ageing. Features include broadcast storm and head-of-line. It has programmability using an

EPROM or an MDIO interface for register and PHY access.
Solid State Supplies
Tel: 01892 836836

Digital video troubleshooter

Tektronix has announced a module for its TDS3000 family of digital phosphor oscilloscopes (DPO). The TDS3SDI module enables broadcast engineers and technicians to view ITU-R

BT.601 digital video waveforms, allowing them to verify source, timing and amplitude. The module operates in TDS3000 series DPOs, and so is intended to provide digital video engineers with a single instrument for qualitative troubleshooting of digital video signals.
Tektronic
Tel: 01628 403446

OCXOs for pico base stations

A range of miniature OCXOs (oven controlled crystal oscillators) for use in mobile phone

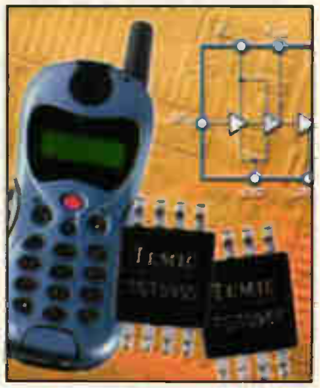
PC functions in a 500MHz DSO

Gould has combined the acquisition power of its 500MHz, 4-channel digital storage oscilloscope (DSO) with the features of a PC. The Ultima's 500MHz bandwidth and up to 2Gsample/s sample rate with 1Mbyte memory per channel is supported with a package of trigger tools to qualify the incoming signal for amplitude, time, pulse, gating and TV. A wide dynamic range in transient mode, with 12-bit resolution up to 2Msample/s, is also included. Featuring a Pentium class computer, and large 10.4in SVGA (800 x 640) colour display, the unit's internal configuration is such that direct data interchange is made to Microsoft Office applications, as well as to other third party PC-based analysis and report generation packages. There is also an internal 8.4Gbyte hard disk and removable 1.44 or 120Mbyte floppy alternatives. With all controls accessible using traditional front panel controls or through mouse and keyboard, the instrument provides a range of real-time analysis capabilities including FFT, differentiation, integration, phase correction and math functions. Up to eight display windows can be viewed simultaneously enabling main and zoom traces or user defined math traces to be viewed independently and scaled to user defined units.
Gould Nicolet
Tel: 020 8501 6604

base stations and Stratum 3 SDH/SONET switching has been launched by C-MAC Frequency Products. The CFPO-6 range is specified to offer stability to within ±0.02 ppm over the –20°C to +70°C temperature range at output frequencies from 4MHz to 40MHz. Alternatively, it can cover frequencies up to 60MHz at ±0.25 ppm stability over temperature range. It is packaged in a low-profile 25.4 x 25.4 x 12.7mm case.
C_MAC Frequency Products
Tel: 01460 74433

900/1900MHz SiGe LNA

Temic Semiconductors has introduced two dual-gain low-noise amplifiers (LNAs) manufactured in SiGe technology. The IC TST0950 can be used for conventional superheterodyne or direct-conversion receivers in



925MHz to 960MHz GSM mobile phones. The companion device, the TST0951 is designed for mobile phone applications in the frequency range of 1800MHz to 2000MHz. The LNAs incorporate a 2-stage amplifier and switchable gain and offer a noise figure of 2.2dB in high-gain mode. The large signal capability is the equivalent of IIP3 equal to –7dBm in low-gain mode, and the reverse isolation is a minimum of –40dB.
Temic Semiconductors
Tel: 00 33 1 30 60 70 68

Miniature baluns for Z conversion

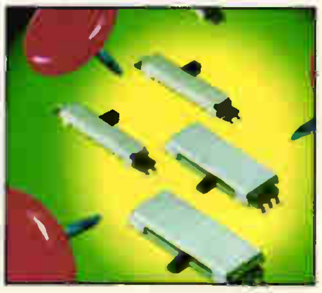
A range of miniature baluns (balanced/unbalanced devices) designed to provide passive impedance conversion between 75Ω coaxial and 120Ω twisted-pair cable has been introduced by Harting. The products are designed to meet the electrical requirements for transmission specified in the ITU G.703 standard. Standard products are available with transmission speeds from 1.55 to 34Mbit/s, with 155Mbit/s devices available to special order. The baluns are available with a variety of interconnection styles, and feature full shielding for EMI/RFI protection. Insertion loss from 1 to 17MHz is less than 0.5dB, return loss is less than –22dB, and crosstalk (with 15mm between centres) is greater than 80dB.
Harting
Tel: 01604 166686

Oven-controlled oscillator

Quartz Pro has introduced a fully automated mass production process for the manufacture of Quartz Pro X-ACT oven-controlled crystal oscillators. The supplier claims that the technique achieves yields exceeding 90 per cent which it says will have a favourable impact on product lead-times.
AEL Crystals
Tel: 01293 524245

Miniature slide and rotary pots

The range of miniature slide and rotary thumb wheel potentiometers available from



USB audio controller

Flint is offering an intelligent audio IC which includes an interface for the universal serial bus. The Micronas UAC3552A contains all the functionality required to balance a loudspeaker enclosure and connect the speakers direct to the PC without involving a sound card. The design integrates a DSP, EEPROM, high-performance digital-analogue converter, operational amplifier and USB controller. The device stores data for human interface device implementation on EEPROM. For example, volume and other parameters may be adjusted on the speaker, then transferred to the PC via the bus for display on control panels.
Flint
Tel: 01539 510333

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Quiller Electronics, the TBM NS range consists of a 3.5mm slide volume control intended for applications such as LCD back-light control of laptop and notebook PCs and a 2.5mm slider that can be used for zoom lens positioning on compact photographic equipment. There is also available a rotary potentiometer with a metal shaft that has a rotational angle of 220° and an outer diameter of 10mm and a footprint spacing of 2.5mm. All units are available in both surface mount and through-hole fixing and they require a small aperture for operation which is intended to minimise ingress of dust into the equipment.
Quiller Electronics
Tel: 01202 436770

Photoelectric sensor

Matsushita has introduced the UZB5/6 single chip optical IC photoelectric sensor with a

sensitivity adjuster integral to the sensor head. This, combined with a red dot LED, speeds up focusing and sensitivity changes. Sensing range is 2m, depending on type. It incorporates M3 mounting screw holes with reinforced metal inserts. It comes in retro-reflective, through-beam, diffuse and convergent reflective variants with either pnp or npn outputs.
Matsushita
Tel: 01908 231555

19in subrack enclosure

A Schroff enclosure for measurement and control suits applications such as process display, data transmission and data input. Available from Pentair, the Protega is rated to IP66 and has a solid plastic body, measuring 20 or 49HP, into which is fitted a metal BMC-shielded subrack that can hold all 19in. components. The 3U product can be supplied in wall-mount or desktop versions. Modules can be installed without the use of any tool. Hinges on the right and left sides let users open the enclosure from either side. The front panel and the whole of the front of the enclosure can be flapped giving the user access to the front and rear of the modules, backplane and cabling. Users can seal the rear, base and front modules by preventing the hinges from opening with lateral barrier locks. The rear section can be supplied with or without a terminal box. When fitted with a terminal box, users have the choice of top or bottom cable entry.
Pentair Enclosures
Tel: 01442 240471

Digital switch

IDT's TSI time slot interchange digital switch is for use in converged voice and data networks. It lets voice-over-IP applications maintain an interface with traditional telephony systems, providing the ability to switch and route digital telephone signals during the transfer of voice

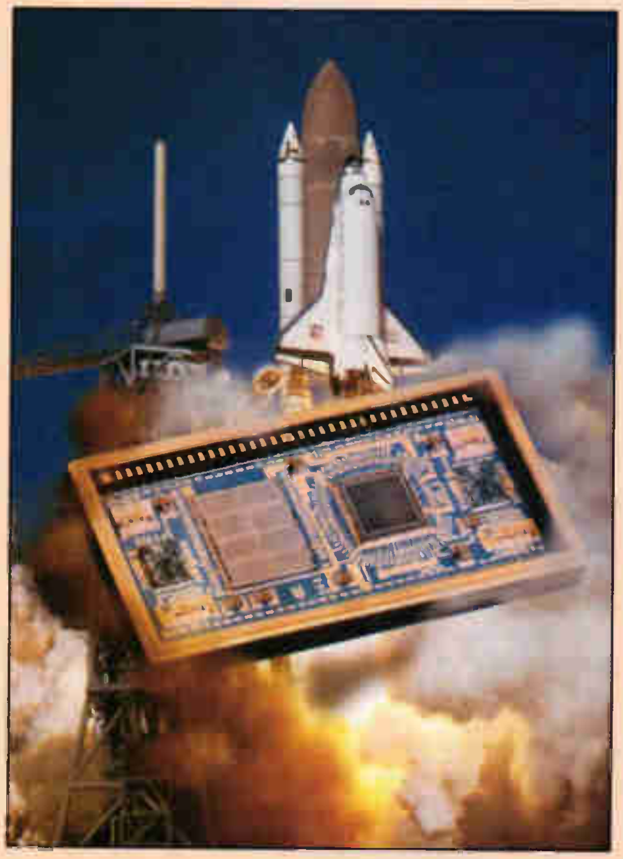
information. It is based on the traditional time-division multiplexed interface standards. Primarily used in public-switched telephone networks, it provides PCM protocol support and is compatible with TDM bus standards. Features include buffered data modes to support voice and data switching, microprocessor interface and a JTAG port. .
IDT
Tel: 01372 363339

Base station antennas

Radiall has introduced antennas for micro cell base station applications. The range includes embedded omnidirectional wire patch, embedded patch array and embedded sectional patch antennas. Bandwidth is 1710 to 1800MHz (DCS) and 880 to 960MHz (GSM), intermodulation –165dBc and VSWR 18:1 maximum. Horizontal beamwidth is 60° (DCS) and 75 to 80° (GSM), and vertical beamwidth 45° (DCS) and 70° (GSM). Matching impedance is 50Ω and isolation between antennas –20dB.
Radiall
Tel: 0208 997 8880

Space databus interface

DDC's BU-61582 radiation hardened version of its ACE MIL-STD-1553 databus terminal is intended for use on launch vehicles, satellites, space stations, the space shuttle and other extra-terrestrial vehicles. Implemented in Honeywell's 0.8μm RICMOS (Radiation-Insensitive CMOS) process, it can be configured as a bus controller, remote terminal or monitor terminal. The device is available with or without bipolar transceivers in voltage options of +5V/–15V or +5V/–12V. Units are designed to withstand a total gamma ray dose of 70Krads. A gamma dose rate as high as 1.0 x 10^{10} is tolerated with 25μs recovery. The 32k x 8 RICMOS RAM is utilised in a shared RAM microprocessor interface as if it were configured as 16k x 16. Package choices are 70-pin DIP or 70-pin flat pack for the BU-61582 or 196-pin ceramic quad flat pack for the BU-65621.
DDC
Tel: 01635 81140



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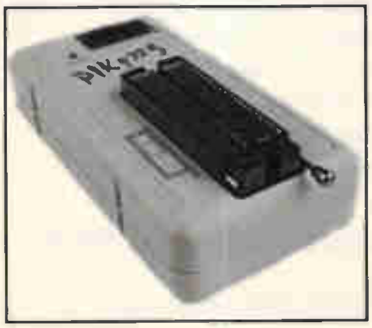

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Crownhill Associates is offering an exclusive 15% discount on any product in the company's programmer range to any reader of *Electronics World* until 31 August 2000. This unique reader discount is of particular interest in that the range of programmers on offer is competitivley priced even without the discount.

For more details on the of range programmers, see Crownhill's web site, detailed below, and the company's advertisement on page 551 of this issue.

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Single/dual input thermometer

Fluke has introduced two digital thermometers, the 50S single input and 50D dual input units. Both work with J or K type thermocouples and have 0.1°C resolution over the measurement range. They are protected by a rugged yellow holster. An offset function can be used to reduce thermocouple errors. Degrees C or F can be selected. A hold mode freezes the display. The 50D has minimum and maximum recording, scan mode to cycle the display and differential T1-T2 mode.

Fluke
Tel: 01923 2216400

315-916.5MHz transceivers

Flint has introduced licence-exempt 3V transmitters, receivers and transceivers in frequencies from 315 to 916.5MHz. The RFM TR3000 surface mount hybrid transceiver uses amplifier-sequenced hybrid (ash) architecture in a metal package measuring 10.7 by 9mm. It covers the European 433.92MHz unlicensed band for security and control applications, supporting data rates up to 115.2kbit/s. For handheld systems, it typically draws 1.8mA from a 3V battery and has a sleep mode. The RX ash receiver has data rates configurable from 2.4 to 115.2kbit/s. The receiver and TX hybrid transmitter are single chips requiring a 3V battery and provide the 433.92MHz waveband for European operation under Etsi I-ETS 300220 and, in Germany, under FTZ 17 TR2100.

Flint Distribution
Tel: 01530 510333

10-bit a-to-d

Maxim has introduced the Max1425 10-bit monolithic bipolar analogue-to-digital converter with 20MS/s digitising rate. It is for imaging, high-speed communication and instrumentation applications. The device achieves a 61dB sinad and 72dB SFDR at 2MHz input frequency. Operating from a +5V analogue supply, it achieves this over a ±2V input range through the use of differential pipeline architecture and a 0.6μm CMOS process. An integrated, fully differential input track-hold amplifier provides wideband dynamic performance with a 5pF input capacitance and a 150MHz full-power bandwidth. An on-chip precision 2.5V bandgap reference supplies additional reference voltages. This allows

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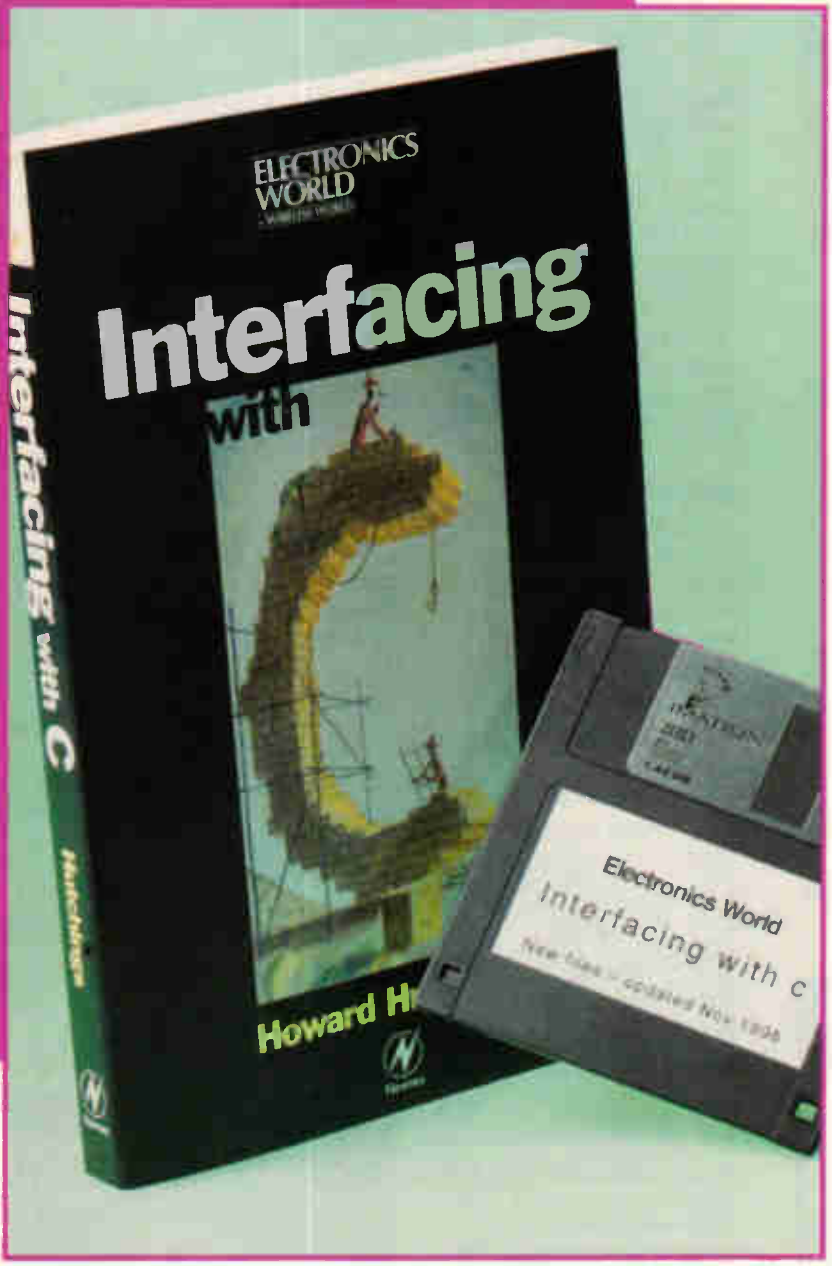

Without an engineering degree, a pile of money, or an infinite amount of time, the revised 289-page Interfacing with C is worth serious consideration by anyone interested in controlling equipment via the PC. Featuring extra chapters on Z transforms, audio processing and standard programming structures, the new Interfacing with C will be especialy useful to students and engineers interested in ports, transducer interfacing, analogue-to-digital conversion, convolution, digital filters, Fourier transforms and Kalman filtering. Full of tried and tested interfacing routines.
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Price £15, or £7.50 when purchased with the above book.

Especialy useful for students, the original Interfacing with C, written for Microsoft C Version 5.1, is still available at the price of £7.50. Phone 0181 652 3614 for bulk purchase price.

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flexible input range options and automatically ensures the correct DC bias levels for AC-coupled applications. A separate digital supply supports 3 to 5V output interfaces. The output data is presented in two's complement format. It comes in a 28-pin SSOP package for the –40 to +85°C range. An evaluation kit is available.
Maxim Integrated Products
Tel: 0188 930 3388

CPU supervisors

Insight Memec has introduced two single-chip user programmable, dual voltage CPU supervisors that integrate power-on-reset, watchdog timer, two low voltage sense monitors and 64kbyte of on-chip EEPROM. The Xicor X40620 and X46402 let system designers tune device parameters to requirements.

Both let the user select the low voltage reset trip points for the independent low voltage monitors. One monitor can be programmed from 1.7 to 3.5V and the other from 2.4 to 3.5V. The X46402 has user selectable security features such as memory partitioning and password protected memory. It also includes user selectable watchdog timeout selection intervals from 200ms to 1.4s. The watchdog timer function can be disabled for system software debugging.
Insight Memec
Tel: 01296 330061

DC/DC converter

Campbell Collins is stocking 2W DC/DC converters from Newport Components. The NML isolated single-output converters are for applications requiring isolation or conversion of DC power rails. They can start up at temperatures down to -40°C with full specification operation up to +85°C without using a heatsink or other external cooling device. Efficiency ratings are about 85 per cent and they come with nominal 5 or 12V inputs. Each input option can be supplied with single-ended outputs of 5, 9, 12 or 15V. They do not require external components for operation and use an internal SMD construction with ceramic capacitors.
Campbell Collins
Tel: 01438 369466

Power amp chipset

National Semiconductor has launched the LM4651 and LM4652 two-chip driver and quad MOSFET amplifier chipset for high-fidelity class D subwoofer designs. This chipset provides up to 170W of audio power at an 85 per cent efficiency. An integrated PWM modulation driver IC, the LM4651 contains short-circuit, undervoltage, over-modulation and thermal-shutdown

protection. A standby function shuts down the PWM and reduces supply current. The LM4652 is an integrated H-bridge power MOSFET. The two ICs require no external components between them. Specifications include an externally controllable switching frequency from 50 to 225kHz, integrated error and feedback amplifiers, turn-on soft start and under-voltage lockout, and a self-checking protected diagnostic. The LM4651N is housed in a 28-lead plastic DIP and the LM4652TF in a 15-lead isolated TO-220. ■
National Semiconductor
Tel: 00 49 8141 351 443

Three-phase power regulator

Tariq Iqbal's three-phase linear power regulator can control resistive loads or induction motors. Drive outputs are opto-isolated and regulation from zero to full load is via a single potentiometer so interfacing to a PC should be easy, assuming all mains isolation is properly implemented.

A three-phase optically-isolated power regulator with linear characteristics is shown in **Fig. 1** over the page. It can be used to operate a three-phase heater load or an induction motor with, say, a fan load.

Load power is controlled from zero to maximum by a single potentiometer producing the control voltage $V_{control}$. If required, control voltage may be supplied from a PC using a d-to-a converter card.

In this power regulator, the power stage and signal stages are isolated via light-activated S21ME3G triacs. Linear transfer characteristics are achieved using the cosine-wave crossing method of triggering the triacs.

In **Fig. 2**, operation of one of the the power regulator's three sections is explained using various waveforms. Waveforms of the other two sections of the power regulator are similar to Fig. 2, but delayed by 120° and 240° respectively.

Figure **2a)** shows typical three-phase line waveforms. These line voltages are stepped down using three transformers. Full-wave rectified output of line the L_1

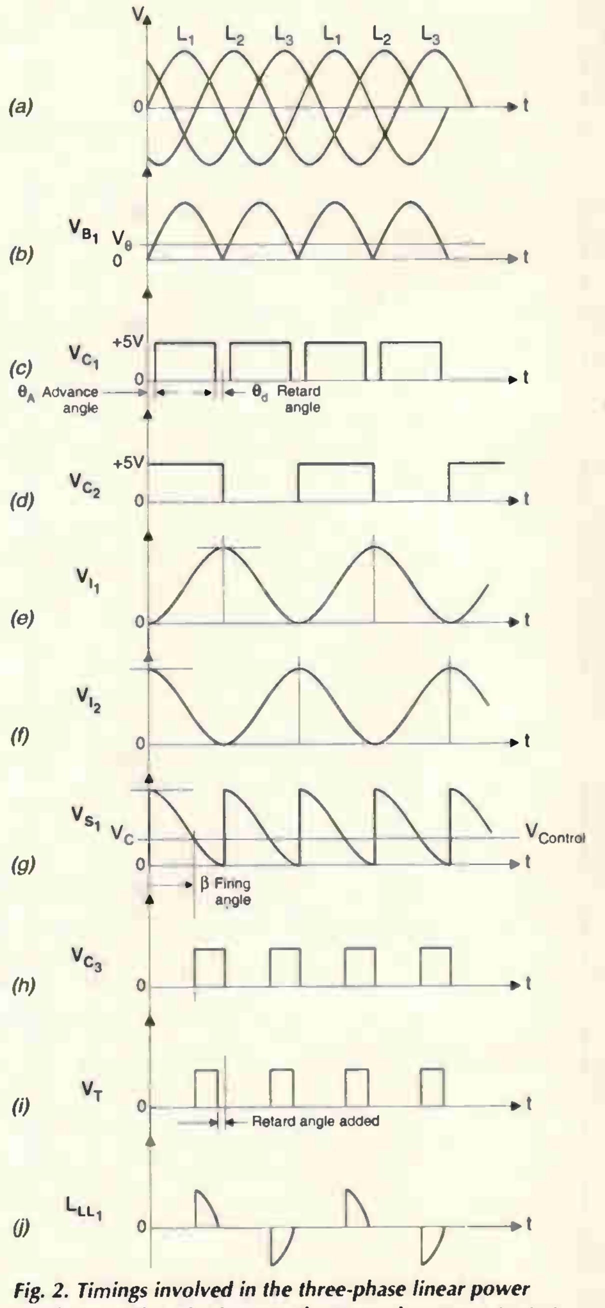
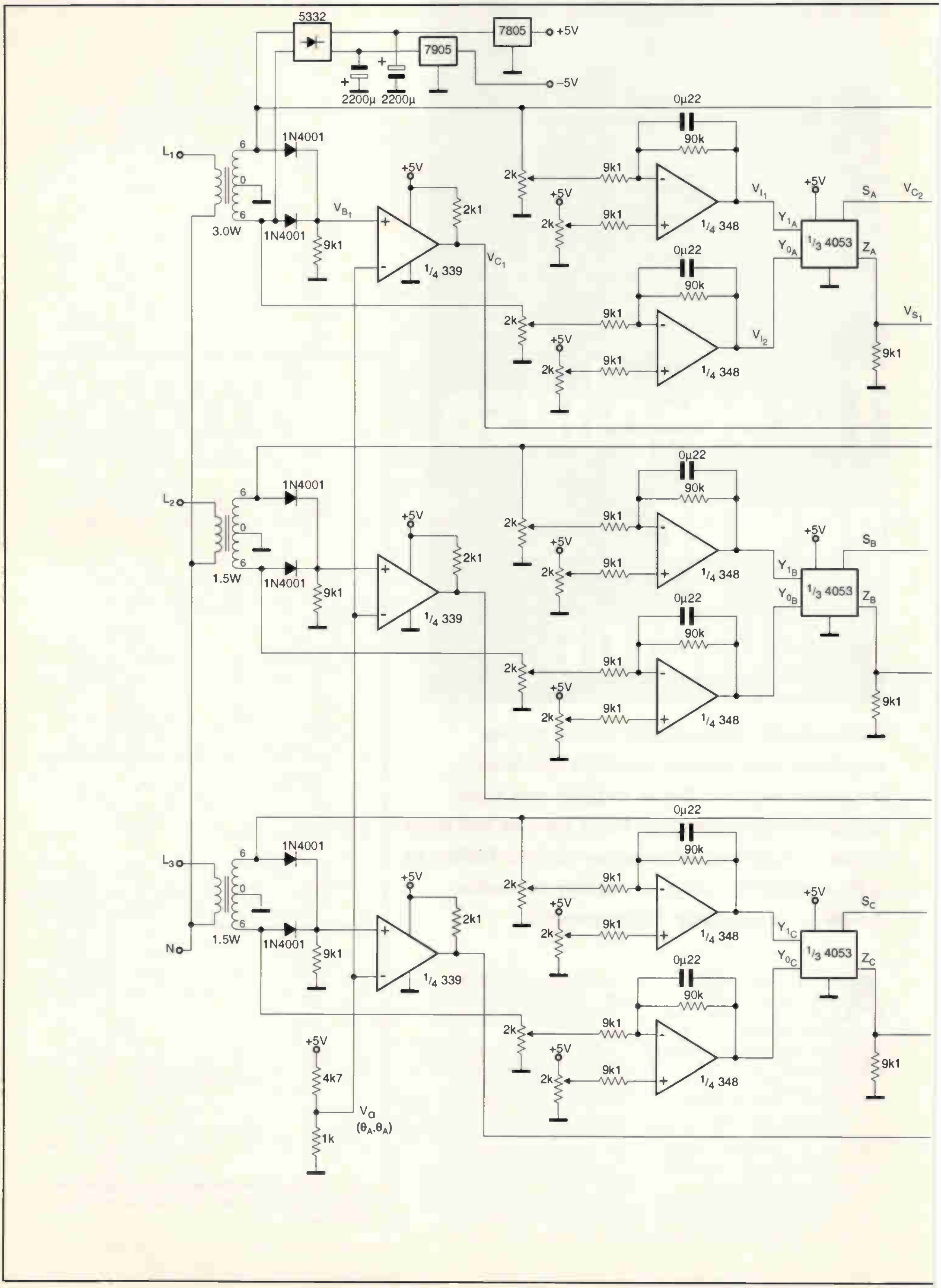


Fig. 2. Timings involved in the three-phase linear power regulator, with mains input at the top and power output at the bottom.



5332
7805
+5V
7905
–5V
2200µ 2200µ
0µ22
90k
1N4001
L_1
6
0
6
3.0W
V_{B_t}
+5V
2k1
9k1
1/4 339
V_{C_1}
2k
9k1
1/4 348
V_{I_1}
V_{I_2}
+5V
S_A
V_{C_2}
Y_{1A}
Y_{0A}
1/3 4053
Z_A
V_{S_1}
L_2
1.5W
S_B
Y_{1B}
Y_{0B}
Z_B
L_3
N
S_C
Y_{1C}
Y_{0C}
Z_C
4k7
1k
V_α
(θ_A.θ_A)

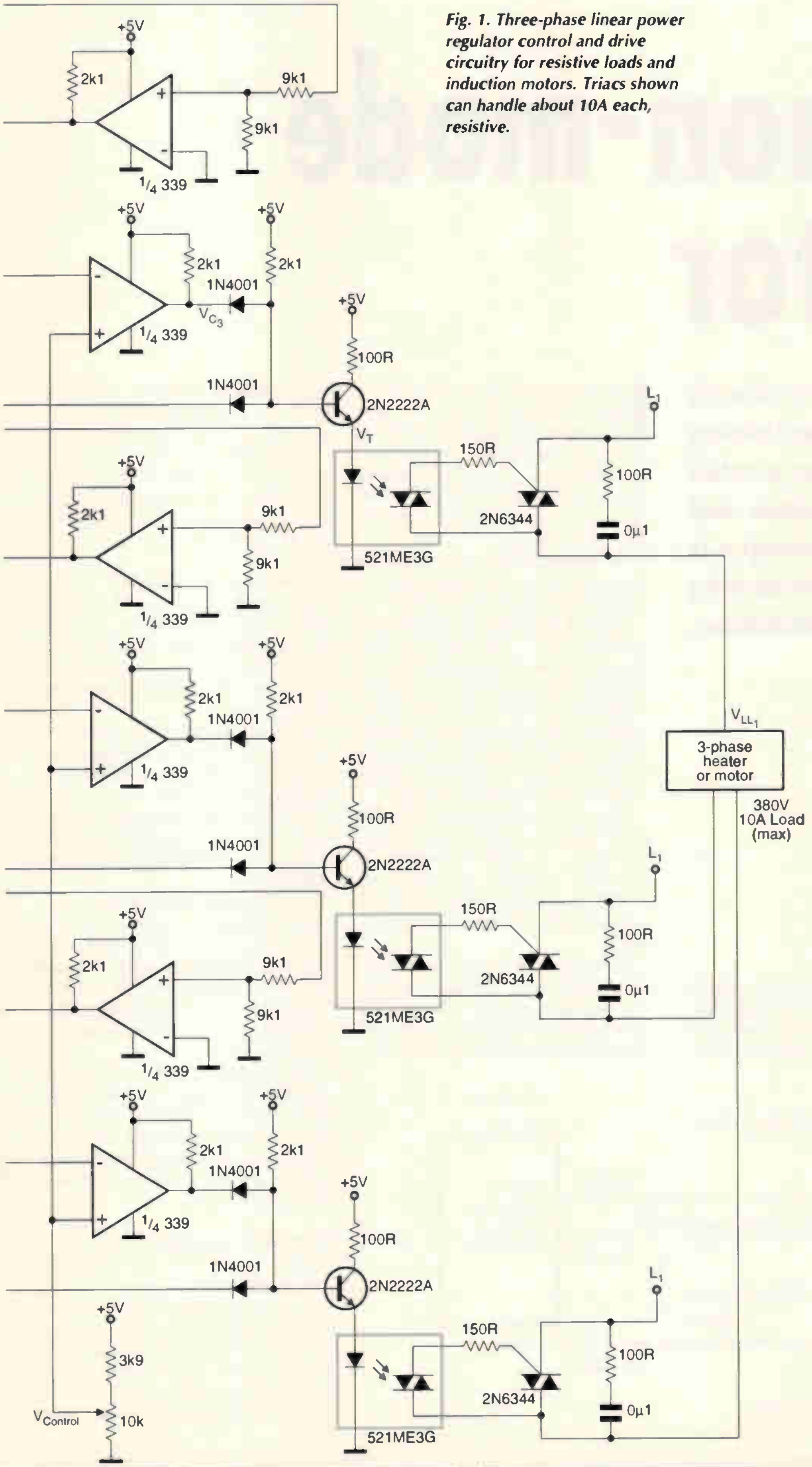


Fig. 1. Three-phase linear power regulator control and drive circuitry for resistive loads and induction motors. Triacs shown can handle about 10A each, resistive.

transformer is shown in Fig. **2b)**. This is compared with a small voltage V_θ to generate advance and retard angles. Advance and retard angles, θ_a and θ_d, are required to make sure that there is some voltage across the triac before it is triggered.

Figure **2c)** shows the advance and retard angle for phase L_1. A further comparator generates a reference waveform, V_{C2}, required for the isolation of two half cycles of cosine waves.

Two 180°-shifted biased cosine waves are generated using two integrators, Fig. 1. Both 2kΩ potentiometers of each integrator should be adjusted to obtain integrator outputs as in Figs **2e)** and **f)**.

Using reference voltage V_{C2} and a 4053 multiplexer, two cosine waveforms are combined to generate a reference cosine triggering waveform Fig. **2g)**. This is needed to achieve the power regulator's linear transfer characteristic.

Voltage V_c and the reference cosine waveform are compared to generate the respective firing angle, β, as in Fig. **2h)**. An AND function using two diodes and a transistor produces advance and retard angles.

The resulting waveform, Fig. **2i)**, is used for triggering the triac connected to phase L_1 of the supply. Three similar sections of the power regulator make sure that all three triacs are fired at exactly equal delay angle β.

I used a 2N6344 power triac in the original design. It is capable of carrying 10A load. If required, high current devices may be used. ■

The common-mode resistor

Ian Darney describes a simple method of minimising interference on a cable connecting a transmitter and receiver. The same technique is valid for reducing EMI coming from the interconnection.

It's frustrating to hear news that an item of equipment that you have just developed has failed its formal test for electromagnetic compatibility, i.e. EMC. It is especially frustrating when a great deal of time and effort has gone into the design to prevent such an eventuality.

The purpose of this article is to explain how such things can happen, and to describe a solution to the problem.

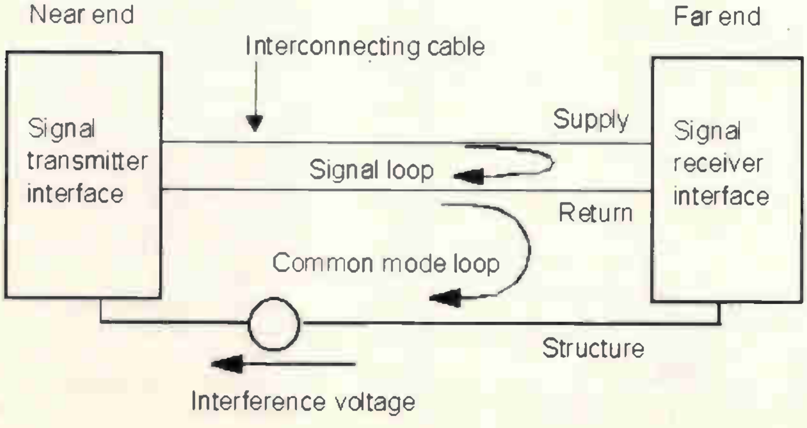
Figure 1 identifies the main factors affecting EMC. The objective is to transmit an electrical signal from a transmitter to a receiver, using cable conductors to carry the current. If the cable is a long one, then two conductors are needed to perform this function efficiently. For a 500MHz signal, 'long' means greater than 150cm.

These two conductors are identified in the diagram as 'supply' and 'return'. Inevitably, a third conductor is involved, variously described as 'earth', 'ground', or 'structure'. This carries any fault current, and acts as a shield.

In the presence of an external electromagnetic field, spurious currents will be generated in the structure, and these will create voltages along the surface. Circuit theory allows this electromagnetic interference – EMI for short – to be represented by a voltage source in series with the structure.

Since there are three conductors, there are two loops, the 'common-mode loop' carrying interference current, and the 'signal loop' carrying signal or supply current.

Fig. 1. EMC design fundamentals. In electronics, signals are carried from one point to another via conductors. The structure of these conductors becomes more important as frequency increases.



Making a circuit model

It is possible to construct a circuit model of this configuration, as in **Fig. 2**. Component values can either be calculated from physical parameters, or measured with general purpose test equipment[1].

In this model, the inductive and resistive values assigned to the return conductor, L_2 and R_2, can be described as the transfer impedance.

Analysis of the coupling between the two loops is a straightforward task, given the availability of circuit analysis software[2].

Such a model can provide a very accurate simulation of the response at low and intermediate frequencies, where resistance and inductance are the significant parameters. The effect of capacitance in creating resonance peaks and troughs can also be predicted with fair accuracy.

It is possible to develop the model to simulate the response at even higher frequencies, by connecting several elements in series, in the same way that a ladder network is formed to simulate a transmission line. But the task then becomes tedious.

Simplified approach

Besides, there is no need to invoke such complexities.

Using the concept of equivalent circuits, the inductors L_2 and L_3 can be combined, to represent the inductance of the cable. A similar process can be applied to the resistive and capacitive parameters of the supply and return conductors. It is then a simple matter to create a single T-network representation of the cable/structure assembly.

Analysing this using transmission-line theory provides a response characteristic for the common-mode loop current.

If this current is assumed to flow in the transfer impedance, the voltage across this impedance can be calculated. This 'transfer voltage' can then be simulated by a voltage source in series with the signal loop, and the currents and voltages in this loop are now amenable to analysis.

Using this method, each loop can be treated as a two-conductor transmission line.

Transfer impedance

The key to this approach is the ability to assign a value to the transfer impedance, and this can only be determined from the initial circuit model. For a wire pair, it is the inductance of the wire in series with its resistance[3].

For a coaxial cable, it is just the resistance of the screen[3]. Any voltage developed by interference current in the screen can be expected to be extremely low, giving a high level of common-mode rejection in the system.

Choice of configuration

The reasoning so far points to the use of coaxial cable to carry signals, and to the choice of the characteristic impedance as the value for the resistors at each end of the line. This gives the most efficient method of transmitting a signal between two locations in a system, and the least risk of interference. No surprises there.

A choice still needs to be made on the best method of interfacing with the structure. Grounding and floating are the two most obvious options.

Floating. The response of the floating configuration when the frequency reaches the first resonance point is shown in **Fig. 3**. When the quarter wavelength of the interference is equal to the length of the cable, an open circuit at the far end of the line looks like a short circuit at the near end. High amplitude current flows in the return conductor, creating a high amplitude transfer voltage, and a high level of interference in the signal loop.

Grounding. Response of the common-mode loop when both ends of the return conductor are grounded is shown in **Fig. 4**. Here, the half-wavelength of the interference is equal to the length of the line. In this situation, there is a high current at both ends of the cable, and again, a high level of EMI.

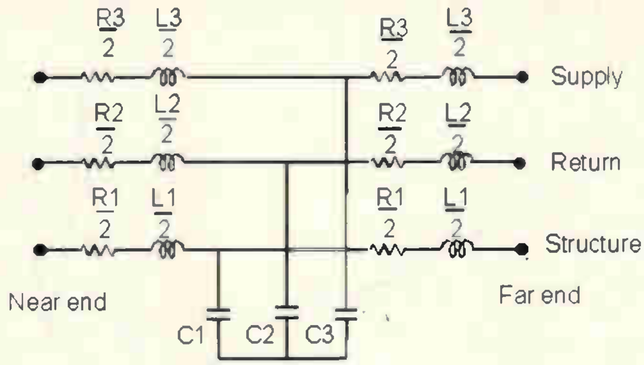
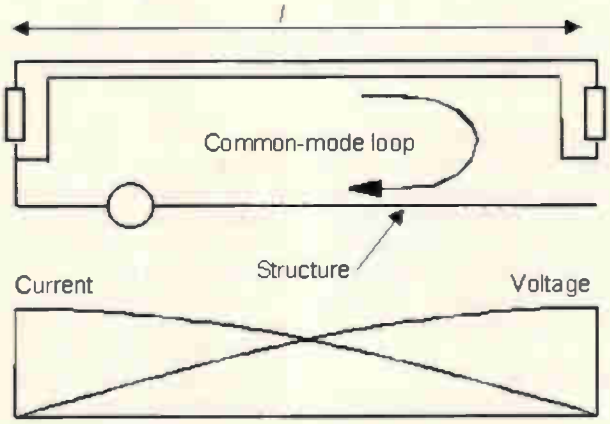
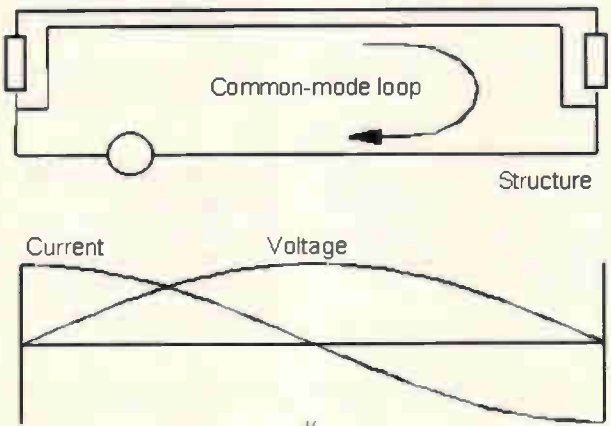


Fig. 2. Basic circuit model of three parallel conductors. This provides fair accuracy up to a quarter-wave frequency.



$v = 2.4\times10^{8}$ m/s $\qquad f_1 = \frac{v}{4\times l}$ Hz

Fig. 3. Floating configuration at its first resonance point – i.e. quarter-wave.



$f_2 = \frac{v}{2\times l}$ Hz

Fig. 4. Half-wave resonance response of the common-mode loop when both ends of the return conductor are grounded.

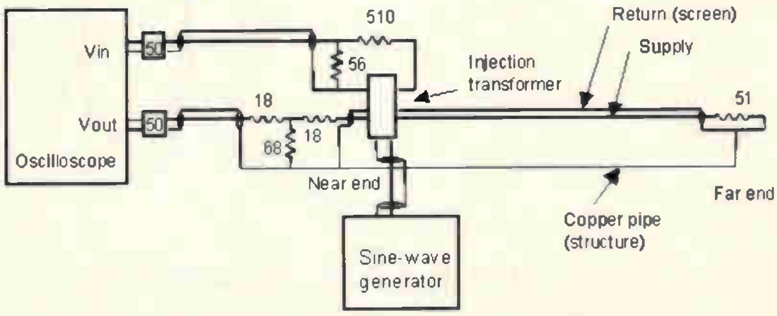


Fig. 5. Test set up for coaxial cable – grounded. Here, the 'supply' conductor is the core of a coaxial cable.

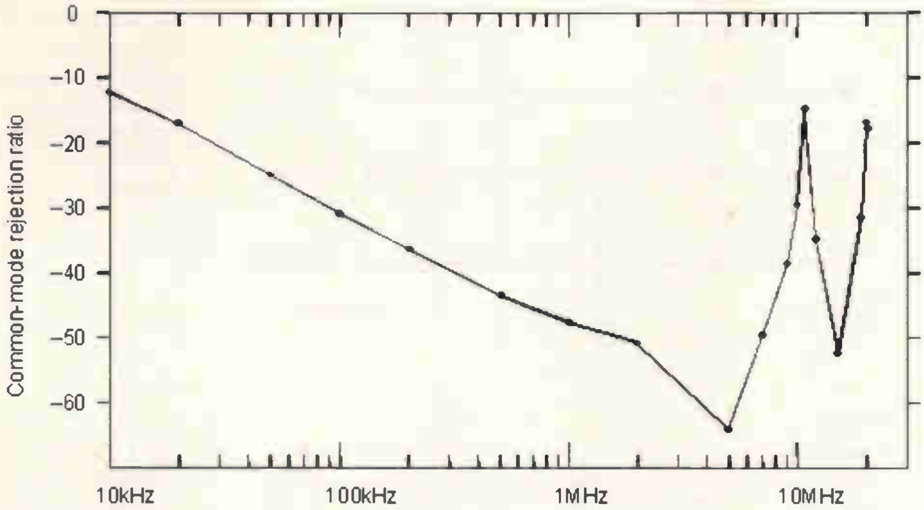


Fig. 6. Frequency response of grounded configuration confirms that common-mode rejection at low frequencies is low, but useful.

Hence, the floating configuration can be expected to cause high levels of interference at the quarter-wave frequency, and the grounded configuration will be subjected to high interference at the half-wave frequency. Transmission line theory indicates there will be many such peaks.

Making a test rig

The best way of determining the height of these peaks is to build a test rig and make actual measurements. So I built one.

This rig consisted of a length of 15mm copper pipe fixed round the walls of a room. I attached a wooden batten along the pipe to provide 12mm separation between cable and pipe. RG59 coaxial cable was fixed along the batten using cable ties.

An aluminium box was clamped to each end of the pipe, to provide a mounting surface for terminal strips. This provided a convenient method of interfacing with the test equipment. The length of cable between the two terminal strips was 11.7m.

Set-up

In the set-up of **Fig. 5**, the 'supply' conductor was the coaxial core, the 'return' was the screen, and 'structure' was the copper pipe.

At the far end of the cable, a 51Ω resistor was linked between supply and return. At the near end, a T-network was used to provide a 50Ω termination, and decouple the cable-under-test from the test equipment.

A toroidal transformer was used to inject a sine wave of about 1V peak to peak into the common-mode loop. An oscilloscope was used to measure the input voltage and the differential voltage appearing between supply and return.

The ratio of the output to input voltage was the common-mode rejection ratio.

Response of grounded configuration

For the first test, the cable was grounded at both ends, the equipment was connected as shown on Fig. 5, and the common mode rejection measured at a set of spot frequencies. Results are shown on **Fig. 6**. These confirm that at low frequencies the common-mode rejection is fairly low – albeit useful.

As the frequency increases, the situation improves, with the level of interference dropping at 20dB per decade. At the quarter-wave frequency, the common-mode rejection ratio is excellent. But that is as far as it goes. Thereafter, things take a turn for the worse, and the interference level reaches a peak at the half-wave frequency, and another at the full-wave frequency.

When the cable is grounded at both ends, the first peak is at 10.8MHz, and the second at just over 20MHz. At these frequencies, the common-mode rejection, even for a coaxial cable, is less than 20dB.

Response of floating configuration

For the second test, the link between screen and ground at the far end was removed, giving a floating configuration, and the test repeated. Results recorded in **Fig. 7** show that this configuration gives excellent performance at low and intermediate frequencies. This is marred by the response at high frequencies, which follows a pattern of peaks and troughs.

The frequency of first peak was measured as 5.3MHz, and the second peak occurred when the generator recorded 15.4MHz.

Problem and solution

Hence, both configurations suffer from the problem of resonance, which causes unwelcome peaks to occur in the frequency response.

But what can be done about this problem? Transmission-line theory points to the use of resistors at the cable interfaces to damp down the level of any resonance. It also points to the method of selecting a resistance value that will eliminate that resonance – namely, the characteristic impedance.

But how can resistance be inserted in the common-mode loop without adding unnecessary impedance to the signal loop? One answer to this lies in the use of a well known circuit component; that is, the common-mode transformer.

The common-mode resistor

I wound a small toroid with five bifilar turns of 22SWG enamelled copper wire and connected it as in **Fig. 8**.

In this set-up, a 240Ω resistor is inserted in the common-mode loop. Any current in the screen of the cable develops a voltage across one winding of the transformer. This creates an equal voltage in series with the supply conductor. The net result is that the interference voltage appearing across

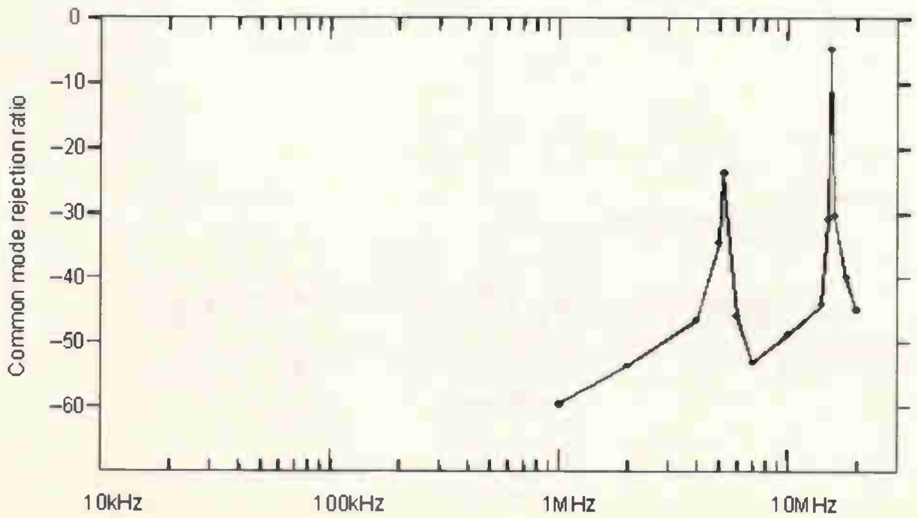


Fig. 7. Frequency response of floating configuration shows excellent performance at low and medium frequencies.

the common-mode resistor is cancelled out, and does not appear at the input terminals of the monitor circuitry.

The presence of a resistor in the common-mode loop acts to limit any current in that loop. Since a resistor is used, there is no dependence on frequency, and the current peaks will be limited, no matter where in the spectrum they occur.

Figure 9 compares the results of testing the grounded configuration with and without the common-mode resistor. Data used for the dotted curve was the same as used for Fig. 6.

Using a linear frequency scale makes the comparison easier. Both peaks are reduced dramatically. Significant improvement in the low-frequency response is also achieved. The transformer is effective at 10kHz and above.

Figure 10 demonstrates that the same action occurs with the floating configuration. The first peak was reduced by more than 20 dB, the second by more than 30 dB.

Review and deductions

I have shown that resonances can occur in the interconnecting cables of systems, creating large currents. Coupling via the transfer impedance induces high levels of interference in the signal loop. The effect occurs whether the configuration is grounded or floating.

It is possible to reduce the amplitude of the current at resonance by inserting a series resistance between ground and cable screen at a grounded termination. It is equally possible to reduce the amplitude of voltage reflections at a floating termination.

I have also shown that a common-mode resistor improves low-frequency performance. Since the braid of a coaxial cable acts both as a screen and as a return conductor, it was necessary to include a common-mode transformer in circuit. If the braid were to act solely as a shield, as with a screened wire pair, there would be no need for the transformer.

The cable/structure loop has been treated as a transmission line. It is equally valid to regard it as an antenna. The floating configuration can be treated as a whip aerial. When the cable is grounded at both ends, it becomes part of a loop antenna. Whatever the configuration, EMI can be absorbed by inserting a common-mode resistor in the loop at each cable/equipment interface.

Thus far, the reasoning has been in terms of minimising the susceptibility of a system. If the objective had been to reduce the level of emissions, the approach would have remained substantially the same, and the same solution would have been proposed. ■

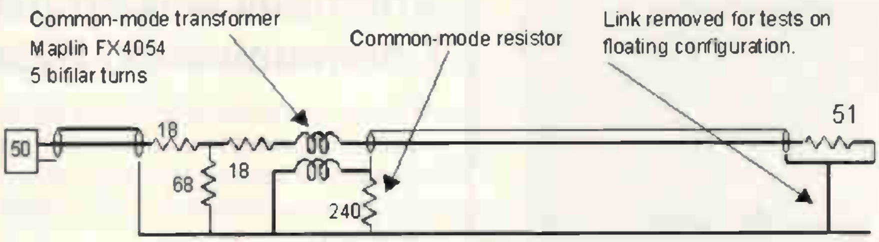


Fig. 8. Inserting a common-mode resistor.

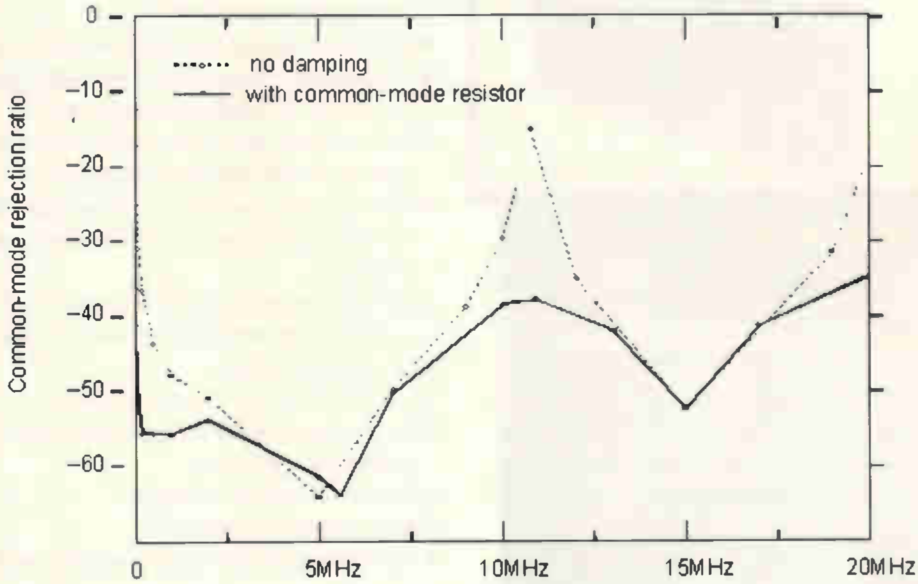


Fig. 9. Frequency response of grounded configuration, comparing results of testing the grounded configuration with and without the common-mode resistor.

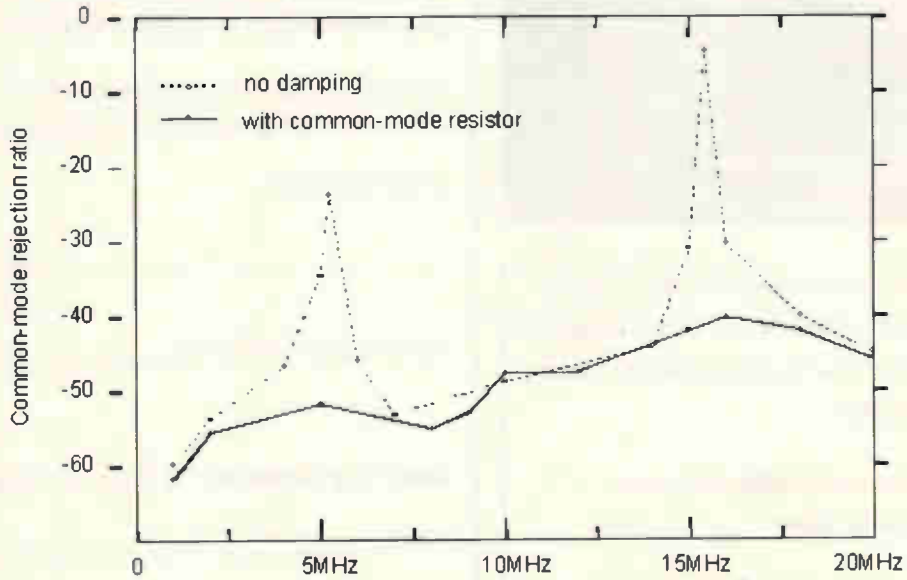


Fig. 10. Frequency response of floating configuration, again with and without the common-mode resistor.

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3. Darney, I. 'Grounding, Floating and Screening', IEE Seminar, 27 January 2000.

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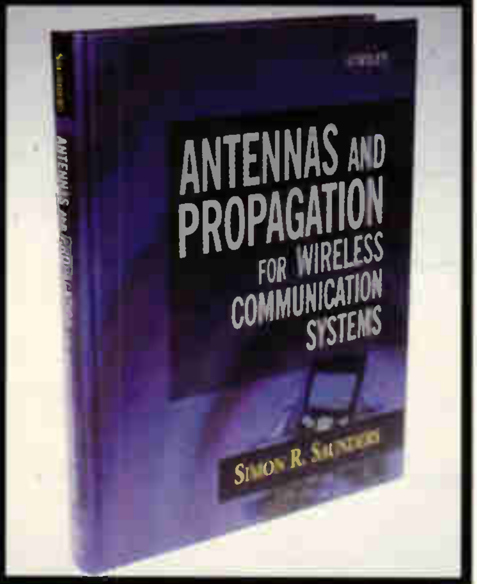

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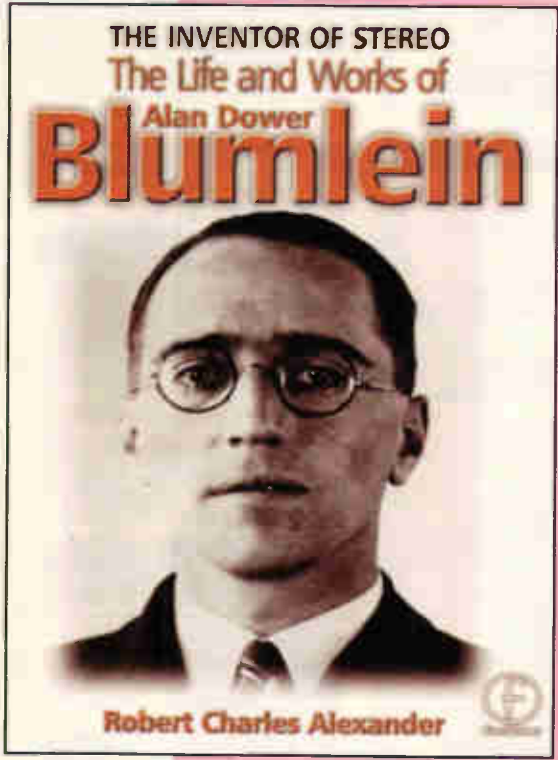

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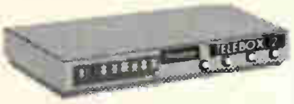

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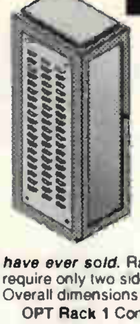

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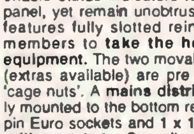

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CIRCLE NO. 121 ON REPLY CARD

Interfacing TINI

Having described TINI – a low cost Java-based controller with Web capabilities – Les Hughes now explains how to go about implementing it.

This second article examines how the tiny TINI controller can fit into a web-based architecture. Made by Dallas Semiconductor, TINI is a Java powered networked controller with a host of features and interfaces, but costing only around $50.

In last month's article, I looked at a simple 'hello-world' style network application. This article explains how to access TINI's external interfaces.

J2EE comes to J2ME

The Java 2 Enterprise Edition – J2EE for short – is a set of application programming interfaces and specifications. They are often employed as the enabling technologies for large scale computing projects.

The set includes database access, transaction management, distributed software components and dynamic content generation for web-enabled applications: EJB, JMS, JTS, JSDK, JDBC, RMI; TLAs galore! By the way, TLAs are 'three-letter acronyms'.

Machines are resource packed, fast and extremely expensive. If you've ever spent more than half an hour browsing .com sites, chances are that you've been using J2EE across half a dozen machines without ever downloading an applet. Powerful stuff indeed.

Java 2 Micro Edition – or J2ME – is at the other end of the spectrum. It is designed for small, memory-limited devices, hidden away in PDAs, cell phones and the like; resources are scarce and devices have to be cheap. Saving 0.005p per device counts when you're producing several million of them.

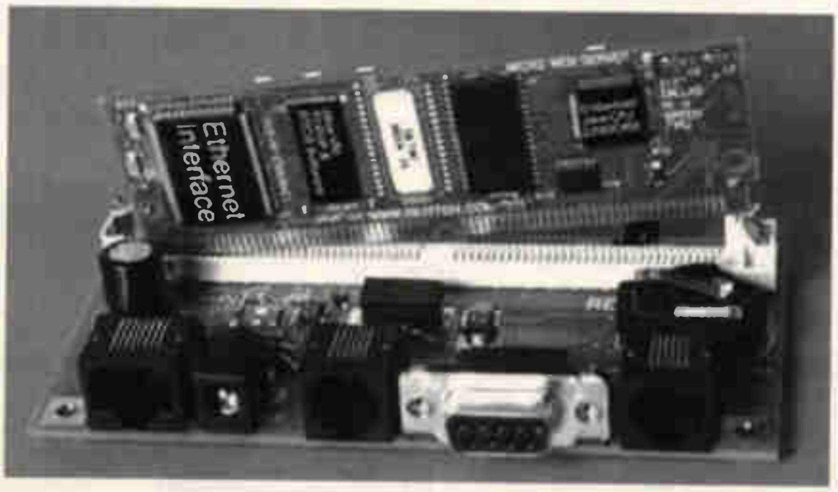
One of the really impressive things about TINI is the way that a small resource-constrained device can steal some of the wind from the big boys in 'enterprise computing land' and use some of the J2EE application programming interfaces effectively in a J2ME environment. The first of these to become available was the Java Servlet API.

Embedding the Web

One of the most important pieces of software that we need in order to create a J2EE/ME hybrid is a web server. Most applications come web-enabled nowadays, and TINI is no exception – it is the Tiny Internet Interface remember.

However, serving up a series of plain, static HTML is of no use to anyone anymore. We want real-time information concerning the devices attached to TINI, and for that we need to get TINI to run code for us on demand.

The J2EE technology for this is the Java Servlet API. Servlets are similar to the Java Applets that you can download onto your machine except that they never leave the server. As a client, you simply point you web browser at the servlet's URL and then sit back and let it run. Output is generated dynamically by the servlet and displayed on your browser's screen.



In order to run Servlets on TINI, you need a web server and a servlet engine. Smart Software Consulting provide a GPL'd webserver and servlet engine for TINI in source code format. As with any open source software, you're free to modify this code as you see fit and you are able to distribute you changes.

Creating a servlet

In their simplest form, servlets have a method that is invoked by the web server when it receives a request to activate the program from a client. If the client is sending data from an HTML form, this method is called 'doPost' and if not, 'doGet'. In order to access TINI's devices across the net, we just need to create a servlet that talks to our hardware in one of these methods.

An example application

A simple application – a basic entry system – will help show how easy it is to web enable an embedded controller with TINI.

A number of companies use Dallas iButtons for entry control. These are most often based on the DS1990A, an iButton that has a unique ID laser cut at the factory. TINI of course has an iButton one-wire interface. It also has an RS232 port that could be connected to an external LCD display, as well as direct digital outputs that could be used to operate a door lock.

We could also use a web browser to access the lock, in order to view entry logs, upload new authorised users etc. as this mode of operation is becoming quite common for system administration tasks on larger systems.

So to summarise, we'll design a lock that,

1. reads the ROM ID from any connected iButton
2. checks whether the user is authorised
3. sends a welcome / 'Go Away' message to the serial port which is con-

List 1. Using a web browser to access a lock, in order to view entry logs, upload new authorised users etc.

```
//std jdk imports
import java.io.*;
import java.util.*;
//std java extension imports
import javax.comm.*;
import javax.servlet.*;
import javax.servlet.http.*;
//Other imports
import com.ibutton.*;
import com.dalsemi.system.Clock;
import com.dalsemi.system.BitPort;
public class EntryServlet extends HttpServlet implements Runnable {
     int[] romID;
     String s_romID;
     Thread runner;
     Vector accesses;
     Hashtable users;
     Access onewire;
     BitPort cpuled;
     Clock clock;
     //initialise the servlet
     public void init() throws ServletException {
          //Initialise our local fields
          accesses = new Vector();
          clock = new Clock();
          cpuled = new BitPort (BitPort.Port3Bit5);
          //Lookup table - put your entries here....
          users = new Hashtable();
          users.put("0x01CF2E12050000D6","The Boss");
          //Start a Thread to listen to the lock
          runner = new Thread(this);
          runner.start();
          System.out.println("EntryServlet: Started Thread.....");
     }
     //Send the log to the Browser
     public void doGet(HttpServletRequest req, HttpServletResponse resp)
          throws ServletException, IOException {
          resp.setContentType("text/html");
                    PrintWriter out = resp.getWriter();
                    //Output HTML Header
                    out.println(
                    "<HTML><HEAD><META HTTP-EQUIV=\"Refresh\" CONTENT=\"3\"><TITLE>Access Logs</TITLE></HEAD>");
                         //Output a simple style sheet for our log page...
          out.println("<STYLE TYPE=\"text/css\"><!—");
          out.println("BODY { background-color: burlywood }");
          out.println("H1 { color: white; background-color: gray; font-family: sans-serif; font-size: 14pt; font-weight:
                                                                                                   normal }");
          out.println("TD { color: green; background-color: white; font-family: \"Lucida Console\", courier, monospaced;
                                                                                       font-size: 10pt }");
          out.println(".heading { color: white; background-color: gray; font-family: sans-serif; font-size: 12pt }");
          out.println("—></STYLE>");
     //Now Output our data
          out.println("<BODY><H1>Viewing Access Log</H1>");
          out.println("<CENTER><TABLE WIDTH=90% BORDER=0><TR><TD CLASS=\"heading\">Log Entry</TD></TR>");
               //Check our access log
          if(accesses.isEmpty()) {
               out.println("<TR><TD>Log is empty</TD></TR>");
          } else {
               for(Enumeration e = accesses.elements(); e.hasMoreElements();)
               out.println("<TR><TD>"+(String)e.nextElement()+"</TD></TR>");
          }
          out.println("</TABLE></CENTER></BODY></HTML>");
}//End of doGet
//Run method, polls lock etc.
public void run() {
          //First, we initialise our hardware, then we loop
          try {
               //Open OneWire interface
               onewire = new Access();
               onewire.SetAdapterType(com.dalsemi.comm.SerialPort.SERIAL0, "port0");
               //Open our RS232 Port and get a PrintWriter to talk to

               CommPortIdentifier portID = CommPortIdentifier.getPortIdentifier("serial0");
               SerialPort port = (SerialPort)portID.open("MessageDisplay", 1000);
               port.setSerialPortParams(115200,
                                        SerialPort.DATABITS_8,
                                        SerialPort.STOPBITS_1,
                                        SerialPort.PARITY_NONE);
               PrintWriter out = new PrintWriter(port.getOutputStream(), true);
```

Continued over page

nected to some kind of display

4. opens the door if appropriate (simulated by the flashing of one of TINI's LEDs)

Also, administrators will be able to connect to the lock using a web browser in order to view access logs.

Basic design

Our servlet needs to do two things at once; listen to the web and watch for people trying to gain access. This calls for a multi-threaded design, where one thread looks after the lock and the other the web.

So, instead of accessing our hardware when asked by a remote web browser, we shall be polling the one-wire interface looking for users wishing to gain entry. This function isn't really anything to do with web servers, servlets and such; we could

```
            for(;;){
                try{
                    //loop until we find a button
                    while(romID == null) {
                    //sleep for a while - give TINI a rest!
                        Thread.currentThread().sleep(500);
                        romID = null;

                        //At this point we could iterate through all connected
                        //buttons but for simplicity we'll assume there's
                        //only one button on the wire
                        //Select the first button on the wire if there is one
                        if(onewire.iBFirst()) {
                                //get the selected button's ROM data
                            romID = onewire.iBROMData();
                        }
                    }

                    //we now have a button - convert int[] into a String
                    StringBuffer romStr = new StringBuffer("0x");
                    for (int i = 0; i < 8; i++) {
                        if (romID[i] < 0x10) romStr.append("0");
                        romStr.append(Integer.toString(romID[i], 16).toUpperCase());
                    }
                    //Get the time of access and convert into something pretty
                    clock.getRTC();
                    int hour    = clock.getHour();
                    int minute  = clock.getMinute();
                    int second  = clock.getSecond();
                    StringBuffer timeString = new StringBuffer(Integer.toString(hour));
                    timeString.append(":");
                    if(minute < 10) timeString.append("0");
                    timeString.append(Integer.toString(minute));
                    timeString.append(":");
                    if(second < 10) timeString.append("0");
                    timeString.append(Integer.toString(second));
                    //Lookup user's romID - this could be a database call (but not on TINI yet!!)
                    String user = (String)users.get(romStr.toString());
                    if (user == null ) {
                        //unauthorised access attempt!!!
                        accesses.addElement("Unauthorised Access Attempt (ID="+romStr+") @
                                                                "+timeString.toString());
                        out.println("INTRUDER ALERT!!!");
                    } else {

                        accesses.addElement(user+"@"+timeString.toString());
                        out.println("Greetings, "+user+" pull the door.");
                        //Let's flash the 'CPU' LED to simulate the lock
                        for(int n=0;n<3;n++) {
                            cpuled.clear();
                            Thread.currentThread().sleep(500);
                            cpuled.set();
                            Thread.currentThread().sleep(500);
                        }


                    }
                }catch(InterruptedException interrupted) {

                }finally{
                    //always reset the romID to null
                    //otherwise we won't try to read again!!
                    romID = null;
                }
            //Now go back and monitor onewire
            }//end of for(;;)
        }catch(Exception setupExcept) {
            System.out.println("Exception in run "+setupExcept.toString());
        }
    }//end of run()
}//EOF
```

factor this code into a separate class to allow other security applications to re-use or logic. However, for the purpose of this example, we will leave everything inside one class.

Finishing off our design exercise, we'll create a servlet that starts up a new thread that is responsible for monitoring the one-wire interface, allowing access to those authorised and logging any access attempts etc. The servlet half of the application will simply turn the log into HTML for display on a browser.

The code

List 1 contains the code for our application. Logically this is divided into two parts; the servlet methods (`init` and `doGet`) and the reader thread method (`run`). The first few lines of code import various library classes. A number of local fields (variables) are next declared. Following these lines, comes the first of the servlet methods – `init()`. This method is called once when the servlet engine loads the servlet and performs initialisation tasks. In our case, this involves creating our local objects and starting the lock polling thread.

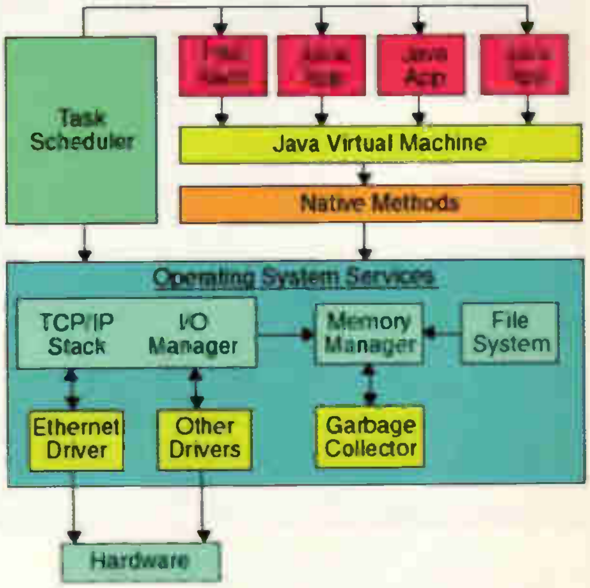
After init in the listing comes `doGet()`. This method simply writes out an HTML log of accesses using println statements.

Further down the listing comes the run method of our polling thread. This method contains all of our 'Security' logic; accessing the one-wire interface, reading the ROM ID of any buttons found, logging and allowing/denying access.

Future enhancements

Obviously, our EntryServlet is basic – to say the least. There's a number of enhancements that could be developed.

- Refactor the code into separate 'security' and 'servlet' classes.
- Enhance the one-wire scanning logic – perhaps two people's iButtons are needed to gain entry to a high-security area?
- Implementing a doPost method that allows new users to be added to the lock from a browser, taking this idea further we could:
- Develop an intra-lock protocol that allows multiple locks to talk to a central service for user verification, central logging etc. ■



Although very low cost, this controller is a complete computer with Internet, network and serial i/o capabilities, giving it huge potential for remote i/o and telemetry applications.

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Input capacitance	40pF+oscilloscope capacitance
Working voltage	600V DC or pk-pk AC

Switch position 2

Bandwidth	DC to 150MHz
Rise time	2.4ns
Input resistance	10MΩ ±1% if oscilloscope i/p is 1MΩ
Input capacitance	12pF if oscilloscope i/p is 20pF
Compensation range	10-60pF
Working voltage	600V DC or pk-pk AC

Switch position 'Ref'

Probe tip grounded via 9MΩ, scope i/p grounded

Variable-gain circuits

Cyril Bateman looks at the trade-offs involved with the various ways of producing variable-gain circuits and reports on what's available on the Internet to help deginers make their choice.

In last month's article, I discussed a selection of circuits for controlling signal amplitude. These used voltage-controlled amplifiers to produce low-distortion, variable-gain or closed-loop AGC circuits.

Many techniques are available. Some introduce distortion; others introduce phase and group-delay changes. A few methods though have no adverse affects at all.

Analogue multipliers can be used to provide a voltage-controlled amplifier or AGC circuit, but they can introduce noise. With increasing levels, the signal can even be crudely clipped using only a simple resistor diode network.

When the signal has a very wide dynamic range, logarithmic amplifiers or converters can be used to compress a 100dB input signal range into an output range of less than 10dB, or just two or three volts. Such circuits introduce distortion but can be effective with high frequency and wide bandwidth signals.

For a recent task I needed to

Bugs

As I write, Microsoft[A] has just released a series of patches to eliminate five security holes in its Internet Explorer and Office software. These bug patches result from criticisms of its security policies – part of the aftermath of the considerable damage caused by the recent 'Love-bug' virus.

Variants of this virus have now appeared that are even more damaging. These arrive as an e-mail with an attached file having a '.vbs' extension. Their subject line varies, so these versions are more difficult to filter than was the original 'Love bug'. Interested readers can find more details by accessing two web sites, BugNet[B] and Cnet[C].

If you access Internet using Internet Explorer software, you are advised to download these security patches from Microsoft, **Fig. A**.

Readers whose computers have mother boards using the Intel 820 chipset together with the new memory translator hub or 'MTH', chip may experience 'boot-up' problems. The MTH chip is used to enable the 820 chipset to address conventional memory pending the arrival of Rambus memory chips.

Intel is expecting to bear the cost of replacing large numbers of affected mother boards[D].

Fig. A. The latest version of e-mail based computer bugs. Developed from the recent 'Love bug' virus, if opened, this version can overwrite many computer files.

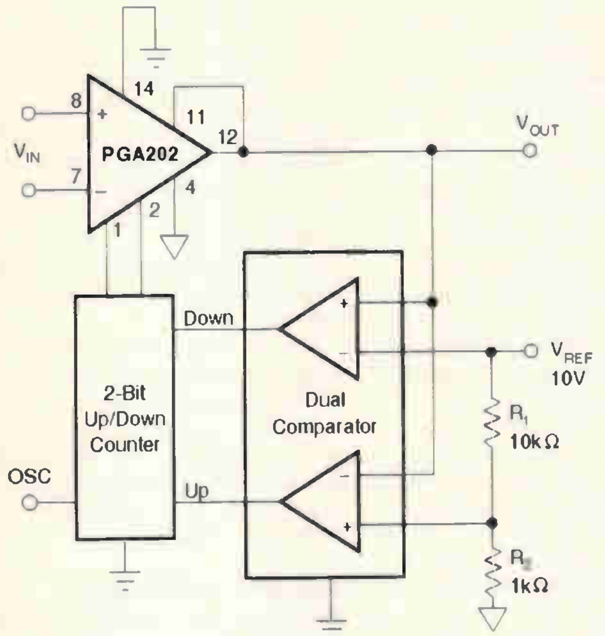


Fig. 1. Using a programmable-gain instrumentation amplifier as an auto-ranging gain control. The PG202 provides decade ranging steps. Its companion PG203, provides binary ranging.

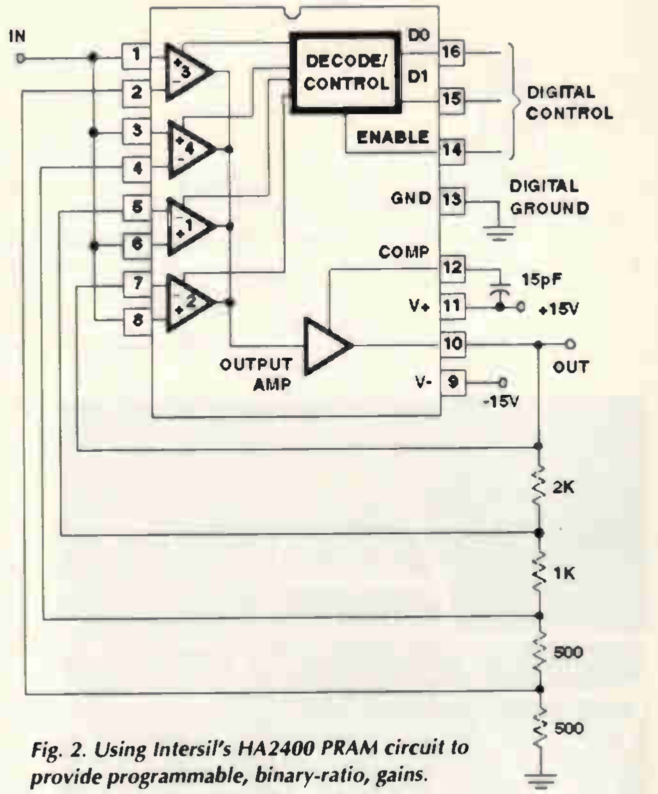


Fig. 2. Using Intersil's HA2400 PRAM circuit to provide programmable, binary-ratio, gains.

measure the time differences between two low level 1MHz sinewaves, having much differing amplitudes. To eliminate zero-crossing slew-induced errors, I needed to raise each signal to a consistent, measurable level, without affecting their time differences.

This article reports my recent Internet findings on variable-gain circuits.

Controlling gain

The input to a voltmeter or a-to-d converter may need to use gain or range switching to optimise its measurement range. A gain-selectable, programmable instrumentation amplifier IC, such as Burr Brown's PGA202/3[1], can provide either decade or binary gain steps, and very high gains, under simple logic control, **Fig. 1**.

Intersil's PRAM, or programmable amplifier, provides a variety of programmable functions besides simple gain blocks[2]. With its 40MHz gain bandwidth and 30V/µs slew rate, the HA2400 comprises a series of four-channel programmable amplifiers. Any amplifier – or indeed no amplifier – may be electronically selected and connected to a single output stage, using TTL-compatible address inputs.

A programmable-gain amplifier can be provided by arranging differing gains for each of the four channels. Input impedance of the device is 30MΩ and all four amplifier inputs may be connected in parallel. The required gain channel is then selected using simple logic, **Fig. 2**.

Where to surf

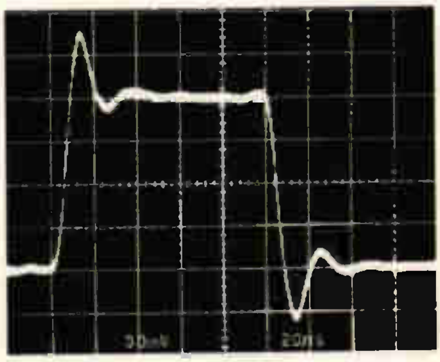
1. Burr-Brown Corporation	http://www.burr-brown.com
2 Intersil Corporation	http://www.intersil.com
3 Linear Technology Corporation	http://www.linear-tech.com
4 Analog Devices Inc.	http://www.analog.com
A Security update	http://www.microsoft.com/windows/ie/download/critical/patch6.htm
B Worm targets exchange e-mail address books	http://www.bugnet.com/alerts/bugalert_000504.html
C New virus more deadly than 'Love bug'	http://www.cnet.com/software/0-1592552.html
D Intel to replace faulty PC mother boards	http://news.cnet.com/news/0-1006-200-1849850.html

Capable of driving more than 20V pk-pk into a 2kΩ load at frequencies up to 400kHz, this amplifier has many applications besides the programmable gain example described.

'Clamp' amplifiers

In many applications, it is desirable to restrict, or 'clamp', the output from an op-amp to prevent saturation of subsequent stages. Simple circuits, such as the resistor diode clipping network mentioned, may suffice. Using a clipping diode in the feedback circuit of an op-amp though could be more effective.

Although simple, this circuit suffers from signal distortions and lacks precise amplitude control. If the input to the output stage of the op-amp is



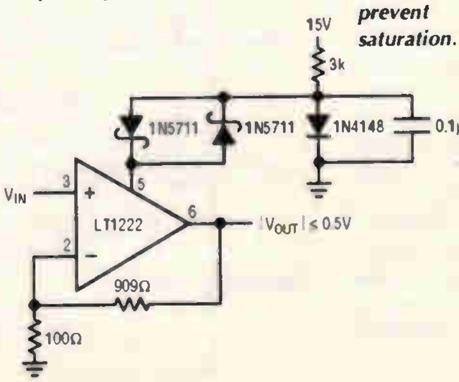
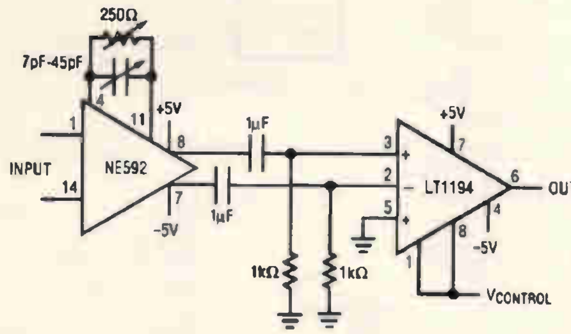


Fig. 3. Compensation input, pin 5, of the LT1222 IC can also be used to insert diode clamping, restricting drive amplitude into its output stage to prevent saturation.



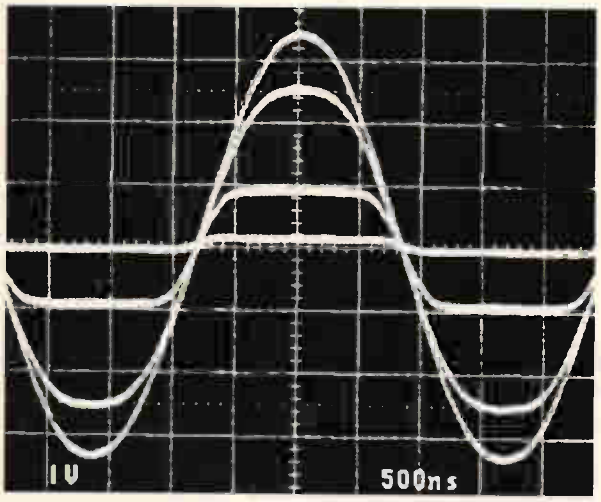


Fig. 4. Adjustable soft-limiting of this 200kHz sinewave was obtained by setting the $V_{control}$ voltage to -5V, -4V, -3V and -2V in turn.

externally accessible, performance can be improved by clipping the drive signal fed into the output stage. This stops the output saturating, **Fig. 3**.

When higher frequencies are needed, the LT1194 can be handy[3]. This is a high-speed video 'difference' amplifier that can be used as a variable limiting amplifier. This is done by varying the control voltage supplied to its two balance/limiting reference pins. Such an arrangement provides a controlled, 'soft' limiting, output voltage, **Fig. 4**.

Analog Devices' AD8036[4] is a low-distortion clamp amplifier designed to provide high-speed signal clipping with minimal pre-clipping signal distortion. Based on a 1200V/μs, 240MHz high voltage gain, unity-gain stable differential amplifier, this device has two high-speed internal comparators used to sense the input signal levels.

These comparators connect to independently-settable high and low reference voltages. Provided the input signal does not exceed these set voltages, the comparators have no effect. The amplifier functions as a normal, low distortion, op-amp.

When the input voltage exceeds one of the reference voltages, the relevant comparator switches, substituting the reference voltage for the input signal. The amplifier input voltage remains fixed at this reference voltage until the input signal becomes smaller than the reference, **Fig. 5**.

Log converters

The conversion of a signal's amplitude into its equivalent logarithmic value involves a non-linear operation. With small-signal inputs, gain of the converter is extremely large.

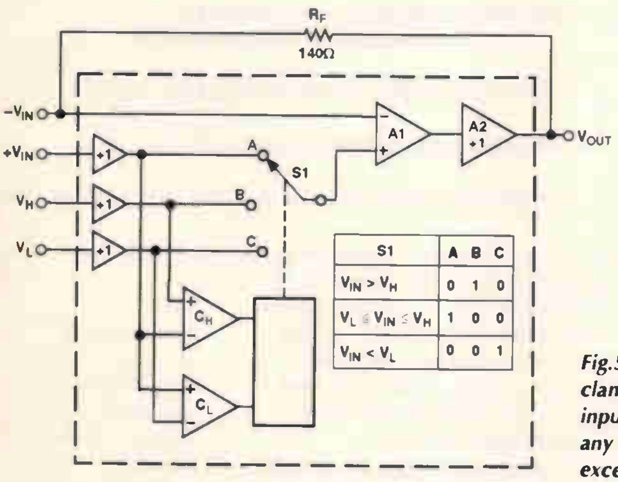


Fig.5. Analog Devices' AD8036 input-clamping amplifier, provides easy control of input drive levels. V_H and V_L can be set to any voltage between ±3.9V, provided V_H exceeds V_L.

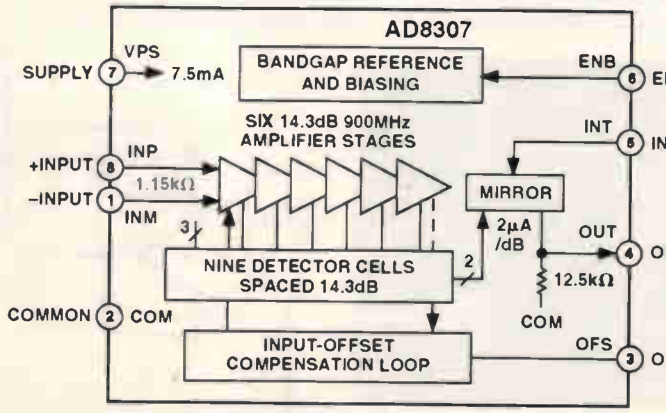


Fig. 6. This high-frequency logarithmic amplifier – the first in an SOIC 8 pin package – can measure signals from audio frequency to 500MHz, producing a voltage output scaled at 25mV/dB.

Any change in input offset appears to the converter output as though a change of input signal has occurred.

It is possible to produce a low-frequency logarithmic conversion using only an op-amp with a log feedback diode. Log converters intended for high frequency, low-level signals though use multiple gain stages. A good description of this construction can be found in the AD8307 data sheet from Analog Devices[4].

Released in July 1997, the AD8307 claimed to be the first low cost 500MHz log amp in an 8 pin SOIC package to measure IF signals. It is able to measure signal strength with a ±3dB accuracy over a 92dB measurement range, improving to ±1dB over an 88dB range, **Fig. 6**.

Exponential amplifiers

Perhaps the most conventional way to vary the effective gain of a circuit is that used for a 'volume' control. Here, the circuit gain remains constant but its input signal is deliberately attenuated as desired.

Analog Devices[4] has contrived this mechanism in silicon within a conventionally-packaged IC. An externally-supplied control voltage can attenuate the input signal over a 40dB range. The attenuated input is then amplified by a fixed gain, very low-noise, broadband amplifier.

The net result is a quick-responding variable-gain amplifier having a response that's 'linear in dB' and low distortion. With V_g set to 0V, the chip is precisely calibrated in production, ensuring accurate gain control.

The resistive input attenuator used with a fixed-gain amplifier assures both a constant bandwidth and near-constant group delay. Analog Devices has classified this novel circuit as an 'X-Amp', which is short for exponential amplifier of course.

The AD600/AD602 designs were published in their 'Special linear reference manual – 1992'. Both provided dual-channel 'X-Amps' in a 16-pin DIL package with a 35MHz bandwidth. They differed only in the gain of their fixed-gain output amplifier. The AD600 has a gain of 41.07dB and the AD602 10dB less. Their input attenuators can reduce the input by up to 42.14dB.

Subsequently, other variations on this circuit have been introduced. The AD603 is a single-channel amplifier with pin-programmable bandwidths of 9MHz and 51dB maximum gain or 90MHz with 31dB maximum gain.

The dual-channel AD604 with 40MHz bandwidth, is similar to the AD600, but has a programmable input amplifier to increase its maxi-

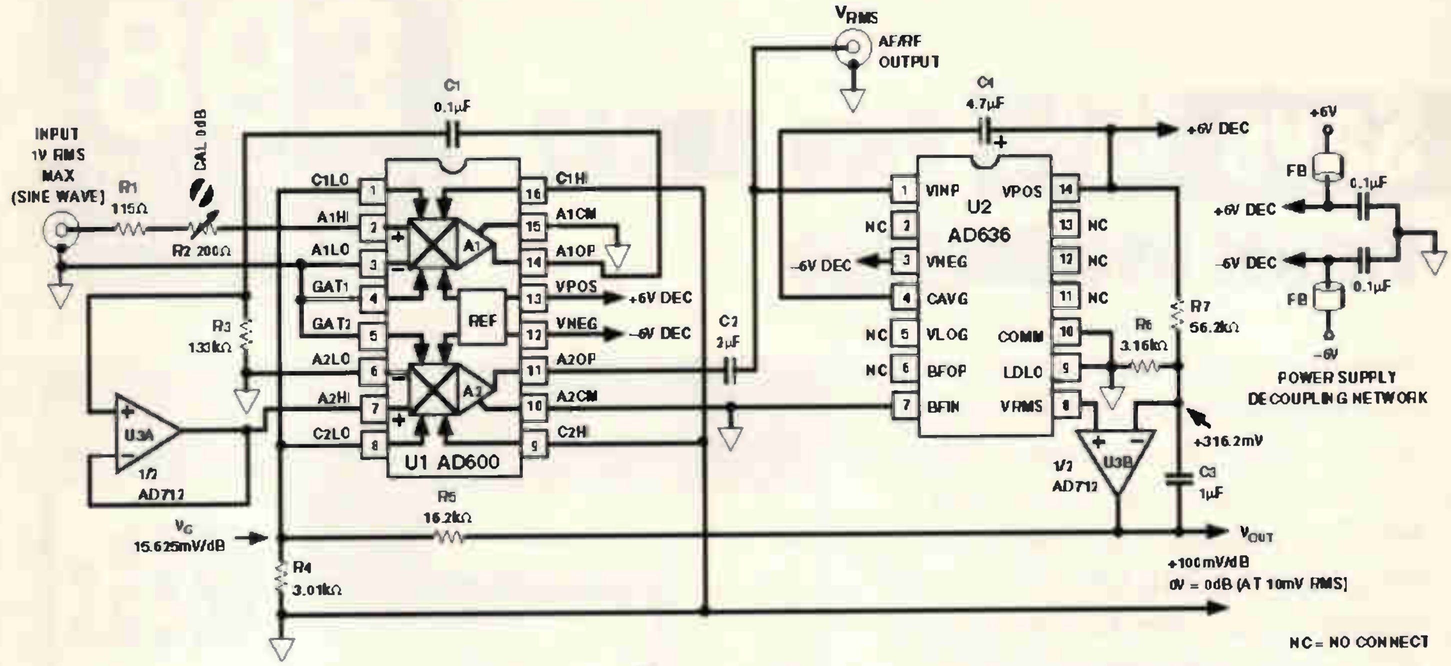
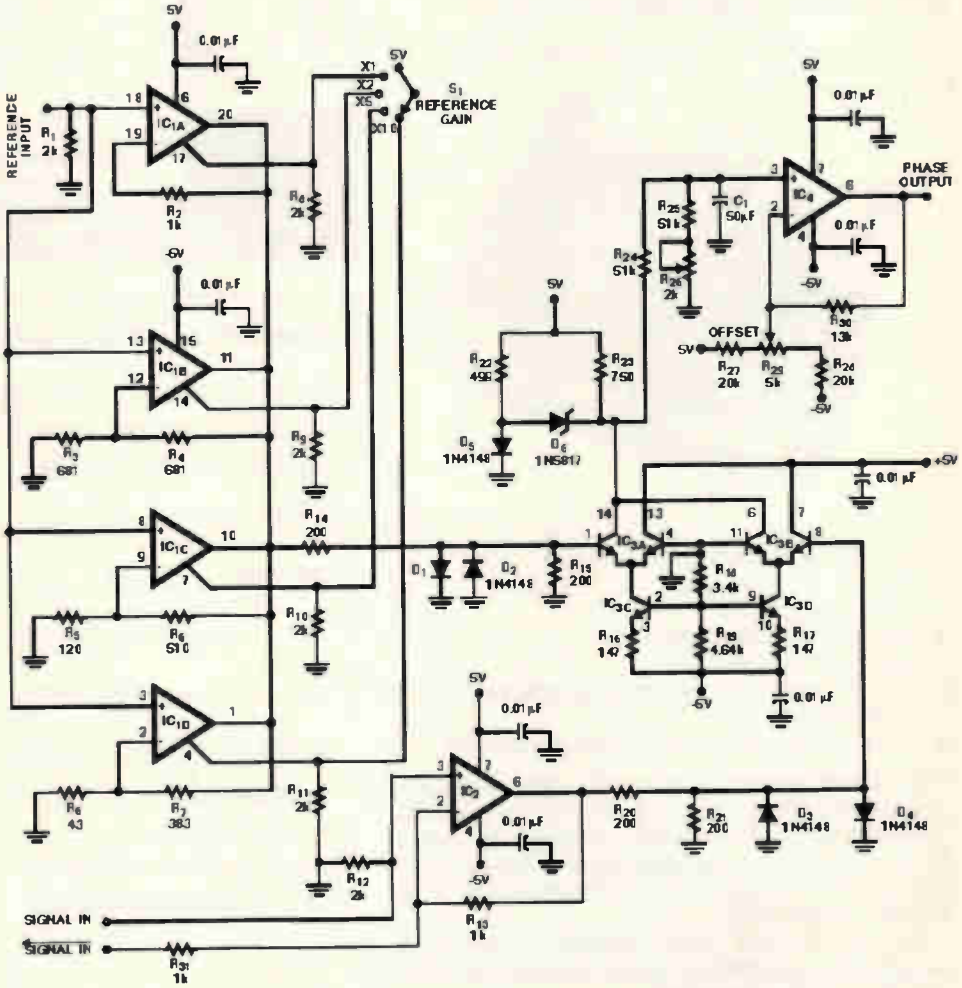


Fig. 7. Deceptively simple, this three-IC voltmeter measures true RMS of high crest factor waveforms from audio to 2MHz with an 80dB dynamic range. Claimed accuracy is ±0.5dB from 80μV to 500mV. Having built this circuit in 1993, I can confirm this performance is achievable.



mum gain to 54dB.

Data sheets for these parts all describe their use to build closed loop, wide bandwidth, AGC amplifiers with 80dB input range and quick response. Additionally, the sheet for the AD600/AD602 also describes a 120dB-range AGC amplifier and an 80dB 'linear-in-dB', 2MHz bandwidth, true RMS responding voltmeter, **Fig. 7**.

Constant propagation delay

For some applications, it may be necessary to vary an amplifier's gain while keeping its propagation time delay constant. With change in signal amplitude, the maximum signal slew rate, which occurs at the zero voltage crossing, varies. When the time difference of two signals having differing amplitudes must be measured, this difference in slew rate causes measurement errors.

One circuit application where this matters is measurement of phase difference. Application note AN9637 from Intersil[2] describes how the company's HA5024 amplifier can be used to adjust signal amplitude while minimising these errors.

The HA5024 is a quad amplifier, and each section can be enabled individually. This allows amplifier gain to be controlled by simple selection of amplifier, rather than switching resistor feedback networks. Each amplifier can be compensated for propagation delay, avoiding switching circuit parasitics.

The HA5020 incurs similar propagation delays. This is a single amplifier having identical schematic circuit and made using the same process as the HA5024. Note AN9637 describes the design of a phase meter using these techniques that is capable of 1% accuracy and usable to 10MHz, see **Fig. 8** ■

Fig. 8. Using the quad amplifier with the HA5020 single amplifier to roughly equalise reference and measure signal levels while maintaining similar propagation delays. This phase meter design can be downloaded as AN9637.

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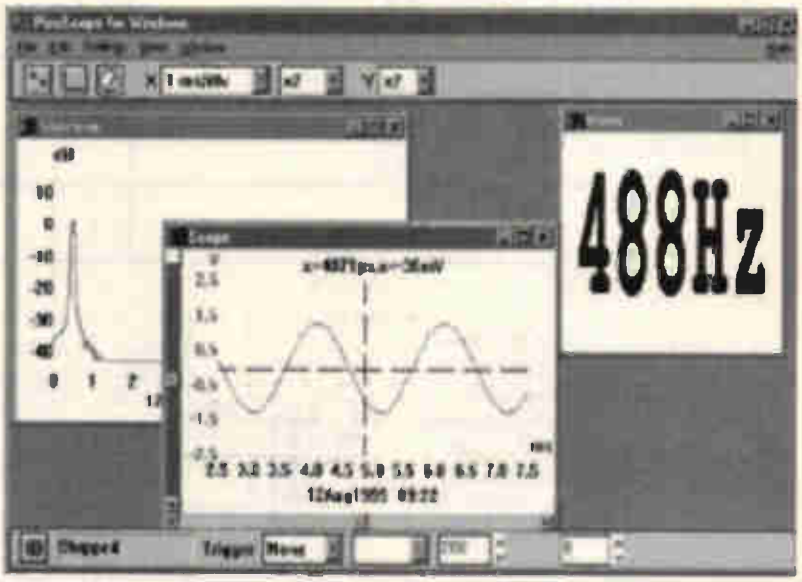

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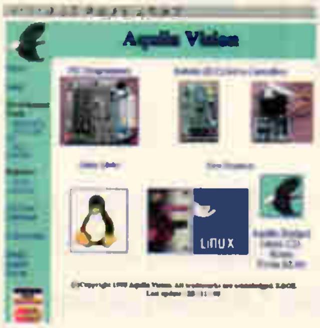

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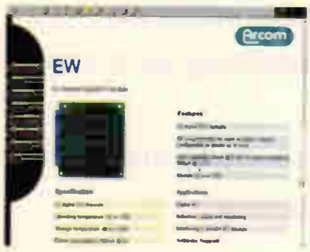

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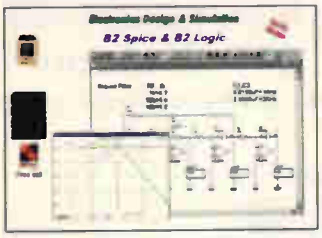

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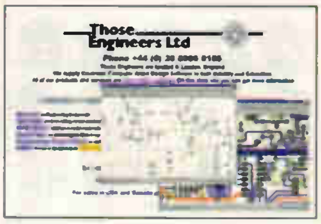

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Letters to the editor

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e-mail jackie.lowe@rbi.co.uk using subject heading 'Letters'.

Linear power supplies

Can anyone help me track down the best source of basic linear power supply design information ever produced.

Many years ago Mullard published the most wonderful design guide booklet on designing, calculating and specifying linear power supplies. Unfortunately I do not have a copy although I do know that parts of it were later reproduced somewhere in a National Semiconductor book.

I know that several historical publications of Mullard origin have been republished: has this happened to this guide? If so, where could I obtain it? If anybody has a copy, would copyright still exist? Or is a photocopy available anywhere?

Alternatively, is there another better or more complete guide to power supply design available without resorting to the fairly simple but very tedious calculations from first principles?

If anyone can point me in the right direction I would be glad to hear from them by e-mail.

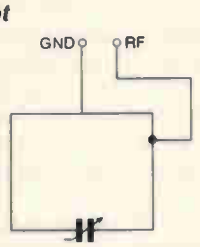
John Paulson
Warrington
Cheshire
john.paulson@aimarketing.co.uk

Radio control

As I am currently experimenting with radio control in the 4/8MHz band, may I say how helpful I found Pei An's article in the June issue.

However, with regard to the loop type aerial shown on page 450, I think it would be preferable to make the ground connection at the neutral point on the loop, i.e. the voltage node, and make a gamma match for the RF connection as shown.

J S Linfoot
Oxford



Teething trouble

I write in response to Ian Johnson's letter, published in the July issue, in which he expresses concerns about both the safety and the reliability of the Bluetooth system described in an article in the May issue.

Unfortunately, I missed this article. However, since I am currently involved in the development of a commercial Bluetooth product, I hope to be able to address some of those concerns.

While my area of expertise is software and digital electronics, my role in the project demands that I have an understanding of the RF operation of the product. What I present here is my own personal opinion, independent of my employer. I am not an expert, but I consider myself well informed on the subject.

First, I would dispute that 2.45GHz is "the very frequency most likely to cause personal injury". Although he doesn't explicitly state it, I assume Mr Johnson is referring to the fact that this frequency is employed in the operation of microwave ovens, since I cannot think of another argument to support this statement. However, I would be very interested to hear from someone more knowledgeable than me in this area.

It is my understanding that this frequency is chosen for microwave ovens not because the frequency itself provides maximum heating effect, but because it allows efficient generation of high output power levels by use of a magnetron.

While the BlueTooth specification does allow for 'high-power' transmitters with an output power of 100mW for certain applications, most devices with which the user will come into close physical contact will fall into the 'low-power' category and be capable of transmitting at a maximum power of 1mW, i.e. just over one millionth of the power output of a really good microwave oven.

Wien-wise

Further to Ian Hickman's discussion of the Wien oscillator in the June edition of *Electronics World*, readers may be interested in the visual analysis of the general Wien network that is summarised in **Figs 1** to 4. Voltages V_G and V_P are in phase when $\phi=\theta$, corresponding to $\omega=\omega_0$.

Bryan Hart
Leigh-on-Sea
Essex

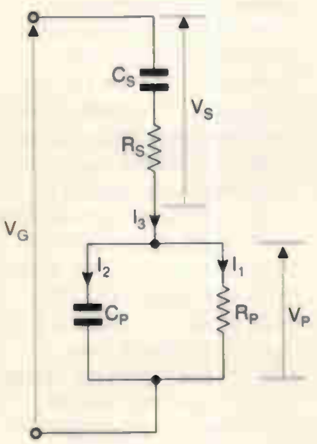


Fig. 1. General Wien network.

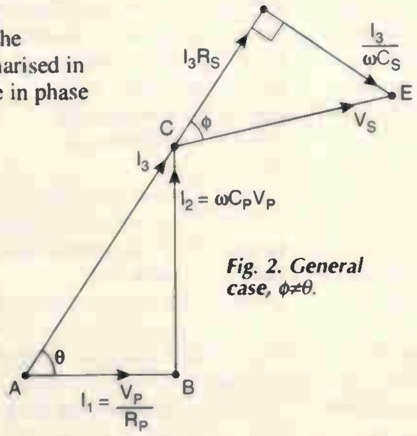


Fig. 2. General case, $\phi\neq\theta$.

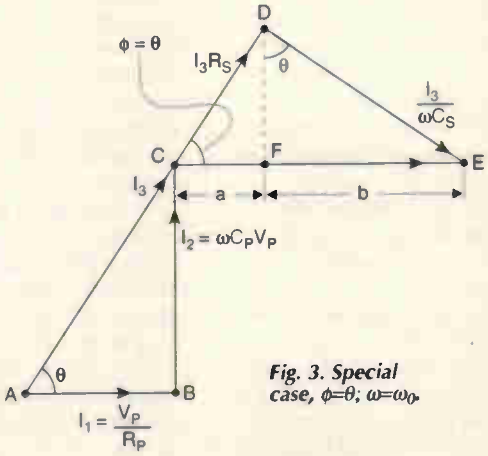


Fig. 3. Special case, $\phi=\theta$; $\omega=\omega_0$.

	Triangle				Equating marked columns
	CAB	ECD	DCF	EDF	
Tan θ	$\omega C_P R_P$	$\frac{1}{\omega C_S R_S}$			$\omega = \omega_0 = \frac{1}{\sqrt{C_P C_S R_P R_S}}$
Cos θ	$\frac{V_P}{R_P I_3}$		$\frac{a}{I_3 R_S}$		$a = \frac{V_P R_S}{R_P}$
Sin θ	$\frac{\omega C_P V_P}{I_3}$			$\frac{b\omega C_S}{I_3}$	$B = \frac{V_P C_P}{C_S}$

$$(a+b) = V_P\left[\frac{R_S}{R_P} + \frac{C_P}{C_S}\right] = V_S$$

$$\therefore V_G = V_P + V_S = V_P\left[1 + \frac{R_S}{R_P} + \frac{C_P}{C_S}\right]$$

Fig. 4. Visual analysis for $\phi=\theta$: ($\omega=\omega_0$).

Given that the alleged health risk associated with 1.8GHz phones, that operate closer to the 2.4GHz band than lower-frequency devices, is not yet so obvious as to be beyond doubt. Note that the output power of these phones is a thousand times that of most Bluetooth devices.

Secondly, "the people who set out the specification," were a consortium that included major mobile phone manufacturers Ericsson and Nokia. I cannot claim to know what was going on in their minds. But in my opinion, they did realise what they were doing.

Thirdly, regarding the style of antenna, I can only comment that the reference to Ericsson's Bluetooth development kit I have seen uses an etched 'F' antenna. Early product photographs released by vendors generally tend to show no apparent external antenna. This could, however, be cleverly disguised – a full wavelength at 2.4GHz is only 12.5cm.

In the case of Bluetooth-enabled phones, it is possible that the phone antenna doubles as the Bluetooth antenna, but I am not qualified to comment on the feasibility of this arrangement.

As for interference from microwave ovens, in a fast frequency-hopping system, it is quite possible for stray radiation from a microwave oven to interfere with one of the seventy-nine channels – or possibly two adjacent channels. The result would be a lost data packet, which would be re-transmitted on the next slot and at the next frequency in the pseudo-random hopping sequence.

Performance degradation due to this interference would occur, but would be slight. The main reason for frequency hopping is to avoid interference between Bluetooth devices and from the environment.

In the event that the microwave interference were strong enough to saturate the receiver such that all channels were blocked simultaneously, I would be more concerned about my health than about any resulting data loss.

I trust this addresses some concerns, but I'm sure it will raise others.

Bill Jones
Via e-mail

Amplifier input filtering

Writing to 'Letters' in the May 2000 issue, J.N. Ellis proposed that a slew-rate limiting input filter would lay the ghost of TID to rest by preventing an internal amplifier stage from over-slew switching. He suggests a series signal path resistor of 10kΩ shunted to ground by 220pF.

I agree that all good amplifiers might sound equally well with such an input filter, but the more professional ones would have their music signal degraded at input, and this in a way that cannot subsequently be corrected.

Together, 10kΩ+220pF form a low-pass filter that introduces 0.3% harmonic distortion at 2kHz, and at all volume levels. Is this the realism that we are attempting to achieve?

A drum-stick 'rim strike' is one of the most dynamic, high spectrum, musical sounds that challenges audio equipment. Frankly, 10kΩ+220pF would ruin its wavefront – even where an amplifier might well be capable of reproducing such fast waveforms at normal listening levels. It is similar to putting a plain 18μH choke in series with an 8Ω loudspeaker, which is audibly deleterious.

I agree that TID can be a problem, and the higher the output voltage swing the greater the problem can be. But I see the use of some of the better output devices or amplifier architectures as being the real answer.

Loudspeaker systems and output devices have much more of a bearing on an amplifier's internal behaviour than do the music waveforms that our amplifiers are expected to reproduce. Failure of an output stage to follow input signal waveform causes an error current to be generated which will drive the intermediate stages to greater currents, but not, in bipolar amplifiers, to slew to appreciably greater voltages. Indeed, when non-clipping hf audio coincidentally rides bass at moments when a dynamically reactive loudspeaker system impedance dips below its nominal value, the conducting output device becomes much harder and slower to turn 'off'. This happens when the negative feedback loop attempts correction. It does so by inducing counter conduction in the opposite output half. There is now a risk of running the amplifier into power sapping, fuse blowing or smoke releasing output stage cross conduction.

Mr Ellis mentions hard switching. A differential input pair can conduct input or negative feedback loop current through their bases but they cannot 'hard switch' like a switching transistor. This is because they never lose their high collector voltages and thus do not slew. Actually they recover extremely quickly – probably in nanoseconds.

Also, a driver transistor fed from an input mirror is easily turned on or off because its base is actively and conductively driven either on or off. Additionally, it usually has constant collector current sinking.

Drive overshoots that look like interstage 'hard switch' conduction are generally due to output device storage effects after these have been driven to maximum driver transistor current, or driver source current, to correct errors. They are not due to Miller effect at the driver transistor affecting that or earlier stages. You can check this by square wave monitoring an amplifier working into say a 100Ω load.

Nor am I sure that driver transistor base-collector capacitance is the most serious amplifier problem; some designers need to add an external fixed-value parallel capacitor at this same point to ensure overall stability at high output levels during the high collector voltage parts of a waveform. Here, driver transistor C_{bc} asymmetrically falls to too low a value.

It is true that the sort of transistors often used for a driver voltage amplification stage can have internal collector-base capacitance values that approach 100pF. But this is only at 0V. As most circuits normally operate above ten volts, their working values tend to be less than 20pF.

This would have a considerable bearing upon Mr Ellis's calculations – by a factor of five. The phase shift that his proposed filter is obliged to introduce could thus be reduced from fifteen degrees to just three degrees at 20kHz. This might be more widely acceptable as a domestic compromise, and could be achieved using 3.9kΩ+100pF. Don't forget though that the power amplifier itself then introduces its own negative feedback controlled phase distortion after this.

Most amplifiers already have a low value shunt capacitor between the base of the first transistor and ground. This component is a vital part of the negative feedback loop. However, it should not be so high in value that with the series input resistor, plus line driver impedance, it causes irreversible signal phase

Put it in the fridge then tap it...

Hot headed Conner drive in the July issue: put the drive in the refrigerator for an hour. Take it out, tap it lightly and re-install. This works most of the time for us.

Ed Dell
Via e-mail

This is not a hoax! I'd be interested to hear if anyone has an explanation for it. Ed.

New – Windows 2000 users' forum

As CAD is an important part of most electronic circuit designers' jobs, and in the light of Rod Cooper's recent review, we will be making space for any comments you have on CAD and operating system compatibility issues. All tips, hints and experiences will be considered, provided that they are relevant to circuit designers using CAD. Peripherals, back-up and printing problems are included.

distortion at audio frequencies.

I believe that the source should determine the response as much as is safely possible. I would not willingly start out with a specification worse than one degree at 20kHz, this being shared between an input filter and the power amplifier.

My view of the problems with asymmetrical, i.e. unbalanced, voltage amplifying driver stages is that, under negative feedback control, they are capable driving output devices into maximum conduction in one push-pull half much more quickly than are the resistor or semiconductor current sources feeding the opposing half. The situation is made much worse when the driver stage then attempts to turn them off again, for whichever was the faster 'on' subsequently becomes the slower 'off'.

The problem is thus much more severe than first apparent, or revealed when using test-bench loading resistors or computer simulation programs. Unless this is specifically catered for, the entire amplifier can become momentarily imbalanced. This creates a zero-referenced error and a transitory, phantom disturbance within the reproduced sound field.

The solution is some form of ensured current balancing or limiting circuitry. Alternatively a much more complex symmetrical arrangement is needed. This would be in the form of complementary first-stage input pairs. Each pair would have mirroring and feed separate positive and negative rail drivers that act together. This is carefully explained in an article by Mr Stochino.

Of course, we could simply use an electronic crossover to bi-amp carefully optimised amplifier designs, like those from Mr Self. These really can sound superb when the mid-treble amplifier is isolated from the transient blunting that is induced by current demanding electro-dynamic impedance reactions within bass loudspeaker-crossover systems. Also, the bass amplifier is then not expected to slew quickly anyhow.

I can confirm that when this kind of a system is driven loudly enough for the bass amplifier to start clipping cleanly, overall reproduction appears as little affected by the loss of bass power, as our brains 'continue to hear' clean mid and upper reproduction instead of being distracted by 'in your ear' distortion.

Surely the whole point is that an amplifier must faithfully follow its signal input, without first modifying it, i.e. distorting it, to ensure that it can subsequently cope without itself distorting?

Whether active or passive in origin, all distortion must be minimised to reproduce sound realistically, and that includes phase distortion!

Graham Maynard
Newtownabbey
Northern Ireland

Valve driver

I was pleased to see my valve amplifier drive circuit published under Circuit ideas in the July 2000 issue. However there are two errors: the collector voltages of Tr_1 and Tr_2 should be 135V not 13.5V. Also, my address is Chale Green not Castle Green.

Keith Cummins
Chale Green
Isle of Wight

A new IM distortion mechanism?

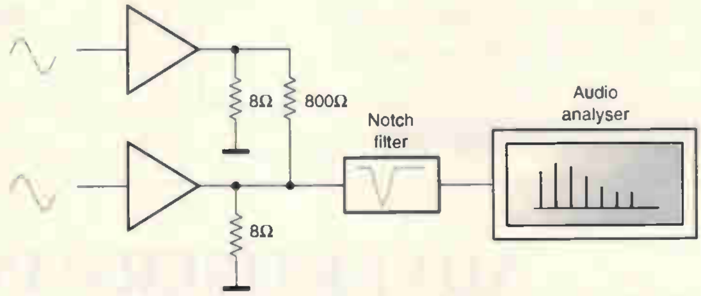
After reading the interesting article by Anthony New in the June issue, THD is meaningless, it occurred to me that another distortion mechanism may also be at work. As I have never heard it described before I have dubbed it LIIMD.

The voltage across a moving coil speaker has two components, resistive – which is linear – and back EMF, which is generated by the movement of the cone. This includes non linearities from the suspension and the cabinet and reverberant effects. This is a low impedance signal which is generally connected directly to the output stage of the amplifier, where it is free to generate intermodulation products with the input signal – especially at the crossover region where the open loop gain of the amplifier may fall, and the output stage is at its least linear.

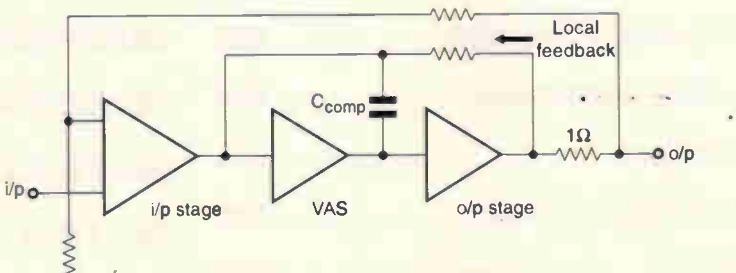
Due to the fact that I have no access to appropriate test equipment I have been unable to measure it. But if anyone is interested I have proposed a way of testing my hypothesis below.

The intermodulating signal is injected into the output of the amplifier under test by means of an attenuator to isolate the signal source. The intermodulation products could be seen on an audio analyser via a notch filter to remove one test signal so that intermodulation products are not generated in the front end of the analyser.

The two amplifiers used in the test should have separate PSUs to prevent intermodulation being generated in the power supply.



A means of measuring 'load-induced intermodulation distortion'.



Configuration of a modern amplifier.

Could the continual witterings of the hi-fi fraternity about speaker cables be related, as a speaker cable is only isolation between amplifier and speaker? That is to say that putting half an ohm between amplifier and speaker may suppress this mechanism.

If so a better solution would be to put the output stage of the amplifier in a local feedback loop and connect it to output global feedback point via a small resistor say an ohm. This is in a spirit resembling triode and ultra-linear valve amplifiers. In these, local feedback within the triodes, or pentodes with feedback taps on the output transformer, are used and the resistance is contributed by the transformer's windings. A more modern configuration is shown above.

Is this a known distortion mechanism? Am I just talking nonsense, or have I finally said something useful?

Jo Atkin
London

As you might have guessed, there's quite a number of letters relating to Anthony New's article on intermodulation distortion. I will try to run them all together in the next issue. Ed.

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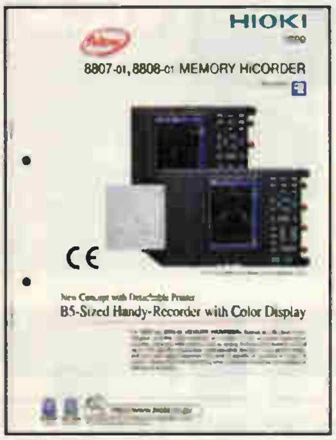

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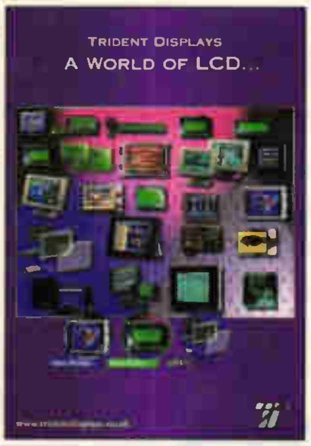

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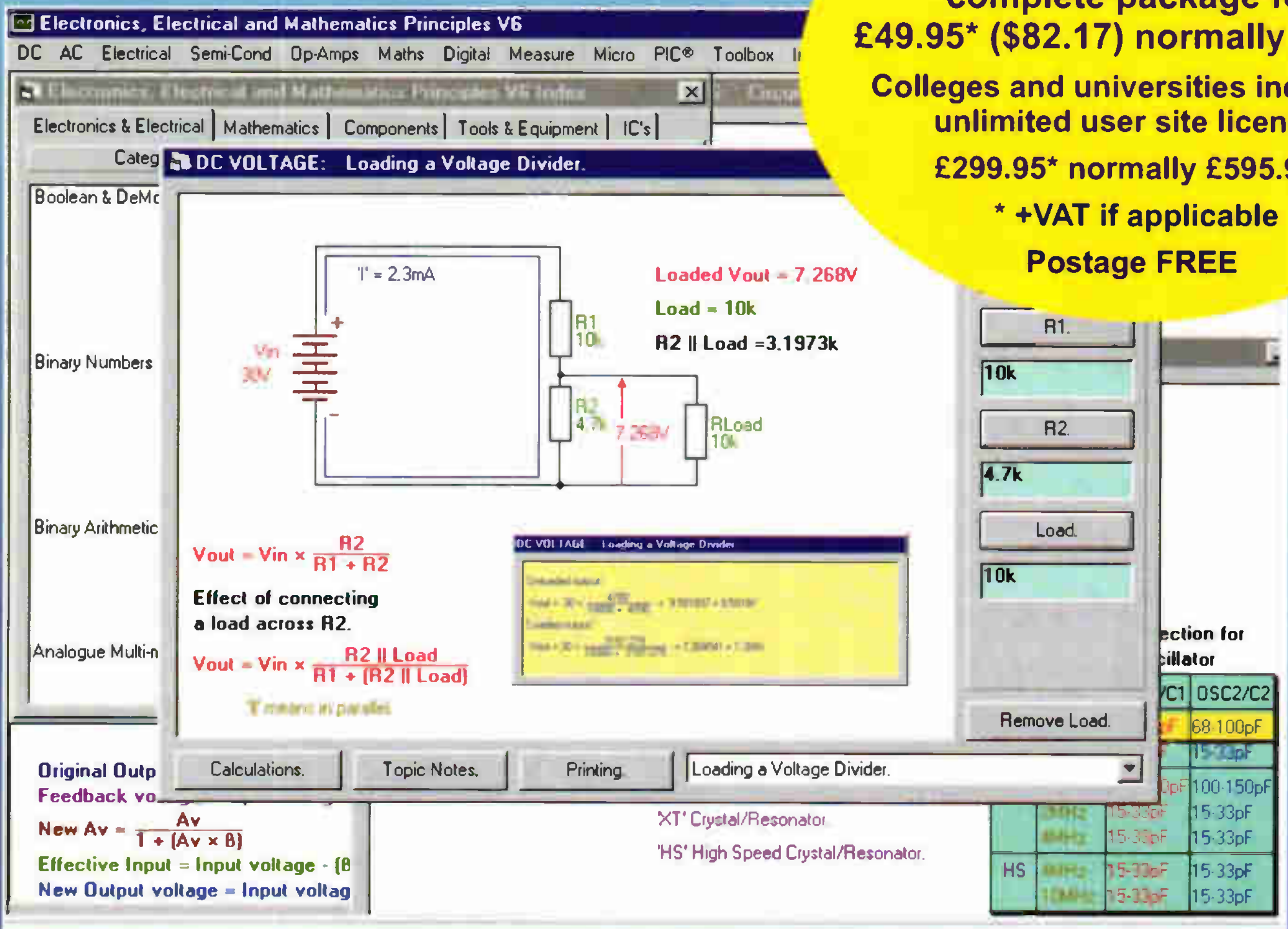

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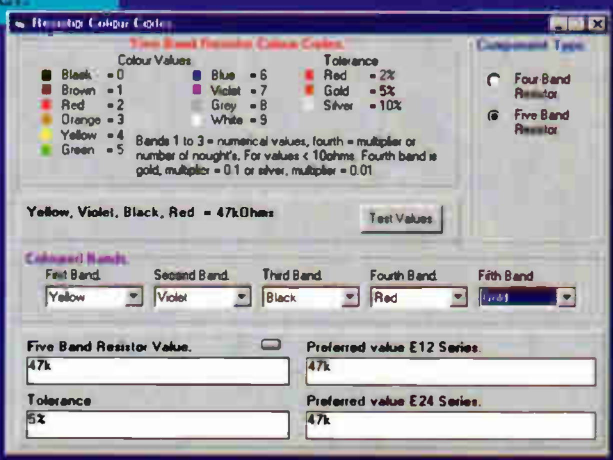

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