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SEPTEMBER 2002 £2.95

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DIY MFB loudspeaker
Spectrum pricing
Capacitor sound part 2

Circuit ideas:

Valve amp turn-on delay, Self-decoding indicators, Water level indicator

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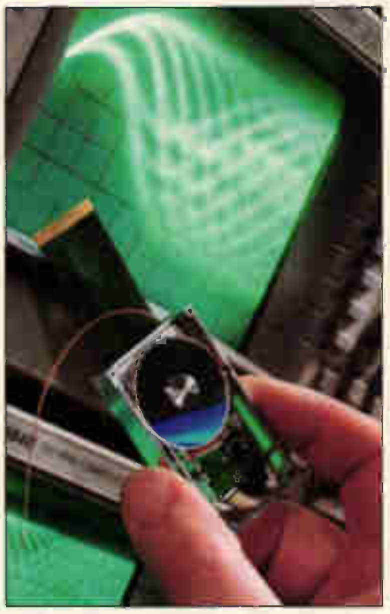

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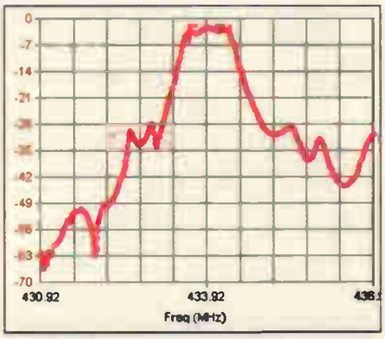

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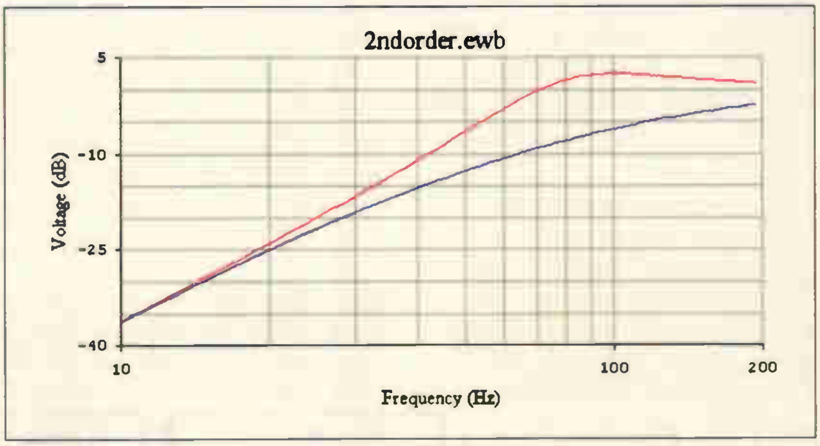

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Reader enthusiam

What an enlightening month! I have never worked for a magazine before who's readers feel so passionately about it.

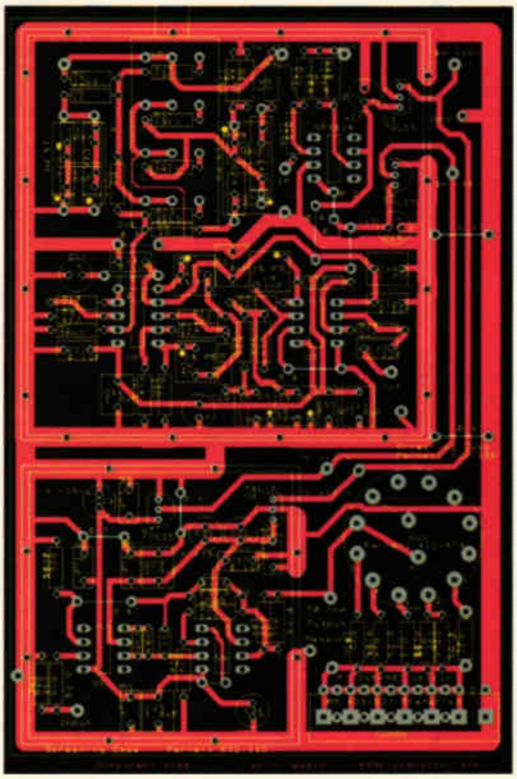
My mailbox was jolly full this month, as you can see from the extended letters pages. It was not all good, as some of you were quite critical about EW – but I certainly do not mind that. And nor do I mind publishing them. My job here is to produce a magazine that you want to read, and as I said last month, all comments are welcome.

I am trying to broaden the horizons of EW and I for one, would to see more in the way of computer articles, perhaps setting up small networks and even software reviews. Whilst I have plenty of traditional EW articles to hand – some of this 'new' stuff is a bit more rare. So, if any of you feel up to the challenge – let me know.

I am also considering running some 'new technology' items, which would take the form of insights into the enabling technologies of new products and services. Some ideas of mine would encompass digital imaging, video and audio compression, Bluetooth and wireless LANs. Again any suggestions gratefully received.

One reader suggested a free 'readers ads' page, where you could ask other readers' help in finding rare electronic items or old articles. Feedback please.

This month sees a mixed bag of articles from the second part (of six) of Cyril Bateman's excellent ideas on distortion caused by capacitors, the final part of David Rudd's spectrum pricing article and part one of a DIY motion feedback loudspeaker project from Jeff Macaulay. Also on the subject of audio (a subject dear to your hearts, I know) comes part 1 of an article on V-I protection in audio amplifiers by Michael Kiwanuka. For the RF aficionados, Marc Isherwood tackles interference problems on 433.92MHz. And of course we have the usual mixed bag of circuit ideas, industry news and new products.



Cyril Bateman's excellent ideas on distortion.

I hope you enjoy it.

Phil Reed, *Editor.*

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Blue laser fits 1Gbyte into 3cm disc

Philips has demonstrated a 1Gbyte fully functional 3cm optical disc drive, claiming it to be the worlds first.

Behind the development is the technology which will be used in Blu-ray optical discs, the generation to follow DVDs.

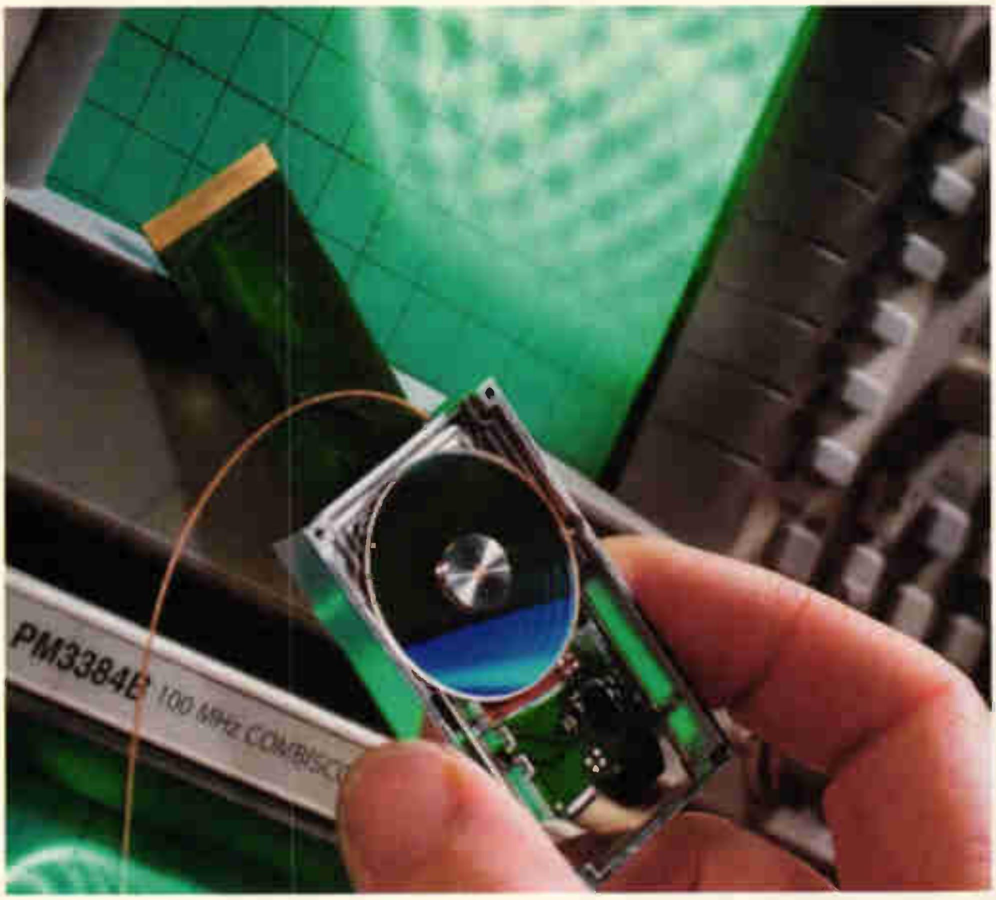
A blue laser is the key to storing so much data on a small disc and an all-plastic objective lens has cut the thickness of the 56x34mm drive to 7.5mm.

The lens makes a difference because it is mounted above the disc and has to have a certain diameter/thickness ratio.

Thinning it means reducing the diameter from the normal 3.5mm to a hard-to-handle 1.3mm. Although Philips has made both glass and plastic lenses, plastic allows extensions to be moulded onto the lens, which are later removed, for production handling.

Electronics are currently in a separate box, but the company sees no problems in fitting them in with the optics and mechanics.

The drive, which has a fully-removable disc, will allow Philips to try out future PDA and personal music player concepts, said the company.



Laser cuts inches of steel

IPG Laser from Germany has built an infra-red fibre laser that has a 2kW single mode, continuous wave output.

Diode-pumped fibre lasers such as this are gaining popularity in applications such as welding and cutting. They do not need any water cooling and the quality of the beam is far better than competing devices such as CO_2 or Nd:Yag lasers.

This 2kW version could easily be used to cut through several inches of steel or weld steel, copper and aluminium. In areas such as automotive, the beam can be delivered by up to 20m of single mode optical fibre.

IPG's design is an example of a standard diode-pumped fibre laser, made from a double clad fibre, with an ytterbium-doped core, said the firm's Dr Sergei Popov. The inner cladding is used to contain the pump light which is launched into the fibre from multiple diodes, each providing a portion of the input power.

The output is at around 1085nm. With a beam size of 50μm, the laser has a power density of around $100MW/cm^2$. The lifetime of the diode pumps is claimed to be over 100,000 hours, another advantage over other solid state lasers.

The entire system fits in a box measuring 1.1x1.2x0.6m. Higher power versions with outputs of 5 and 10kW are planned before the end of the year.

Superconductors yield another secret

Researchers at Ohio State University have nailed an answer to one of the remaining high-temperature superconductor (HTS) questions.

The mystery is associated with cuprate-type HTSs.

Scientists, according to professor Thomas Lemberger at Ohio, are still arguing whether cuprates exhibit d-wave or s-wave superconductivity.

S-wave is more desirable because this gives better properties at high temperatures. Unfortunately, most HTS cuprate compounds appear to exhibit d-wave behaviour.

The Ohio team has shown ceramic cuprates switch between s and d-wave under certain circumstances and by doping certain cuprates with enough cerium s-wave superconduction can be promoted.

"It seems that the mechanisms for both kinds of behaviour are always present in these materials," Lemberger said. "So if you do something to suppress one, a cuprate will automatically switch to the other."

UK firm ousts copper with superconductors

A superconductor company has been set up in Cambridge to exploit magnesium diboride, the material unexpectedly found to have superconducting properties by Japanese scientists in 1999.

Called Diboride Conductors, the company already has £10,000 from the Cambridge Virtual Incubator Fund. "In mid-Autumn we plan on going for £300,000, which will be a mixture of angel and VC [venture capital]," said the companys founder and managing director Dr Philip Sargent.

According to Sargent, the big interest in MgB_2 conductors is their low cost, cheaper than copper. He puts high-temperature superconductors (HTSs) at an expensive $3,000/kA.m, non-superconducting copper at $15/kA.m, followed by MgB_2 at an estimated $5/kA.m.

Sargent is a material scientist with a background in running companies. Diboride Conductors already has two patents on MgB_2 preparation for superconducting wires, with more on the way.

Although magnesium diboride is a true superconductor, having no resistance at all below its operating temperature, it is not a metal, but a black powder.

MgB_2 wires are prepared like HTSs, by filling a metal tube with superconductor powder, then drawing it into a wire.

"You pull it down [in diameter] at room temperature," said Sargent. "The powder gets [crushed] finer and finer."

Unlike HTS, which need a silver tube, MgB_2 can be drawn inside cheaper iron or copper.

MgB_2 superconducts around 20°K, too low for liquid nitrogen but "you can buy cryo-coolers that get to this temperature in one stage", said Sargent. A motor, for instance, could be cooled using gaseous helium.

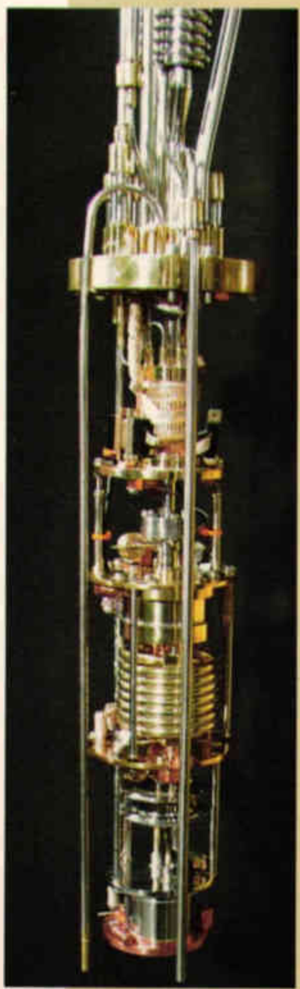
Using post-drawing heat-treatment, MgB_2 wires able to carry $7x10^5A/cm^2$ have been made, said Sargent, compared with $10^3A/cm^2$ for copper in room-temperature motors.

A drawback of MgB_2 is its poor performance in strong magnetic fields, but this is changing. "It has a pretty ropey resistance to high fields in wires made now. You could use it at 2T in an ordinary motor and could get up to 10T in some years," said Sargent. "Typical working conditions are 25K and 1T, so it can be used in transformers, motors and generators, but not in most MRI scanners."

Some small MRI scanners use weaker conventional copper-iron electromagnets and Sargent thinks MgB_2 will be used to make superconducting versions of these. "I expect to see demonstration machines in 2003," he said, "either low-field scanners or electric motors."

Continuously-used big electric motors are another obvious target as, over a 30-year life, electricity cost for these are 100 times the initial motor cost. Superconducting motors are also 80 per cent smaller, so ships could be another market.

Europe invests in nanotech



As part of its funding programme, the European Union is planning to invest €700m in nanotechnology in the coming four years.

"The challenge is so big that it has to be faced by solid public-private partnerships," said EU Research Commissioner Philippe Busquin.

"With private sector contributions this amount should rise to €1bn," he said.

Typical projects that could benefit include those in energy storage, processors and display technologies, bio-analysis and drug delivery, robotics and prosthetics.

Funding would cover projects such as COUNT, run by six European metrology institutes, including NPL in the UK. It is creating a single electron current source and current meter. The equipment pictured allows the circuits to work at temperatures below 50mK.

However, the EU's investment is still far short of the US funding for nanotech. "The US government is pouring 600 to 700 million dollars per year into this sector," Busquin said.

Disabled man helped by Aliens

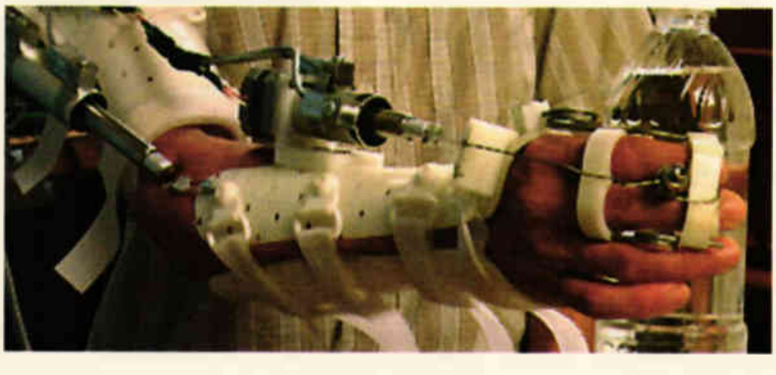
Using two motors, speech-recognition software and an exo-skeleton inspired by the film Aliens, undergraduates at the US Johns Hopkins University have designed and built a device to help a disabled man grasp and lift.

The project was started after the disabled man, with progressive muscle deterioration, approached Volunteers for Medical Engineering, a non-profit organisation, which got in touch with the university.

The team rejected electromagnets and air pressure systems, finally settling on two stepper motors for motive power. These move fingers and elbow in small, slow, increments and hold position without power.

The words "arm" and "hand" wake up the device, the elbow motor then responds to "raise", "down" or "stop", and hand motor responds to "open", "close" and "stop".

A waist-carried rechargeable 12V lead-acid battery powers the whole set-up. With the device a user can clasp a cup firmly without crushing it. "The students did a wonderful job," said Jan Hoffberger, executive director of Volunteers for Medical Engineering.



Batteries get Guardian angel

A four-person firm from Manchester is developing a chip that monitors standby batteries, giving constant updates on their condition and potentially extending their lifetimes.

Guardian Link said its technology could be used in thousands of sites across the UK, including hospitals, air-traffic control and other important sites where standby cells provide backup in case of mains outages.

"But nobody knows what condition these cells are in, because there's no fluid in them anymore," said Nigel Scott, MD of Guardian Link.

Also, said Scott, "they are more sensitive than the old batteries used to be".

Scott developed a technique to analyse the cell from the waveforms produced when varying frequency currents are placed across the cell's terminals. He worked with UMIST which formulated the algorithms needed to analyse the data.

"We wanted to use spectrum analysis to break cells down into electrochemical components," said Scott.

In 1999 this resulted in a benchtop instrument, sold to AVO in the US. Now Scott is working on a single chip version where the IC is integrated into the cell. The cell's condition is monitored constantly, with data delivered over a CAN-bus.

The mixed signal chip is in design at the University of Glamorgan. "It's fairly well advanced into its design. We should have it within nine to 12 months," said Scott.

Moreover, Scott said the technique is applicable to any battery chemistry. "The chip could go in a nickel metal hydride pack and tell you the condition of each cell in the pack."

Such a system could help paramedics in life-threatening situations when using defibrillators, for example, or with battlefield comms systems.

By controlling terminal voltage and avoiding under and over-charging, the chip can also extend battery life, Scott claimed.

This is because all standby cells have slightly differing terminal impedances, so charging at constant voltage causes slightly different charging levels. Over-charging causes heat damage, while under-charging results in sulphation.

Thus a battery with a rated lifetime of ten to 12 years might only manage six. "The chip can actively do something for the cell itself," Scott pointed out.

Robot technology steps forward

This is Honda's ASIMO robot, designed eventually to assist humans.

"I think it will take another dozen years until the robots will be used at home," said Toru Takenaka, chief engineer for the ASIMO project. "Until then there will be some use in the public areas, such as the robots working as a guide at the museums or robots working as a body guard to some persons. Or could be used in the dangerous situations instead of human beings actually being there."

It stands 1.2m tall and weighs 52kg. Currently ASIMO can walk, wave, grasp and lift things on its own, without human assistance.

Power comes from internal nickel-zinc batteries, brushless servomotors provide movement and the body is magnesium alloy.

Honda's biggest contribution to robots is probably its research into balance and walking which has produced some of the best biped movement algorithms that exist.

These are pulled together under Wind River's VxWorks real-time operating system.

"In the program at any given time more than ten tasks are running and if you include all of the programs, there should be something like some hundreds of thousands of commands in it," said Takenaka. "For instance, a task to take control of leg balance and task to operate arms and also the wireless communication with external systems. Also, there is communication between the motors that actually move joints."

Old idea cleans new telescopes

The huge hi-tech Gemini optical telescopes are being cleaned using surprisingly old-tech materials.

After much experimentation, detergents normally used to wash horses and natural sponges have proved to be the best combination.

Horse soap, apparently, is a strong detergent but non-abrasive and leaves no residue. Using it and sponges to clean the 8m mirrors was originally developed by technicians at the Gemini South telescope in Chile, but it has now been adopted at Gemini North in Hawaii.

The mirrors are 20cm thick, weigh 24 tonnes, and have to be lowered to the ground for cleaning.

Bright white LED breaks all records

Lumileds, the Agilent-Philips joint venture, has announced what it claims it the brightest white LED in production.

Called the Luxeon 5-Watt, it emits an incredible 120 lumen, four times the previous record holder which is also a Lumileds product.

Although the LEDs are available un-mounted, heat-sinking is said to be so critical that Lumileds will only be supplying most customers with 5W Luxeons pre-mounted to a aluminium-PCB sandwich. Even then, the devices must only be operated with this PCB in good contact with further heatsinking.

Several beam patterns are available, including a 'side-emitting' version which looks to be designed for use with a reflector.

Lumileds' 1W devices are available in a /O variant which has in-built beam-forming optics. Unfortunately for experimenters, this will not be the case with the 5W LEDs.

Green, cyan, blue and royal blue 5Watts are also planned, delivering outputs from 30l (blue) to 120l (green and cyan).

A roll in the PARC

US researchers at PARC, the former Xerox Palo Alto Research Center, have developed a technique for rolling copper into micro-scale out-of-plane inductors.

The manufacturing process is simple, they say, and the resulting devices have high Q-factors, up to 70 at 1GHz.

To make the coils, copper is electroplated onto a stress-engineered film. The stress in the underlying layer forces the copper to curl into shape.

Inductances range from 1 to 50nH, while the Q on a standard CMOS process is over 40.

PARC became an independent company in January this year.

Greek legend protects PC

Software giant Microsoft is developing an operating system that adds hardware encryption to block viruses, spam and limit software piracy.

Called Palladium, the system uses its combination of hardware and software to check data from the Internet, CD-ROMs, and even inside the PC, such as from the keyboard or to the monitor.

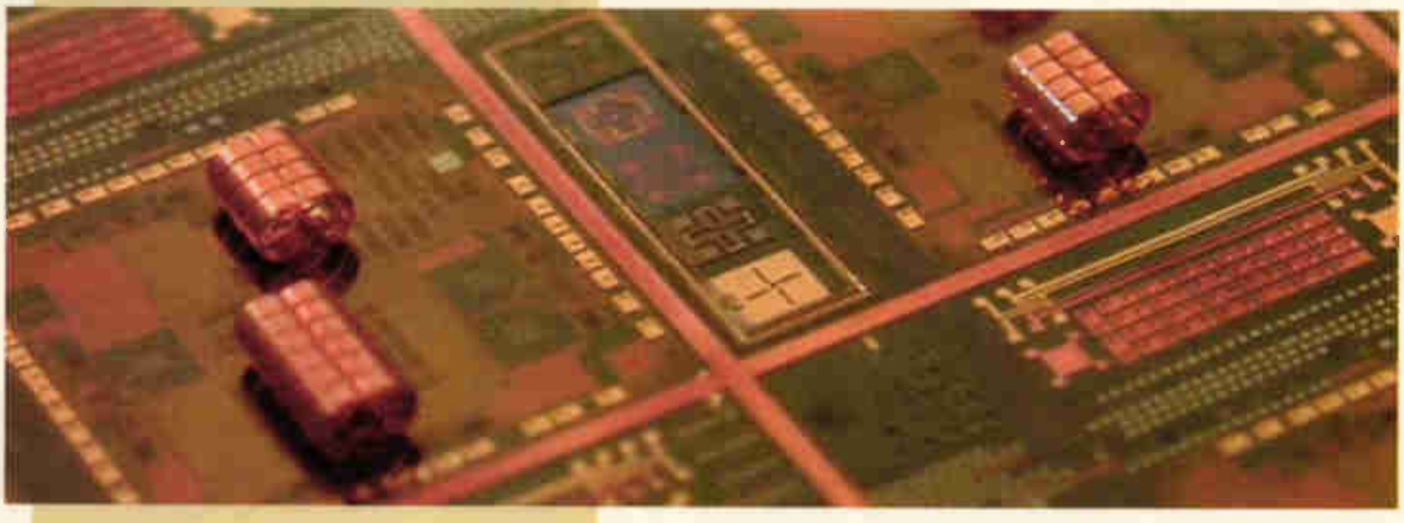
Palladium checks all applications for author and source before installing, which should stop viruses from loading. It could also apply encryption to documents, so e-mails could only be read and not modified, copied or forwarded by the recipient.

The OS will also be liked by the music and video industries, as it might allow CDs and files to be copied once for personal use, but not more, ensuring copyright is maintained.

Processor firms AMD and Intel are supporting the development of Palladium with dedicated chips for the encryption of data.

The OS could be included in an update to Windows within two or three years.

It is named after the statue of Athena that stood at the gates of Troy, allegedly protecting the city.



Women suffer in science

Women scientists have fewer opportunities than men, drop out of the profession sooner, while their careers progress slower, claims a report from the European Commission.

The EC survey, covering 30 countries' found discrimination against women in all the European nations.

Philippe Busquin, the EU's Research Commissioner, said: "The evidence demonstrates unquestionably that women scientists are under-represented in key scientific and research positions."

This is despite the fact that women outnumber men at undergraduate level in science, especially in fields such as biological and medical.

However, by the time they reach the top jobs, women are in the minority.

"This issue has to be addressed if we are to improve the position and role of women in scientific research," said Busquin.

He described the situation as a waste of women's skills and knowledge, as women drop out of scientific careers at every level.

Antenna cuts RF radiation

UK aerial specialist Sarantel has produced a mobile phone antenna that has a specific absorption rate (SAR) just six per cent that of competing designs.

The Wellingborough-based firm's PowerHelix design has an SAR of 0.07mW/gm, much lower than the 1.26mW/gm typical of existing mobile phone antenna.

New legislation is set to force mobile phone makers to state the SAR of their handset on packaging.

"The issue of mobile phone health is escalating rapidly," said Professor Yiannis Vardaxoglou of Loughborough University's Centre for Mobile Communications Research (CMCR). The centre has worked with Sarantel to model the near field from the antenna, which uses ceramics and a double-helix moulded aerial.

"The inspiration for a radical redesign of the wireless antenna hit me while I was sending a fax through to our patent agent," said inventor Dr Oliver Leisten, technical director at Sarantel. "Our double-helix design provides lower SAR values for mobile users and reliable connections for wireless LAN and Bluetooth applications."

The firm has 12 patents filed on its designs.

Magnetic RAM reaches 1Mbit

A magnetic memory device with a density of 1Mbit has been fabricated by Motorola. The achievement is a significant step toward commercial MRAM ICs.

Several firms are pursuing MRAM as it is dense, low power and non-volatile. Motorola and IBM have the major research programmes, although Infineon Technologies and some Taiwanese firms are also interested.

"MRAM has the potential to become the prevalent memory of choice for the vast majority of digital consumer applications," said Saied Tehrani, director of MRAM technology at Motorola.

All the research efforts with a reasonable chance of success are using a single transistor combined with a magnetic tunnelling junction to form the memory cell.

Both IBM and Motorola are expecting to go into commercial production of magnetic memories in 2004, by which time they hope to reach densities of 256Mbit.

Digital radio breaks £100 barrier

Digital audio broadcasting has come one step closer to the mainstream with the launch of a sub-£100 DAB radio. VideoLogic Systems' EVOKE-1 radio has gone on sale in the high street priced at £99.99. The radio is both mains powered and portable.

Magnetic sensor is most sensitive

Researchers at the State University of New York at Buffalo claim to have developed the world's most sensitive nanoscale magnetic sensor.

"The magnitude of the magnetic effect created surpasses all previous records," claims the US National Science Foundation which funded the work. Sensor resistance changed "more than 3,000 per cent", said the foundation.

Made of nickel and only a few atoms in diameter, the sensor works with ultra-small magnetic field, at room temperature, said inventors Harsh Deep Chopra and Susan Hua.

The effect is spintronic and called 'ballistic' magnetoresistance (BMR).

Chopra predicts the sensor will be used for data storage and the ultimate capacity will be around a terabit per square inch, far above today's densities

BT tests mesh radio in Wales Into the Valley

BT is running a wireless broadband trial in Pontypridd, South Glamorgan using technology from Radiant Networks.

The 100 home trial run by BT Wholesale will test the technology's ability to supply interactive TV and video-on-demand from Yes Television.

Radiant uses a mesh structure to deliver wireless data, allowing the system to be used in areas where point-to-multipoint networks suffer from multipath fading and non-line of sight problems.

"This trial, with Radiant Networks' mesh radio solution, will give us an extended look at the technology in a real-life situation and let us explore its suitability for BT to deliver broadband services in areas beyond the reach of ADSL," said Paul Reynolds, CEO at BT Wholesale.

Pictured alongside a Radiant node are Tom Kressner (back), Geoff Butcher (front left), and Paul Reynolds, the chief executives of Yes TV, Radiant and BT Wholesale.

Tachos to go digital

One of the UK's most archaic forms of technology, the tachograph, is set to be junked in favour of all-digital units logging data onto smartcards.

Switching to digital tachographs on trucks and commercial vehicles across Europe will result in millions of smartcards being issued to professional drivers.

By April 2004, all new vehicles over 3.5 tonnes will have the digital units. The ruling follows the European Commission's adoption of technical specifications for the tachograph.

"We're replacing an instrument that been in use since the 1970s in a way that's frankly prehistoric," said a spokesperson at the Freight Transport Association.

Existing tachos scratch through a layer of wax on their paper disc to record speed.

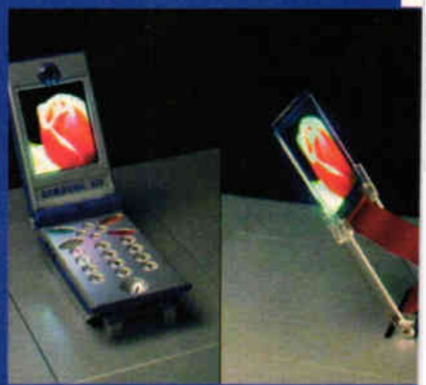
"It's a natural progression, but like any change it costs time and money," said the FTA spokesperson. "I'm told it will be 20 to 25 per cent more than existing tachos, and the cost relative to the vehicle is small."

The new tachos will interface better with vehicle electronics, said Steve Wilson from manufacturer Siemens VDO: "Vehicles are advancing by using CAN-bus systems, and the tacho must link to that."

The unit itself will include a small alphanumeric LCD and thermal printer.

US organic LED developer Universal Display Corporation has demonstrated a full-colour, active-matrix OLED mobile phone prototype with a 5.6cm display.

Samsung made the phone using UDC's proprietary phosphorescent OLED (PHOLED) technology. UDC came out of Princeton University.



Curved lithium ion cells are being produced by Philips. The 'Lithylene' batteries aimed at handheld products. To physically stabilise the cells, Philips has developed a technique of creating microscopic holes in the electrodes and separators which are then filled

Pico scopes out car diagnostics

PC instrumentation firm Pico Technology has introduced an automotive diagnostics package for its oscilloscopes.

The Automotive Diagnostics Package can measure car systems such as starter current, signals from ignition, injectors and ABS, and RPM sensor outputs.

The system "affords automobile technicians and engineers the ability to make accurate and high-resolution measurements and compare them to... expected results", said Alan Tong, Pico's technical director.

The system uses the firm's ADC-212 oscilloscope, which uses a standard PC to display traces. The kit includes a 600A current clamp, secondary ignition pickup and test clips, leads and probes. Software includes drop-down menus of common automotive tests and expected waveforms - including common faults.

"The test package is everything you need to measure and record the majority of the signals, currents and voltages responsible for whether a vehicle is functioning or not," said Tong.

The complete package costs £599, or £200 without the ADC212. ■

Tackling interference problems on 433.92MHz

Marc Isherwood, RF specialist at contract design and manufacturer, EMA, asserts that migrating from 433.92MHz to 868MHz to escape interference is only a short-term fix for Short Range Devices (SRDs) and that improving receiver design is a better long-term solution.

Since the days of the earliest transmitters, designers have continually migrated to higher frequencies in order to escape the interference caused by ever-increasing user numbers. Now, with the spectrum fully allocated from 9kHz to 300GHz, designers of Short-Range Devices (SRDs), such as security and automotive remote entry systems, are finding that they have nowhere left to run. There are no more green-field sites in the RF spectrum.

The harmonisation of radio frequencies in the EEC should have made life easier for manufacturers. Instead of producing devices for 224MHz, 315MHz, 418MHz and 433.92MHz the single Europe-wide standard was defined by ETSI 300-220 as 433.92MHz[6]. But 433.92MHz is not exclusively reserved for SRDs and the frequency is available for radiolocation as well as amateur and military users.

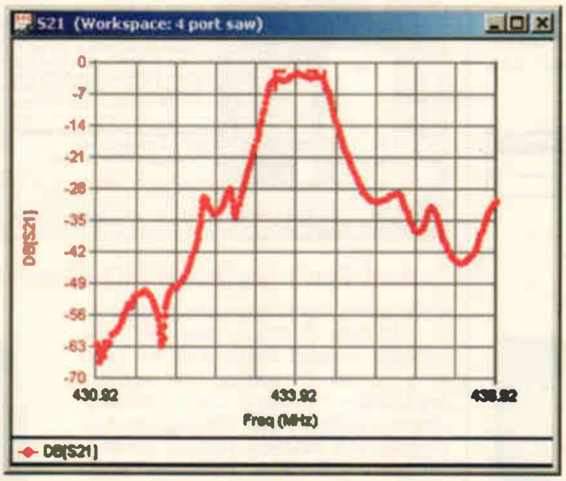


Figure 1: Correctly wired 4-pin SM package with no shared vias

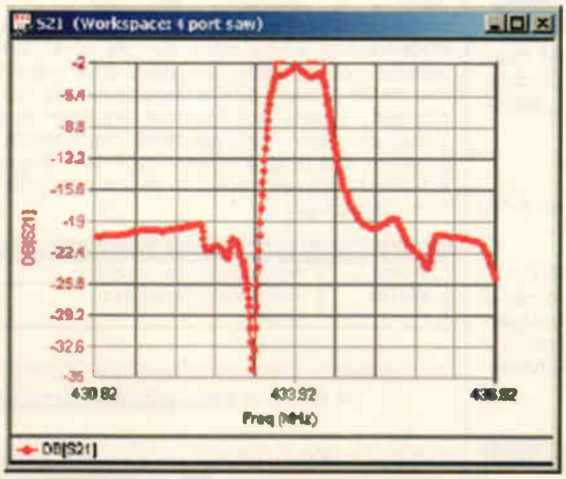


Figure 2: Degradation in performance caused by common ground pin

More traffic means more interference, and many designers of SRDs are abandoning 433.92MHz and have begun an exodus to the newly allocated 868MHz band[4]. But already the cycle is repeating itself: the level of interference on 868MHz is rising and it will only be a matter of time before 868MHz is as crowded as 433.92MHz.

Of course, modulation techniques such as the various spread spectrum systems[7] make extraordinarily good use of the available bandwidth, but older, less efficient systems are firmly entrenched. Even with the government's desire to turn off analogue TV transmissions to free up much needed space, interference will be a constant challenge for designers. The only viable long-term solution is to make best use of the space that is available. This means better receiver design and, when required, use of the more exotic modulation systems.

At the receiving end

Efficient front-end filtering is critical to overall performance. Today, SAW front-end filters are available with bandwidths of ±150kHz, making best use of the improved stability of the SAW manufacturing process. The major manufacturers such as RFM, Murata and Epcos make parts available in very small surface mount packages with all four connections for the input and output ports available. This does not however mean that things are simple for the designer.

To ensure there are no extraneous coupling effects between the input and output ports there must be no sharing of vias connecting to the ground plane. PCB tracks must be arranged to minimise inductive and capacitive coupling. Fortunately manufacturers are generally rather specific as to the correct layout for their parts.

To demonstrate the effects of extraneous coupling on the theoretically achievable performance of a typical SAW filter (Epcos B3555) I have performed simulations using the Eagle Gensys simulator.

In **Fig. 1** the Epcos B3555 is matched as suggested in the manufacturer's data sheet with no shared vias. The strange lumps in the response are mostly due to internal reflections in the filter substrate, which vary considerably from manufacturer to manufacturer.

A second simulation used a common ground pin and the degradation in performance caused by introducing this shared via is clearly visible in **Fig. 2.** Whilst the degradation in the bandpass characteristics might be considered minor, the ultimate rejection has suffered badly, losing some 40dB at 430.92MHz. This is due to the 1nH inductance now appearing as a shared ground between the input and the output ports. Even with an inductance as low as 0.1nH in the common ground pin, there is a loss of some 20dB in rejection at the same point.

A good approximation for the inductance of a PCB via is given by the following formula where h is the height and d is the diameter of the via in mm.

$$L \approx \frac{h}{5}\left[1+\ln\left(\frac{4h}{d}\right)\right] nH$$

Even with 0.8mm PCB material and a diameter of 0.3mm, the via's inductance will still be around 0.5nH.

In a similar manner, capacitive coupling between the input and output ports of the filter is equally bad. A capacitance of only 0.2pF will produce an equally poor result as shown in **Fig. 3**.

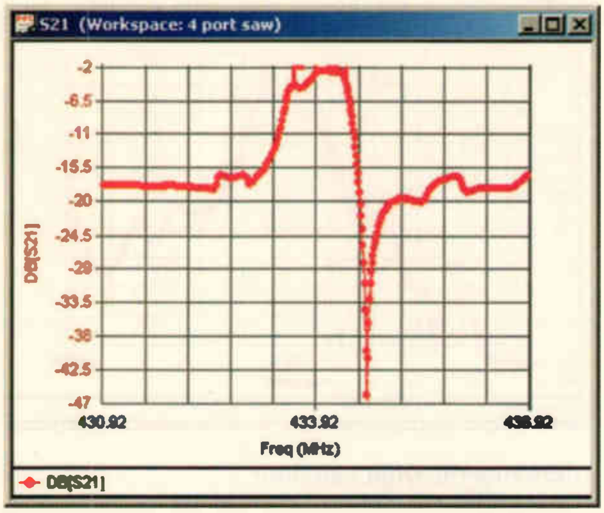


Figure 3: Effect of capacitive coupling between input and output ports

Another common problem arises with testing the system. Typically, all manufacturers test their RF filters in a 50Ω system, but in the actual application it is most unlikely that the impedances seen by the filter will be 50Ω. Impedance matching of the filter will be required. Some designers assume that the antenna will be 50Ω and only re-match the output of the filter to their system, **Fig. 4**.

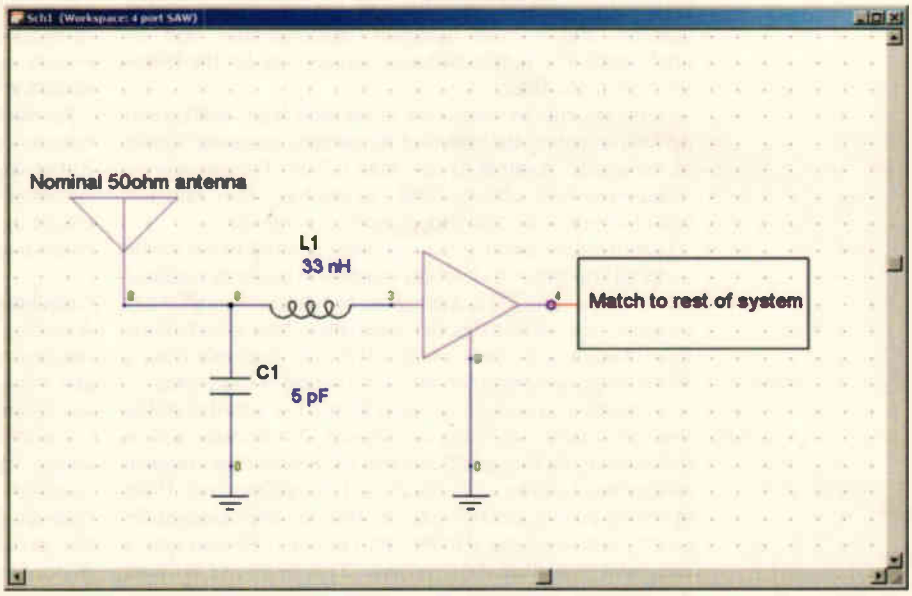


Figure 4: Typical layout of filter circuit

This will work perfectly when tested on the bench with a 50Ω signal generator or a 50Ω test antenna. Unfortunately, in the wild, antennas are often not the impedance they are supposed to be, particularly when simple whip or loop antennas are used. Even worse, some vehicle manufacturers use a length of wire embedded in the vehicle loom; take your pick for antenna impedance at this point!

Due its small geometry, the SAW filter is sensitive to static discharge. Although manufacturers are often rather coy about ESD ratings, SAWs can typically only withstand a maximum discharge of a couple of hundred volts. If you really must connect the SAW directly to the antenna then the notes on ESD protection on the RFM website will give useful guidance[9].

In addition, this arrangement provides no isolation between the aerial and the input of the receiver. Any local oscillator leakage present at the input is therefore directly coupled to the antenna and will contribute adversely to the EMC compliance of the product. Whilst a pure resistive error in matching to the filter simply results in a small loss of signal, a reactive mismatch alters the filter's bandpass characteristics.

A 10mm error in the effective length of a wave whip can reduce the filter rejection at 433.25MHz from a typical –19dB to –16dB and add another couple of dB to the pass-band ripple. This can easily occur when track or connection pin lengths between the SAW and antenna are not taken into account. Similar problems occur with stray capacitance on the antenna tracks. The variation in bandpass response with mismatch depends on the manufacturer of the SAW. Some are far more susceptible to mismatch than others.

Depending on the type of mismatch (inductive or capacitive), the filter either produces the classic triple hump in the passband, with rather sharper shoulders than normal, **Fig. 5**, or a substantially flat passband with rounded shoulders as shown in **Fig. 6**, Both forms exhibit a substantially greater in-band loss when compared with the typical –3dB that can be achieved with a properly matched part. (N.B. The mismatch has been exaggerated on both these graphs to make the effect clearer).

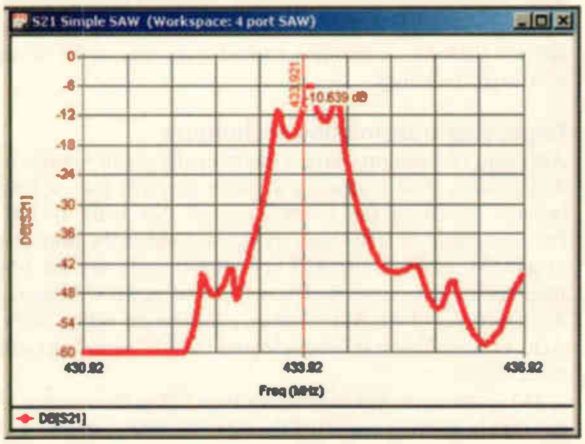


Figure 5: Effect of inductive mismatch on bandpass

These figures only take into account mismatch at the input of the filter. As a SAW filter is a symmetrical part, this mismatch has a direct effect on the impedance seen at the output side of the filter, which can further degrade the response of the system.

The same problems in matching exist with 868MHz systems, but the effects of stray capacitance and inductance are doubled. In addition, SAW filters for use at 868MHz typically have 1MHz bandwidth whilst the equivalent 433.92MHz filters are available down to 300kHz bandwidth. Although manufacturers are looking to remedy this, at present it is easier to make receivers more selective at 433.92MHz than at 868MHz.

Figure 6: Effect of capacitive mismatch on bandpass

Buffering the front end filter

Clearly what is needed is a buffer between the antenna and the SAW. The impedance presented to the SAW will then be essentially constant regardless of the impedance of the antenna, and so the bandpass characteristics will be unaffected. The buffer will also compensate for the losses inherent in the filter.

Common emitter stages are often used here, and careful design can reduce the effects of the antenna mismatch shown in the above example to less than 0.5dB. Typical reverse isolation will be about –20dB, so we have also gained a useful attenuation of any local oscillator leakage.

Care must be taken to ensure that the amplifier is stable under all conditions, as it is only too easy to create an oscillator rather than an amplifier. Attempting to squeeze the maximum possible gain out of the buffer is certain to lead to instability. Even though more than 14dB is typically available from a single stage, designing for an S_{21} of 6-10dB is much safer.

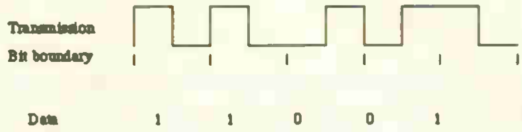
A common base stage can also be used as a buffer and in this application has some advantages. A transistor with a lower transition frequency than that required for the common emitter stage can be used, which will also reduce cost. If the operating point is carefully chosen, then the impedance at the emitter can be arranged to be very close to 50Ω with only a minimum of additional components. Typical reverse isolation will be about –28dB and the effects of our example mismatch will be much the same as with the common emitter stage. Although stability is again crucial, this is usually easier to achieve than with a common emitter stage, due to the lower transition frequency.

Improving transmission techniques

Although FM systems have a theoretically greater range, at 433.92MHz, AM systems will often perform just as well because much of the interference in this band is FM. Provided the level is not high enough to saturate the front end of an AM receiver, the AM system will only see the FM interference as a constant carrier. The FM signal will reduce the sensitivity of the AM receiver, either by the action of the AGC or logarithmic IF response, but the AM signal can still be detected.

AM systems are typically noisier than FM in the absence of an input, however a simple squelch arrangement can eliminate this problem with a loss in sensitivity of only a few dB. Careful design of the decoding software is essential to recover data in a noisy environment.

When sending data, consideration must be given to the fact that an RF link is not perfect. Software at the receiving end must have some idea of when a message starts, and should easily be able to extract the clock from the data stream. It must also be able to cope with missing data. All these points have already been given thorough consideration some time ago by designers of tape streamers, which use a form of Manchester encoding. In this form of encoding the clock and data are combined in such a way that there is always a transition at the centre of the bit window. The direction of this transition defines a 1 or 0. (Other forms of Manchester coding also work in a similar manner).



To give the receiver software some idea of when a message is about to arrive, a consistent preamble is placed at the front of the message. In AM systems the mark space ratio of this preamble should be the same as the average mark space ratio for the data. This is because in most receivers the AM signal is detected by measuring the difference between the actual signal level and the average signal level. If the average level in the preamble differs from that in the data, then the receiver will appear less sensitive when tested with real data than with a bit error rate tester. This is due to the fact that it takes a finite time for the average level to adjust to that in the data stream, causing the first few bits of the data stream to be missed at low signal levels.

As most SRDs are not transceivers, the receiver cannot request a data packet again if it was not decoded correctly. Either an error correcting code can be used or, more commonly, the message is transmitted more than once. To ensure the best chance of successful transmission, the length of the message should be kept as short as possible.

Conclusions

Avoiding interference by migrating to higher, less cluttered frequencies is a short-term solution to a long-term problem, and a front-end filter is essential for any receiver to work successfully in today's crowded spectrum.

It is tempting to believe that that a receiver can be made by simply by connecting an aerial directly to the input of a 'complete' receiver chip. However, even 'complete' receiver chips do not include front-end filters and, in the real world, the performance of such circuits will not achieve the performance guidelines published by RAKE (Radio Activated Keyless Entry).[8]

The only long-term solution to tackling interference lies in careful front end design, efficient software management and the correct choice of modulation techniques. These factors have a far greater effect on overall system performance than the operating frequency and allow designers to build circuits which operate effectively at 433.92MHz. ■

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Further reading:

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Socket testing is possible too thanks to the special patch cables included in the outfit.
Swapped lines, missing lines, shorted lines are all uniquely identified together with the full connection pattern!
What's more, if you want to know how to make up a special network cable, the Atlas can instruct you, even down to cable colours!
The Atlas IT is supplied in a robust carry case complete with a spare terminator and spare battery.
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Capacitor sound 2

Many capacitors introduce distortions onto a pure sine wave test signal. In some instances this distortion results from the unfavourable loading which the capacitor imposes onto its valve or semiconductor driver. In others, the capacitor generates the distortion within itself.

Output Buffer and Twin-Tee Notch/Pre-amp.

Most properly designed power amplifiers measure less than 0.01%, or 100 PPM distortion when sine wave tested at 1kHz. Such small distortions are believed inaudible, yet users often claim to hear distortions from these amplifiers when listening to music.

As a result many articles can be found on internet and in specialist magazines, claiming to have identified differences in sound, between different capacitor types. Not by measurements, but by listening tests, having upgraded a capacitor. This has led to a retrofit upgrade market supplying 'better' audio grade capacitors, at substantially elevated prices compared to mass market types.

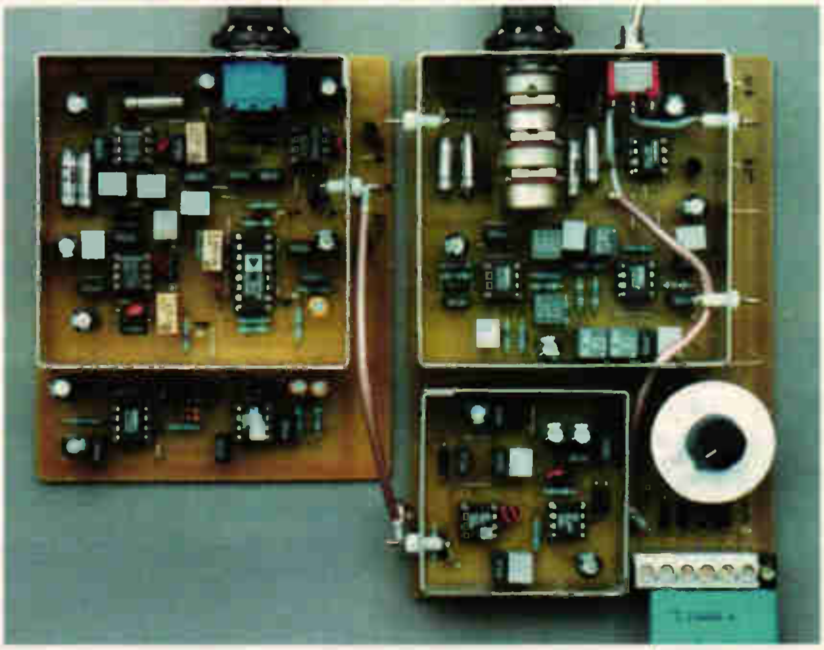
A common subjectivist claim is that oil impregnated paper capacitors sound better than film types in valve amplifiers. Others claim that a PET capacitor sounds 'tubby' while a Polypropylene sounds 'bright' and that all ceramics sound awful. Naturally these claims have no supporting measurements.

A year ago, a particularly acrimonious letters page dispute arose regarding capacitor distortions. It seemed some of the issues raised could only be resolved by providing proof positive, that many capacitors do cause distortion. I offered to perform some comparative distortion measurements.

Commitment honoured.

To measure the distortion level for most capacitors, a very low distortion generator

Figure 1: Very low distortion, low output impedance buffer amplifier, with passive Twin Tee notch filter, bandpass filters and 40dB gain preamp printed circuit board (right). This arrangement used with my low distortion 1kHz oscillator(left), can measure capacitor distortions down to -130dB.



complete with a matching low output impedance, low distortion, buffer amplifier must be used. An easily replicated, low cost, extremely low distortion test generator was described in my last article. Ref.1

This article describes a matching very low distortion, low output impedance, buffer amplifier needed to generate a pure sine wave voltage across a test capacitor. Having a near 600Ω input impedance, this buffer amplifier could equally be used with many commercial generators as well as with my design. **Fig. 1.**

To facilitate measuring capacitor distortions using low cost instrumentation, the 1kHz test fundamental should first be attenuated some 65dB in a passive Twin Tee notch filter. Reducing the dynamic range to be measured.

Using a typical 3 volts test signal, this attenuated test fundamental plus distortion components, is reduced to a few millivolts. This small signal should be bandwidth filtered and pre-amplified by 40dB, to allow measurement using a 16 bit computer soundcard or the 12 bit Pico ADC-100 converter.

An easily built, low cost buffer amplifier together with a notch filter/pre-amplifier, has been designed on a second PCB. Together with my 1kHz test generator Ref.1 these two provide a complete system able to measure distortions as small as -130dB, 0.3 PPM or 0.00003%, below a 5 volt test signal.

To replicate common circuit drive voltages, this buffer should be able to generate up to seven volts RMS across a 1μF capacitor, fed via a 100Ω current limiting source resistor.

Test Requirement.

Perhaps you already have a low output impedance test generator. The simple method I used to decide when my equipment was suitable for capacitor distortion measurements, will determine whether your existing equipment can be used.

Using a 100Ω source impedance, connect a 511Ω resistor to ground. Increase the generator output so as to measure 3 volts or more across this 511Ω using a DVM. Remove the DVM and perform a distortion measurement across the 511Ω resistor.

If one PPM or less, replace the resistor by a good, nearly perfect 1μF capacitor and without changing the generator output voltage, perform a distortion measurement across the capacitor. If less than 1 PPM, the equipment can be used to measure capacitor distortions.

The best test capacitor for this would be either a COG ceramic or an extended foil/Polystyrene. These are not distributor items so are impossible to obtain in small quantities.

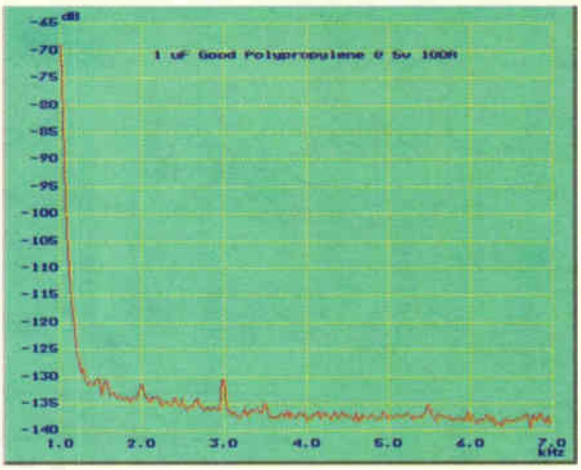


Figure 2: Plot of a near perfect 1μF foil and Polypropylene capacitor tested at 5 volts in series with a 100Ω source impedance. This plot includes not only any capacitor induced distortion but also that of my test system.

Next best is an extended foil and film Polypropylene, closely followed by extended metallised film electrodes with unmetallised Polypropylene dielectric. This last, manufactured by BC Components (Philips) is stocked by Farnell as part 577-881, 0.47μF 250v. I used two of these, type 376 KP 0.47/250v connected in parallel. **Fig. 2.**

If you have a generator able to provide suitably low distortion into a 600Ω resistive load, then my buffer amplifier may allow your generator to be used, however it is important to note that the series input resistance seen by my buffer, some 1120Ω inclusive of the 511Ω R38, is essential for its low distortion. This total value should not be changed.

Buffer amplifier design.

The buffer amplifier must not itself contribute measurable distortions. Since distortion levels measured in good capacitors are -130dB, 0.3 PPM or less, designing a suitable generator and buffer amplifier was no simple task. Designing a suitable buffer amplifier required almost as much development time as was needed for my low distortion oscillator. Ref.1

To drive 7V RMS into a 100Ω/1μF capacitor combination using my generator, a buffer was required with a gain of 2.

Many potential buffer amplifier configurations were breadboarded and rejected. While able to drive a resistive load, they were not able to develop a few volts across a 1μF capacitor without distorting.

An open loop buffer IC, the Burr Brown BUF634P used with an OPA604 in the makers suggested circuit, worked

Constructing the notch filter boards.

To provide a degree of notch filter tuning, a four gang variable resistor is needed, ideally it would be a well matched conductive plastic part. To fit within the screening case it cannot be larger than 18mm diameter.

I could not find a suitable four gang conductive plastic potentiometer. Alps do list a more modest four gang carbon track design, but again I did not find a supplier. Glancing through an old price list from Falcon Electronics. Ref.2 I found a four gang 4x50kΩ Alps potentiometer at £1.75, used in active crossover filters.

I ordered five potentiometers for evaluation. Apart from being rather old stock needing re-tinning of the terminal pins, they worked well and all were ganged closer than 1dB. I used these pots in both my 1kHz and 100Hz notch filter builds.

Since then a regular and valued correspondent, Juan from Spain, has written to me suggesting I look at the Sfernice P11 four gang 100 kΩ linear control stocked by Selectronic in France. Their part number 22.5700-1 is priced at 22.71 Euro. (http://www.selectronic.fr.) I visited their web site several times, but the web page will not accept a UK postal code. Without a postal code, their catalogue cannot be requested. I eMailed my request, but so far with no success.

The increased resistance of the P11 should not be a problem. To minimise potentiometer distortion, its tuning range is restricted by a 38k3 series resistor, then bridged by a 22k6 shunt resistor. With the exception of this variable control, to minimise noise and distortion and for easy replication, all resistors used in the twin tee notch filter signal path up to the first amplifier input, used 0.5% Welwyn RC55C, seen as black in the photo. To save space the four 38k3 series resistors are mounted between the potentiometer and PCB, so are hidden in the photo.

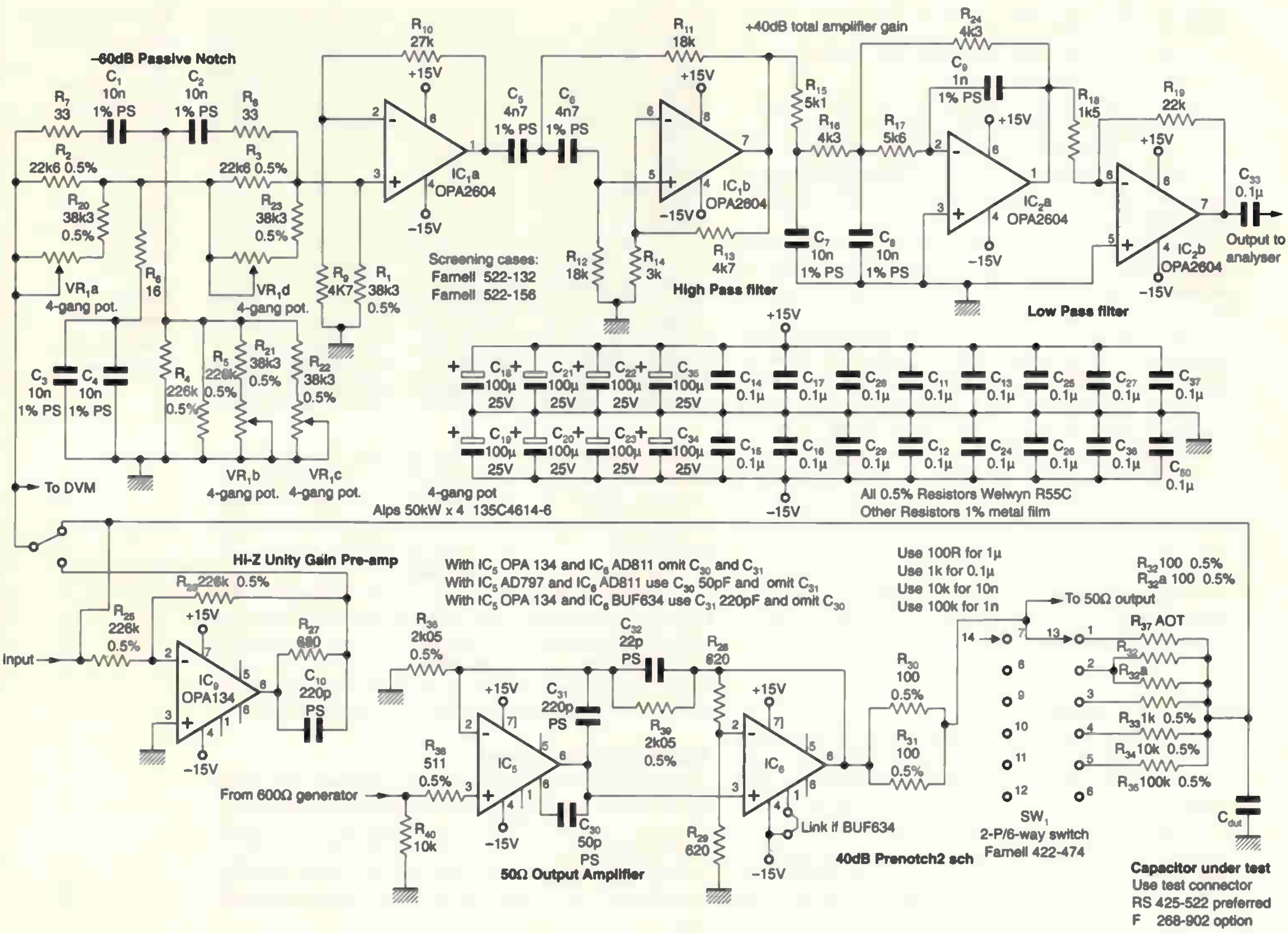


Figure 3: Schematic drawing of the low distortion buffer amplifier, pre-notch filter/pre-amplifier circuits shown in figure 1. The buffer amplifier at bottom, can drive more than seven volts at very low distortion, into a 100Ω/1μF test combination.

well at low drive voltages or with smaller capacitors. Loaded with a 100Ω/1μF capacitor test load, it distorted at increased drive levels. By closing one link, this combination can be used on my PCB.

The most nearly suitable circuit I tried was described in the Analog Devices AD797 datasheet. With an AD811 as the output driver, this combination claimed to be able to drive a 600Ω load to 7 volts RMS at 100kHz with less than -109dB distortion.

When breadboarded, this design produced less distortion driving into my capacitive test load than did the BUF634P circuit. For minimum distortion however, the circuit required critical matching of the impedances seen at both AD797 inputs. I was working to ensure suitable matching in November, when my only spare AD797 was damaged. Replacements not being available until February, I was forced to try other IC options. This combination of AD797/AD811 can be used in my PCB.

A low cost NE5534A worked quite well with this AD811 output stage, but again required careful input matching to minimise distortion. An OPA604 distorted at high drive, but the OPA134/AD811 worked best of all the combinations I tried.

Performance plots in this and my earlier article, were made using this OPA134/AD811 buffer amplifier.

With maximum drive into a 1μF load, the AD811 heats up, so should be fitted with a small heatsink, half of Maplin RN69. To minimise noise pickup, the circuit was screened using a small 50mm x 50mm Perancea solder mounting screening can and lid. To reduce heat build up, eight 8mm holes were distributed around the box sides with twelve 6mm holes in the lid.

Capable of more than seven volts output, I found this buffer circuit sufficient to measure distortions produced by capacitors from a few hundred picoFarads up to 1μF, at 1kHz, **Fig. 3.**

Notch filter/pre-amplifier design.

To ensure minimal distortion of the test signal, a passive Twin Tee notch filter, with a nominal input impedance of 10kΩ is used. To track the oscillator frequency, this notch is tuneable by some ±10% from its nominal 1kHz frequency. Measuring source impedances greater than 1kΩ, the loading of this passive notch filter is excessive. A high input impedance unity gain, low noise low distortion pre-amp can then be switched into circuit.

The notch filter is followed by four stages of low noise, low distortion, amplification and bandpass filtering. To minimise hum pickup, the filtered input is 50dB down at 100Hz. To reduce high frequency input into the measuring ADC, output is 20dB down by 22kHz. Amplified by 40dB, harmonics from the 2nd to 9th are maintained flat within 0.5dB

All measurements shown in this and the previous article, were made using this pre-notch filter/pre-amplifier as the input into my ADC-100 converter.

While care was taken to minimise noise and distortion in this notch filter/pre-amplifier, its contribution is included in all my test results. Using this notch filter/pre-amplifier, the distortion of my oscillator, built using AD797 IC's and the OPA134/AD811 buffer, driving 5V into my 100Ω/1μF test capacitor load, measured -130dB, or 0.3 PPM, **Fig 3.**

In most circuit applications, a capacitor is used either connected as shunt to ground or in series with the signal either to tailor the frequency response or simply block DC. Our test method should permit testing capacitors in either configuration.

Capacitor jigging

To avoid soldering the capacitor under test, some form of test jig, permitting easy exchange of various size capacitors, is required. The test jig must provide very low resistance and secure connections to the test capacitor.

I tried a number of spring contact terminal blocks. All but one required excessive capacitor lead lengths to ensure secure connections and that needed at least 5mm wires (Farnell part 268-902.) My PCB accepts this terminal block as well as the cage type below, **Fig 4.**

Ultimately for my own use I choose a 5mm centres, cage type, screw terminal strip, able to measure capacitors having 4mm long wires (RS part no 425-522.)

Designed to accept thick wires, it easily accepts 2.5 and 7.5mm spaced leads within its cage mouth. These cage terminals grip a wire tightly but without bending or damaging the capacitor leads. This terminal strip 'jig' was used for all 1kHz measurement plots.

The buffer amplifier/test jig shown can be used to test either series or shunt connected capacitor configurations. My preference is to shunt test, exactly as shown in the photo. The switchable current limiting resistor in series with the test signal, the capacitor being connected between signal and ground, Fig 1.

This provides two benefits:-

1) A good capacitor acts to slightly reduce any test generator harmonics, while a bad capacitor clearly shows much increased harmonic amplitudes.

2) The capacitor test voltage can be measured directly, using a high impedance meter attached to the DVM output test point. This test point measures the voltage at the input to the passive Twin Tee notch filter.

A test capacitor connected in series with the test signal, depresses the lower frequencies while slightly increasing higher harmonics, relative to the shunt connection. The test voltage can only be measured by connecting a DVM directly across the capacitor. This DVM must be removed before the capacitor can be tested.

Harmonic levels between the two methods differ by only one or two dB for the same capacitor voltage. A good capacitor looks good, and bad capacitors look bad, regardless of testing in the series or shunt connection.

By way of comparison, using a 1kΩ source impedance, I plotted test results of a known bad, 0.1μF metallised PET capacitor, measured in both series and shunt modes at 5 volts. In comparison, the third harmonic distortion peak of a good 0.1μF metallised PET capacitor tested at the same voltage, measures substantially lower, around -125dB. **Figs 5&6.**

Series tests.

To test in the series mode, the test capacitor and current limiting resistor are simply interchanged. The test capacitor is connected to the A.O.T resistor Vero Pins and the switch is set to the A.O.T position. The current limiting resistor is fitted to the test jig terminals, replacing the test capacitor shown in the Figure, Fig 1.

Test Capacitor Source Impedance.

The buffer amplifier output switch provides selection of four values of current limiting, or source impedance resistors. In principle any resistance value can be used to test any capacitance. However this resistor value determines the maximum test voltage which can be developed across the capacitor and the test's sensitivity.

By way of illustration I plotted test results for a 220pF Y5P 50v ceramic capacitor, Farnell 896-524, using each value of current limiting resistor in turn. At 1kHz a 220pF capacitor has an impedance around 720kΩ.

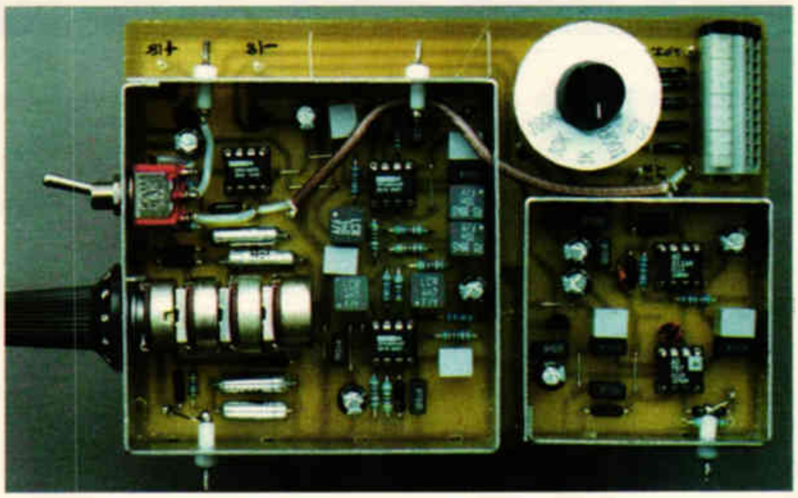


Figure 4: The PCB is double pierced so as to accept either the screw cage terminal test jig, as shown in figure 1, for capacitors with lead spacing up to 30mm. Alternately this 'spring contact' terminal strip, accepts lead spacing up to 27.5mm centres.

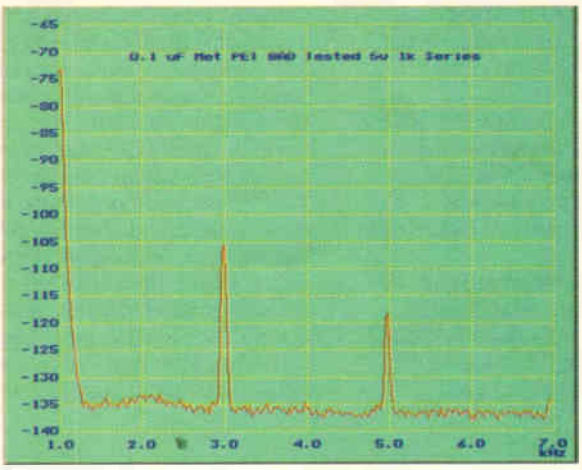


Figure 5: Distortion plot of a known 'bad' 0.1μF metallised PET capacitor tested at 1kHz with 5 volts across the capacitor, using the optional 'series mode' connection. The capacitor is in series with the test voltage, the 1kΩ current limiting resistor, is to ground.

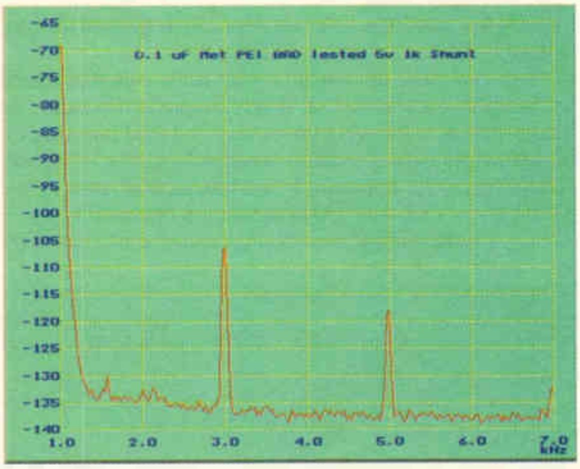


Figure 6: Distortion plot of the figure 5 capacitor and with the same 5 volts 1kHz signal, using my standard 'shunt' connection. The 1kΩ current limiting resistor in series with the test voltage, the capacitor connected to ground as in figure 1. Almost identical distortion was measured in both configurations.

Other measuring methods

In the sixties, engineers at Ericsson believed that non-linearities in capacitors and resistors could be detected. They measured the level of third harmonic distortion generated in a component subject to a very pure sine wave test signal. Ref.3 Non-linearities were believed to result of badly ground resistor spirals, poor electrical contacts and non-linear materials. At that time poor contacts, especially in capacitors, were commonplace. Fortunately today, with improved techniques, poor contacts in capacitors are now quite rare.

Their original non-linearity detector design produced low distortion test signals at 10 and 50kHz. Third harmonic distortion generated by the component under test was passed through bandpass filters for measurement. Subsequently the 50kHz test frequency was dropped and a commercial instrument, the CTL1 component linearity tester, was produced by Radiometer of Denmark. Ref.4

To accommodate the range of component impedances and test voltages needed, a low distortion output transformer was used. Having seven adjustable tappings, it was used to tightly couple the instrument to the component under test. Component impedances from 3Ω to 300kΩ could be directly measured, using source impedances from 0.05Ω to 500Ω respectively.

When testing lower impedance capacitors, the CTL1 datasheet which I still have, claimed to be able to output 0.58 A maximum. Resulting in a maximum test voltage around 100mV at 10kHz testing a 100μF capacitor. In my view this is not sufficient to reveal the true characteristics of such an electrolytic.

Today an updated version can be obtained from Danbridge A/S, Denmark, a specialist manufacturer of capacitor test instruments. Some specialist audio suppliers quote distortion levels for Electrolytic capacitors, measured using the CTL1 meter. Because of the capacitance values measured and the 10kHz test frequency, these results usually are based on extremely small test voltages. Such small test voltages will not harm the capacitor and will reveal any shortcomings in the metallic connections used in an electrolytic capacitor. However, in my experience, today these are at such low level as to be unimportant.

Most important and relevant to audio in my view, are the inherent distortions which result from the electrolytic capacitor's diode characteristics. This diode characteristic is easily measured. Ref.5 From my test measurements at 100Hz and 1kHz, I find significant and measurable distortions when testing electrolytics, using voltages above 0.5 volts, but less so at very low test voltages. This is exactly the result to be expected from consideration of the constructions used to manufacture these capacitors.

Extremely tight coupling between the test capacitor and the linearity tester is implicit in the CLT1 equipment design. From my early work measuring capacitors, I found it necessary to loosen this coupling in order to clearly reveal anomalies, now found in many modern capacitors. By trial and error, measuring known good and bad capacitors at 1kHz, I found that 100Ω in series with a 1μF capacitor provided the best compromise between measuring current and capacitor voltage. Adjusting this resistance value according to the capacitors impedance, at the test frequency used.

Only the 100kΩ and 10kΩ plots are shown. These clearly show that as the capacitor is more and more closely coupled, then its distortion peaks look smaller. Tested with 1kΩ the third harmonic peak had fallen to -121dB and with 100Ω to -127dB. **Figs 7 & 8**

Readers may recall it was use of a 220pF capacitor 10kΩ resistor low pass filter combination, which sparked off considerable reader discussions last year.

The actual value of current limiting resistor used or source impedance, determines how tightly coupled is the capacitor to the source. Using very low source and load impedances, makes even a badly distorting capacitor look relatively good. This is my main objection to the test method used by the CTL1 tester. (See box 'Other measuring methods'.)

This is a measurement quirk, the capacitor still generates the same distortion currents, but the measurement cannot see them. Similarly when testing with reduced voltage, the distortion still exists, but can be lost in the noise floor and so not seen. From many measurements of known good and bad capacitors, I found that a compromise between these impedance extremes should be used. Using a 100Ω current limiting resistor with a 1μF capacitor gave the best and most consistent results. Good capacitors looked good and bad capacitors looked very bad.

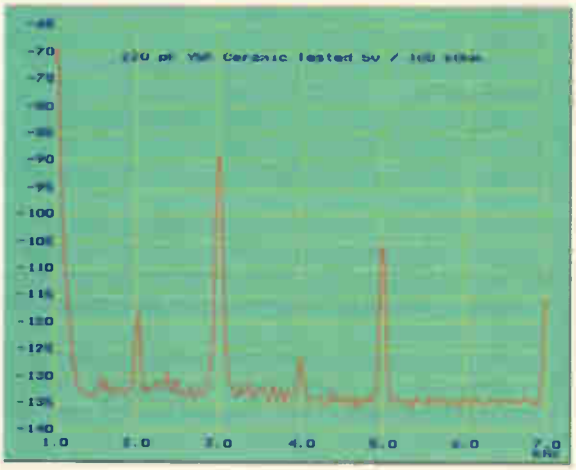


Figure 7: Distortion plot of a 220pF Y5P disc ceramic capacitor tested using a 100kΩ current limiting resistor and with a 5 volts 1kHz test signal across the capacitor. Clearly shows significant distortion products when tested using this source impedance.

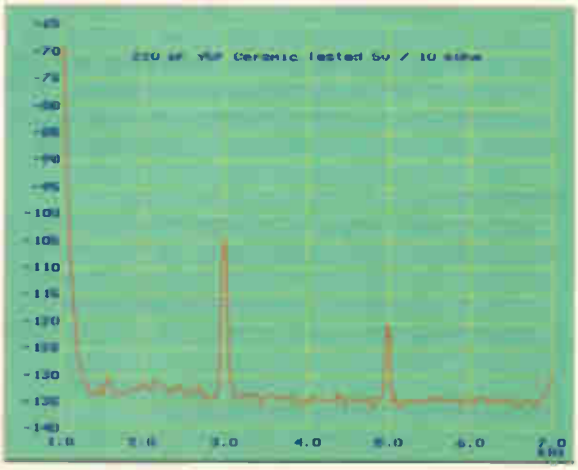


Figure 8: Distortion plot of the figure 7 capacitor, tested exactly the same except for the current limiting resistor, now 10kΩ. Because the capacitor is more tightly coupled to the very low distortion test source, its distortions are partially decoupled, so appear much smaller.

Alternate IC's/Components.

To produce a low distortion notch filter it is important to use resistors having a small voltage coefficient. To ensure an easily reproducible design, I used 0.5% Welwyn RC55C metal film resistors, visible as black in the photograph, in the signal path. These are marked as 0.5% on the schematics. These resistors use plated steel endcaps, which I prefer for reliable long term end contact stability. Many subjectivists claim non-magnetic endcaps are better. I do not subscribe to that belief.

Having emerged from the notch, the fundamental signal has been reduced to a few millivolts, so my usual 1% resistors can be used. Amplified by 40dB, the maximum output signal is still less than 0.5 volts.

Low distortion, low noise ICs must be used in this amplifier circuit. In my tests I found the OPA134 worked better than the OPA604 for high input levels, but found the reverse when amplifying the tiny voltages output from the notch filter. For my builds I used OPA134 for the high input impedance, high level, switchable pre-amp U9 and OPA2604 dual IC's for the low level amplifier stages U1, U2. In each case my preferred IC choice is the first type listed on the schematic drawing. To facilitate evaluating IC's I used Harwin turned pin sockets for each position.

Similarly for capacitors, those used in the notch filter must be low distortion and for the 1kHz version, 1% COG ceramic or extended foil/Polystyrene types only should be used. At 100Hz which requires 100nF, such capacitors are not easily obtained. Foil/Polypropylene then metallised Polypropylene, in order of preference, can be used.

Thus I would normally use the 100kΩ source impedance when measuring test capacitors of 1nF and below. Whether these measured capacitor distortions are audible or not depends on the capacitor's location in the circuit, the subsequent gain of the circuit, capacitor voltage drive levels and whether the capacitor is inside or outside the negative feedback loop. Since I cannot determine that, my object was simply to prove absolutely, using easily repeatable methods, that many capacitors can and do distort a very pure sine wave test signal.

Intermodulations.

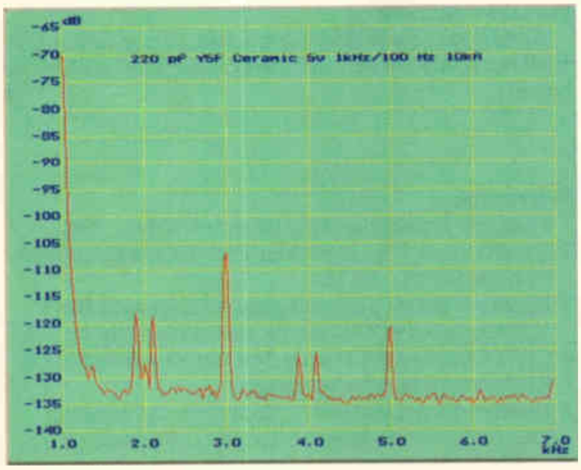
Is it not possible that any measurable capacitor distortion using a single tone test signal, say distortion greater than -120dB, will be made many times worse, when subject to a multiplicity of signals?, thus contributing notable intermodulation distortion.

Intermodulation distortion measurements of such capacitors using just two pure tones, 100Hz and 1kHz, do show a multiplicity of distortion products, almost regardless of dielectric. Similar intermodulation distortions have been measured in bad metallised film capacitors, i.e. those which show significant distortion above -120dB, using a single tone. Testing good capacitors with the same two tones, resulted in no intermodulation products being seen.

Comparing the single tone test in figure 8 with the dual tone test in figure 9, we see distortion products around 2kHz and 4kHz in this dual tone test. They are not visible in the single tone test, even though both tests used the same capacitor, voltage levels and source impedance. **Figs 8 & 9**

The level of distortion measured is naturally dependant on capacitor style, construction and the AC voltages present across the capacitor terminals.

Figure 9: A dual test frequency intermodulation distortion plot, 100Hz and 1kHz, of the capacitor shown in figure 8. Made using the same voltage and source impedance. Notice the appearance of new distortion products around 2kHz and 4kHz, not present when using the single test frequency. Bad metallised film capacitors exhibit similar distortions.



Measurement equipment

I have designed a second printed circuit board, similar to that housing my test oscillator which provides both the buffer amplifier and notch filter/pre-amplifier needed to complete a measurement system. The buffer amplifier section is designed so it can be easily separated from the notch filter/pre-amplifier if desired. **Figs 10**

For values above 1μF, it is common practise to change to using electrolytic types, both tantalum and aluminium. To avoid overstressing such capacitors while maintaining simi-

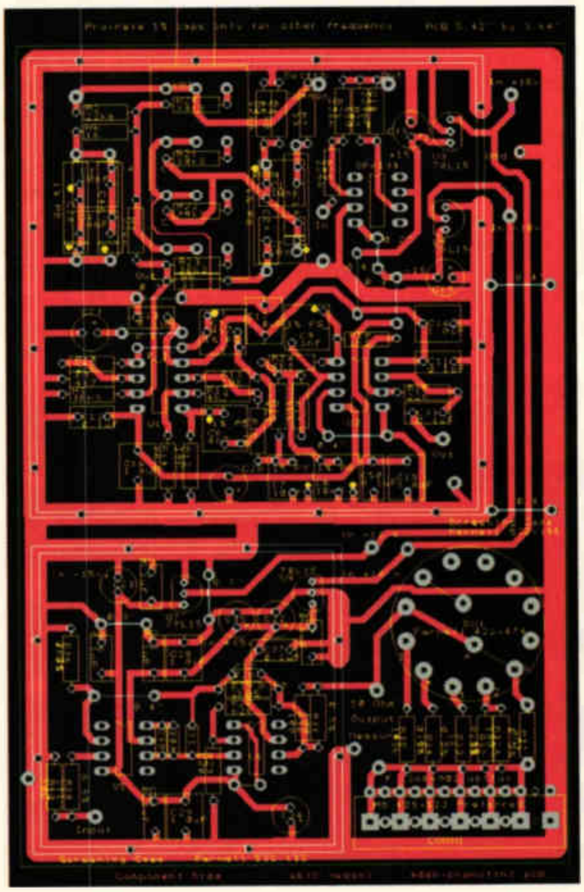


Figure 10: The version II printed board designed for my 1kHz notch filter/pre-amplifier and low distortion buffer amplifier. This arrangement was used for all measurement plots in this and my previous article. The board is multi-pierced to allow the widest possible choice of Twin Tee notch and band-pass filter tuning capacitors.

Soundcard FFT Software.

In this and my earlier article I used my Pico ADC-100 for all measurements, with the latest software downloaded from their site. However many readers will not have this ADC and wish to use a soundcard instead. A modern low cost PCI card with FFT software can provide improved capability, measuring even smaller distortions using my instruments, than is possible using the ADC-100. The software I choose to use for the remainder of this series, is the 'Spectra 232Plus' FFT software. It can be downloaded from: www.telebyte.com/pioneer

Should you have only an older ISA soundcard, some software may not work. One that will, is FFT.EXE, a DOS program by Henk Thomassen. This can be found on the internet, also the Elektor 96-97 software CD-ROM.

Users having a modern PCI soundcard will find a very large variety of programs, often available as freeware, on the internet. One site which links to some of the better packages is: www.pcavtech.com/links/index.htm.

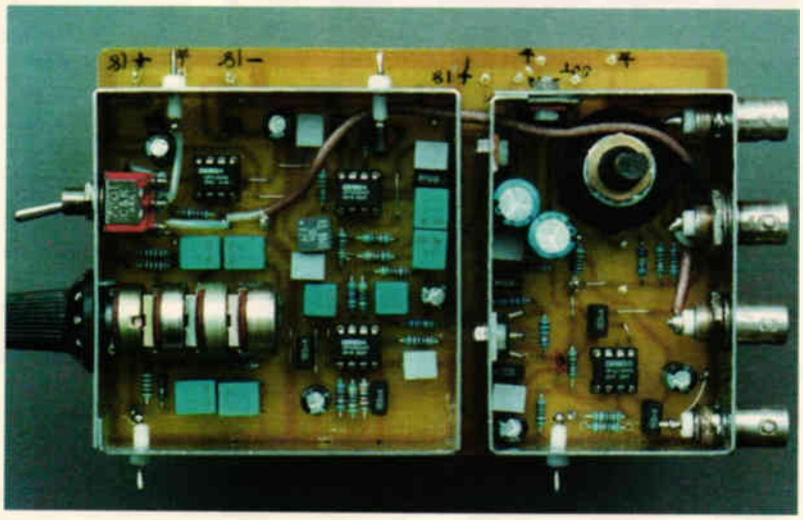
lar test voltages, a reduced test frequency must be used. I developed an alternative buffer amplifier, able to drive up to 7 volts and 400mA at 100Hz, albeit with slightly greater distortion than for my 1kHz design. Since electrolytic capacitors distort more than the lower valued better quality film and ceramic types, this small increase in distortion is acceptable.

The printed circuit boards for my 100Hz and 1kHz generators are identical. The only component differences are the three low loss tuning capacitors, C1, C2 and C3 which are 100nF 1% for 100Hz. One resistor value, R16 is 1kΩ for 1kHz but 0Ω for 100Hz. Pads for a wire link have been provided.

The 100Hz notch and bandpass filters are also based on the 1kHz design and need ten times capacitance values for 100Hz. The board layout accepts the Vishay 100nF 1% MKP capacitors (Farnell 303-8609), also 47 nF (Farnell 303-8380.) Smaller capacitances were provided using the same capacitor types used for the 1kHz design. However, as can be seen in the photo, the buffer amplifier section of this PCB layout is quite different. **Figs 11**

The full schematic and PCB layout for this 100Hz version will be included in a future article, 'Testing Aluminium and Tantalum electrolytics'.

Figure 11: Photograph of the 100Hz version printed board assembly complete with BNC sockets allowing use with Hewlett Packard test jigs or four separate coax cables. The board is identical to figures 4 and 10, except for the tenfold increase in tuning capacitor values and the higher output current buffer amplifier, designed around an Elantec EL2099C integrated amplifier.



Capacitor Tests

Having tested one capacitor of a make and type, what guarantee does this give about harmonic distortions generated by other similar capacitors in the same batch? In my view that depends totally on the method of manufacture and the particular dielectric used. For the audio perfectionist however, perhaps every signal path capacitor should be distortion measured.

For example, COG ceramic is probably the most stable and most nearly perfect of all commonly used dielectrics. COG disc and multi-layer ceramic capacitors do not rely on pressure contacts or metal spray connections onto their electrodes. One maker's products should measure consistently and with remarkably low distortion. Those from a different maker may measure slightly differently, but again should be consistent from batch to batch.

Polystyrene is another of the best performing capacitor materials. Capacitors made using the extended foil technique and with their lead out wires soldered directly onto the extended foil electrodes, should be consistently nearly perfect. Distortions in capacitors made using metal spray end contacts to their metallised film dielectric electrodes, for any one film type, will vary more from maker to maker. Worse still, from my measurements, they can also differ considerably even within a small capacitor batch.

Some film capacitor makers however do seem remarkably consistent within a batch and from batch to batch. With other makers I have measured some 20-30dB different harmonic levels, in quite small batches, even when the capacitors have been supplied taped to card strips.

Having provided a usable, repeatable test method and easily assembled, low cost test equipment, my next articles will explore which capacitor types produce the least harmonic distortion, according to capacitance value. When possible I shall try to explain how different capacitor constructions can affect the harmonic distortion generated in the capacitor.

With so many capacitor suppliers available, I cannot provide a best buy list. This measurement hardware, which allows repeatable capacitor distortion tests, I feel should be more than sufficient.

My next article will discuss capacitors having values up to 10nF and soundcard FFT measurement software available on Internet. ■

Technical Support

Professionally produced printed circuit boards for the 1kHz low distortion signal generator, the 1kHz low output impedance buffer amplifier/notch filter/pre-amplifier boards and the 1kHz DC bias buffer will be available.

Full details of price and availability will be provided in my next article of this series, which will also include details of my DC bias buffer circuit and PCB.

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Spectrum pricing's uncertain future

Since 1998 spectrum pricing has been applied successfully to many allocations to private operators, but continuing failure to understand its underlying principles has marred its future prospects.

My article, 'Spectrum pricing comes of age' in the July issue, broke off with the UK Department of Trade and Industry (DTI) white paper on spectrum management in 1996. That white paper professed to recognise that the prices charged to spectrum users should not be set to maximise the revenue, but should reflect the value of the spectrum and it proposed initial tariffs for the first affected services in the private sector. But the preceding consultations with 400 spectrum users over two years had exposed a conflict between 'administrative pricing' and auctions.

The UK government preferred auctions for users in the private sector on the supposed grounds of economic efficiency, transparency and speed, but the responders to the consultations evinced a marked preference for 'administrative pricing'. They were the people who had experienced the technical problems of spectrum management at first hand and would have to cope with the new regime and they preferred the prices to be set 'administratively', which meant by the Radiocommunications Agency (RA, the new name for the old RRD), rather than by a newly invented, untried process. Perhaps those users preferred the devil they knew.

That white paper also announced that defence and the emergency services should have the same incentives for spectrum efficiency as the private sector and would pay charges "on a comparable basis", but it did not explain how that basis would be calculated. I reviewed that white paper for Electronics World in September and October 1996 [1].

It was of course gratifying to learn that my proposals for dealing with what had become a multi-billion pound national problem, which had been discarded 13 years earlier, looked like reaching the statute book at last. But there followed a general election in 1997 and the new government evidently regarded the radio spectrum as a legitimate source of revenue. It was therefore to be expected that it would probably try to maximise that revenue, as I predicted in Chapter 8, 'Technology and Politics', of my book, 'Prosperity from Technology', published in 1999[2].

Following that white paper, the Wireless Telegraphy Act 1998 duly gave the RA wide powers to introduce spectrum pricing and they forged ahead with administrative pricing in the ensuing four years - first for mobile networks, then point-to-point fixed links, embracing the whole private business radio (PBR) sector in July 2000 and extending to earth satellite stations a year later[3]. They repeated the commitment to charging the armed forces and emergency services on a comparable basis, but explained that "the details are under negotiation with the departments concerned." In other words, those departments were (and probably still are) dragging their feet and so increasing the costs of licences unnecessarily in the private sector.

Where angels feared to tread

Pursuing that DTI preference for auctions over administrative pricing, the RA sold five licences for third-generation (3G) mobile phones in a controversial, high-profile auction in 2000. It raised £22.5 billion for the Treasury (10 times the fee per MHz charged to the TV broadcasters and 200 times the fee per MHz for PBR). The sale was subsequently staunchly defended by the National Audit Office (NAO)[4]. The stated objectives were optimum efficiency of spectrum utilisation, promotion of competition and realisation of the full value of the spectrum, but those are weasel words. Optimum utilisation efficiency often differs substantially from optimum economy and the meaning of 'realisation of full value' is at best ambiguous.

The essence of spectrum pricing as the means of making the best practicable use of a physically scarce resource is that the users should be induced to compare the charges for spectrum and their other costs of using spectrum with their benefits of using it and consider their options. More efficient utilisation would be one way of using less spectrum but it might be better to move to a less congested band or revert to land lines. Different users and potential users would come to different conclusions. Some competition would ensue but competition is not the object and should not be artificially stimulated. Charging the users more than the minimum necessary to prevent overcrowding undermines the essential principle of spectrum pricing.

On the other hand 'realising the full value' in an auction, if it means anything, means that the greatest possible revenue is to be realised. It did not seem to occur to the organisers of the auction, nor to the NAO, that the much higher prices per MHz obtained for the band allocated to 3G licences was an indication that perhaps that band should be somewhat wider and the prices correspondingly lower, with the bands on either side somewhat narrower and the prices higher. The advantage to the 3G licensees would outweigh the disadvantage to the users in the neighbouring bands.

The NAO repeatedly denied that the object had been to maximise the proceeds for the Treasury, but the organisation of the auction could hardly have been better crafted if that had been the avowed intent. The RA reserved one licence for a new operator (denying it to any of the four incumbents - Vodaphone, BT Cellnet (now O2), Orange and One 2 One (now T-Mobile) in order to stimulate competition. They then employed N. M. Rothschild as financial advisers and paid them a success fee of £700,000 for persuading nine new bidders to participate - at their great expense.

The bidders were required to put in simultaneous sealed bids with £50-400 million bidding bonds in a succession of bidding rounds. The highest bidder for each licence in each round was excluded from the next round for that licence but could resume bidding in subsequent rounds. The bidding continued in that way for seven weeks until all but the four incumbent operators and one new bidder had dropped out voluntarily. In total there were nearly 500 bids of which more than half were by the eight losing bidders. They were left without licences and had to recoup their bidding costs, including those expensive bid bonds, from their other operations.

That organisation was devised explicitly to simulate auctions of other goods and services. But auctions (e.g. for pictures, houses and second-hand furniture) have evolved and been honed over centuries precisely to maximise the proceeds, beginning with maximum publicity. If there had to be an auction as a matter of policy there was at least no need to drum up the maximum interest in advance nor to conduct it during a surge of euphoria in which the participants were speculating about highly uncertain future prospects. But their inherent tendency to maximise the proceeds is the principal objection to having auctions at all. And it was a distortion of spectrum pricing to try and use it to force the four incumbent operators to accept a new entrant against the advice of the Universal Mobile Telecommunications Service, the main body representing the telecommunications industry.

In fact, the whole operation was a very risky political intervention in the radio communication industry which may yet prove to have damaged it irrevocably. The NAO let the cat out of the bag by reporting that "The timing of the auction was conducive to maximising the proceeds". and now the Financial Times newspaper has reported on a survey in June this year, revealing that "more than four in 10 of the population have no interest in third-generation (3G) services . . . [which] will be a serious blow to mobile operators that spent billions on third-generation technology and need a rapid return on their investment"[5]. Thus a possible disaster has come a step nearer.

A persistent blind faith in deregulation

On 6th March of this year the Department of Trade and Industry and HM Treasury published the final report of a 'Review of Radio Spectrum Management' by Professor Martin Cave. It is concerned with spectrum valuation, pricing and trading as tools for allocating and assigning scarce radio spectrum across virtually the whole range of frequencies from 88MHz to 60GHz. It runs to 260 A4 pages and makes 47 recommendations.

That review was preceded by a consultation which elicited 80 written responses, including one from myself, but it is impossible to say how much attention, if any, the review team paid to them individually. There are no explicit counter-arguments to any of the responses in the final report.

Most of the report's 47 recommendations are pretty bland, but the report remains wedded to the economists' pathetic faith in deregulation and competition as the only effective way of preventing private operators from exploiting their customers' dependence on their services for the benefit of their shareholders. And it advocates extensive trading in spectrum between private operators with scant regard for the practical problems of preventing interference, less in fact than was evinced in the DTI economists' paper in the Merriman report 20 years ago.

In reality the mere existence of four or five operators cannot ensure effective competition beyond the short term. They are an oligopoly and even before the FT survey they expected a process of consolidation in a few years time when the euphoria for high-technology industries in general and mobile telecommunications in particular has evaporated, as even the NAO recognised. The forces, familiar to engineers, that will impel such consolidation are the economies of eliminating or reducing the duplication (or now quintuplication) of the expensive transmitting stations that are springing up all over the country. There will have to be regulation to prevent exploitation and I proposed a new approach to that subject in 1992[6] but, like spectrum renting/pricing nine years earlier, it was discarded.

Government horizons are inevitably restricted to a few years, but one might hope their independent advisors would recognise that competition cannot be a panacea for all industrial problems in the long term. They might also admit that there is little, if any, difference in practice between realising the full value of spectrum and maximising the proceeds of selling or leasing it, and that beneficial trading in it depends on all the traders thoroughly understanding the technical complexities. Acquiring that necessary expertise would be very expensive and experience has shown that the traders in any competitive industrial field will not go to any expense they hope they can avoid.

For the public services the report recommends a policy of reserving specified bands of spectrum for such purposes and basing the charges on "the value of that spectrum to users". But it again ducks the problem of how to calculate that value and obscures the anomaly of public and private users being charged at greatly differing rates in adjacent bands. Because the spectrum is physically continuous it follows that ideally the rate per MHz should not change substantially at any such boundary. If at a boundary between such bands the lower rate is high enough, the higher rate in the adjacent band must be too high. In practice such discrepancies will inevitably continue for many years, but meanwhile they should be valuable indicators of how much both the charges for the public services and the sizes of their allocations should be gradually adjusted as and when the opportunities to do so occur.

The report preaches again that spectrum pricing should not be a device for raising government revenue, but it does not criticise the auction of third-generation mobile phone licences in 2000 for becoming just such a device. Moreover it professes to be "concerned that the current regime, in which inflexible allocations to particular purposes, generate artificial scarcities which . . . ultimately lead to higher prices paid by consumers", without apparently realising that its recommended policy of reserving specified bands of spectrum for specified public services is an example of just such inflexibility.

As with many applications of technology, spectrum management is not a matter of common sense, nor of applying economic theories without fully understanding the practical problems of the technology itself There may not be many people in Britain who are capable of understanding both the technology and the commercial economics of spectrum pricing in the necessary detail (other countries may be better placed). But there are some and they will have to be found and put in charge if we are to get the full benefit of radio communications in the long term. ■

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TYING THE KNOT

The notion of using some kind of 'servo feedback' to control speaker cone movement occurred to engineers a very long time ago. The first references that I can find date from the early 1950's and in fact Philips produced some domestic models in the 80's.

Back in those Halcyon days ideas that we now take for granted, servo theory and negative feedback for example, were only just being widely disseminated. Hi-Fi was very much an enthusiast's activity, especially in this country. Nevertheless, several references can be found for increasing speaker 'damping' in 'The Radio Designers Handbook'[1]. Before discussing these ideas in some detail, it would be as well to define just what motional feedback is and how it can be employed to improve loudspeaker response.

The main problem in speaker design is that of obtaining a flat frequency response. A typical loudspeaker has the response of a band-pass filter with a load of ad hoc resonances thrown in for good measure. Even when carefully designed, a tolerance of ± 3db within the driver's pass-band is considered good. Any power amplifier with a similar response variation would be discarded as seriously substandard! The nub of the problem is that a loudspeaker is a mechanical device and that it is passively driven. The only sensible way of linearising the response would be to actively control the cone's motion. This, however, is beset with further problems. Many forms of motional feedback control have been proposed and tried. All of them are expensive to implement, especially in the context of an existing audio system. Still the prize of a truly flat speaker response is sufficiently alluring to make the effort worthwhile.

Measuring the position of a moving cone is surprisingly difficult and implies the use of some kind of sensor. Among the methods that have been applied include piezo-electric accelerometers, dual voice coils, photo-electric and capacitive proximity detectors and measuring the acoustic output directly with a microphone. All of these methods also require signal-processing circuitry. Furthermore, with the possible exception of using a microphone, these methods are only useful when the driver is operating as a piston. With a typical 8" speaker this region extends only up to 1-2kHz. At higher frequencies the phenomenon of cone break-

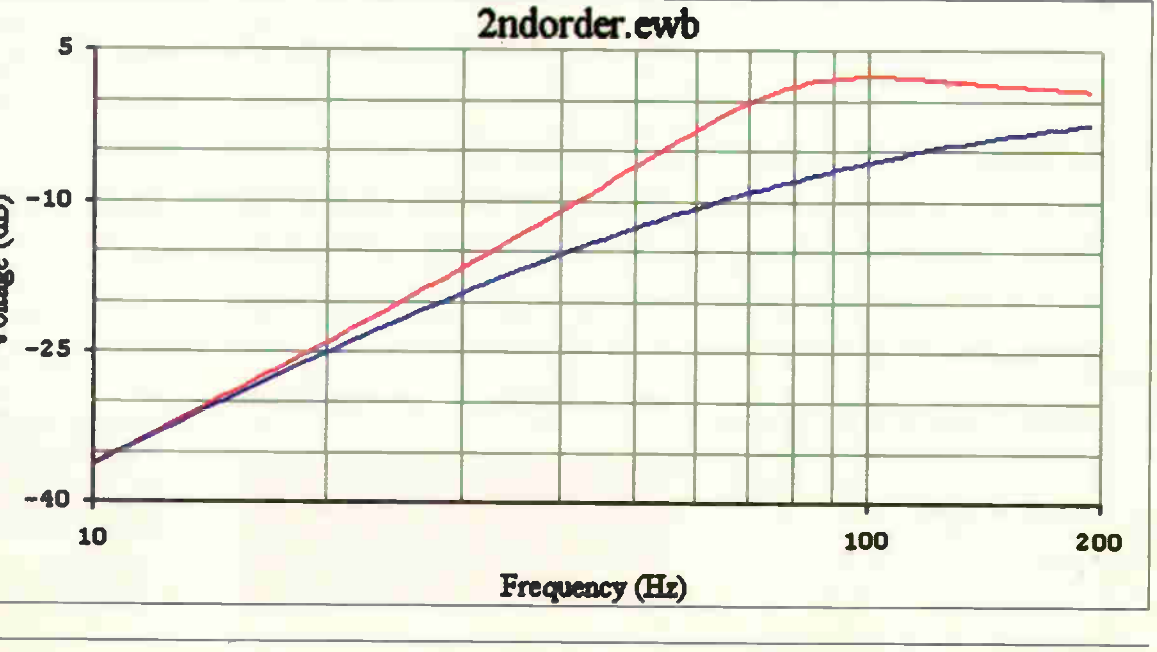


Fig. 1: The effect of driving an experimental IB enclosure

up occurs. In this mode different parts of the cone move at different velocities and the servo's signal will become unrepresentative of the driver's output.

There is one area in which motional feedback can be employed to advantage - namely the improvement of bass response, and this can be done without the use of expensive transducers. The basis of the idea is that a speaker system is itself a transducer and as such it is a two-way device. Apply a current and the cone moves. Apply a force to deflect the cone and a back emf is generated which is proportional to the cone's velocity. All that is required is to extract the back emf, process it and apply it as a feedback signal in the power amp stage. The purpose of this article is to tie the knot between this form of motional feedback and Thielle/Small (T/S) theory.

At first sight motional feedback speaker systems seem to fly in the face of existing T/S theory. The usual method of designing a speaker system is to take a suitable driver and calculate the enclosure size for a flat response. Because of the interactions between the cabinet and speaker parameters this is not an easy task and the resulting system is usually a compromise between domestic acceptability and performance. However, T/S theory can tell all you need to know about any conceivable speaker/enclosure combination. The difference between a motional feedback system and a normal system is that the open loop gain of the amplifier is used to straighten the response at the bass end, in the same way that a negative feedback circuit uses open loop gain to reduce its non-linearity. In the design of a motional feedback speaker system the enclosure has to be chosen so that the driver cannot self-destruct due to large bass excursions. The resulting system non-linearity is taken care of by the electronics.

If you measure the impedance curve of a driver around the bass resonant frequency you will observe that the impedance is that of a tuned circuit peaking at resonance. The rise in impedance is entirely due to the back emf generated by the driver in response to the mechanical resonance between the compliance of the surround and the cone mass. This rise in impedance can be modelled by an equivalent electronic circuit (see box). In order to suppress the rise in acoustic output around resonance, power amplifiers are designed to have zero output impedance. Unfortunately this doesn't work very well because the speaker's coil resistance is in series with the reactive components. If the output resistance of the power amp were made negative and of the same value as the voice coil resistance then the resonance could be perfectly damped by the amplifier. The result is a response that rises at 6db/octave. This is because the drivers' output would be made directly proportional to cone velocity. In practice this is difficult to do since the adjustment is critical and the voice coil impedance is temperature sensitive.

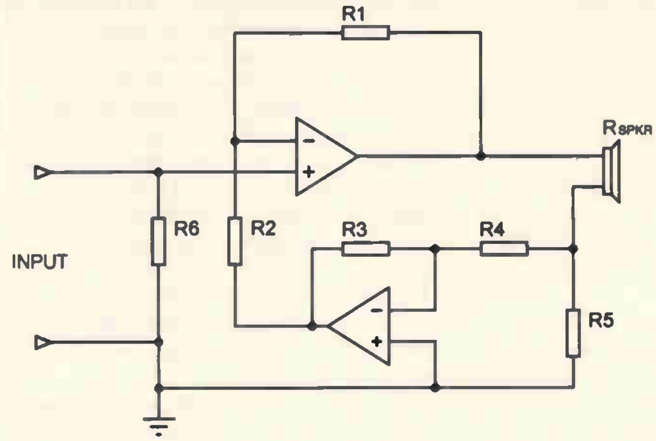


Fig. 2: The modern equivalent to Clements' PF circuit in outline

However by adjusting the value of the output impedance the Q of the speaker can be controlled and this relates directly to standard T/S theory. Furthermore the feedback voltage is derived directly from the driver's back emf and so qualifies for the term 'Motional Feedback'.

In order to understand the technique we must consider the three main T/S parameters that define the behaviour of a driver at resonance. These are Vas, the volume of air that has the same compliance as the speaker surround, Fs, the free air bass resonant frequency of the driver and Qts the 'Q' of the bass resonance. Qts is essentially the factor that can be varied by the use of negative output impedance.

The definition of Qts includes the output impedance of the drive source the equation being,

Qts = ((Qes.Qms)/(Qes + Qms)),

Where Qes is the electrical Q and Qms is the mechanical Q of the resonance.

Qes = k ((Re + Rg)

Where Re is the voice coil resistance and Rg the generator impedance. As previously discussed the output impedance of a power amp is designed to be zero and so Rg is zero. By altering the value of the output impedance Qes, Qts can be varied over a wide range. The relevance of this is that the flat response enclosure size for both sealed and vented speaker enclosures depend critically on the value of Qts. So being able to vary this parameter means that making that ideal system becomes a lot easier!

Varying the Q of the driver's resonance is obviously useful, but to make a viable feedback speaker system requires that some other factors are taken into account. The rest of this article is going to concern itself with the design of sealed enclosures, the so-called

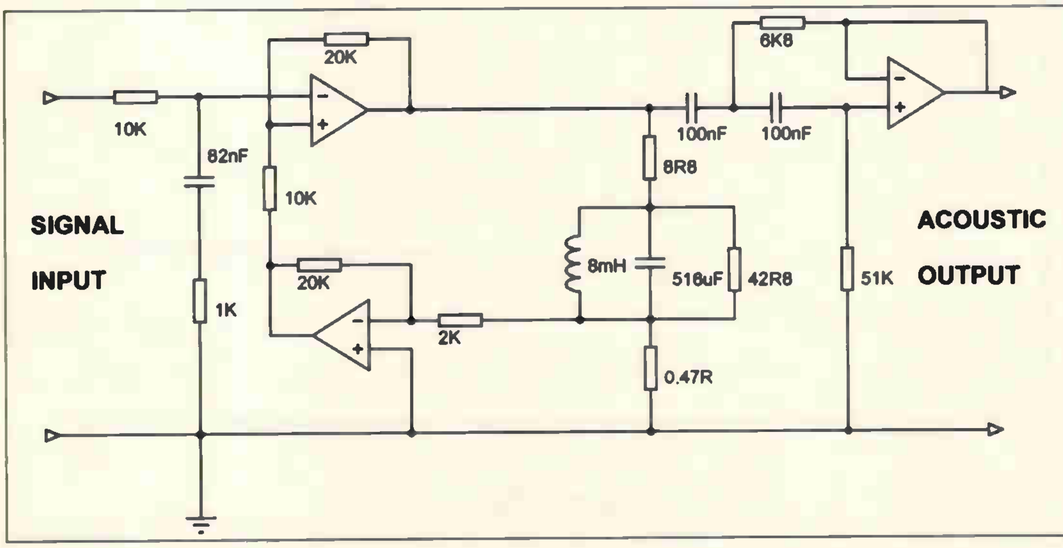


Fig. 3: The PF circuit with EQ

infinite baffle or IB type. The major advantage of the IB is that you only have to worry about one resonance and that the rate of bass rolloff is limited to 12db/octave. Fig. 1 shows the effect of driving an experimental IB enclosure. The upper trace shows the response obtained when driven from a normal amplifier, zero output impedance. The lower trace shows the response obtained from the same enclosure when driven from an amplifier modified to give an output impedance of 3.5R.

As you can see the resonant peak is far better damped. The enclosure Q has been lowered from 1.2 to 0.4. The importance of having a well damped <0.5 Q response is twofold. Firstly, the original Q of 1.2 response had a lousy transient response and would overshoot when transient inputs were applied. Secondly, equalisation to a flat response would be difficult. With the Q at 0.5 or less, transients don't overshoot and the response rolls off initially at a sedate 6db/octave making equalisation a doddle, relatively speaking.

It's also easy to see that, with the IB, the amount of bass that you can generate is limited to the volume of air that the driver can move. That is to say the product of the moving cone area and the peak excursion that the speaker can provide. Although you can simply pick a box size at random and fit the driver in, the results are unlikely to be as desired. To get good results some design work needs to be done. When trying to obtain the maximum bass from an enclosure, attention must be paid to ensuring that the driver doesn't self-destruct! The reason is that for a flat bass response the cone excursion will increase fourfold every time you half the input frequency. Without correct acoustic loading, the driver's excursion limit will be exceeded and the driver's coil will become de-centred or worse.

Luckily, the air inside an enclosure acts as a mechanical load to the driver that will limit excursion to tolerable levels providing the correct volume has been chosen. Below the driver's resonant frequency, the response will drop off at 12db/octave for a given electrical power input, the same, luckily, as the bass roll off of an IB enclosure. On the other hand, if the cabinet is too small, the full acoustic output will not be reached. In order to find the right enclosure volume you first need to know the electrical power rating, peak excursion and the radiating diameter, D. This can be determined by measuring the diameter of the driver plus half the surround. The last factor required is the sensitivity of the driver expressed as so many db/W @ 1m. All the T/S parameters should be available in the driver data sheet.

With this information the enclosure can be designed from the following equations. The proof of which was given in Russel Bredon's 'Roaring Subwoofer' article.[2]

$$Mo = SPL + 10Log(P) \text{ db/W/m}$$

Where Mo is the maximum acoustic output in the driver's pass band, SPL is the sensitivity of the driver in db/W/m and P is the power rating of the amplifier used in Watts RMS.

$$M1 = 40Log(D) + 20Log(App) - 83 \text{ db/W/m}$$

where M1 is the maximum acoustic output at 1Hz, into half space, D is the effective diameter of the driver and App the peak to peak cone excursion expressed in mm. The resonant frequency of the driver in the enclosure, Fc can then be determined from:

$Fc = 10\text{^}((Mo - M1)/40)$.Qts, the Q of the driver enclosure volume can be determined by :

$$Qtc = Qts(Fc / Fs)$$

Finally the enclosure volume, Vb, can then be determined from:

$$Vb = Vas / ((Fc / Fs)^2 - 1)$$

Having found our optimum Vb the frequency response will be far from flat. This is where the motional feedback techniques come in. A driver, mounted in a sealed box, will have a frequency response exactly the same as a 2nd order high pass electronic filter of the same Fc and Q. The classic 2nd order high pass can have a variety of response shapes dependent only on the Q of the filter. As previously discussed the Qtc, the 'Q' of the speaker system response can be directly varied by manipulating the output impedance of the amplifier.

In his 1952 articles[3,4] Warner Clements outlined a simple technique to alter the output impedance of an amplifier to a negative value. To understand the technique it must be realised that even an infinite amount of negative feedback applied to a power amplifier circuit can only reduce the output impedance to zero. To get a negative output impedance, positive feedback must be used. At first sight the idea of positive feedback in an audio circuit is anathema. However it has a long history. For example, many valve amp designs were improved by the judicious use of positive feedback (PF). The idea was to increase the gain of the front-end valve, which typically had little THD by using PF. This improved the open loop gain of the amplifier and thus improved the response and THD when global feedback was applied.

The modern equivalent circuit to Clements' is shown in outline in **Fig. 2.** This shows a conventional power amp, A1, with a feedback loop consisting of R1 and R2. The positive feedback is applied via the shunt amplifier A2. A small resistor, R5 is connected in series with the speaker and monitors the current flow through it. The voltage drop across this resistor is thus proportional to the current flowing through the speaker. As the impedance rises this voltage will decrease proportionally. Normally, feedback from this point would be negative. The inverting amplifier produces

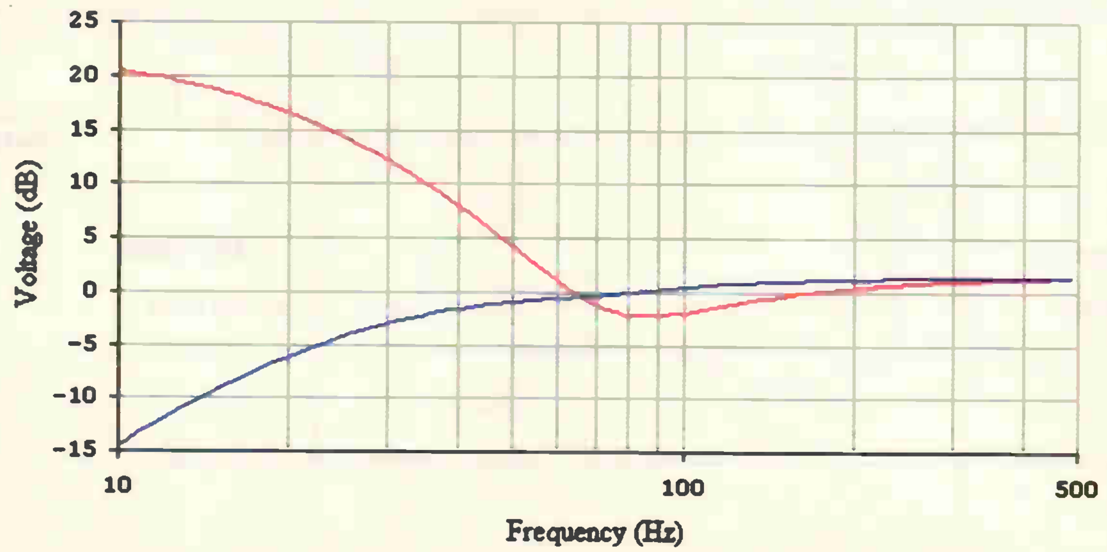
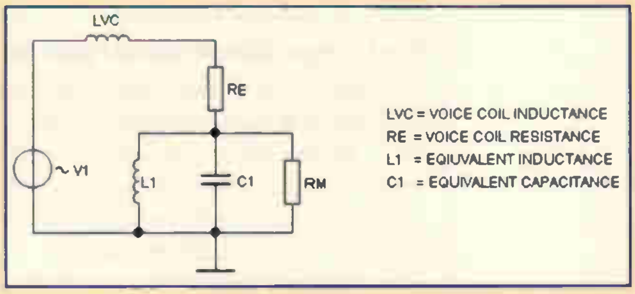
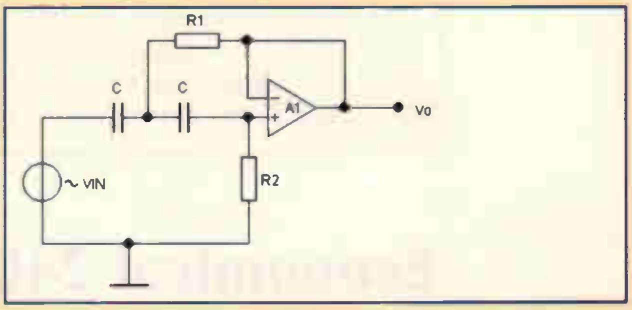


Fig. 4: Acoustic response of the entire system

The rise in speaker impedance around resonance is entirely due to the excess motion of the cone. It can, however, be modelled by an equivalent tuned circuit as shown right. The parameters can be gleaned from the information and T/S parameters supplied with the driver. Otherwise the parameters can be measured directly[5]. The equivalent circuit will exhibit the same impedance curve as the driver mounted in a sealed box. Note that although the parameters quoted here are measured in an IB enclosure the same arguments apply to measurements taken unmounted in free air.



RE is the voice coil resistance.
RM is the maximum impedance at resonance, RM = RE [(QMC / QEC) + 1]
FS is the resonant frequency.
C1 = 106 [QTC / (2 P FC)] / [(RE RM) / (RE + RM)] value is in µF.
L1 = $(2\pi FC)^{-2}$/C1, C1 EXPRESSED IN FARADS



The 2nd order high pass filter has a frequency response identical to the bass response of a speaker of the same F and Q. As such it is useful as a method of simulating the acoustic response of the speaker when driven by a power amplifier. To do this the components must be selected as follows.
Choose a convenient value for C. 100nf, say. Then fo =FC and Q=QTC. Let D = 1/Q, then,
Rn = 159155 / fo / C, where C is expressed in µF and R1 = D / 2 * Rn and R2 = 2 / D * Rn.

the 180∞ phase shift required to produce positive feedback and the output operates as an earth return for R2. By inspection you should be able to see that as speaker impedance rises the positive feedback will fall, reducing the amplifier's gain and damping out the bass resonant peak.

In order for the circuit to work as advertised R5 should be << than the speaker impedance and 0.47R is an acceptable value. The negative output impedance obtained can be calculated quite easily.

Rout = - R5((R1.R3) / (R2.R4)) Ohms.

Note the negative sign. The output impedance is always negative. When fed in this way the system Q will be reduced according to the formulas presented. As with any electronic system there are limits that should be observed. For example, the extra circuitry can be added to an existing amplifier without fear of oscillation because the feedback is positive. The op-amp is not critical; a TL072 is fine, and commendably cheap. Knowing the values of R1 and R2 will allow a large range of negative output impedance to be generated. To avoid problems the op-amp should be operated from a pair of voltage regulators.

From the driver point of view it should be remembered that voice coil resistance is temperature sensitive. This means that if you try to remove all the impedance you are likely to experience oscillation. The maximum amount of damping that you can apply is probably limited to about 80% of the voice coil resistance that is more than enough to transform the bass response of an enclosure.

So far, I have discussed the basic circuit. This can be extended to model actual speaker systems. The technique is to replace Rspeaker with the tuned circuit network derived from the in box measurements of Fc, Qmc and Qec. The output from the circuit is fed through a second order filter with fo = fc and Q =Qtc. As you will recall such a filter will have the same response as the T/S speaker response. The output of the filter will be the same as the acoustic response of the speaker system. This allows multiple 'what if' acoustic simulations to be made of the system's acoustic response via a suitable Pspice package.

To render the bass response flat, some external equalisation is required. Look at **Fig. 3**, that shows a simulation circuit of the type just discussed. Using the ubiquitous 'ideal' op-amps on Electronic Workbench to simulate an actual system. Referring to this R1, R2 and C1 provide passive bass boost between 20Hz to 200Hz. Components Re, L1, C and RM simulate the speaker's impedance curve. A2 in conjunction with R6 and R7 provide a positive feedback signal that is injected into the main amplifier feedback loop comprising R3 and R4.

This particular system has a Fc of 75Hz and Qtc of 1.2. The acoustic output of the driver and amplifier system is produced by feeding the output of the amplifier into the 2nd order Sallen and Key High pass filter built around A3. This filter has the same characteristics of the driver and box combination. Examining its output will therefore reveal the acoustic response of the entire system. **Fig. 4.** shows the result. The red trace shows the output from the amplifier, this being the sum of the input filter and the damping applied to the driver. The bottom curve shows the acoustic response of the system. As you can see the response is flat down to low audio frequencies being only 6db down at 20Hz. In part 2, I will describe a system for DIYers with an extended bass response and domestically acceptable size based on these principles. A kit will also be available once the second part has been published.

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*Reprinted in the 'Audio Anthology' Vol 2, Available from 'Old Colony Sound Lab, P.O.BOX 243, Peterborough, NH 03458, USA.

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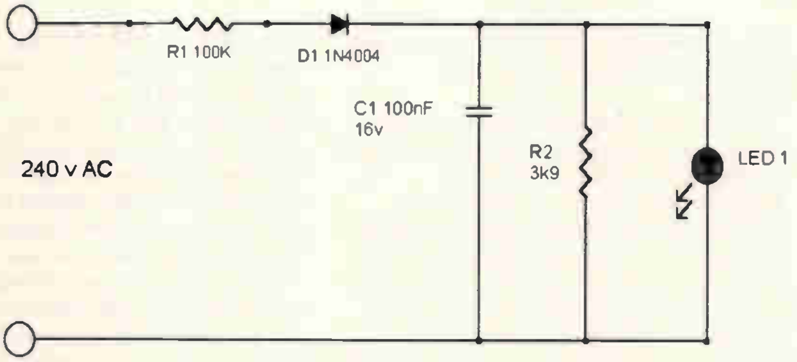
Economical 240V AC mains power LED

With the availability of high efficiency LEDs, it is now possible to power LEDs directly from the 240v AC mains. Normally this was possible through the use of a mains power capacitor or a mains transformer. In those cases the costs are higher and they would be not be feasible should space becomes an issue. All the resistors used in this circuit are 0.25W.



Overall, the space utilised by the components is very tiny in this design. The LED used should be around 3000mcd capability.

The above circuit was able to power a BLUE LED with maximum brightness. In the case of lower efficiency LEDs, the value of R1 should be reduced to 47K and R2 to 2K2.

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Two channels – one gain pot

Reading the February 2002 issue, I was interested in the article 'Two channels, one gain pot'. I recall that analogue multipliers have been used to solve the same problem.

The circuit shown uses LM13600s but it could also be implemented with an LM13700. It allows you to drive two or more voltage-dependant preamplifiers with only one potentiometer. Gain goes from 0 to 900 when the voltage is varied from –15 volts to +15 volts. I think this idea was originally from a National Semiconductor application note.

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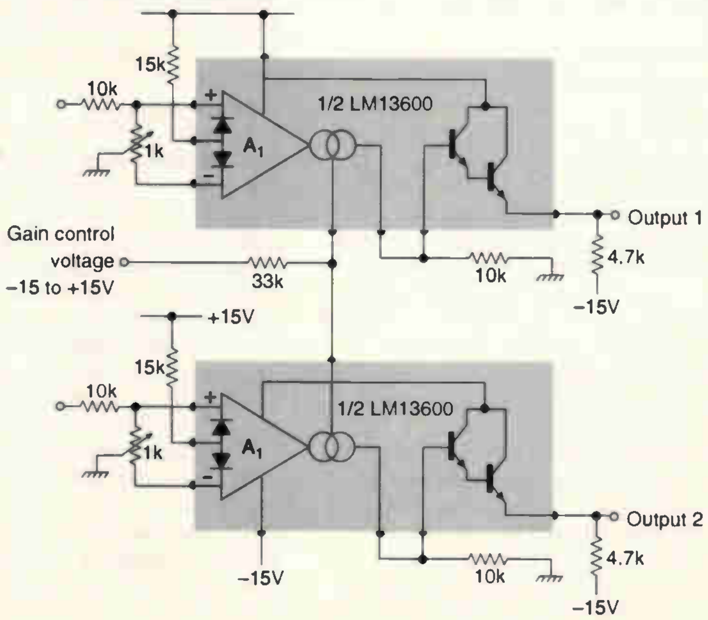
Saint-Laurent-du-Var
France



An alternative solution for controlling two audio channels using one potentiometer.

Valve amp turn-on delay adds automatic protection

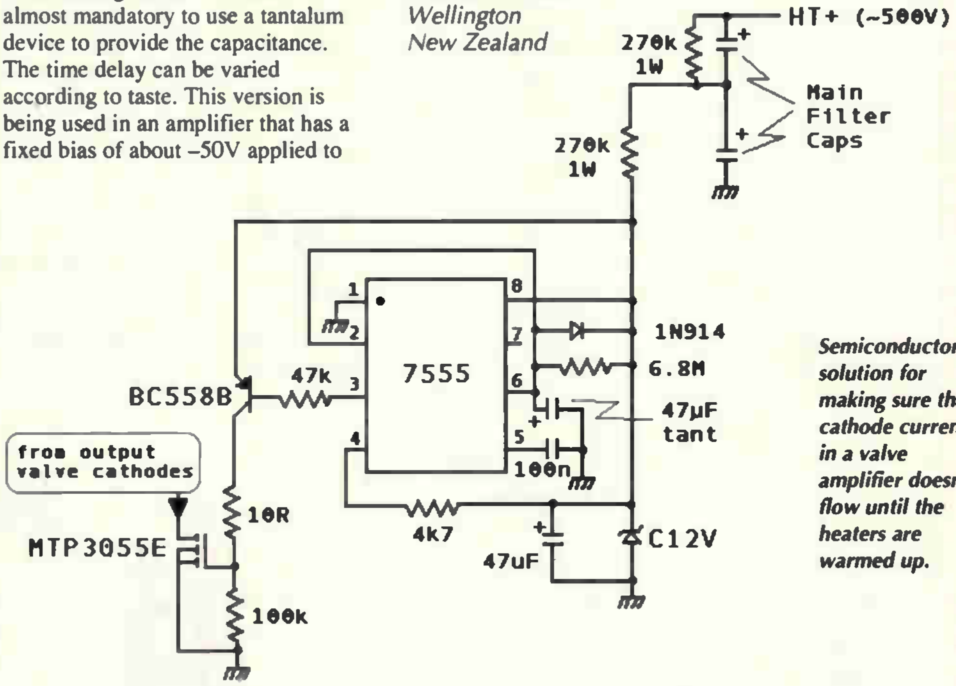
While building a valve stereo amplifier, I wanted to prevent cathode current being drawn in the output valves until the heaters had reached a suitable operating temperature. It is standard practice to include a 'Standby' switch in guitar amplifiers, but the idea presented here was developed to remove the need for a separate switch.

It is not new to allow the cathodes of the output valves to float, rather than delaying the application of HT voltages, to prevent current being drawn while the valves are warming up. The circuit shown uses a CMOS 555 with an *RC* time constant and power MOSFET to perform the function.

In amplifiers with high HT voltages, it is common practice to include voltage equalising resistors across the series filter capacitors. This design makes use of the bleed current for its power source by drawing power from the lower bleed resistor, instead of connecting it to ground. While there is now a difference of 12V across the upper and lower filter capacitors, this is deemed quite tolerable.

Given the value of resistance used in the timing circuit – 6.8MΩ – it is almost mandatory to use a tantalum device to provide the capacitance. The time delay can be varied according to taste. This version is being used in an amplifier that has a fixed bias of about –50V applied to the output valve grids but it can also be used when cathode resistor/s are used for biasing.

Malcolm Watts
Wellington
New Zealand



Semiconductor solution for making sure that cathode current in a valve amplifier doesn't flow until the heaters are warmed up.

Self-decoding one-of-four indicator

I was using a two-bit binary ripple counter to electronically select one of four signal sources and I wanted a visual indication of the active source. Instead of adding an extra chip – gates, decoder etc. – to decode and drive the LEDs, I made use of the inverting and non-inverting outputs of the bistable devices to do the decoding, resulting in the circuit shown.

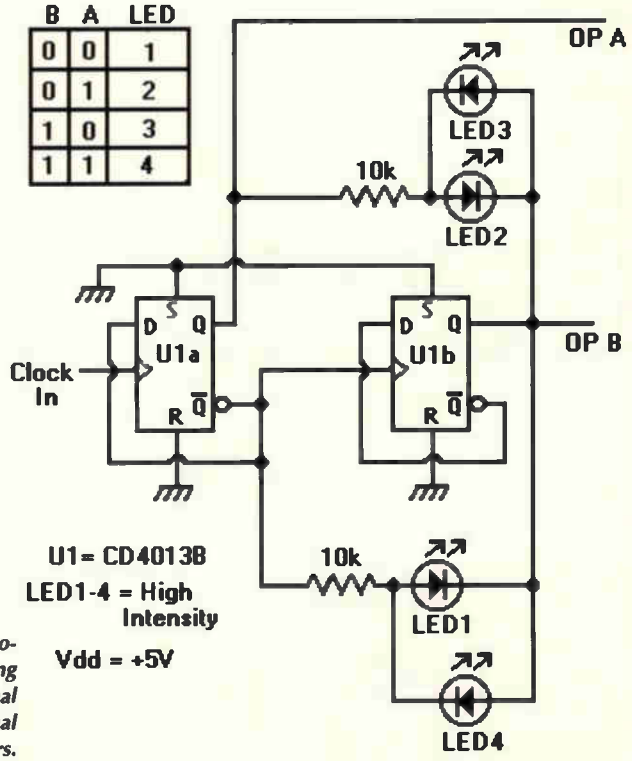
A switch-activated monostable – not shown – toggles the clock input of the first bistable device.

The LEDs are high-brightness types with clear lenses. A useful light source can result from a LED current of less than 500µA. Here, the series resistors restrict bistable-device output currents to somewhat less than this current to make sure that logic high and low levels on the output pins are not affected.

Although this circuit may be implemented in several ways, designing for back-to-back connection of the LEDs ensures that reverse-bias ratings are not exceeded. Higher current output capabilities of HCMOS devices, like the 74HC74, would allow higher LED currents to be used without logic level specifications being violated.

Malcolm Watts
Wellington
New Zealand

B	A	LED
0	0	1
0	1	2
1	0	3
1	1	4



Clocked by a signal from a monostable device, this circuit produces a two-bit binary signal for selecting one of four signal sources. Normally, adding four LEDs to indicate which source was selected would mean additional two-to-four-line decoding. The scheme shown avoids any additional components apart from current-limiting resistors.

Water level indicator with audible high and low warnings

£50 winner

A versatile, effective and yet simple and low-cost water level indicator is described here.

This circuit represents different water levels in meaningful English characters. It has the added advantage of giving an audible warning at the lowest water level. When the water reaches the uppermost level, the electric motor that pumps the fluid is automatically stopped.

The circuit is also flexible in that it can even be made to produce an audible alarm at both the lowest and highest levels.

At the heart of this circuit is programmable array logic device, which is programmable and can be programmed using PALASM assembler. Alternatively you could use a GAL16V8.

The INDICAT.PDS file is the firmware for the PAL. Here a PAL16V8 is used as PAL16L8, i.e. its output will be active low.

Connection to the PAL device can be made in two different ways, represented in **Figs 1** and **2** respectively. In Fig. 1, the circuit beeps continuously, while when the connections are made as in Fig. 2, the device beeps in pulses whose pitch can be controlled using the potentiometer.

The circuit displays five different characters on seven-segment display corresponding to five different levels. They are F for full, A for above half full, H for half, L for low and E for empty.

The electric motor is stops automatically when the water reaches the highest-level only if the motor-enable switch, labelled 'MES', is closed. The circuit connects to the motor via a normally-closed relay. The ratings of this relay will depend on the motor used.

I have tested the circuit on a Vinytics PAL kit. You can request the .PDS file in electronic form by e-mailing me at tejinder_marwah@hotmail.com

Tejinder Singh
New Delhi
India

E-mail J.Lowe at highburybiz.com for a copy of the firm ware file

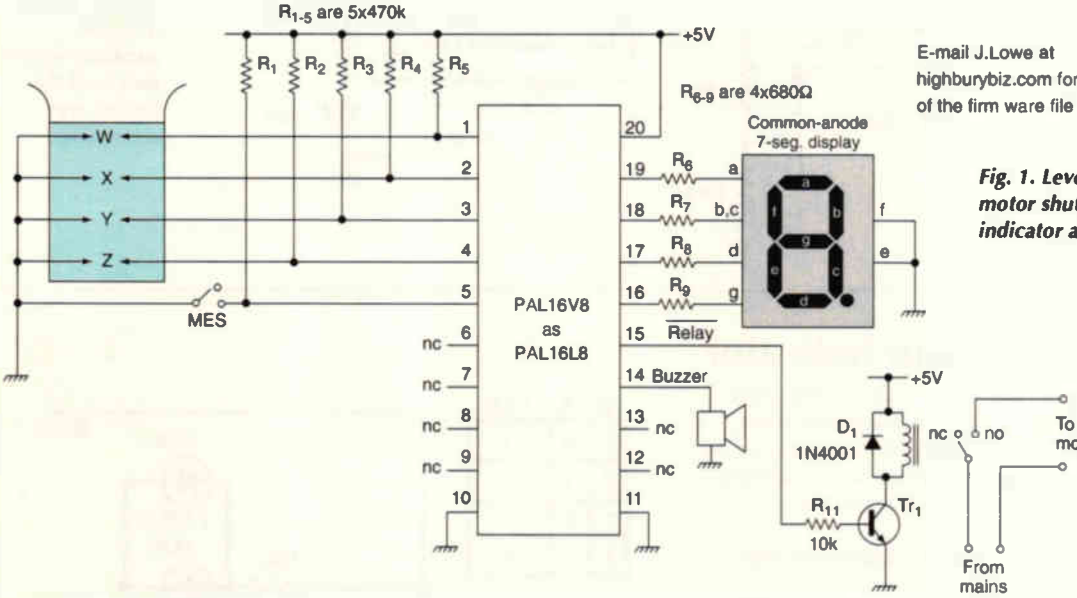


Fig. 1. Level indicator featuring motor shut-off, five-level indicator and 'level-low' alarm.

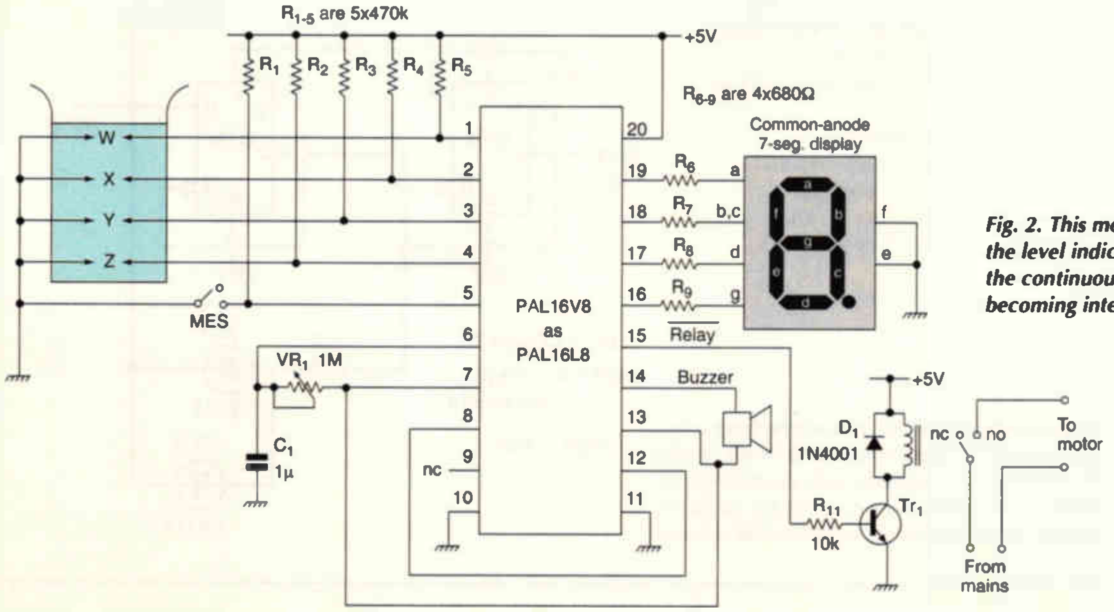


Fig. 2. This modification to the level indicator results in the continuous alarm beep becoming intermittent.

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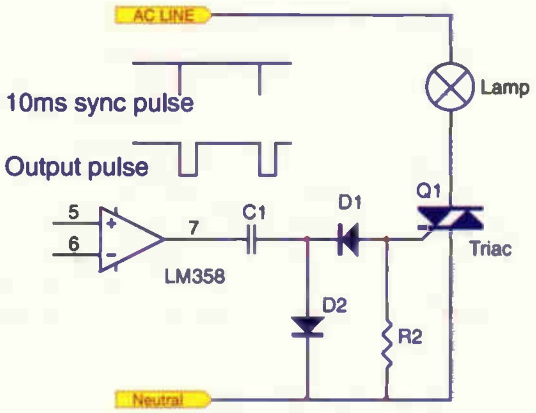
Triac drive features AC coupling

This circuit makes use of the AC synchronous pulses and a DC voltage to control the delay of an AC phase-control circuit. As there is no floating ground in the circuit and the ground reference is linked to the neutral line, the potential for getting an electric shock from the circuit is greatly reduced.

Fig. 1. AC-coupled triac driver circuit with associated waveforms.



Typical triac triggering circuits operate at quadrants I and IV. They normally use an *RC* delay circuit with a bilateral triggering device such as a diac. Unfortunately, the triggering currents in quadrants I and IV are not even. The driving circuit would usually need a high margin to avoid missing the odd pulse.

Triggering currents in quadrants II and III on the other hand are relatively the equal. In this case, only a negative drive of similar magnitude would be needed. But there is a drawback here in that it is necessary to drive the gate negatively, that is, the flow of triggering current should be sunk out from the gate of the triac.

In order to produce a negative gate drive, most circuits tie the positive supply terminal to the ground terminal, which is shared by MT2, so that the negative current can drive the gate in quadrants II and III. Most triac control circuits in this configuration have 'hot' and floating references all over the circuitry. Such a set up can be hazardous.

In the design presented here, the triac driver circuit eliminates such hazards by sharing the common ground of the circuit with the neutral power line. The hazardous locations would only be the power line input and the terminals across the load.

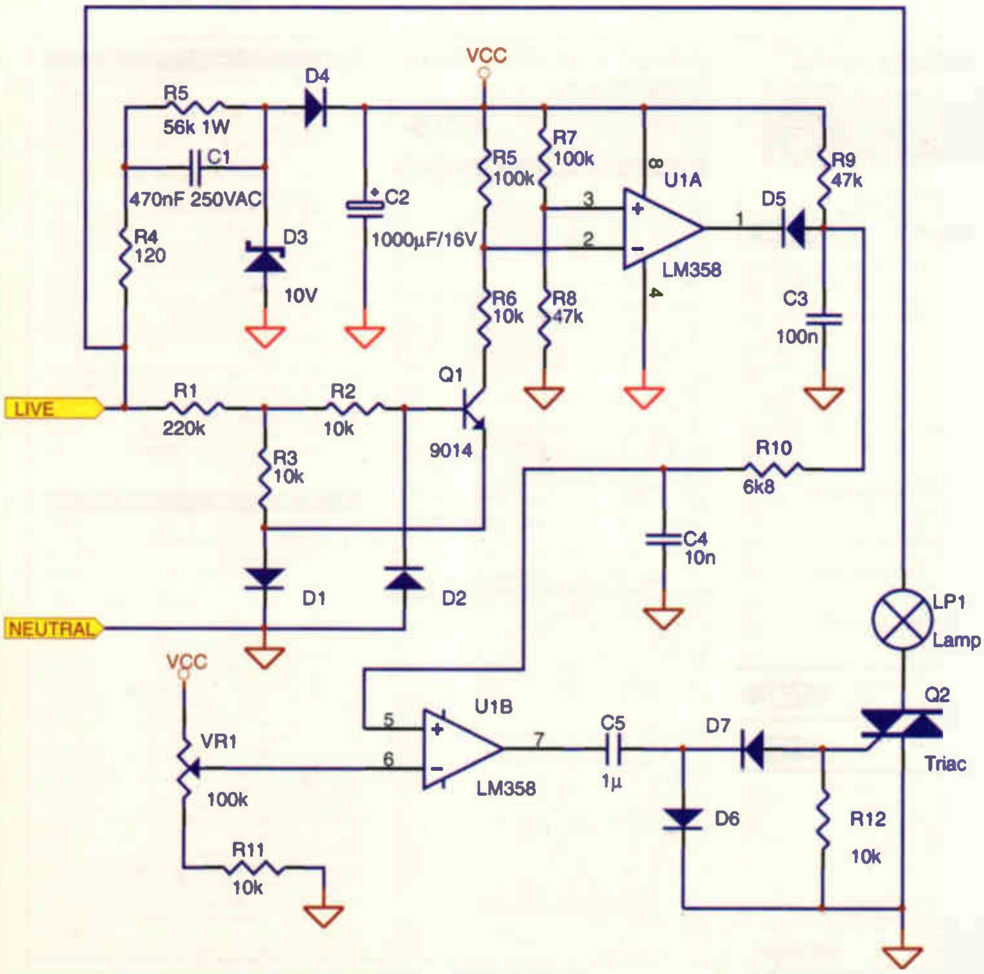
The driving circuit

The circuit is a typical inverter running from synchronous pulses from the zero-crossing detector circuit published on page 36. It does not need to link up to the live line for triac control as other commercial units do. With the introduction of the AC coupling capacitor, it allows additional and adequate protection for the driver circuit.

The basic principle of getting negative drive with a common ground connection is to make use of a charge and discharge circuit. A complementary output driver would be needed.

The capacitor is charged up most of the time and then discharged at 1 to 2ms after the zero-crossing point. This generates a negative pulse current passing through the gate during the discharge cycle.

Note that in using the operational



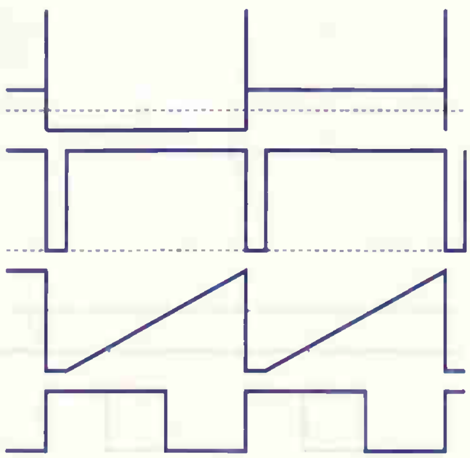


Fig. 2. AC lamp on/off control circuit.

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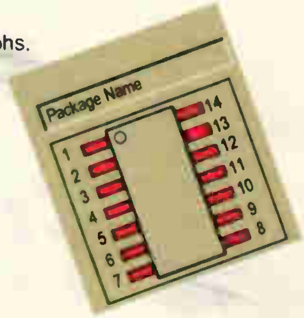

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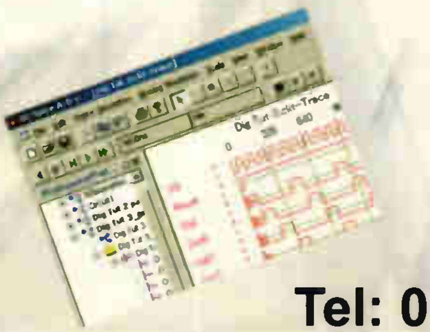

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amplifier for the driver, there are, in general, three types of output level swing:

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- 0V to V_{cc}–0.6V (rail-to-rail) output, such as TLV2462.

In this application, the output should be close to ground level during discharge so only types the fist two types can be used.

Control circuitry

The control circuit has the following novel aspects: AC coupling gate drive; power saving; independent DC supply voltage.

AC coupling gate drive: The output is virtually isolated from the gate. It provides additional protection to the circuit should someone accidentally short the gates to MT1 or MT2 during while experimenting. This assumes that the capacitor can withstand the voltage.

Power saving: Since the triggering circuit requires a short duration of discharge current throughout the 10ms period of every half cycle, the triggering pulse may be designed to have a discharge pulse width of, say, 2ms, taking 8ms for the capacitor to charge up.

Assume that the charging and the discharging rates are linear and perfect. For such a 4:1 current ratio, a triac of 10mA trigger current would only need 2.5mA charging current.

Theoretically, if the triac responds in 1ms during the discharge cycle, this gives a ninefold charging advantage. In other words, it is possible to charge at 1.1mA for a 10mA triggering device, provided triggering occurs at the right time in the cycle. Hence the supply current can be adjusted according to the mark:space ratio of the driver.

Independent DC supply voltage: Since the triac driver is triggered negatively, it is isolated from the charging current delivered from the operational amplifier, which does not necessary depend on the supply voltage. Thus the operational amplifier circuit may have a power supply voltage closed to its maximum limit. This would provide more charge to the capacitor for the next discharge pulse, resulting in a further saving in the charging current.

If the current consumed during the operating half cycle is limited, not only can the capacitance be reduced, but also the current consumption of the whole unit.

The complete circuit is shown in **Fig. 1**. The control signal is a direct voltage selected by an external variable resistor. This circuit can easily be adapted for use with microprocessor control as the i/o interface can redesigned to suit various voltages.

Lamp controller

In this lamp controller, **Fig. 2**, control is set by varying the voltage at the inputs of U_{1B}. The sawtooth wave is delayed by R_{10}/C_4 for control at lower conduction angles. These parts can be ignored as they give an improvement in brightness of around 1 to 2dB.

The design is easily interfaced to digital circuitry since there is no worry of floating ground. A mark:space ratio of 4 to 5 could be easily obtained. Higher ratios would require more precise component values and further circuit optimization, such as the addition of logic gates.

The power circuit may also have room for optimization. The standby current could be further reduced down below 10mA by adding a pre-regulating stage of 33V zener and 10μF/35V capacitor.

Since the major current consumer is the triac trigger and the DC power circuit is a shunt regulator, which requires around 3 to 5mA to maintain the regulation, a higher voltage would shorten the charging time and increase the discharge current for triggering. In fact, regulation is not essential as the triggering pulse is generated from the short discharge pulse irrespective to the supply voltage.

A hint for optimizing the performance is to select the right triac first and then calculate the required currents and the pulse widths at different points of the circuit.

Yeuk-chi Chan
Kowloon
Hong Kong

Zero-crossing detector for triac control

Developed for use in lighting control circuits, this zero-crossing detector is compact and costs much less than any ASIC of similar function. It also functions as a frequency doubler.

Most triac control circuits require a zero-crossing detection circuit for zero phase reference – especially when the circuit is linked to logic control units.

If the circuit is designed to drive the triac in quadrants II and III, the it needs either a negative supply or an isolated triggering circuit. Using a negative supply, greater care is required in analysing the circuit since the reference ground is floated to the live terminal. With an isolated trigger, it the transformer would be expensive and may make the product too bulky for small household applications.

This original idea is derived from an old trick for making full wave rectification using only *one* diode. The answer is to use a bridge circuit of three resistors and one diode. If the reference point is taken from the voltage divider of two identical resistors, then the voltage coming out

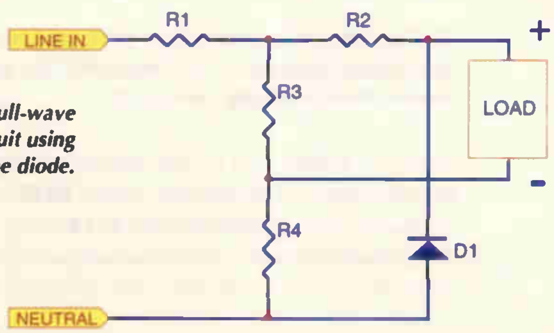

Fig. 1. Full-wave rectifier circuit using just one diode.

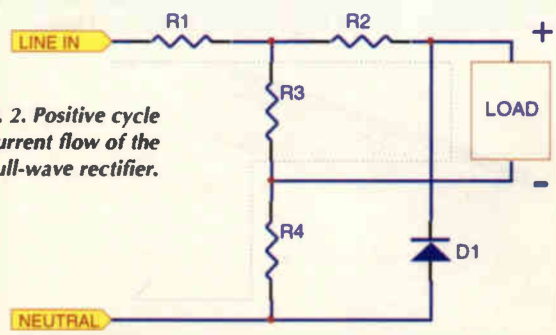

Fig. 2. Positive cycle current flow of the full-wave rectifier.

of the diode should have a full wave rectified waveform.

Single-diode full-wave rectification

A full-wave rectifier is also a frequency doubler where the period of the power line is changed from 20ms to 10m. The old ways of getting 10ms synchronous pulse could use Exclusive-Or gates or A transformer. These methods would have higher cost and size as compared to the circuits described here.

Figure 1 illustrates how to full-wave rectification out of a single-diode. Here, the reference point is shifted up by half of the incoming voltage. This circuit fits our requirement since we don't have to consider the efficiency of the rectifying circuit. **Figures 2** and **3** add details.

Zero-crossing detector

The zero-crossing detection circuit is derived from the single diode detection circuit. The load is replaced by the base-emitter connection of an n-p-n transistor.

In order to fix the voltage and allow even pulse width, a second diode is added to balance the waveform. This ensures that the transistor base would remain at a potential difference of two p-n junctions above and below ground level during turn on.

Output from the transistor is a regular 10ms negative-pulse waveform. The selection of the resistors may vary the slope of the edges and the pulse widths. **Fig. 4**.

The circuit requires just a few components to get the AC synchronous pulses, which in turn could be linked up to a suitable delay circuit for phase control.

It is important to have the supply voltage common ground connected to the neutral line. A simple single wave rectification is enough for V_{cc}.

Yeuk-chi Chan
Kowloon
Hong Kong

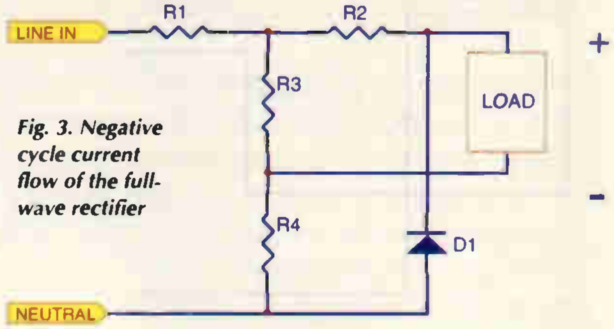

Fig. 3. Negative cycle current flow of the full-wave rectifier

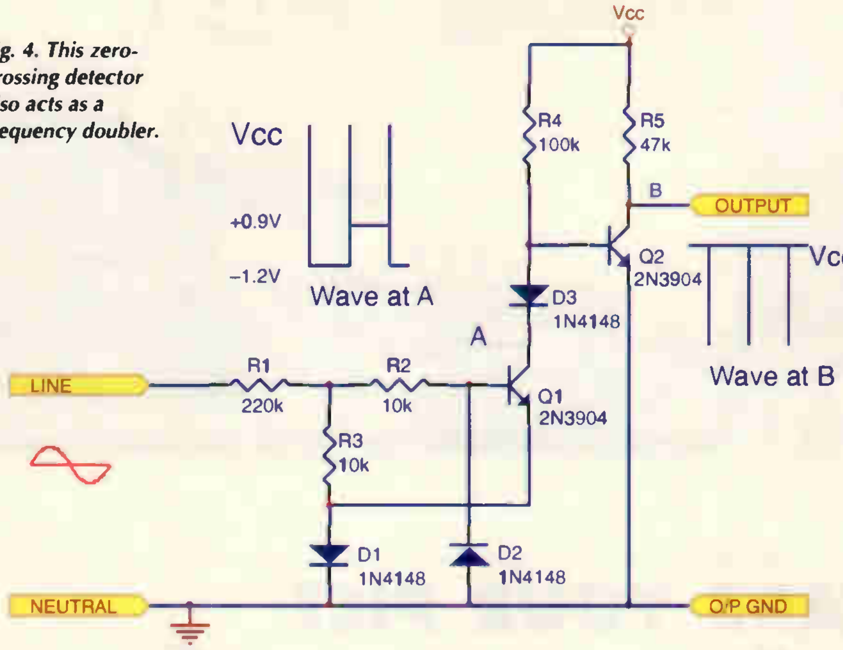

Fig. 4. This zero-crossing detector also acts as a frequency doubler.


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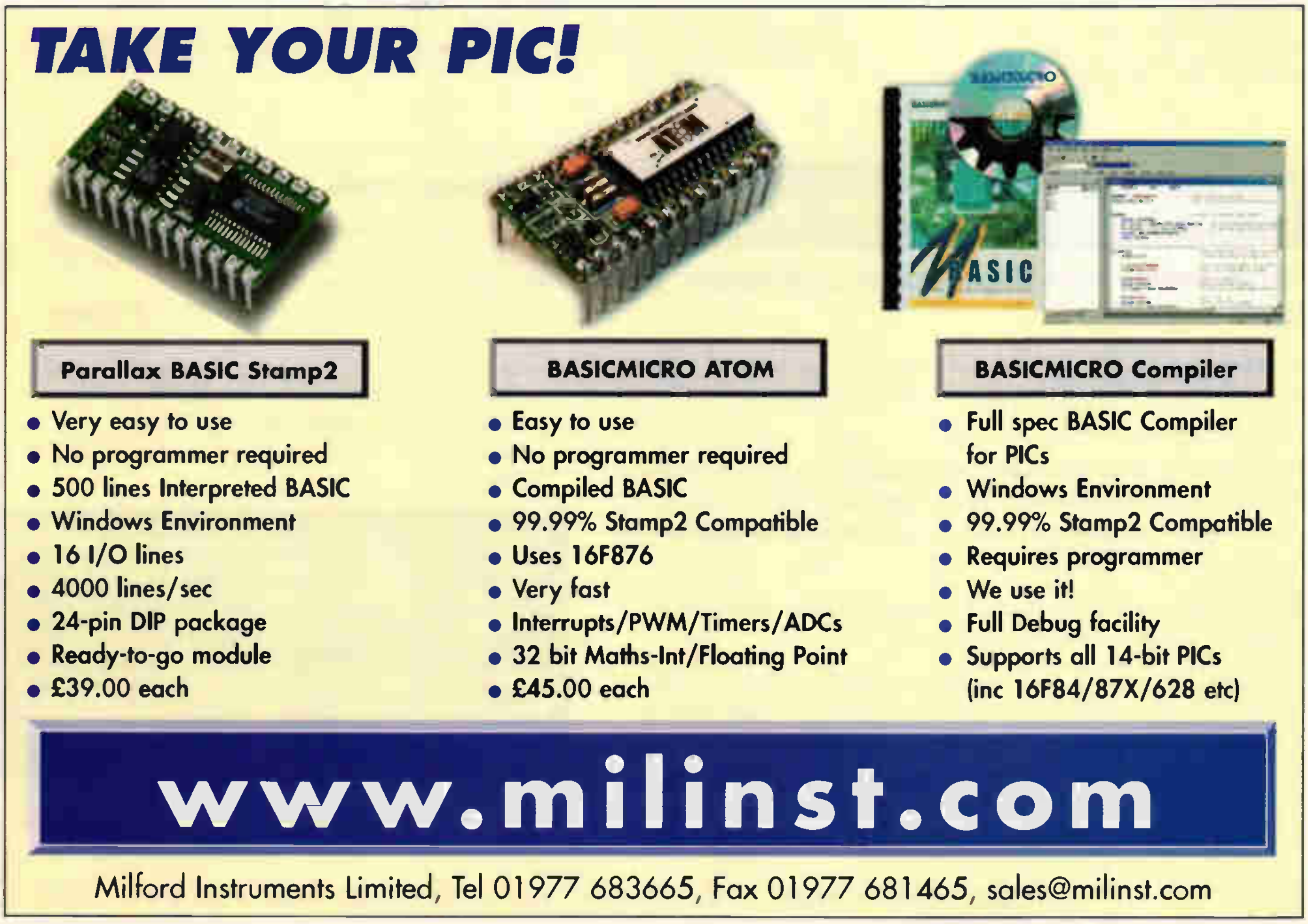
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NEWPRODUCTS

Please quote *Electronics World* when seeking further information

Optical connectors and assemblies

Radiatron have added the Fujikura range of optical connectors to their product line. The firm's products are primarily used in fibre optic networks. The standard range of over 45 types includes multifibre connectors, field installable versions, backplane connectors, adapters, mechanical splices and fibre optic cord assemblies. Where applicable the connectors are fully compliant with international specifications including JIS and IEC standards. Cable assemblies are all produced in Fujikura's ISO9002 accredited facility.

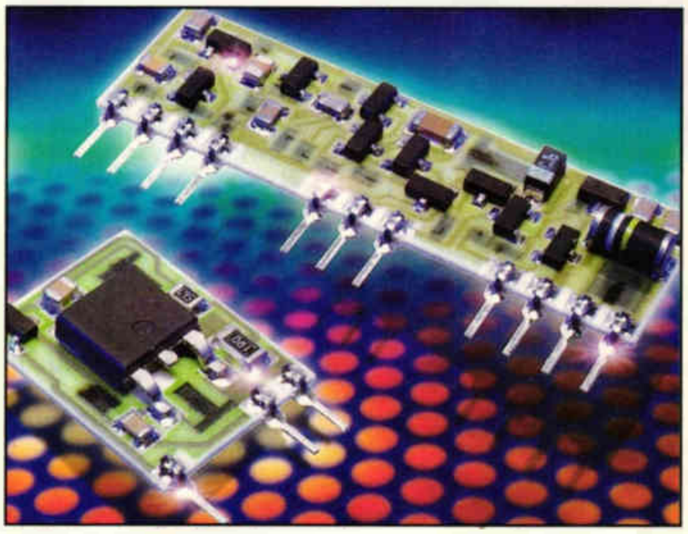
Radiatron
Tel: 01784 439393
www.radiatron.com

Connector housings with metric cable entries

Harting has introduced metric threaded versions of its Han range of heavy-duty connector hoods and housings. The connectors conform to the recently published international DIN EN 50262 metric thread specification standard, and the firm's existing PG Series (PG7 to PG48) will be replaced by the metric series M12 to M63. The adoption of metric threads is intended to simplify the understanding and specification of glands, since the product type description contains the thread dimension (eg. M16 refers to 16mm thread diameter), said the company. The firm has produced a new catalogue as a supplement to its main heavy-duty Han connectors catalogue in order to support the change from PG threads to metric threads.

Harting
Tel: 01604 766686
www.harting.co.uk

Infrared daughter boards open doors



The IRTI and IRDI infrared pulse transmitter and detector from RF Solutions are designed for applications such as automatic door opening and building security systems. Both transmitter and detector boards have SIL pins to enable vertical mounting into customer PCBs. The transmitted infrared beam has a typical frequency of 400Hz and can be detected by the IRDI module at a distance of up to 20m. If the beam is broken, then both high and low output signals are generated on IRDI pins nine and seven respectively. The thick film hybrid technology used on both modules results in stable electrical characteristics and high RFI immunity. Supply voltages required are 9VDC for the transmitter, and either 12 or 24VDC for the detector.

RF Solutions
Tel: 01273 480661
www.rfsolutions.co.uk

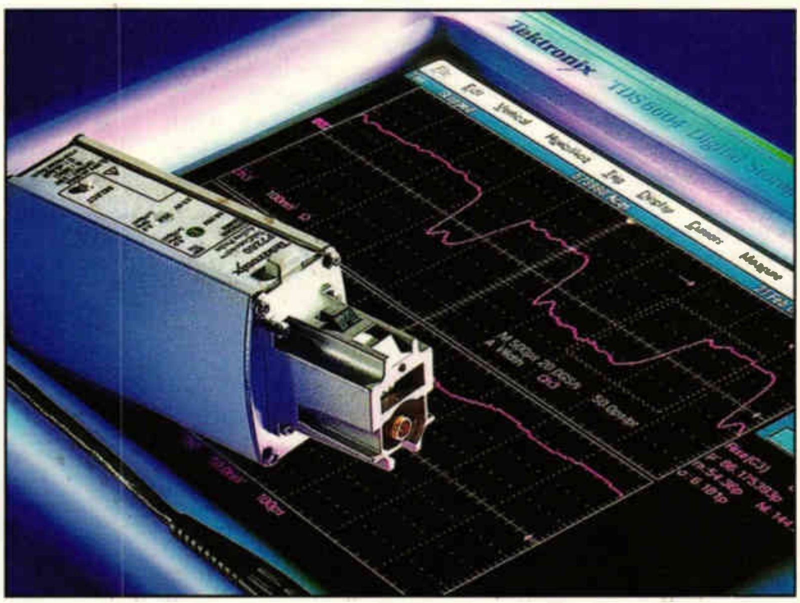
Low power SRAMs run at 1W

Apta group has introduced a range of monolithic SRAMs designed to minimise power in handheld appliances. Featuring maximum operating and standby power dissipation values of 1W and 82mW respectively, the MSM8512CB devices, from the company's HMP range, are 512k x 8 SRAMs with access times of 15ns and 20ns. The devices are packaged in a 48-ball (1mm pitch) 'chip size' bga (ball

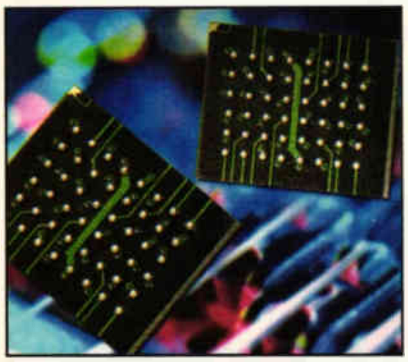
Fast probes for 5GHz scope

Tektronix has announced two oscilloscope probes and accessories for the acquisition and analysis of high-speed signals in computer, communications and advanced electronics. It is the firm's first oscilloscope probe to implement IBM's silicon germanium (SiGe) technology. Called the P7260, the active probe has a 6GHz bandwidth and better than 75ps rise time capability. It features a toggle switch that offers the choice of using the probe with a dynamic range of 6V peak-to-peak with x25 attenuation or at 1.5V peak-to-peak with x5 attenuation with increased sensitivity. An input capacitance of less than 0.5pF. The PPM100 Probe Positioner employs 3-axis control that holds a probe in position, the tool is available with either a high-mass base camp or a clamp. The TCA-IMEG high-impedance buffer amplifier provides selectable input coupling, selectable bandwidth limit, and a 1MΩ input.

Tektronix
Tel: 01344 392000
www.tektronix.com



Please quote *Electronics World* when seeking further information



grid array) measuring 10 x 8 x 1.4mm thick. According to the supplier, these are less than 50 per cent the board area of conventional SOP, SOJ and TSOP 11 packages. Devices with a 5V supply voltage provide a standby supply current of 60mA and an operating supply current of 185mA. There is also a 3.3V version with the same package and dimensions.
Apta
Tel: 0191 2930500

Motor drive gets intelligent power module

International Rectifier (IR) has launched its first programmable isolated-intelligent power module (PI-IPM) in the iNTERO range. The device integrates the power stage with the embedded driver board containing a programmable DSP on board. It is designed for three-phase, AC induction and brushless DC industrial and servo motor drive up to 15kW. The PIIPM50P120-B004 is rated at 1200V and 50A and includes all power semiconductors for a three-phase inverter power stage, including non-punch-through IGBT, as well as a programmable 40MIPS DSP, current sensing, isolation, gate driver and power stage protection for direct interface with the motor drive host and the input stage. The module includes an asynchronous 2.5Mbit/s serial port compatible with the scalable coherent interface/IEEE 1596-1992 (SCI), service provider interface (SPI), and controller area network (CAN). A joint Action Test Group (JTAG) IEEE 1149.1 standard port for DSP interfacing and an isolated serial port input with strobe signal for quadrature encoders or SPI communication are also included. The device is housed in the EMP package that has the mechanical outline of the industry-standard EconoPack 2 package.
International Rectifier
Tel: 020 8645 8003
www.irf.com

Power buffer IC gets in a tri-state

Torex Semiconductor has a pair of tri-state output buffer ICs with built-in input amplifier and divider circuits. Designed for

voltage-controlled crystal oscillator circuits, the XC2300/1 are CMOS buffers with maximum operating frequencies of 70 and 160MHz respectively.
Torex Semiconductor
Tel: 01509 269992
www.torex.com

Packet processor for 3G basestation

Broadband Technology is offering an access packet processor from WinPath designed to handle data-path and control-path packet processing functions required in 2.5G and 3G basestations and digital subscriber loop access multiplexers. The design incorporates a MIPS 64bit Risc processor core, which can be removed and used with an external PowerPC bus compatible microprocessor. With 18 resident protocols available are the WIN777 which includes both control and data path functionality and the WIN770 for data path only (no MIPS core). Each product is offered in both 166MHz and 200MHz versions.
Broadband Technology
Tel: 01727 791000
www.broadband.uk.com

Microwave or optical test gets custom switch system

Racal Instruments has developed a switching platform for custom RF, microwave and optical switching, which offers users custom configurations. The aim has been to tackle the requirement to pre-condition signals from a test object as well as switch them. The system, the 1257, incorporates elements of control, conditioning and signal routeing, as determined by the needs of the test engineer for each application. The lifetime of a switch is typically around one million cycles - which can be exceeded in as little as two months. This makes 'serviceability' a primary consideration, said the supplier. Removable drawers allow fast access to all components, both diagnostics in development applications, and for ease and speed of service. Removable top and bottom covers allow for maintenance whilst the unit is under power, thus avoiding the need for re-calibration of the unit. The channel capacity is up to 240 channels. It has GPIB and RS-232 remote interfaces in 19in. rack mount 4U, 5U, or 6U chassis.
Racal Instruments
Tel: 01628 604455
www.racalinstruments.com

TV chip combines set-top with cable modem

Broadcom's combined cable TV set-top box/cable modem is designed to provide a dedicated digital cable television channel as well as a DOCSIS 1.1 cable modem supporting interactive TV functionality and broadband Internet access through the same device. The BCM7110 also incorporates multiple features considered important for next-generation digital cable set-top boxes, including personal video recording (PVR) functionality, a

Controller with built-in TCP/IP

Epson's latest network controller incorporates the protocol stack functions and other elements needed to connect household and industrial appliances to a network.
As protocols required for TCP/IP connectivity are processed internally, only simple command and data need to be sent from the host MPU. The SIS60000 enables devices that use 8- or 16-bit MPUs to change into network-connectable devices without any advanced operating system or commercial protocol stack. The company also intends to provide IPv6 support for faster connections and intends to develop an interface module that incorporates the SIS60000 controller. Compatible protocols include ARP, ICMP, IP, TCP, UDP, HTTP, DHCP, TFTP, SNMP, with IPv6 under planning. Applications are expected to include home electronics products, home security, home gateways, electric power and gas control terminals.

Please quote *Electronics World* when seeking further information

MIPS processor and graphics engine. In addition to the cable modem/set-top box interfaces, set-top boxes based on the BCM7110 enable the distribution of voice, video, and data services throughout the residence. It supports home networking through Ethernet, 802.11b Wi-Fi wireless, HomePNA-over-Coax, and various other options. The chip includes a complete DOCSIS 1.1 media access controller that supports quality of service (QoS) for broadband interactive services such as Voice over IP (VoIP) and video conferencing. The chip's dual tuner architecture allows for simultaneous viewing of Internet and video, or two independent programme streams for watch and record PVR. It can also be configured to provide true watch and record and DOCSIS Internet browsing simultaneously with an external QAM demodulator.

Broadcom
Tel: 001 949 450 8700
www.broadcom.com

Low noise VCO for wireless Internet systems

Z-Communications is offering a VCO specifically for wireless Internet systems, which generates frequencies between 2535-2675MHz within 0.4-2.9V DC of control voltage.
According to the supplier, this low noise VCO exhibits a clean spectral signal of –84dBc/Hz, typically. The SMV2560A provides the end user with 73dBm of output power into a 50Ω load and suppresses the

second harmonic to better than –12dBc. This device draws 21mA of current from a 3V DC supply. Available from Eurosource Electronics, the device comes in a surface mount package measuring 0.30 x 0.30 x 0.08in.

Eurosource Electronics
Tel: 020 8878 5355
www.eurosource.co.uk

Silicon support for software in cars

Philips Electronics has launched an automotive software radio semiconductor design, the SAF7730 is a dual IF car radio and audio Digital Signal Processor (DSP).
According to the supplier, user specific functions can be configured to match the software platform required for the car radio manufacturer's system.
The chipset includes audio features including adaptive ultra-bass 11 and music elevation, combined with a high sound quality. The software radio also includes an iDAB interface, enabling digital radio.

Philips Electronics
Tel: 0031 40 272 2091
www.semiconductors.philips.com

Trimmers for tight corners

A small trimmer chip capacitor measuring 1.5 x 1.7 x 0.85mm is available from Murata.
According to the supplier, the TZR1 series requires one third of the board mounting area of a 2mm device. The trimmer is available in two capacitance value ranges: 1.5 to 4.0pF and 3.0 to 8.0pF. Both are rated at 25V DC, operating temperature range from –25 to +85°C.
Primarily designed for high frequency applications, these miniature trimmers are suitable for RF and wireless equipment including Bluetooth devices, remote keyless entry systems, portable communications equipment, pagers, crystal oscillators and filters. The lead-free TZRI design is constructed with no plastic material, which provides superior heat resistance to enable high temperature lead-free soldering.

Murata
Tel: 01252 811666
www.murata.co.uk

100W quarter-brick DC/DC converters deliver 1.2V

The Lambda PAQ series of DC/DC converters, available from Powerline, is a range of industry-standard quarter brick profile devices with ratings from 50 to 100W. designed for use in distributed power systems where the voltage is converted at board level, the units feature up to 90 per cent efficiency and do not require heatsinks. Nominal input voltage is 48V DC,and output voltages of 1.2V, 1.8V, 2.5V, 3.3V and 5V are available at current ratings from 10A to 25A depending on the model. Line and load regulation are 10mV. Overcurrent and overvoltage protection are fitted as standard. Remote sensing, remote on/off control and series operation are available as options.

Powerline
Tel: 01494 753800
www.powerline.co.uk

Alphanumeric LCD with CMOS drivers

The SC series of alphanumeric dot-matrix LCD modules from Wimbledon-based CTL Components have been designed with built-in CMOS driver devices to reduce component count and simplify the interface circuitry. Available in formats ranging from 2 lines of 8 characters to 4 lines of 40

150W switching PSU in 1U enclosure

Ultimate Renaissance is offering IU form factor switchers packing 150W and available in output voltages from 1.8V to 48V. The Cortech CSS150 is available in quad, triple, dual and single output versions. Output voltage is selectable from 3.3V to 48V from output 1 and 2, at 150W maximum from output 1 and 100W maximum from output 2. Choices for outputs 3 and 4 allow any value from 3.3V to 24V, at 45W. Outputs 3 and 4 can also be combined for a maximum 60W continuous delivery. Features include remote sense, single wire current sharing on outputs 1 and 2, OVP, over current protection and thermal shutdown. Active power factor correction is also included and all units are packaged in 4.2 x 8 x 1.5in. u-channel suitable for system enclosures.

Ultimate Renaissance
Tel: 01793 439310
www.ur-home.com

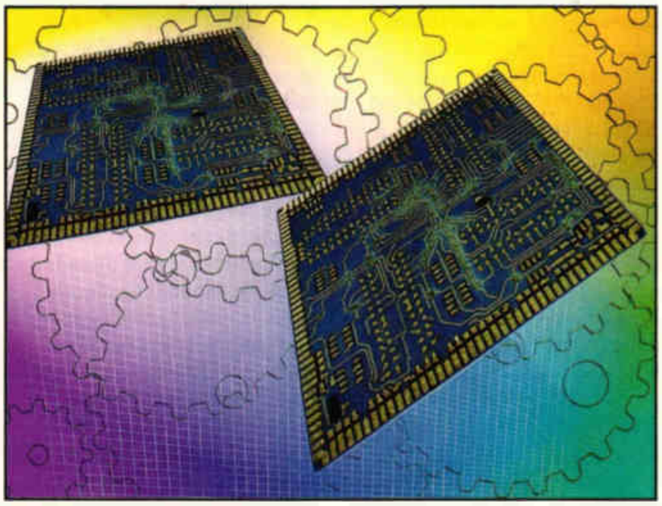
Please quote *Electronics World* when seeking further information

characters, the modules can be specified with either twisted-nematic or supertwist LCD technology, according to the requirements of the application. Backlit versions are also available, utilising a choice of green, blue/green or white electroluminescent illumination or yellow, orange or red LED backlights. Module sizes range from 58.0 x 32.0 x 8.6mm to 190.0 x 54.0 x 9.0mm and all use a 5 x 7 dot-matrix format with cursor.

CTL Components
Tel: 020 8545 8700
www.ctl-components.com

High reliability memory wins size upgrade

Apta, formerly HMP Europe, has upgraded its range of high reliability memory modules making use of its pinned uncommitted memory array (Puma) design, which allows for the addition of higher order address lines. The 4 x 512k x 8 Puma2 SRAMs feature the same footprint as the first generation 4 x 8k x 8 devices. The hermetically sealed, ceramic part is housed in a 66-pin PGA package. The range can now offer 8, 16 and 32-bit wide memory devices. They are user-configurable, and upgradeable and can feature a combination of memory types, such as EEPROM, flash and SRAM in a single footprint. For example, a 512k x 16 flash and 128k x 16 SRAM die can be supplied in one device. There is also a surface mount equivalent constructed as two stack layers each with two memory die.



Apta
Tel: 0191 2930500

Real-time design for Ada

Artisan Software Tools has released version 4.2 of its Real-time Studio which has expanded Ada support that includes reverse engineering and round-trip synchronisation. Users can also translate their legacy Ada source code into UML models, and support round-trip engineering for Ada and Spark. Also the C++ synchroniser is substantially re-written to better support re-engineering of legacy code. It uses the same new tool structure as the Ada synchroniser, and supports visual differencing of model elements against source code constructs. The parsing technology is capable of reversing incomplete source-code, and provides technology to cope with partially formed constructs and missing types. The code generation technology in both Synchronisers is entirely template-based so users can control the form and content of the generated code, allowing them complete code generation flexibility.

Artisan Software Tools
Tel: 01242 229300
www.artisansw.com

Low power microcontroller

Hitachi has introduced the first member of its H8/300L low power series of flash microcontrollers. The H8/38024F is also the first member of the series to offer an on-chip debug interface, which allows users to debug their applications in-circuit using the firm's own E10T debugger. The microcontroller incorporates 32kbytes of flash memory that can be programmed and erased with a single power supply. The device operates at maximum frequency of 10MHz using a 3V power supply, and offers on-chip peripherals including a 32kHz sub-clock, timers, a special asynchronous timer, watchdog timer, asynchronous/synchronous serial interface, 10-bit analogue/digital converter, 32 x 4 LCD controller/driver and high-current pins. The device is supported by the E6000 real-time in-circuit emulator with 4Mbyte of emulation RAM, 256 PC break points and 12 cascadable and complex hardware break points (events). An evaluation board is available. The H8/38024F is available in two 80-pin Quad Flat Packages, the 14 x 14mm QFP-80A and the 14 x 20mm QFP-80B, as well as a 12 x 12mm 80-pin Thin Quad Package, the TQFP-80C.

Hitachi
Tel: 01628 585163
www.hitachi-eu.com

Controller with built-in TCP/IP

Toshiba Electronics has launched a range of miniature chip scale power Mosfet packages that it claims are over 60 per cent thinner than conventional TO-220SM (D2PAK) devices and require 30 per cent less PCB mounting area. Called the Slim TFP package, it has a board mounting area of 10.7 x 9.2mm and a profile of 1.7mm. The devices have high DC current handling capabilities and are rated for power dissipation of up to 125W. A four-pin structure (gate, drive source, main current source and drain) is intended to minimise the effects of parasitic inductances while offering electrical noise immunity. The first Mosfet in the new package will be a 500V, 8A device with an on resistance of 750mΩ. Future devices will include 200 and 250V Mosfets.

Toshiba
Tel: 01276 694730
www.toshiba-europe.com

Power converter raises Mosfet efficiency

Texas Instruments is offering a family of low-voltage pulse-width-modulated (PWM) synchronous DC-to-DC

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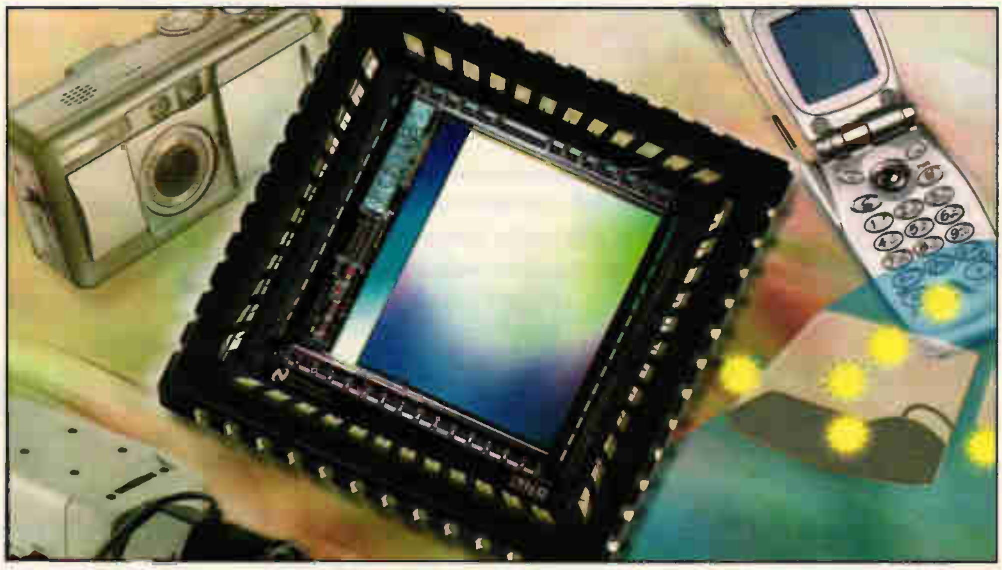
controllers. The TPS4000x PWM step-down DC-DC controllers feature a patented predictive gate drive technology and are designed to achieve overall converter efficiencies greater than 95 per cent, said the supplier. PGD technology is designed to provide the highest efficiency in rectifier control by driving Mosfet power transistors at two to four per cent greater efficiency than conventional adaptive gate drive technology. Two internally fixed operating frequencies are available, at 300 and 600kHz, as well as source and source/sink options. A programmable closed-loop soft start, short circuit disable and thermal shutdown provide protection for system circuitry.
Texas Instruments
Tel: 01242 229300
www.ti.com

200W power module baseplate cooling

Lambda's latest halfbrick DC-DC power module has a nominal input voltage of 48V. The single output 200W PAH200H48 series offers a range of output voltage of 1.8, 2.5 or 3.3V. The DC-DC converter has a power density of $72.4W/in^3$ and power efficiency of 90 per cent at an output of 3.3V. According to the supplier, there is no need for baseplate cooling under normal temperature conditions. Module footprint is 58 x 61mm and profile is 10mm. The open modules can be convection or forced air cooled. Operating temperature range is –40 to 85°C.
Lambda
Tel: 01271 856666
www.lambda-gb.com

CMOS monochrome and colour CMOS image sensors

National Semiconductor has expanded its range of CMOS image sensors which are expected to find applications in wireless access devices, entry-level digital cameras and industrial products, including security cameras and scanners. The monochrome LM9618 and colour LM9628 VGA CMOS image sensors feature the wide linear and non-linear dynamic ranges and operate over industrial temperature ranges, so are suitable for outdoor applications. The sub-QCIF (quater common interface format) LM9630 supports frame rates exceeding 600frames/s. The 1.3Mpixel monochrome LM9638 and colour LM9648 sensors offer an optical format of one half inch.
National Semiconductor
Tel: 0870 242171
www.national.com

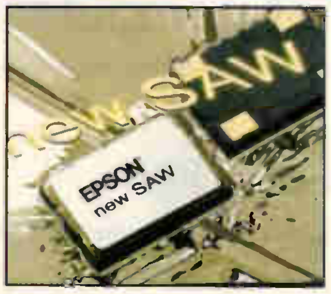


Multimeter has reference accuracy

Fluke's 8508A reference multimeter is an 8.5-digit resolution instrument designed as a precision measurement tool for calibration laboratories. In addition to measuring voltage and current for both AC and DC, and frequency, the meter also includes other features. A ratio feature, with front and rear connectors supporting all 8508A functions and GPIB control, allows for automated measurement transfers. There is also current measurement up to 20A. One year accuracy is specified as 3ppm and 20-minute transfer uncertainty is 0.12ppm.
Fluke
Tel: 01603 256689
www.fluke.co.uk

1GHz SAW filter for networking

Epson's quartz-based precision SAW (surface acoustic wave) resonator for networking applications is claimed to



improve on the temperature characteristics compared to conventional SAW resonators by 50 per cent. The firm also offers prototype-free custom device development. The SAW resonator has an improved temperature coefficient of –1.7 x $10(-^{8})$. Fundamental oscillation up to 1GHz is supported.
Epson
www.epson-electronics.de

Processor board has PCI bridge to 133MHz

Radstone Technology's latest G4DSP signal processor board based on the 7410 PowerPC Risc processors with AltiVec between 400MHz and 550MHz includes a processor and memory upgrade. The design means that the board provides each 7410 PowerPC compute node with 256Mbyte L2 cache and an IBM-CPC710 64bit/66MHz PCI host bridge. The board exercises the same crossbar like PCI ring architecture and takes advantage of a PCI bridge upgrade to take the processor bus and memory bus interfaces from 100MHz to a peak of 133MHz. As with the hardware architecture, software support is also open standard to protect software investment and ease development. The G4DSP supports VxWorks/VxMP and VSPWorks DSP OS from Wind River Systems and also MPI message passing libraries and VSIPL signal processing libraries from MPI Software Technology. It is available in five ruggedisation build levels for industrial or extended temperature, shock and vibration environments in either air-blown or conduction-cooled formats. ■
Radstone
Tel: 01327 359444
www.radstone.co.uk

Transparent V-I protection in Audio Power Amplifers

(Part 1)

The desirability or lack thereof, of over-voltage and over-current protection for power semiconductors in audio power amplifiers remains a point of contention in the field. Michael Kiwanuka explans.

For example, Nelson Pass[2] appears to recommend multiple-transistor complementary output stages, as mandated by class-A operation, to circumvent the need for V-I protection of bipolar devices, while Rod Elliot[3] suggests that V-I limiters can be dispensed with altogether by adopting e-MOSFETs.

These views appear to be rather more widely accepted than they should, and constitute a charter for near heroic unreliability in amplifiers so designed. The zener diode-clamping of gate-source voltage for e-MOSFET's is thought by some[4,5] to be all that is required with regard to protection. While the zener diodes are mandatory, (ideally with $10V<V_{zener}<20V$ to prevent premature clamping), they only serve to protect the e-MOSFET gate oxide insulation from over-voltage destruction[6], and do nothing whatever to protect the device from accidental short circuits and forbidden voltage-current combinations that may occur when the amplifier is called upon to drive reactive loads.

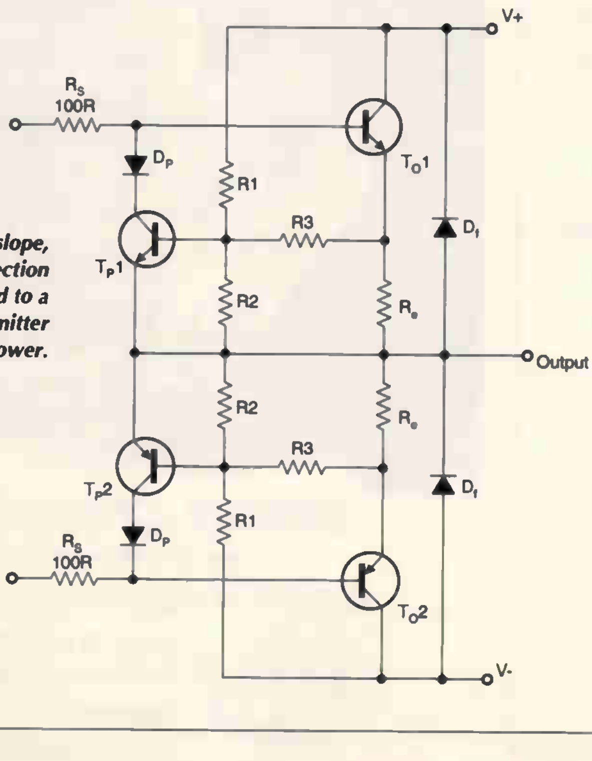


Figure 1: Single slope, linear foldback protection circuit applied to a complementary emitter follower.

The positive temperature coefficient of on-resistance[7], (and therefore negative temperature coefficient of drain current), enjoyed by e-MOSFETs eliminates the secondary breakdown phenomenon which is the bane of bipolar transistors, but does not constitute a licence for wilful violation of power dissipation limits in linear, audio-frequency applications. This is in contrast to ultrasonic switching usage, where e-MOSFET dissipation bounds can be blissfully ignored, and adherence to drain current and drain-source voltage limits will suffice.

All output stage semiconductors used in complementary, or quasi-complementary, (full or half bridge), linear audio power amplifiers, without exception, require V-I protection for reliable operation. However, such circuitry must be carefully designed to prevent premature activation during normal amplifier operation.

Single-slope, linear-foldback limiting

Many low to medium-power, (sub-100W), commercial audio amplifiers incorporate a single slope, linear foldback, voltage-current protection circuit, (figure 1), attributed to S.G.S. Fairchild Ltd. by Dr A.R. Bailey[8]. In practice the complimentary output transistors, T_{o1} and T_{o2}, may each consist of a compound arrangement of at least two transistors in series. The collector-emitter voltage, Vce, across T_{o1} is sensed by R_1, and R_3, while the output current, in the guise of a voltage developed across emitter resistor Re, is simultaneously monitored by R_3, and R_2. The voltages are thus summed algebraically at the base of the protection transistor, T_{p1}, which is driven into conduction, shunting voltage drive to T_{o1}, in the event of an over-voltage, over-current or simultaneous occurrence of both conditions in the output device.

The series resistor, Rs, (typically 100R), expedites this process by limiting the current required by T_{p1} to shunt voltage drive to T_{o1}. The freewheeling diode, D_F, protects the output device from excessive base-emitter reverse bias[9], due to over-rail voltage spikes generated by inductive loads, while D_F performs the same function for the small-signal protection transistor, by preventing its base-collector junction from being forward biased[10].

If the output approaches the negative supply rail while driving a sufficiently low impedance, the current sunk by T_{o2} generates an appreciable voltage drop across its emitter resistor. The output is therefore at a significantly higher potential than the common input to the complementary output stage. Transistor T_{o1} is reverse biased, and T_{p1}'s base-collector junction, in the absence of D_F, would be forward biased, resulting in current flow from emitter to collector.

Diode D_F prevents this form of spurious, inverse-active mode limiter activation by decoupling T_{p1}'s collector as T_{o1}'s base-emitter is reverse biased. The potential at T_{o1}'s emitter is then equal to the output voltage since, contrary to Duncan[11], T_{o1} is non-conducting and no current, except negligible leakage, flows through its emitter resistor. By symmetry, the explanation above also applies to the negative half of the circuit.

A small-value capacitor is sometimes connected across the base-collector junction of each protection transistor[1], with a view to eliminating benign parasitic oscillation[12] that may occur sporadically in the network during the limiting process. These capacitors appear in parallel at A.C, and are entirely unsatisfactory, as they create an ill-defined and therefore undesirable feed-forward path around the output stage, shunting it out of the global feedback loop at high audio frequencies, precisely where the amplifier is most vulnerable with respect to non-linearity. Such vulnerability is due to a necessarily diminished feedback factor at high audio frequencies in the interest of Nyquist stability. Connecting the capacitor, (of the order of 1*n*0), across the base-emitter junction of each protection transistor is the preferred solution.

The resistor values for the arrangement in **Figure 1** are obtained by drawing the desired protection locus onto a linear scale graph of the output transistor's safe operating area, (S.O.A). One of the three resistors (usually R_3), is assigned an arbitrary value (typically $100R \leq R_3 \leq 1k$), and the two remaining resistors calculated from simultaneous equations developed from two convenient points on the protection locus.

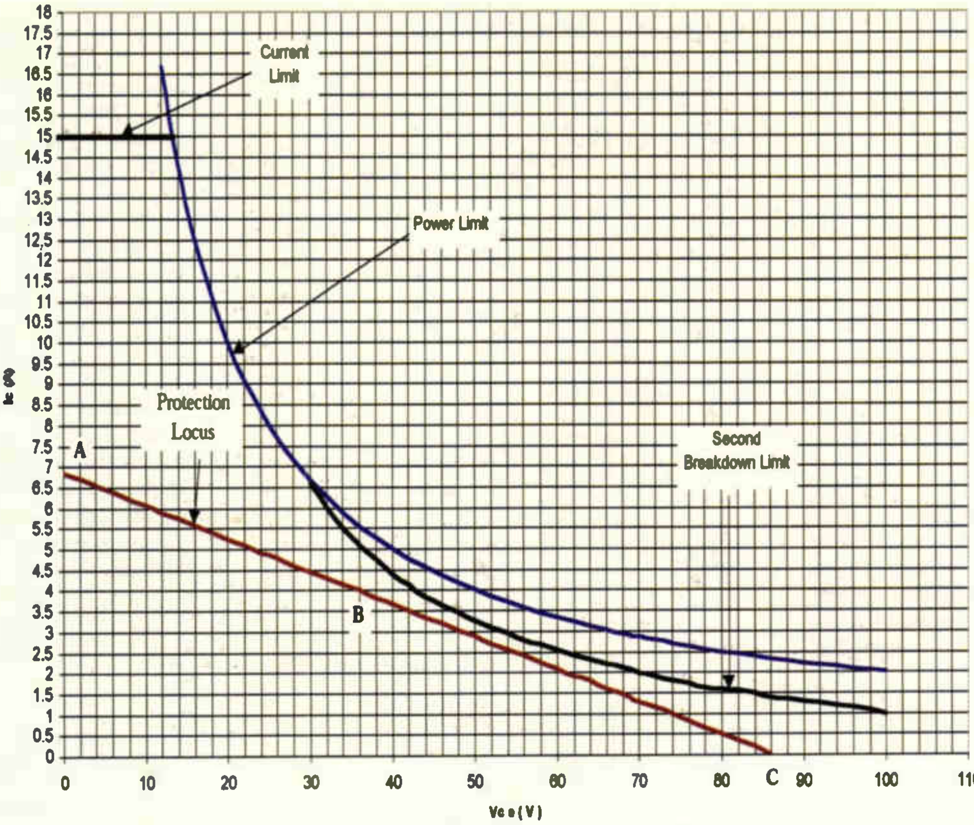


Figure 2: MJL3281A safe operating area with single slope, linear foldback protection locus drawn to intersect the Vce-axis at a value greater than $2|V_{cc}|$*, to prevent premature limiting.*

This arrangement requires that the linear protection locus intersects the S.O.A's Vce axis at a value greater than the sum of the moduli of the amplifiers voltage supplies, otherwise T_{p1} turns on under normal loading when the output swings negative, even with the output open-circuit. Similarly T_{p2} would be activated under normal output loading when the output swings positive. This effectively short-circuits the small signal circuit preceding the output stage directly to the output, causing gross and very audible distortion. Failure to adhere to the above condition appears to have caused some designers to erroneously abandon electronic S.O.A protection of any form altogether[1,13].

This requirement however, constitutes a significant limitation with regard to efficient utilisation of the comparatively large S.O.A in the low-Vce region of the graph, especially at high supply-rail voltages where, in the case of bipolar transistors, secondary-breakdown severely curtails flexibility in optimal placement of the protection locus. This is graphically illustrated in figure 2, for an amplifier with ±40V supply rails, using Motorola's excellent[14] 200W, MJL3281A-MJL1302A complementary power transistors.

Only the positive half, **Fig. 3.**, of the circuit in figure 1 needs be used to calculate the required component values. Ideal devices are assumed, with infinite input impedance, zero saturation voltage, and zero ohmic resistance, the error thus accrued is negligible. Let V_{be}=0V6, R_3=220R, Re=0R22. Taking two arbitrary points, A and B on the locus, such that,

$$0.6 = \left\{0 \leq V_{ce} < \left(2|V_{cc}| = 80V\right)\right\}$$

where for point A, Ic=6.85A; Vce=0V, and for point B,

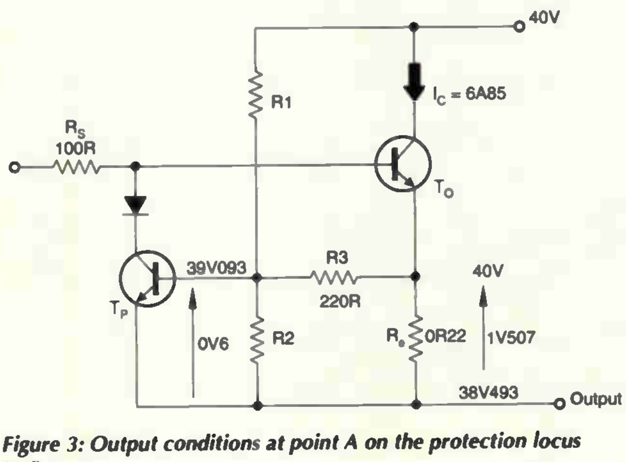


Figure 3: Output conditions at point A on the protection locus in figure 2.

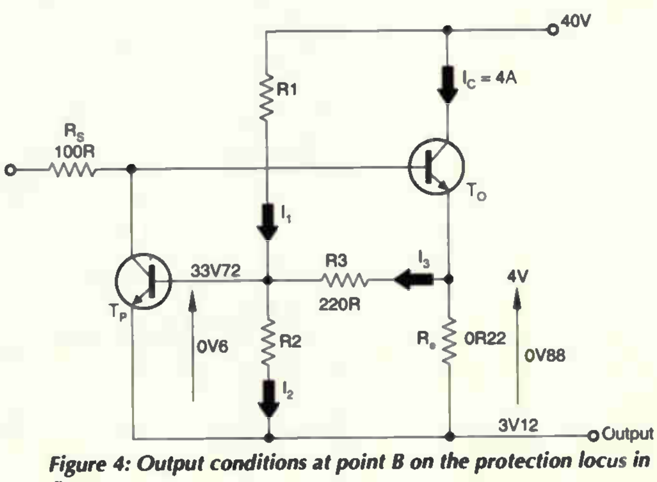


Figure 4: Output conditions at point B on the protection locus in figure 2.

Ic=4A; Vce=36V, it follows from figure 3:

$$\frac{1.507R_2}{R_2 + R_1 220/(R_1 + 220)} \qquad (1)$$

With reference to figure 4:

$$I_2 = I_1 + I_3 \qquad (2)$$

$\Rightarrow$

$$0.6/R_2 = (40 - 3.72)/R_1 + (4 - 3.72)/R_3$$

$\Rightarrow$

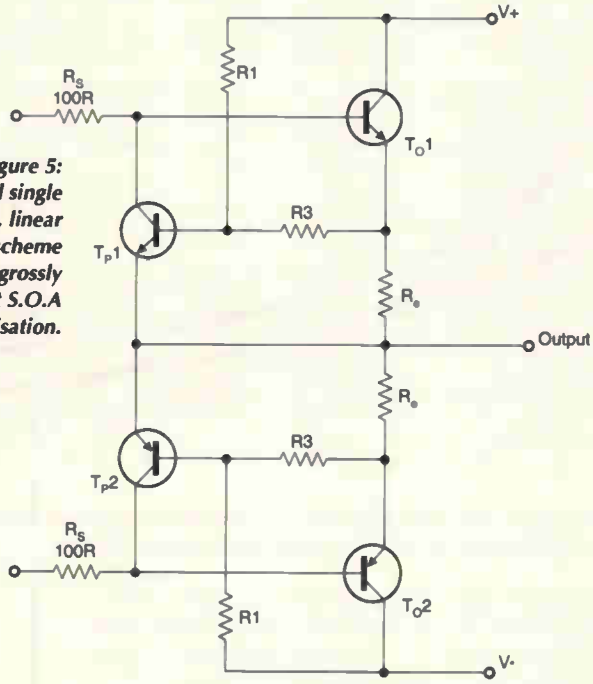


Figure 5: Compromised single slope, linear foldback scheme resulting in grossly inefficient S.O.A utilisation.

$$0.6 = R_2(36.28/R_1 + 0.28/220) \qquad (3)$$

Solving (1) and (3) simultaneously gives $R_1 \approx 12K4$ and $R_2 \approx 143R0$. To afford an acceptable degree of precision, it is recommended where necessary, that these values be made up from series, or parallel combinations of 1% resistors.

When the output swings to –40V, then 80V appears across R_3 in series with $R_2//R_3$, to a good first approximation. Therefore the voltage present at the base of the protection transistor, T_p, is given by:

$$V_{be} \approx \frac{80(R_2 // R_3)}{(R_2 // R_3) + R_1} \approx 0V55$$

It follows therefore that subject to instantaneous collector current, i_c, being less than the maximum permissible collector current, $I_{C(MAX)}$, at $V_{ce} \approx 2|V_{cc}|$, spurious activation of T_p cannot occur. A general expression which allows the rapid verification of the compliance of any amplifier using single slope, linear foldback limiting may developed:

$$\left|\frac{2V_{cc}(R_2 // R_3)}{\{(R_2 // R_3) + R_1\}}\right| < 0V6$$

$\Rightarrow$

$$\left|\frac{2V_{cc}R_2R_3}{(R_2R_3 + R_1R_2 + R_1R_3)}\right| < 0V6 \qquad (4)$$

Equation 4 is valid subject to the following condition:

$$i_C < I_{C(MAX)}\Big|_{V_{ce} \approx 2|V_{cc}|} \qquad (5)$$

This condition is invariably fulfilled during normal operation, as no practical loudspeaker system would demand that the output transistor sustain $V_{ce} \approx 2|V_{cc}|$, while providing any appreciable current.

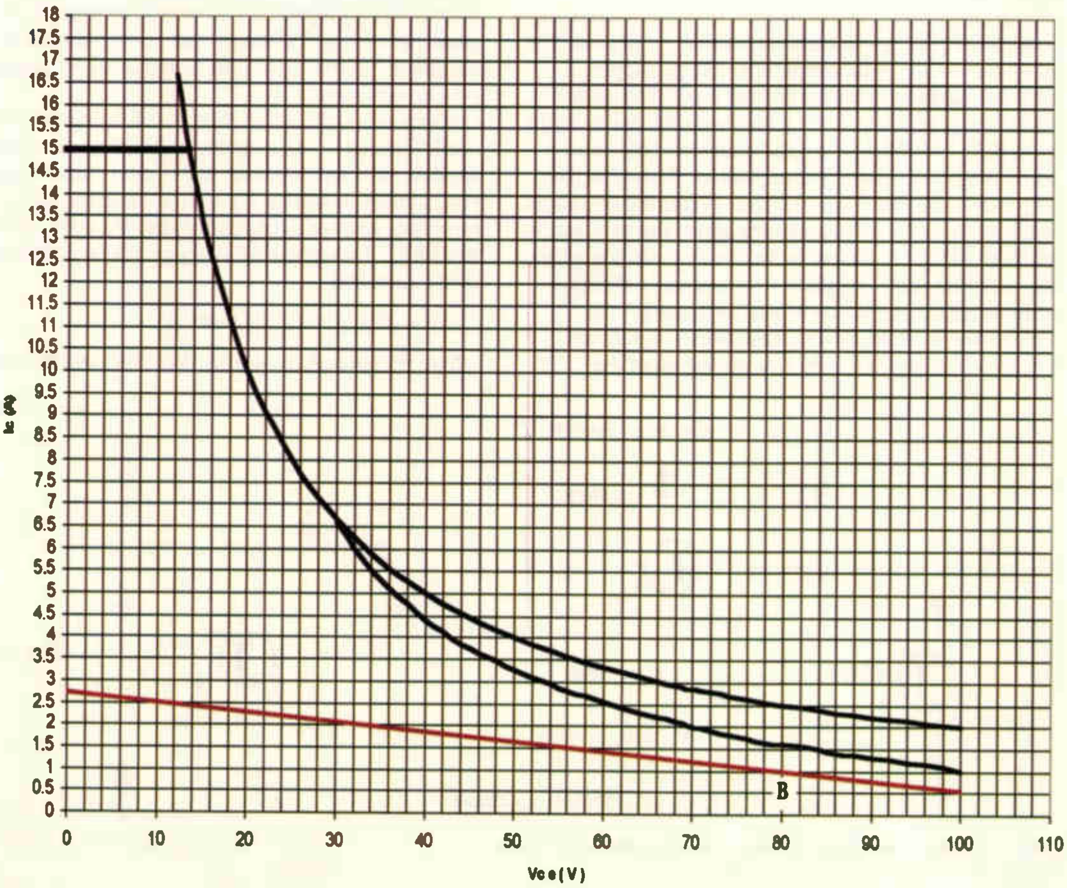


Figure 6: Linear protection locus clearly shows inflexibility of scheme in figure 5.

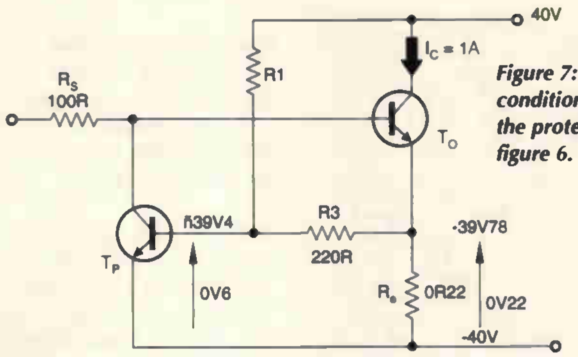


Figure 7: Output conditions at point B on the protection locus in figure 6.

Figure 5 shows a common variation[11,14,15], on the single slope, linear foldback limiter of figure 1, with resistor R_2 excised, so that from equation 4:

$$\left|\frac{2V_{cc}R_3}{\left\{R_3 + R_1 + \left(R_1\frac{R_3}{R_2}\right)\right\}}\right| < 0V6$$

Since $R_2 \to \infty$, then:

$$\left|\frac{2V_{cc}R_3}{(R_3 + R_1)}\right| < 0V6$$

The optimal protection locus for this network **Fig. 6.**, is plotted so that calculated resistor values comply with the above condition. This scheme is atrociously inefficient, as for a nominal V_{ce}≈0V and R_e=0R22, resistors R_1 and R_3 are in parallel, and collector current, I_c, is perforce prematurely limited to,

$$\left\{I_c\big|_{V_{ce}\approx 0V} < \left(V_{be}/R_e \approx 2A7\right)\right\}$$

A value of R_e=0R1 gives a modest improvement, with,

$$\left\{I_c\big|_{V_{ce}\approx 0V} < \left(V_{be}/R_e \approx 6A0\right)\right\}$$

The protection locus is realised by deriving output stage conditions **Fig. 7.** for a single, arbitrary point, B on the locus, subject to, {0<V_{ce}<2|V_{cc}|}. With V_{cc}=40V, R_e=0R22, R_3=220R and noting that R_1, R_3 constitute a simple voltage divider:

$$R_1 \approx \frac{(40 + 39.4)}{(-39.4 + 39.78)/220R} \approx 46K$$

This unwarranted dependence on the value of R_e is unacceptable, as in some applications such as output stages comprised of paralleled, complementary e-MOSFET pairs, (0R1<R_e≤1R0), may be required to ensure equable current sharing.

Driving reactive loads.

A clear appreciation of the nature of the amplifierís load is required to establish the bounds within which the V-I limiter must remain inactive. **Figure 8** shows an ideal complementary emitter follower, (in Electronics Workbench's excellent Multisim professional simulator[16]), used to drive a standard (8Ω∠0°) test load to ±40V supply rails.

The plots obtained in **Fig. 9.** show that the voltage v_{ce}, across T_{o1} is precisely 180° out of phase with the current, i_c,, its required to source; the voltage across the device is a minimum when its collector current is at a maximum, and vice versa. Instantaneous power dissipation is merely the product of instantaneous device voltage and current. Peak transistor dissipation, $P_{d(max)}$≈50W, occurs twice in T_{o1}'s conducting half-cycle, at half the peak load voltage, ($V_{out}/2$≈$V_{cc}/2$) and half the peak load current, $i_{c(peak)}/2$.

As the (8Ω∠0°) load line lies well below the linear

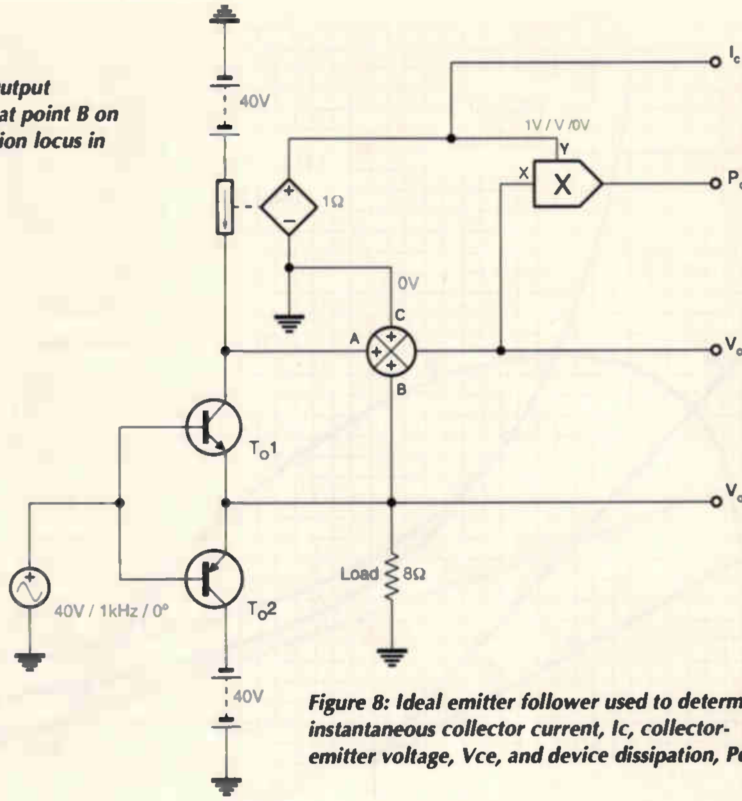


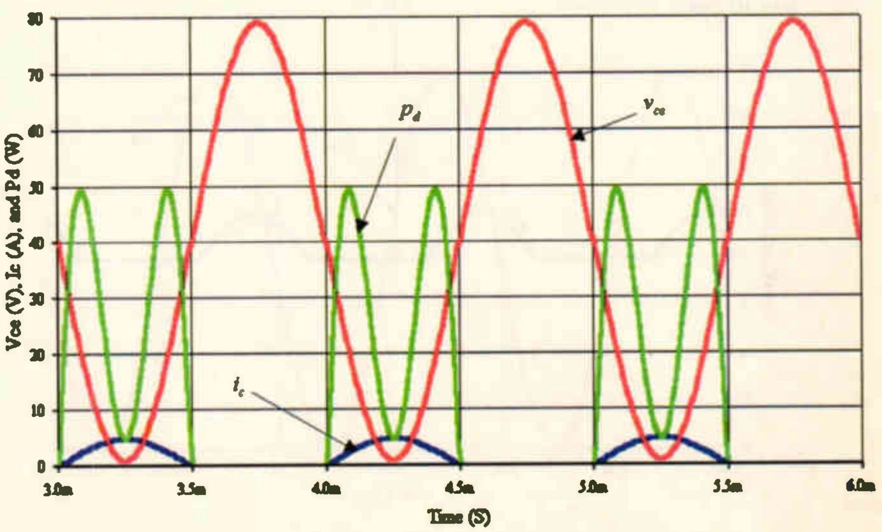
Figure 8: Ideal emitter follower used to determine instantaneous collector current, Ic, collector-emitter voltage, Vce, and device dissipation, Pd.

protection locus in figure 10, (reproduced from figure 2), it is clear that a single pair of MJL3281A-MJL1302A power transistors, operating from ±40V rails will comfortably drive an 8Ω dummy load to clipping without V-I limiting. This however, will certainly not be the case with loudspeaker loads, which are invariably reactive[17,18]. An amplifier with 'high-fidelity' aspirations, intended to drive full-range, multiple-transducer loudspeaker systems, including electrostatics, should at least be capable of driving a (4Ω∠±60°) impedance without V-I limiting.

A (4Ω∠–60°) impedance was devised by driving a 2Ω0 resistor in series with a 45μ 9441 capacitor at 1Khz with the ideal complementary emitter follower in figure 8. The traces thus obtained, **Fig. 11.**, were used to plot the (4Ω∠±60°) load line in figure 10. Peak transistor dissipation, $P_{d(max)}$≈352.9W, occurs at v_{ce}≈45.97V, and i_c≈7.68A.

In other words **Fig 12.**, because current leads voltage in a capacitive impedance, the NPN transistor, T_{o1} in figure 8, is required to source ≈7.68A when the output swings

Figure 9: Instantaneous Vce, Ic, and Pd in sourcing output transistor, driving 100W into (8Ω∠60°).



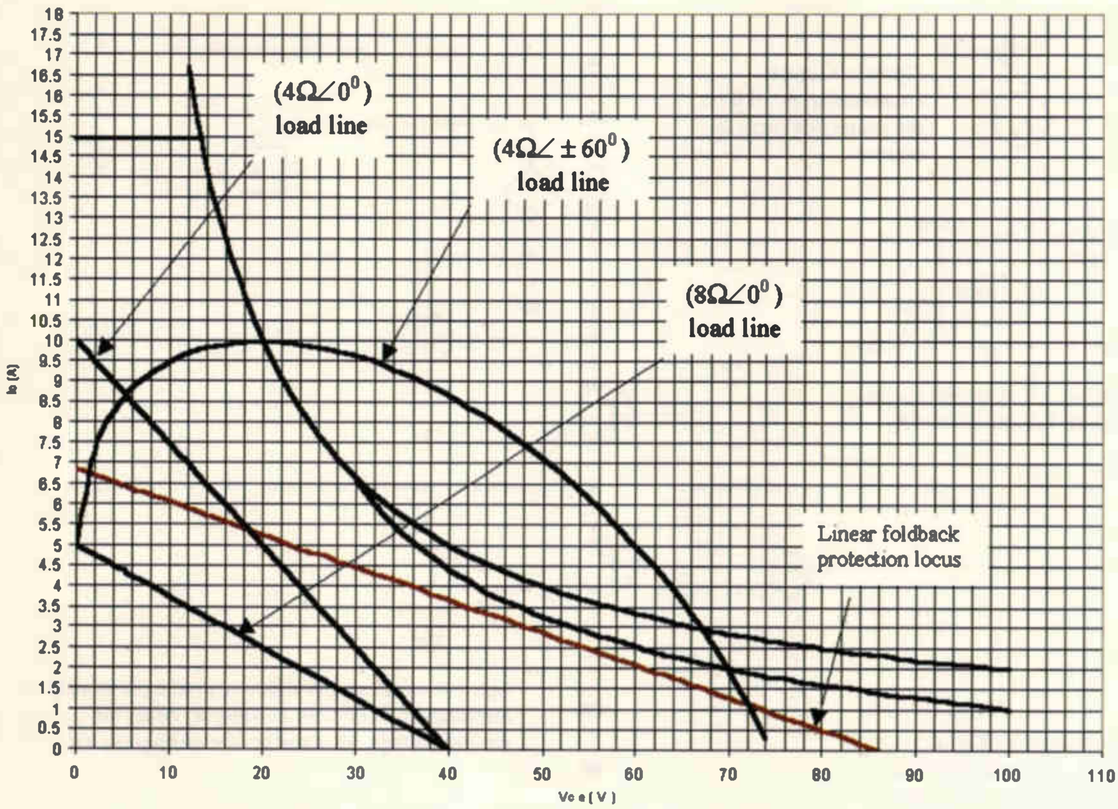


Figure 10: Reactive load gives rise to an elliptical line, resulting in more than seven times greater peak device dissipation than for the (8Ω∠60°) case.

away from the negative supply rail to ≈–5.97V. Similarly, the PNP device, T_{o2}, must sink ≈7.68A when the output swings to +5.97V from +Vcc. Note that the crossover discontinuity in the output voltage characteristic (figure 12), now precedes zero crossing by 60°, at $|V_{out}|$≈35V.

For a (4Ω∠+60°) inductive impedance, in which current lags voltage, the output conditions are reversed, with the load demanding 7.68A from T_{o1} when the output swings from the positive supply to –5.97V. Regardless of the nature of the load however, device voltage, v_{ce}, and load voltage, v_{out}, are always 180° out of phase, and being a voltage follower, the input voltage is always in phase with v_{out}.

The linear foldback protection locus of figure 10 only permits 3.1A at Vce=45.97V, therefore a minimum of three, (ideally four), output pairs are required to drive a notional (4Ω∠±60°) loudspeaker system from ±40V supply rails without intrusive limiter activation. On this basis and using other established techniques[12,19], including D.C. offset, and thermal overload protection, a reliable, low distortion, 100W into (4Ω∠±60°) class-B amplifier may be constructed.

As the cost of power transistors is significant, there is a compelling financial incentive to minimise the number of devices used by utilising the S.O.A as efficiently as possible. To this end it has been suggested[11] that ideally the protection locus should closely match the bounds of the S.O.A. This is unnecessary, as reactive load drive primarily requires that current delivery in the $|V_{cc}| \leq V_{ce} < 2|V_{cc}|$ region be maximised without violating D.C safe operating limits. In general an optimally located, non-linear protection locus with no more than one breakpoint should suffice.

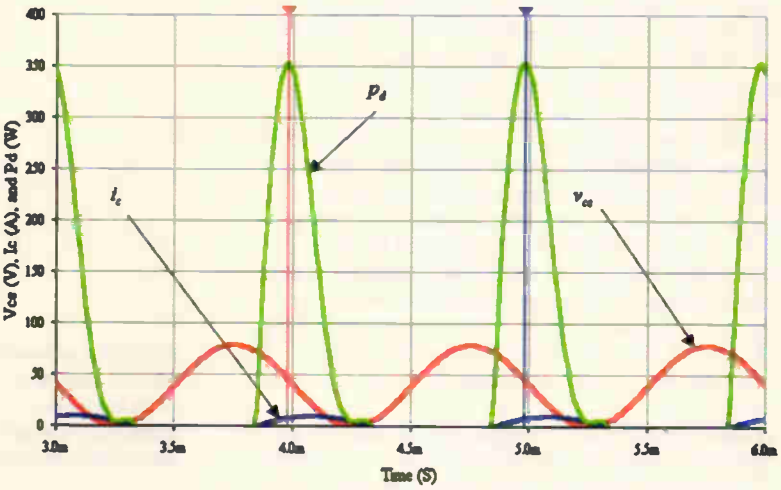


Figure 11: Instantaneous Vce, Ic, and Pd in sourcing output transistor, driving 150W into 4Ω∠60°).

Single slope, single breakpoint non-linear foldback limiting.

Introducing a zero-gradient segment **Fig. 13**, at some optimal point in the protection locus permits the enhancement of current delivery at the low-Vce end of the S.O.A., without significantly compromising available current at higher device voltages. The single slope, linear foldback 'protocol', (equation 4), is made redundant, as the protection locus does not cross the Vce-axis at any point. This scheme is briefly mentioned in reference [20], where it is dismissed in favour of the comparatively

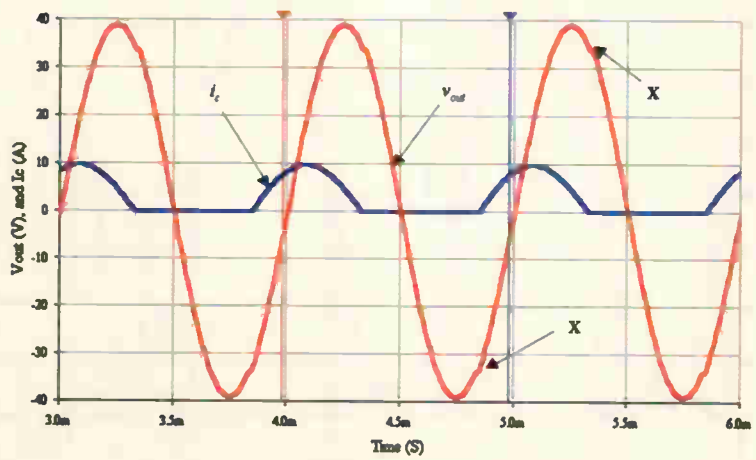


Figure 12: Transistor T_{o2} delivers 7.68A to the (4Ω∠60°) load when output swings away from -Vcc to -5.97V. Note that the crossover discontinuity marked X precedes zero voltage crossing by 60°.

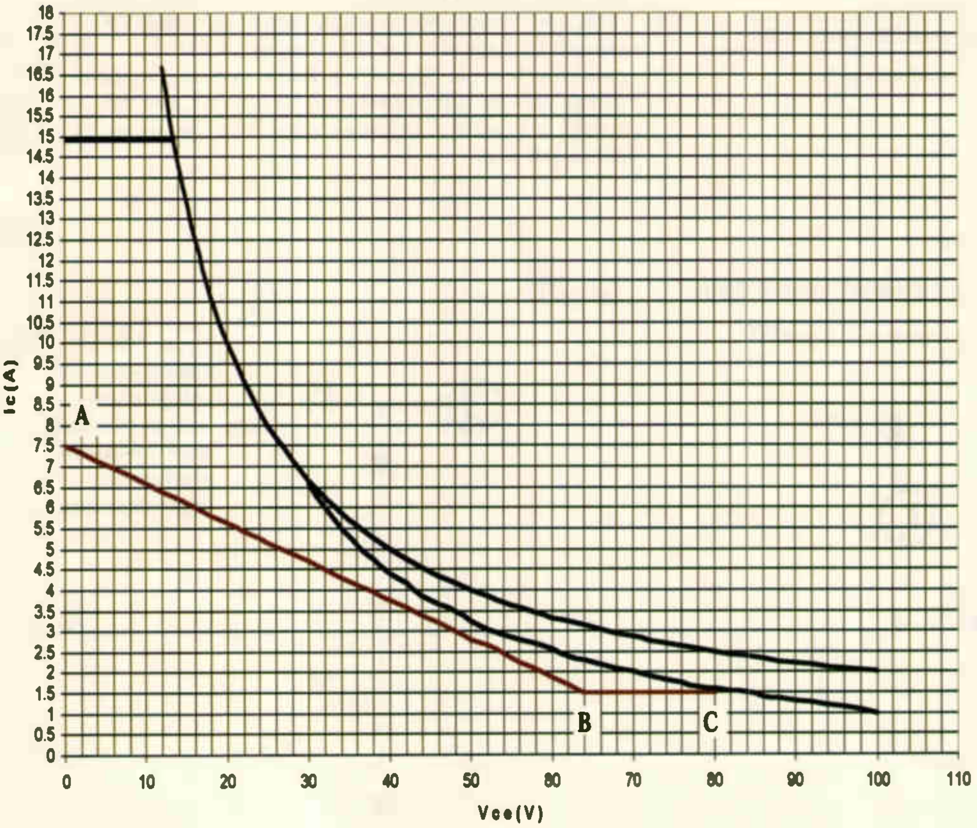


Figure 13: Single slope, single breakpoint non-linear foldback protection locus.

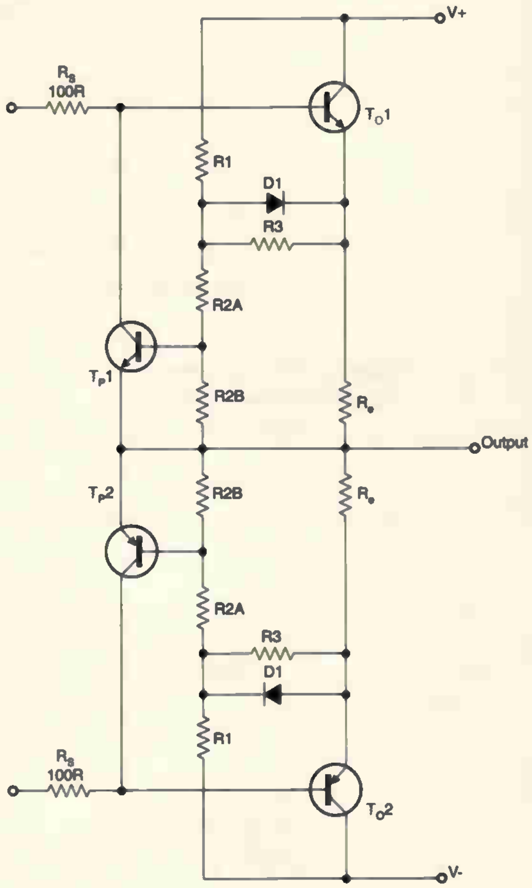


Figure 14: Single slope, single breakpoint non-linear foldback protection circuit as applied to a complementary emitter follower.

inferior single slope, linear foldback method.

The zero-slope segment, B-C, is realised by splitting R_2 in figure 1 into voltage divider, R_{2A} and R_{2B} **Fig 14.**, and shunting R_3 with a fast recovery diode, D_1. The diode applies a constant voltage, $V_f \approx 0V6$ (to a first-order approximation), across R_3, for,

$$\{64V \le V_{ce} < (2|V_{cc}| - 80V)\}$$

Therefore, for,

$$\{64V \le V_{ce} < (2|V_{cc}| - 80V)\}$$

subject to $I_c < 1A$, the diode effectively clamps the voltage across R_{2A}, and R_{2B}, preventing the development of sufficient voltage across R_{2B} to turn on the protection transistor.

However, for,

$$\{64V \le V_{ce} < (2|V_{cc}| - 80V)\}$$

and $I_c \ge 1A$, the increased potential drop across Re with I_c results in a net increase in voltage across R_{2A}, and R_{2B},, inducing a large enough voltage drop across R_{2B}, to trigger the protection transistor.

For ($0V \le V_{ce} < 64V$) the diode is off (open-circuit to a first-order approximation) and the circuit reverts to a linear foldback, single slope regime.

Resistor values are calculated by developing simultaneous equations for segments B-C, and A-B, at points B and A respectively **Figs. 15, and 16**. Resistor, R_1 is selected with a view to minimising diode power dissipation when R_1 and the diode are exposed to the magnitude sum of the supply rails.

With reference to figure 15, let R_1=8K2, and I_d=1mA. Assuming $V_{be} = V_f \approx 0V6$, then:

$$I_1 = I_d + I_2 + I_3 \qquad (6)$$

And,

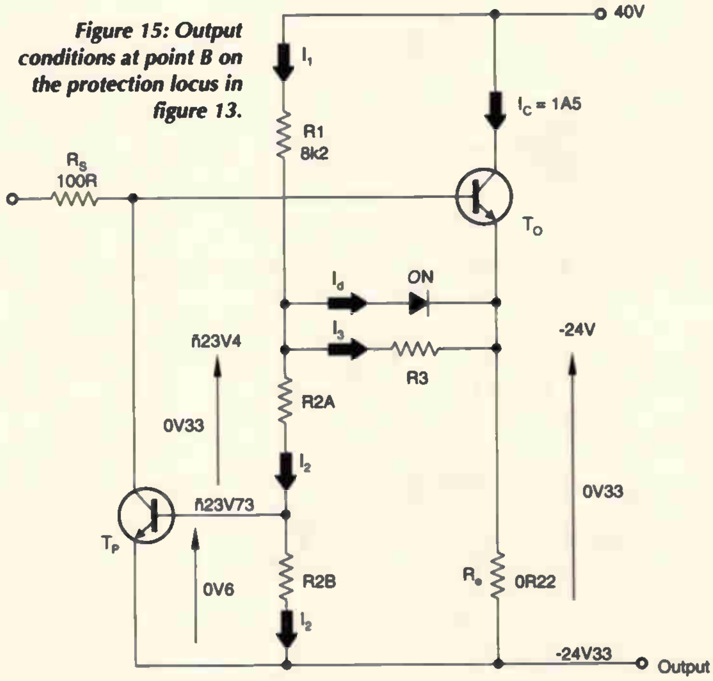
$$R_{2B} = \left(\frac{0.6}{0.33}\right) R_{2A} \qquad (7)$$

From equation 6:

$$\frac{(40 + 23.4)}{8K2} = 1mA + \frac{0.33}{R_{2A}} + \frac{0.6}{R_3} \qquad (8)$$

With reference to figure 16, and invoking equation 7:

Figure 15: Output conditions at point B on the protection locus in figure 13.



$$0.6 = \frac{1.65R_{2A}(0.6/0.33)}{R_{2A}(0.6/0.33) + R_{2A} + 8K2R_3/(8K2 + R_3)} \quad (9)$$

Solving equations 8, and 9 simultaneously:

$R_3 \approx 198R7$

$R_{2A} \approx 88R9$

And,

$R_{2B} = R_{2A}(0.6/0.22) \approx 161R6$

As was the case with the linear foldback locus of figure 20, a minimum of three output pairs is required to drive a (4Ω∠±60°) load, since available current at Vce≈45V97 remains unchanged at Ic≈3A1. However with the protection locus in figure 13, available current per output pair at Vce≈4V is increased from 6A4 to 7A1, and the current at $V_{ce} \approx 2|V_{cc}|$, increases from 0A5 to just under 1A5 per output pair.

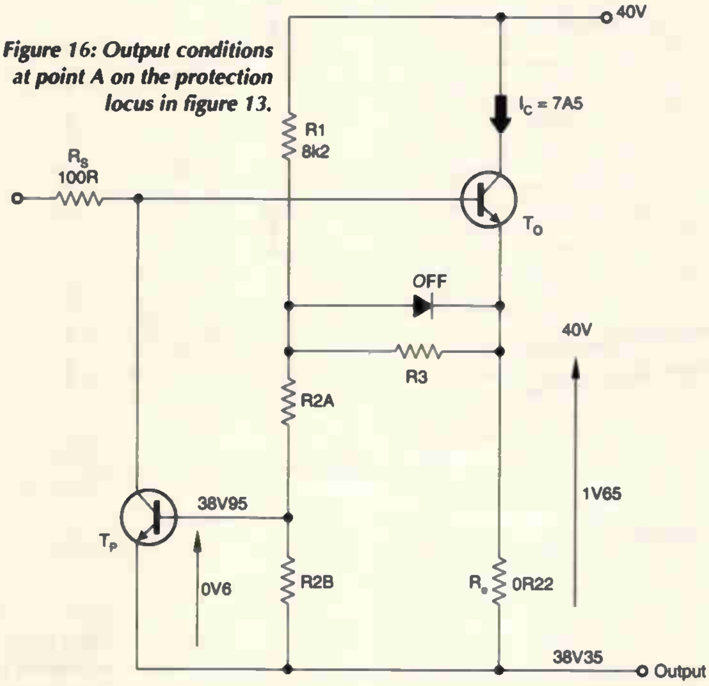


Figure 16: Output conditions at point A on the protection locus in figure 13.

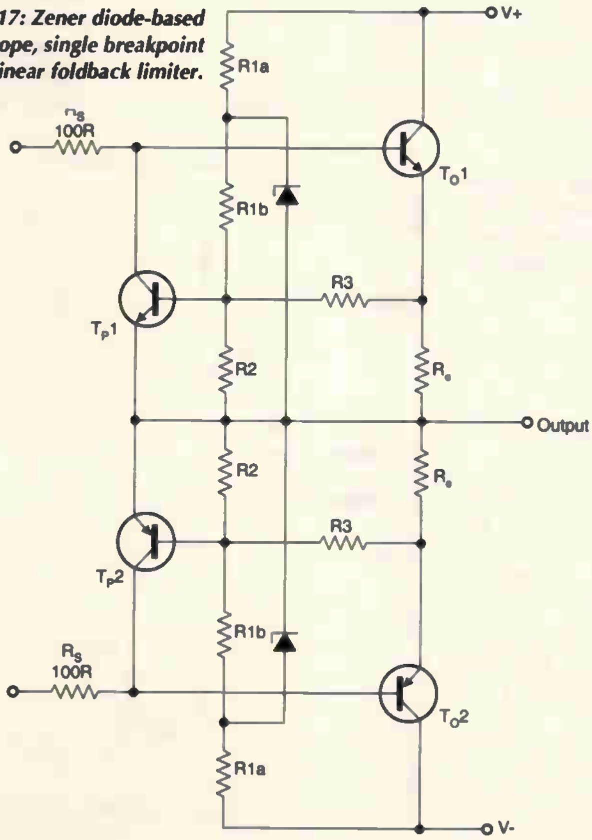


Figure 17: Zener diode-based single slope, single breakpoint non-linear foldback limiter.

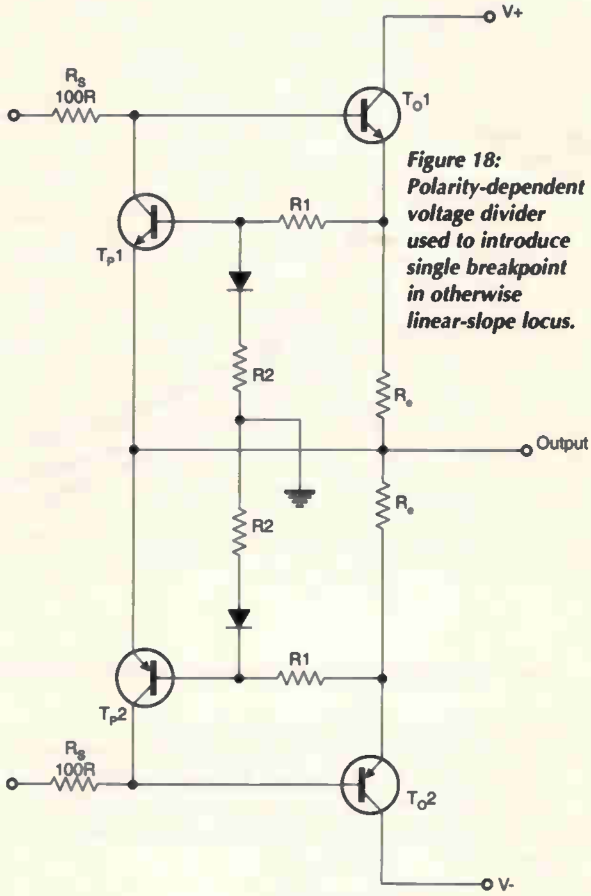


Figure 18: Polarity-dependent voltage divider used to introduce single breakpoint in otherwise linear-slope locus.

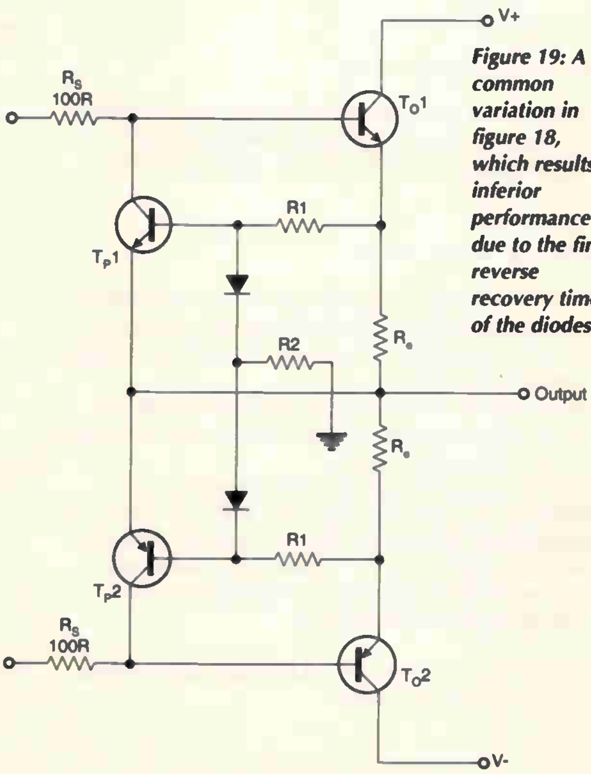


Figure 19: A common variation in figure 18, which results in inferior performance due to the finite reverse recovery time of the diodes.

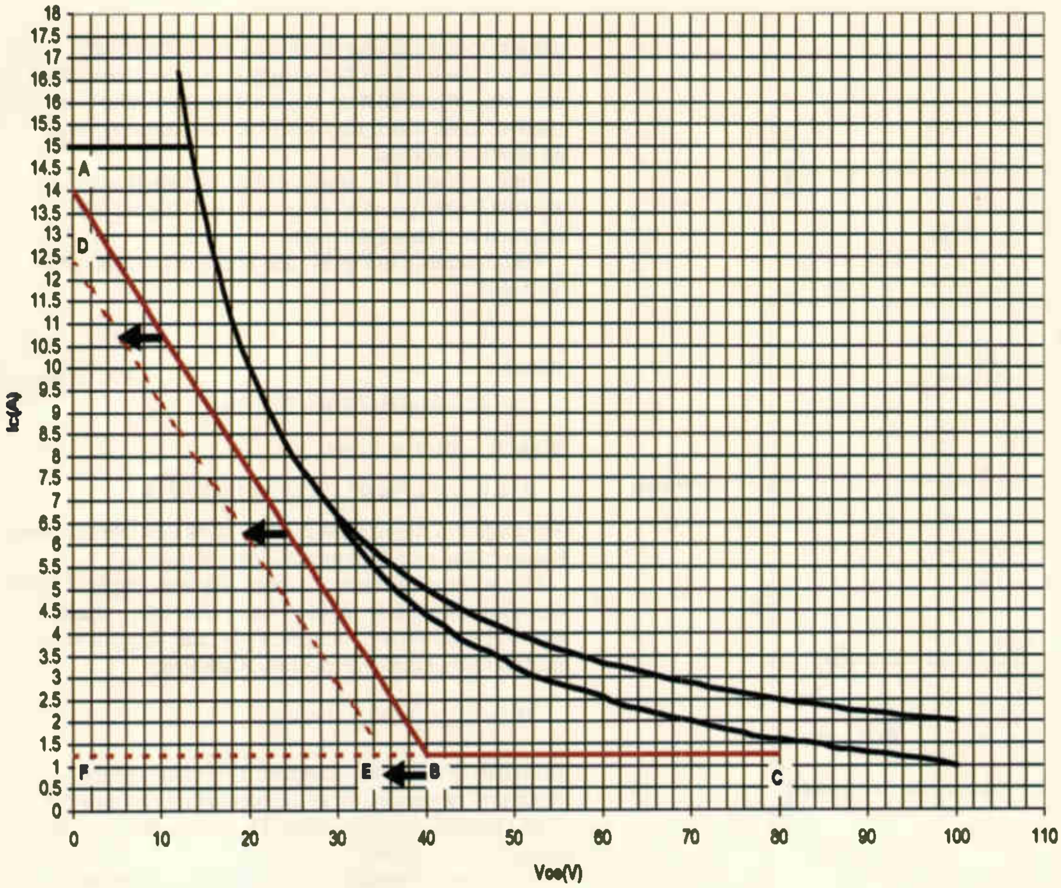


Figure 20: Single slope, single breakpoint, non-linear protection locus described by network in figure 18. A notional 5V drop in the supply rail causes an equivalent horizontal translation of segment A-B to D-E.

Since the locus is non-linear, caution must be exercised to ensure that, while pursuing the secondary objective of enhancing current delivery in the low-Vce region of the S.O.A., available current in the critical higher device voltage region, (i.e., $|V_{cc}| \leq V_{ce} < 2|V_{cc}|$), is not simultaneously compromised by the location of the breakpoint.

The circuits in **Fig. 17, and 18** are frequently used[10,21,22], to realise single slope, single breakpoint non-linear foldback protection. The zener diode in figure 17 is used to establish the flat portion of the locus. This is a rather unsatisfactory solution as in practice the zener breakdown voltage[23] can vary about its nominal value with current by as much as 25%, as the diode is driven in and out of breakdown, as is the case here. The breakpoint in the protection locus would therefore be ill defined in practice, significantly enhancing the potential for S.O.A. violation.

The more dependable p-n diode is used (figure 18), to effect a single slope, single breakpoint regime by means of a simple, voltage polarity-dependent divider[24]. However this scheme (beloved of American manufacturers) is sub-optimal with respect to flexibility in breakpoint placement as diode commutation can only occur at Vout≈0V, (i.e. $V_{ce} \approx V_{cc}$), so that the nominally zero-slope portion of the locus is solely defined by the voltage drop across R_e being equal to the protection transistors base-emitter voltage, Vbe.

The locus in **Fig 20**, requires a nominal R_e=0R47, more than doubling gain-step distortion[12,pg256] generated by a class-AB amplifier, relative to the circuit in figure 14 for which R_e=0R22. A smaller value for R_e cannot be employed as this would result in a commensurate and necessarily unsafe vertical displacement of segment B-C. Thus segment B-C is fixed for $|V_{cc}|$=40V, and results in even more inefficient S.O.A usage in the crucial $|V_{cc}| \leq V_{ce} < 2|V_{cc}|$ region than the compromised single slope,

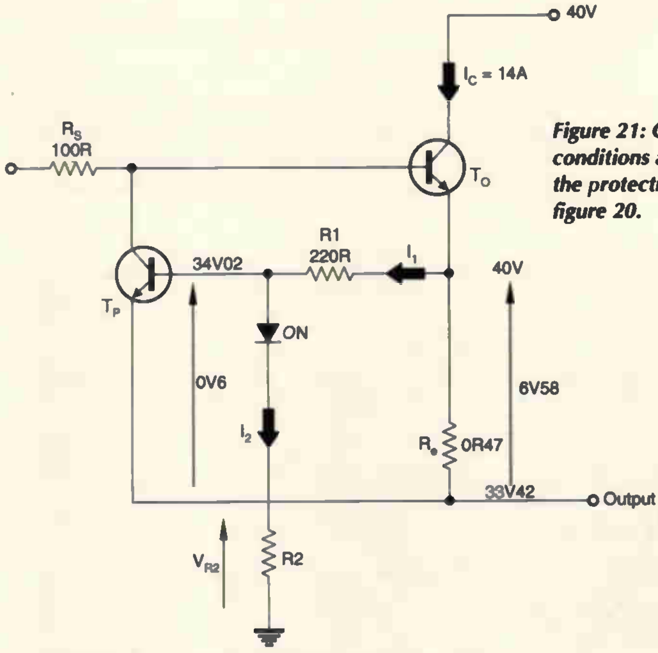


Figure 21: Output conditions at point A on the protection locus in figure 20.

linear foldback arrangement in figure 5.

Further, using a fixed reference voltage (zero volts in this case) independent of the floating collector-emitter voltage, V_{ce}, as the basis for VI protection is rather optimistic, as it presumes equally invariant supply rails that do not sag under load. A nominal 40V supply rail which sags by 5V under load would effect a 5V horizontal displacement (figure 20) of segment A-B to D-E. Conversely, a primary supply surge could cause a potentially disastrous horizontal translocation along B-C of segment A-B into and perhaps well beyond the transistor's S.O.A limits.

Since the diodes in figure 18 are in theory never forward biased simultaneously, the modification in figure 19 is often adopted[19] in what may at first appear to be an elegant

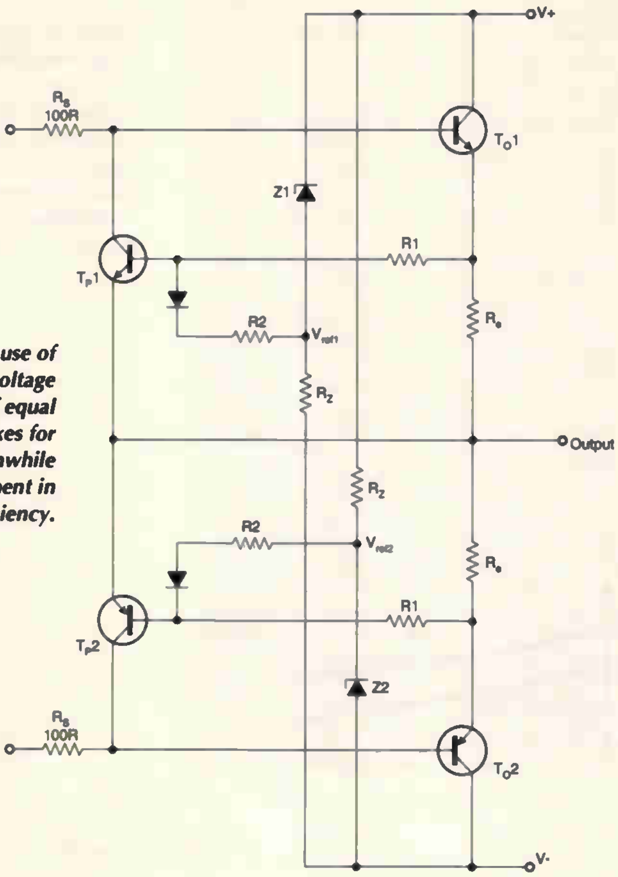


Figure 22: The use of arbitrary voltage references of equal magnitude makes for a worthwhile improvement in efficiency.

simplification. The excision of one of the resistors in this fashion is alas a false economy at best, as the performance of the circuit is now significantly compromised by the finite reverse recovery time of the diodes, with minority carrier storage causing the diodes to conduct briefly when reverse biased. This often results in minute, intermittent zero-crossing oscillation at the output, particularly with a reactive load, which may easily be misdiagnosed as crossover distortion.

Since segment B-C is established by merely selecting R_e=0R47, only point A on locus A-B-C (figure 20) is required to obtain a solution.

With reference to **Fig 21**, and letting R1=220R, and Vcc=40V,

$$I_2 \approx I_1$$

Where,

$$I_1 = (40 - 34.02)/220R \approx 27.18mA$$

With $V_f \approx 0V7$ at 27mA,

$$R_z = V_{R2}/I_2 \approx (34.02 - 0.7)/27.18mA \approx 1K2$$

The circuit in figure 18 is capable of modest improvement, however and therefore merits closer scrutiny.

This scheme can be made more efficient, (figure 22), by changing the voltage dividers fixed reference voltage from zero to two arbitrary voltages, V_{Ref1}, and V_{Ref2}, of equal magnitude but opposite polarity, such that;

$$\left\{0V < \left(|V_{Ref1}| = |V_{Ref2}|\right) < |V_{cc}|\right\}$$

Nominal 40V rails are assumed. This enhances the flexibility of the circuit, as the breakpoint can now be freely located along C-F, **Fig. 23.**, giving rise to a more efficient locus, B-E-F.

The reference voltage is generated by a zener diode, which in contrast with figure 17, is acceptable, as the current established by the diode's current limiting resistor, R_z, is reasonably constant, which makes for a substantially invariant voltage drop across the diode. A depletion mode MOSFET configured as a current regulator could be used instead of R_z to firmly establish quiescent conditions. This is expensive, and therefore probably unjustifiable in a commercial unit.

The reference voltage, is equal in magnitude to the output voltage, V_{out}, at the breakpoint in locus B-E-F, **Fig. 24.**, i.e.:

$$|V_{Ref1}| = |V_{Ref2}| = |V_{out}|_{V_{ce}=60V} = 20V6$$

with V_{Ref1}=–20V6 and V_{Ref2}=+20V6. This calls for a nominal 60V6 zener diode. It is recommended however, that the required voltage drop be realised with multiple low-voltage devices, ($6V \le V_z \le 12V$), as these posses a significantly lower series impedance[23]. Therefore Z_1 and Z_2 may in fact consist of six Motorola 1N5240B 10V zeners, in series with a forward biased 1N4148 diode, the whole quiescing at a nominal 10mA established by R_z

Crucially in figure 22, the cathode of diode Z_1 is connected directly to $+V_{cc}$, effectively bootstrapping V_{Ref1} to the supply rail, so that any anomalies on the supply are directly impressed on the reference voltage. This substantially eliminates the potentially fatal tendency of segment B-E to migrate back and forth along C-F with non-ideal supply rail variation. Similarly V_{Ref2} is bootstrapped to the supply rail by connecting the anode of Z_2 to $-V_{cc}$.

With reference to **Fig. 25**, and taking R_1=22R, and V_{cc}=40V:

$$I_2 \approx I_1$$

Where,

$$I_1 = (40 - 35.9)/220R \approx 18.64mA$$

With $V_f \approx$0V7 at 20mA,

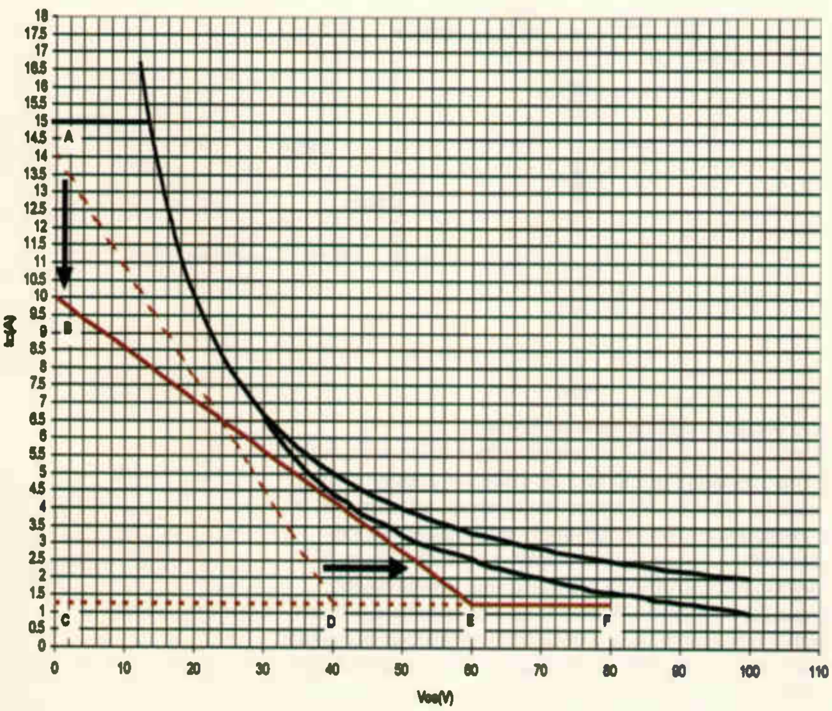


Figure 23: Improved single slope, single breakpoint locus, B-E-F, realised by using an arbitrary voltage reference.

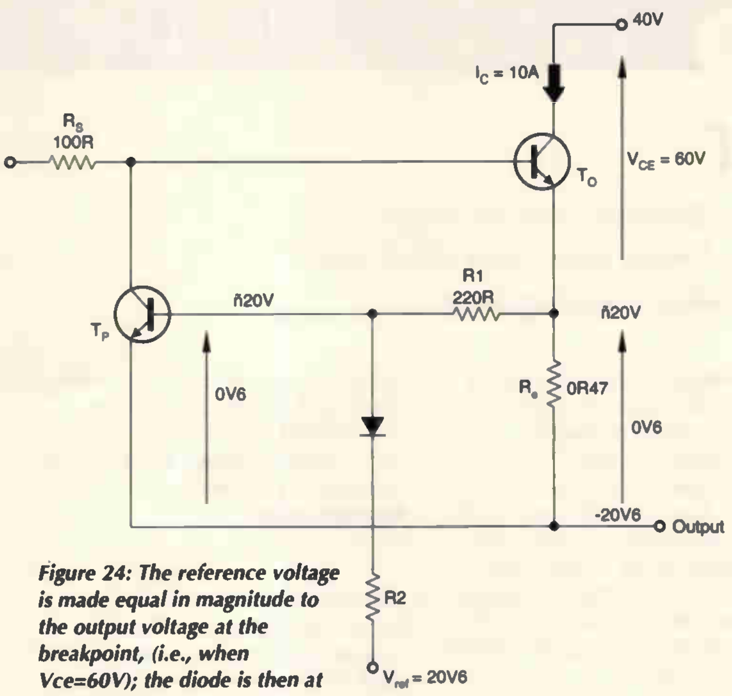


Figure 24: The reference voltage is made equal in magnitude to the output voltage at the breakpoint, (i.e., when Vce=60V); the diode is then at the threshold of conduction.

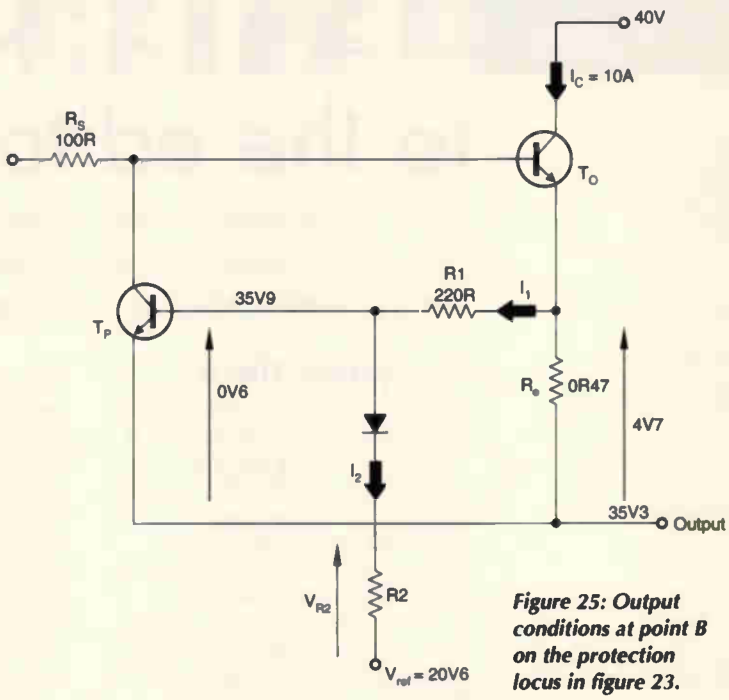


Figure 25: Output conditions at point B on the protection locus in figure 23.

$$R_2 \approx V_{R2}/I_2 \approx (35.9 - 0.7 + 20.6)/18.64mA \approx 3K0$$

The dependence of segment E-F on the value of R_e for the circuit in figure 22 remains its Achilles heel. The singular advantage of the network in figure 14 therefore, is that it permits the arbitrary location of a breakpoint in the protection locus without undue reference to the value of R_e.

Moreover, because the entire network of figure 14 floats between the supply and output rails, the position of the locus in the S.O.A remains resolutely invariant in the face of deviant power supply behaviour, without recourse to a bootstrapped voltage reference. The accuracy of such a reference is necessarily compromised by its dependence on zener diodes, which are only available in discreet, preferred values. ■

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LETTERS to the editor

Letters to "Electronics World" Highbury Business Communications, Anne Boleyn House, 9-13 Ewell Road, Cheam Road, Surrey SM3 8BZ
e-mail j.lowe@highburybiz.com using subject heading 'Letters'.

Conspiracy theory

The town of Valley, near Munich, Germany, has petitioned the Federal District Court in Washington to remove the nuisance of Radio Liberty, which has had the less than satisfactory result of the German government deciding not to renew its lease which expires in 2005.

This has happened because almost half the citizens of the town now have cancer and they have nothing else to blame it on but the powerful transmitters of Radio Liberty, aimed to reach Eastern Europe and Central Asia. It has been there for some decades, but I take leave to doubt if there is even the slightest necessity for it these days.

You notice that citizens come second in this ruling, else the station would be closed immediately. The same phenomenon comes up in this country with regard to toxic hazards from rubbish tips and incinerators. And of course with GM plants. I don't have to drag in, presumably, the chemical town of Cubatao in Brazil, where there are no winds and infant mortality is variously quoted at 27% and 43%, to suggest that just as Tony Blair hails science as our benefactor, it is turning into something else.

Not quite irrelevant is the scare in America over someone supposedly prepared to spread radioactive material around the place. What sort of person would do a thing like that? Well, the Americans did it over 1955-65 with remarkable efficiency, halting the decline in infant mortality there and also in the UK. Thanks to the jet stream and the prevailing winds, they covered the whole contiguous United States bar Vermont.

That people are systematically betrayed is shown by the encouragement of mobile phones in this country. They encourage brain cancer and leukaemia but these may not be seen for 25 years. Though the matter can hardly be researched, the case of Valley should be borne in mind. So should the man whose brain was affected by using his mobile 5 hours a day for propagation studies by BT. He forgets to take the baby upstairs at night, can no longer play the guitar, and puts his speech together quite hesitantly; he can no longer work.

I have always brought a certain pessimism to environmental and occupational health issues but on investigation things have always turned out to be worse than I dreamt.

Looking back at the last century, the first thing I seem to see is asbestos dust covering a whole district in Nottingham, a great hazard that never should have been permitted.

Bernard Jones
London SW1

Editor's comments:
I think Bernard has struck a chord with this one. Some other things to think about are the hazards of living and/or working under power lines. There was in fact a report about this in Wireless World over 20 years ago and a recent report stated that there was in fact no risk. I would be interested to hear other readers' views on governments hushing up scientific facts of life, something we as engineers should be trying to stop.

500Mhz sampling front end - June 2002 issue

You have probably received a fair number of comments from other readers concerning Mr Hickman's interesting and informative article. Nevertheless, I thought I would write to you with an observation of my own.

With regard to producing shorter gate 1 sampling pulses, I suspect that the avalanche pulse generator employed has already reached the limit of its capability in this direction. Some improvement may be indeed be achieved by using a shorter delay line, L1, and a transistor having a higher transition frequency than the BFR91. Unfortunately, this will inevitably be at the expense of pulse amplitude, since most commonly available low cost transistors with higher transition frequencies also tend to have lower avalanche voltages. Therefore, it might be worth considering an alternative method of generating shorter sampling pulses. The method I have in mind is a variation on the theme of the classic step recovery diode (SRD) impulse generator. However, instead of employing an SRD - which is an unusual device that readers are unlikely to find in the majority of mainstream electronic component distributors' catalogues - try using an inexpensive and readily available PIN switching diode. With suitable biasing some short lifetime epitaxial PIN diodes exhibit behaviour very like that of SRD's. For example, Agilent Technologies' HSMP-3820 PIN diode or similar would probably make a suitable candidate for experimentation. In principle it should be possible to generate sampling pulses, having sufficient amplitude, of around 300ps or less using this method.

Douglas R. Taylor
(By email)

M. G. Scroggie

M. G. Scroggie was a well known contributor to Wireless World for many years, writing under his own name and as 'Cathode Ray'. He was the author of many books, such as the Radio Laboratory Handbook and also compilations of his WW articles. A technical biography of him and his work would be a fitting tribute and make an interesting article for EW. So, is there anybody out there with sufficient information and knowledge of him to write it?

PS. Noting the current article on 'super-regens' in EW April 2002, Scroggie wrote at length in WW and in his books on their operation and showed waveforms very similar those in this article.

Howard Miles
Melton Mowbray
Leicestershire, UK.

Any takers? – Ed.

More PCB ideas.

The May issue of your excellent publication has just been received here in the Antipodes. The article by Cyril Bateman on PCB-making was, like all of Cyril's articles, excellent. Cyril suggests some simple ways of producing boards but there is another, apparently little known or used, technique for producing one-off PCBs.

For several years, I have been producing boards by direct-plotting to the copper-clad stock, then etching etc. as usual. Boards of at least the complexity of the examples shown in Cyril's article and, often of rather greater complexity are easily and quickly produced.

I usually use Easytrax for design (antiquated, I know, but adequate for most applications and far more intuitive than the Windows-based programs which I also have) and plot-out to a HP7475A plotter. I also use a HP7550A on occasions, but other types of plotter (e.g. Roland) should be useable. A plot on paper is first made, then a suitably-sized piece of copper stock, cleaned as below, is taped-down by its edges so as to cover the plot. The pen is changed to an etch-resist type and the plot done again onto the copper. The board is then etched etc.

Some tips for those who wish to try this technique; Staedtler Lumocolor 313 pens, made for overhead-projector use, make excellent (and cheap) etch-resist pens; I think the black ones are best, although I have used the red ones successfully. The pen can be mounted with tape in a drilled-out body from a standard plotter pen so that the machine can pick it up. Cleaning the board with a 'polishing' motion, using very fine 'wet & dry' paper and a drop of detergent under running water, seems best as it gives the surface a pattern of tiny 'scratches' which help the pen make a clean line.

Occasionally a track or two will require retouching with an etch-resist pen, but I think that this is due to carelessness on my part in cleaning the copper. In this application the necessity for preparing the copper surface properly is vital and is at least one level of importance above the usual. Watch the orientation of the printout as it may be necessary to 'flip' the artwork to avoid a mirror-image product, depending upon whether your board is SMD or through-hole. Of course I have never felt like a fool after making a mirror-reverse board and then trying to assemble it!

The reference below for Easytrax is worthwhile, both for the support given for this 'obsolete' program and for other information that the delightful Bob Barnes makes available. Here in Australia plotters are to be had for the asking; try your local drafting supplies shop or building plan draftsmen.

With regard to Cyril's notes re: drills, I have found it unwise to use carbide drills in a hand-held machine as they snap very easily. A drill stand is to be preferred; I use a Dremel but I think it a little crude and am saving up for a 'proper' unit!

Kerry Power,
Coffs Harbour
Australia.

References;
Easytrax;
www.cia.com.au/rcsradio/ Drills;
Toolsandtunes (Joseph Moran) on Ebay has a good range of drills

Exposure

In my experience exposing ether side separately invariably ends in tears, or at least misalignment. So here is my double-sided method.

A pair of 500W halogen flood lamps, with glass removed, are good U.V sources for less than five pounds each, at 2'6'' (750mm), defined by the height of the shelves in my kitchen, results in a 15 min. exposure time. One lamp is placed above and one below a pair of shelf brackets on which a polycarbonate carrier simply rests on. The polycarbonate carrier is simply two pieces of non-U.V stable polycarbonate hinged at one edge with 'gaffer' tape and clipped shut with 'bulldog' clips.

Simply take top foil and trim approx. 1''(25mm) off 3 edges, then align it against the bottom foil on a light box and secure with masking tape along the three trimmed edges, forming an envelope that the P.C.B board is placed inside, that is in turn placed inside the polycarbonate carrier. Thin cardboard strips can be used to centre the edges (vertically) of the envelope in the sandwich. The whole assembly then rests on the shelf brackets and is exposed.

Jo Atkin
(by email)

Another UV source

I would like to add the following information to the recent articles from Cyril Bateman about printed circuit making that appeared in EW.

A powerful UV source for photo-resist exposure and also for EPROM erasing, can be made from medium-pressure mercury arc lamps, such as the Philips HPL series. They consist of a discharge tube containing mercury vapour, supported within a glass bulb, coated on the inside with fluorescent material and have metal rods, which also double up as current leads. The space between the discharge tube and the bulb is filled with an inert gas, probably nitrogen. By carefully breaking the neck of the bulb around the base, it can be removed and the exposed discharge tube used by itself, in the same socket and with the same current limiting inductor as the whole lamp. This tube emits strongly in the 436 and 365nm mercury lines and also in the visible 546nm line. Two safety considerations apply: firstly, the discharge tube is made of fused quartz and gets quite hot when running. So, it must be handled with the same care as an halogen lamp bulb, that is, no organic contamination shall be allowed on its surface. Secondly, the tube support rods are now exposed and at mains potential, so they must not be touched by hand or conductive objects while the lamp is on. These lamps are being replaced in outdoor lighting by metal arc lamps, but are

EW improvement ideas

I think the magazine might well benefit from a change of direction (larger page count, lower cover price, rename to Wireless World, OK only joking!) although I suspect there is a loyal core of readers who like it more or less as it is. Personally, I feel the readers' letters are the best part, especially when there is violent (but well argued) controversy. I also feel there should be a historical feature every (or nearly every) issue, as this broadens the appeal and crucially differentiates the magazine from so many others. You don't want to become just a clone or look-alike of every other electronics mag, as this is a highly competitive arena and would be dropping the old values in any case.

Andy Emmerson
UK

A quick comment on EW content recently....

One thing that's rather annoying, is when several pages of a mag are taken up with software listings. I think that these days it's highly improbable that anyone thinking of building a project won't have internet access - if not themselves then via someone they know. Stuff like this belongs on a website so it can be downloaded, error-free by those who want it. There's no need for an all-singing flashy corporate site - just a URL in the article to get the file. Obviously there may be instances where parts of a software listing may be useful or informative, but the inclusion of full listings is just wasteful and unnecessary. The inclusion of two pages of HEX listing on June 2002's issue was particularly ridiculous, and I would question whether this was of any use at all to any single reader! It served no purpose to illustrate the article, and I doubt anyone would think of typing it in! OK that issue was pretty thin but if that's all you could find to pad it out things must be getting pretty desperate!

Mike Harrison
UK

Suffice to say that in future code will be available by email in the short term and I'm hoping to get a rudimentary website up in the not too distant future. There won't be anymore HEX listings........unless they are VERY short!
Ed.

still quite easy to find and not expensive, at least here. I think RS carries them in the UK.
Jose A.Senna
Brazil

Electrocution I

The June and July issues of Electronics World included letters regarding the safety of the current in an EHT circuit which appeared in the May issue. I think that further information is required to clarify the physiological effects of 50 Hz currents and direct currents.

The authoritative document is IEC 60479: Effects of current on human beings and livestock. The following values are from IEC (60)479-1, Third edition,1994.

Italicised text are my notes.

(i) There is considerable variation between individuals.

(ii) D.C. is allowed to include up to 10% ripple.

(iii) Requirements for different frequencies, some non-sinusoidal waveforms and charged capacitors are covered separately.

Threshold of reaction: 0.5mA at 50Hz; 2.0mA d.c.

Reaction can be dangerous, e.g. in confined spaces or on ladders.

Threshold of let-go: 10mA at 50Hz.

For 50Hz above, the let-go threshold is where the victim may be unable to release grip on a source of current. The threshold is lower for women and lower again for children.

For D.C., only making and breaking of the current leads to painful and cramp-like contractions of the muscles.

Threshold of Ventricular fibrillation: 40 - 500mA at 50Hz; 140 - 500mA D.C.

Values depend on exposure time (inversely), current path through the body, state of health, etc.

Ventricular fibrillation is considered to be the main cause of death from electric shock.

Note. IEC publishes product safety standards covering most aspects of safety. These are usually adopted (sometimes with modifications) by the European organisation for electrical standards, CENELEC, and published by BSI. Probably of most interest to EW readers are:

BS EN 60065 - Safety requirements for audio, video and similar electronic appliances;

BS EN 60950 - Safety of Information Technology Equipment (this includes a small section on safety principles); and

BS EN 61010 - Safety requirements for electrical equipment for measurement, control and laboratory use.

At well over £100 each, hobbyists may be put-off by the price. If so, it may be possible to gain access via a library or college.
Ted Smith
UK

Electronics as an occupation

People who are self-employed, e.g. doctors, electricians, plumbers, etc., do tend to earn more than electronic professionals who mostly work for a salary. It is difficult for a design engineer to work for himself, but we (in the electronic profession) could start, manage and own more electronic manufacturing companies ourselves. Those who are self-employed could learn from medical doctors. For example, if you have a repair shop, let your customers wait, to make sure you always have work. Charge them for fault finding, regardless of whether you can identify the problem or fix it. If your repair operation turns out to be failure, still charge for it.

In a free market economy, salaries should be determined by supply and demand. If electronic professionals are not happy with their pay or status, they should seek other occupations. Students tend to take courses that interest them, without considering the demand for such qualifications. Electronics is an interesting subject, and unfortunately too many people think so. Many occupations increase their income beyond the free market level by having a union recommend fees and by reserving work. For example, even if you know all about the prescribed medicine you need, you cannot get it without paying a doctor for a consultation (The patient usually has to discover the side-effects for himself). Nor can you get the results of some medical tests without seeing a doctor. Only a certified electrician may work on the electrical wiring of your home (at least in some countries). Homes may only be sold by a registered estate agent. Engineering associations should recommend salary levels. If all else fails, consumers could be required to obtain a 'prescription' from an electronic engineer or technician to buy electronic equipment. After all, we need to make sure that the equipment is right for them. It could also be dangerous in the wrong hands, e.g. a microwave oven used incorrectly.

Electronics is too successful. If the medical world were able to design medicines which could cure any problem (not just treat the symptoms) and if the medicines were imported cheaply from Asia and sold at supermarkets, doctors would have little work. If laws were written more clearly and decisively (making a court judgement unnecessary), there would be less demand for legal professionals. We need to design products that fail more often and go obsolete. Microsoft has demonstrated how you can force the public to continuously buy new hardware and software. Just make the operating system ever larger, requiring a bigger, faster computer, and convince the peripheral equipment manufacturers to only supply drivers for the latest Microsoft product, and you've got it made. For those who are in doubt, this letter is not be taken entirely seriously.
Dewald de Lange
South Africa

PS. Through the decades that I have read Electronics World, you have had some very unique and well illustrated articles. With electronics becoming more integrated, and dominated by large international companies, the demand for electronic magazines is likely to shrink. Let's hope Electronics World survives.

With more readers like you, I don't think we'll have a problem. Ed.

Electrocution II

Re: Switching EHT generator for ES loudspeakers (Circuit ideas) and Shock Hazard Reduced (letters).

Whilst the shock hazard of Mr Vincent's EHT circuit on the bench may be seen to be minimal, would Mr Bennett say the same for an infant crawling around the back of an electrostatic loudspeaker and attempting to put 'something interesting' into its mouth?

For example, the internals of a Quad ELS-63 (lots of them around - I have a pair myself) have warning notices "DANGER - VERY HIGH VOLTAGES - TAKE EXTREME CARE" and the EHT circuit includes a 10M resistor and a neon indicator in series at the output.

I would suggest that the simple and prudent expedient of a general warning for EHT supplies would not be aims, particularly in a 'Circuit Ideas' section where only a minimal description is given and where substitution of components that happen to be to hand is quite possible. I also note that the Quad ELS-63's four stage voltage multiplier uses 10n and 20n capacitors (versus Mr Vincent's three stage and 1n capacitors), which might be retained and the switching circuit 'tweaked' to drive it (maybe without changing the turns ratios of the transformer).

I accept that my concern may be seen as to be somewhat 'in extremist', but the EHT generator circuit was put forward for electrostatic loudspeakers which would normally be sited in a domestic environment. Furthermore it is likely to be built by persons interested in audio electronics but not necessarily familiar with voltages above the (?)60V or so used in audio power amplifiers. One can experiment with line level audio circuits with relative impunity (one is probably at greater risk of burns from a soldering iron than anything else), but multi-kilovolt EHT DC power supplies are another matter and should always be treated with 'Extreme Care' (even if they are theoretically only able to supply a few mA).

Since it is now understood that once past the skin barrier electrical currents can travel easily within the body favouring nerves and blood vessels as conductors, I would have thought that the safety aspect of EHT supplies would be the first concern in any design. And one perhaps shouldn't overlook the modern reality that without a warning (i.e. showing due diligence) it is possible that both Electronics World (I have been reading this learned journal for over thirty years) and the author might be seen as negligent should an injury or fatality occur directly or indirectly because of this circuit.

I would also note that many dwellings are not fitted with RCD protection (mine is, but only on the 13 amp power sockets - not on the lighting or other circuits) so Mr Bennett's comments on the protection afforded by these devices should be taken in context. And I remain unconvinced as to the safety of touching a television's EHT supply, particularly one in situ! Nowadays all such supplies that I have seen are encased in metal shielding with big 'DANGER' signs and red flashes, unlike the old valve televisions where that bit was just in a corner (and parts of the metal chassis was liable to be live too!)

Susan J.L. Parker AMIEE
London, UK

Lightning

I enjoyed Joe Carr's lucid exposition in the June issue on how to observe lightning and its effects.

Long term readers of EWW may recall a feature (Lightning Strikes!) I wrote on recording the polarity and waveforms of lightning strokes. As a geophysicist, I have been recording variations in the earth's electric field for many years, and would like to add some points to Joe's feature.

For the last 20 years I have used a triode valve (6Q7GT) antenna electrometer (described in the feature) to record the electric field. The valve was replaced once about 10 years ago, when its emission failed. It is easy to get the gigohm input impedances needed using semiconductors, but they fail when exposed to the high voltage pulses from nearby strikes. My observatory has had two nearby lightning strikes which took out the power supply to my earth current geomagnetic field recording system, vaporising the 10A protection diodes across the incoming lines from the earth electrodes, and killing the cordless phone terminal connected to an overhead BT line.

Protecting electronic gear from lightning spikes is tricky. Anything connected to an outdoor antenna or large frame aerial needs to be protected - expensive gear like 'scopes will only take 300-500volts across the input. As well as the obvious direct lightning voltage spike, there is also an earth voltage spike caused by the local dissipation of the stroke energy, which elevates the ground potential nearby and adds to the equipment protection problem. Equipment inputs can be protected by silicon diodes or neon tubes to shunt the induced voltage spike. Inductors or chokes are also helpful to blunt pulses coming in by the back door from the mains. Joe's wire ended neon indicator circuit can be turned into a charge measuring instrument by blacking out daylight to stop neon conduction until the antenna voltage reaches the strike voltage (about 85V). This turns the neon into a sawtooth oscillator whose pulses can be tapped off to a counter or recorder or audio input to give quantitative measurement of incoming charge rate. During thunderstorms, the device conducts almost continuously and positively screams at every lightning stroke. The antenna only needs to be about 40 feet long and about 15 feet above ground. With good insulation, the detector makes a continuous plopping noise in a crystal earpiece at the earthy end of the chain, and is rightly called a 'farter'.

Many books have been published about lightning. Works by Oliver Lodge provide a background to the design of lightning conductors. Watson-Watt (the inventor of radar) and Bainbridge Bell published a treatise on the recording and direction finding of lightning storms using large frame aerials in the 'Applications of the Cathode Ray Oscillograph in Radio Research' as long ago as 1933. More recent works on ionospheric physics and lightning induced whistlers like 'Natural VLF Radio waves' by Okada and Iwai (1988) are worth reading. A trawl through the Web will find hundreds of other references. Above all, heed Joe's advice when recording lightning- it bites!

Anthony Hopwood. *Worcestershire, UK*

Electrocution III

Paul Bennet (Letters, July 2002) in referring to 1mA as the threshold of feeling and that it is the current which is the determining factor in electrical shock hazards is quite correct; however his figure of "1A plus to interfere with heart operation" is a rather dangerous suggestion. Most sources with which I have come into contact, if you will pardon the pun, suggest a much lower value, the usual levels being set out as follows:-

1mA: Threshold of sensation.
5mA: Maximum 'harmless' current.
10mA: Muscle contraction (muscles 'in circuit').
15mA: Loss of muscle control.
20mA + Severe electric shock, depending on path.
100 - 250mA Fatal electric shock (ventricular fibrillation.)
250mA + Severe electric shock, heart stops.

Apparently, patients who receive over 250mA stand a chance of recovering provided CPR is rapidly applied, since fibrillation did not occur. Levels such as these are only a guide, and vary a little according to patient health, and the nature (D.C., 50-60Hz, HF) of the current involved.

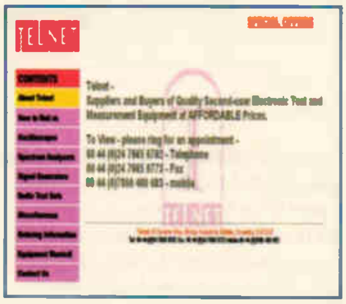
All this is very fine, but the reaction to electric shock (whether due to loss of muscle control or conscious rapid reaction) can cause very much more serious harm than the physical effects of the shock on its own. Serious cuts and lacerations can often result when working on equipment that contains sharp objects - protruding self-tapping screws, glass dials, compartment screening etc.

Stored energy in capacitors does not usually present a problem, for they are normally screened, and in many CRT applications, formed from the coaxial EHT cable itself. However, they are the most dangerous points in most SMPS units such as those used in small computers, etc., retaining typically a 380V charge for a very long time after switch- off. Admittedly the capacitors in the original design by John Vincent present a fairly low series value and these are effectively 'bled away' by virtue of R4, R5, R6 and R7.

My purpose here is not to argue with Mr. Bennet nor Susan J.L. Parker (June 2002), but rather to bring some clarity to the issue of electrical safety. There is a tendency today to think that all things electrical are much safer than they were, especially by the people who actually work on computers, radios, and the like; this must surely be due to the lower voltage rails used (you can touch computer wiring while it is working...!) which when added to the increased equipment safety standards in some manner actually decreases the respect for electricity.

The wall socket is still the wall socket, it's still around 230Vac there, and human physiology has not changed much in the last 100 years. ■

F.C. Trevor Gale.
The Netherlands.



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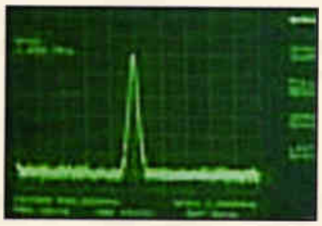

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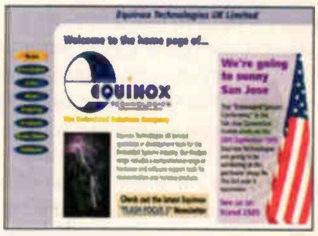

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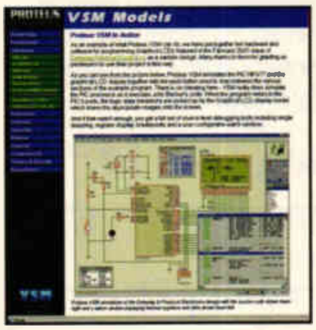

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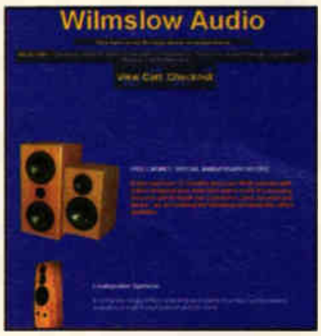

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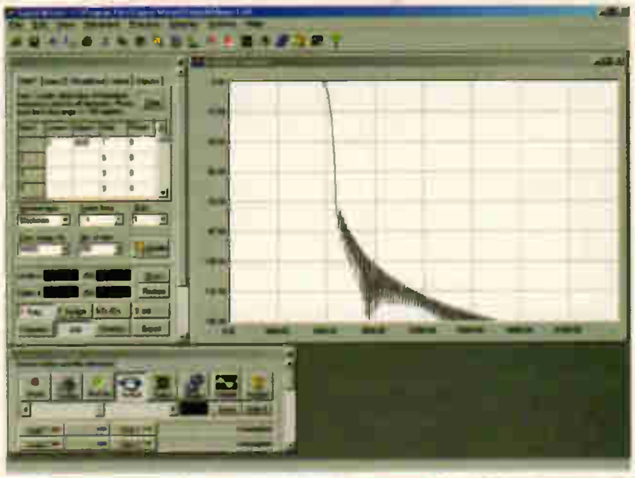

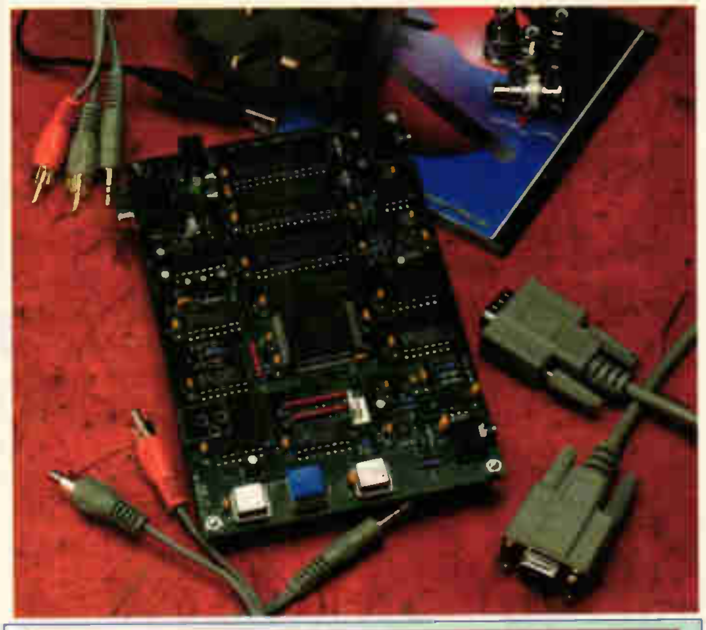

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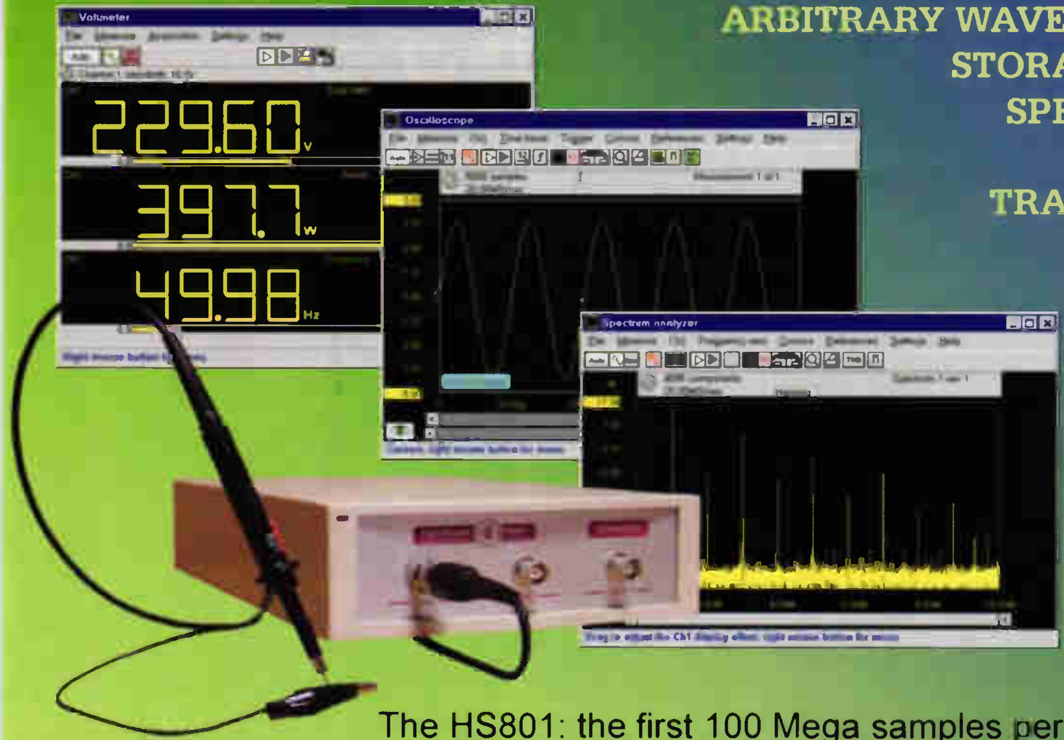

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