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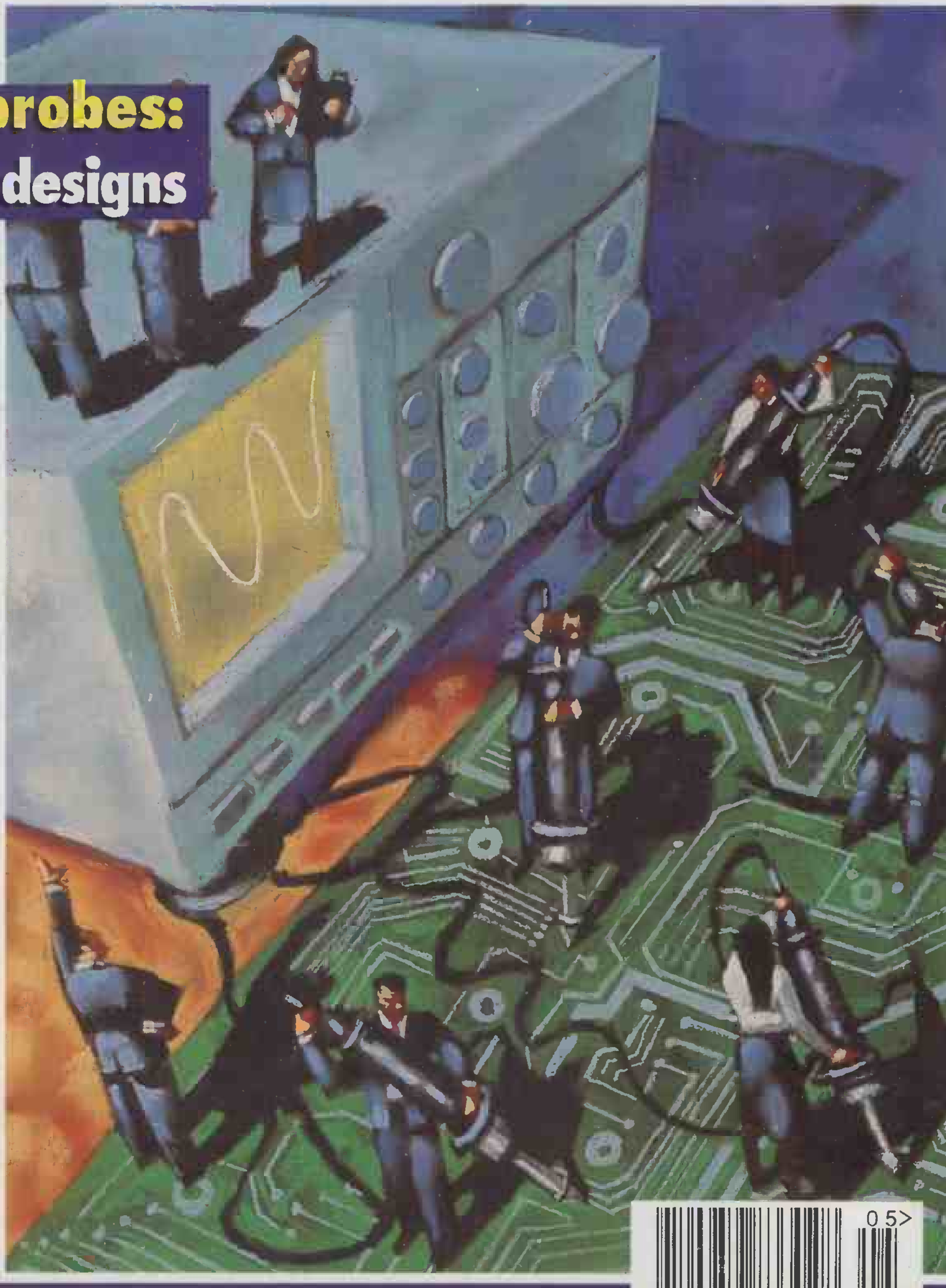

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FREE to EW readers: ultra-fast buffer

Maxim is offering a free high-performance buffer IC to the first 1000 EW readers sending in the reply card, page 372.

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This feedback temperature-controlled soldering station is being offered to EW readers at a special discount price of just £55 – fully inclusive – page 373.

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What a tangled web we weave...

Last year was the year of the Internet – the year browser specialist Netscape was valued at a couple of billion dollars; when the number of Internet users supposedly topped forty million; when companies overdosed on having 'Web Sites' and when the idea of the 'Internet Computer' was born.

1995 was also the fiftieth anniversary of the publication of George Orwell's novel 1984'. Remember Winston Smith's 'telescreen'? He couldn't turn it off. He couldn't cover it up. It could see everything Smith was doing even while 'Big Brother' exhorted him to embrace some nonsense new enthusiasm. There was no escape from the screen's surveillance or its propaganda. It was a nightmare.

Now, fifty years later, the cable and tv companies are gearing us up for 'interactive TV'. We should be so lucky.

Some technology products tend to be personally liberating. Some add to the powers of 'them' – those who would dominate us: governments, companies, cults and the like.

Can a technology box be 'good' or 'bad'? Which is good or bad out of mainframes, minis, televisions, pcs, telephones, faxes, set-top boxes, Internet Computers, personal communicators and PDAs?

Some boxes favour the big battalions. Mainframes and minis are affordable only by organisations. Therefore they are commonly used to replace, monitor, keep records on, or otherwise control, people.

Other boxes, such as pcs and tvs, are personal items, controlled by people for their own pleasure or empowerment.

Is it fanciful to suggest that the pc is a democratising influence spreading power among the many, while the mainframe is an autocratising influence adding to the powers of the few?

In this respect there's something Big Brotherly about the concept of Internet Computing. The idea was proposed last year by Larry Ellison, chairman of the database company Oracle.

Eliison suggested that the future of personal computing is not via the pc but via dumb terminals connected by a the telephone line to a mainframe or server which provides the terminal with its processing power, programmes and storage.

It's a concept diametrically opposed to that of the personal computer, with its key attribute of local

. . . governments around the world are already making noises about controlling the Internet. . .

control – having its own processing power, containing its own programmes and storing its own data. And a pc still works fine whether or not its connected up to anything, or anyone, else.

Of course the access services insist that the Internet is a globally-flung network of distributed computers connected by randomly routed telephone links. But that does not stop it from being controllable. They tap our telephone lines with abandon.

Indeed governments around the world are already making noises about controlling the Internet. Usually they say they are concerned about controlling pornography, though surely only the most dedicated pervert is prepared to sit around waiting for dirty pictures to be downloaded at 28.8 kilobits a second. It's more likely that the pornography issue is being used as an acceptable way to start controlling the Internet. And when that happens how secure can your data be when stored on a remote database?

How much can be found out about you by knowing which programmes you download? Or which Web Sites you visit? Or with whom you exchange e-mail?

Is Internet Computing a better way forward for personal computing than the PC? Or is it an open invitation into our lives addressed to Big Brother?

David Manners

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REED BUSINESS PUBLISHING

EC recommends 'bit-tax' on Internet data

A 'bit-tax' on information sent over the Internet has been recommended in a report for the European Commission.

The motivation for the report is that insufficient revenue from taxation methods such as VAT is predicted in future as data sent by conventional means diminishes.

Chairman of the EC's study group, Luc Soete, said: "A larger share of our production and economic activity is focused on information and communication. We must make sure we have a national tax base which includes these activities."

Soete thinks a bit-tax would eliminate the problem of offshore tax havens – companies based outside the EU do not have to pay VAT.

David Barrett, head of corporate communications at Pipex Ltd, a UK service provider, said: "With around ten million hosts on the Internet, where do you put the counters? How much is data worth? Taxation is a good idea but it has to be dealt with carefully. We would welcome a move whereby the politicians in Brussels talk with people such as the service providers."

Insect robots in space

Tiny robots modelled on insects the size of 50p pieces, called pixelsats, are being tested by NASA. Developed by Mark Tilden, the robots could be used in 'swarms' to carry out complex tasks in space at a fraction of the cost and risk of a conventional satellite.

Modelled on insects, the robots use at most 12 transistors and hence are dedicated to a particular task such as part of measuring arrays or passing data between larger satellites.

Thousands of the pixelsats would be deployed in space. Even if many are destroyed, enough would remain to carry out the allotted tasks.

New mobile phone antenna may reduce brain radiation

Researchers from the University of Stuttgart have designed a mobile phone aerial that could reduce the amount of transmitter energy that is radiated into the user's head.

A paper by H Ruoss and F Landstorfer in *Electronic Letters* describes a double-T slot antenna, resonant at 1.8GHz. This has been modelled and is predicted to dissipate only 5.7 percent of output power into the head. With monopole antennas the head absorbs around 30 percent.

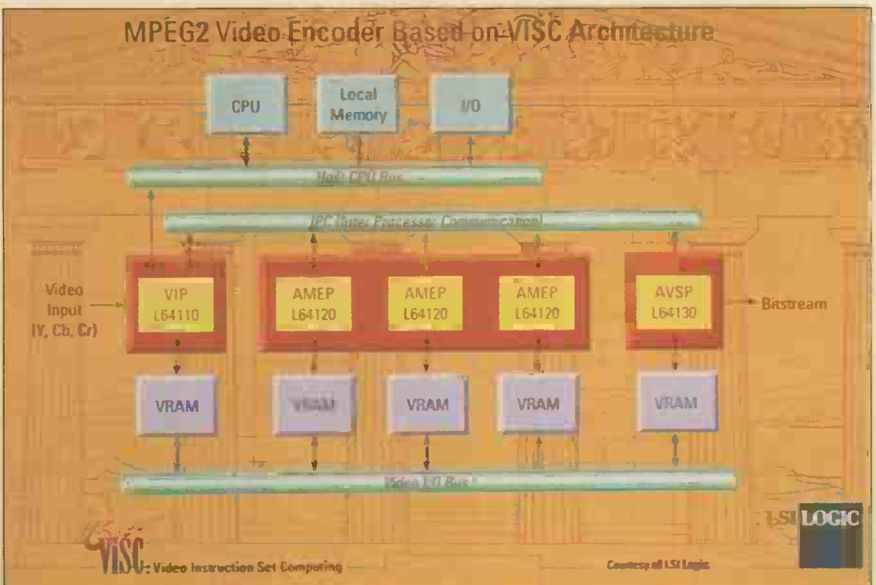
In the model, the head is represented by a lossy dielectric sphere and the holding hand by a rectangular solid.

New architecture looks set to slash MPEG encoding costs

LSI Logic has developed an encoding architecture which aims to reduce cost for real time video encoding by a factor of five.

Called Video Instruction Set Computing, or VISC, the architecture is designed to address a range of applications including real time encoding for cable transmission and direct broadcast by satellite and multimedia.

Jean-Luc Droitcourt, marketing director for the consumer segment, Europe said: "Today real time encoding for MPEG-2 costs $100,000. With the VISC devices, this is reduced to $20,000."



Using as few as 3 chips, Video Instruction Set Computing – VISC – is likely to reduce the cost for real time video encoding by a factor of five.

VISC can execute MPEG-1, MPEG-2 for video encoding, DigiCipherII for broadcast quality video and H.261 video conferencing encoding, using between three and five chips. An industry source commented that the expense of the VISC chipset – in thousands of dollars – would limit its use to broadcasting applications. For MPEG-1 encoding and video conferencing in pcs, for example, it would not be commercially viable.

"We were able to integrate all the key functions of encoding into only three complex microchips by using LSI Logic's CoreWare methodology," said Brian Halla, executive VP of the company's products group. "This involves combining high-level, pre-tested building blocks together to form highly integrated devices. Each chip in the VISC chipset carries out its own dedicated function to efficiently process video data."

Each of the three devices in the chipset incorporates a Mips Risc core running at 40Mips. The VxWorks real time operating system is used, with the software written in C or C++.

Manufactured in a 0.5µm process, the chips contain two million transistors. LSI is predicting a single chip implementation of the VISC architecture by 1998 using 0.3µm technology. **Richard Ball,** HART

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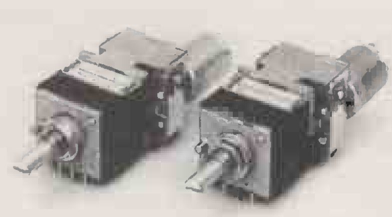

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CIRCLE NO. 109 ON REPLY CARD

All-plastic lcds are a step nearer

Workers from the UK and Brunei have developed a thin film carbon transistor technology which could make possible the deposition of semiconductors directly onto plastic substrates.

The incompatibility between silicon processing and plastic is the biggest obstacle to making all-plastic lcds. Current silicon semiconductors require processing at around 250°C minimum. This is too high for plastics which would otherwise make good low cost substrates.

All, but one, of the stages required to produce the new transistor are performed at less than 100°C. Dr Bill Milne, a University of Cambridge based member of the team, said: "Depositing the silicon nitride gate insulator is the only high temperature part of the process, and we are working on that.

The semiconductor used is tetrahedrally bonded amorphous carbon (ta-C). This is a form of carbon that has a high proportion of diamond-like (sp^3) bonding and whose conductivity is readily affected by external electric fields.

Other groups have suggested that this material type is unsuitable for thin-film transistors. The reasoning is that a more conductive graphite (sp^2) bonded layer forms on the surface, excluding the controlling gate field from the channel.

The geometry of the experimental device is such that the (sp^2) layer can be etched away and prevented from reforming using the pacifying nitride layer.

To make it a semiconductor, the carbon is p-doped with boron – the easiest way to dope carbon based devices.

Milne said: "It is the first of its kind and it isn't a great transistor at the moment. We have a lot of different ways to improve it though, including our nitrogen based n-type doping process.

Steve Bush, *EW*

Twisted pairs – to screen or not?

Cable system suppliers disagree over the relative emc performance of screened and unscreened twisted pair cabling. The situation is not helped by the lack of independent standards guidelines.

Last month's meeting of the ISO's cabling standards committee failed to resolve the situation, which calls into question the emc suitability of unscreened twisted pair cabling for 100Mbit/s high speed lans.

"This issue must be sorted out at a standards levels, but I am concerned about a tendency to knock unscreened twisted pair at the standards level," said Arthur Green, marketing manager for Nortel Cable Networks.

Nortel presented to the ISO, the results of tests which it believes demonstrates that category 5 unscreened twisted pair-based lans can meet the EMC Directive emissions rules – even at 155Mbit/s ATM data rates.

The suitability of unscreened twisted pair for data rates above 100Mbit/s, which put high frequency signals on the cable, has been questioned by European cable system supplier Alcatel Cable Systems. "I am not saying unshielded cat. 5 is a bad product, I sell it, but it has its limits and in terms of emc that limit is 30MHz," said Gunther Gubbelmans, Alcatel's business development manager. Alcatel maintains that data rates which put frequencies higher than 30MHz on the unscreened twisted pair could fail the Directive.

The lack of an ISO ruling means that the inevitable conflict between opposing commercial interests makes it difficult for users to obtain an independent assessment of the cabling situation.

Pressure sensors for car tyres

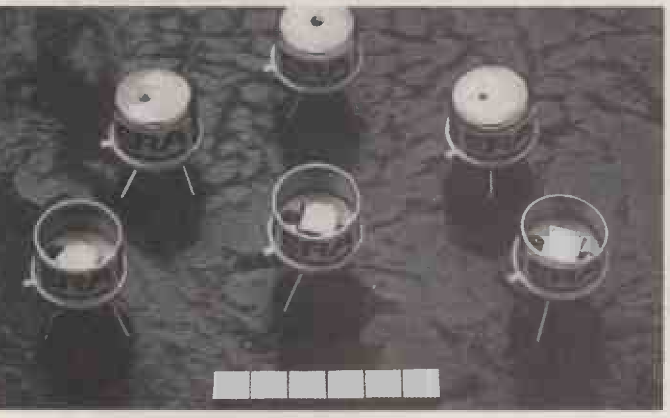
A remote pressure sensing system has been developed by Surrey-based ERA Technology in conjunction with Otter Controls in Derbyshire.

ERA believes the battery powered sensor microsystem has possibilities in remote monitoring applications such as tyre pressures. Both companies are looking for discussions with end-users on potential applications.

The system consists of a capacitive pressure sensor, interface electronics and a radio transmitter, all mounted onto a single substrate. It is available as a component or as a stand-alone system.

The capacitive sensor, designed by ERA, is claimed to offer advantages over silicon piezo-resistive technology. These include low power, temperature operation above 125°C, lower intrinsic temperature coefficient, higher stability and over pressure capability.

Checking tyre pressures and/or temperatures would be accomplished by installing a sensor in each tyre and a receiver in the dashboard. This would hopefully reduce the risk of blow-outs and improve fuel economy.



Capacitive pressure sensors like these could be used to indicate tyre pressure on a car's dashboard.

'Digital paper' for MIT's electronic book

Researchers at the Massachusetts Institute of Technology (MIT) are developing digital paper for use in an electronic book.

Project leader, Dr Joseph Jacobson says the book will have the feel and weight of a few hundred pages-sized hardback, with each page being reconfigurable. Data could be downloaded from a database on the Internet and pages displayed by pressing buttons on the spine.

"We are probably looking at about two years until we have a prototype, but an important point is that every process we are developing is scalable," said Jacobson.

The enabling technology is digital ink particles only 50μm in diameter that are black on one side and white on the other. Similar to toner in laser printers, the particles adhere to a paper-like synthetic surface. When subject to an electronic charge, the particles flip showing either white or black. So far, the researchers have managed to flip particles but have yet to form words. ■

RESEARCH NOTES

Jonathan Campbell

Transparent motives for hdtv

Development of organic light-emitting devices that have transparent contacts is being heralded as a leap forward in the design of display devices – with implications for everything from high definition televisions to head-up display units.

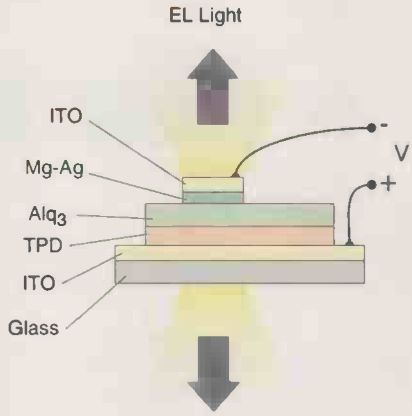
Developed by researchers at Princeton University and the University of Southern California, the new class of devices is said to be 70% transparent when turned off. While turned on, they emit light from both top and bottom surfaces with 75% quantum efficiency. Because the devices themselves are transparent, red, blue and green ones could be stacked on top of each other, giving a simple way to fabricate a very high quality screen.

The devices are grown on a conducting substrate precoated with a transparent indium tin oxide (ito) thin film. Then a thicker layer of TPD – a 'hole' conducting compound – is deposited, followed by a thick layer of the electron conducting and highly-electroluminescent organic compound tris (8-hydroxyquinoline) aluminum (Alq3).

The action that creates colour takes place in the electron conducting layer (Alq3 in this case emits green light). An electron injected from the contact onto the molecules in the layer is attracted by the positive hole and gives off a photon with the frequency of either green, red or blue light. What determines the various colours is the molecular structure of the light emitting material – the greater the separation of electron and hole in energy, the more energetic the photon released and the bluer the light.

On top of the Alq3 layer goes the top contact – the most innovative component of the device built by the Princeton and USC scientists because it is transparent. This electron-injecting contact is made by depositing, through a shadow mask, a thin layer of magnesium-silver alloy (MgAg), onto which is sputter-deposited a thick ito layer.

Stephen Forrest, director of Princeton's Advanced Technology Center for Photonics and optoelectronic materials (Poem) explains the attraction of toleds: "What we can now do with these organic devices is to place the three primary colour emitters in a single, very small stacked structure. By having intervening transparent contacts, we can energise to different extents the red, green and blue devices all in a single stack. The light penetrates through all transparent contacts and other organic layers and out comes a mixture of colours."



Five films are laid down on glass to construct a toled. Stack a red, green and blue toled together and you have the basis of a high definition colour display.

The making of a single pixel with three colours could have profound implications for manufacturing because only one type of pixel need be made, instead of three – thereby reducing three separate fabrication steps to one.

Could wireless video link cut the need for cable?

The head-long rush towards bringing video, voice, data and interactive services into every house is still being held back by one fundamental limitation – the absence of an installed high-capacity broad-band home link.

But two researchers at the University of California at Berkeley have announced successful testing of a system that at a stroke removes some of the drawbacks of domestic broad-band networks. J Park and KY Lau have developed a millimetre-wave fibre-wireless transmission broad-band system that could be quickly installed while promising a cost-effective method of delivering services.

Up to now we have tended to think in terms of using fibre links to a distribution node, then perhaps coaxial cable and amplifiers to carry signals into the home.

Park and Lau's system ("Millimetre-wave 39GHz fibre-wireless transmission of brad band multi-channel compressed digital video", J PArk and KY Lau, *Electronics Letters*, Vol 32, No 5, pp. 474-476) uses a wireless system for this last stage, retaining the optical fibre links for distribution of signals to remote antennas.

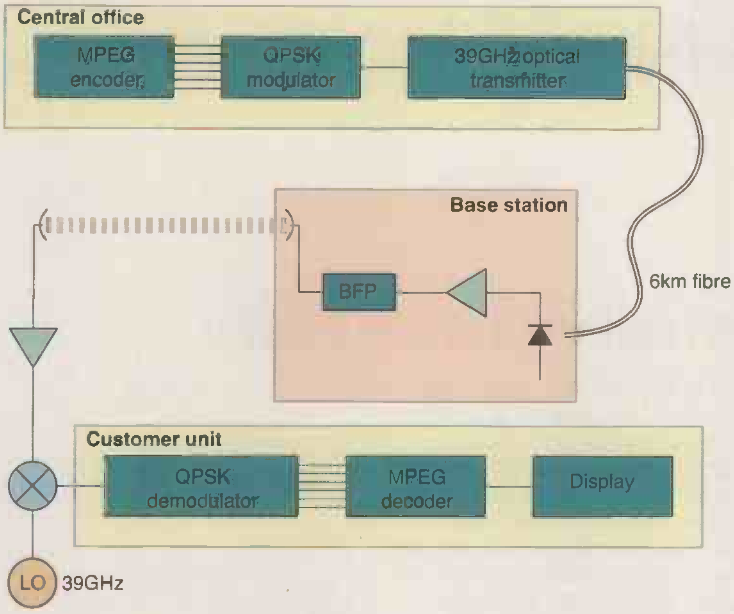
As in the conventional approach, the wireless system begins with a central office or head-end set up for each 300,000 users. From here, five to 15 fibre links connect to distribution hubs, each servicing 20,000 users. Each distribution hub uses 10-40 fibre links to distribute signals over the last 15km to fibre nodes. A fibre node would be needed for every 500 to 2000 customers.

It is from this last link that Park and Lau's system begins to make a difference. Installing a wireless connection to each consumer would be much quicker and less expensive for the service provider to install – especially in urban areas and regions over difficult terrain. A wireless system also accommodates a degree of mobility.

Crowding at lower frequencies, mean that only the millimetre-wave frequency range really offers the band width required for free space transmission of broad band spectra between 1 and 2GHz.

So far in tests, a broad-band

Bringing the information superhighway into the home without fibre links?



millimetre-wave optical transmitter has been used to transmit multi-channel digitally-compressed (MPEG-2) video over a 39GHz millimetre-wave fibre-wireless link.

The complete set-up used to test the system simulated both the fibre link and the wireless connection. To demonstrate fibre distribution of the mm-wave signals, 6km of single-mode fibre was used between the optical transmitter and the base station. The base station itself consisted of a high-speed photodiode, a 39GHz band pass filter, a high-power mm-wave amplifier and an antenna. A 500MHz-wide broad-band spectrum of channels, centred at 39GHz, was amplified to 5dBm and transmitted through an antenna.

Results showed a wireless link loss of 55dB, which at this frequency corresponds to a free space propagation path of over 1km when high gain (35dBi) transmit and receive antennas are used.

At the receiver end, the 39GHz signal was down-converted back to the original intermediate frequencies of 300-800MHz.

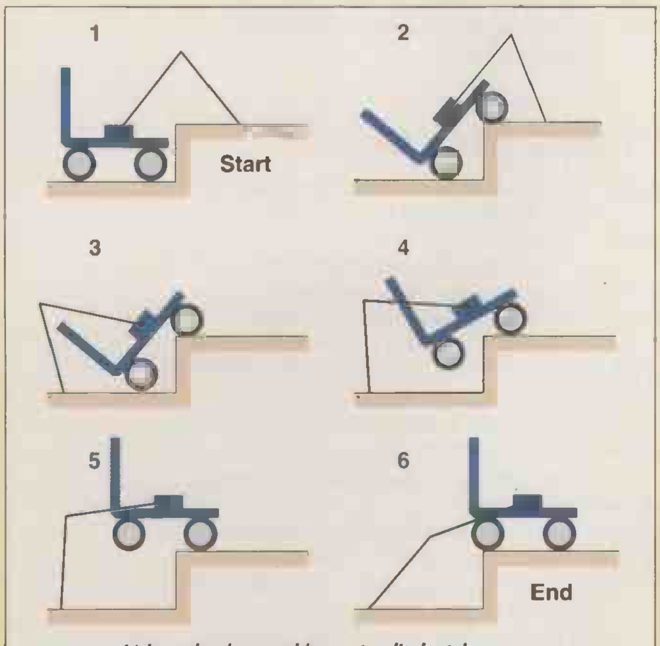
The result, according to the researchers, was that 70 digital video channels were observed using the optical transmitter, and that the video was good, with no decoding errors in any of the video channels.

More information from the Department of EECS, University of California at Berkeley, Berkeley, CA 94720, USA.

Wheelchair users take a step up

Stairs. Most of us take them for granted, stepping up or down, without thinking. For wheelchair users, life is not so simple. Steps can become major barriers, and shopping, using public transport or simply crossing the road can turn into an obstacle course – unless your wheel chair is able to walk up and down steps too.

This is the aim of a 'wheelchair with legs' currently being designed by a US team from the University of Pennsylvania and the AI du Pont Institute.



Using wheels – and legs – to climb stairs.

So far, a prototype chair with wheels and legs and has successfully negotiated uneven terrain and circumvented obstacles. One of the legs can also be used as a manipulator to perform simple tasks such as reaching for objects or pushing open doors.

Starting point for the project is that the wheelchair must exploit the capabilities of the human operator and must be safe. The eventual design ('Design of a Wheelchair with legs for people with motor disabilities', P Wellman *et al*, IEEE Transactions on Rehabilitation Engineering, Vol 3, No 4, pp. 343-353) was for a hybrid wheeled/legged chair with four wheels (two powered) and two legs. This was felt to give the best compromise between capability, cost and consumer acceptance.

A manoeuvre similar to walking is accomplished using the legs to drag the vehicle forward or backward, and the wheelchair (and operator) can climb a 300mm high step while still being able to pass through a 760mm doorway.

The two motors used to move the arms are driven by 20kHz pwm switching amplifiers that are configured to clamp to the motor current – determined by the control signals received from the IBM 486 control computer. System feedback is accomplished through incremental optical encoders which give the position of the legs and also strain gauges that indicate the forces at the feet.

Up to now the wheelchair has successfully completed a range of tests, with a 75kg rider, and the researchers are hoping to develop a modular system that can be bolted onto a conventional wheelchair.

Despite any technology advances, one problem still remains: that of acceptability. Wheelchair users are already angry that people too often only see the chair, not the individual. Unfortunately, that situation is unlikely to be improved by bolting on even more hardware and electronics.

More information from Parris Wellman, Department of Mechanical Engineering, Pennsylvania University, Pennsylvania, Philadelphia, PA 19014, USA.

Now you see it now you don't: glancing into the past

A group of astronomers has identified the most distant population of normal galaxies yet found. Using a new technique designed to isolate large numbers of extremely distant, young galaxies, the scientists have discovered what are very likely the progenitors of the bright galaxies – spirals and ellipticals – seen today. They observed the galaxies at a time very soon after they first formed, roughly 10 billion years ago. These objects show that galaxies were already forming in large numbers at an epoch when the universe was only 10 to 20% of its current age.

In the past, astronomers have had difficulty finding young galaxies. They are very faint and no-one was sure exactly what to look for. The new method involves taking images of the sky using three custom-made colour filters, allowing light of only red, green, or uv wavelengths to be seen. Young galaxies have a strong blue or uv tint, but when they are very distant, the uv wavelengths are strongly absorbed by hydrogen atoms both in the galaxy itself and in any gas that might be present between the galaxy and us. If a galaxy is within a particular range of high red-shifts (corresponding to large distances from us), its uv light will be completely absorbed by the intervening hydrogen. By screening digital images of the sky through these filters, and watching for objects present in both red and green but vanishing in uv, the astronomers have located many objects that are likely to be distant galaxies.

The picture shows a small portion of three images of the same piece of sky, taken with the 200-inch Hale Telescope at the Palomar Observatory in California.

The object at the centre of the circle is clearly present in both the red and the green image, but disappears in the uv image. Typical full-size images, of which these are small portions, will contain 30 to 40 of these objects that are likely to be extremely distant galaxies. Photo: Chuck Steidel, Caltech

For more information contact Chuck Steidel, assistant professor of astronomy at Caltech, California.

Agents of progress – or doom?

What connects World Cup soccer with a simulated war in Europe. If you thought the answer was 'football fan' you'd be wrong. The real solution is that they are both scenarios used for testing out a new generation of computer 'agents' designed to interpret and use human strategies to achieve their targets.

Some of the latest work is being carried out at the University of Southern California's School of Engineering's Information Sciences Institute (ISI), where agents have been created that can match wits with top human jet-fighter pilots in simulated dog-fights conducted in virtual computer environments.

The aim of the project, called Soar and funded by the US government's Advanced Research Projects Agency, is to develop what project leader Paul S. Rosenbloom calls "a basic architecture for intelligent systems."

The project has already explored military modes of decision-making (for example in a 1994 Simulated Theater of War – Europe war game) and is now being extended into entertainment.

Next year, a team of silicon soccer players will compete in a virtual tournament, RoboCup '97, to be held in Japan.

In the war game, the pilot agents were able to post some victories, particularly in multi-plane environments, where human pilots could be more easily distracted.

This is notable because modern high-performance aircraft combat – conducted at long distance by missiles and sensors – is more than a simple test of reflexes, a task at which the computer might be expected to excel. Rather, it is a chess-like game of cat-and-mouse, in which success depends on thinking through conflicting and ambiguous clues and deciding – in time – what an adversary is doing, and how best to counter it.

The agents are quite different to arcade game creations which have a very limited repertoire of heavily scripted behaviour, and little or no adaptability. They can behave in a more autonomous, more complicated way. In conflict, agents watch the behaviour of their adversary, attempting to understand it the way a human would – as actions aimed at accomplishing a goal. Similarly, they themselves act in order to accomplish a mission with specific goals.

Soar helicopter-pilot agents engage enemy fighting vehicles The agents must follow terrain not just to avoid crashing, but for tactical purposes, and must find areas where they can hide from enemy view and pop up at the right moment to attack. Programming covers not just flying, but how to attack, how to choose weapons, how to fly in formation, how to break formation, when and how to go to a firing position, all of which have to be dealt with in real time.

Creation of such agents is only now becoming possible because the amount of computation necessary is formidable, and programs must run fast enough to mimic the speed of human thought. Even the best decision is worthless if it comes too late.

Simulating team behaviour is another challenge, and each agent has to ask itself what its role is in the group, what commitments does it have to the group, even the occasional dysfunctional behaviour of anarchist agents must be possible.

At present, the agent fighter pilots are being tested in war games that link artificial-intelligence laboratories with military computers around the world. Rosenbloom and his team members say they look forward to a full schedule next year, with a major test scheduled for 1997. ■

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Traditionally, active oscilloscope probes have been very expensive. But today's high-speed op-amps make it possible to extend the measurement range of an oscilloscope significantly – and at very low cost.
Ian Hickman explains.

PROBES GO *active*

Theoretically, an oscilloscope shows what is actually going on in a circuit. But this assumes that connecting the oscilloscope to a circuit node does not change the waveform at that node.

To minimise loading effects, oscilloscopes are designed with a high input impedance. The standard value is 1MΩ, in parallel with some capacitance, which is usually about 20-30pF.

As far as the power engineer working at mains frequency is concerned, this high value is insignificant. Generally, the same goes for the audio engineer. One exception is when examining the early stages of an amplifier, where quite high impedance nodes may be encountered. But the oscilloscope's high input impedance exists at its input socket, to which the circuit of interest must be connected. So some sort of lead is needed.

Between circuit and oscilloscope

Connecting a circuit to an oscilloscope with leads of near zero length is difficult and tedious – and often impossible. Sizeable low frequency signals from a low impedance source present no difficulty; any old bit of bell flex will do. In most other cases a screened lead will be needed, to avoid pick-up of hum or other extraneous signals.

A screened lead of about a metre or a metre and a half proves to be convenient, and such a lead would add somewhere between 60 and 150pF of capacitance to that at the scope's input socket. But the reactance of just 100pF at even a modest frequency such as 1MHz is as low as 1600Ω – a far cry from 1MΩ and not generally negligible by any stretch of the imagination.

The usual solution to this problem is the 10:1 passive divider probe. This provides at its tip a resistance of 10MΩ in parallel with a capacitance of around 10pF. This is not ideal, but a big improvement over a screened lead, at least as far as input impedance is concerned. But the price paid for this improvement is a heavy one. Sensitivity of the oscilloscope is effectively reduced by a factor of ten.

Passive divider probes

Figure 1a) shows the circuit of the traditional 10:1 divider oscilloscope probe, where C_O represents the oscilloscope's input capacitance, its input resistance being the standard value of 1MΩ.

Capacitance of the screened lead C_C plus the input capacitance of the oscilloscope form one section of a capacitive potential divider. Trimmer C_T forms the other, and it can be set so that the attenuation of this capacitive divider is 10:1 in volts, which is the same attenuation as provided by the 9MΩ of R_A and the 1MΩ input resistance of the oscilloscope. When this condition is fulfiled, the attenuation is independent of frequency, **Fig. 2a)**.

Defining the cable plus oscilloscope input capacitance as C_E, where C_E is C_C+C_O, Fig. **1b)**, then C_T should have a reactance of nine times that of C_E, i.e. C_T is $C_E/9$. If C_T is too small, high frequency components such as the edges of a squarewave will be attenuated by more than 10:1. This results in the waveform of Fig. **2b)**. Conversely, if C_T is too large, the result is as in Fig. **2c)**.

Input capacitance of an oscilloscope is invariably arranged to be constant for all settings of the Y input attenuator. This means that C_T can be adjusted by applying a squarewave to the oscilloscope via the probe using any convenient Y sensitivity, and the setting will then hold for any other sensitivity.

Circuit Fig. 1a) provides the lowest capacitive circuit loading for a 10:1 divider probe. This circuit has the disadvantage that 90 per cent of the input voltage – which could be very large – appears across variable capacitor C_T.

To take care of this, some probes use the circuit of **Fig. 1c)**. Capacitance C_T is now a fixed capacitor and a variable shunt capacitor C_A is fitted, which can be set to a higher or lower capacitance to compensate for instruments with a lower or higher input capacitance respectively. Now, only 10 per cent of the input voltage appears across the trimmer. As a bonus, the trimmer is also conveniently located at the oscilloscope end of the probe lead, permitting a smaller, neater design of probe head.

Even if a 10:1 passive divider probe – often called, perhaps confusingly, a ×10 probe – is incorrectly set up, the rounding or pip on the edges of a very low frequency squarewave, e.g. 50Hz, will not be too obvious. This is because with the slow time base speed necessary to display several cycles of the waveform, it will appear to settle instantly to the positive and negative levels.

Conversely, with a high-frequency squarewave, say 10MHz, the probe's division ratio will be determined solely by the ratio C_E/C_T. Many a technician – and chartered engineer – has spent time wondering why the amplitude of a clock waveform was out of specification, only to find out eventually that the probe has not been set up for use with that particular oscilloscope.

Waveforms as in Fig. 2 will be seen with a squarewave of around 1kHz.

Probing at high frequencies

At very high frequencies, where the length of the probe lead is an appreciable fraction of a wavelength, reflections would occur, since the cable is not terminated in its characteristic impedance. For this reason, oscilloscope probes often incorporate a resistor of a few tens of ohms in series with the inner conductor of the cable at one or both ends. Alternatively they may use a special cable with an inner made of resistance wire. Such measures are necessary in probes for use with oscilloscopes having a bandwidth of 100MHz or more.

While a 10:1 passive divider probe greatly reduces the loading on a circuit under test compared with a similar length of screened cable, its effect at high frequencies is by no means negligible. **Figure 3** shows the typical variation of input impedance versus frequency of such a probe, when connected to an oscilloscope.

Another potential problem area to watch out for when using a 10:1 divider probe is the effect of ground-lead inductance. This is typically 150nH for a 15cm lead terminated in a miniature 'alligator' clip, and can form a resonant circuit with the input capacitance of the probe. On fast edges, this results in ringing in the region of 150MHz. For high frequency applications it is essential to discard the ground lead and to earth the grounded nose-ring of the probe to circuit earth by the shortest possible route.

Probing actively

Figure 3 shows that over a broad frequency range of, say, 30kHz to 30MHz, the input impedance of a 10:1 passive divider probe is almost purely capacitive. This is illustrated by the almost 90° phase angle. But it is evident that at frequencies well beyond 100MHz, the input impedance of the probe tends to 90Ω resistive – the characteristic impedance of the special low capacitance cable used.

At frequencies where C_T is virtually a short

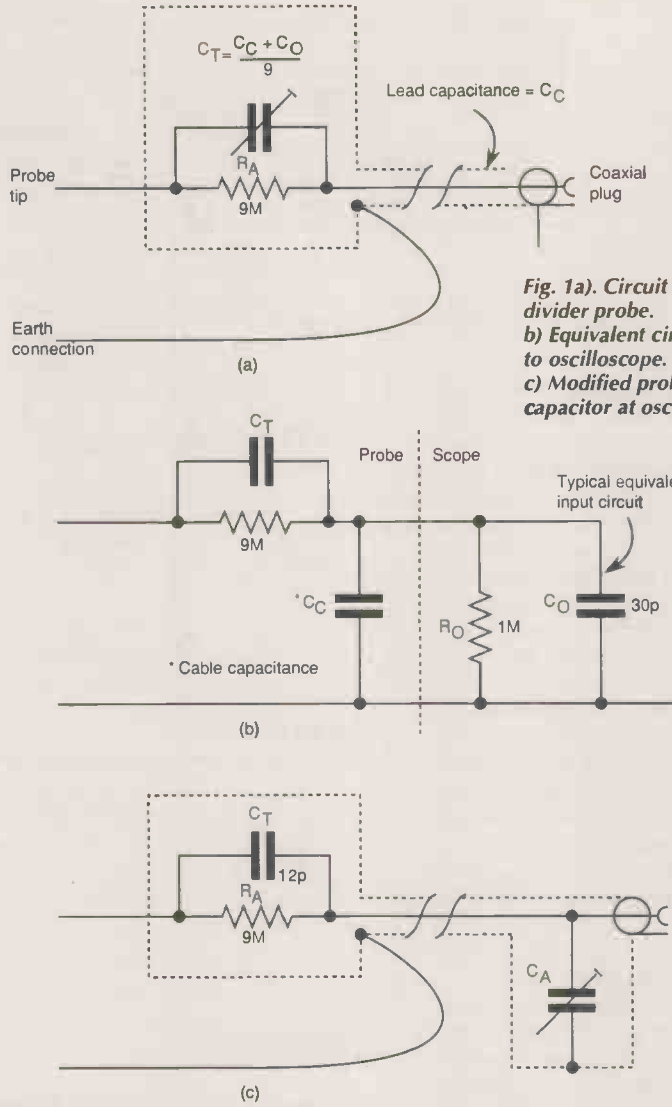


Fig. 1a). Circuit of traditional 10:1 divider probe.
b) Equivalent circuit of probe connected to oscilloscope.
c) Modified probe circuit with trimmer capacitor at oscilloscope end.

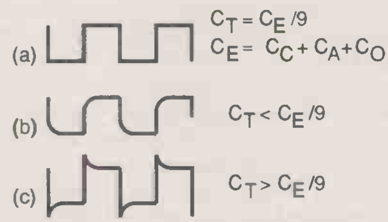


Fig. 2. Displayed waveforms with probe set up a), correctly, b), undercompensated, c), overcompensated.

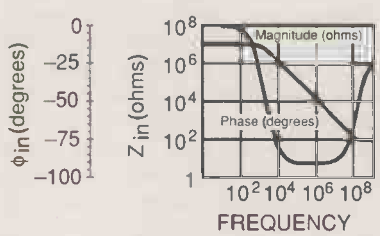
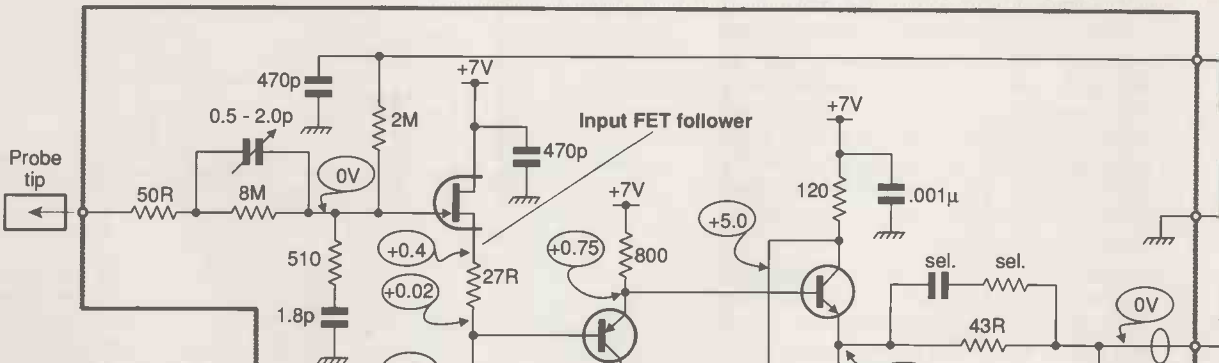


Fig. 3. Variation of impedance with frequency at the tip of a typical 10:1 passive divider probe, Courtesy Tektronix UK Ltd.



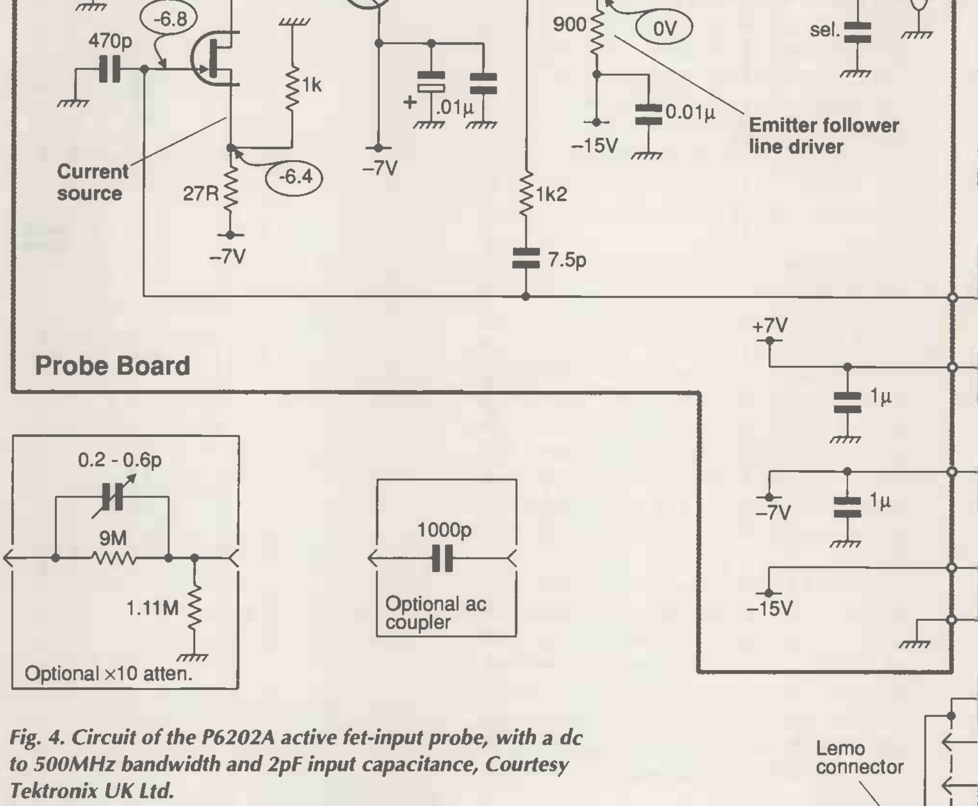



Fig. 4. Circuit of the P6202A active fet-input probe, with a dc to 500MHz bandwidth and 2pF input capacitance, Courtesy Tektronix UK Ltd.

circuit, the input of the probe cable is connected directly to the circuit under test, causing heavy circuit loading. The only way round this is to fit a buffer amplifier actually in the probe head. In this way, the low output impedance of the buffer drives the cable, isolating it entirely from the circuit under test.

Such active probes have been available for many years for top-of-the-line oscilloscopes from the major manufacturers. In many cases, their oscilloscopes are fitted with appropriate probe power outlets.

Figure 4 is the circuit diagram of such an active probe – the Tektronix *P6202A* providing a 500MHz bandwidth and an input capacitance of 2pF. It has stackable clip-on caps to provide ac coupling or an attenuation factor of ten to increase the dynamic range.

The circuit illustrates well how, until comparatively recently, when faced with the need to wring the highest performance from a circuit, designers were forced to make extensive use of discrete components.

Note that such an active probe provides two important advantages over the passive 10:1 divider probe. Firstly, the input impedance remains high over the whole working frequency range, since the circuit under test is buffered from the low impedance of the output signal cable. Secondly, the factor of ten attenuation of the passive probe is eliminated.

While high performance active probes are readily available, at least for the more expensive models of oscilloscope, their price is high. The result is that most engineers are forced to make do, reluctantly, with passive probes. These cause heavy loading on the circuit under test at high frequencies, and cause a loss of a factor of ten in sensitivity.

Affordable passive divider probes for oscilloscopes with a bandwidth of 60 to 100MHz are readily available, but active probes of a similar modest bandwidth are not. But with the continuing improvements in op-amps of all sorts, it is now possible to design simple active probes without resorting to the complexity of a design using discretes such as Ref. 1 or Fig. 4.

Designing an active probe

To provide a 10MΩ input resistance – the same as that of a passive 10:1 divider probe – an active probe built around an op-amp must use a mos input device.

For optimum performance at high frequencies, it is desirable that the op-amp should drive the coaxial cable connecting the probe to the oscilloscope as a matched source. In the jargon of the day, the cable is described as being 'back-terminated'. This, together with a matched termination at the oscilloscope end of the probe lead, divides the voltage swing at the output of the op-amp by two.

So for a unity gain probe, the op-amp must provide a gain of two. For this purpose, an op-amp which is partially decompensated, for use at a gain of two or above, is convenient.

An active probe using such a mos-input op-amp, the SGS-Thomson *TSH31*, is shown in **Fig. 5a)**. This op-amp has a 280MHz gain-bandwidth product, achieved by opting for only a modest open loop gain; large-signal voltage gain, A_{vd}, is typically ×800 or 58dB for V_o of ±2.5V and R_l at100Ω. At a gain of 2, it should therefore provide a bandwidth approaching 140MHz.

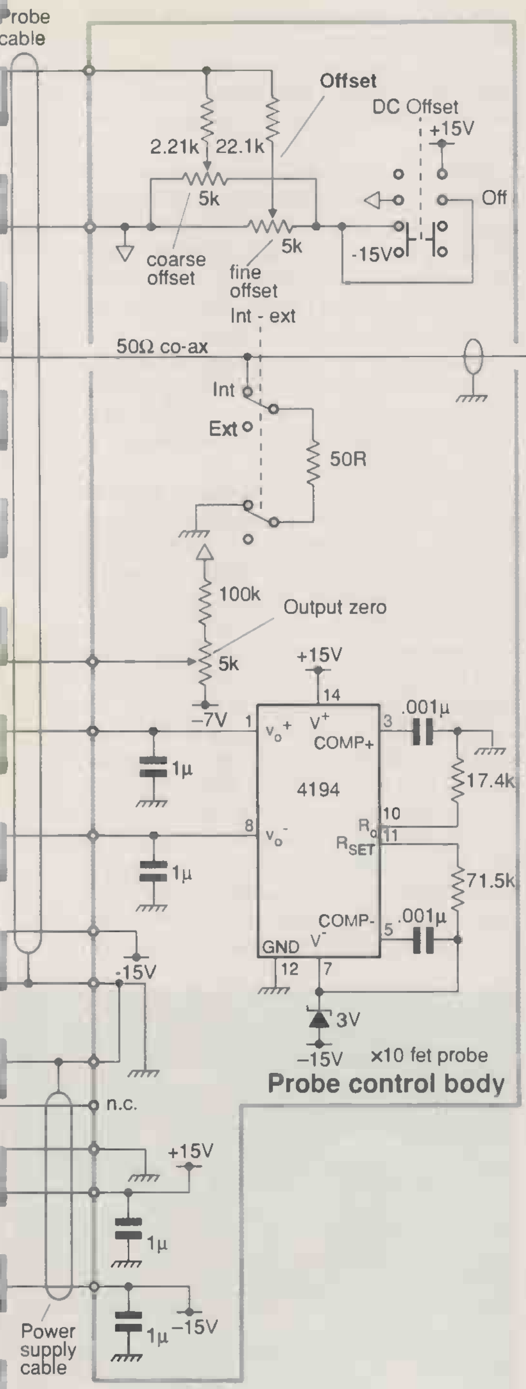
Take care with the layout to minimise any stray capacitance from the non-inverting input to ground. This would result in high-frequency peaking of the frequency response. If need be, a *soupçon* of capacitance can be added in parallel with the 1kΩ feedback resistor from pin 6, to control the settling time.

A zero offset adjustment is shown, but in most cases this will be superfluous. Even with a device having the specified maximum input bias current I_{ib} of 300pA, the offset due to the 10MΩ ground return resistor at pin 3 is only 3mV. The typical device I_{ib} is a meagre 2pA.

With the omission of the offset adjust circuitry, the circuit can be constructed in a very compact fashion on a few square centimetres of copper-clad laminate or 0.1in matrix strip board. The output signal is routed via miniature 50Ω coaxial cable.

Supply leads can be taped along side the coaxial cable to a point near the oscilloscope end of the probe. Here they branch off, allowing a generous length for connection to a separate ±5V supply, assuming such is not available from the oscilloscope itself.

Note the use of a commercially available 50Ω 'through termination' between the oscilloscope end of the probe signal lead and the Y



input socket of the oscilloscope itself.

For ac applications, where it is necessary to block any dc level on which the signal of interest may be superimposed, a blocking capacitor can be incorporated in a clip-on cap to fit over the probe tip. A similar arrangement can be made to house a 10:1 divider pad, to extend the dynamic range of the unit. Without such a pad, the maximum signal that can be handled is clearly quite limited.

Bear in mind that ±2.5V peak-to-peak at the output of the op-amp will provide the oscilloscope input with only ±1.25V. As a result, an attenuator cap will be needed if looking at, for example, clock pulses. But for this purpose, a conventional 10:1 passive divider probe will usually suffice.

Where an active probe scores is when looking at very small signals, which are too small to measure with a 10:1 passive divider probe. Another application where an active probe scores is when looking at high frequency signals emanating from a high impedance source. Clearly, the heavy damping imposed by a passive divider probe at 100MHz and above precludes its use to monitor the signal across a tuned circuit. On the other hand, the active probe provides much reduced damping, in addition to enabling much smaller signals to be seen.

An active probe to the circuit of Fig. 5a) was made up and tested. As miniature $^1/_{16}$W, lkΩ resistors were not to hand, 1.2kΩ resistors were used instead. This, together with the use of a DIL packaged amplifier in a turned pin socket, rather than the small outline version, meant that some capacitance between pins 2 and 6 was needed.

A 0.5-5pF trimmer was used: it was adjusted so that the probe's response to a 5MHz squarewave with fast edges, see Fig. 9a), was the same as a Tektronix *P6106* passive probe, both being used with a Tektronix *475A* oscilloscope of 250MHz bandwidth. Advantages of an active probe are illustrated in Fig. 5b), where all traces are effectively at 100mV/division, allowing for the unity gain of the active probe, and the 20dB loss of the passive probe. All four traces show the 100MHz cw output of an inexpensive signal generator, the Leader *LSG-16*.

Measurements were made across a 75Ω termination, the top trace being via the active probe and the next one via the *P6106* passive probe. Both show an output of about 280mV peak to peak, agreeing well with the generator's rated output of 100mV rms.

The third trace shows the same signal, but with a 470Ω resistor connected in series with the tip of the active probe, while the bottom trace is the same again but with the 470Ω resistor connected in series with the tip of the passive probe.

The effect of the 470Ω resistor has been to reduce the response of the passive probe by 12dB, while that of the active probe is depressed by only 4.5dB. Thus the active probe not only provides 20dB more sensitivity than the passive probe, but exhibits a substantially higher input impedance to boot.

Providing gain

An active probe can be designed not merely to provide unity gain, avoiding the factor of ten attenuation incurred with a passive divider probe, but actually to provide any desired gain in excess of unity. **Figure 6a)** shows a circuit providing a gain of times ten, which as before requires a gain of twice that from the op-amp.

Again, in the interests of providing the con-

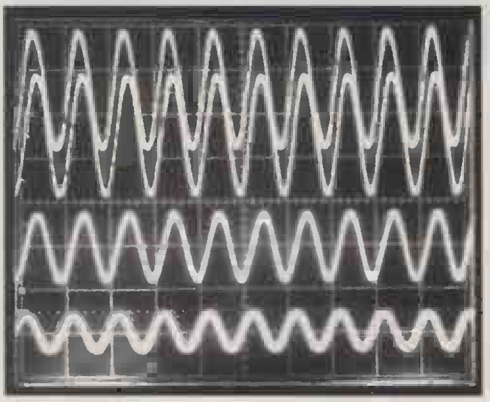


Fig. 5b) Performance of the active probe, compared with a P6106 passive probe. Signal generator output, 100MHz cw at 100mV, viewed at 100mV/div vertical and 10ns/div horizontal. Top and third traces are active probe without and with a 470Ω resistor in the tip, respectively. Second and bottom traces are the same but for the passive probe.

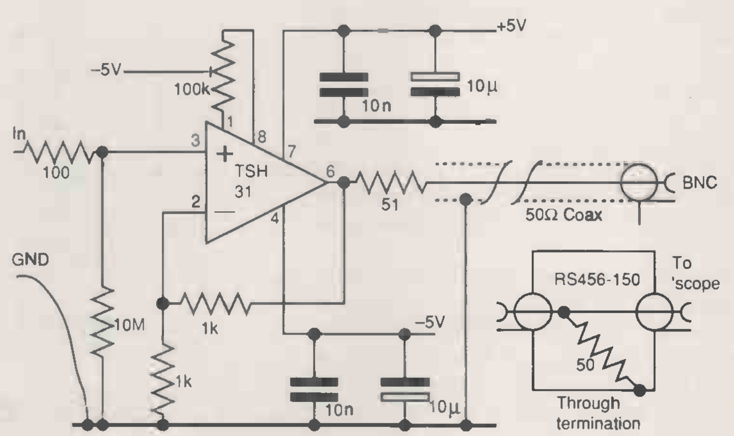


Fig. 5a). Circuit of a unity gain active fet input probe, using a decompensated op-amp designed for use at gains of ×2 or greater. Bandwidth should be well over 100MHz.

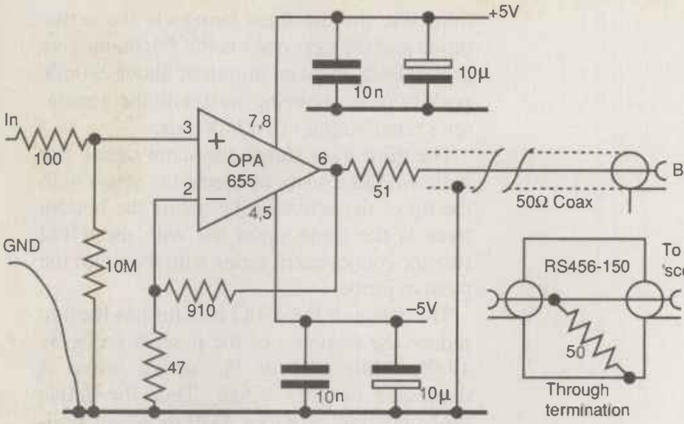
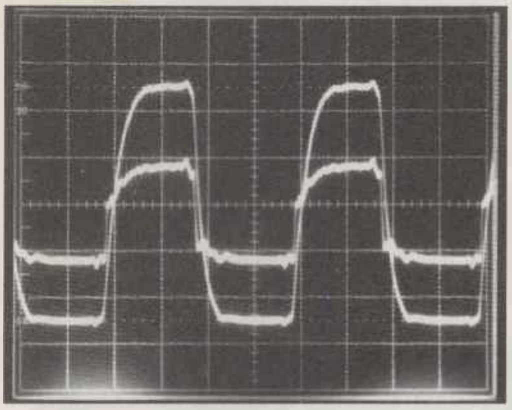


Fig. 6a). Circuit diagram of an active fet input probe providing a net gain of ×10. b) 5MHz 100mV test squarewave input, smaller trace, at 50mV/div. Output at oscilloscope, larger trace, is 1V peak-to-peak at 200mV/div. Time base is 50ns/div.



ventional 10MΩ probe input resistance, a fet input op-amp was chosen, in this case the Burr Brown *OPA655*. This device is internally compensated for gains down to unity, and provides a 400MHz gain-bandwidth product.

In this application it is required to provide a gain of twenty, so clearly a decompensated version of the *OPA655* would improve performance. But despite persistent rumours of the imminent appearance of such a version, I have not managed to get my hands on one.

At a gain of times twenty or 26dB, the *OPA655* might be expected to provide a bandwidth of 400/20, or approaching 20MHz. Note however that as more and more gain is demanded of a unity-gain compensated voltage feedback op-amp, the bandwidth tends to reduce rather faster than *pro-rata* to the increase in gain.

Figure 6b) records the performance of the times ten gain active probe of Fig. 6a), tested with a 100mV peak-to-peak 5MHz squarewave.

Rise and fall times of the test squarewave were 4ns, and of the oscilloscope 1.4ns. The smaller waveform is the 100mV squarewave recorded with a passive 10:1 divider probe with the oscilloscope set to 5mV/division. Effectively this is 50mV/division allowing for the probe.

The larger waveform is the 1V peak-to-peak output of the active probe, recorded at 200mV/div. Rise and fall times of the active probe output are 25ns and 20ns respectively; it is not uncommon to find differing rise and fall times in high performance op-amps, though here the result is influenced also by the shape of the positive-going edge of the test waveform.

Taking an average of 22.5ns and reducing this to 22ns to allow for the risetimes of the oscilloscope and test waveform, gives an estimated bandwidth for the active probe of 16MHz. This is equated using the formula risetime t_r=0.35/BW, where t_r is in microseconds and bandwidth *BW* is in megahertz.

This probe would be useful with any oscilloscope having a 20MHz bandwidth, the instrument's 17.5ns risetime being increased to 28ns by the probe.

A much faster probe with a gain of ten can be produced using that remarkable voltage feedback op-amp, the Comlinear *CLC425*. The *425* is a decompensated device, for use at gains of not less than ten. It is an ultra low noise wideband op-amp with an open-loop gain of 96dB and a gain-bandwidth product of 1.7GHz. At the required gain of ×20 therefore, it should possible to design an active probe with a bandwidth approaching 85MHz.

Figure 7a) was made up and tested using a 5MHz squarewave with fast edges, produced with the aid of *74AC* series chips, as shown in **Fig. 9a)**. The result is shown in Fig. **7b)**. Here the smaller waveform is the attenuated test waveform viewed via a 10:1 passive divider probe at 50mV/division.

The test waveform was intended to be 50mV, but the accumulated pad errors resulted in it actually being 55mV. The larger trace is the 550mV output from the ×10 active probe, recorded at 100mV/div. with the oscilloscope's variable Y gain control adjusted to give exactly five divisions deflection, for risetime measurements.

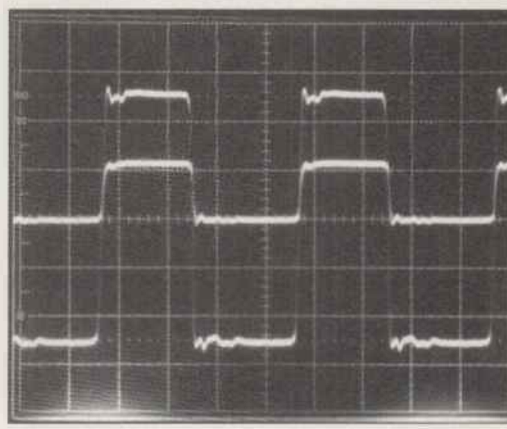
The two traces were recorded separately, only one probe at a time being connected to the test waveform, Fig. 7 b) being a double exposure.

With the timebase speed increased to 10ns/div. the rise and fall times were measured as 4.5 and 4.0ns respectively. This implies a bandwidth, estimated by the usual formula, of around 80MHz – even before making corrections for the risetimes of the oscilloscope and test waveform.

But there is a price to be paid for this performance. The *CLC425* is a bipolar device with a typical input bias current of 12μA. This means that the usual 10MΩ input resistance is quite out of the question.

In the circuit of Fig. 7a), however, a 100kΩ input resistance has been arranged with the aid of an offset-cancelling control. In the sort of high speed circuitry for which this probe would be appropriate, an input resistance of 100kΩ will often be acceptable. The need to adjust the offset from time to time is a minor drawback to pay for the high performance provided by such a simple circuit.

As described in connection with the unity gain active probe of Fig. 5, the two ×10 versions of Figs 6a) and 7a) can be provided with clip-on capacitor caps for dc blocking. Clearly, with an active probe having a gain of ×10, the maximum permissible input signal, if overloading is to be avoided, is even lower than for a ×1 active probe. But it is not worthwhile making a 20dB attenuator cap for a ×10



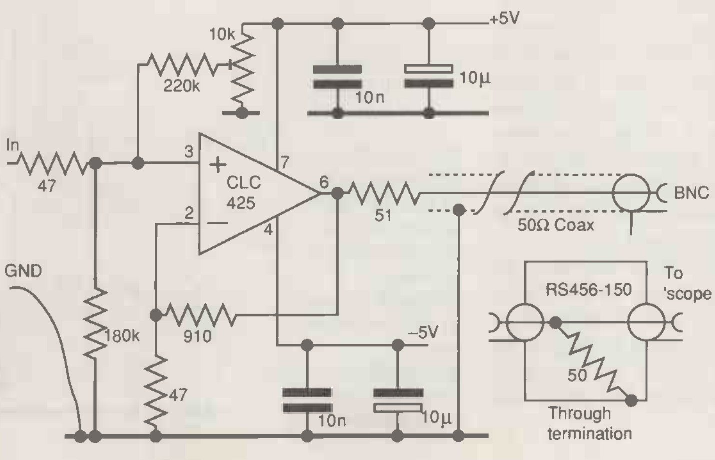


Fig. 7a). Circuit diagram of an active bipolar probe providing a net gain of x10. b) 5MHz 55mV test squarewave input (smaller trace, at 50mV/division), 550mV peak to peak output at oscilloscope (larger trace, at >100mV/division). Time base is 50ns/division. Output rise and fall times measured at 10ns/div. but not shown, are 4.5 and 4.0ns respectively.

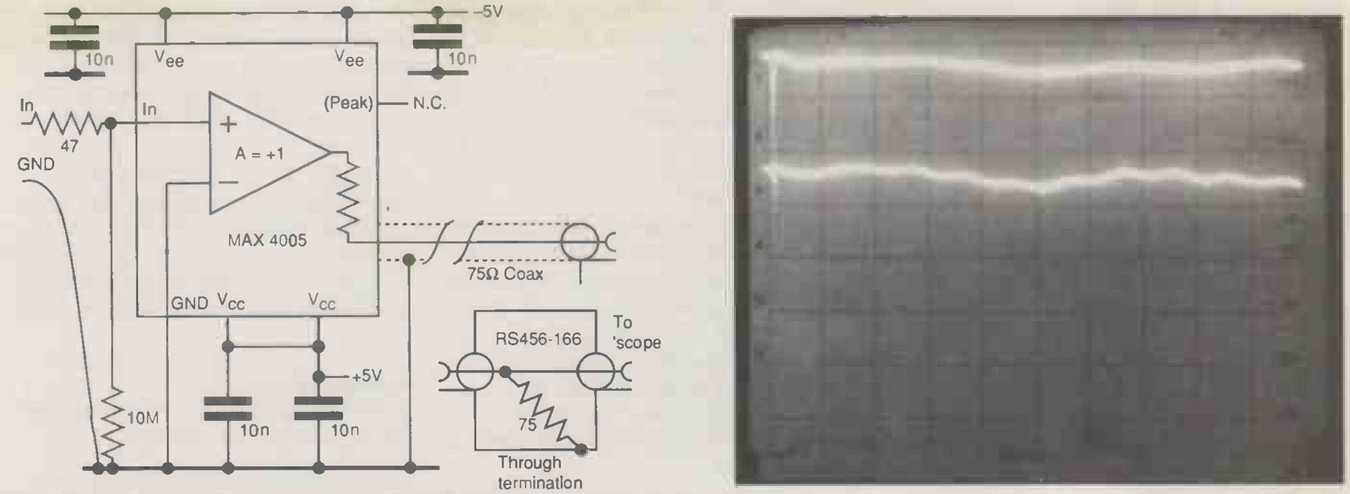


Fig. 8a). Circuit of a wideband fet input probe with a gain of ×0.5.
b) Roughly level output of a sweeper use to test the probe circuit of a), upper trace, and output of probe, lower trace. Span 0-1000MHz, intermediate frequency bandwidth 1MHz, 10dB/division vertical, ref. level, top of screen, is +10dBm.

active probe; with the probes described being so cheap and simple to produce, it is better simply to use a ×1 probe instead.

An interesting possibility for the circuit of Fig. 6a) is to fit a miniature single-pole changeover switch arranged to select either the 47Ω resistor shown, or a 910Ω resistor in its place. This provides an active probe switchable between gains of ×1 and ×10.

In the ×1 position, the bandwidth should rival or exceed that of Fig. 5a). This scheme is not applicable to the circuit of Fig. 7a) however. While the *OPA655* is unity-gain stable, the *CLC425* is only stable at a gain of ×10 or greater.

For a really wideband active probe...

The three probes described so far all use op-amps with closed loop feedback to define a gain of two, giving a gain of unity net at the oscilloscope input. But another possibility is to use a unity gain buffer, where no external gain setting resistors are required. This provides the ultimate in circuit simplicity for an active probe.

Devices such as National Semiconductor's *LH0033* or *LH0063* fet-input buffers could be considered. Having some samples of the Maxim *MAX4005* buffer to hand, I made an active probe using this device, which claims a 950MHz –3dB bandwidth and is designed to drive a 75Ω load.

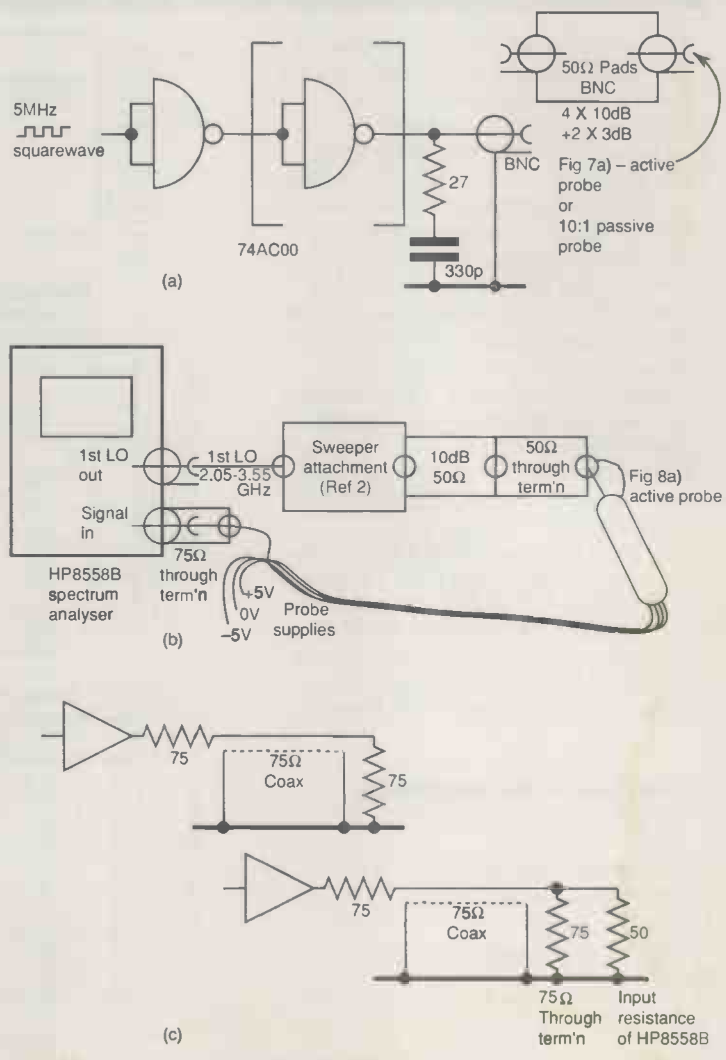
The usual 10MΩ probe input resistance is easily achieved, as the *MAX4005* is a fet-input device. **Figure 8a)** was made up on a slip of copper-clad laminate 1.5cm wide by 4.0cm long. I mounted the chip near one end of the board, most of the length being taken up with arrangements to provide a firm anchorage for the 75Ω coaxial cable. The chip was mounted upside down on four 10nF chip decoupling capacitors connected to the supply pins and used as mounting posts.

Note that to minimise reflections on a cable, the *MAX4004* contains an internal thin-film output resistor to back terminate the cable. This means in practice that the net gain from probe input to oscilloscope input is in fact ×0.5. In turn, this means that the 5 and 10mV input ranges on the oscilloscope become 10 and 20mV respectively. While slightly less convenient since the 20mV range becomes 40mV per division, this is acceptable for most jobs.

For this probe, of course, a 75Ω coaxial lead was chosen, terminated at the oscilloscope input with a commercial 75Ω through termination.

Since the expected bandwidth of this active probe was far in excess of the 250MHz bandwidth of my *TEK 475A* oscilloscope, some other means of measuring it was required. My *HP8558B* spectrum analyser was pressed into service. Unfortunately, this instrument does

Fig. 9a). Test circuit used to produce a 5MHz squarewave with fast edges, to test the probe of Fig. 7. The 27Ω plus 330pF snubber at the output suppressed ringing on the test waveform. b) Test set-up used to test the wideband probe of Fig. 8. c) Showing how the 6dB signal reduction in normal use becomes 11dB in the test set-up of b) above. Together with the 10dB pad at the sweeper output, this accounts for the 21dB separation of the traces in Fig. 8b).



not provide a tracking generator output, but a buffered version of the swept first local oscillator output covering 2.05 - 3.55GHz is made available at the front panel.

In an add-on unit as described in Ref. 2, this is mixed with a fixed frequency 2.05GHz oscillator to provide a swept output tracking the analyser input frequency. Mixer output is amplified and low pass filtered, providing a swept output level to within ±1dB or so, at least up to 1GHz, at a level of around +6dBm. This is shown as the top trace in **Fig. 8b)**.

The active probe was then connected to the output of the sweep unit, via a 10dB pad to avoid overloading, and a 50Ω through termination to allow for the high input impedance of the *MAX4005*. Significant care needed to be taken with grounding arrangements at the probe input, Fig. **9b)**.

Output of the probe – including the 75Ω through termination shown in Fig. 8 a) – was connected to the input of the spectrum analyser. As a result, the 75Ω coaxial cable was in fact terminated in 30Ω. This mismatch explains the amplitude variations in the probe output, Fig. 8 b), lower trace, corresponding to the electrical length of the 75Ω coaxial lead.

These apart, the level follows that of the sweeper output, upper trace, up to just under 1GHz, where the expected roll-off starts to occur. The level is about 20dB below that of the sweeper output which is explained by the 10dB pad, and the additional loss above the expected 6dB, due to the mismatch at the analyser input, Fig. **9c)**.

You may have been asking, "What is the use of a 950MHz bandwidth active probe when the 75Ω termination at the oscilloscope is in parallel with a input capacitance of around 20pF?" After all, the effective source resistance seen at the instrument's input is 37.5Ω. The oscilloscope bridges both the source and load resistors, which are thus effectively in parallel, while the reactance of 20pF at 950MHz is 8.4Ω.

But remember that the figure of 20pF is a lumped figure, measured at a comparatively low frequency. In fact, this capacitance is typically distributed over a length of several inches. The input attenuator in the *475A*, for example, is implemented using thick film pads. These are connected in circuit or bypassed as required by a series of cams on the volts-per-division switch.

Because of this construction, the 20pF is distributed over some kind of transmission line, the characteristics of which are not published. It is therefore likely that the effective capacitance at 950MHz is less than 20pF: the only way to be really sure what bandwidth the probe of Fig. 8a) provides with any given oscilloscope is to measure it. But given the 370ps rise time of the *MAX4005*, this exceedingly simple active probe designed around it is likely to out-perform the vast majority of oscilloscopes with which it may be used. ■

References

1. 500MHz high impedance probe, J. Dearden, *New Electronics*, March 22, 1983 page 28.
2. Simple tracking generator for spectrum analyser, Ian March, *Electronic Product Design*, July 1994 page 17.

Acknowledgments

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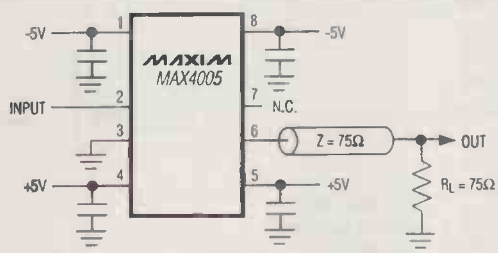

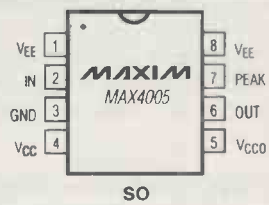

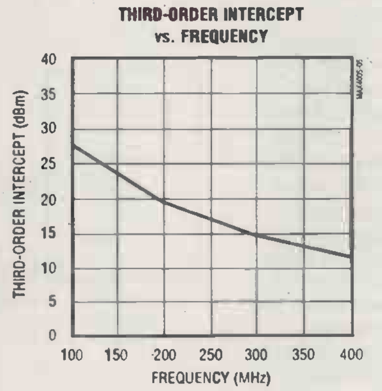

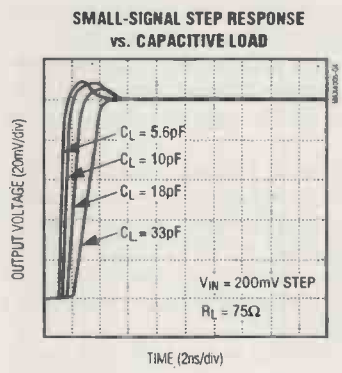

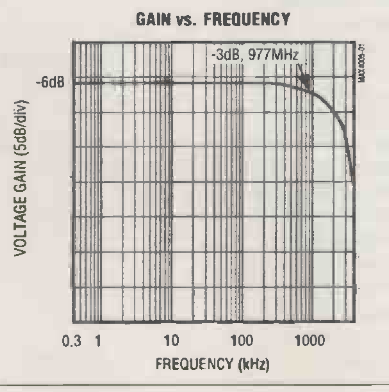

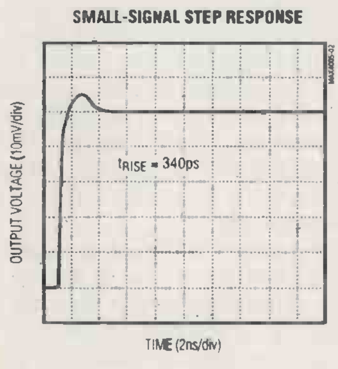

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Smart batteries

Geoff Lewis shows how intelligent control is solving battery recharging problems.

Portable equipment powered by rechargeable batteries can create many awkward problems for the operators of devices such as cell phones, camcorders and laptop computers. If NiCd batteries are recharged too soon, their future capacity might be jeopardised. There is a need to know fairly accurately the current state of charge, and if there is sufficient capacity to carry out the next operation before recharging.

Battery-capacity is specified by the product of discharge time and a constant discharge current, expressed in ampere hours (Ah). The discharge period is limited by the end-point or end of discharge voltage (eodv). The rated capacity *C*, is commonly quoted for a 5, 10 or 20 hours discharge period and *C5* is commonly used for small rechargeables such as A or AA cells. Capacity *C5* is therefore the rated capacity for a constant current discharge to the EODV in 5 hours. Similarly *I5* would be 20% of the rated capacity value in Ah. The power dissipated during this period may also be specified as *P5*. The battery recharge rate typically varies from 150% to 300% of the ampere-hour capacity.

Rechargeable batteries should not be discharged beyond the end-point voltage too often as this form of abuse can reduce the useful lifetime. If care is taken to control the charge and discharge cycles, then it is possible for a cell to easily withstand between 500 and 1000 recharge cycles. Abuse can easily reduce this value by a factor as high as 100. Efficient management of battery power can now be achieved by a system known as Smart Battery. This describes a pack of rechargeable batteries equipped with a microchip circuit that collects, calculates, and predicts battery information to the host system, under software control.

This is just one example of the many potential auses for a Smart battery, illustrating information that can be provided and the manner in which it might be displayed.

The Smart Battery concept was originally developed through cooperation between Intel Corp. and Duracell Inc during 1993-94. Since that time, other battery and semiconductor manufacturers such as Philips, Sequoia, Exar and Maxim, together with original equipment manufacturers and software houses, have combined efforts to create systems that provide the user with:

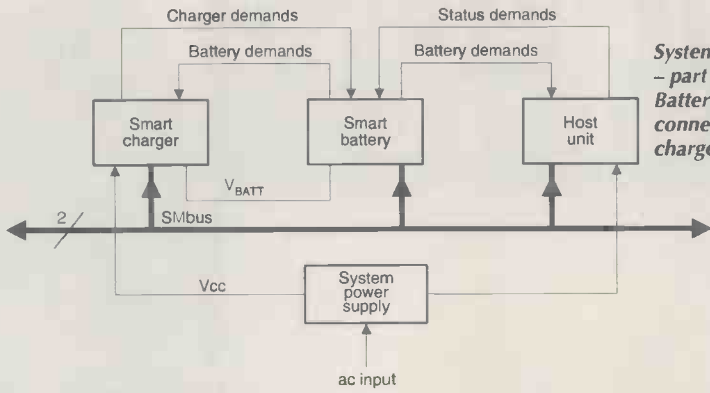
- An assessment of the current state of charge.
- Accurate prediction of remaining operational time.
- Controlled discharge-recharge cycles.
- Controlled charging and operation within safe limits.
- Operation with any battery technology/chemistry.

This close and accurate control of the battery environment produces longer life and run-times, typically by as much as 20%. Such an intelligent system which is constructed around asics and surface mounted components, can occupy a space of about 350mm^2. A figure that is commensurate with the size of the batteries that it is intended to manage.

Currently the operation of Smart Battery systems are restricted to relatively light current applications and are rather more costly than conventional technology. However, it is envisaged that the concept will soon be adapted for use with portable television cameras that require larger batteries, and this should lead to a significant increase in usage, leading in turn to competitive costs.

Battery packs that are intended for use in Smart Battery systems are equipped with a rom that carries embedded code that identifies the battery maker, the date of manufacture, the battery serial and model number, the battery chemistry – NiCd, NiMH etc – the theoretical capacity and the current number of recharge cycles. This information is added to that stored within the Smart Battery system during use to provide a complete on-board battery history.

As shown in the diagram on the right, the host unit, which can be anything from a mobile telephone to a laptop computer, the battery and the charger, are all linked together via the two wire System Management Bus or



System Management Bus – part of the Smart Battery concept – connects the battery, charger and host.

SMBus. This bus structure is very similar to that of the IC bus in that either element of the system can act as master or slave depending upon the system needs.

However, the SMBus employs a clock circuit that is specified to run at a frequency between 10kHz and 100kHz and uses fixed voltage level signals. A logic low level is specified to lie between –0.5V and +0.6V, with a high between 1.4V and 5.5V. Since most c-mos devices pull down below 0.4V and open collector/drain devices pull up to the supply level of 3V or 5V, well above 1.4V, the system is c-mos and ttl compatible.

The battery charger periodically polls the battery to obtain its charging characteristics and then adjusts its output to match the requests. The charger also receives notice of critical events such as alarms for overcharging, over-voltage and over temperature.

In a similar way, the host device requests information from the battery in order to advise the user of the current status. This includes remaining capacity, future run time, required recharge time and the predicted final discharge point. Parameters that may be displayed for the user are shown in the handset diagram.

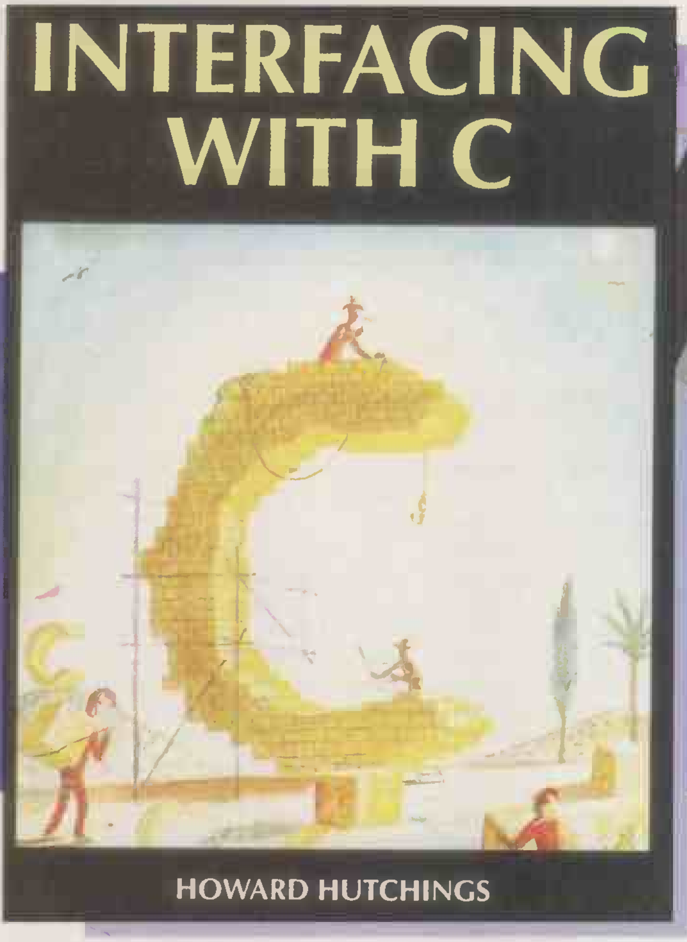
In the interests of power efficiency, sensors measure the voltage dropped across very low value resistors – typically 0.01 to 0.05Ω – which are wired in series with the negative lead of the battery. This allows for the monitoring of both charge and discharge currents to an accuracy of 0.2%.

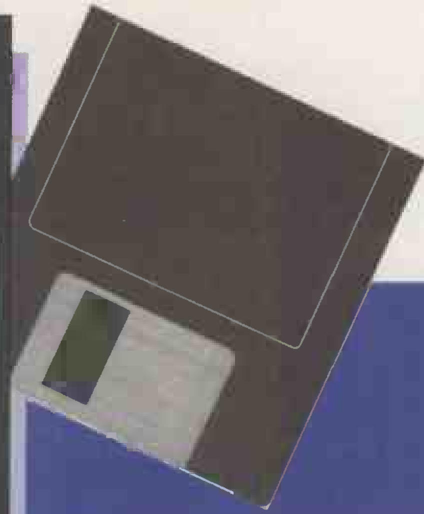
By using rolling average values, the system integrates the charge and discharge currents to allow it to compensate for the changes that occur in efficiency with different battery loads. The Smart Battery system uses an internal temperature and voltage stabilised clock circuit to avoid the need to use either external crystals or thermistors. Battery temperature can be sensed in 0.1° steps from –40 to +85°C to an accuracy better than 3°.

The system continually monitors the battery terminal voltage in order to evaluate the eodv. Because battery self discharge which is time and temperature dependent can be a significant feature of these rechargeables, the circuit monitors this process during power down, to maintain an accurate record of the state of charge.

The system may also allow the ac power source to supply the host unit during periods in which the battery is fully charged and still connected to the mains, thus prolonging the battery's lifetime.

Further information about Smart Battery systems can be obtained from either Intel Corporation, Intel Architecture Labs Technical Support, Sacramento, California or Duracell Batteries Ltd, at Crawley. The writer is particularly indebted to Mike Dixon, OEM Business Manager, Duracell (UK), for his help in the preparation of this article. ■





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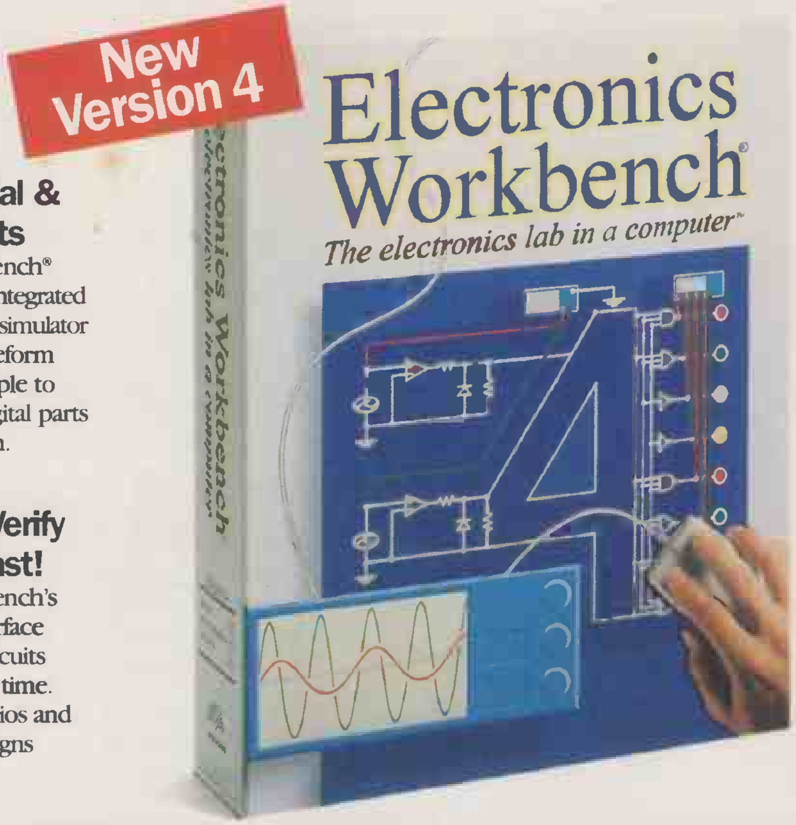

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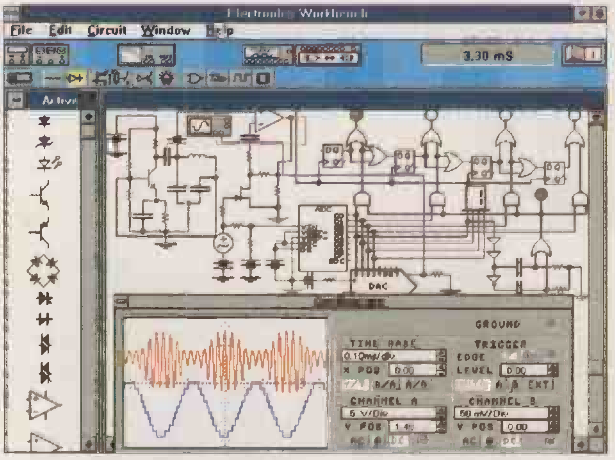

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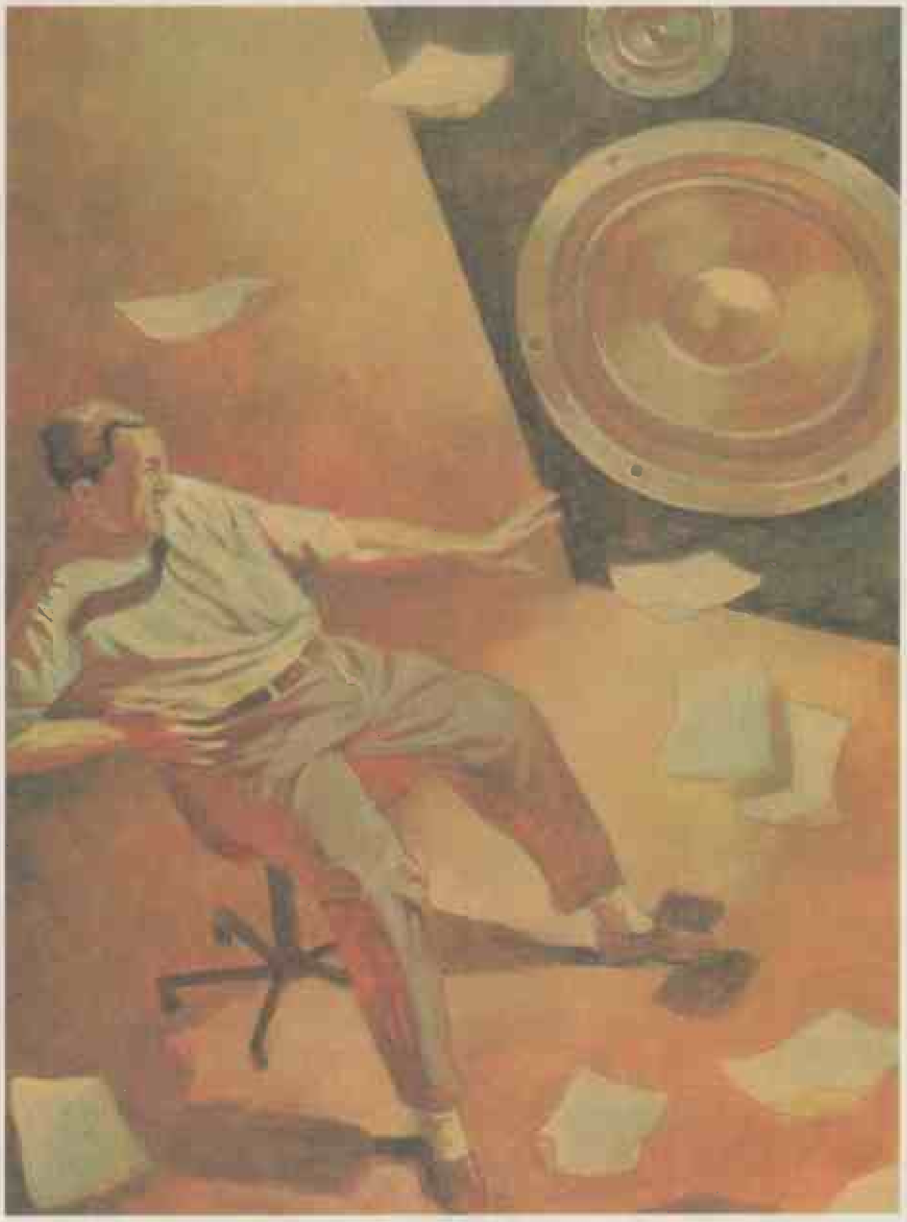


SPEAKER Feedback

Ian Hegglun demonstrates how feedback – in particular current feedback – can be used to enhance loudspeaker performance.

Motional feedback would seem to be the obvious solution to the main shortcomings of the voice coil loudspeaker – namely distortion and resonance. Philips brought out a in small cabinets motional feedback system using an acceleration transducer in the early 1970s. This system gave clean low-end bass[1]. Another system appeared in *Wireless World*[2], also using a Philips speaker with a factory fitted acceleration transducer, allowing distortion to be reduced by a factor of 5 at 25Hz.

With such impressive results, why hasn't motional feedback become more popular? A few reasons are given in reference 3. My version of the problems are:

- Speakers at normal levels do not produce very much distortion. It is only when they are driven hard – particularly at low frequencies – that distortion rises rapidly. Since low bass distortion is not very noticeable by ear[4], why reduce it?
- Motional feedback is not easy to retrofit to existing speakers.
- Transducers usually require complicated equalisation. For example, in reference 2, twelve break frequencies were used to compensate the transducer.
- Nyquist stability places a limit on the amount of feedback that is available at high

frequencies. Feedback is limited by the frequency where an additional 180° phase lag occurs. In reference 2 this was at 5kHz. This allows 12dB of servo-loop feedback with a 10dB margin for safety up to 300Hz.

- Conventional speakers can usually be run on almost any amplifier and speakers can be interchanged with almost any speaker.
- Several other techniques can give as good bass in small enclosures with standard speakers.

Two of these alternative techniques, negative resistance damping – called Active Servo Technology or AST by Yamaha – and equalisation produce a good low frequency response in a small enclosure. Both techniques can deal with the problem of speaker resonance. Their main advantage over motional feedback is they use standard speakers.

Some observations and circuits tested show feedback around speakers can overcome some of the problems with equalisation.

Equalising loudspeaker response

Equalisation, as applied to speakers, is a technique that boosts the signal to the power amplifier to compensate for frequency roll-off by the speaker and enclosure. Macaulay's recent articles[5,6] are good examples of how low-end compensation and a small enclosure of less than 20 litres can achieve a 20Hz roll-off.

An understanding of speaker basics is helpful when applying equalisation. Placing a speaker in a small enclosure raises the free-air resonant frequency and the Q. **Figure 1**[7] shows the effect of reducing enclosure volume on Q and the –3dB frequency f_c. Compliance ratio α is simply the ratio of the speaker's V_{as} to enclosure volume V_b. Speaker compliance, V_{as}, is expressed as an equivalent volume of air and compliance is the inverse of the spring constant k.

For example, an enclosure with a compliance ratio of 3, means the enclosure has a spring constant 3 times that of the speaker. Figure 1 shows an α of 3 raises speaker Q by a factor of 2. In speaker design, the target Q is usually 0.5 for best transient response. Most speakers start with a Q (Q_{ts} in data sheets) of around 0.3 with lower cost speakers having Qs in the range 0.5 to 1.0. For small enclosures, loudspeaker Q increases to at least 0.6 depending on the driver used. Low cost speakers can end up with a Q of 2, which sounds unacceptably 'boomy'.

Although an equaliser is usually used to boost low-end roll-off, an equaliser can also be used to add damping to a speaker system, as described in the panel. This allows smaller cabinets and/or lower cost drivers. However, compensation usually requires measurement of f_o and Q so the correct amount of compensation can be applied. Although most equalisers f_o and Q parameters are not easily changed, a circuit in reference 8 does allow independent setting of f_o and Q.

There is a practical limit to how high Q can go with equalisation, and hence how small the enclosure can be made. This is because of the sensitivity to errors that creep in at manufacture and changes with time, temperature, etc. There is also the limit of speaker excursion which limits the speaker size that is needed to move the amount of air needed to reach a design SPL at the lowest frequency. For example, a 200mm speaker with typically 6mm travel can only produce 86dB at 1m but this becomes 92dB with two speakers in cabinets near a wall.

Excursion limits apply to motional feedback systems. However, if the transducer is linear with large excursions, speaker travel can be doubled giving a 6dB improvement over other systems.

Equalisation of sealed enclosures need a +12dB/octave boost starting from cabinet resonance f_o. Most vented reflex enclosures require +24dB per octave boost, eg two 12dB per octave equalisers. The Microflex concept appears to be a logical step in the right direction. By tuning the port to 30Hz, slightly above the required roll-off frequency, a small enclosure has a –12dB per octave roll-off, rather than 24dB/octave, down to the port's roll off at 20Hz[6]. This simplifies equalisation while giving a few decibels advantage over a sealed enclosure by reducing cone excursion at very low frequencies.

Negative-resistance damping

To reduce excessive Q of a small enclosure it is possible to add negative resistance from the power amplifier to reduce the effective speaker resistance and give more electrical damping. If the negative resistance could completely cancel speaker resistance then the effect of speaker resonance on sound output is suppressed.

Better damping via an equaliser *before* the power amp?

Strange as it may sound, your loudspeaker can be damped by adding an equaliser *before* the power amplifier. This is because the overall response of a system is the mathematical product of the transfer function for each stage. We can arrange the transfer function of an equaliser to have the same numerator as the speaker's denominator so the numerator of the equaliser cancels the denominator of the speaker, like cancelling in fractions. The denominator of the equaliser now sits under the speaker's numerator, see appendix. This allows the equalised speaker to have a lower roll-off and a low Q for good transient response.

With a 12dB per octave speaker roll-off the equaliser needs to be a second-order function with adjustable f_o and Q. Ideally, f_o and Q should be adjustable with a single variable resistor and non-interactive with Q. The circuit in figure 8[8] allows this. Settings for equaliser Q (k_1) and f_o (k_2) can be calculated from the equations given in the appendix.

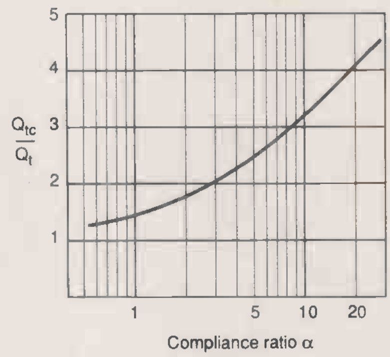


Fig. 1. Effect of enclosure compliance on enclosure Q.

Figure 2 shows the equivalent circuit of a speaker in a sealed enclosure driven by a negative resistance voltage source. If $-R_g$ can be made equal to R_{vc}, then the voltage source in Fig. 2 is effectively directly controlling speaker back-emf at low frequencies. Hence speaker velocity is controlled by the amplifier since emf is directly proportional to voice coil velocity in a constant magnetic field. This means negative resistance damping is a form of motional feedback where voice coil velocity is being controlled at the low frequency end – a discovery of my own. Since this discovery, I came across the same technique[9] from the fifties, to increase damping using a voltage feedback proportional to voice coil velocity.

The pioneers, see 'Further reading', call this technique 'positive current feedback', 'dynamic negative feedback' as well as 'motional feedback' and the generation of negative resis-

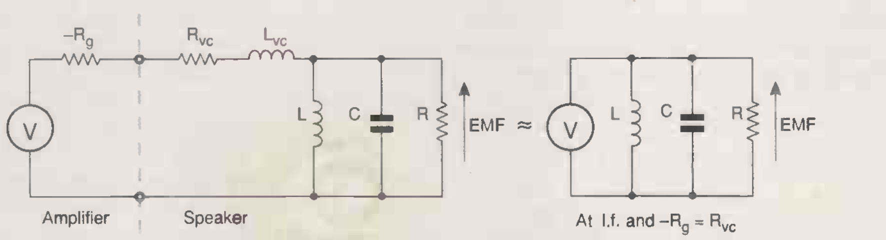


Fig. 2a). Equivalent circuit of a speaker in a sealed enclosure fed by a negative resistance source. Reference 6 defines L, C and R terms. In 2b), right, when $-R_g$ equals R_{vc} the amplifier input voltage effectively controls the voice coil's back emf at low frequencies. This prevents resonance between L and C affecting cone movement.

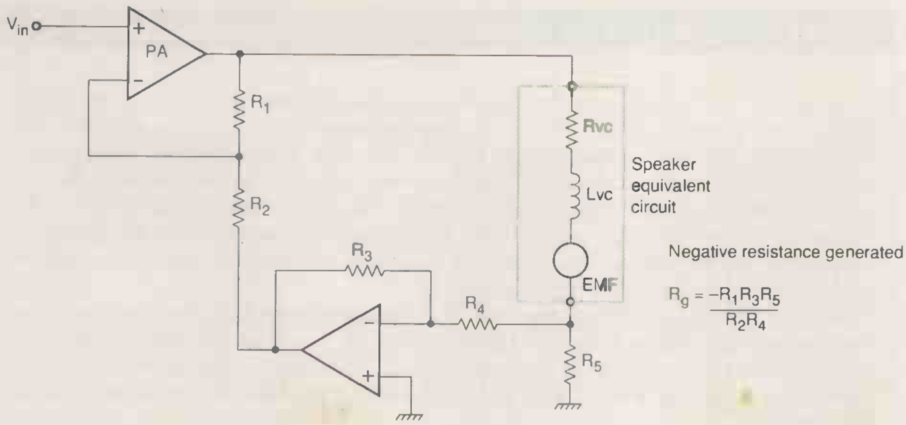


Fig. 3. The usual circuit[10,11] used to generate negative resistance for damping speaker resonance.

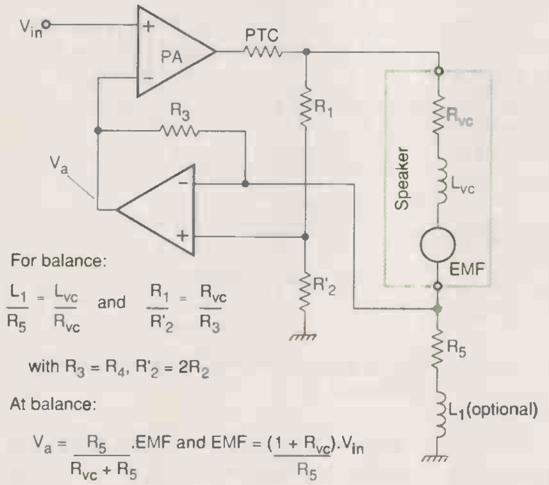


Fig. 4. A rearranged version of Fig. 3 clearly shows it is a form of motional feedback since the feedback signal (V_a) contains only velocity information when the bridge circuit is balanced. In practice 100% balance cannot be achieved because R_{vc} and L_{vc} vary, limiting the amount of servo feedback loop gain.

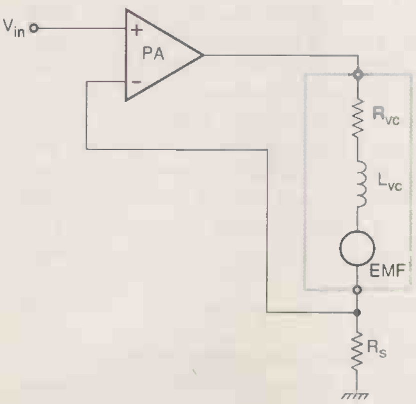
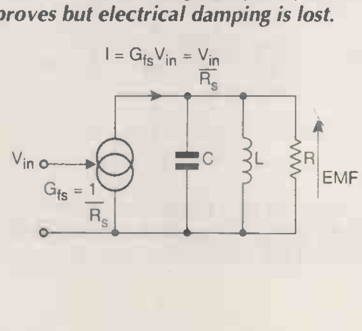


Fig. 5a. Current feedback around a speaker effectively drives the speaker with a current source as in Fig. 5b. High frequency response improves but electrical damping is lost.



tance was also understood then. The effectiveness of this type of servo loop depends on how close the negative resistance can be made to the voice coil resistance.

Figure 3[10,11] shows the technique used to generate negative resistance using positive current feedback from the current sensing resistor R_5. Resistors R_1 and R_2 provide negative feedback which obviously needs to be greater than the effect of the positive feedback to remain stable. This circuit can be rearranged to **Figure 4** by moving the op-amp and doubling R_2.

When the bridge formed by R_1, R_2, R_{vc} and R_5 is balanced, the voltage V_a becomes proportional to the voice coil's back emf. Voltage V_a is then compared to the amplifier's input voltage. Error voltage is amplified and fed to the speaker. This forces the speaker back emf to follow the amplifier's input voltage. Since the back emf is proportional to the voice-coil velocity, assuming constant flux through the coil, then this produces motional feedback where voice-coil velocity is controlled.

For constant sound pressure – i.e. flat response – a constant cone acceleration is needed. This means a velocity controlled system will need to have the input signal compensated by boosting low frequencies as frequency falls at +6dB per octave from the frequency where velocity feedback becomes effective.

Voice coil inductance limits this form of velocity feedback to under 1kHz. Because speaker resonance is suppressed and the rate of compensation is predictable, velocity feedback greatly simplifies equalisation and reduces sensitivity to speaker and enclosure variations. Of course, this assumes velocity control itself is stable.

Temperature change in the voice coil places a practical limit on the amount of servo loop feedback. In references 10 and 11, around –50% of the voice coil resistance is used. Although this does not provide very much servo loop gain it does allow a smaller cabinet.

For example this –50% resistance factor halves the net resistance for damping the speaker. This halves the Q which allows the enclosure volume to be reduced by a factor of 3. Most small enclosures typically have a roll-off around 100Hz and need an equalisation slope of +12dB per octave. With this boost the equaliser gain reaches around 20dB at 30Hz. My observations show this level of boost does not generally increase the peak level of the music since most music, apart from organ works does not contain high levels of these very low frequencies.

Temperature compensation

Yamaha uses temperature compensation in their systems to stabilise the damping resistance. As the amount of negative resistance approaches 100% of R_{vc} the sensitivity of damping to temperature increases rapidly.

Fortunately, when R_{vc} increases the system still remains stable. As an example, with 80% negative resistance, set at room temperature, and assume the speaker's voice coil rises by 100°C. Voice coil resistance rises by 40% and the net damping resistance rises from 20% to 60% of R_{vc}. Therefore the enclosure Q increases by a factor of three.

Assuming the original Q was 0.5, then when hot it will rise to 1.5, making it sound boomy. So, to operate at these levels of damping, some form of thermal compensation is needed.

A simple yet effective form of compensation was developed by winding resistor R_5 with copper wire. It is self heated by the speaker current flowing through it. Complete compensation would be achieved if the wire heats to the same temperature and at the same rate as the voice coil.

I wound about 300mm of 36swg (0.2mm) enamelled copper wire on a 0.5W high-value resistor and soldered to the 'cold' (–) speaker terminal. This provided cooling from speaker movement and improves thermal tracking. Effectiveness of temperature compensation can be checked by operating close to 100% negative resistance and running at medium power to check if the system remains stable.

Under-compensation, rather than over-compensation, should be aimed at to avoid a positive feedback situation. With 100% negative resistance the system is so sensitive to offset voltage that the dc operating point cannot be stabilised. Eventually the amplifier goes to one of the rails, delivering full supply voltage (dc) to your speaker and destroying it in seconds unless protection is provided.

Always provide some protection when adjusting the negative resistance by varying either R_1 or R_4. A polyswitch can be mounted at the output of the power amp just prior to the feedback take-off point by R_1. I used two 12V 12W car lamps. Inserted at this point, its additional resistance will not upset the balance. With protection in place you can find the limit for stability, then reduce the setting for a safety margin or to the level of damping required for your enclosure.

A better method of temperature compensation has been devised by winding R_5 adjacent to the voice coil. A non-inductive resistor can be formed by looping the wire back on itself. Since the centre of the voice coil gets hotter than its ends, compensation will be close but still slightly under-compensated.

Only three wires need to be run to the outside world, the two existing current carrying wires plus a third low current feedback wire which can be twisted around the 'cold' speaker flexible then soldered to a third speaker lug.

Manufacturers could do this at little extra cost. Yamaha provides compensation modules to suit various speakers.

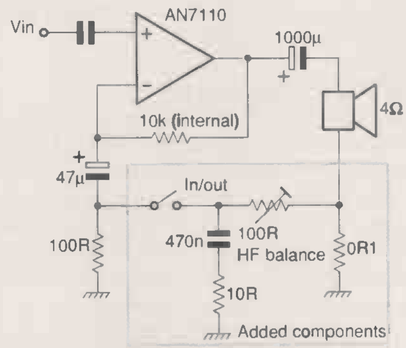


Fig. 6a. Modifications to a radio-cassette player are easy to retrofit.

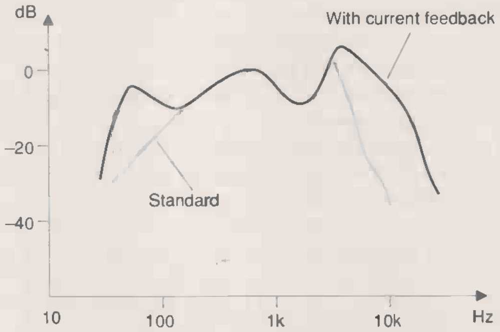


Fig. 6b. Response is improved at both high and low frequencies.

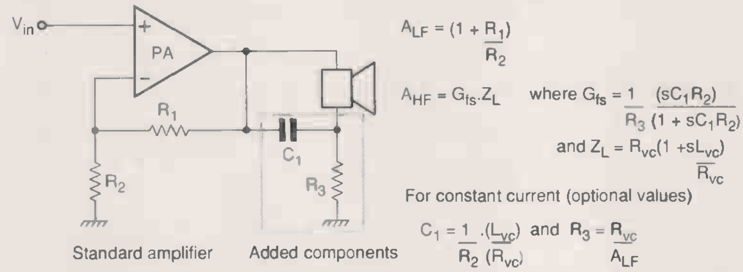


Fig. 7. Current feedback can be disabled at low frequencies for electrical damping of speaker resonance. A conventional power amplifier can be easily modified by adding R_3 and C_1. Reduce R_3 until the high frequency end is sufficiently boosted.

Negative current feedback

Although this technique does not give motional feedback or reduce damping, it does improve the high-frequency response of the speaker.

Also, as suggested by J.R. Allison[12], it gives inherent short-circuit protection and changes in voice coil resistance with temperature and changes in voice coil inductance with low frequency excursion, have no effect on speaker response. This is because speaker current is controlled rather than speaker terminal voltage.

The technique is very easy to implement, **Fig. 5**. It requires only one resistor for current sensing. The feedback loop forces the current through the speaker to follow the input voltage. The equivalent circuit of the speaker shows R_{vc} and L_{vc} effectively overcome by the current feedback loop.

At resonance, the impedance of the parallel *LC* components rises and hence the emf rises at resonance. At this point cone movement is limited only by mechanical losses mainly in the speaker suspension plus cabinet acoustic losses and cone radiation. Effectively, there is no electrical damping with this technique. Before mounting, *Q* will be in the range of 2 to 6. It can be reduced using an equaliser but enclosure size will be severely limited by sensitivity to variations and therefore seems unsuited to small enclosures.

Figure 6 shows a circuit used on a small radio-cassette to improve the speaker's high frequency response. Current feedback provides 6dB/octave boost. Another 6dB/octave is added using the 470nF capacitor. Together these give a similar effect to adding a tweeter.

More bass was noticeable. Overall these modifications gave a remarkable improvement. There was a problem with hum due to the way the tracks were run to the power supply and speaker. This was reduced by adding more capacitance to the supply reservoir capacitor. This highlights the need for more care with current feedback system earthing.

Figure 7 allows good electrical damping at low frequencies and current feedback at high

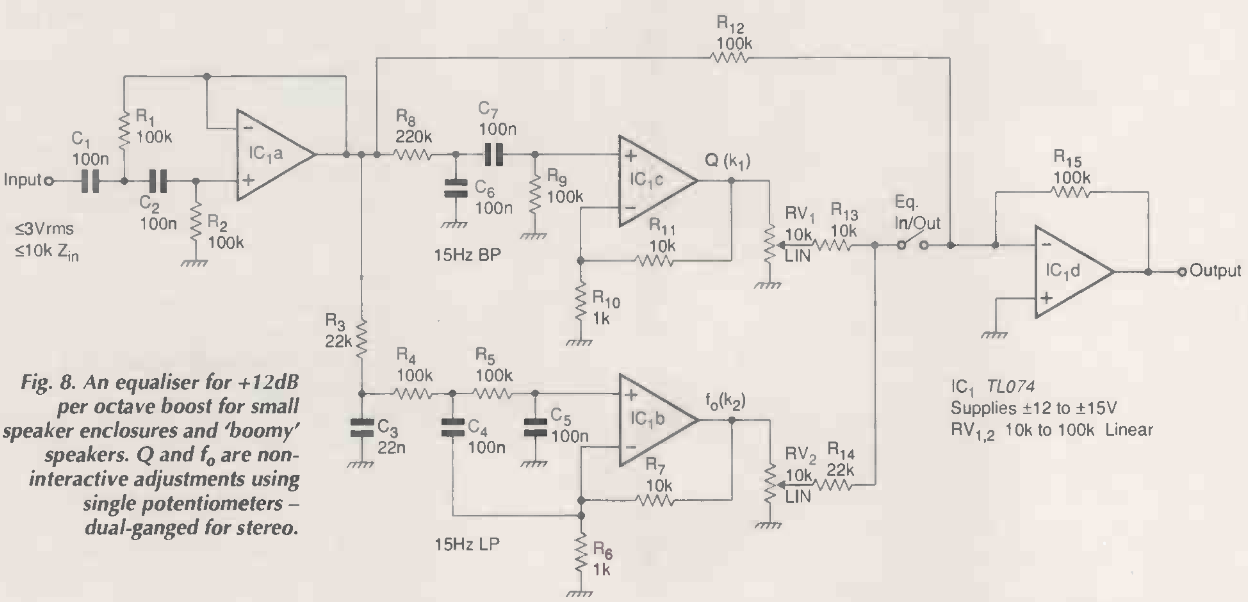


Fig. 8. An equaliser for +12dB per octave boost for small speaker enclosures and 'boomy' speakers. Q and f_o are non-interactive adjustments using single potentiometers – dual-ganged for stereo.

frequencies for improved high frequency bandwidth. Unfortunately, inherent current limiting is lost, so the power amplifier needs the usual output protection. Earthing problems are reduced since C_1 disables current feedback at low frequencies.

Current feedback pushes the midrange output level higher starting from where voice coil inductance normally causes current to fall. This means a flat response speaker will show lift at mid frequencies. However, speakers that normally showed slow roll-off in the midrange can be compensated to near flat with current negative feedback.

Wide-range dual-cone speakers and those with a metallised dust cap work well with current feedback – some to 15kHz – so the tweeter can be omitted. Since the dispersion angle narrows at high frequencies a high frequency diffuser mat be needed.

Voltage versus current

A voltage driven speaker can also be boosted to extend the high frequency response. However, with current feedback, unlike a voltage driven circuit, voice coil inductance variations due to large excursions from low frequencies do not amplitude modulate the high frequencies. Doppler distortion, where large excursions frequency modulate higher frequencies, can in theory be removed by modulating the clock of an audio delay line with V_a in Fig. 4.

Figure 8 is the circuit of an equaliser similar to B.J. Sokol's but modified for lower noise by boosting the gains of the band- and low-pass filters.

Figures for Q of 0.2 to 5 can be compensated and the frequency range for f_o is about 30 to 150Hz. Resistor R_3 and C_3 have been added to give a few degrees more phase shift at f_o to give a better null with RV_2. Set up is described in reference 8.

I have found set-up can be done by ear. First set RV_1 to minimum, then adjust RV_2 for the best notch, the greatest loss of low frequencies. Next, increase RV_1 until ‘booming’ becomes noticeable, then reduce the setting slightly for the desired effect. ■

Calculations for high-pass filtering and equalisation functions.

Given a second order speaker function and equaliser functions:

$$\text{Speaker} = \text{HP} = \frac{S^2}{S^2+\left(\frac{\omega_{01}}{Q_1}\right)S+\omega_{01}^2}$$

$$\text{Equaliser} = \text{EQ} = \frac{S^2+\left(\frac{\omega_{02}}{Q_2}\right)S+\omega_{02}^2}{S^2+\left(\frac{\omega_{03}}{Q_3}\right)S+\omega_{03}^2}$$

Let $\omega_{02}=\omega_{01}$ and $Q_2=Q_1$ to allow cancellation

$$HP\times EQ = \frac{S^2}{S^2+\left(\frac{\omega_{01}}{Q_1}\right)^2 S+\omega_{01}^2}\times\frac{S^2+\left(\frac{\omega_{01}}{Q_1}\right)S+\omega_{01}^2}{S^2+\left(\frac{\omega_{03}}{Q_3}\right)S+\omega_{03}^2}$$

$$\text{Hence, } HP\times EQ = \frac{S^2}{S^2+\left(\frac{\omega_{03}}{Q_3}\right)^2+\omega_{03}^2}$$

The equaliser function can be synthesised using a LP, BP and flat functions[6] where

$$LP = \frac{\omega_{03}^2}{S^2+\left(\frac{\omega_{03}}{Q_3}\right)S+\omega_{03}^2}$$

$$BP = \frac{\omega_0 S}{S^2+\left(\frac{\omega_{03}}{Q_3}\right)S+\omega_{03}^2}$$

$$Flat = \frac{S^2+\left(\frac{\omega_{03}}{Q_3}\right)S+\omega_{03}^2}{S^2+\left(\frac{\omega_{03}}{Q_3}\right)S+\omega_{03}^2}$$

So, $EQ = Flat + k_1 BP + k_2 LP$

$$EQ = \frac{S^2+\left(k_1+\frac{\omega_{03}}{Q_3}\right)S+(1+k_2)\omega_{03}^2}{S^2+\left(\frac{\omega_{03}}{Q_3}\right)S+\omega_{03}^2}$$

Equating to EQ above gives

$$k_1+\frac{\omega_{03}}{Q_3}=\frac{\omega_{01}}{Q_1} \text{ and } (1+k_2)\omega_{03}^2=\omega_{01}^2$$

$$\text{Hence } k_1=\left(\frac{\omega_{01}}{Q_1}\right)-\left(\frac{\omega_{03}}{Q_3}\right) \text{ and } k_2=\left(\frac{\omega_{01}}{\omega_{03}}\right)^2-1$$

Note: k_1 sets damping of the speaker and k_2 is set to suit the speaker's resonant frequency. LP and BP should have equal ω_0's and Q's and ω_{03} needs to be lower than $f-3dB$ of the equalised system.

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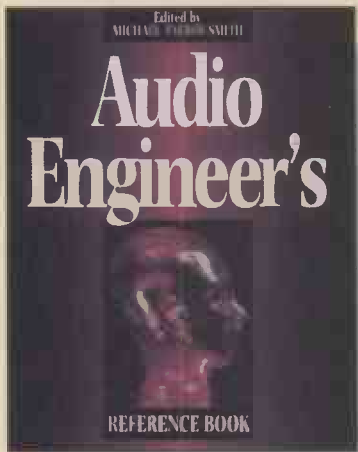

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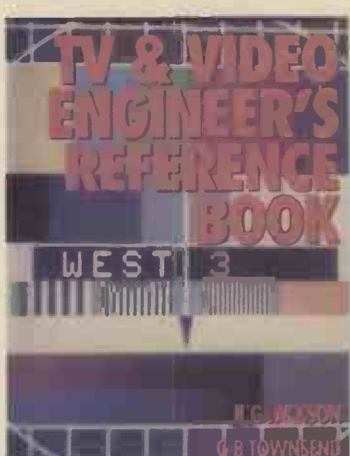

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Understanding emi filters

Cyril Bateman expels the myths surrounding emi filters and highlights the inadequacies of emi filter CAD alternatives.

With the arrival of the CE mark emc regulations, most equipment previously exempt from compliance is now included. Due to the all-embracing nature of the EC proposals, electronic circuit designers need to be emc aware. A good starting point is to obtain a copy of the DTI's Business in Europe booklet[1].

For many products, use of improved screening or changing layout will render the system compliant. But whenever connection is made between a circuit and the 'outside world', the use of low-pass electromagnetic interference filters becomes an essential part of the design process.

Unfortunately emi filters, and the measurement methods used to derive their performance claims, are not generally well understood. This resulted in the comment by one respected engineer, that "fitting filters is more an act of faith, than one of certain benefit".

What then is an emi filter?

A conventional emi filter is a collection of capacitors and or inductors, designed to pass the required signal or power, but to attenuate higher frequency unwanted energy. It cannot dissipate energy, hence can only attain its insertion loss performance by reflecting this unwanted energy back to its source.

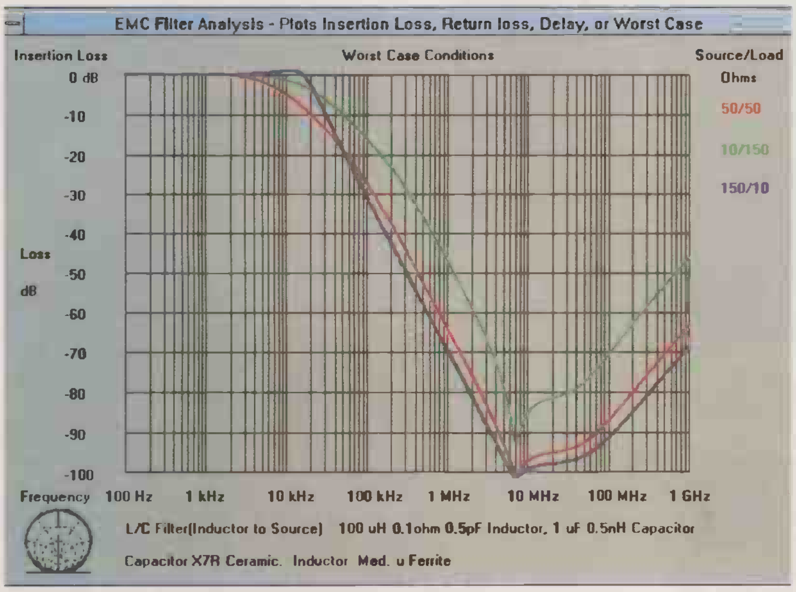


Fig. 1. Effect on insertion loss of L/C filter in differing source/load impedances. This demonstrates significant change of insertion loss from the 50/50Ω specification nominal.

This reflected energy is expressed as 'return loss,' being the ratio of reflected voltage to incident voltage. It is expressed in decibels.

Performance claims for emc filters are conventionally based on source and load impedances of 50Ω each, and the specified load current. Change of source or load impedance changes the filter's insertion loss, **Fig. 1**.

Load current considerations

Load current is important in filter design. Most interference filters include at least one inductor, and except when twin-line 'bucking mode inductors' are used, the value of this inductor can be load current dependent, thus changing the insertion loss of the filter.

While it is good practice to underrun an emi filter to improve reliability, the resulting insertion loss changes should be taken into account. For size and cost reasons, many filters are designed such that at the specified dc current the inductance is half that of the no current inductance. Consequently performance will differ from the catalogue claims, depending on the actual source, load impedances and current loading applied.

With certain combinations of filter construction and end use conditions, the –3dB cut-off frequency and cut-off rate can change such that the filter becomes unduly 'peaked', giving insertion gain. A wrong filter choice can *increase* the transmitted emi level, Fig. 1.

Measurement versus simulation

The best way to gain understanding of any device is to measure it. Assuming access to a suitable spectrum analyser and generator or network analyser, this is feasible. For frequencies higher than 10MHz and/or insertion losses greater than 50dB are an exception however. Design of the test jig and set-up requires specialist knowledge, to house the filter and maintain an acceptable 50Ω system[2]. Measurement of filters in differing source/load impedances is also possible, but difficult. It requires the use of wide-band impedance converting transformers which are not easily obtained.

With all these practical and cost difficulties, why not just simulate the filter's behaviour?

To be meaningful, any simulation must be based on real life parts. And the models used must be supportable against actual measured results. With the wide frequency and dynamic ranges required, the capacitors and inductors used to build practical filters simply do not behave in an ideal manner. Both component value and losses are very much frequency dependent. These variables cannot be supported by established simulation software, resulting in overly optimistic insertion loss predictions above 100kHz and overly pessimistic insertion loss predictions above the capacitor series/inductor parallel resonance frequencies, Figs 2,3,4,5.

S parameters

The S parameter measurement technique was developed to measure high frequency transistors, giving distributed rather than lumped results. It also facilitates subsequent circuit analysis and avoids the need to provide open and short circuit conditions during the actual measurement.

S parameters are measured using a vector network analyser. The test jig is inserted within a 50Ω transmission line, previously characterised at this insertion point by calibrating using known open, short, and matching loads.

Fundamental to this concept, the number of measured parameters for each frequency is the square of the number of device ports. For example, a capacitor to earth or series inductor has two ports, hence four pairs of magnitude and phase parameters per frequency.

Example 1. 1μF capacitor at 1MHz.
Example 2. 100μH inductor at 1MHz.

	S_{11}	S_{21}	S_{12}	S_{22}
1.	0.996-179.6	.0007-57.0	.0007-57.0	0.996-179.6
2.	0.988 9.0	0.157-80.9	0.157-80.9	0.988 9.0

S parameters provide the only way to simulate a filter using actual measured frequency dependent variables. However the vector network analyser is expensive, not easily managed, and properly characterised component test jigs are essential.

Interested readers should obtain copies of Hewlett Packard application notes 95-1 and 154, and also ref. 2.

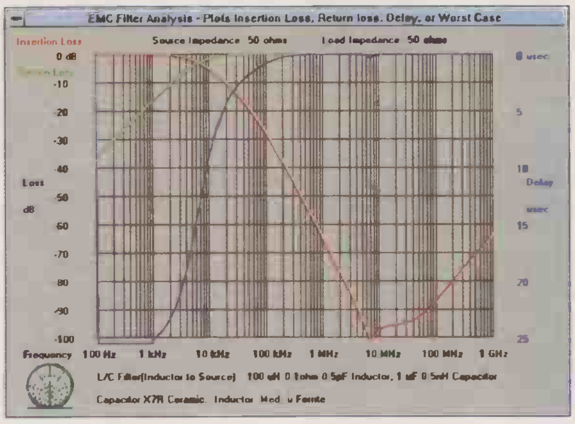


Fig. 2. Response of LC filter – inclusive of capacitor/inductor losses in 50/50Ω source/load. Insertion loss plot, in red, shows realistic attenuation attained at all frequencies. Group delay plot, in blue, results from calculating 50 points in each decade of frequency. The return loss plot, green, clearly shows that insertion loss is attained only by reflection.

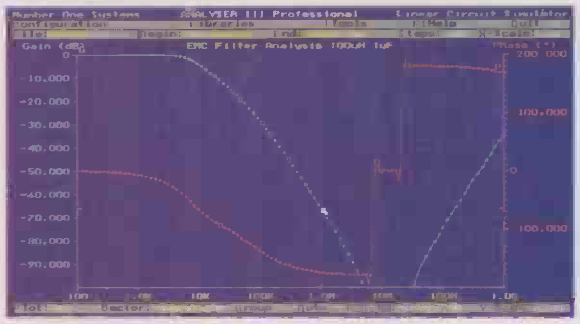


Fig. 3. Response of L/C filter – exclusive of capacitor/inductor losses in 50/50Ω source/load. Insertion loss plot in blue shows overly optimistic attenuation above 1MHz. Analyser III can calculate relative insertion loss (gain) or S_{21}, by menu selection. Since the plot styles are pre-configured, the phase response is automatically drawn. Analyser III plots a maximum of 100 frequencies, group delay must be three decades scan maximum.

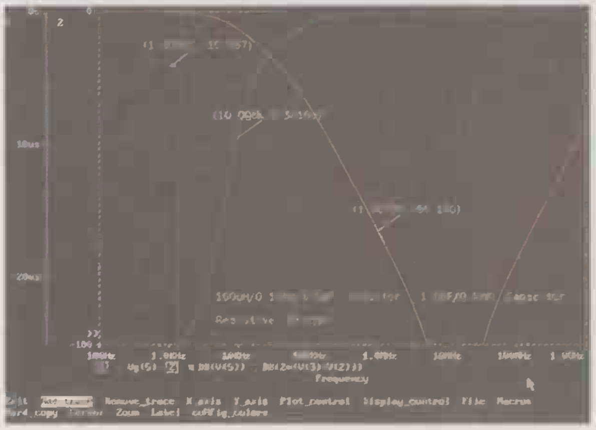


Fig. 4. Response of L/C filter – excluding capacitor/inductor losses in 50/50 source/load. This plot has been calculated using the resistive reflection bridge as described. Insertion loss plot, in red, shows overly optimistic attenuation with frequency above 1MHz. Since PSpice calculates only node voltages, insertion loss must be instructed as 'DB(V(5)'. Similarly, to plot return loss (green), the instruction 'DB(2(V(3)-V(2)))' must be issued. Group delay, blue, is calculated from the insertion loss node, instructed as $V_g(5)$.*

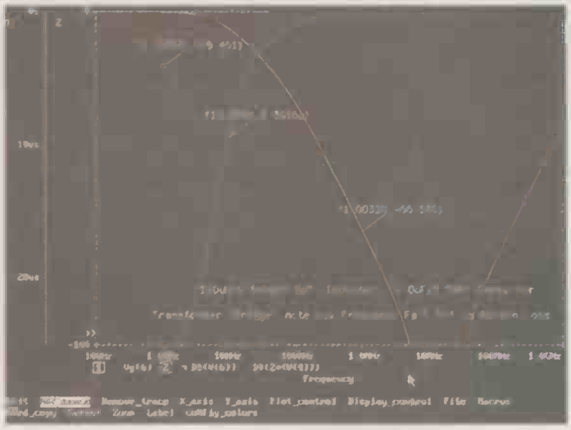


Fig. 5. Response of L/C filter – excluding capacitor/inductor losses in 50/50Ω source/load. This plot is calculated using the transformer coupled reflection bridge as described. Notice the low frequency limit on return loss due to the transformer's primary impedance. The insertion loss plot, in red, shows overly optimistic attenuation with frequency above 1MHz. Since PSpice calculates only node voltages, insertion loss must be instructed as 'DB(V(6)'. Similarly to plot return loss, green), the instruction 'DB(2(V(4)))' must be issued. Group delay, blue, is calculated from the insertion loss node, instructed as $V_g(6)$.*

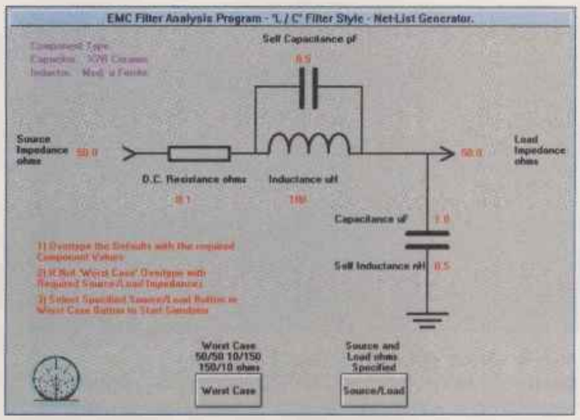
For consistency, all simulations in this article model a filter having the same component values, a typical 1.0μF surface mount *X7R* ceramic chip (0.5nH self inductance), and a 100μH inductor wound for minimal self capacitance of 0.5pF on a medium μferrite toroid, **Fig. 6**.

Specially designed discoidal ceramic capacitors and feedthrough capacitors have smaller self inductances.

High dielectric constant ceramic capacitors have resonant modes which result from physical size and dielectric constant, above 10MHz. In practice these must be measured using 'S parameters', or ignored for simulation.

Spice derived simulators are designed to model in the time domain and cannot accept frequency dependant variables.

Fig. 6. Illustrating the user friendly 'Net-List' generating screen in my emi filter calculation program. The only instructions needed are on screen. Just overtype the defaults with required values. Simple prior menu selection provides other similar screens prepared for each filter style. Simulation is started by 'clicking' on either 'Worst Case' or 'Source/Load' buttons as required.



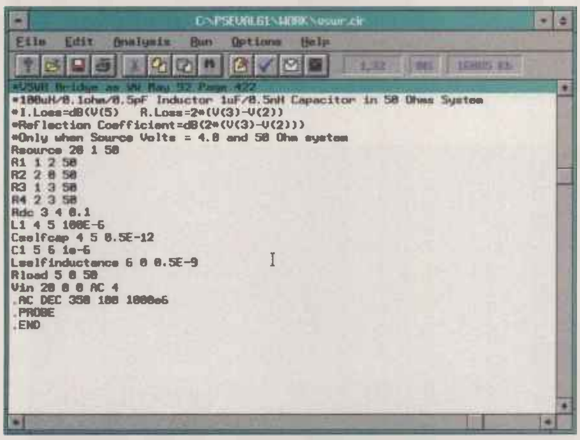


Fig. 7. Net-List used with PSpice to model the resistive bridge and L/C filter. The filter being simulated commences at node 3, insertion loss is calculated as 'DB(V(5)'. Return loss calculation is the voltage drop across R_4, calculated as 'DB(2*(V(3)–(V(2)))'. Group delay is calculated from the insertion loss node, instructed as $V_g(5)$.

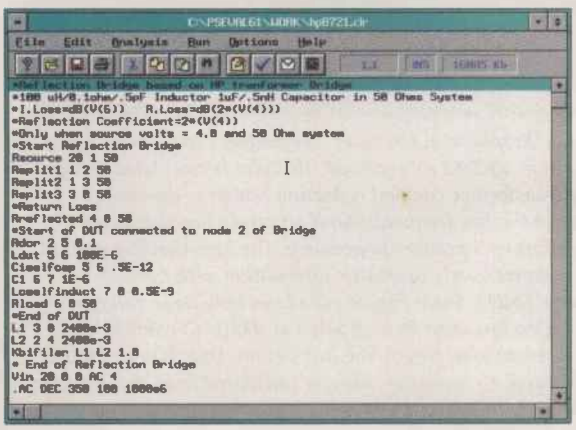


Fig. 8. Net-List used with PSpice to model the transformer bridge and L/C filter. The filter being simulated commences at node 2, insertion loss is calculated as 'DB(V(6)'. The return loss calculation is the voltage drop across R_4, calculated as 'DB(2*(V(4)))'. Group delay is calculated from the insertion loss node, instructed as $V_g(6)$.

Most available frequency domain simulators make no provision, or require the use of measured S parameters[3,4], see box 'S parameters'.

Simulation using measured S parameters, making true allowance for all component variables, as used for microwave design, is undoubtedly the ideal method. Unfortunately such parameters are not published for the components normally used to make a filter.

Measurement of filter component parts S parameters requires much skill and time to characterise and de-embed the test jigs needed for the various components. Due to the extreme mismatching of these parts to the test system source, a vector network analyser with 12-term error correction also called 'full 2-port correction' is essential. From personal experience, the resulting costs are difficult to justify – even for the component maker.

To solve these problems, I have written a new emi filter analysis program for Windows. It is able to accept a database of equation models for the frequency dependant variables, tanδ and *K* for capacitors and *Q* and μ for inductors.

Using this new method of frequency domain simulation gives predictions close to measured values, automatically provides the required insertion loss, return loss and group delay results, yet requires no prior knowledge from the user or use of S parameters. It avoids using convergence techniques, hence unlike Spice based simulators, it cannot misconverge. It has a user friendly net-list generator screen, providing an easy to use, realistic and cost effective simulation.

Loss models for *X7R* ceramic capacitors and medium μferrite toroids were used, Figs 1, 2, 6.

EMI filter fundamentals

Traditionally, emi filters are characterised for insertion loss by frequency in 50Ω systems, by circuit style – for example *C*, *L/C*, Pi, etc – total capacitance value and total inductance value at the specified load current. These are the end-of-line test parameters used in manufacture.

Insertion loss (S_{21}) being the major consideration, emi filters traditionally are not designed to conform to recognised characteristics, such as Butterworth or Bessel. It is common practice to use equal value capacitors for Pi filters and equal value inductors for T styles.

To predict in-circuit performance, the user requires return loss and insertion loss at the actual source/load impedances and load current used. Both vary with end use conditions and the filter's style. Published data in this detail is not generally available, but it can be simulated, Fig. 2.

Applications using multiple-frequency or non-sinusoidal signals, also need 'group delay' to minimise phase distortion.

Envelope degradation of digital signals transmitted through the filter is estimated by combining Fourier analysis of the source waveform with the filter simulation results, followed by reverse FFT.

Unlike resistive attenuators, which provide insertion loss by dissipating energy, emi filters are designed to provide a zero loss 'low-pass' characteristic while passing the required current. Resistive elements are minimised thus energy cannot be dissipated.

Insertion loss in low-pass emi filters results from mismatched impedances, with the filter reflecting the emi back to its source. With a poorly chosen filter, this reflected energy combined with the incident energy, can be much greater than with no filter. Insertion loss depends on the filter component values and the source/load impedances and can result in gain rather than loss at certain frequencies, hence the quote mentioned earlier, Fig. 1.

This reflected energy can be measured or simulated either as return-loss or reflection co-efficient. The return-loss concept is the more useful, being the attenuated level of the reflection compared to the energy incident on the filter. The sign of this reflected energy relative to incident energy, depends on the input impedance of the filter and the circuit source impedance[6].

With short cable lengths and pass-band frequencies, the reflected wave generally adds in phase with the incident wave at the filter and continues voltage additive back to the source. Unlike some recently repeated claims[7], this aspect of transmission lines does not disappear with short cable lengths. It can be measured or simulated, Fig. 2,4,5.

To measure or simulate 'return loss' the forward and return signals must be separated. Ideally using a network analyser with an S-parameter test set to measure both signals. If phase can be ignored, you can use a variation of the Wheatstone bridge, preferably with a spectrum analyser or less accurately, using an oscilloscope, see box 'Bridges'.

My Hewlett Packard *8721A* directional bridge, specified for use from 100kHz to 100MHz, was used as a basis for the transformer bridge simulation model. By winding larger bifilar transformers, successful audio frequency versions have been produced.

With the reflection port terminated by its characteristic impedance, this bridge is calibrated by applying in turn at the load port, open/short circuits and the load impedance to be used. Respective voltages at the load and reflection ports are noted. Return loss is measured at the reflected port in decibels relative to the calibration open/short voltages. Insertion loss is measured, at the load impedance, with the filter inserted immediately between the bridge and this load, in decibels relative to the calibration load voltage noted without the filter, see box 'Bridges'.

Insertion loss measurement - method

For a detailed discussion on filter measurement methods, read 'Measuring insertion loss of lowpass rfi filters'[2]. It discusses the MIL-STD method and the use of S parameters. The original test specification for emi filters was MIL-STD-220[8], based on concepts developed by Beattie[9]. It forms the basis for all subsequent filter specifications.

Fundamental to this method is the use of two 10dB attenuators[8]. These define the source/load impedances and reference plane, close to the filter being measured. The system is calibrated for 0dB loss by connecting these attenuators together and measuring the load voltage at the required frequencies, called the 'filter out' condition.

The filter to be measured is inserted between these attenuators and the measurements are repeated, called 'filter in' condition.

Insertion loss of the filter is defined as 'the ratio of voltage measured immediately beyond the point of insertion with and without the filter inserted', expressed in decibels, ie

20log(*filter in/filter out*).

Any connecting cables between these attenuators, to be less than 0.05 of a wavelength – less than 10cm for 100MHz – and common to both filter-in and filter-out conditions.

Due to the high vswr of typical emi filters – many thousands to one in the stop band – this measured result obviously includes the cable and jig mismatch losses with that of the filter. This is because the 'filter out' condition provides 50Ω matching. True measurement requires test jigs to be electrically short and 50Ω impedance. It is not possible to de-embed the filter from this jig/filter measured value.

Insertion loss simulation

Insertion loss on its own can be calculated by any circuit simulator using source and load resistors with the required filter circuit, hence replicating the MIL-STD method, and plotting relative gain (–6.02), Fig. 3. Accuracy depends on the simulator's models.

Depending on the chosen circuit simulator, simultaneous return loss and insertion loss can be simulated using one variation of the Wheatstone reflection bridge. Four differing simulators were used for this article.

Dos

Hewlett Packard 'RF & Microwave AppCAD' – S_{21}, S_{11} etc but not return loss[4], box 'S Parameters'.

PSpice 6.1 eval downloaded from Internet. Both model bridges[10], Figs 4,5,7,8.

Analyser III Professional from No. One Systems. Gain, S_{21} and phase, not return loss[3], Fig. 3.

Windows

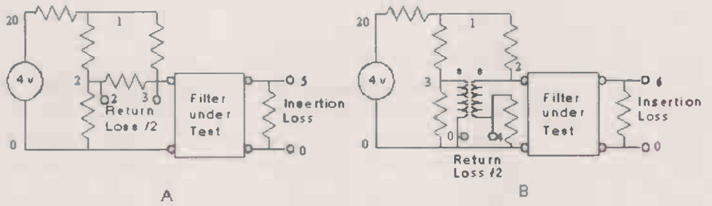
PSpice 6.2 evaluation downloaded from Internet. Either model bridge[10].

My own 'EMC Filter Analysis' program. Insertion loss, return loss and delay, all standard[5], Fig. 1,2,6.

Bridges in emi filter measurements

A. If three limbs of a Wheatstone bridge are made exactly 50Ω and the unknown is the remaining limb, then the unbalanced voltage of this bridge can be measured between the centre point of the two 50Ω resistor arm and the unknown, hence the resistive bridge circuit as used in Fig. 4. See also *Electronics World & Wireless World* May 1992 p.422.

B. In an S parameter co-axial measurement system the required voltage is balanced to earth hence difficult to measure, as for Fig. 5. A bifilar 1:1 balun transformer converts it to an unbalanced, easily measured voltage. This same arrangement works for a simulator.



Technical support

The emi filter calculation software used to produce plots shown in Figs 1, 2 & 6 of this article is available from the author at £100 fully inclusive. This price includes VAT, postage, and technical support. A demonstration disk is also available at £7 fully inclusive, the price of which will be refunded on purchase of the full package. Note that this software runs under Windows version 3x. Please send a cheque or postal order payable to Cyril Bateman Engineering, to Cyril Bateman at Nimrod, New Road, Acle, Norfolk NR13 3BD.

Measuring filter components

Capacitance measurement to ground can be influenced by the presence of inductance in series with the through terminals. Regardless of filter style, to measure total capacitance, link the through terminals together and measure from them to the common ground. Remove this link.

Series inductance measurements can be influenced by capacitance to the ground terminal especially with pi filters.

Regardless of filter style, to measure series inductance, connect the filter common ground terminals to the *LCR* meter guard terminal and measure inductance between the through terminals. If your meter has no guard terminal, this measurement is not possible.

If your *LCR* meter has no current bias facility, then an adaptor, comprising series capacitors to block the bias from the *LCR* meter together with isolating inductors to supply the required bias, is needed. This technique has been built and used by the writer for up to 20A dc bias and 250MHz frequency measurements.

Simulators such as PSpice, with ability to subtract node voltages, can use either bridge model, see PSpice netlists, **Figs 7,8**.

Simulators having single ended outputs only should be able to use the bifilar-wound transformer version, to generate an unbalanced output node.

In simulations the resistive model bridge, if not real life, has unlimited frequency range. The transformer version can only be used at frequencies where the impedance of the primary is large compared to the source impedance, Fig. 5. With either model bridge it is essential the correct source voltage and output instructions are used, to ensure a calibrated bridge, Fig. 7,8.

Filter component values

For standard catalogue filters, the required component values should be obtained from the maker. However, assuming the component parts or a sample filter is to hand, these values can be measured, by following correct measurement techniques.

Obtaining the actual inductance value at the required load current is more difficult, due to the need to bias the inductor being measured, see box 'Measurement of filter components'.

In summary

This article describes the behaviour of emc filters and demonstrates workable means to measure or simulate this behaviour, and thus gain a working knowledge of emc filters.

By making measurements or simulations as described, many of the peculiarities of emc filters will be understood, resulting in better application of these essential components.■

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Load protection

Frantisek Michele explains the benefits of protecting sensitive loads from power anomalies using a combination of series and parallel devices.

The operating speed and sophistication of electrical and electronic devices have increased enormously during the past decade. Unfortunately, the vulnerability of these devices to ac power line anomalies has increased at an even faster rate.

This article discusses the nature of these power line anomalies, their effect on electronic devices, the kinds of protection available, and choosing effective protection.

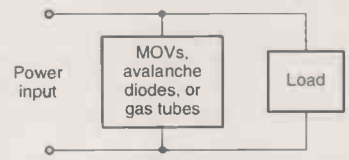


Fig. 1. Single-stage surge control circuit consists of a high energy device installed in parallel to the line. This device clamps the voltage at a predetermined level.

What causes the damage?

The most common cause of large, high voltage spikes or surges is lightning. Although a close lightning strike can cause such immediately obvious damage as blown fuses, scorched insulation, and smoking equipment, many less obvious kinds of damage can result.

Damage from any kind of ac power line anomaly falls into one of three categories: destructive, disruptive, or degrading. The first and most obvious – destructive – could result from a high voltage spike generated by lightning, or by a utility fault.

The second level of damage is disruptive and may be far from obvious. A disruptive spike might result from utility switching or from turning heavy electrical loads on/off in the vicinity of the affected devices.

Disruptive spikes are difficult to assess. They are evidenced by unexplained errors in operation of the disrupted device, inconsistent results, excessive downtime, and unusually frequent maintenance requirements.

Even more subtle is the long-term degrading, caused by the accumulation of many small spikes over time. Excessive downtime and significantly shorter product life can result from the degrading spikes. Although subtle and sometimes hard to detect, degrading spikes are no less costly in the long run than a single destructive spike.

Causes of anomalies in power mains environments

Potential anomaly	Source: Public utility	Load switching	Lightning	Shop/field equipment	Office equip	Atmospheric	Automobiles
Undervoltage	•	•					
Energy surges	•	•	•	•			
Single phasing	•		•				
RFI		•		•	•	•	•
EMI				•			
Noise				•	•	•	
Induced transients			•				

Various power line anomalies and their causes are listed in the Table.shown on the prevous page.

Spike protection

The traditional single-stage surge control circuit, Fig. 1, consist of a high energy device installed in parallel to the line to divert or bleed off the energy of the spike to ground. This device usually comprises metal oxide varistors, or MOVs, avalanche diodes, or gas tubes. These components are designed to clamp the voltage at a predetermined level.

Although inexpensive and easy to install, these devices have several disadvantages. First, their clamping level is a function of the spike rise time. The higher and faster spike, the higher their clamping voltage. A large spike can exceed the safe voltage level and cause degradation, disruption, or destruction.

Single-stage parallel surge suppressors have further limiting characteristics. If this component has a large surge capacity, it is relatively slow to react. Conversely, if it has a fast reaction time, it has a low capacity or short life.

Parallel protectors can be improved by adding stages of different types of protective component. But multi-stage parallel designs are still ineffective against major spikes.

Series-parallel alternatives

The most important characteristic of a series-parallel circuit, Fig. 2, is that it has an inductor directly in the path of an incoming spike. This element offers very little insertion loss to the wanted ac signal, but reacts in proportion to the spike, holding up the energy surge until it can be dissipated harmlessly by two parallel stages of high energy devices and fast-acting transient energy protectors.

The inductor allows the series-parallel circuit to attenuate noise effectively and requires no resetting and little or no maintenance.

Selecting proper surge protection

There are four important factors to consider in selecting the proper surge suppressor for a given application – energy capacity, speed, clamp ratio and life.

Energy capacity of the suppressor should be sufficient to withstand any surge that does not burn out the wiring. Speed in the 10ns reaction time range is generally sufficient for all high voltage spikes – with the exception of electromagnetic pulses. Yet some manufacturers claim reaction times in the picosecond range. Unfortunately, the rating of a component as measured in the laboratory is not usually representative of how fast a whole protective circuit will react when installed in a real world application.

With 1 to 10ns components, series inductors can delay high speed surges until the parallel protective components can react, making raw speed less critical.

Clamp ratio is the voltage at which a protector clamps fast, high-current surges in real life, divided by the slow rise, laboratory-tested clamp voltage at which protectors are usually rated. This ratio should stay near one for full protection. Yet most parallel protectors will have clamp ratios up to five or more when faced with severe, lightning-induced strikes, allowing potentially destructive surges to pass.

Series-parallel circuits provide greater protection because the in-line inductor slows down the slew rate of surges before they enter the load.

Life refers to the number of surges that a surge suppressor can withstand before it needs to be replaced. Some protectors such as MOVs are inexpensive to replace but need to be replaced frequently. While it is difficult to estimate exactly, a series-parallel surge suppressor can have a design life of at least ten years, even under the most adverse conditions.

If less destructive but more frequently experienced transients and emi/rfi noise are a concern, in addition to high voltage spikes, further protection is necessary. Transients and emi/rfi noise, while rarely destructive, are definitely disruptive and degrading – and can be costly in the long term.

Inexplicable errors, inconsistent results, and increased maintenance and downtime are clues that these smaller, but frequent, power line anomalies may be disrupting your operations and degrading your equipment.

Since traditional parallel circuit surge suppressors provide little protection in this respect, additional protection must be provided. This usually takes the form of isolation and filtering transformers.

Protection levels

A series-parallel circuit will often protect equipment from lightning strike effects and other high voltage spikes that make up approximately 40% of all power line anomalies. These high voltage spikes are usually destructive.

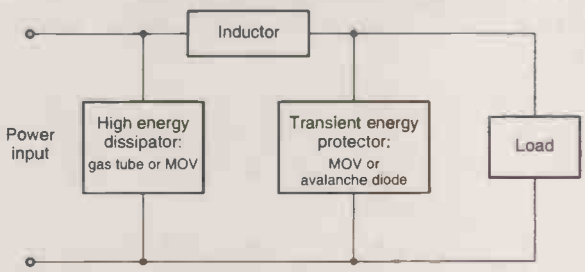


Fig. 2. Unlike a parallel circuit, series-parallel circuit has an inductor directly in the path of an incoming spike. This component holds up the energy surge until it can be dissipated by two parallel stages of high energy devices and transient energy protectors.

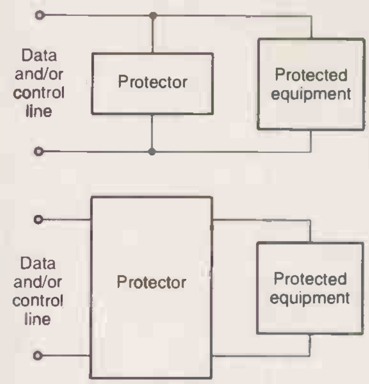


Fig. 3. A parallel transient suppressor can handle very high current levels but only for short periods. In the second diagram, series suppressors reflect incoming transient energy, reducing dissipation in any following voltage clamp wired in parallel.

Transients and noise, which are usually disruptive and degrading, make up another 45 to 50% of all power line anomalies. It is relatively easy to add filtering to series-parallel devices to eliminate transients and emi/rfi noise, providing protection against 85 to 90% of all power line anomalies. The remaining 10 to 15% of anomalies comprise voltage sags.

For lower voltage systems

As with ac power line surge suppressors, data and control line surge suppressors fall into two groups, namely parallel and series-parallel protectors.

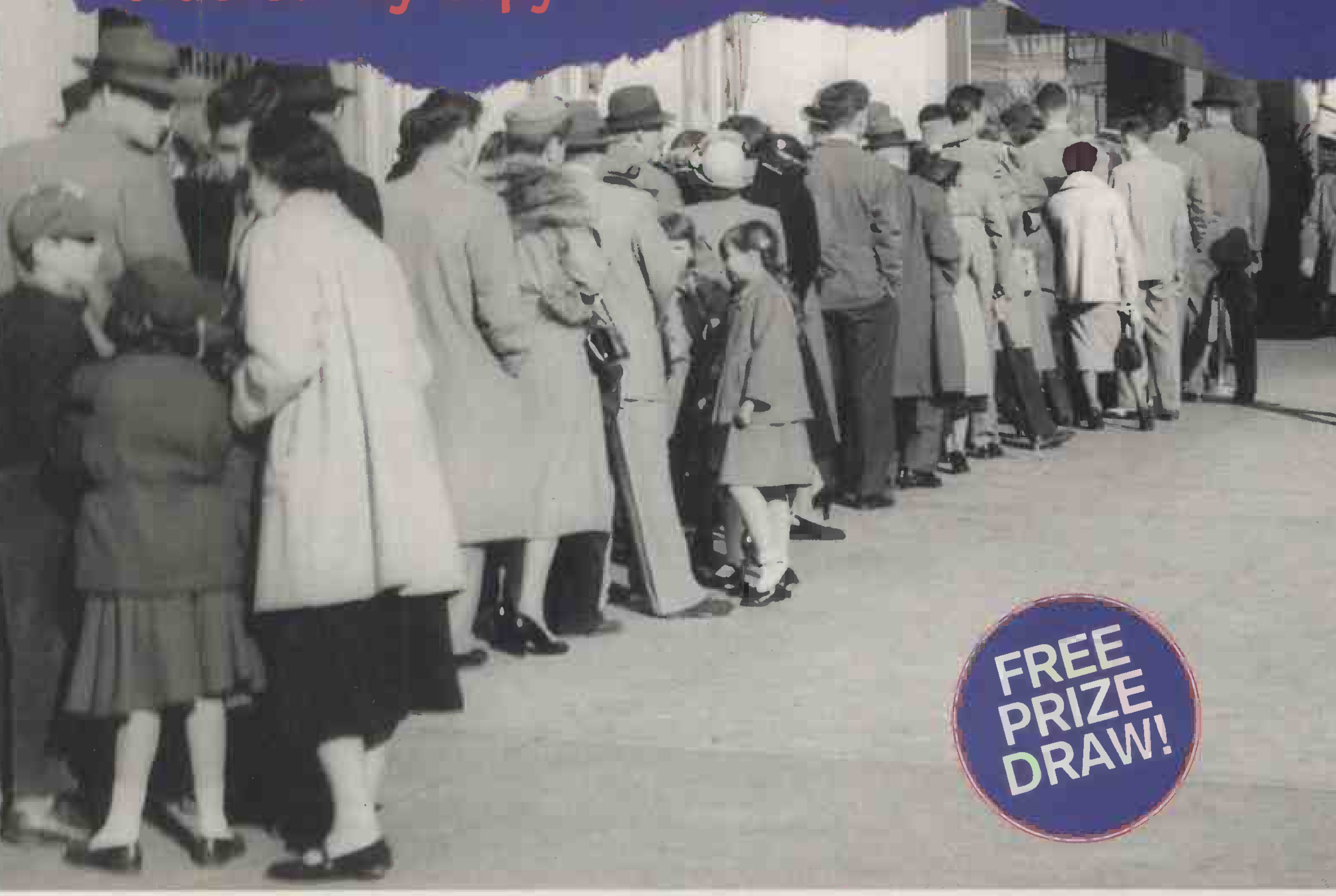
Parallel gas breakdown devices such as spark gaps and gas discharge tubes may be used for clamping. They are rugged – but slow – and are generally limited to applications involving supply voltages above 90V. In the case of a fast rise surge or impulse, over 1200V may be exceeded prior to clamping. Spark gaps and gas discharge tubes can handle such very high transient current levels for short periods.

Avalanche diodes are fast and will hold an accurate clamp voltage. However, because of their limited physical volume, a high energy transient of 1 or 2 joules can heat and destroy the diode junction, leaving the protected circuits vulnerable.

Selenium diodes, thermistors, and MOVs are also used in parallel circuits to protect the control and data lines. But, none of these devices can cover the complete spectrum of transients. Either they respond quickly but have limited power handling capability, or they can handle very large energies but do not clamp at an acceptable level. This limits their uses to low energy applications or to situations in which high voltage spikes can be tolerated.

Figure 3 compares parallel and series-parallel suppressors in data-link applications. ■

I knew I should have ordered my copy of Electronics World



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Designing an SSB outphaser

In Part 2 of this rare analysis of ssb outphasing, David Gibson investigates the effects of component tolerances on performance. David also looks at multi-section filters and digital filter implementation.

As it stands, Fig. 9 of last month's article makes no provision for adjusting the parameters. In addition to component tolerance problems there will be drifts due to temperature, ageing and so on. You need to know if these errors are significant and, if so, how they can be trimmed.

What is apparent from the Basic programs I used to derive the graphs, is that component tolerances are critical. Even slight deviations from the required values give rise to noticeable increases in phase and amplitude error. Having said that, most outphaser designs do not make any provision for trimming the component values.

To analyse the errors I will assume that the gain-setting resistors, Fig. 3d of last month's article, are exactly right, with $R_1=R_2$. You can either use tight-tolerance parts or trim the gain, if need be. Trimming to unity gain gives you a degree of orthogonality, which helps analysis as well as performance.

Having done this you will notice that the tolerances of the *RC* pairs can be applied to either component. That is, using ±1% resistors and ±1% capacitors is the same as using exact resistors and ±2% capacitors. The analysis is easier because you now need only consider the *C* tolerance when running the simulation program.

Although each component can drift independently, it is likely that temperature will affect all parts similarly. Equally, it is unlikely that you will obtain four capacitors which combine to produce the worst possible effect, as demonstrated in **Fig. 1**.

The diagram shows 24 possible responses due to combinations of capacitor error. It is based on a capacitor tolerance of ±2.5%, or ±1% resistors and ±1.5% capacitors. It shows that the original ±1.5° error has now become over ±4.5°. Equation 8 (last month), demonstrates that with no amplitude errors this increases the power in the unwanted

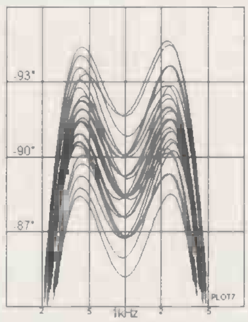

Fig. 1. Effect of component tolerances of 2.5%, for 24 scenarios.
Difference of two pairs of first-order sections, component values from Example 2a of last month's article.

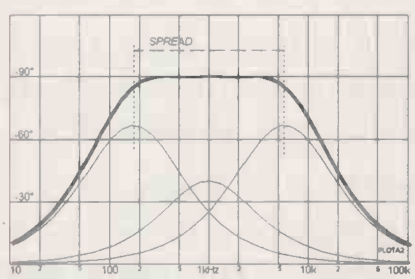

Fig. 2. Difference of three pairs of first-order sections, example 4.

Table 1. Performance data for examples given in text.

Example	1	2	2a	3	4	5	6
Pairs	2	2	2	2	3	4	4
Span	4.36	4.08		3.62	3.447	3.67	4.50
					2.036	2.25	3.06
Spread	14.0	12.0		9.00	33.0	6.89	12.0
					–	5.00	9.0
±3° (Hz)	216	254	257	350	194		22
	4620	3940	3890	2850	5160		46200
±1 (Hz)	240	288	292	417	246	39	
	4170	3470	3470	2400	4240	6460	
±0.5° (Hz)	245	299	302	442	256	41	
	4070	3450	3350	2260	3910	6100	
Phase	+2.66°	+1.54°	+1.4°	+0.29°	±0.13°	+0.49°	±≈3°
ripple	–2.52°	–1.55°	+1.5°	–0.22°		–0.47°	
			–1.6°				

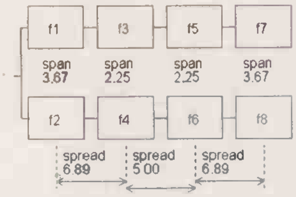

Fig. 3. Difference of four pairs of first-order section.

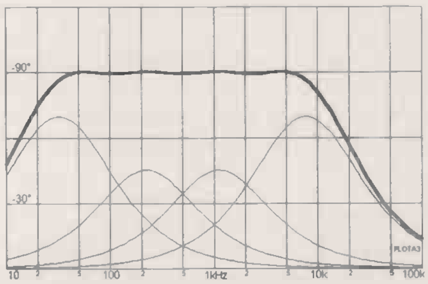

Fig. 4. Difference of four first-order sections showing phase ripple <0.5° from 40Hz to 6100Hz. Parameters from Example 5.

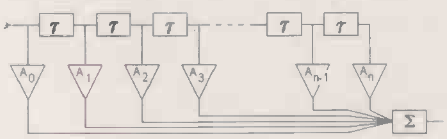

Fig. 5. One form of a finite impulse response, or FIR, digital filter. There are 'n' delays of value and 'n+1' coefficients.

sideband from –38dB to around –28dB. This is a large reduction, but still adequate for some applications. Any error in the gain-setting resistors will make this even worse.

Difference of three first-order pairs

Example 4. The filter described above should be adequate for many applications. You can improve the performance however, at little expense, by adding a third pair of filters.

Take a twin-pair filter with a span of 3.447 and a spread of 33. A third pair of filters is added at the centre frequency, in this case 1000Hz, with a span of 2.036. Result are shown in **Fig. 2**. Ripple is an excellent ±0.125° and the 0.5° 'bandwidth' is 256Hz to 3910Hz, **Table 1**.

Difference of four first-order pairs

The trial and error approach is long-winded for four pairs of filters. There are too many variables to juggle. However, it is still possible to arrive at a filter with a reasonable flat top.

Such a filter is probably over-specified for radio work. One exception might be if the radio had little rf selectivity, as is the case with a specialised very-low-frequency device I am working on. Here, the outphaser has to provide a tight phase response from a few hertz to around 5 or 6kHz.

A more probable use for a four-stage filter would be in a frequency shifter for preventing howl-around caused by microphone feedback. This can be prevented by using a small shift of around 5Hz over the 0-20kHz spectrum*. The ear is very sensitive to the 'warble' caused by inadequate suppression of the unwanted sideband resulting from a frequency-shift.

Example 5. Centre frequency of the four pairs of filters, **Fig. 3**, is 500Hz. This means that the four pairs are at 32.45, 223.6, 1118.0 and 7703Hz. The f_3/f_4 and f_5/f_6 pairs have a spread of 5.00. Overall spread between the f_1/f_2 and f_7/f_8 pairs is 237.4.

The filter has low ripple, Fig. 4. Phase ripple is under ±0.5° while the error is within ±0.5° from 41Hz to 6100Hz, and ±1° from 38Hz to 6460Hz. With a change in centre frequency to 1250Hz the 'bandwidth' becomes 95Hz to 16.2kHz, which is suitable for many audio applications.

In terms of 'bandwidth' and flatness, this represents very good performance. These features will, of course, be degraded by increased component tolerances. The values could probably be tweaked to reduce the ripple, or to aid the use of E24 components, but this is a lengthy exercise.

Wide-band filters

Example 6. Assuming the following for Fig. 3,

f_1/f_2 span 4.50 spread (to f_3/f_4) 12
f_3/f_4 span 3.06 spread (to f_5/f_6) 9.0
f_5/f_6 span 3.06 spread (to f_3/f_4) 9.0
f_7/f_8 span 4.50 spread (to f_5/f_6) 12

produces a filter which maintains a flat passband to ±3° over a bandwidth of 22Hz to 46200Hz. This is the equivalent of over 2000:1, or 3.3 decades, or 11 octaves. I have not attempted any fine adjustments to minimise phase ripple. Even so, this represents an unprecedented response. It far exceeds the performance of the old polyphase filters.

A ripple of 3° is –32dB power attenuation. This could be improved on, but the example serves to demonstrate some of the possibilities of filter design.

* For example, shift up to a high intermediate frequency using an outphaser, and then – because there is no out-of-band noise – you can down-shift in a simple remodulator.

Table 2. Coefficients for the 32-tap FIR filter shown in Figs 6 and 7.

Coefficient	Rectangular Window	Hamming Window
A(1)	–0.042441	–0.003770
A(3)	–0.048971	–0.007714
A(5)	–0.057875	–0.016462
A(7)	–0.070736	–0.031849
A(9)	–0.090946	–0.057272
A(11)	–0.127324	–0.101294
A(13)	–0.212207	–0.195756
A(15)	–0.636620	–0.630993
A(17)	0.636620	0.630993
A(19)	0.212207	0.195756
A(21)	0.127324	0.101294
A(23)	0.090946	0.057272
A(25)	0.070736	0.031849
A(27)	0.057875	0.016462
A(29)	0.048971	0.007714
A(31)	0.042441	0.003770

Using integrated filters

The fact that component tolerance is crucial indicates the use a filter *IC*. There are two types of integrated circuit filter, one involving switched-capacitors, the other continuous time techniques. Unfortunately, switched capacitor filters have too restricted a range of clock frequencies, and there is the question of clock noise too.

Continuous-time filters are state-variable designs using conventional op-amps. They incorporate closely matched on-chip resistors and capacitors. However, many such designs appear no better than would be achievable using ±1% tolerance components.

Component costs

Using discrete components allows you to tighten up the tolerances. A ±1% 50ppm/°C resistor costs around £0.03 while a ±1% –100ppm/°C capacitor is around £0.35.

It might be possible to trim the gain. If not,

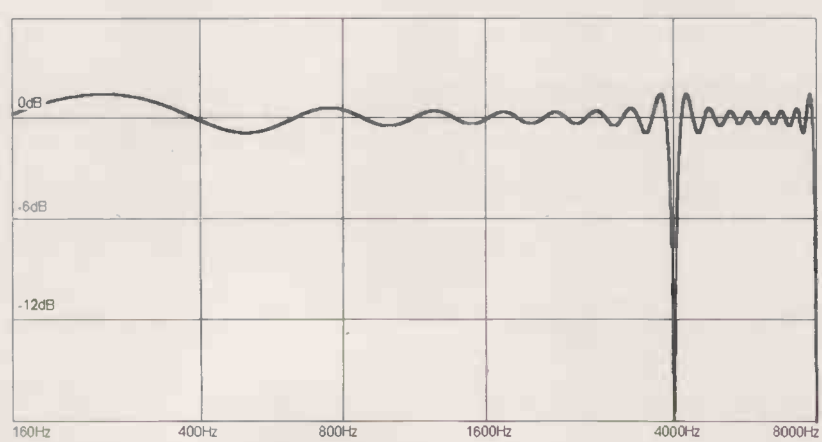

Fig. 6. Amplitude response of 32 tap FIR filter with 0-4kHz bandpass, quadrature phase and 8kHz sampling. Phase is a constant –90° while amplitude ripple is within 0.5dB (6%) from 580Hz to 3.4kHz.

you could consider using ±0.1% 10ppm/°C resistors here, and for the *RC* filters. These cost around £0.80 each. But, since they are available in E96 values, you will probably only need one per filter section; since the gain-setting resistors are all the same value you could save money too.

Digital filters

I cannot discuss outphasers without mentioning the possibilities opened up by digital filters. A 3kHz audio outphaser is well within the scope of even a modest digital filter algorithm. You can also use dsp techniques to implement a very good wideband response, as in Fig. 4 – provided you can cope with the sampling rate required.

You could easily implement the all-pass filters digitally since they are based on 'simple' first-order sections. However, there is another approach which generates the phase-shift directly.

In a finite impulse response (FIR) filter, **Fig. 5**, each of the 'τ' blocks is a delay of one sample. The design process involves our specifying both the amplitude and the phase response. Normally, you would not bother to specify the phase response, and the finite-impulse response filter would give a linear phase shift.

In summary,

$$G(\omega) = 1: \ 0 \le |\omega| \le \frac{1}{2}\omega_s$$

$$\theta(\omega) = \begin{cases} -90^\circ : \omega > 0 \\ 0 \ : \omega = 0 \\ +90^\circ : \omega < 0 \end{cases}$$

This could be described as a brick-wall band-pass filter from 0 to $\omega_s/2$. It has a quadrature phase response instead of the more usual linear phase response.

Calculating the filter coefficients involves a Fourier transform, and is well-covered in digital-signal-processing textbooks. For a filter with *M* taps, i.e. coefficients 0 to *M*, and *M* is even, the coefficients can be derived to be,

$$A_n = \begin{cases} 0 & : n = \frac{1}{2}M, \text{ else} \\ \dfrac{1 - \cos\left(n - \frac{1}{2}M\right)\pi}{\left(n - \frac{1}{2}M\right)\pi} & : 0 \le n \le M \end{cases}$$

so for a 32 tap filter the coefficients are,

$$0, \frac{-2}{15\pi}, 0, \frac{-2}{13\pi}, 0, \frac{-2}{11\pi}, 0, \ldots, \frac{2}{13\pi}, 0, \frac{2}{15\pi}, 0$$

Note that there are only eight distinct coefficients and that the even coefficients A(0) to A(32) are zero. There is a common factor of $2/\pi$, leaving the coefficients as simple ratios. This could help to speed up the operations when the filter is implemented in a microprocessor or digital-signal processor.

Within the constraints of sampling theory and 'windowing' the filter has a 'perfect' 90° phase shift and a flat amplitude response. Windowing is the effect caused by the finite length of the filter. In practice, the phase shift is 90°, but there is some amplitude ripple. The only remaining point to note is that in addition to the 90° phase delay there is a sampling delay of $\frac{1}{2}M\tau$ which must be matched by the in-phase channel.

Amplitude response of a 32-tap filter is shown in **Fig. 6**, with the coefficients listed in **Table 2**. Ripple is around 0.5dB, which is ±6%. This can be improved dramatically by implementing a Hamming window as shown in **Fig. 7**. I will not explain windowing here, suffice to note that the coefficients are modified by the windowing function,

$$A_n \leftarrow A_n \left\{ 0.54 + 0.46\cos\left(\frac{n - \frac{1}{2}M}{\frac{1}{2}M} \right)\pi \right\}$$

This has the effect of flattening the amplitude response without affecting the phase response.

The graphs were generated by running a simple Basic program. Firstly, the program generates the coefficients and then executes an inverse transform to produce the values of amplitude and phase which would occur.

The programs were based on those given in Lockhart & Cheetham in 1989, see last month's article for a list of references. Finite impulse response filters and windowing are covered in any number of digital-signal processing books, though quadrature-phase filters are not widely discussed.

Using the filters in a receiver

Schematics discussed so far have implicitly given the configuration for a transmitter. The same module is used in a receiver, but the quadrature signals from the demodulator drive the inputs to the two filter chains. These are summed, or differenced, at the output to recover one of the sidebands. Just as in the modulator of Fig. 2 last month, you can reverse the order of the components and do the outphasing at rf if desired.

Variations on a theme

I compared the outphaser and Weaver methods, and found them similar. Other choices facing the designer are whether to do the phasing at rf or af, and whether to use low or high-pass filter sections.

Swapping the position of the *R* and *C* in last month's Fig. 3d, does not alter the operation of the circuit. I have deliberately avoided giving any definite recommendations here, since the design route you take depends on your precise application.

Using first-order sections and the parameters of span and spread eases the analysis and allows us to adapt the outphaser concept for other uses. ■

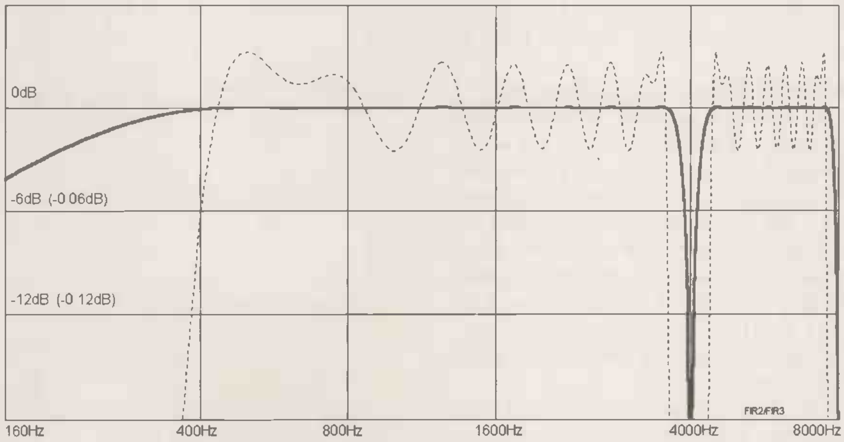

Fig. 7. Effect of Hamming window. The same 32 tap filter with modified coefficients shows virtually no amplitude ripple in the pass band. Scale for the dotted line is 100 that of the solid line, and shows that the ripple is 0.035dB (0.4%) from 420Hz to 3.6kHz.

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Access

VISA

Learn 8048

Jim Whitehouse examines a hard and software training kit designed to help teach how 8048 family controllers work.

Kanda's Microcontroller Training system is a complete package aimed at introducing students and newcomers to the hardware and software intricacies of the *8032/8052* family of microcontrollers. It is specifically aimed at the Btec syllabus, but it should be of use to anyone wanting a basic knowledge of microcontrollers.

The kit is housed in an attractive lockable case, inside which is all the equipment required for the course – apart from the host computer. It consists of four teaching manuals, one reference manual, ten hardware units, software on disk and the necessary interconnecting cables.

A 'Connecting the system' leaflet helps student to check that they have all the necessary parts and the correct connections. For the lecturer there is a short introduction explaining what the training system is about.

First reactions

My initial reaction to the kit was that considerable thought had been given to the packaging of the hardware. The units were robust and clearly labelled.

On the other hand, the manuals, were *not* well made and some of the pages were already falling out. In addition, reading the manuals did not enthuse me to become an avid reader due to the meandering style of writing.

However, Kanda explained that the text was written in accordance with the requirements of Btec NIII syllabus and had already been in use for two years. The manuals covered most of the topics that you would expect to find. It would have been useful to know what was coming next by way of an index. I found having to wait until page 35 to discover what was going to happen a little unhelpful.

Within the manuals

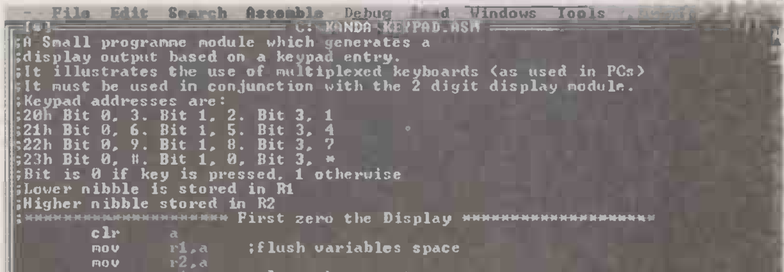
Each manual splits into six teaching blocks with sections for ease of teaching. The training system assumes little knowledge of electronics. Although not intended to teach electronics, the package does briefly explain what is happening where necessary.

Simple digital techniques, binary and hexadecimal mathematics and Boolean algebra are covered first in the manuals. Later binary coded decimal concepts and conversions are discussed.

Useful exercises are included to ensure that the student has understood the points made in each section. If the manual was followed in its entirety then the student would cover all the functions – as opposed to every code – involved in the instruction set. This is a sound basis for further learning.

Unlike the teaching manuals the reference manual is well written and contains invaluable information. The teaching manuals contain excellent technical coverage on factual issues, but where they express opinions, they tend to fail to appreciate that there are other views than those of the software engineer.

All elements of hardware and software of the microcontroller training kit are housed in the same lockable case.



The system incorporates a full function editor with cut and paste, search, multiple file capability etc. for writing the source code. All operations take place in this single desktop which gives an excellent working environment.

Loading up

The software loaded easily and performed well. Programs can be written in source code mnemonics, assembled and simulated. In addition, there is provision for downloading the object code to the microcontroller.

Based on pull down menus, the package allows the option of selecting via a mouse. It has facilities to read, write and modify programs in the source code and store them.

Source code can be assembled, and debugged if necessary, into the required machine code. It can then be simulated, using single step, step back, step over or run combinations, after which the program can be loaded into its targeted device. The targeted device could be an emulator, or one of a wide range of eprom types.

There is a central interface unit communicating with an RS232 serial port. Because the software takes control of the host computer serial link, no problems were experienced due to incompatible bit rates, etc. Cabling the host computer to the central interface and the 9V supply was easy thanks to the circuit diagram provided.

In practice

Training starts with the ubiquitous traffic light system and is user friendly. Next, modules show how to input digital signals in the form of switches and digital outputs, illuminating leds, creating sound with different notes and multiplexing the outputs onto a seven segment display. Inputting numbers from a keyboard and handling analogue inputs and outputs are also covered.

One section deals with how special-function registers and interrupts are handled. A prototype board included allows for the student's own circuitry. One additional module emulates an eprom. This allows the user to modify programs more easily and more quickly. A further module programmes the eprom with the finalised code.

Teaching blocks deal with the important procedure of writing programs. Equally importantly they help assessing what is required with the aid of such tools as the flow diagram.

In summary

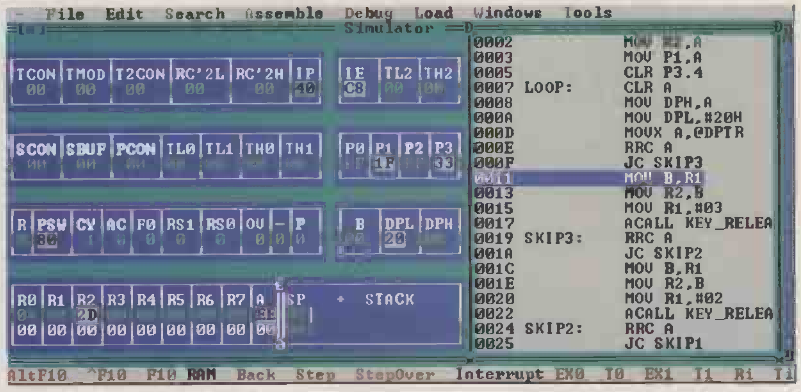
As a practising engineer, I found the package easy to use and understand. I believe that any electronics engineer would be able to use this package as an introduction to microcontrollers.

The whole package was far better than other manufacturer's low cost systems. I believe that it would enable a reasonably intelligent student to understand the workings of the microcontroller sufficiently well to enable that person to build a small unit and go on to learn more about computer systems.

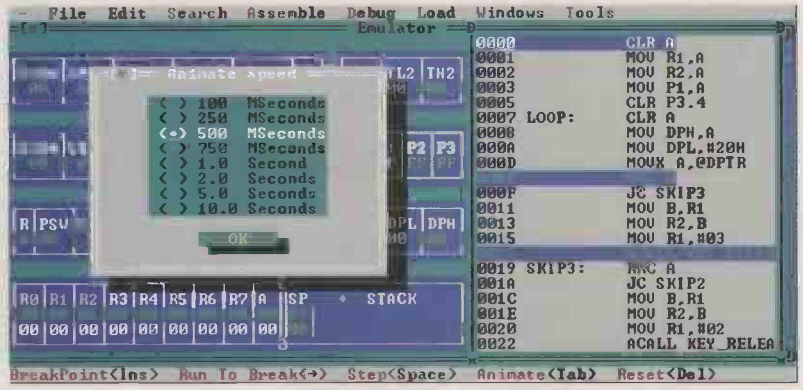
Since this is intended as an educational package, it must be marked accordingly, so I give it 9 out of 10. Presentation of the manuals caused the lost point. ■

Availability

Kanda's Microcontroller Training system is priced at £595, exclusive. The training kit plus development kit is £795 and includes eprom emulator and programmer. The prototyping board is separate at £80. Quantity discounts available. Call Kanda on 01974 282670, fax 01974 282356, or write to Pendre Hafod, Pontrhydgroes, Ystrad Meurig, Dyfed SY25 6DX.

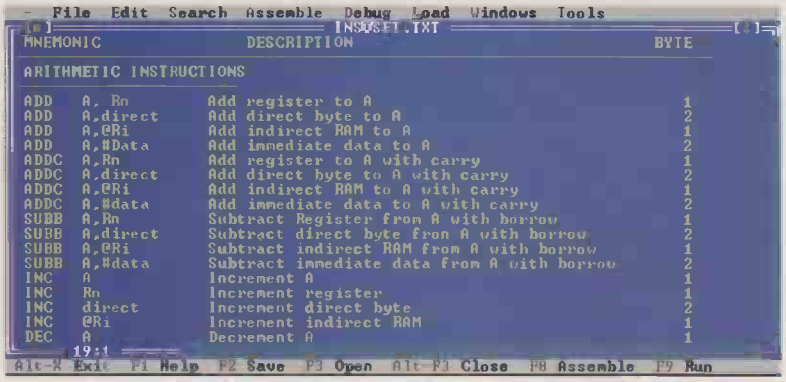


The simulator is available with one key press and provides a clear picture of the processor values as you step through the code enabling you to find the inevitable bugs. The values in each register can be altered very easily to simulate different conditions or inputs.



In-circuit emulation allows your code to run on the processor itself for more advanced debugging when you add the hardware. Breakpoints, single step and animate give the control needed to isolate your problems.

The assembler is integrated with the editor and highlights errors as they occur, rather than producing a separate error listing as this simplifies the assembly process.



The instruction set is included for easy reference although full explanations are given in the reference manual that forms part of the comprehensive course work.

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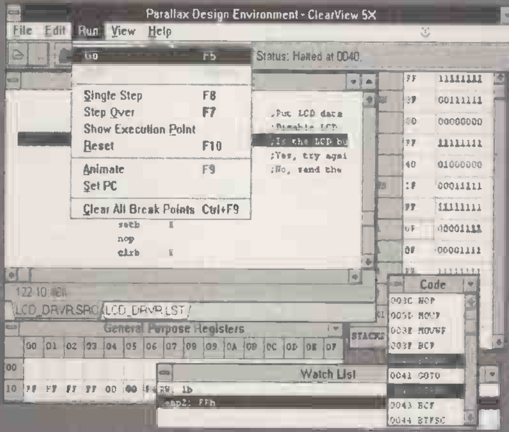

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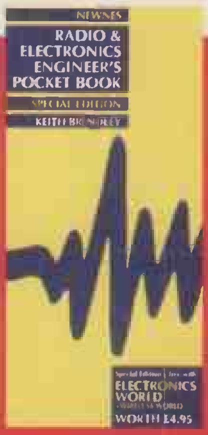

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'Scissors' overcome earth loop problems

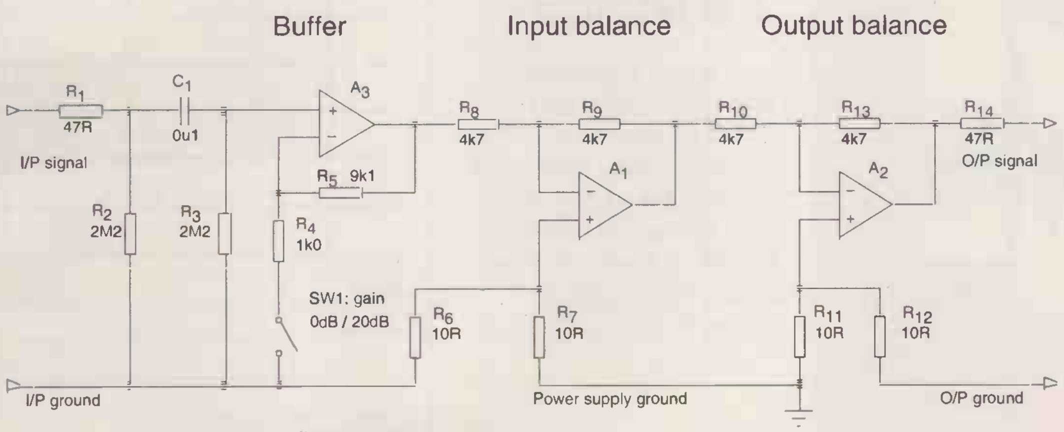
This is a transformerless circuit for overcoming earth loop problems in the interconnection of equipment. Amplifier A_1 is connected as the usual differential amplifier except that, as the ground side impedance is low, the ground side resistors can be small to reduce noise.

Amplifier A_2 is also configured differentially, but here the differential input cancels any series mode signal appearing between the supply earth of the circuit and the input earth of the of the destination equipment. Thus you can have many 'earth scissors' running from a common power supply without making more earth loops. This second section is sometimes featured in well designed audio equipment whose outputs are described as 'ground-compensated' and is highly recommended for general use.

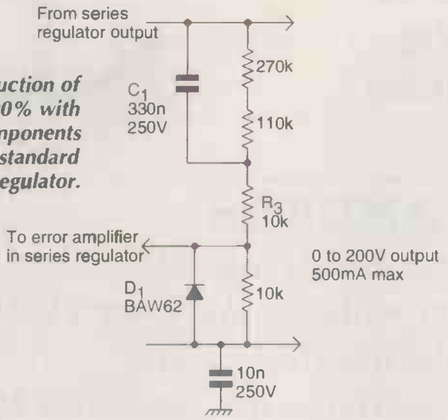
Low-level sources may require pre-amplification, or a high-impedance load. Inclusion of A_3 accommodates these sources, providing buffering and gain referred to the source ground, before feeding A_1.

Simon Bateson
Hutton Rudby
Yorkshire

Many of these 'earth scissors' can be run from a common power supply without causing earth-loop problems.



A_1, A_2 = 1/2 NE5532 or similar A_3 = 1/2 TL072 or similar FET op-amp   All supplied with ±15V



Ripple reduction of around 90% with these few components on a standard regulator.

Reduce power supply ripple

As a means of obtaining a 10:1 reduction in ripple, this could hardly be less complicated.

It consists merely of by-passing the sensing network from the power supply output to the error amplifier or, to put it another way, of providing better coupling for the feedback.

Most of the sensing network is by-passed by the 330nF capacitor and R_3 helps to maintain stability. Testing with 1kHz load switching showed a much faster regulator response with no sign of instability. Diode D^1 prevents the error amplifier input being taken negative in the event of a short-circuit.

The circuit was used in an otherwise completely standard regulator and gave quite dramatic results.

Gregory Freeman
Nairne
South Australia

£100 WINNER

Spectrum analyser for audio

Operating in real time, this circuit allows the frequency response of a speaker to be viewed on an oscilloscope using X-Y mode. The X-axis indicates log frequency and the Y-axis dB, where a sound level meter provides decibel output.

The sweep vco is based on a *4046*. A single sweep covers the audio range 20Hz to 20kHz with logarithmic voltage-to-frequency relationship.

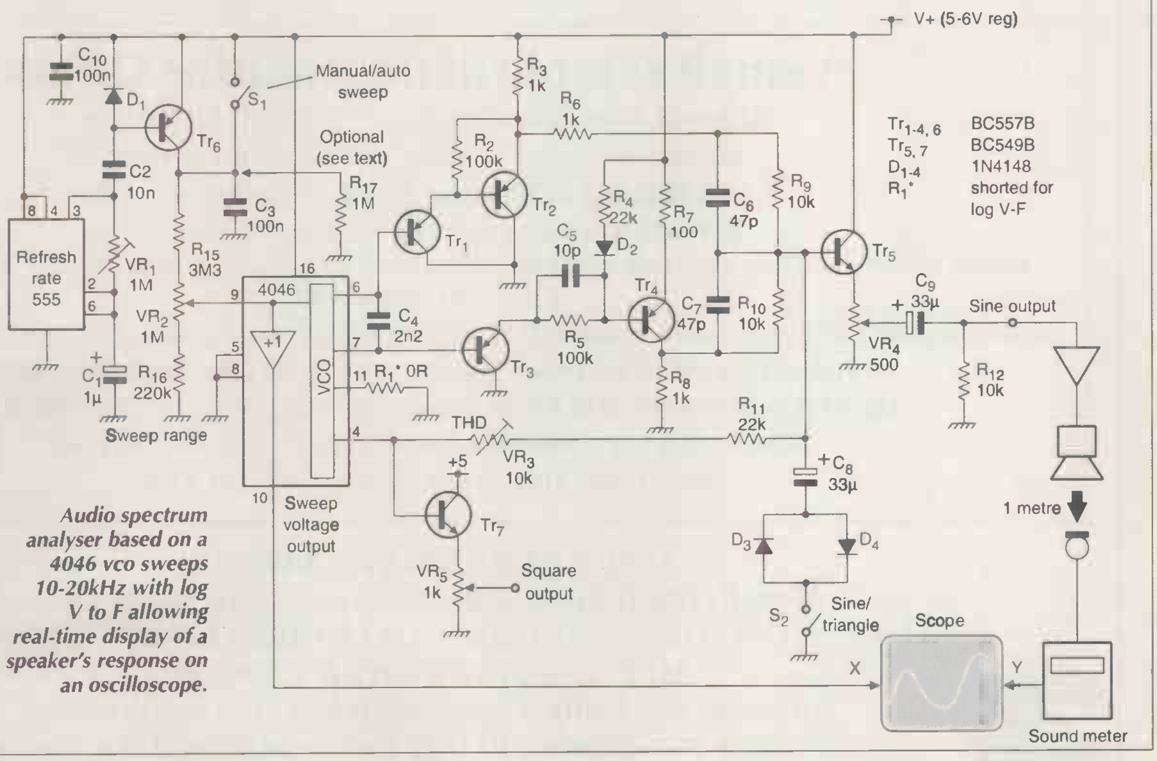
Darlingtons Tr_{1-4} buffer the ramp across the timing capacitor C_4. The waveform from pin 7 is inverted using Tr_4 and summed with the waveform from pin 6 giving a triangle wave plus a squarewave component which is removed by trimming VR_3. Capacitors C_{5-7} provides compensation for Tr_{1-4} to remove switching spikes.

Sweep range is set with VR_2. The screen refresh rate is set by VR_3 at about 8Hz to avoid flicker. An oscilloscope X input sensitivity of 50mV/div gives two divisions per decade, up to 20kHz. A sound meter ac output gives fastest response.

Calibrate the vertical scale using the sound level meter and VR_4 to find dB/div. Long cone excursions caused by frequencies below 30Hz can be avoided by increasing VR_3 or by reducing sweep time via R_{17}.

For manual sweep, short Tr_6 (c-e) and vary VR_3. This allows levels to be recorded manually if a scope is not available. The vco can also be used as a sine, triangle, square signal generator to cover 10Hz-100kHz in one range with 1% sine thd, or for sound effects. For minimum thd the supply voltage needs to be about 5.8V and resistor R_4 may also need trimming for a lower second harmonic component.

Ian Hegglun
Goodna
Australia

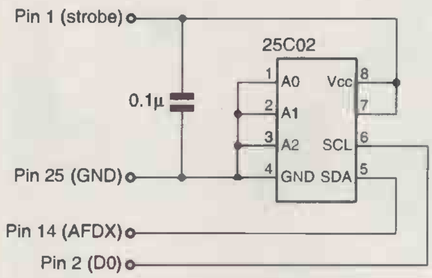


Audio spectrum analyser based on a 4046 vco sweeps 10-20kHz with log V to F allowing real-time display of a speaker's response on an oscilloscope.

Ultra simple I²C interface for eeprom

Only one capacitor is needed to access the *24C02* I²C programmable eeprom via a pc. The AUTOFDX line is bidirectional. It has an open collector output and needs no external pull up. I used the strobe, pin 1, for power supply, which is enough for this cmos IC.

Verdieri Giordano
Montanara
Italy



Programming and reading the 24C02 eeprom via a pc's parallel port needs only one capacitor since the LPT bidirectional lines used have an open-collector structure with integral pull-up.

C listing demonstrating how the I²C eeprom is accessed via the pc's parallel port.

```
#define SDA  0x2      // Autofdx  pin 14
#define VCC  0x1      // Strobe   pin  1
#define CLK  0x1      // D0       pin  2
                      // GND      pin
25
#define prn  0x378
inpsda(void);
delay(void);
delay1(void);
stope(void);
starte(void);
clack(void);
clk_0(void);
clk_1(void);
sda_0(void);sda_1(void);
tst_sda_1();
rdeer(unsigned char ,unsigned char *);
wreer(unsigned char ,unsigned char );
write8(unsigned char );
unsigned char read8(void);
main()
{
   outp (prn+2,inp(prn+2) & ~VCC) ;
   for(;;)
   {
    unsigned char addr,dat1,car;
     printf("\nR=Read  W=Write  Q=Quit:
");
     car = getch();
     switch(car)
     {
        case 'r':
          printf("\nRead address (hex):
");
          scanf("%x",&addr);
          if (rdeer(addr,&dat1))
            printf("\nError ! \n");
          else
            printf("              Value=
%02X \n ",dat1);
        break;
        case 'w':
          printf("\nWrite address
(hex): ");
          scanf("%x",&addr);
          printf("              Value:
");
          scanf("%x",&dat1);
          if (wreer(addr,dat1))
            printf("\nError ! \n");
        break;
        case 27:
        case 'q':
        case 'Q':
          exit(1);
        default:
        break;
     }
   }
}
clk_0() { outp (prn,  inp(prn)   &
~CLK) ; }
clk_1() { outp (prn,  inp(prn)   |
CLK) ; }
sda_0() { outp (prn+2,inp(prn+2) |
SDA) ; }
sda_1() { outp (prn+2,inp(prn+2) &
~SDA) ; }
inpsda() { sda_1(); }
rdeer(unsigned char addr,unsigned char
*dat1){
    *dat1 = 0;
    if (tst_sda_1()) return(1);
    starte();
    write8(0xa0);
    if (clack()) return(1);
    write8(addr);
    if (clack()) return(1);
    starte();
    write8(0xa1);
    if (clack()) return(1);
    *dat1 = read8();
    stope();
    return(0);
}
wreer(unsigned char addr,unsigned char
dat1)
{
    if (tst_sda_1()) return(1);
    starte();
    write8(0xa0);
    if (clack()) return(1);
    write8(addr);
    if (clack()) return(1);
    write8(dat1);
    if (clack()) return(1);
    stope();
    delay1();
    return(0);
}
delay()
{
    int i;
    for(i=0;i<200;i++);
}
delay1()
{
    int i;
    for(i=0;i<10000;i++);
}
starte()
{
      clk_1();
      delay();
      sda_1();
      delay();
      sda_0();
      delay();
      clk_0();
      delay();
}
stope()
{
      sda_0();
      delay();
      clk_1();
      delay();
      sda_1();
      delay();
}
clack()
{
      inpsda();
      clk_1();
      delay();
      if ((inp(prn+2) & SDA) == 0 )
return(1);
      clk_0();
      delay();
      return(0);
}
tst_sda_1()
{
    sda_1();
    delay();
    if ((inp(prn+2) & SDA) != 0 )
return(1);
    return(0);
}
write8(unsigned char dat1)
{    int i;
    unsigned char mask;

    mask = 0x80;
    for(i=0;i<8;i++)
    {
      if ((dat1 & mask) == mask)
sda_1(); else sda_0();
      delay();
      clk_1();
      delay();
      clk_0();
      mask >>= 1;
    }
}
unsigned char read8()
{
    unsigned char a,dat1;
    int i;

    a= 0x80;
    dat1 = 0;
    for(i=0;i<8;i++)
    {
      clk_1();
      delay();
      if ((inp(prn+2) & SDA) == 0 )
dat1 |= a ;
      clk_0();
      a >>= 1;
    }
    return(dat1);
}
```

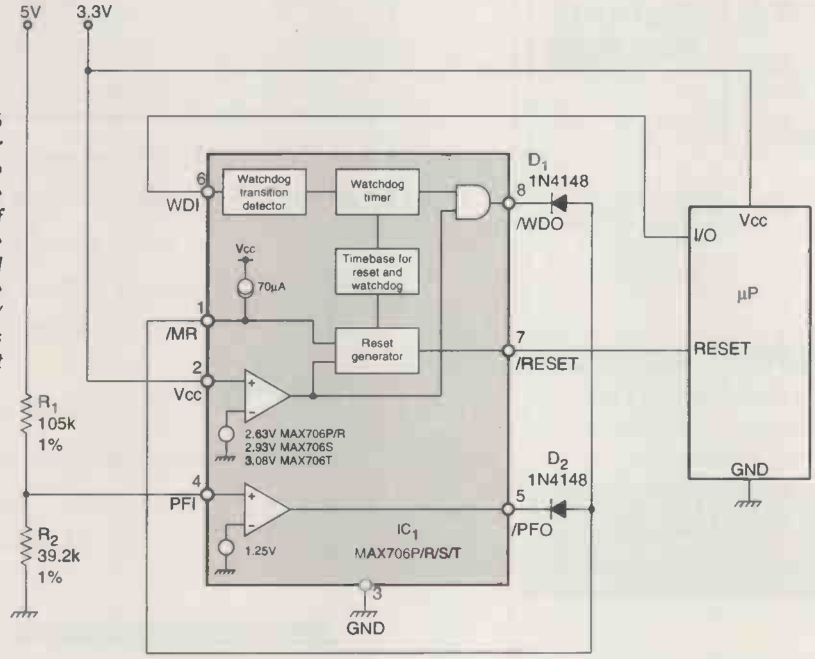
Dual function microprocessor supervisor

In addition to monitoring the 3.3V and 5V supplies to a microprocessor, this *MAX706* supervisor ic circuit also keeps a check on software execution.

The software check relies on the expectation of transitions on a selected i/o line at least once every 1.6s. If these do not occur, IC_1 sends a reset pulse to the microprocessor.

One of three reset thresholds for the 3.3V monitor, from 2.63V to 3.08V, depending on the part number suffix, trigger resets directly via the internal op-amp, while the 5V monitor triggers externally via the /PFO output and the /MR (manual reset) input, the level depending on R_1,R_2 and the PFI input switching threshold of, typically, 1.25V. Resets caused by either input are maintained for as long as the supplies remain low and for 200ms after a supply is restored to normality.

Diodes D_1, D_2 are wire-Ored to allow the software watchdog output /WDO to share control of the /MR input, but if the 3.3V supply is derived from 5V, an early warning of 5V failures can be obtained by removing the diodes, connecting /WDO to /MR and taking /PFO directly to an interrupt pin on the microprocessor.



Maxim 706 microprocessor supervisor used to monitor the progress of software operation, as well as to provide resets when 5V and 3.3V supplies fall below set thresholds.

Dana Davis and Craig Falkenham
Maxim Integrated Products Ltd
Theale
Berkshire

Serially accessed memory is expandable

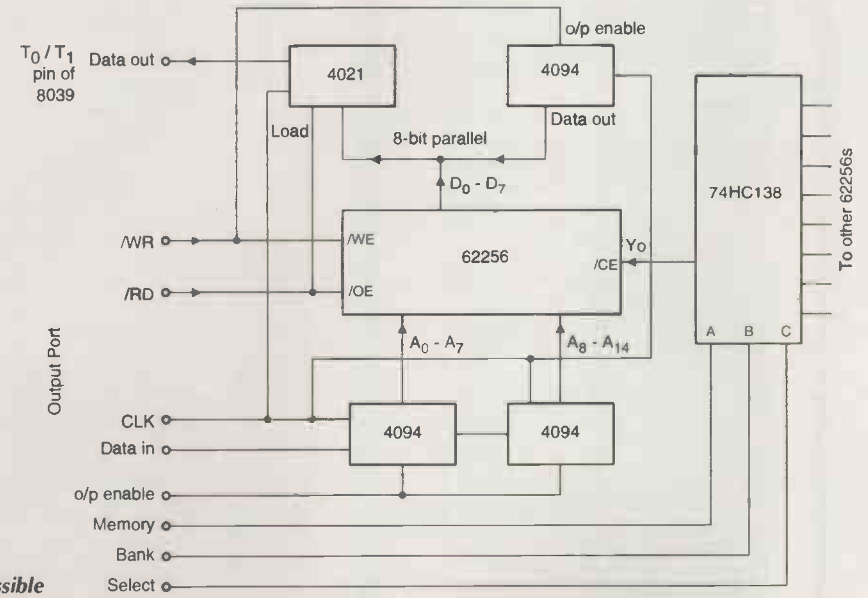
A 3-to-8-line decoder allows this serially-accessible static ram circuit to be expanded to accommodate up to eight *62256* memory ICs.

Writing to the memory is carried out as follows.

Select data bank
Set data bit at DATA IN
Apply a clock pulse at CLK
Set next data bit at DATA IN
Repeat the previous two steps until address and data is clocked in
Apply O/P ENABLE high
Apply a pulse for WRITE

To read from the memory,

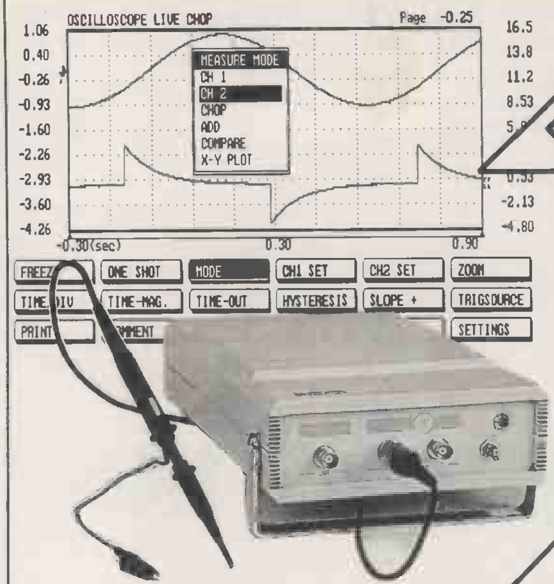
Select data bank
Set WRITE output high
Set data bit at DATA IN
Apply a clock pulse at CLK
Set next data bit at DATA IN
Repeat previous two steps until address is clocked in
Apply O/P ENABLE
Apply a READ pulse
Apply O/P ENABLE low
Check T0/T1
Apply a CLK pulse
Repeat previous two steps until byte is assembled

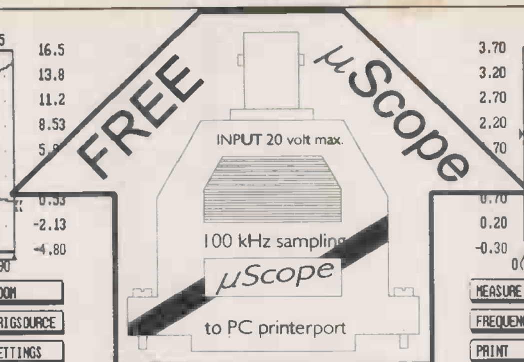


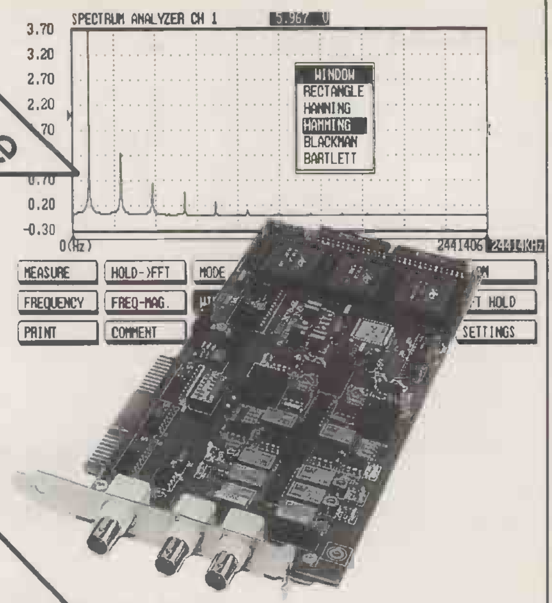
4094: 8-bit shift register/latch with tri-state o/p (serial in/parallel out)
4021: 8-bit static shift register (parallel in/serial out)
62256: memory
74HC138: 3-to-8 decoder

Serially accessible memory is bank selectable for up to eight 62256 static rams.

Jayant Kathe
Bombay
India







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TP508

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- *sampling speed* 50 Msamples/sec
- *resolution* 8 bits
- *input range* 5 mV/div - 20 V/div
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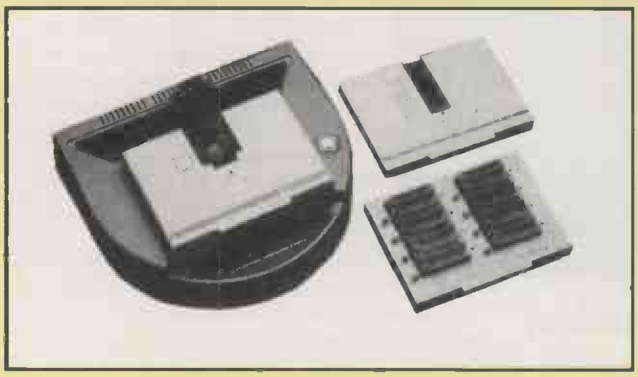

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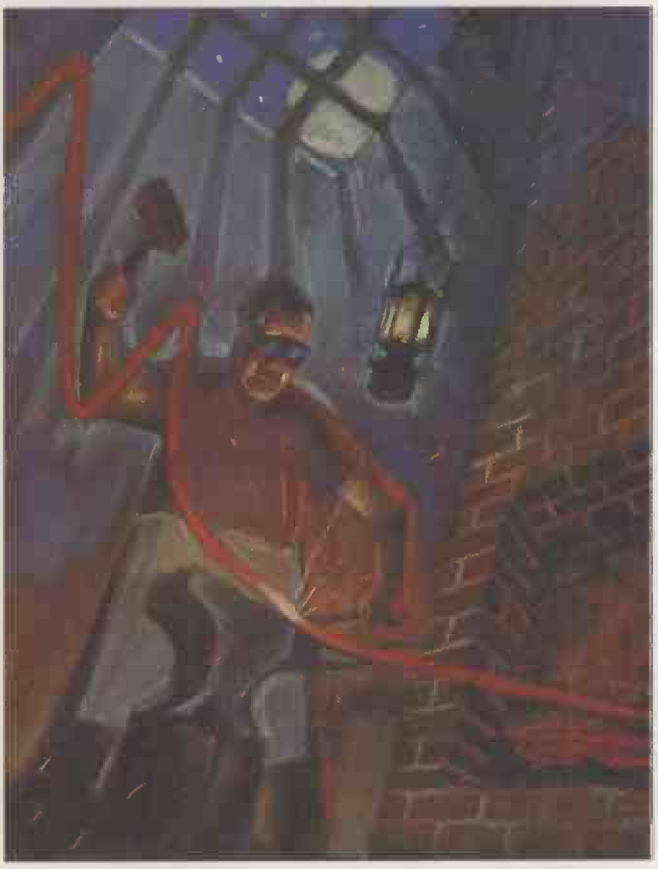
Thermal dynamics in audio power



While analysing thermal dynamics in high performance power amplifiers, Douglas Self finds unexpected results when comparing bias errors in emitter follower and complementary feedback pairs.

The most intractable problem in Class B power amplification is crossover distortion in the output stage. High order harmonics generated by crossover gain fluctuations are poorly linearised by negative feedback. This is because the amount of feedback applied at high frequencies must be restricted for Nyquist stability.

Some earlier work of mine suggests that the amount of crossover distortion produced is largely fixed for a given configuration and devices. As a result, the best you can do is ensure that the output stage runs under optimal quiescent conditions. Schemes for controlling quiescent current via direct servo control have been mooted[1]. But all suffer from the difficulty that the quantity we wish to control is not directly available for measurement. This is because the quantity is swamped by Class B output currents, unless there is a complete absence of signal.

In contrast, the quiescent current of a Class A amplifier is easily measured, allowing very precise feedback control. Ironically, its value is not critical to distortion performance[2].

Quiescent current considerations

So just how accurately must quiescent current be held? This is not easy to answer, not least because it is the wrong question. Reference 1 established that the crucial parameter is not quiescent current, hereafter I_q, as such, but rather the quiescent voltage drop V_q across the two emitter resistors R_e. This takes a little swallowing. After all, people have been worrying about quiescent current for 30 years or more, but it is actually good news, as the value of R_e does not complicate the picture.

Voltage across the output stage inputs, V_{bias}, is no less critical. Once R_e is chosen, V_q and I_q vary proportionally. The two main types of output stage, the emitter-follower, ef, and the complementary feedback pair, cfp, are shown in **Fig. 1**. Their V_q tolerances are quite different.

From measurements, I take the permissible error band for V_q in the ef stage as ±100mV, and for the cfp as ±10mV. These figures are not definitive; I only suggest that they are reasonable. In terms of total V_{bias}, the ef needs 2.93V ±100mV, and the cfp 1.30V ±10mV. Voltage V_{bias} must be higher in the follower as four base-emitter voltages are subtracted from it to get V_q. In the cfp on the other hand, only two driver base-emitter voltages are subtracted.

The cfp stage appears to be more demanding of V_{bias} compensation than follower, needing 1% rather than 3.5% accuracy, but things are not so simple. Stability of V_q in the follower stage depends primarily on the hot output devices, as emitter follower driver dissipation varies only slightly with power output.

Voltage V_q in the cfp depends almost entirely on driver junction temperature. This is because the effect of output device temperature is reduced by the local negative feedback. However, cfp driver dissipation varies strongly with power output[3] so the superiority of this configuration cannot be taken for granted.

Driver heatsinks are much smaller than those for output devices, so the cfp V_q time constants promise to be some ten times shorter.

Thermal compensation

In Class B, the usual method for reducing quiescent variations is 'thermal feedback'. The V_{bias} is generated by a thermal sensor with a negative temperature coefficient, usually a V_{be} multiplier transistor mounted on the main heatsink.

This system has proved workable over the last 30 odd years, and usually prevents any possibility of thermal runaway. However, it suffers from thermal losses and delays between output devices and temperature sensor. These make maintenance of optimal bias rather questionable, and in practice quiescent conditions are a function of recent signal and thermal history.

Thus the crossover linearity of most power amplifiers is intimately bound up with their thermal dynamics. It is surprising this area has not been examined more closely. Reference 4 is one of the few serious papers on the subject – though the conclusions it reaches are unworkable.

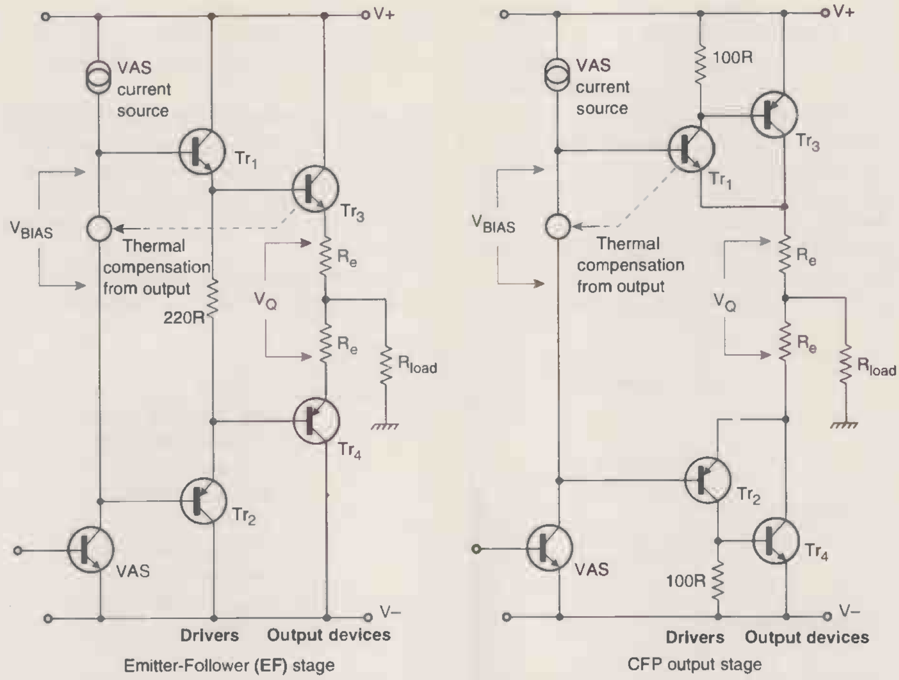


Fig. 1. Emitter-follower and complementary feedback pair configurations, showing V_{bias} and V_q.

As is almost routine in audio design, things are not as they appear. So called 'thermal feedback' is not feedback at all. This implies the thermal sensor is in some way controlling the output stage temperature. It is not. It is really a form of approximate feedforward compensation, as shown in **Fig. 2**.

The quiescent current I_q of a Class B design causes a very small dissipation compared with the signal. As a result, there is no meaningful feedback path returning from I_q to the left of the diagram. This might be less true of Class AB, where quiescent dissipation may be significant.

Instead, this system aspires to make the sensor junction temperature mimic the driver or output junction temperature. It can never do this promptly or exactly though because of the thermal resistances and thermal capacities that lie between driver and sensor temperatures in Fig. 2. It does not place either junction temperature or quiescent current under direct feedback control, but merely aims to cancel out the errors. From now on, I will simply call this 'thermal compensation'.

Assessing the bias errors

Temperature error must be converted to millivolt error in V_q, for comparison with the tolerance bands suggested above. In the cfp stage this is straightforward.

Both driver V_{be} and the halved V_{bias} voltage decrease by 2mV/°C. As a result, temperature error converts to voltage error by multiplying by 0.002. Only half of each output stage will be modelled, exploiting symmetry, so most of this article deals in half V_q errors, etc.

To minimise confusion, this use of 'half amplifiers' is adhered to throughout. The only exception is at the final stage, when the calculated V_q error is doubled before comparison with the tolerance bands quoted above.

Error conversion in the emitter follower is more subtle. The follower V_{bias} generator must establish four times V_{be} plus V_q. Consequently, the V_{be} of the temperature sensing transistor is multiplied by about 4.5 times, and so decreases at 9mV/°C.

The cfp V_{bias} generator only multiplies 2.1 times, decreasing at 4mV/°C. The corresponding values for a half amplifier are 4.5 and 2mV/°C.

However, the emitter-follower drivers are at near constant temperature. After two driver V_{be} values have been subtracted from V_{bias}, the remaining voltage decreases faster with temperature than does output device V_{be}. This runs counter to the tendency to under compensation caused by thermal attenuation between output junctions and thermal sensor. In effect the compensator has 'thermal gain', and this has the potential to reduce long term V_q errors.

I suspect this is the real reason why the emitter follower stage, despite looking unpromising, can in practice give acceptable quiescent stability.

Simulating thermal performance

Designing an output stage requires some appreciation of how effective the thermal compensation will be, in terms of how much delay and attenuation the 'thermal signal' suffers between the critical junctions and the V_{bias} generator.

It is necessary to predict the thermal behaviour of a heatsink assembly over time, allowing for things like metals of dissimilar thermal conductivity, and the very slow propagation of heat through a mass compared with near instant changes in electrical dissipation. Practical measurements are very time consuming, requiring special equipment such as multipoint thermocouple recorders. A theoretical approach would be very useful.

For very simple models, such as heat flow down a uniform rod, it is possible to derive analytical solutions to the partial differential equations that describe the situation. The answer is an equation directly relating temperature to position-along-the-rod and time. However, even slight complications, such as a non-uniform rod, involve rapidly increasing mathematical complexities, and anyone who is not already deterred should consult reference 5, this will deter them.

To avoid direct confrontation with higher mathematics, finite element and relaxation methods were developed. The snag is that finite element analysis is a rather specialised taste, and so commercial element analysis soft-

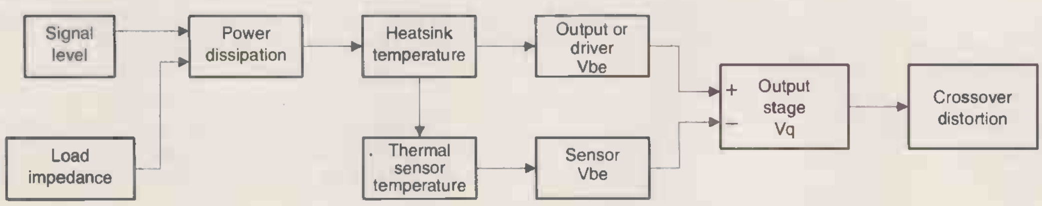


Fig. 2. Thermal signal flow of a typical power amplifier, showing that there is no 'thermal feedback' to the bias generator. There is instead feedforward of driver junction temperature, so that the sensor V_{be} will hopefully match the driver V_{be}.

ware is horrendously expensive. Writing your own program is not practical for most of us.

I therefore sought for another method, and found I already had the wherewithal to solve problems of thermal dynamics. The use of electrical analogues is the key. If the thermal problem can be stated in terms of lumped electrical elements, then a circuit simulator of the Spice type can handle it. As a bonus it has extensive capabilities for graphical display of the output.

The work here was done with PSpice. A more common use of electrical analogues is in the electromechanical domain of loudspeakers, see reference 6 for a virtuoso example.

A simple model of the follower stage

This approach treats temperature as voltage, and thermal energy as electric charge, making thermal resistance analogous to electrical resistance, and thermal capacity to electrical capacitance.

Thermal capacity is a measure of how much heat is required to raise the temperature of a mass by 1°C. And if anyone can work out what the thermal equivalent of an inductor is, I would be interested to know. With the right choice of units, the simulator output will be in volts, with a one to one correspondence with degrees Celsius, and amps, similarly representing watts of heat flow, **Table 1**. It is then simple to produce graphs of temperature against time.

Since heat flow is represented by current, the inputs to the simulated system are current sources. A voltage source would force large chunks of metal to change temperature instantly, which is clearly wrong. The ambient is modelled by a voltage source, since it can absorb any amount of heat without changing temperature.

Consider first the popular emitter follower output stage, in which output device junction temperatures dominate V_q dynamics. The drivers have near constant dissipation regardless of output power[4] and will initially be ignored.

Figure 3 shows a TO3 output device mounted on a thermal coupling bar, with a silicone thermal washer giving electrical isolation. The coupler is linked to the heatsink proper via a second conformal material. This need not be electrically insulating so highly efficient materials like graphite foil can be used. This is representative of many amplifier designs, though a good number have the power devices mounted directly on the heatsink, the results hardly differ.

A simple thermal analogue model of Fig. 3 is shown in **Fig. 4**. The situation is radically simplified by treating each mass in the system as being at a uniform temperature, ie isothermal, and therefore represented by one capacity each. Boundaries between parts of the system are modelled, but the thermal capacity of each mass is concentrated at a notional point. In assuming this capacity elements can be given zero thermal resistance, eg both sides of the thermal coupler will always be at the same temperature.

Similarly, elements such as the thermal washer are assumed to have zero heat capacity, because they are very thin and have negligible mass compared with other elements in the system. Thus the parts of the thermal system can be conveniently divided into two categories – pure thermal resistances and pure thermal capacities. Often this gives adequate results, if not, more subdivision will be needed. Heat losses from parts other than the heatsink are neglected.

In a real output stage

Real output stages have at least two power transistors. The simplifying assumption is made that power dissipation will be symmetrical over anything but the extreme short term, and so one device can be studied by slicing the output stage, heatsink, etc, in half.

It is convenient to read off the results directly in °C, rather than temperature rise above ambient. As a result, Fig. 4 represents ambient temperature with a voltage source V_{amb} that offsets the baseline (node 10) 25°C from sim-

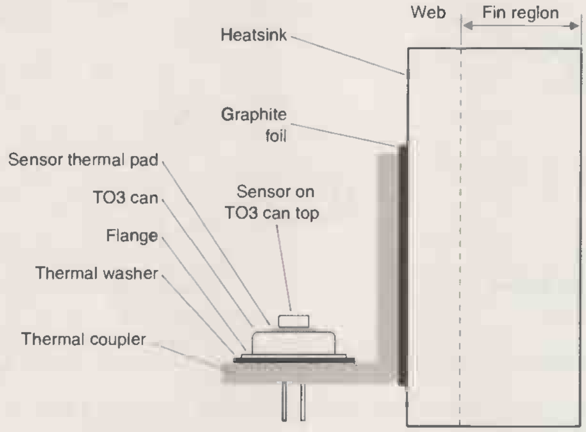


Fig. 3. A TO3 power transistor attached to a heatsink by a thermal coupler. Thermal sensor is shown on can top, more usual position would be on thermal coupler.

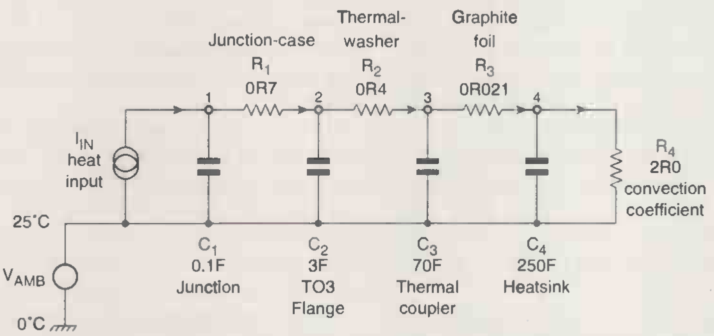


Fig. 4. Thermal/electrical model of Fig. 3, for half of one channel only. Node 1 is junction temperature, node 2 flange temperature, and so on. Voltage V_{amb} sets the baseline to 25°C. Arrows show heat flow.

Table 1. Circuit simulation values for EF stage.

	REALITY	SIMULATION
Temperature	Degrees C	Volts
Heat quantity	Joules (watt-seconds)	Coulombs (amp-seconds)
Heat flow rate	Watts	Amps
Thermal resistance	°C/Watt	Ohms
Thermal capacity	°C/Joule	Farads
Heat source	Dissipative element, eg transistor	Current source
Ambient	Medium-sized planet	Voltage source

ulator ground, which is inherently at 0°C (0V).

Values of the notional components in Fig. 4 have to be filled in with a mixture of calculation and manufacturer's data. Thermal resistance R_1 from junction to case comes straight from the data book. So does the resistance R_2 of the TO3 thermal washer, also R_4, the convection coefficient of the heatsink itself, otherwise known as its thermal resistance to ambient. This is always assumed to be linear with temperature, which it very nearly is. Here R_4 is 1°C/W, so this is doubled to two as the stage is cut in half to exploit symmetry.

Resistor R_3 is the thermal resistance of the graphite foil. This is cut to size from a sheet and the only data is the bulk thermal resistance of 3.85W/mK, so R_3 must be calculated. Thickness is 0.2mm, and the rectangle area in this example was 38×65mm. You must be careful to convert all lengths to metres. Heat flow per °C is,

$$\frac{3.85 \times Area}{Thickness}$$

$$= \frac{3.83 \times (.038 \times .065)}{.0002}$$

$$= 47.3W/°C$$

So thermal resistance is,

$$\frac{1}{47.3} = 0.021°C/W$$

Thermal resistance is the reciprocal of heat flow per degree, so R_3 is 0.021°C/W, which just goes to show how efficient thermal washers can be if they do not have to be electrical insulators as well.

In general all the thermal capacities will have to be calculated, sometimes from rather inadequate data, thus *thermal capacity* is *density×volume×specific heat.*

A power transistor has its own internal

Table 2 Parameters for TO3 and TO-225AA with SW38-1 heatsink.

		Output device	Driver
C_1	Junction capacity J/°C	0.1	0.05
R_1	Junction-case resistance °C/W	0.7	6.25
C_2	Transistor package capacity	3.0	0.077
R_2	Thermal washer res	0.4	6.9
C_3	Coupler capacity	70	—
R_3	Coupler-heatsink res	0.021	—
C_4	Heatsink capacity	250	20.6
R_4	Heatsink convective res	2.0	10.0

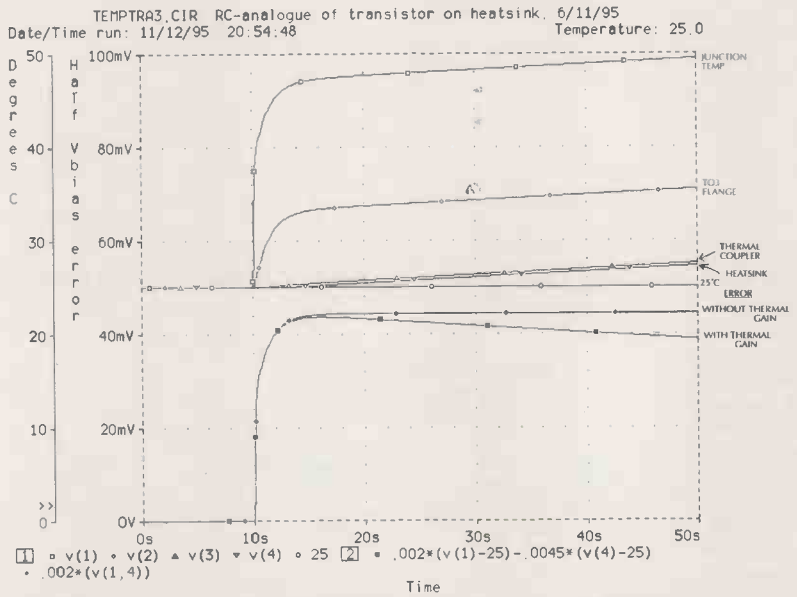


Fig. 6. Results for Fig. 4, with step heat input of 20W to junction initiated at time=10s. Upper plot shows temperatures, lower the V_{bias} error for half of output stage.

structure, and its own internal thermal model, Fig. 5. This represents the silicon die itself, the solder that fixes it to the copper header, and part of the steel flange the header is welded to. I am indebted to Motorola for the parameters, from an *MJ15023* TO3 device[7]. The time constants are all extremely short compared with heatsinks, and it is unnecessary to simulate in such detail here.

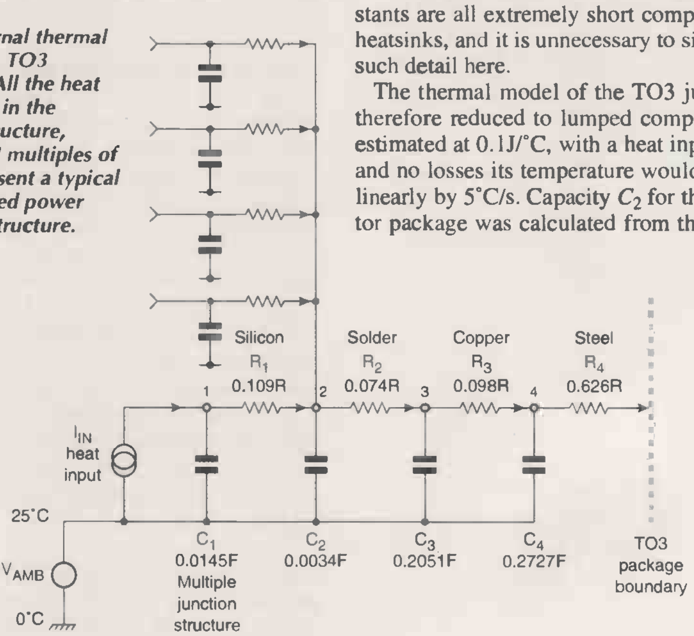
The thermal model of the TO3 junction is therefore reduced to lumped component C_1, estimated at 0.1J/°C, with a heat input of 1W and no losses its temperature would increase linearly by 5°C/s. Capacity C_2 for the transistor package was calculated from the volume of the TO3 flange, representing most of the mass, using the specific heat of mild steel.

The thermal coupler is known to be aluminium alloy, not pure aluminium, which is too soft to be useful, and the calculated capacity of 70J/°C should be reliable. A similar calculation gives 250J/°C for the larger mass of the aluminium heatsink.

Our simplifying assumptions are rather sweeping here, because we are dealing with a substantial chunk of finned metal which will never be truly isothermal.

Derived parameters for both output TO3 and T0-225AA drivers are summarised in **Table 2**. The drivers are assumed to be mounted onto small individual heatsinks with an isolating thermal washer. The data is for the popular Redpoint *SW38-1* vertical heatsink.

Fig. 5. Internal thermal model for a TO3 transistor. All the heat is liberated in the junction structure, shown as N multiples of C_1 to represent a typical interdigitated power transistor structure.



Thermal transient effects

Figure 6 shows the result of a step function in heat generation in the output transistor. Twenty watts dissipation is initiated, corresponding approximately to a sudden demand for full sinewave power from a quiescent 100W amplifier. The junction temperature V(1) takes off near vertically, due to its small mass and the substantial thermal resistance between it and the TO3 flange.

Flange temperature V(2) shows a similar

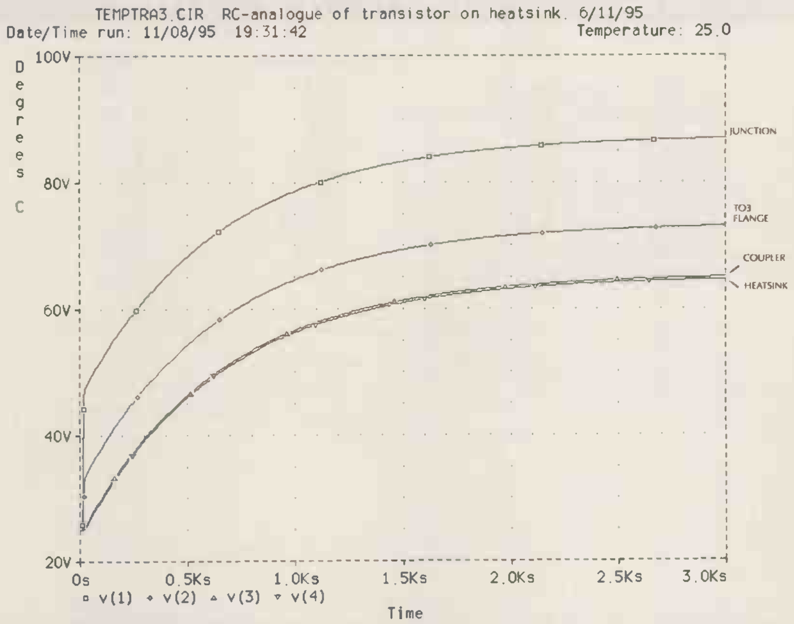
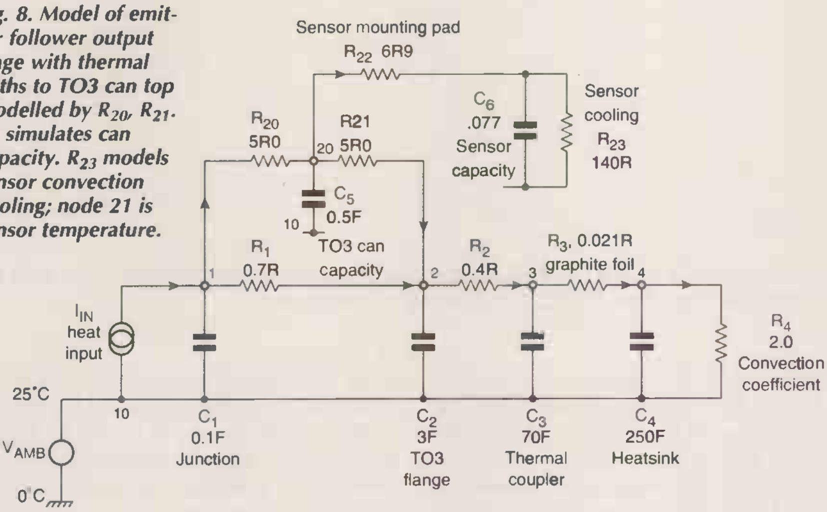
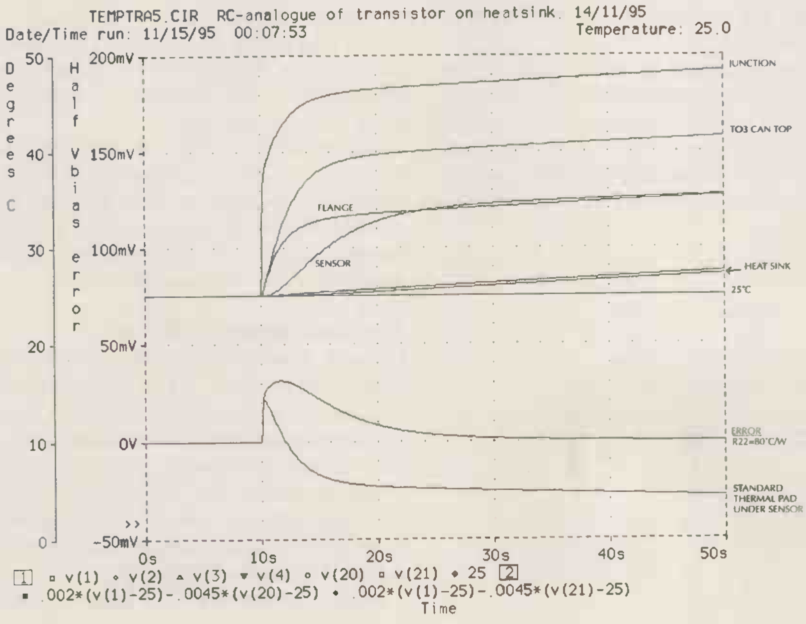


Fig. 7. Long term version of Fig. 6 shows that it takes over 40min for the sink to get within 1° of final temperature.

Fig. 8. Model of emitter follower output stage with thermal paths to TO3 can top modelled by R_{20}, R_{21}. C_5 simulates can capacity. R_{23} models sensor convection cooling; node 21 is sensor temperature.





but smaller step as R_2 is also significant. In contrast, the thermal coupler, which is so efficiently bonded to the heatsink by graphite foil that there might almost be one piece of metal, begins a slow exponential rise that will take a very long time to reach asymptote. After the effects of C_1 and C_2 have died away the junction temp is offset by a constant amount from the temp of C_3 and C_4, so V(1) also shows a slow rise. Note the X axis must be in kiloseconds, because of the relatively enormous thermal capacity of the heatsink.

This shows that a temperature sensor mounted on the main heatsink can never give accurate bias compensation for junction temperature, even if it is assumed to be isothermal with the heatsink. In practice there will be some 'sensor cooling' which will make the sensor temperature read slightly under the heatsink temperature V(4).

Initially the temperature error V(1)–V(4) increases rapidly as the TO3 junction heats, reaching 13° in about 200ms. The error then increases much more slowly, taking 6s to reach the effective final value of 22°.

If you ignore the 'thermal-gain' effect mentioned above, the long term V_q error is +44mV, ie V_q is too high. When this is doubled to allow for both halves of the output stage we get +88mV, which uses up nearly all of the ±100mV error band, without any other inaccuracies.

Hereafter all V_{bias}/V_q error figures quoted have been doubled and so apply to a complete output stage. Including the thermal gain actually makes little difference over a 10s timescale, the lower V_q error trace in Fig. 6 slowly decays as the main heatsink warms up, but the effect is too slow to be useful. Amplifier V_q and I_q will therefore rise under power, as the hot output device V_{be} voltages fall, but the cooler bias generator on the main heatsink reduces its voltage by an insufficient amount to compensate.

Figure 7 shows long term response of the system. At least 2500s pass before the heatsink is within a degree of final temperature.

As to where to mount the sensor...

In the past I have recommended that emitter follower output stages should have the thermal sensor mounted on the top of the TO3 can – despite the mechanical difficulties. This is not easy to simulate as no data is available for the thermal resistance between junction and can top. There must be an additional thermal path from junction to can, as the top definitely gets hotter than the flange measured at the very base of the can. In view of the relatively low temperatures, this path is probably due to internal convection rather than radiation.

A similar situation arises with TO3P – a large plastic package, twice the size of TO220, for the top plastic surface can get at least 20° hotter than the heatsink just under the device.

Using real thermocouple data[3], I have esti-

Fig. 9. The simulation results for Fig. 8; lower plot shows V_{bias} errors for normal thermal pad under sensor, and 80°C/W semi-insulator. The latter has near zero long term error.

mated the parameters of the thermal paths to the TO3 top. This gives **Fig. 8**, where the values of elements R_{20}, R_{21} and C_5 should be treated with considerable caution, though the temperature results in **Fig. 9** match reality fairly well. The can top (V20) gets hotter faster than any other accessible point. Resistor R_{20} simulates the heating path from the junction to the TO3 can and R_{21} the can-to-flange cooling path, C_5 being can thermal capacity.

Figure 8 includes approximate representation of the cooling of the sensor transistor, which now matters. Resistor R_{22} is the thermal pad between the TO3 top and the sensor, C_6 the sensor thermal capacity, and R_{23} is the convective cooling of the sensor, its value being taken as twice the data sheet free air thermal resistance as only one face is exposed.

Putting the sensor on top of the TO3 would be expected to reduce the steady state bias error dramatically. In fact, it overdoes it. After factoring in the thermal gain of a V_{be} multiplier in an emitter follower stage, the bottom most trace of Fig. 9 shows that the bias is overcompensated.

Following the initial positive transient error, V_{bias} falls too low giving an error of –30mV, slowly decreasing as the main heatsink warms up. If thermal gain had been ignored, the simulated error would have apparently fallen from +44, Fig. 6, to +27mV, apparently a useful improvement, but actually illusory.

Since the new sensor position overcompensates for thermal errors, there should be an intermediate arrangement giving near zero long term error. I found this condition occurs if R_{22} is increased to 80°C/W, requiring some sort of semi-insulating material rather than a thermal pad, and gives the upper error trace in the lower half of Fig. 9. This peaks at +30mV after 2s, and then decays to nothing over the next twenty. This is much superior to the persistent error in Fig. 6, so I suggest this new technique may be useful.

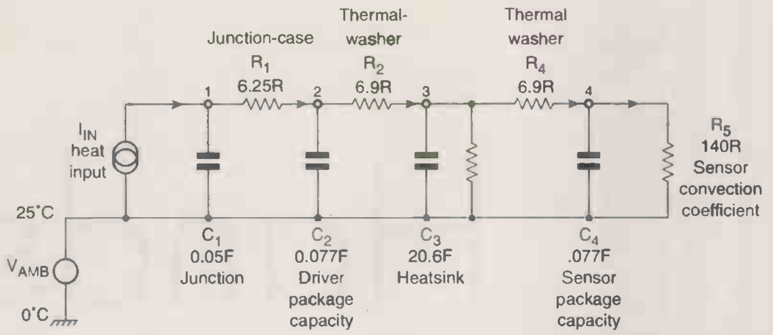


Fig. 10. Model of a cfp stage. Driver transistor is mounted on a small heatsink, with sensor transistor on the other side. Sensor dynamics and cooling are modelled by R_4, C_4 and R_5.

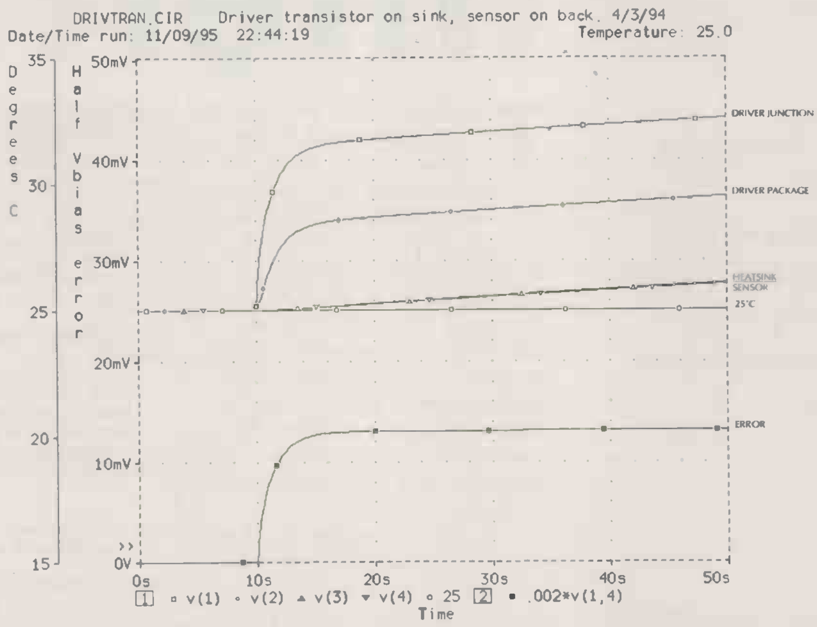


Fig. 11. Simulation of cfp stage, with step heat input of 0.5W. Heatsink and sensor are virtually isothermal, but there is a persistent error as driver is always hotter than sink due to R_1, R_2.

Modelling the cfp output

Turning to the complementary feedback pair output stage it is the driver junctions that count, output device temperature has little effect and is neglected.

Thermal parameters for a *TO-225AA* driver, for example *MJE340/350* on an *SW38-1* vertical heatsink, are shown in **Table 2**. The drivers are on individual heatsinks so their thermal resistance is used directly, without doubling.

In the simulation circuit **Fig. 10**, V(3) is the heatsink temperature. The sensor transistor, also *MJE340*, is mounted on this sink with thermal washer R_4, and has thermal capacity C_4. Resistor R_5 is convective cooling of the sensor. In this case the resulting differences in **Fig. 11** between sink V(3) and sensor V(4) are very small.

You might expect the feedback pair delay errors to be much shorter than in the emitter follower. However, simulation with a heat step input suitably scaled down to 0.5W Fig. 11, shows changes in temperature error V(1)–V(4) that appear rather paradoxical. The error reaches 5° in 1.8s, levelling out at 6.5° after about 6s. This is markedly slower than the emitter follower case, and gives a total bias error of +13mV. After doubling to +26mV, this is well outside the feedback pair error band of ±10mV.

The initial transients are slowed down by the much smaller step heat input, which takes longer to warm things up. The 'final' temperature however, is reached in 500 rather than 3000s, and the timescale is now in hundreds rather than thousands of seconds. The heat input is smaller, but the driver heatsink capacity is also smaller, and the overall time constant is less.

It is notable that both timescales are much longer than musical dynamics.

In summary

For these simulations at least, the results are unexpected. I thought that the complementary feedback pair would show smaller bias errors than the emitter follower, but it is the follower that stays within its much wider tolerance bands, with either heatsink or *TO3* top mounted sensors. The thermal gain effect in the emitter follower stage seems to be the root cause of this.

It is clear that thermal attenuation and delay between transistor and sensor still cause significant errors in both stages. Ways to further reduce these shortcomings will be presented in the next article. ■

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Hands-on Internet

Cyril Bateman discusses the benefits and pitfalls of active browsers for the World Wide Web.

The World Wide Web[1] as it exists today, forms the most popular entry point into Internet resources. As a concept, from its restricted beginnings as a scientific information linking tool at CERN in Switzerland in 1980, it has developed in its latest incarnation using Java into the HotJava and Netscape2 browsers. These potentially offer the most useful and desirable Web access tools, but if not properly controlled, Java could provide the opportunity to be developed into a potentially dangerous hackers' tool.

In essence, to access the Web resources requires use of 'browser' software[2] on your personal computer. Versions are now available for all operating systems and platforms. The normal browser is benign and largely dumb software, used to passively interpret the HTML commands buried within the transmission. Just as with printer commands buried within a word processor document, these commands remain invisible to the user, unless the document is viewed by a different editor, when they become revealed for all to see, **Fig. 1**.

Java began development four years ago as a new language, from Sun Microsystems[3]. It was intended to be used for controlling embedded systems or smart appliances. While still in its Beta stages, it provides a method for seamlessly integrating small programs called 'applets' into your system, **Fig. 2**.

On its own Java, previously known as Oak, was useful enough. However, with the release of the NCSA Mosaic 1.0 Web browser in mid 1993, its potential use within a Web browser became apparent, culminating in Sun's browser HotJava and Netscape's Navigator2[4]. Versions are available for Windows'95, Windows NT,

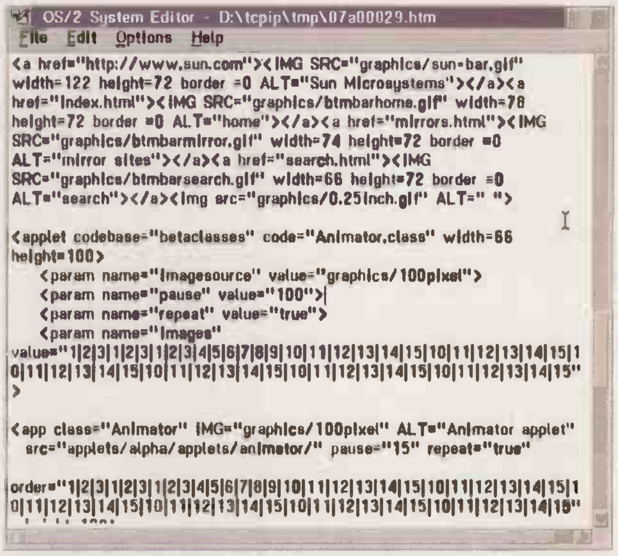

Fig. 1. Viewing the 'hidden' HTML script from http://java.sun.com home page. Lines 1 to 8 illustrate hidden html coding revealed when viewed using editor. Lines 11 onward illustrate hidden Java applet coding now revealed when viewed using editor.

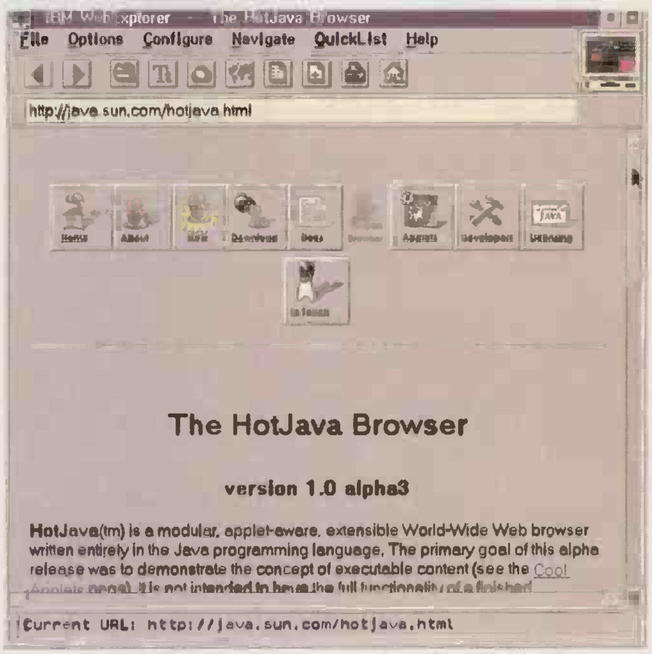

Fig. 2. Sun Microsystems HotJava page from http://java.sun.com. This explains the intentions and current status of the HotJava software.

Fig. 3. The home page of the Netscape Navigator 2 browser http://home. netscape.com .Outlines the current status and platform availability of Navigator2. Provides facility to download their latest browser software. Notice 'hatched out' applet sections above main graphic.

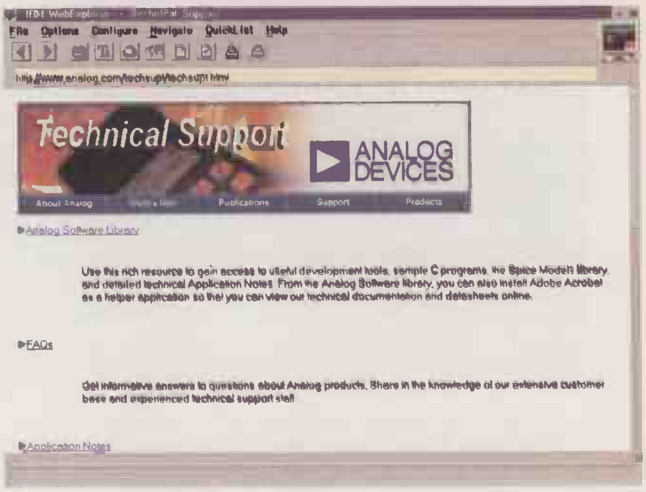

Fig. 4. Technical assistance offer from Analog Devices page http://www.analog.com. Provides download of their software libraries, also Spice macromodels. There's technical design assistance in abundance.

Fig. 5. Harris Semiconductors' home page at http://www.semi.harris.com. Many pages of design assistance are on offer under their 'Design Made Simpler' topic. Provides design support software as well as their excellent macromodel library.

Sparc and Solaris, Fig. 3.

Java and HotJava suddenly became the hottest Internet topics. An Archie search on Java reveals hundreds of active topics, including not a few recipes. Obviously, the word Java on the Internet is not exclusive to Sun's software.

Why all this intense interest in Java and HotJava? Well, they provide the facility to transmit 'executable content' within a Web page or computer program. And how is this significant to electronic designers? Well maybe at present it has just as much significance to all computer users, but consider the implications.

Older non-Java Web browsers were simple program interpreters of the HTML instructions received. Java aware browsers have the ability to transparently accept and action an applet. This is a short computer program which could, for example, perform an animated logo, then discard it when its task is complete. On the other hand, without the necessary controls on the logo originator, this applet could perform any other computer task, good or bad, offering the potential to change the face of software purchase as known today. Or it could perhaps be used to introduce a virus – or even reformat one's hard disk, Fig. 1.

To protect the user, all Java aware Web browsers automatically check the authenticity of each applet downloaded. They do this by first making sure it has been compiled using an official Java compiler. Then a sophisticated checksum routine is invoked to ascertain that the originator is a registered Java programmer. Thus at present, Java is simply yet another way to safely liven up your Web pages.

Non-Java aware browser software 'hatches out' any Java animated applets. But just as with the printer instructions analogy, examination of a Java HTML script in an editor reveals the applet program, Figs 1, 3.

The potential for good outside a Web page is the potential future availability of low cost software applets. For example, such applets may be designed for say the new self-assessment tax return calculations, downloaded and paid for on the Web. The applet is then discarded when its task is finished.

I advise those of you interested to visit Sun's Java page[3] where much information is available from the Java and HotJava FAQ, and the browser software can be downloaded and tried for real.

Fig. 6. National Semiconductor's home page at http://www.national.com. clearly offering technical support and Spice macromodels for download. The quick search by part number is worth trying.

Fig. 7. LSI Logic Corporation's home page at http://www.lsilogic.com. Click on 'What's New' to find their press releases at /mediakit/unit3_lxhtml.

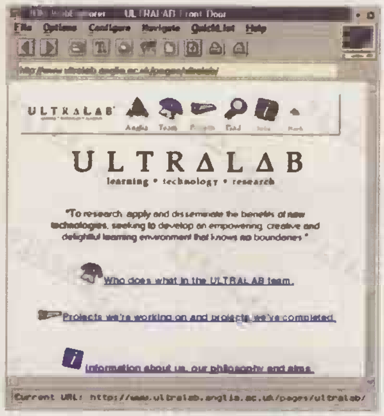

Fig. 8. University of East Anglia at Norwich at http://www.ultralab.anglia.ac.uk also offers downloads of this department's issued reports. Much reading here.

Following last month's PSpice topic, in the past year, many North American semiconductor houses have been developing their own Web pages. National[5], Analog[6], Burr-Brown[7] and Harris[8], all have now established pages. In its own unique way, each has something special to view, in addition to offering downloadable application software and Spice macromodel libraries.

Try downloading Netscape from the Burr-Brown[7] page. It might be less busy than Netscape's own site, **Fig. 4,5,6**.

A visit to LSI Logic Corporation's site[9] revealed a December 1995 press release which promises a dramatic impact on Internet access by offering a single chip Internet architecture able to make a sub-$500 tv based access system. This chip uses 0.25μm process, with up to 100 million instructions a second at a quantity price around $50, **Fig. 7**.

You might be forgiven for believing that the only useful Internet information is US based. Not so. The UEA at Norwich[10] has been working with BT, Apple and the BBC on interface issues for BT's interactive television trials, **Fig. 8**.

Questions or comments on this article can be sent directly to me at the following email addresses.
cyrilb@ibm.net or
76251,2535@compuserve ■

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High-performance mic preamplifier

Simon Bateson believes that he's produced the ultimate mic preamp.

Although many articles have appeared in various journals on the subject of microphone preamplifiers, there have been relatively few attempts at viewing the problems of real-life use and abuse, where the highest standards of performance are required along with a complete immunity to the all too common phantom-power accident, **Fig. 1**.

Far from being a completely solved and extinct topic, the microphone preamplifier has become an increasingly exotic, specialised and expensive item in the recording studios. Valve designs possessing 'warmth' and 'character' compete with up-to-date rack units with inbuilt matrices for MS stereo, filters, limiters and level meters. I have seen such a unit reviewed very recently. It is a straightforward design using the low cost *SSM2017* chip – yet costs well over £400.

My design evolved around the *SSM2016* differential amplifier IC. This device has a much higher specification than the *2017*. It is specifically and solely designed for low-impedance low noise applications such as microphones and virtual earth busses in mixing consoles. It has some remarkable properties, **Table 1**. Its output figures endow it with tremendous dynamic range when employed as a bus mix stage. Emphasis in this design, however, is on minimal input noise and distortion, so it is run at a moderate ±18V to stay cool.

Microphone amplifier topology

An ordinary differential op-amp circuit is shown in **Fig. 2**. It has several problems which limit its ability to reject common-mode signals: in particular, rejection depends on perfectly balanced resistors and on the differential source having zero impedance. Two resistors need changing to alter gain, and the input resistors add to the circuit noise. All in all, very unsatisfactory. The instrumentation amplifier, **Fig. 3**, has several advantages over the ordinary differential amplifier:

- There is a high and equal input impedance at both inputs, making common mode rejection independent of source impedance.
- Gain is adjustable from unity upwards by a single resistor.
- Very high gain and cmrr are available without careful resistor matching

The gain/cmrr benefit occurs because all the differential gain is obtained before the differencing stage. Suppose a common mode signal is applied to both inputs, by op-amp action the inverting inputs are also both at the same voltage. Hence there is no voltage across R_I and no current flows through it. It can be ignored making the com-mon-mode gain of the first amplifiers just 1. This eases the rejection of common-mode signals by the differencing stage which therefore needs less carefully matched resistors.

In a conventional microphone preamp, the

Table 1. Microphone preamplifier specifications.

Maximum supply voltage	±36V
Maximum output current	±40mA
THD at 10Vrms out 1kHz	0.009%
	10kHz
0.015%	
Bandwidth	500kHz
CMRR	100dB
Input noise (150Ω source)	0.11μV in a 20kHz bandwidth (0.8nV per √Hz)

Note: all figures typical at 1000 gain.

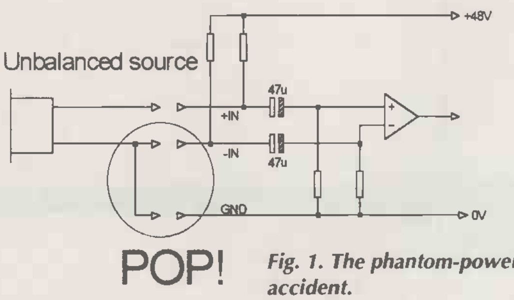

Fig. 1. The phantom-power accident.

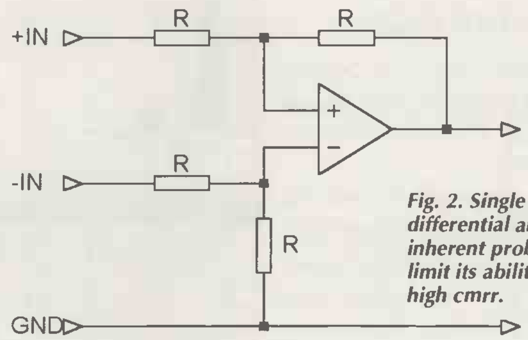

Fig. 2. Single op-amp differential amplifier has inherent problems that limit its ability to provide high cmrr.

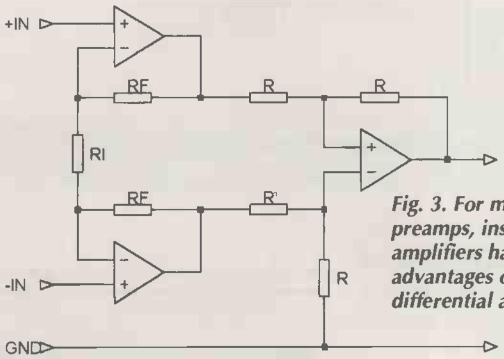

Fig. 3. For microphone preamps, instrumentation amplifiers have several advantages over the common differential amplifier.

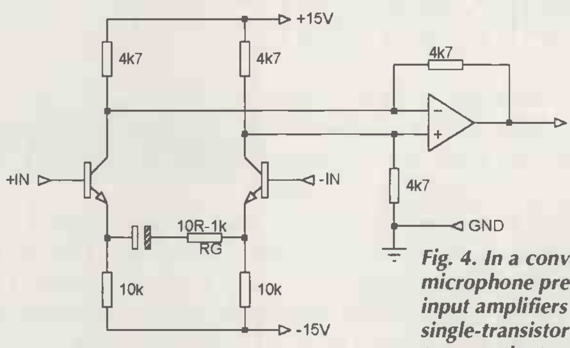

Fig. 4. In a conventional microphone preamp, the input amplifiers are just single-transistor transconductance stages.

input amplifiers are not op-amps. They are just single-transistor transconductance stages, which are fairly linear over small differential excursions, followed by a differential current to voltage converter, Fig. 4.

Gain considerations

The gain setting resistor is connected between the emitters and so is effectively in series with the signal path. It has a low value just when we need the lowest circuit path resistance, namely, at high gains.

It is very important to realise that, as this resistor changes, it affects both the open-loop and the closed-loop gain equally, so we can overcome the usual op-amp limitation of the fixed gain-bandwidth product. In the *2016*, we can have total stability at low gains and still have a 500kHz bandwidth at a gain of 1000.

To be sensible, we can increase the feedback capacitor to reduce the bandwidth a little, ensuring gain and phase flatness without encouraging radio reception or exposing subsequent circuitry to excessive out-of-band signals.

In the *2016*, very large geometry input transistors, fabricated in a 'super-matching' process, act as the input amplifiers. They have very low base spreading resistance and when fed from the low optimum source resistance of 150Ω, have a noise figure of just 1dB.

Extra circuit details in the *2016* prevent the transistors suffering from gain modulation, and hence distortion, at high input levels. Note that, in the circuit of Fig. 4, a high value electrolytic capacitor is required to keep the dc differential gain down. It is better to omit the electrolytic, particularly since it is not polarised properly. It is possible do this with the *2016* because the input transistors are super-matched and closely linked thermally.

In most mic preamps, the gain is varied continuously with a potentiometer and series stopper resistor. This is unsatisfactory for many reasons. The contact resistance of a typical pot is variable and noisy, and since the gain is inversely proportional to resistance, the resulting calibration is hopelessly non-linear and non-repeatable. There is no need for continuous control anyway, since a fader always appears later in the signal path.

I have used switched resistors to give accurate gains from 20 to 60dB in 10dB steps and this has been perfectly adequate and trouble-free. The overall design of the gain stage, Fig. 5, requires little extra comment other than to say that I have not used the manufacturer's optional output dc offset trim. The output is ac coupled anyway, (see below) and so you only need to trim the input devices with PR_2, to prevent dc thumps occurring when the gain is switched.

Input/phantom power facilities

There are two main types of microphone which will be used with this amplifier. The first is the dynamic moving coil or ribbon microphone. This type has a low impedance of 150 to 600Ω and low sensitivity, hence low voltage output. To give you an idea of the signal levels involved, a popular typical microphone, the Shure *SM58LC*, has a quoted sensitivity of –77.5dB. Here, 0dB is referred to 1V/μbar, and 1μbar (1 atmosphere) is about 10^5Nm^{-2} so the reference level is 0.1Nm^{-2}.

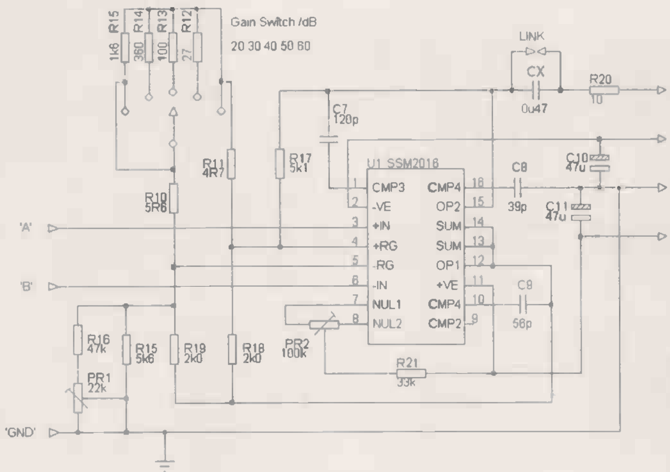

Fig. 5. The SSM2016 gain stage has an optional trim for dc offset.

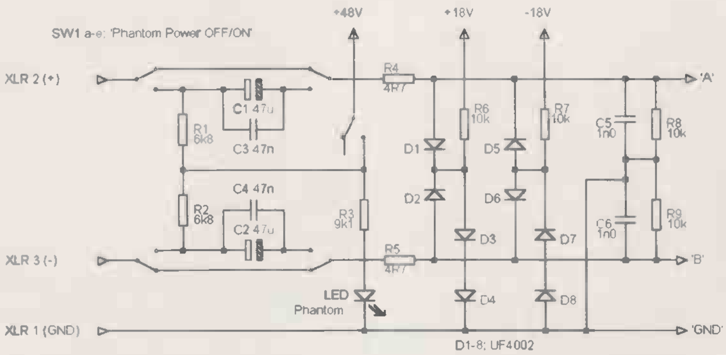

Fig. 6. Input phantom power switching and protection of the new microphone preamplifier.

The commonly accepted threshold of hearing, 0dBA, is $2\times10^{-5}\text{Nm}^{-2}$ so 1μbar is 5000 times higher sound-pressure level than this, namely, 74dBA. This is the equivalent of loud conversation or the sound of a vacuum cleaner from a few metres.

At this sound level the microphone delivers a voltage 77.5dB below 1V. That is, the princely sum of 0.13mV.

Clearly, the amplifier must have a very low voltage noise figure and we must minimise circuit impedances to minimise current noise contributions.

In addition, since the signal may have travelled through many metres of hopefully good quality microphone cable, a high level of common mode interference may be present. This interference will start at 50Hz (or 60Hz) and can contain many high level harmonics, especially if phase-controlled lighting is in use. Clearly the best way to connect this signal is directly to the amplifier input stage, without intervening impedances or anything which could lower the amplifier's common-mode rejection ratio.

The second common microphone type is the condenser, either 'real' or electret. These mics contain on-board buffers and are phantom powered. They are far more sensitive than dynamic types and their noise figures are usually limited by their internal electronics, rather than by the mic amp they feed. The phantom power feed is defined by a DIN standard and is commonly implemented with 6.8kΩ resistors from a 48V supply. Isolation between the line and the preamp at dc is ensured either by transformer or capacitors.

Capacitive coupling

High quality audio transformers are very expensive so the majority of solid-state microphone amplifiers use capacitors. These must have a very high value, at least 47μF, to offer sufficiently low impedances in the audio band. At least these capacitors are well and truly polarised so you don't need to worry too much about them being electrolytic. However, there are two other problems.

Firstly, these input capacitors are incapable of being matched, even the best quality modern electrolytics have a ±20% tolerance. The capacitors interact with the amplifier input bias resistors. These usually have a low value, around 4.7kΩ, to form a high-pass filter.

For low frequency signals, the capacitor mismatch causes an attenuation mismatch and a phase shift, which renders the 100dB common-mode rejection ratio of the amplifier stage rather helpless. Often, high quality preamps will have a common-mode rejection ratio trim control so you can make an adjustment at 50Hz but what about 150Hz?

Technical support

A set of five circuit boards – input switch and preamp ×2, dual led meter, dual filter, three rail psu – is currently being prepared. Please send an sae for details to Electronics World, Room L333, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.

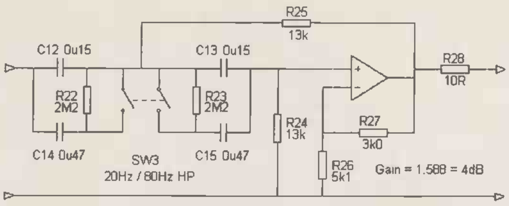

Fig. 7. Contrary to popular belief, bass is a nuisance in microphone applications, hence this 20/80Hz high-pass filter.

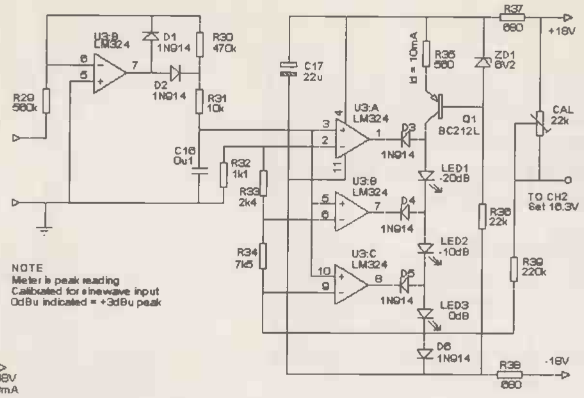

Fig. 8. Three-led peak level meter helps choose the right gain setting.

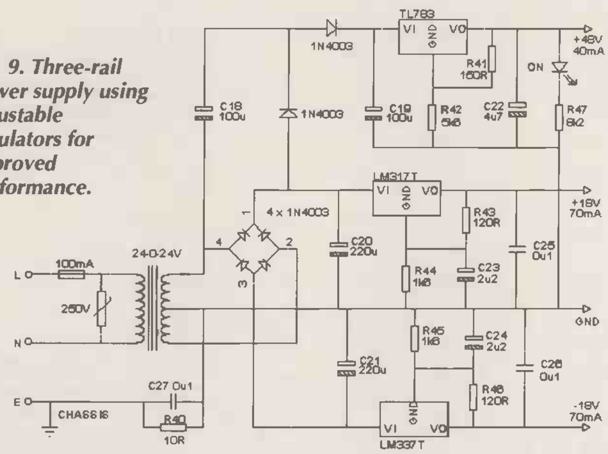

Fig. 9. Three-rail power supply using adjustable regulators for improved performance.

Input protection

Secondly, these capacitors are charged up to quite a high voltage. If an unwary user plugs in an unbalanced or faulty microphone lead or a lead from other devices, the input terminals can be suddenly shorted. The capacitors, one or both, must then discharge, Fig. 1, through the nearest available path, either the input transistors of the preamp, causing damage, or through some protection route.

The most common protection components are a pair of zener diodes connected between each input line and ground. However I am dubious about this practice. Zener diodes have a very non-linear junction capacitance when unbiased and this could be a source of distortion. Also, with less subtlety, ordinary zeners often just do not survive capacitor discharge and fail short-circuit.

So, my search was for a fairly economical circuit which would permit the best possible input route for dynamic microphones, coupled with adequate protection for the rather expensive *2016*. The resulting design is shown in **Fig. 6**.

The multipole switch completely bypasses the coupling capacitors. When in circuit, they are paralleled with polyesters to maintain a low impedance at high frequencies.

Protection is assured by commonly available ultra-fast rectifiers *UF4002*s. These are reverse biased by a couple of volts. Reverse biasing them by direct connection to the supply rails would provide insufficient protection for the *2016* and it would encourage noisy leakage current.

Zero reverse bias would permit non-linear capacitance variation and forward conduction with large input signals. As it is, these diodes will not limit the input until it reaches around ±1.8V, by which time the *2016* output will be well and truly clipped. The 4.7Ω resistors limit the maximum fault current to 10A – well within the surge rating of the diodes – without adding too much series resistance to the input signal path.

The effectiveness of this protection can be judged by the way in which you can draw many, many sparks from the input terminals without the slightest distress to the preamp. I considered removing the diodes from the circuit for dynamic mics but concluded that the signal levels were too low to cause any problems and that the diodes would always be useful if the mic input was fed from a speaker cable, which can happen.

These two sections of circuitry fit neatly onto a single pcb and I have to emphasise that the exact layout, especially in terms of earthing, of high-performance circuitry is far more important than whether all the capacitors are made of polyester, polycarbonate or diamond-impregnated Teflon.

Extra facilities

The preamp boards need a certain amount of support circuitry. A power supply, obviously, and I have found a simple level indicator and high-pass filter most useful.

Bandwidth and low-frequency phase-shifts in audio systems have been discussed in these pages before. But it has to be admitted that an extended lf response in vocal and most instrument microphones is a bad thing. It permits boominess, causes a general lack of clarity and wastes amplifier and loudspeaker power.

If we can clearly define the lf extension of the system at the input we can save a lot of discussion further on. This design, **Fig. 7**, breaks no new ground in the use of an equal-value Butterworth response Sallen-Key filter. This is a good circuit in high-pass form, and is switchable between 20Hz 'full bandwidth' and 80Hz roll off.

As far as level indication is concerned, I believe that this design will find most applications in direct DAT recording or in studios. Here, there will be plenty of facilities for level monitoring. The function of metering here must therefore be as an aid to choosing the right gain setting, in 10dB steps, so a three led circuit is ample.

Although not entirely new, the design, **Fig. 8**, is worthy of comment because it is specifically designed not to cause disruption to the power supplies or earth, which a lot of careless designs will do. Supply current runs between the rails, not to ground, having first been reduced in voltage and decoupled. A constant current source feeds the led chain and supply current is diverted into the amplifiers in order to extinguish them. As a result, the total current drawn is barely affected by the number of leds lit.

Level indicators using standard comparators such as the *311* can induce clicks into other circuitry even if the supply is decoupled. This is due to the high comparator switching speed. The *324* or *3403* quad op-amps, with their leisurely slew rate, low supply current, low input offset and low price are perfect for this.

A further board holds the transformer and regulators for a three rail power supply, **Fig. 9**, and this is dimensioned to supply ample current for a stereo application. The expense of an special transformer is saved by using a simple voltage doubler, along with a *TL783* high-voltage regulator, for the 48V rail.

If more than about four preamps are to be constructed it may prove more economical to use a separate transformer and regulator board for the 48V supply.

Implementation and setting up

Board layout for the microphone preamp is important to its success.

The output capacitor on the preamp board should be replaced with a wire link if the high-pass filter circuit is going to be used. There are just two presets to adjust.

Input offset adjustment should be trimmed to eliminate the clicks which occur on changing gain. Then, a signal of a few volts at around 150Hz can be applied in common mode to both inputs, with the phantom switch off, and the common-mode rejection ratio adjusted for maximum signal rejection. Then you should have a preamplifier beyond criticism.

After about 15 years of experimentation this is, for me, the final microphone preamp. ■

NEW PRODUCTS CLASSIFIED

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ACTIVE

A-to-D and D-to-A converters

20-bit A-to-D. From Asahi Kasei, the *AK5350* 20-bit, 64-times-oversampling, two-channel analogue-to-digital converter, which has a 100dB dynamic range and sinad of 94dB; maximum sampling rate is 54kHz. Power needed is 115mW at 5V. The companion *AK5391* is a similar design, but is a 24-bit type providing 115dB of dynamic range. DIP International Ltd. Tel., 01223 462244; fax, 01223 467316.

Discrete active devices

Power mosfets. Four fifth-generation Hexfets in SOT-223 from IR exhibit up to 80% less on resistance than earlier devices in this package. As an example, the 55V *IRLL2705* has an on resistance of 40mΩ and the 30V *IRLL3303* 31mΩ. The range includes both n-channel and p-channel devices. International Rectifier. Tel., 01883 713215; fax, 01883 714234.

Digital signal processors

DSP development. From Analog Devices, the *EZ-KIT Lite* development system for the *ADSP-2100* family of processors, which allows evaluation, development, debugging and prototyping of digital signal processing applications. Kits include a development board with 16-bit stereo audio i/o, assembly, linker and simulation software, pc host software, dsp algorithm source code and accessories, together with demonstration programs. A 16-bit *ADSP-2181* board in the kit has 32Kword of ram, dma ports and power management and, running the MPEG audio decode demonstration, it plays around 7s of audio without external ram. Analog Devices Ltd. Tel., 01932 266000; fax, 01932 247401.

Linear integrated circuits

1100V/µs amplifier. *MAX477* from Maxim is a ±5V, 300MHz amplifier that is in two stages, namely a current-mode feedback amplifier for wide bandwidth and high slew rate, and a voltage-feedback amplifier for low offset volts, low noise and distortion. The result slews at 1100V/µs, is flat to within 0.1dB to 130MHz, puts out 100mA and drives 100pF without drama. Differential phase and gain are 0.01%/0.01°, thd −78dBc at 10MHz, voltage noise density 5nV/√Hz and settling time 12ns. Maxim Integrated Products UK Ltd. Tel., 01734 303388; fax, 01734 305511.

12-bit a-to-d converters. Analog's *AD9220* 12-bit analogue-to-digital converter is the first in a new series of all-cmos devices offering lower cost and high-speed, high-resolution. It uses one 5V supply, taking 280mW at 10Msample/s, and has an on-chip s/h amplifier and settable 1/2.5V reference, the s/h amplifier being configurable for single-ended or differential working. Sinad ratio is 68dB, dynamic range 75dB and differential non-linearity 0.25 of the least significant bit. Analog Devices Ltd. Tel., 01932 266000; fax, 01932 247401.

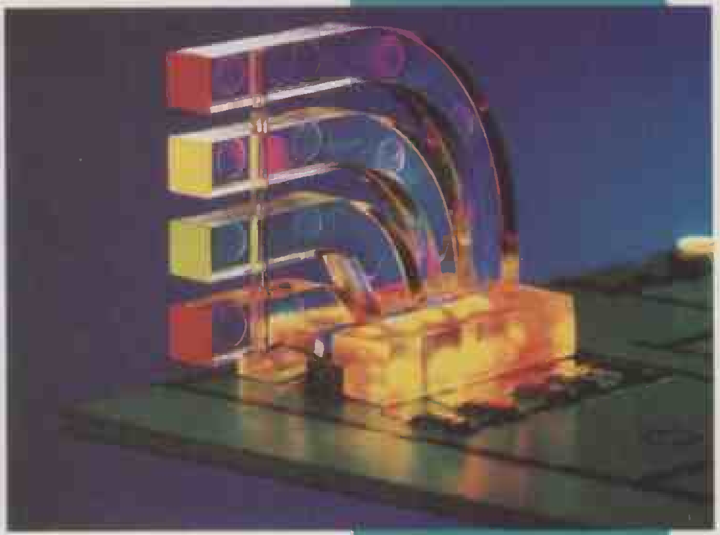
Single-supply video amplifier. *EL2150* from Elantec is a single or dual-supply op-amp that will work on a single 2.7V supply, offering a 125MHz bandwidth and 255V/µs slewing. With a single supply, the output can swing completely to ground without saturating and to within 1.2V of the supply voltage, also sensing voltages below the bottom supply rail and to within 1.2V of the top rail. Reactive feedback is permitted; differential gain is 0.05% and differential phase 0.05°. The output stage delivers ±100mA. Elantec. Tel., 0171-482 4596; fax, 0171-267 1026.

Memory chips

2.7V, 8Mb memory. Fujitsu has the first of a new family of single-supply, 2.7V flash memories, the *MBM29LV*. It is readable, programmable and erasable at 2.7V and is intended for digital cellphones and portable computers. The memory is organised as 1Mb by 8 and 512K by 16 and gives a 120ns access time. Fujitsu Microelectronics Ltd. Tel., 01628 76100; fax, 01628 781484.

Mixed-signal ICs

Stepper driver. Allegro's *A3961SB* stepper motor driver is a dual full-bridge type providing pwm current control of bipolar steppers, delivering continuous output of ±750mA at 5 to 45V. Internal, fixed off-time control circuitry and an internal 2.5V reference need only an external resistive divider and current-sensing resistors, the off-time duration being set by external timing resistors. Full protection is provided. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.



Light pipes. Dialight offers an addition to its *Optopipe* range of light pipes, which carry light from surface-mounted leds to a front panel. The units have a wide viewing angle and, since one unit gives four indicators, reduce cost. *Optopipes* are in plastic, are mounted on the printed board after reflow and carry the light, usually at right angles to the board, to the panel. Dialight. Tel., 01638 662317; fax, 01638 560455.

Optical devices

Laser diode amplifier. Elantec's *EL6251* laser diode amplifier incorporates a sense amplifier responding to input from a laser-power monitor diode, voltage-controlled driver and 5V power-supply monitor, providing pulsed write/erase current and dc or pulsed read current to a common-cathode laser diode. Read and write currents are 160mA each with transient times of 2.4ns. The read output may be inductively isolated from both the write output and laser diode. METL. Tel., 01844 278781; fax, 01844 278746.

Dual opto-isolator. *ISP827* from Isocom contains two independent optical isolators in an eight-pin dil pack. Three versions provide current transfer ratios of 70% at 0.5mA (100% at 1mA) for the *ISP827-3* to 50% at 1mA for the lower-cost *ISP827-1*. Specially selected units can be provided giving 200% ctr. Transistor output saturation voltage is 0.4V, breaking down at 70V. Power dissipation is 250mW per isolator. Isocom Components Ltd. Tel., 01429 863609; fax, 01429 863581.

Oscillators

Dil saw clocks. C-MAC announces the *CMD 5000 Series* of dual-in-line, surface acoustic-wave, ecl/pecl clock oscillators, designed for use as processor clocks in workstations or as pixel clocks in graphics cards. frequency coverage is 250-800MHz, generated directly, use of the saw technique reducing jitter to a low level. Stability in response to all hazards is ±200ppm. C-MAC Quartz Crystals Ltd. Tel., 01279 626626; fax, 01279 454825.

PASSIVE

Passive components

Chip resistors. Thick-film chip resistors in Welwyn's *LRC* series have values in the 0.02Ω-1Ω range in 1% or 20% tolerance and are meant for current sensing and other power-management uses. These 0.5W and 1W components are an alternative to wire-wound low-value types and, being surface-mounted, show an advantage in small, portable equipment. A thick-film element on a ceramic substrate with solder-plated wrap-round terminations reduce hotspots and provide a lower profile on the board. Welwyn Components Ltd. Tel., 01670 822181; fax, 01670 827434.

Connectors and cabling

Locking connectors. *MATE-N-LOCK* multiple-lead connectors from AMP have up to 15 positions and both

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plug and cap have locking housings to prevent them coming apart when used as panel-mounting connectors or in loose cables. They have a large number of keying combinations and pins and sockets can be mixed in the same housing. Maximum current is 16A, rated voltage 380V, dielectric voltage 5kV or 10kV and insulation resistance 1GΩ. Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095.

Wire-to-board connectors. Molex has 2mm connectors for wire-to-board connection, the *Mill* family. Connectors are available as single and dual row straight and right-angle headers and single/dual row crimp and idt receptacles. The tin-plated brass pins of the through-hole headers are kinked for better board retention during soldering. Polarised, single-row headers come in sizes from 2 to 15 and have extra friction-lock retention, while the dual row headers are available in sizes from 6 to 60, ejection latches being standard for sizes over 28. Application tools are available. Molex Electronics Ltd. Tel., 01420 477070; fax, 01420 478185.

Engraving service. EAO-Highland offers an engraving service for pushbuttons and indicators. Customers specify what they want from a range of characters and symbols, including graphics and engineering symbols, Greek characters and super/subscripts in various sizes, whereupon a computer controls an engraving machine to engrave on the front of the lens or on the rear diffuser for longer life. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Backplane connection. Robinson Nugent's *Metpak 2* backplane connector system is modular and based on a 2 by 2mm pitch. Features include offset, beryllium-copper contacts for reduced insertion force, available with press-fit or solder terminals; early mate/late break contacts arranged by row or programmed; and a six-point, low insertion-force hold-down clip. The system offers double the contact density of *DIN41612* connectors. Available connectors include cable-to-board signal, high power, coax and optical fibre, shielded cable, board-to-board filtered and unfiltered. Robinson Nugent (Europe) Ltd. Tel., 01256 842626; fax, 01256 842673.

Filtered power distribution. Vero has added to its range of mains power distribution panels new filtered and rccb types for 19in racks and Imrak racks, the filtered units having two-stage *RC* filters. The panels use UK, French or CEE 22 style sockets, either switched or unswitched, with five outlets in 2Y models for 19in horizontal mounting or with six, nine or 13-way vertical types. Vero Electronics Ltd. Tel., 01703 266300; fax, 01703 256126.

Shielded cable. *Suprashield* cables from Alpha Wire reduce or eliminate radiation over a wide range of frequencies, particularly relevant in view of the latest European standards. They are made with a triple-laminated tape in aluminium-polyester-aluminium, bonded in one layer and in contact with a stranded, tinned-copper drain wire, equal in size to the conductors. A new catalogue describes these products and 30,000 others. Alpha Wire Ltd. Tel., 01932 772422; fax, 01932 772433.

Molex connectors. Electrospeed is now a distributor for *Molex* connectors in a range containing, for example, Mini and Maxi KK crimp and IDT types, data communications connectors and the Snapper PCMCIA cardframe kit with input and output connectors. These and others are described in Electrospeed's 1996 catalogue. Electrospeed. Tel., 01703 644555; fax, 01703 610282.

Crystals

PCMCIA crystal. ACT announces the *DNC-13* series of low-profile, surface-mounted crystals covering the 8-72MHz range and designed for use in GSM and PCMCIA products, its case measuring 5 by 7mm and 1.3mm high. These crystals age at less than ±3ppm/year and are stable to within ±50ppm/°C. Advanced Crystal Technology. Tel., 01635 528520; fax, 01635 528443.

Filters

2400A filters. Very high-current power line filters from *MPE* to cope with currents in the 800-2400A range at 250/440Vac, 50/60Hz. These are for use on three-phase and neutral supplies and have low leakage with current compensating inductors for low heat dissipation, low running cost and better safety. One of the two types meets MIL-STD 188-125 for emp protection systems and both provide full insertion loss with or without load. A catalogue is available. MPE Ltd. Tel., 01371 875071; fax, 01371 875037.

Snap-on mains filter. Schurter's *Multifit* range of mains input modules now includes a line filter that clips on the back of the power inlet, thereby saving board space and assembly time. Filters are available in 1, 2, 4 and 6A versions, use the board's own ground and provide attenuation comparable to separate filters. Crosstalk is eliminated. Radiatron Components Ltd. Tel., 01784 439393; fax, 01784 477333.

Hardware

Floating toolkit. Should your interests lie in that direction, Jensen can supply you with the *JTK-87WP* field engineers' toolkit, which is said to float, its cycolac resin case being airtight and, of course, watertight. It contains more than 100 tools, half of which are Jensen's own products. Jensen points out that a catalogue of thousands of tools, kits and test gear is available. Jensen Tools. Tel., 0800 833246 (free); fax, 01604 785573.

Cable ties. Thomas and Betts has acquired the US company Catamount of Massachusetts, which also has operations in the UK and Scandinavia. This further reinforces the T&B claim to be a leading supplier of these components. Thomas & Betts. Tel., 01582 677049; fax, 01582 608816.

Emc shielding. Finger strips of beryllium copper alloy 25 for emc shielding are now added to James Walker's range of conductive seals and adhesives. The strips have a spring action and are designed for racks and doors in frequent use in which a wiping action might tear conductive elastomer. Protection against radiation is 120dB E-field in the 1-100MHz range. James Walker & Co, Ltd. Tel., 01483 757575; fax, 01483 755711.

Four-in-one test set. At a cost of about £450, the SJ Electronics *MkII Universal T&M System* offers four commonly used pieces of test gear: a 1.3GHz counter/timer; a 3.5-digit, autoranging dmm with an RS 232 interface and software, giving *R* and *C* measurements as well as ac/dc voltage and current at 400mV/40mA; a swept-frequency, 2MHz function generator giving ttl levels, sine, square and triangular waveforms; and a triple-output power supply providing variable 0-30V at 2A, 15V at 1A and 5V at 2A, all floating. These are full-function instruments intended to form a basic workstation and are finding ready acceptance in universities and manufacturers' test stations. S.J Electronics Ltd. Tel., 01376 562004; fax, 01376 562215.

Rack-mounted pc chassis. IMS has a new industrial PC chassis, the *IPC-620*, which has 20 full-length ISA PCbus slots. It has been designed to take many expansion boards or disk drives and for combining up to four pcs in one chassis, each with its own system controls. It is equipped with a 350W power supply, four fans and a removable air filter on the front panel. It comes as chassis only or as a complete single or multi 486/Pentium system. Integrated Measurement Systems Ltd. Tel., 01703 771143; fax, 01703 704301.

Test and measurement

Gas sampling. From CBISS comes the *MkII Intelligent Sampling System*, which has up to 68 sampling channels for gas and liquid monitoring, required in gas leak detection, pollution monitoring and the like. The MkII is rack-mounted and controlled by an internal single-board computer for eight channels, expandable to 68 by way of five twelve-channel units. It is usable with any pc running the CBISS Windows-based C-DAS software for automatic data acquisition and report generation. CBISS Ltd. Tel., 0151-343 1543; fax, 0151-343 1847.

Portable waveform analyser. A portable version of Nicolet's *2580* waveform analyser has appeared, the *2580-P*, which works alone or in remote-controlled data acquisition. It uses a *P120* processor, a colour tft display and Windows 95-based software, is about the size of a desk-top computer and weighs 33lb. There are up to 24 channels with diff. amplifiers, anti-aliasing filters, digitisers with 12-bit resolution at 10MHz, a store length of 8Msample/channel, multiple and split timebase and various trigger modes. Nicolet Technologies Ltd. Tel., 01908 679903; fax, 01908 677331.

Computing multimeter. Tti has the *1906* multimeter, which is a fully automatic, 5.5-digit true-rms instrument, for control by RS-232 or an optional GPIB interface. Sensitivities are 1μV, 1mΩ and 1nA, input impedance being 10MΩ on direct voltage ranges except on the lower two, which may be selected to 1GΩ. Calibration constants are kept in eeprom and six control set-ups can be held in non-volatile memory. Computing facilities include linear scaling with offset, limits comparison, deviation, min/max storage, log. measurement and data logging. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Insulation testers. *Danbridge JP15* and *JP30A* dc non-destructive insulation testers are 15kV and 30kV testers for quality-control and production, both having variable output with dual meters for current and voltage. The dc technique is said to possess advantages over ac methods in that it allows more quantitative analysis and will show how rapidly a condition is deteriorating. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

GPS clock. Datum has announced the *StarTime* clock that takes its reference from the global positioning satellite system. It produces standard IRIG (Inter Range Instrumentation Group) B time code in modulated and unmodulated form, a 1p/s 50μs pulse and comes with a 9600baud RS-232C interface. With GPS restrictions currently in use, timing accuracy is better than 2μs and a velocity of 400m/s is allowable. Time to first fix is under 90s after a short power-down and 5-15minute from cold. Sematron UK Ltd. Tel., 01256 812222; fax, 01256 812666.

Multimeter/component tester. As well as the normal functions of hand-held digital multimeter, the *Wavetek DM16XL* measures frequency to 15MHz and capacitance to 20μF, tests diodes, transistors, cmos and ttl logic circuits and continuity. There is provision to freeze the display, which has 0.7in numerals, for later reference. Wavetek Ltd. Tel., 01603 404824; fax, 01603 483670.

Interfaces

RS-232 interfaces. Harris's range of ttl/cmos/RS-232 interface ics now includes eight standard types needing only low-cost 0.1μF external charge-pump capacitors. The *HIN200/1/2/4/6/7/8/11* parts operate from a single 5V supply or dual +5/+7.5 to 13.2V supplies, offering a variety of arrangements of receivers and drivers. Harris Semiconductor UK. Tel., 01276 686886; fax, 01276 682323.

Literature

Rf/microwave semiconductors. *M/A-COM* has a new catalogue featuring discrete, monolithic and multi-function technology, including chip *Cs*, pin diodes, variable-capacitance diodes, Schottky and Gunn diodes, transistors and power modules. There are also amplifiers, power splitters/combiners, mixers and semiconductor materials. BFI IBEXSA Electronics Ltd. Tel., 01622 882467; fax, 01622 882469.

Capacitors. 225 pages of NEC's capacitor data book give product information and applications data on the company's small solid tantalums and 'Supercapacitors', which come in values up to 3.3F at 5.5V to provide reserve power of better reliability than batteries. As an example, a 3.3F capacitor will hold up a 256bit ram for 70 hours. NEC Electronics (UK) Ltd. Tel., 01908 691133; fax, 01908 670290.

Displacement sensors. Non-contacting sensors using the eddy current method of operation are described in a brochure from *Monitran*. Units described work in the 0-2.5mm up to 0-8.5mm range and come in threaded or flange mountings. Monitran Ltd. Tel., 01494 816569; fax, 01494 812256.

Materials

Cleaning solvent. Electrolube has *Electronic Cleaning Solvent Plus* (ECSP), which is a fast-drying solvent for cleaning contacts, tape heads and pcbs and other devices of a similar nature. It replaces CFC solvents such as 113 and leaves a clean dry surface. It is harmless to most plastics, rubber, elastomers and surface coating and is applied by immersion or spray from 200ml CO_2 propelled aerosols. Electrolube Ltd. Tel., 01734 403014/031; fax, 01734 403084.

Electron gun supply. Applied Kilovolts has an isolated filament and grid power supply to drive electron guns for surface analysis. These guns are usually at –30kV, the anode being grounded, so that the gun supply must be well isolated; this device supplies 15W to the filament and 1kV to the grid of the gun and the controls and filament/grid monitor signals are at ground voltage. Ripple in the *K1/62* is 0.1Vpk-pk, temperature coefficient 0.1°C and output stability with load and input variation 1%. Applied Kilovolts Ltd. Tel., 01273 439440; fax, 01273 439449.

Production equipment

Liquid dispensing. For the precise dispensation of glop such as cyanoacrylate, lubricant, solvent and other low-viscosity fluids in discrete amounts, Fisnar offers the *PPD-120* peristaltic pump dispenser (the squeezy kind), which takes the said glop straight from its own container, the amount being controlled by a built-in timer. If you take too much, the device sucks it back and the risk of operator contamination is greatly reduced – you won't get the stuff all over your trousers. Intertronics Ltd. Tel., 01865 842842; fax, 01865 842172.

Power supplies

Plug-top adaptors. XP's *TSA* and *TRA* plug-top mains adaptors, the *TRA* linear regulator series producing regulated voltage of 5, 6, 9, 12, 15 and 24V at up to 8.5W at 24V, plugging directly into a mains outlet. *TSA* units are switched regulators and produce up to 15W at 77% efficiency, regulation being 1% and stabilisation 0.1%. Both types can be fitted with UK or European plugs and a 2.1mm battery-eliminator jack is standard. XP plc. Tel., 01734 845515; fax, 01734 843423.

Sil step-down regulators. International Power Devices produces the *SIP* series of step-down buck regulators in single-in-line packs. These non-isolated dc-to-dc converters produce a stable voltage in the 1.2-3.5V range at 6A from existing board inputs. They are optimised for 5V input and use a synchronous rectifier buck regulator for high efficiency. Overload protection is provided. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 01273 570215.

Converters for notebooks. Linear's new *LTC1438/9* voltage converters exhibit constant frequency and high efficiency over a 100:1 load range; their characteristics are applicable to use in notebook computers. Output is more than 5A at over 90% efficiency, an auxiliary low-dropout regulator driver supplying 12V at 200mA with an external p-n-p transistor. Control functions are included and supplies with up to four outputs may be configured with only a few inductors. Linear Technology (UK) Ltd. Tel., 01276 677676; fax, 01276 64851.

Supply monitors. Zetex offers a range of supply voltage monitors for low-power, lower voltage supplies, now including five new devices for 3V and 3.3V working. The *ZM33* series provides good noise immunity with a hysteresis level of 60mV, The *ZM33164-3* is for 3V systems and has a 2.68V threshold, a 4.3V threshold making the *ZM33164* the choice for 5V use. *ZSM300/330/500* monitors have hysteresis of 20mV and have thresholds of 2.63/3.2/4.3V. Open-collector outputs sink 10mA and the devices take very little current on standby. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5467.

NiMH cells. Varta has extended its range of nickel metal hydride cells by the addition of sintered dry roll types and *Combat*, which is a prismatic cell. *Combat* cells are intended for telecomms use, where medium drain is to be expected, and are made using mass plate button cells instead of the conventional spiral-wound electrode. The sintered dry roll types provide 50% more capacity than NiCd cells and are resistant to abuse. Varta Ltd. Tel., 01460 73366; fax, 01460 72320.

Radio communications products

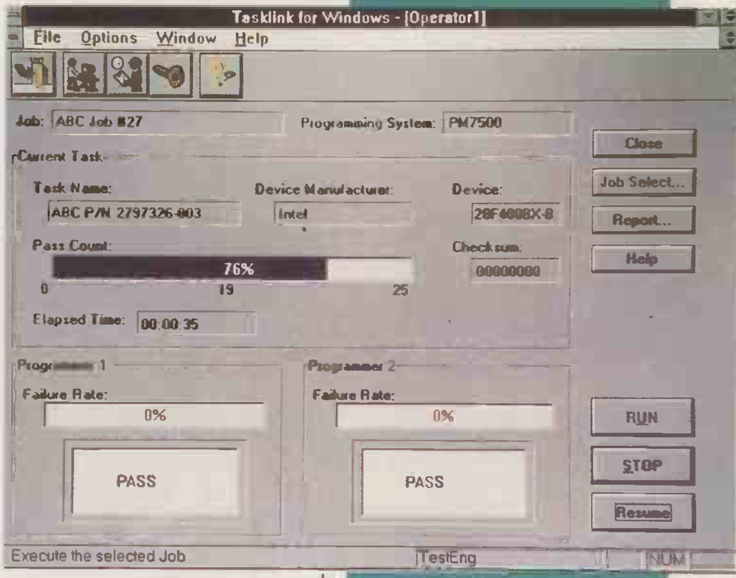
Low-power receiver. A low-power, am superhet receiver for use in radio-based alarms and controls, the *Neohm SHR* is available for either 418MHz or the new European 433.92MHz (see the entry under 'Car alarm filter'). It is a board-mounted, single-in-line module measuring 38 by 17mm and is based on a saw resonator; sensitivity is 1.4μV for 6dB s:n ratio. With a suitable antenna and the appropriate transmitter, this level of performance gives around 50m range. The device needs a 5V supply, but will also take dual supplies to interface with decoder logic at higher voltages; current drain is 6mA and under-bonnet temperatures are supported. LPRS offers much applications data and a range of transmitters at both frequencies. Low Power Radio Solutions Ltd. Tel., 01993 709418; fax, 01993 708575.

LabVIEW+LabWindows/CVI. National's *LabVIEW* graphical programming software and *LabWindows/CVI* visual development package for virtual instrumentation are up-graded to v.4.0 in such a way that they now work together to combine graphical and C programming in one system, both of the packages having new tools, one of which, *CodeLink*, allows C code from *LabWindows/CVI* to be used in *LabVIEW*. Both packages now have 32-bit versions for Windows 95 and NT. National Instruments UK. Tel., 01635 572400; fax, 01635 523154.

Broadband antenna. Intended for pre-compliance emc test, the Seaward antenna complements the company's range of test equipment, but is also fully usable on its own. Frequency range is 30-450MHz and, when uses with Seaward's spectrum analyser, makes use of the antenna factor compensation and ambient noise reduction of the instrument. The antenna is suitable for screened-room and open-site use, the package including 10m of lead and a tripod. Seaward Electronic Ltd. Tel., 0191-586 3511; fax, 0191-586 0227.

GPS tracking. Racal Survey has introduced the *Tracs 2000*, a GPS-based system to monitor the position of vehicles and ships, to monitor and report fuel reserves of vessels, cargo deployment, personnel movement for safety monitoring, emergencies and much other data. The time-slot software used in the system allows the central control to allocate reporting intervals for each mobile to automatically transmit, the system monitoring up to 20 units per second, with no operator input. Position reports are accurate to within one metre, since differential GPS is used. Racal Instruments Ltd. Tel., 01734 669969; fax, 01734 262121.

Digital frequency measurement. *Wide Band Systems*'s digital frequency discriminators for use in military instantaneous frequency measurement receivers are now available from Anglia Microwaves. The dfds provide extremely rapid measurement on pulsed or cw rf input over octave or more bands. Dfds use an array of temperature-compensated microwave correlators after an rf limiter, output being taken to a processor to convert this information to a digital rf output. Anglia Microwaves Ltd. Tel., 01277 630000; fax, 01277 631111.



Protection devices

Thermal fuses. Designed for rapid installation and removal, *Orient* temperature protection and thermal fuses are mounted by flange and connected by 6.3mm spade terminals. The units contain an integral current fuse and, since the thermal fuse carries no current, temperature derating is unnecessary. Temperature range is 78-190°C. Microtherm Ltd. Tel., 01483 450100; fax, 01483 451816.

Car alarm filter. Since the adoption of the new pan-European frequency of 433.92MHz by car alarm makers, baffled drivers have found themselves either locked out or with immobilised ignition. This comes about because the new frequency is in the amateur 70cm band. To overcome the problem which, one would have thought, was fairly predictable, Siemens has produced the *B3530* front-end saw filter, which is a *TO39* device needing only a resistor and capacitor, both surface-mounted, for matching. It has a 600kHz bandwidth and 2.3dB insertion loss. No tuning is needed. Quantelec Ltd. Tel., 01993 776488; fax, 01993 705415.

Switches and relays

Photorelay. The *AQZ PhotoMOS* relay by Matsushita switches 4A at 60Vdc or 0.5A at 400Vac/dc in one normally cut-off channel. It has a 'negligible' output offset voltage compared to the normal solid-state device's 0.7V, a 1μV thermal emf, 1.5kVac i/o isolation and total silence in operation. Matsushita Automation Controls Ltd. Tel., 01908 231555; fax, 01908 231599.

Automotive relays. Fujitsu's twin relays, mounted in one enclosure, switch up to 30A at 12Vdc and are meant for use with all the motorised functions in a car, the twin design being used for up/down and forward/backward motion to save space and assembly time. The relays are in hermetically sealed enclosures and cope with the –40°C to 85°C range of temperatures. Inelco Ltd. Tel., 01734 810799; fax, 01734 810844.

Production control. Data I/O has produced *TaskLink for Windows*, a process-control software package for device programmers in production and automated handling systems, to be initially available on the company's *ProMaster 2500, 3000, 7000* and *7500* handling systems. It allows the automation and setup of a manufacturing session, including blank device selection, quantity of parts programming data, finished part number, device serialisation and labelling information. It tracks and reports on yield and operational statistics. Data I/O Ltd. Tel., 01734 440011; fax, 01734 448700.

Transducers and sensors

Absolute encoders. *TWK*'s range shaft encoders of the absolute type comprises precision, single and multi turn models and pc-programmed multi turn types with options of synchronous serial, asynchronous serial or Interbus-S interfaces, all with up to 32-bit resolution. Output is either fixed on one type of code or can be programmable to give Gray, binary-coded decimal, natural binary, Gray tree and binary tree code. All can be supplied with environmental protection to IP67. Compact Instruments Ltd. Tel., 01204 532544/5; fax, 01204 522285.

Optical sensors. Omron's *Photomicrosensor* range of optical switches now contains convergent-beam reflective types for improved accuracy, the use of Fresnel lenses further enhancing performance. In the *EE-SY190/1* devices, the axes of emitter and detector are inclined towards each other, so increasing the level of light and also accurately defining the range of an object, even a transparent reflective one, with no effect from the background. The sensors measure 18mm long, 9mm in height and are 6mm wide. Omron Electronics Ltd. Tel., 0181-450 4646; fax, 0181-450 8087.

Pressure sensors. *Novasensor* NPC410,1210 and 1220 series ceramic substrate pressure sensors by Lucas are lower-priced, 'drop-in' replacements for competing products. According to Lucas, they produce a better performance. The three types are uncompensated, the *410*; compensated with gain-set resistor, the *1210*, and compensated with current-set resistor. All three types handle the 5 to 100lb/in^2 full-scale range in gauge and differential pressure and 15 to 100lb/in^2 full scale in absolute pressure. Lucas Control Systems Products. Tel., 01535 661144; fax, 01535 661174.

IR photo-interrupter. Isocom's *H22* photo-interrupter is a direct equivalent of the standard opaque infrared type. It is a single-channel switch consisting of a gallium-arsenide infrared diode, an n-p-n transistor and a 3mm gap between them, all in a polycarbonate encapsulation. Six versions give 0.6 to 5.5mA output current for 5 to 30mA drive, with collector voltages from 30V to 55V. Non-standard mounting arrangements are available. Isocom Components Ltd. Tel., 01429 863609; fax, 01429 863581.

COMPUTER

Development and evaluation

Z380 In-circuit emulator. Signum Systems's *USP380* emulator for the Zilog *Z380* processor offers a point/click interface. A source-level C debugger enables source-code line stepping, display of local variables and support for all types of variable including nested types and arrays. Uploading and downloading of 64K takes 14s over a serial port at 115kbaud. There is 1Mbyte of overlay program ram for positioning anywhere in the 4Gbyte address space and a 32K trace buffer provides 80-bit width, incorporating filter controls and a real-time, 100ns stamp. An application can be debugged without stopping the processor. Noral Micrologics Ltd. Tel., 01254 682092; fax, 01254 680847.

Programming hardware

Gang/set programmer for 3.3V and 5V. *Speedmaster GLV-32* from ICE Technology provides high-speed programming for eproms, eeproms and up to 8Mbit flash memory at 3.3V and 5V, in a gang of eight or in sets. Programming is carried out at voltages from 3V to 25V in 0.1V increments to protect devices and maintain yields. The programmer plugs into a parallel pc port and also offers two-button stand-alone working with a master eprom, which only needs reconnection for new chips or different algorithms. Eight devices can be programmed at once in a matter of seconds. Ice Technology Ltd. Tel., 01226 767404; fax, 01226 370434.

Software

VisSim v.2. Adept Scientific's *VisSim* Windows-based package for control system simulation and development is now in version 2, having 32-bit capability to allow processing at twice the speed of earlier versions. VisSim is very simple to use, needing no programming, by connecting graphical blocks together to model a system; inputs are entered and the response charted. New features include an improved user interface, including 'virtual instrumentation'; better animation to give improved dynamic model simulation; new design tools including fir and iir filter design; and more facilities for producing reports. Adept Scientific Micro Systems Ltd. Tel., 01462 480055; fax, 01462 480213.

Formulas. *Gieck's Engineering Formulas*, a well known reference book, is now available on cd or floppy, courtesy of MathSoft and McGraw-Hill, the publishers. The disks run under Windows and, although the software uses *MathCad* techniques, *MathCad* itself is not needed and the pc need not be especially exotic. More than 300 standard formulae may be solved by button-pushing rather than head-scratching and users' own variables can be used. Adept Scientific Micro Systems Ltd. Tel., 01462 480055; fax, 01462 480213.

Visual Designer for education. A low-cost version of *Visual Designer*, the Windows-based package for data acquisition, test, measurement and control in chemical, environmental, electrical, medical and mechanical applications, is now available and is mainly intended for universities. It adopts a flow-chart approach with a new set of function blocks, allowing applications development by drawing block diagrams – no code generation being needed. Display functions allow plot modification and the representation of instrument control panels. Intelligent Instrumentation. Tel., 01923 896989; fax, 01923 896671.

LETTERS

Letters to "Electronics World"
Quadrant House, The Quadrant,
Sutton, Surrey, SM2 5AS

Excuse me

It was a genuine thrill to see my book, The Tube Preamp Cookbook, listed as supplementary literature in Morgan Jones' article 'Designing valve preamps' in the March 1996 issue of *EW* – its first such mention.

I like what Mr. Jones writes but I find it unusual that in discussing RIAA equalisation in preamps he ignores a key point I made in my book.

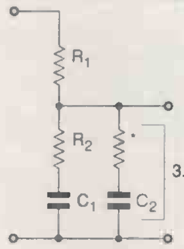
Morgan states (p194) "...high frequency attenuation must continue indefinitely," and while I agree that this is the accepted norm, I consider it to be non optimum. The inflated ego of preamp designers notwithstanding, the RIAA replay curve should not be considered an absolute in itself. Rather it should mirror the actual equalisation curve of the recording process.

If the playback is to decrease indefinitely with frequency, the record high frequency would need to have been increased indefinitely. This is of course impossible. No record was ever cut this way.

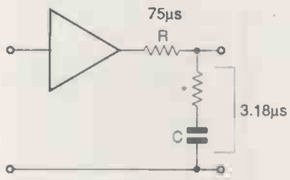
I suggest in my book that adding a further step in the curve at 3.18µs, to flatten the falling playback response gets much closer to the way long-playing records were – and still are – actually cut. This figure was chosen after consulting a number of record cutting equipment manufacturers to see what they actually did, and some 20 years of client and personal experience.

To add this 3.18µs step, Morgan's schematics need only the addition of one resistor, but what an unexpected sonic improvement this makes in an otherwise optimum system.

Allen Wright
Munich
Germany



Adding a resistor to Morgan Jones' RIAA circuits, March 1996 issue, as shown in these three diagrams, provides an "unexpected sonic improvement". Above is Fig. 8 from Morgan's article.



Morgan's Fig. 9 with additional resistor.

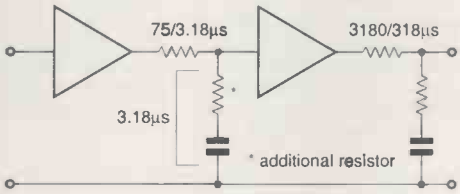


Fig. 10 with additional resistor.

10mV diode proof?

I would like to issue a challenge to Douglas Self.

In his 'Creative Fiction' letter in the same issue, Douglas scoffs at Ben Duncan's proposition of '10mV diodes' in multistrand copper cable and gives measurements to prove his point.

I applaud doing tests rather than relying on simulations but let's take this a step further. I would like to offer a different test – one that opened my eyes to the sonic properties of cables in audio, and which has led to my second book – which is certain to upset many people – The SuperCables Cookbook.

Douglas:
1. Please connect one of your blameless power amplifiers to your best speakers with some good quality copper braid naturally using separate lengths per terminal. Such braid is the shield of readily available *RG59* for example.
2. Play some music at normal levels to reassure yourself that this inexpensive cable option sounds noticeably better than lamp cord.
3. Now reduce the volume to a very low level, ie milliwatts, as if you were listening at 3am in a sleeping house, and take a minute or two to become accustomed to this new level.
4. Now, without changing the volume level, replace the braids with 0.25mm diameter single strand wire-wrap wire and listen again. No fancy twisting or braiding – just hook it up, one thin solid core strand per terminal.

Your challenge is to come up with a better concept than Mr. Duncan's 'diodes' to explain the resultant increase in clarity and reduction of distortion coming from this unusual wire substitution. And please do this listening test before offering measurement proofs that no change occurred or is possible.

Now while I'm not advocating using 0.25mm wire for speaker connections – under most conditions anyway – something is certainly going on here and to me, Ben's '10mV diodes' are as good a description as any of the sonic effects I clearly hear.

Of course you'll need an amplifier with absolutely no crosssover artifacts to hear these cable effects and if 'crystal' amps do not allow this, perhaps Morgan could help out with one of his rather nice class A valve amps.

AW

What day is it?

In Table 1 of the 'Building blocks of time' *EW+WW* March 1996 the day-of-week code used by the MSF slow code is presented as '1-7, bcd'. In fact the code used is 0 for Sunday to 6 for Saturday to avoids the possibility of imitating the framing code carried by bits 52 to 59. This framing code comprises six consecutive long pulses (ones) between two short pulses (zeros). The framing sequence given in the article has the wrong polarity.

During a 61-second minute containing a leap second the extra second is 'inserted' before second 17 so all the information for the next minute will appear to be one second late.

The article makes brief mention of the 'fast code', this is more difficult to receive and decode and it may soon be withdrawn to give a full 500ms minute marker. So the fast code should not be used in new designs. The slow code travels well, I have MSF domestic clocks operating reliably in a house in Finland, 1888km (only 378 wavelengths) from Rugby, with the hour hand advanced by two.

As well as the time and date information provided by on/off keying, the 60kHz MSF carrier itself provides a very accurate frequency reference, (±2 parts in 10^{12}), traceable to the national standard at the NPL.

John Chambers
Head of Time and Frequency Services
National Physical Laboratory

Ancient valve myth

I was fascinated to find an ancient valve myth propagated in the hybrid jungle of Morgan Jones' valve preamplifier.

In 'Practicalities and Performance' he talks about the dangers of stripping valve cathodes if HT is applied before they are hot. Not so. This problem was confined to power valves with thoriated tungsten bright emitter filaments and gas filled thyratrons and rectifiers. Cathode degradation due to instant HT was never a problem with the smaller valves used in domestic equipment.

The lie is given to the myth by nearly 40 years of silicon diode rectifier usage. These put HT on to the valve anodes long before the cathodes are even warm. Incidentally, there should be surge limiting resistors between the rectifier bridges and reservoir capacitors in Figs 6 and 7, or does the winding resistance serve in a design where everything else has been carefully arranged?

In the late fifties, plug-in diode rectifier modules on valve bases were offered by valve makers as replacements for valve rectifiers in domestic and industrial equipment, and don't think this was a crafty ploy to sell replacement valves! By the seventies, wired in diodes had supplanted the thermionic rectifier, without apparent ill effects on valve life.

I can see little point complicating things by keeping valve heaters on stand-by with all that active stabilisation built into the amplifier – its certainly not needed to protect the cathodes.

Anthony Hopwood
Worcestershire

Panic attack

Soft-core pornographers are rightly limited as to what they can put on their covers, in order not to corrupt tender minds via the newsagent's shelves. What a pity then that *EW* is free to display frightening disinformation on its covers. I refer to the latest dose of dubious research from Anthony Hopwood. Somebody with political power, but no scientific insight, might well see your cover, take Hopwood seriously, and start a panic.

I shall never take Mr Hopwood seriously. This is not because of any vested interest, or a natural disdain for amateurs, but because I always recall Mr Hopwood's *New Scientist* article (20/27th December 1979). In this, he claimed to be able to 'dowse' whether power was flowing through an elevated wire. He also noted that the results were affected by sunlight (well I never – the gleam of a startling idea!).

When a group of skeptics descended upon him, and made him carry the tests out properly, the strong effect which he had claimed completely disappeared (*New Scientist*, 22nd October 1981).

Could you please apply a modicum of skepticism yourself before publishing any more of Hopwood's 'evidence'. In fact, I would advise you to double-check the work of anyone who is known to believe in dowsing, ball lightning, 'free energy', and/or anti-gravity.

Dr. David J. Fisher
Cardiff

Mile high mine?

Anthony Hopwood, in his article on 'Power lines particles and cancer', mentions the city of Denver twice. Neither he nor anyone else, as far as I can recollect, has referred to the fact that Colorado is the centre of USA's uranium mining, and other metals are extracted there. Denver originated as a mining area and probably is perched on rocks with a high radon content.

Using figures given in the current edition of Whittaker's Almanac, leukemia, as a percentage of the total neoplasms, is 2.5% in England and Wales, 1.83% in Scotland and 2.3% for N. Ireland. As percentages of the total populations in the three regions the figures are 0.007%, 0.005% and 0.005%.

Apart from the almost equality of the results, are the low population percentages sufficient evidence on which to form clear opinions?

S.F. Brown
Shropshire

And Anthony replies...

Dr. Fisher's intemperate letter is a classic of the 'don't frighten the horses' genre.

He is obviously far more interested in something I did nearly 20 years ago than looking at the science behind my observations on power line effects which have already been replicated by others.

On the derided electric field experiments, I would remind him that the tests he mentioned were carried out indoors, whereas, I specified outdoor conditions to match the original series of observations taken over many months. Decades later no-one has had the scientific curiosity to try and replicate a simple experiment to show humans have a weak electric field sensitivity, and thus provide a physics based explanation of water dowsing.

These experiments could be done by a school sixth form and need no megabuck apparatus. Come on Doctor, try some experimental science for a change instead of making false accusations of my involvement in topics that I have never investigated or even written about. You must be confusing me with someone else!

Now to power lines. My work is based on text book physics – no more – no less. Even now, there is far more international epidemiological evidence for a link between electric power systems and disease. Far, far more than is available to prove CJD comes from BSE infected beef.

The implication of ionising radiation intensification as the link between electrodynamic fields and cell damage is simple and needs no physics rewrite. What is needed is properly funded research, not medieval blinkerdom.

Anthony Hopwood
Worcestershire

Best rf article '95

Entries for this challenge are still being evaluated. Apologies – we *will* make an announcement about the winner in next month's issue.

HELP wanted

Have you any queries?

If you have any electronics-related questions that you have been unable to find an answer to, why not see if other readers can answer them? Simply write to me, the editor, at the address on page 267, fax 0181 652 8956, or e-mail martin.eccles@rbp.co.uk.

Last month, Terence Heatley asked...

Could one of your readers explain to me a phenomenon connected with the distribution of lines of magnetic flux around a single length of wire carrying a dc current of 1A? With this wire passing through a card at right angles to the wire; if soft iron filings are sprinkled around the wire magnetic lines may be observed which form concentric circles around the wire with spaces between them.

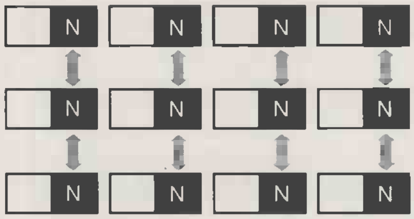
My question is this: has some form of standing wave been set up in the spacing between 'crests'?

Two similar explanations submitted by readers are...

The question of lines of magnetism struck me a few years ago, and I think that the explanation is thus.

If you take a handful of bar magnets, like poles repel and unlike poles attract. You can join the magnets into a chain or line. I think that you will find that the lines or chains will repel each other apart.

For the problem of lines around the wire, for soft iron filings, they form bar magnets by the flux flowing through them, with same pole at the clockwise end.



Douglas Rice
Ipswich

Terence Heatley (April 1996) proposes a vibrating lycopodium powder analogy to explain the rings seen in iron filings scattered on a card through which a current carrying wire passes perpendicularly, believing that standing waves are involved. If the current is steady, no waves are involved and the rings seen in the iron filings are not due to electrons travelling in bunches along the wire making noise, propagating as rings.

When an iron particle falls through the field, it becomes magnetised parallel to the field, and like a little bar magnet, influences the way in which magnetised particles nearby fall onto the card. The energy of the system is reduced by the particles forming chains as they settle, the lycopodium powder analogy is thus false. This emphasises the value of the course 'Engineering Science' which I teach to young electronic engineers here at the University of Hertfordshire. When they find out more about natural phenomena, this helps to balance studying all the complicated man-made devices like silicon chips with finding out how nature itself works.

Guy S M Moore
Division of Physical Sciences
University of Hertfordshire.
Hatfield

Ivan Eamus replied with a similar explanation but his reply was too late for publication.

SURVEILLANCE TELESCOPE Superb Russian zoom telescope adjustable from 15x to 60x! complete with metal tripod (imposible to use without this on the higher settings) 66mm lense, leather carrying case £149 ref BAR69

RADIATION DETECTOR SYSTEM Designed to be wall mounted and connected into a PC, ideal for remote monitoring, whole building coverage etc. Complete with detector, cable and software. £19.95 ref BAR75.

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CCTV CAMERA MODULES 46X70X29mm, 30 grams, 12v 100mA, auto electronic shutter, 3.6mm F2 lens, CCIR, 512x492 pixels, video output is 1v p-p (75 ohm). Works directly into a scart or video input on a tv or video. IR sensitive. £79.95 ref EF137.

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REMOTE CONTROLTANDATA TD1400 MODEM/ VIEWDATA Complete system comprising 1200/75 modem, auto dialler, Infra red remote keyboard, (could be adapted for PC use?) psu, UHF and RGB output, phone lead, RS232 output, composite output. Absolute bargain for parts alone! £9.95 ref BAR33.

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28vdc 100 watt 50-200mhz 50 ohm in/out FM/AM £75 ref RF4
A heat sink will be required, ring for price and availablity.
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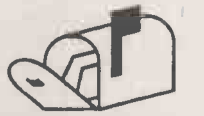

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RACAL MODEM BONANZA! 1 Racal MPS1223 1200/75 modem, telephone lead, mains lead, manual and comms software, the cheapest way onto the net! all this for just £13 ref DEC13.

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PLUG IN ACORN PSU 19v AC 14w , £2.99 REF MAG3P10

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RATTLE BACKS Interesting things these, small piece of solid perspex like material that it you try to spin it on the desk it only spins one way! in fact if you spin it the 'wrong' way it stops of its own accord and go's back the other way! £1.99 ref GI/J01.

GYROSCOPES Remember these? well we have found a company that still manufactures these popular scientific toys, perfect gift or for educational use etc. £6 ref EP70

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RECHARGE ORDINARY BATTERIES UP TO 10 TIMES! With the Battery Wizard! Uses the latest pulse wave charge system to charge all popular brands of ordinary batteries AAA, AA, C, D, four at a time! Led system shows when batteries are charged, automatically rejects unsuitable cells, complete with mains adaptor. BS approved. Price is £21.95 ref EP31.

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PHOTOGRAPHIC RADAR TRAPS CAN COST YOU YOUR LICENCE! The new multiband 2000 radar detector can prevent even the most responsible of drivers from losing their licence! Adjustable audible alarm with 8 flashing leds gives instant warning of radar zones. Detects X, K, and Ka bands, 3 mile range, 'over the hill' 'around bends' and 'rear trap facilities. micro size just 4.25"x2.5"x.75". Can pay for itself in just one day! £79.95 ref EP3.

SANYO NICAD PACKS 120mmx14mm 4.8v 270 maH suitable for cordless phones etc. Pack of 2 just £5 ref EP78.

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CIRCLE NO. 139 ON REPLY CARD

Circuit reflections

Gems and oddities from early reference works – including the Admiralty Handbook – presented by Ian Hickman.

With the advance of electrotechnology, Degree syllabuses have become so specialised that new graduates enter industry with a good grounding in their own sphere, but an almost complete lack of knowledge of other areas of electronics. Many a new engineer has wasted time trying to perfect a circuit that an old hand could have told him – at a glance – could not possibly work.

A great deal of material basic to the practice of electronics is becoming more difficult to find. Often, the best sources of material are second-hand bookshops, where invaluable copies of 'Terman', 'Langford Smith' and other classic texts on electronics can be picked up for a few pounds.

Admiralty Handbook surprises

Old technical books often surprise you by revealing that many aspects of science were well known much earlier than one realised. The 1925 Admiralty Handbook is no exception.

Figure 1 reproduces Table VI from page 6 of the Handbook. From this it appears that millimetric waves with frequencies up to 200GHz had already been produced at that date, though not exploited till long after. Figure 3 on page 15 of the Handbook reveals that among atomic forces, the strong short-range force was already part of the corpus of knowledge in 1925.

Other items from old technical books, while not surprising, are nonetheless of interest, and occasionally of practical use. For example, the question of units is well handled by the Admiralty handbook, which gives the relationships between practical units in electrical science and the absolute (c.g.s.) units, as well as others. Thus on page 75, one learns that 1 farad equals 9 times 10^{11} absolute units of capacitance, or cms.

I had come across capacitors with their values marked in centimetres decades ago, while repairing pre-war valve radios, and been told that taking the value as meaning picofarads would not be too far adrift. In fact, as can be seen, 1cm equals 1.11pF, this being the capacitance of an isolated metallic sphere of 1cm diameter or radius, I forget which, situated at infinity.

At that time, the Navy, used Jars as its measure of capacitance, this presumably being the nominal capacitance of a Leyden jar. The Handbook defines 1 Jar as equal to 10^3 cms, making 1 Jar equal 1.11nF. Fascinating stuff, if of limited use nowadays, except to collectors and restorers of vintage radios.

Of more use is the very handy chart facing page 65, which is reproduced here as **Fig. 2**. This provides the factor F for use in the formula for self inductance of an air-cored coil, $L = r \times N^2 \times F$, where,

r = the mean radius of the coil in inches
l = the winding length in inches
d = the depth of the winding in inches, or the thickness of wire used in a single layer coil
N = the number of turns in the coil
F = the 'form factor' of the coil, and is found from the curve in Figure 40 of the Handbook,

Titles to look out for

Of particular interest are the various handbooks issued by the Air Ministry, the Admiralty and various branches of the Army. For many aspects of telephony, the Handbook of Line Communication, Vol.1, prepared for The Royal Signals and published by Her Majesty's Stationery Office in 1947, is of considerable interest.

Very common in second-hand bookshops is the Admiralty Handbook of Wireless Telegraphy, published in two volumes by His Majesty's Stationery Office in 1938. Despite its title, Volume 2 also covers radio telephony. Appendix D of Volume 1 is titled 'W/T Text Books, Works of Reference and Journals', and the first entry under Journal is 'Wireless World.' (Iliffe & Sons. Weekly. -/4d.).

The two volume 1938 edition superseded the earlier 1931 edition, of the same title, which I have not seen. However, I found a single volume edition of the 'Admiralty Handbook of Wireless Telegraphy 1925' in a subterranean second-hand bookshop in a small town in Yorkshire, although my copy is from the 1928 reprinting. This was prepared by Captain W.G.H. Miles, R.M., and cost just 5 shillings – getting on for £20 in today's money – from HMSO.

The title page states that it supersedes 'The Admiralty Handbook of Wireless Telegraphy, 1920', a copy of which I would dearly like. Like the later editions, the 1925 edition also covers radio telephony, but not having seen the 1920 edition, I can't say whether that did. Like many publications of the period, the 1925 edition includes advertisements from various manufacturers active in the field at that time, such as Dubilier, Claude Lyons, Exide, The British Ebonite Co., Ltd. and, of course, Marconi's Wireless Telegraph Co., Ltd.

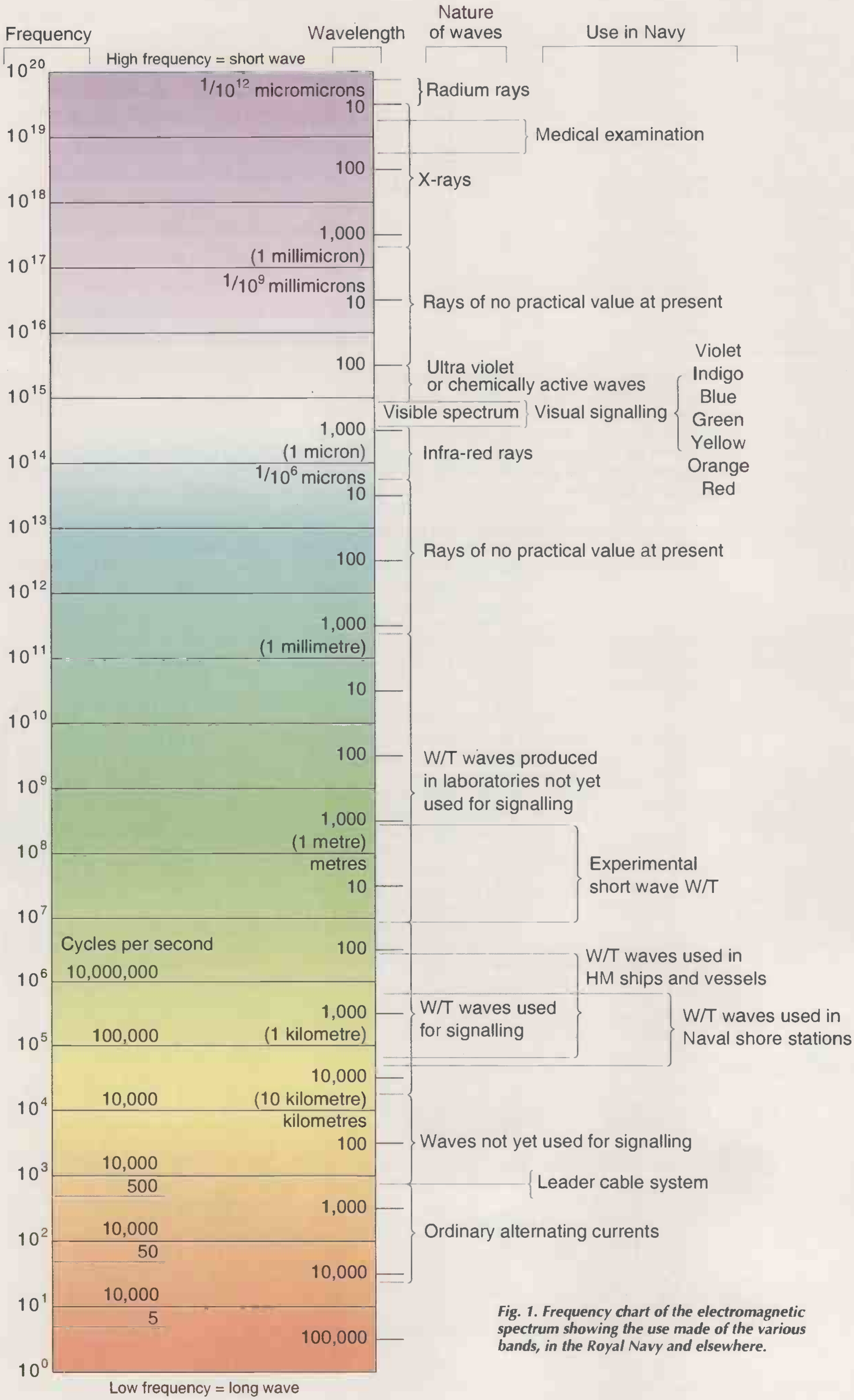


Fig. 1. Frequency chart of the electromagnetic spectrum showing the use made of the various bands, in the Royal Navy and elsewhere.

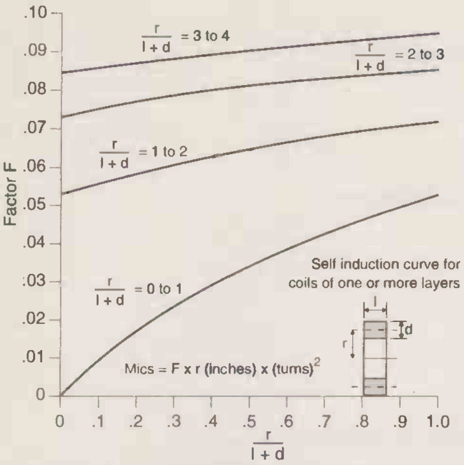


Fig. 2. Chart providing the factor F for use in calculating coil inductance.

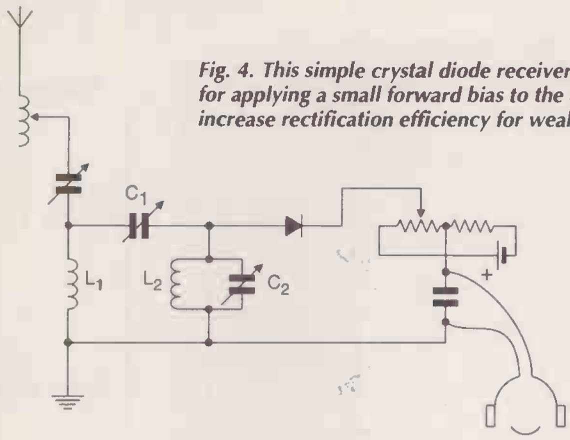


Fig. 4. This simple crystal diode receiver had means for applying a small forward bias to the detector, to increase rectification efficiency for weak signals.

after the ratio $r/(l + d)$ has been determined. Inductance L is in 'mics', mics being the then colloquial abbreviation for microhenries, the equivalent of modern day 'puffs' for pF.

From the Navy's point of view, working in mics and jars gave a convenient formula for the resonant frequency, ω – where ω is $2\pi f$ – of a tuned circuit:

$$\omega = \frac{3 \times 10^7}{\sqrt{(LC)}}$$

Circuits galore

The Handbook describes and analyses many circuits, including parallel resonant tuned circuits, with the aid of vector diagrams. Although rather on the small side, these greatly help the reader to visualise and understand circuit action.

Turning to active circuits, the Handbook contains many. Those associated with radio telephony do not look too unfamiliar to the mature engineer – or even to the younger engineer keen enough to have found out a little about valves, which are currently making a comeback in audio amplifiers. But one which looks strange to modern eyes is shown in Fig. 3. This depicts a Poulsen Arc Transmitter.

Current is supplied by a motor-generator set and the arc burns in a powerful transverse magnetic field produced by pole-pieces, fitted with coils energised by the arc current itself. This causes the arc to bow, increasing its length, the more so the more the current. In conjunction with a ballast resistor, this stabilises the mean or dc value of the arc current, for like any discharge phenomenon the arc exhibits a negative resistance.

But at ac, the negative resistance buffered from the damping of the ballast resistor by the effect of the polepiece chokes, is capable of supplying the losses in a resonant tuned circuit, maintaining continuous-wave oscillation which is keyed and applied to an aerial circuit.

Compared to a simple spark transmitter, the Poulsen Arc transmitter can radiate much more power from a ship's antenna, which is necessarily restricted in size. This is because, with a spark transmitter producing heavily damped oscillatory wavetrains, the maximum voltage that the antenna can support without arcing over is only present for a small proportion of the time.

By comparison, the Poulsen Arc transmitter can radiate continuously the maximum power that the antenna can handle.

Crystal-diode receiver

Turning to more familiar-looking circuitry, Fig. 4 shows a crystal-diode receiver,

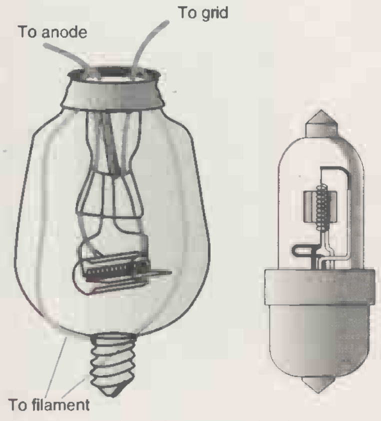


Fig. 5. Triode valves of 1925 vintage used a directly heated filament rather than an indirectly heated cathode.

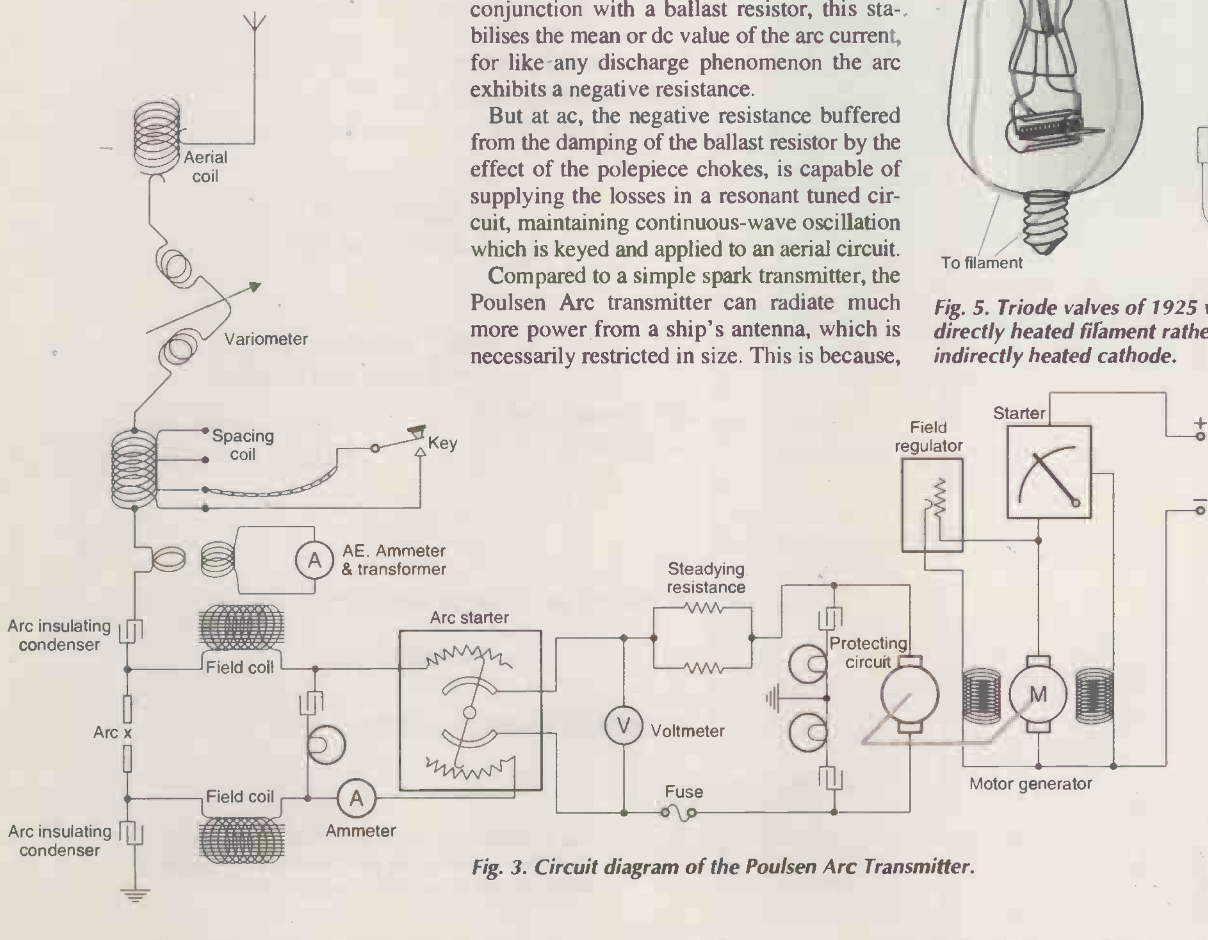
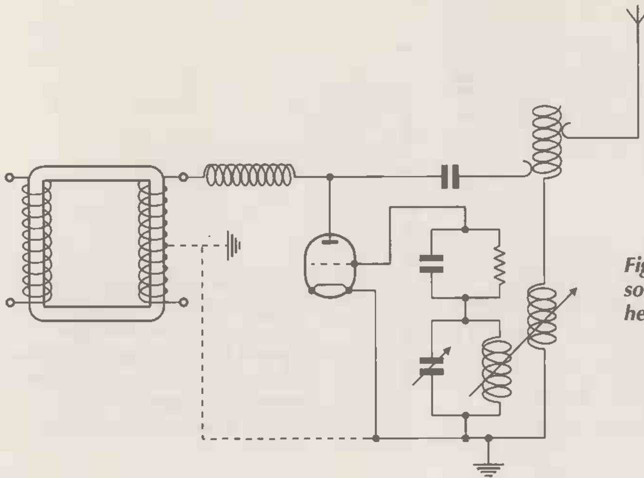


Fig. 3. Circuit diagram of the Poulsen Arc Transmitter.



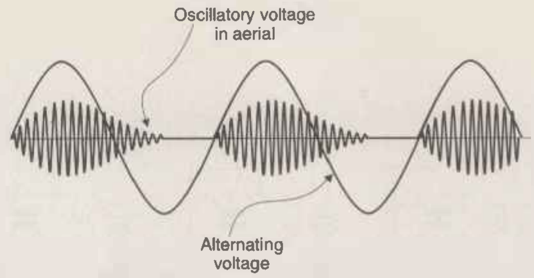
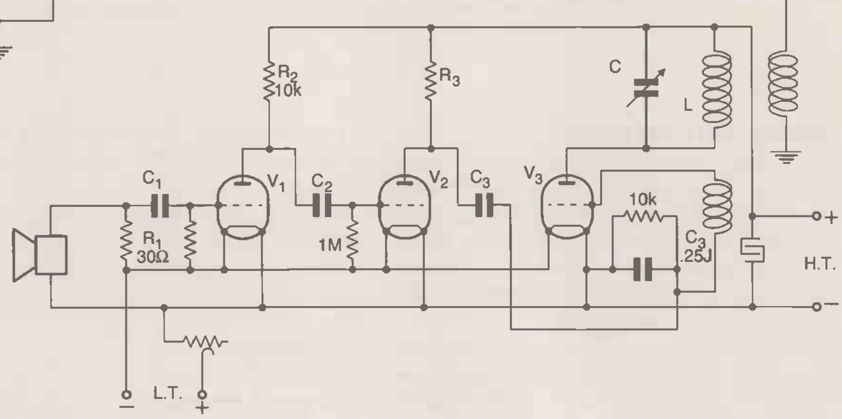


Fig. 8. Simple circuit providing ICW (interrupted continuous wave), so that Morse signals could be received on sets without a heterodyne oscillator (BFO).

Fig. 9. Telephony transmitter using grid modulation of the oscillator, which also doubled as the 'final'.



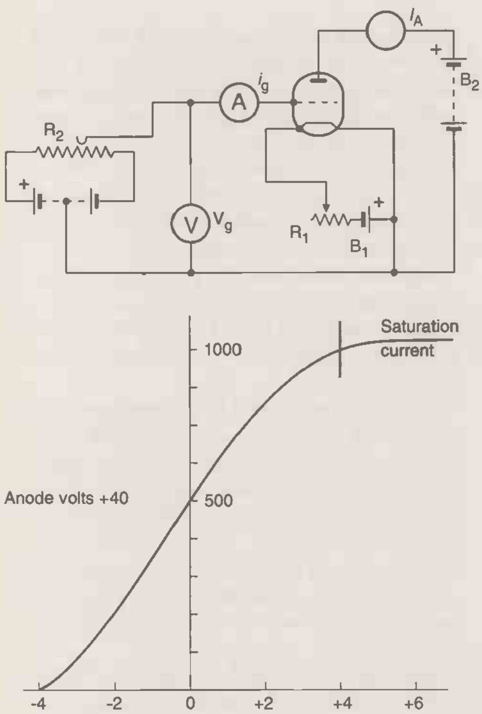


Fig. 6. Valve action was thoroughly understood, as illustrated by the test circuit A and typical characteristic B.

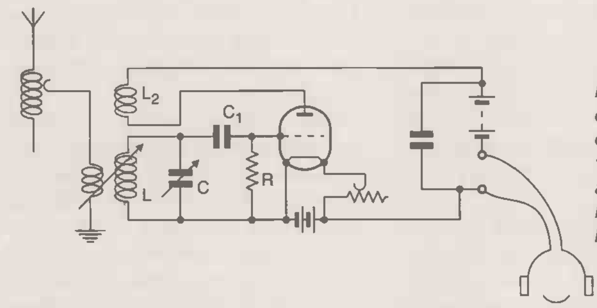


Fig. 10. Circuit of a one valve receiver employing 'regenerative amplification', which later came to be known as reaction.

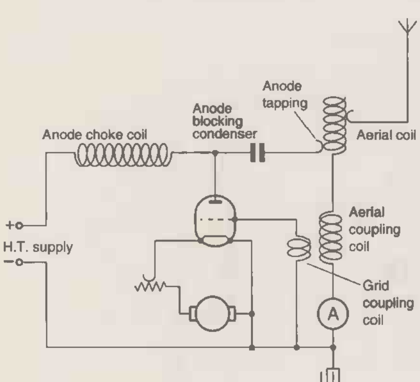


Fig. 7. In this simple transmitter, the frequency was determined by the resonant circuit formed by the aerial coil and the aerial's capacitance to ground.

designed for headphone reception. The variable inductor, variable capacitor and L_1 act as an antenna tuning unit, or ATU. Capacitor C_1 determines how tightly this is coupled to the main tuned circuit, L_2/C_2.

Of interest is the arrangement for applying forward bias to the diode to site the incoming rf at the 'bendiest' point of the characteristic; less than 100mV for a Bornite-Zincite detector, 300mV for tellurium-zincite or 700mV for carborundum-steel.

With the arrangement shown, if the particular detector you are using works best as a backward diode, so be it. Simply adjust the potentiometer for best results.

In 1925 valves were still at an early stage of development, **Fig. 5**, but their operation was well understood, **Fig. 6**. The earlier 'bright emitters', using a pure tungsten filament and consuming considerable filament current from a 6V accumulator, were being replaced, at least in receiver applications, by the new 'dull emitters'. These used a filament of thoriated tungsten and could consume as little as 60mA at 1.8V, though 60mA at 3V was a more common rating.

Both dull and bright-emitter valves were used in transmitters at this time. **Figure 7** shows a simple transmitter circuit. In this configuration, the transmitted frequency is determined by the resonant frequency of the variable antenna coil, together with the antenna capacitance.

Naturally, as the ship heeled, the wind blew and the gun turrets rotated, the antenna capacitance changed and so did the transmitter frequency. This made life difficult for another ship trying to receive the signal. Consequently, Fig. 277 of the handbook, a few pages further on, shows another transmitter circuit. This configuration has a separate parallel tuned circuit in the valve's grid circuit, resulting in much reduced pulling of the transmitted frequency by variations of the aerial capacitance.

At that time, some ships did not yet have heterodyne receivers. Their simple diode or valve receivers were fine for receiving Morse continuous-wave from a spark transmitter, or radio telephony. But continuous-wave from a valve transmitter, being simply a keyed unmodulated carrier, produced nothing but clicks in such a receiver, making the reading of Morse difficult or impossible.

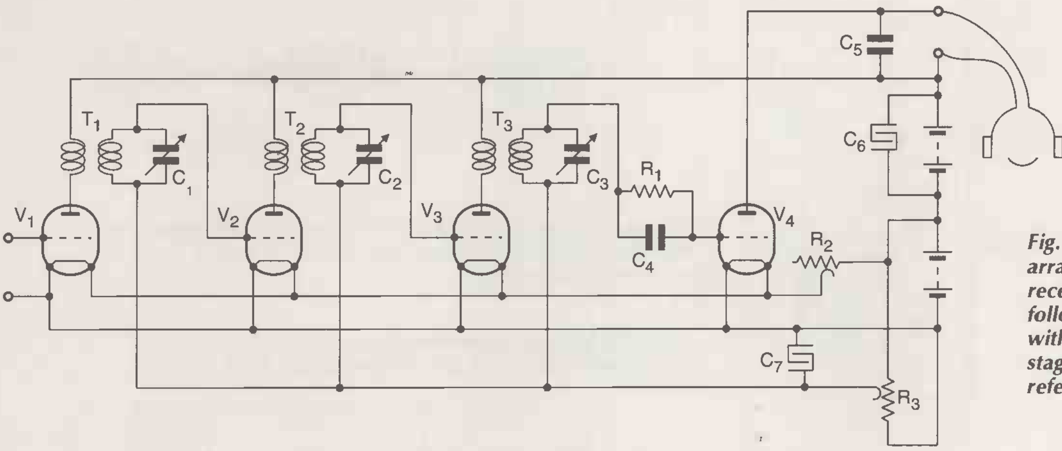


Fig. 11. Circuit (less aerial arrangements) of a complete receiver with three rf stages followed by a leaky grid detector, with no further af amplification stages; an arrangement later referred to as a 3V0 receiver.

Sparks, dots and dashes

By contrast, the dots and dashes from a spark transmitter consisted of a series of damped oscillations at the spark repetition frequency, resulting in an audible 'tone' in the headphones. **Figure 8** shows a simple transmitter modification to cope with this situation, without the keying arrangements.

The two diodes of the full wave rectifier, and the reservoir capacitor have all been disconnected. The valve anode circuit supplied direct from one half of the fullwave secondary winding of the supply transformer.

Now, the output is modulated at the supply frequency. An improved arrangement retained the fullwave rectifier circuit, omitting just the reservoir capacitor. As a result, the modulating or 'tone' frequency was twice the supply frequency.

A power supply complete with reservoir capacitor was used in the case of radio telephony transmissions, such as those produced by **Fig. 9**. This is a three valve transmitter where the oscillator V_3 – which doubles as the 'final' – is grid modulated by the two stage audio amplifier V_1, V_2, then called a 'note magnifier'. The carbon microphone is energised from the filament supply via R_1.

Figure 10 shows a one valve receiver using what later became universally known as 'reaction', except in the USA, where it was called 'tickling'.

The antenna could be resonated with the variable inductor, and its coupling to the tuned circuit *LC* was adjustable. In addition, reaction was controlled by varying the spacing between L_2 and L – known as 'swinging choke reaction'.

Complete receivers

Figure 11 shows a complete receiver – less antenna input arrangements. It uses four triodes, three as rf amplifiers and one as a 'leaky grid detector'.

This is surely an ambitious scheme. Indeed the book goes on to state under section 490, 'Prevention of Oscillations', that "One of the greatest difficulties... is to prevent the establishment of unwanted heterodyne oscillations." It lists a number of steps which can be taken, noting that most of them result in a reduction of amplification. These include using positive grid bias, leading to grid current and consequent damping of the tuned circuits, and the use of 'negative reaction' or neutralising. This later term is not mentioned.

The difficulty of obtaining a large measure of amplification at rf in the days before the appearance of tetrode screened grid valves and pentodes, led to the development of the 'superheterodyne' circuit, now universally used in receivers of all sorts. **Figure 12** shows the Armstrong short-wave circuit, designed, in this instance, to receive signals at 5MHz. Valve V_1 is a self-oscillating additive mixer, whose output is coupled to the three stage intermediate-frequency amplifier, V_2 to V_4.

Amplified IF signal is applied to a leaky grid detector. This additionally oscillated at a frequency removed by 1kHz from the 30kHz IF, to give a 1kHz tone for receiving Morse transmissions.

The three tuned 30kHz IF transformers gave great selectivity, while at the very low IF there was no problem of self oscillation of the three triode IF stages. Of course, with only a single tuned circuit at the wanted 5MHz signal frequency, and an IF as low as 30kHz, there would be little rejection of any unwanted signal at 5.06MHz. Nowadays, this is known as the image response, but then as the 'second channel'.

But in those halcyon days, signals in the hf band were very few and far between. For receiving really weak signals, a two valve af amplifier or 'note magnifier' would precede the headphones. ■

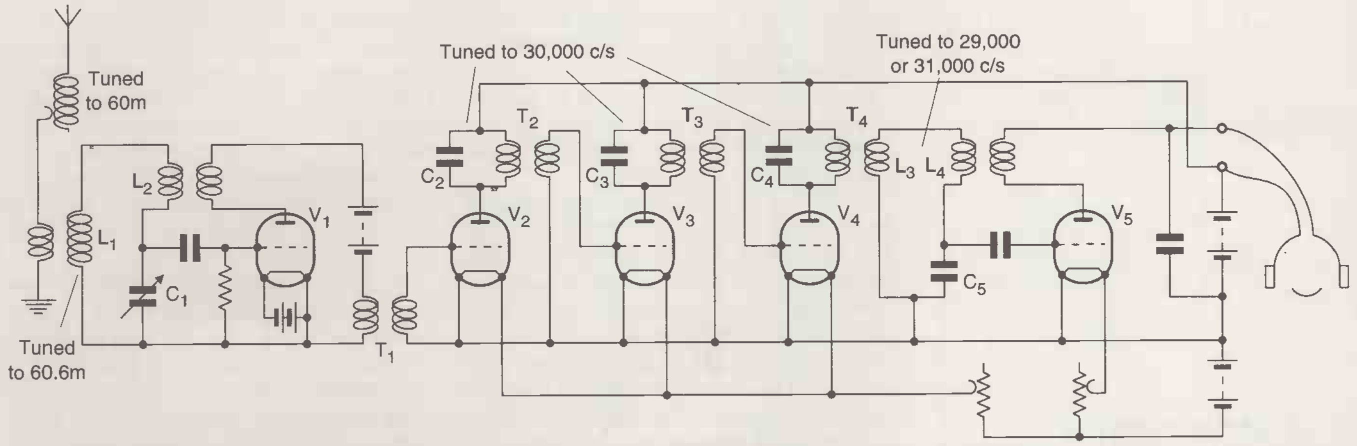


Fig. 12. The Armstrong Short-wave Receiver, an early version of the superhet, for reception of cw signals.

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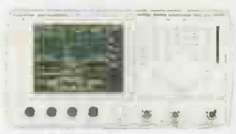

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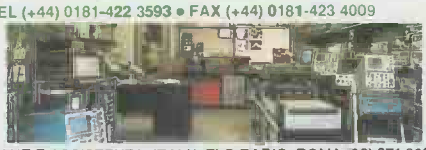

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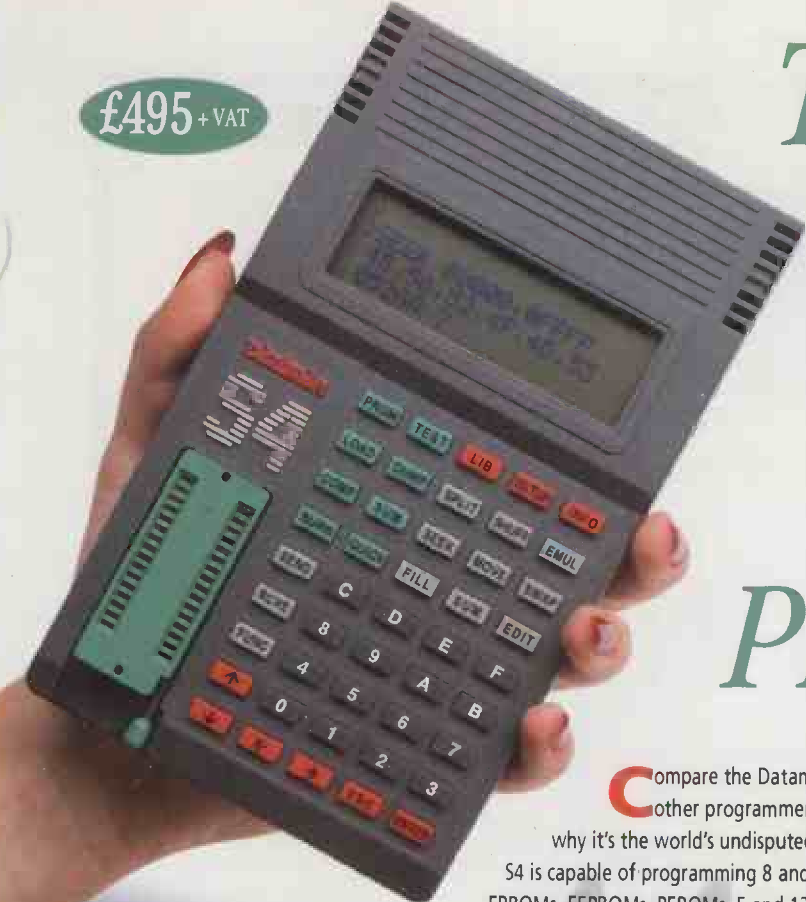

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