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C. Bateman Engineering

CIRCLE NO. 108 ON REPLY CARD
Memory for the masses

There should be a number of benefits resulting from the collapse of the price of d-ram, which has fallen to a quarter of its 1995 price.

Nowadays, d-ram is so ubiquitous that its price affects the retail price of many electronic products – though the one it affects most is the personal computer.

The pc has been a revelation – the most important industrial product of the second half of the century. But how much further does it have to go?

We’re looking at 30 per cent household penetration levels in the West and it could be that all those who want access to a pc, already have it.

But how about the other 70 per cent who don’t want access to a pc? Can anything be done for them?

This looks like a silly question, but the truth is that many people who are at first indifferent to new products - like cars, tvs or mobile phones - often come round to buying them when they become cheap enough and easy enough to use.

At the moment the pc is still the province of the educated middle-class. If it is to have a market outside the educated middle-class it probably has to change its current evolutionary course.

Instead of blindly pursuing faster processing, it may have to start looking at what people want to do with it. Furthermore, if the pc is to become as universal as the telly, it has to be made as easy to use and as cheap as the telly.

But, at the moment, the future of the pc is driven by the supercharged techies of Silicon Valley. To them the absolutely overriding issue is more performance.

To techies, the pc is still a vehicle for fulfilling the same urge that drove the pc’s pioneers – to impress their peers – rather than to make something saleable to the masses.

In the early days of the motorcar, the pioneers similarly vied with each other for more and more power. Enzo Ferrari once famously dismissed W.O. Bentley’s cars as ‘torries’. But then came Henry Ford.

He had a different priority. Ford’s vision was to make cars affordable – cheap enough for Joe Soap to buy.

That was the original promise of the microprocessor which, in the words of its inventor Ted Hoff, “democratised the computer.”

The microprocessor was invented 25 years ago this November. Ever since then, the evolving price/performance ratio of the computer should have been linked to the chip learning curve, i.e. doubling in performance or halving in price every eighteen months.

For a while it did. By now the power of a 1980s pc can easily be incorporated on a single chip. That should mean that the High Street shops are stuffed with affordable computers. But they are not. Instead pcs tend to cost from £800 to £1500. That’s because the chip and computer industry prefer the option of doubling in power every two years rather than halving in price. That way they keep computer prices and chip prices high and margins fat.

And they don’t get any easier to use. The average Joe Soap despairs of the unfriendliness and non-intuitiveness of pc software and finds it a nightmare trying to add on a modem or even connect up a printer. If the electronics industry wants pcs to reach the other 70 per cent of the West’s potential pc market, then pcs will have to be made as easy to use as calculators.

Two things might persuade people to make computers for Joe Soaps. First a number of microprocessors are getting to be powerful enough to emulate the ubiquitous 486, and Pentiums without losing significant performance, and secondly, the cost of d-ram is now low and sinking. So there is a good opportunity to supply that 70 per cent of the Western market for cheap, easy-to-use pcs which the computer industry is currently unwilling to supply.

David Manners
ULTiboard’s interactive strength has always been the major selection criterion of professional Printed Circuit Board designers. Now that every ULTiboard Designer system will be supplied with a SPECCTRA SP4 Autorouter, ULTiboard designers now get the best of both worlds.

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Intel is announcing its 200MHz Pentium this week while privately telling PC vendors that volume production using the processor will only be possible later this summer. This contrasts markedly with previous Intel processor launches which have always shipped in volume once announced. Intel is telling its customers that yields of the 200MHz Pentium are still very low. Sources close to Intel say that the 200MHz Pentium introduction was rushed to prevent Cyrix from getting too much attention with its 200MHz 6X86 P200+ processor (EW, June 5, p3).

Despite the shortage of the Intel device, major PC companies will this week announce 200MHz Pentium-based systems. These include Compaq Computer, Dell Computer, Hewlett-Packard and IBM.

**New infra red camera**

The Jet Propulsion Laboratory and Amber in Goleta, California are collaborating to develop a cooled long wavelength infra red (LWIR) camera. The aim is to design an infra red imaging system based on using a quantum well infra red photodetector (QWIP) and Amber’s 256x256 readout multiplexer. Infra red (IR) detectors made from materials already tested in lasers promise superior IR cameras and night-vision devices. These QWIP devices are made from alternate layers of semiconductor materials arranged atom by atom into layers just a few atoms thick, such as Gallium Arsenide (GaAs) and Indium Gallium Arsenic Phosphide (InGaAsP).

New infra red camera technologies which have the advantage of running on existing telephone lines. "The most compelling reason to use a cable modem is speed, but speed alone won’t be enough to win the battle," said Dataquest industry analyst Lisa Pelgrim. Although ISDN and XDSL will be available to greater numbers of people, Dataquest points to the advantages of cable modems: they can easily transmit data types like video and two-way audio, they offer lower cost per transmitted bit, and also support constant connectivity down the line.
Wireless interactive services – via satellite

Satellite operator Astra plans to make satellite transmission a two-way affair, with businesses and home users having a narrow-band return path over which they can engage in all the interactive opportunities already planned for cable.

Astra plans to make satellite transmission a two-way affair, with businesses and home users having a narrow-band return path over which they can engage in all the interactive opportunities already planned for cable.

Sky, have been restricted to the prospect of a fairly crude version of interactivity relying on a telephone link between the user and the service centre at the broadcaster's uplink. Users are restricted by the low bandwidth of telephone lines and the geographic limitations of cable penetration.

Few technical details are as yet available from SES. Questions raised at the conference – but not answered – involved the design and power of domestic satellite systems which also operate as transmitters, whether users would need radio licences, and how the transponders would cope with thousands of signals from different users. On this last point, the best guess was some form of cellular multiplexing, as with mobile phones.

It is perhaps significant that the newly-appointed technical director, Aldis Grinbergs, who was with Astra until 1988, returns after five years with Motorola, working on its Iridium satellite-based mobile communications project.

Astra has also announced a second orbital slot, at 28.2° E, for digital broadcasting. The first satellite in this position, Astra 2A, is scheduled for launch next summer, and BSkyB has already booked 14 transponders – representing half its capacity. Astra is in "advanced negotiations with major public and private broadcasters' on the remaining transponders, according to director general Romain Bausch. These involve "an important amount of services primarily targeting the the UK market." Two satellites are so far planned, the second due up in late 1998, giving a total capacity of 56 transponders, or around 400 digitally-compressed channels.

Transponder output will be 100W, and signals will be receivable on 50cm dishes in the British Isles. Existing 60 or 80cm dishes, pointed at 19.2° E, should be able to receive signals from the new satellites, provided they are equipped with appropriate lnbs. However, neither Astra nor its broadcasters is recommending this approach. Few existing Sky subscribers, for instance, have the requisite lnb fitted.

A probable scenario will be for Sky to replicate existing analogue services as part of the new digital package – they would all fit onto three of its 14 transponders – enabling subscribers to flaunt the new, smaller dish.

Sky also has some digital capacity booked on Astra 1E and 1F. This results from long-standing bookings and is to be used for continental broadcasting, said an Astra spokesman.

SES/ASTRA MULTIMEDIA SERVICE PLATFORM

[Diagram showing satellite communication network]
UK centre of Europe for set-tops

The UK is set to become the European centre for volume manufacturing of digital set-top boxes as consumer electronics makers line up with announcements.

Last week, Sony started shipping digital set-top boxes from its Pencoed Technology Centre in Wales. The boxes are designed and manufactured in Wales and are Digital Video Broadcast (DVB) compliant. The first generation boxes will be delivered to Canal Plus for its 24 hour service in France.

Matsushita and the US receiver developer ComStream announced recently their production of digital set-top boxes in Cardiff, Wales, this year. In the start-up phase, scheduled to begin in July, Matsushita will produce around 500,000 units. From August much higher volumes will then be produced.

Pioneer is also close to a decision with regards to volume manufacturing digital set-tops in the UK, due to start at the end of this year.

“The first generation of set-top boxes will be done in Japan and then we’ll switch the production to Europe. The UK is a serious candidate,” said one Pioneer spokesman.

These announcements follow the joint declaration from Digi Media Vision and Mitsubishi earlier this year for the development and manufacture of such systems in Scotland. In addition, Pace, the indigenous set-top box volume manufacturer manufactures in excess of 800,000 units per year.

Matsushita and ComStream will supply set-top boxes for satellite, cable, MMDS (Microwave Multipoint Distribution System) and digital terrestrial television broadcast. The boxes will be designed by ComStream, but manufactured by Matsushita under the Panasonic brand name.

Svetlana Josifovska
Electronics Weekly

DMX wins satellite award

DMX was voted runner-up in the ‘Most innovative satellite product/application’ sector at the 1996 European Cable & Satellite Awards. DMX, the revolutionary new digital audio subscription service, was launched in the UK on March 20 of this year.

As the UK’s first digital audio subscription service, DMX is available on satellite and cable, offering continuous music 24 hours a day with no interruptions from DJs or advertising. There is something for everyone with 62 channels available on the service, each one dedicated to a particular genre of music – from rock to reggae, from jazz to dance and from opera to mariachi.

Ceramic laser

Japanese workers have developed what is claimed to be the first neodymium-doped yttrium-aluminium-garnet (Nd:YAG) laser based on ceramic materials.

Nd:YAG lasers are used for micro fabrication and, through fibre-optic light pipes, for medical work. Normally the lasers are made from monocrystalline material. The size, and thus the power, of these is limited by the size of the single crystals that can be grown. Growing large crystals can take over a month and defects often develop during the growth time.

The bulk material for the ceramic laser, developed by Krosaki Corporation and Prof K. Yoshida at Osaka Institute of Technology, can be made by sintering in a few hours.

Another advantage of ceramics is that up to 10 percent of Nd can be incorporated. This contrasts with only one percent of Nd included in monocrystals. The group found that 2.4 percent gives the maximum power output, twice that of the monocrystalline material.

Talking your way through the web

In future, computer nerds will surf the Web by talking to their terminals. So claim researchers at the Massachusetts Institute of Technology (MIT).

The spoken language systems group at MIT, led by Dr Victor Zue, is working on its third generation of voice recognition and understanding systems called Galaxy. Six researchers have spent over two years programming the software based system.

Zue says: “It is not just a case of linking the recognition and understanding bits together. In speech it can be difficult to tell where one word ends and another begins. An example is ‘euthanasia’ and ‘youth in asia’.” Zue claims to have solved many of the problems of speech taken out of context.

In an Internet system, the local user will have a client program that ferries the digitised speech, such as “What are the flight times from London to New York?”, to a central server. The server attempts to recognise and understand the speech, forms the correct response and delivers the answer back across the Net to the client, either through a text transfer or a voice synthesis system.

Zue thinks a working system for tailored applications using specific knowledge bases could be in use within five years.

Apple Computer’s PlainTalk system already provides plug-ins for Netscape Navigator that allows control of the navigator via speech. Users can speak hyper text links, bookmarks and commands such as ‘Go Back’ and ‘Reload’.

Global GSM network

Iridium, the prospective satellite phone operator owned by Motorola and a number of international telephone operators, has placed an order with Siemens for GSM switching systems to be used in the ground segment of its global network. Iridium’s network, which is scheduled to begin service by 1998, aims to be compatible with existing terrestrial GSM networks while offering intercontinental coverage using its own satellite base stations.

Tethered satellite report released

NASA and the Italian Space Agency (ASI) have recently released the report of the investigative board appointed to determine factors which resulted in the January 13 tether break and loss of the Tethered Satellite during the STS-75 Space Shuttle mission.

Findings of the board identified primary causes which accounted for the tether break during deployment of the Tethered Satellite.

“The tether failed as a result of arcing and burning of the tether, leading to a tensile failure after a significant portion of the tether had burned away,” the report concludes.

(Turn to page 538 for more details)
Lateral thinking turbo-charges submicron bicmos

Conventional bicmos technology—ideal for low voltage low-power applications—has a limitation when it comes to trying to achieve the ever higher packing densities required for VLSI. The speed of the bicmos gate shows a rapid deterioration as it is scaled down to 0.5-0.25µm levels. But by adding on a lateral bjt to the standard bicmos design, to create additional charging and discharging paths, two circuit electronics engineers at Nanyang Technological University, Singapore, have created a circuit that gives speeds and voltage swings much greater than has been obtained by bicmos at the submicron level ('Novel low-voltage bicmos digital circuits employing a lateral p-n-p bjt in a p-mos structure', S S Rofail and Y K Seng, IEE Proc - Circuits Devices Syst, Vol 143, No 2, pp. 83-90).

The principle of the new design is that the additional lateral p-n-p bjt traps charge during the pull up cycle, and uses it to speed up the pull down cycle. Tests on 0.25µm technologies are reported to show a good comparison with other circuits in terms of speed, output voltage swing and power dissipation.

According to the workers, large voltage swings at high speeds are easily achievable under 2.2V operation. NAND gate implementation of the new design also seems to demonstrate the best compromise of performance characteristics compared with conventional sub-micron bicmos and CMOS technologies.

More information from S S Rofail, Microelectronics Centre of the School of Electrical & Electronic Engineering, Nanyang Technological University, Nanyang Avenue, Singapore 2263.

Giant leap in magnetic sensor design?

A cheap, simple sensing system based on the principles of giant magnetoimpedance has been developed by a team of researchers in Spain. The device does not use optical technology so should be suitable for dusty industrial atmospheres, yet it is much less complex than other magnetic-field sensing techniques.

Giant magnetoimpedance (gmi) describes the effect where ferromagnetic materials subjected to an ac current exhibit a strong decrease in their impedance in the presence of a dc magnetic field. R Valenzuela and colleagues have used gmi as the basis for a magnetic-field sensor, adapted to monitor the passage of moving pieces or vehicles in industrial processes. The gmi mechanism in wires is actually quite complicated but is now well understood and accepted. Impedance in the Spanish

X-ray spectacular: The first global x-ray image ever obtained of the Earth’s aurora shows a hot spot of x-rays emanating from the atmosphere near midnight at the onset of a small magnetic disturbance and a wide band of weak x-ray emissions extending through the night and morning hours to noon. It was taken on March 20, 1996 by the Polar Ionospheric X-ray Imaging Experiment (Pixie) aboard the Nasa Polar spacecraft.

The image is presented in false colour with the colour corresponding to the measured x-ray intensity from blue (weakest) through red (strongest). The x-ray energy range covered by this image is from about 2000eV to over 10,000eV, and the x-rays were emitted when energetic electrons from the Earth’s magnetosphere struck the upper atmosphere. Intensity of the x-rays is directly related to the intensity of the precipitated electron flux. Asymmetry in the emissions between the local time regions corresponding to early morning (over Siberia and Alaska) and late afternoon (over northern Canada and Greenland) is the result of the natural motion of energetic electrons in the Earth’s magnetic field. The field causes electrons to drift to the east from their source region, probably far from Earth and near the equator, across the Earth’s magnetic field. As the drift is caused by large-scale electric fields, the electrons are lost into the atmosphere, producing the wide blue band of x-ray emissions. When the electrons reach the day side near noon, they can be swept out of the magnetosphere by the effects of large-scale electric fields.

Goddard Space Flight Center, Greenbelt, MD.
Tethered satellite was not a loss

Embarrassing breaking free of the satellite linked by a tether to spacec...
Ball lightning comes down to earth

The phenomenon of ball lightning has long puzzled scientists. Reports of floating luminescent globes mysteriously zigzagging above the ground then disappearing with a bang go back centuries — without explanation. Now a scientist at the CSIRO research institute in Australia has developed a theory that not only explains what ball lightning is, but also suggests why it moves as it does.

J J Lowke, in the Division of Applied Physics, took as his starting point the fact that reports of ball lightning usually follow a local lightning strike. The luminous ball, which can be up to 25cm in diameter, is seen to glow with the intensity of a 20W lamp and travel about 1m above the ground at a speed of around 3m/s. It can float for up to 10s — inside houses and even aircraft — after which time it extinguishes, sometimes silently, and sometimes with a bang.

Ball lightning has also been seen to pass through glass panes without affecting them, though some observers have reported wood singes and the smell of ozone and nitrogen oxides.

Previous theories have ranged from an optical illusion, to a standing wave of electromagnetic radiation, to antimatter, to the latest that the ball is the manifestation of complex chemical phenomena involving water vapour.

But Lowke’s explanation (‘A theory of ball lightning as an electric discharge’, J Phys D: Appl Phys, vol 29 (1996), pp. 1237-1244) begins by noting that when lightning strikes a point on the earth’s surface, an amount of charge, usually negative, is transferred via the lightning arc from the cloud to the ground. Positive charge is then transferred from the ground to cloud. Previous calculations of space charge effects have assumed that the Earth is a perfect conductor for the dispersing charge. But Lowke points out that earlier experiments carried out by other researchers show that lightning can produce filamentary arcing along the surface of the ground to distances of 20m and over. The sequence of events, according to Lowke, begins with the development of strong negative electric charges in the base of a thunder cloud, for example due to the interaction of wind and freezing supercooled rain drops. Next, there is the rapid transfer of charge through the highly conducting arc of a lightning strike — positive charge going to the cloud and negative charge to the Earth, to distances of many metres. Finally, there is the very much slower further dispersion of negative charge along fingers of relatively high electrical conductivity on the earth, in which the field at the heart of the advancing charge in the earth will be less than 1MV/cm. This produces an electric field above the earth which is the source of the power and motion for the ball lightning.

Lowke’s contention is that in the air above this charge there will be occasions when the field will be greater than 5kV/cm and so able to sustain ball lightning. Normally the field for electric breakdown in air is about 30kV/cm. But once a conducting plasma has been formed it can be sustained at the much lower figure.

Crucially, Lowke refers to two eyewitnesses who, observing ball lightning at night, told him they saw a faint luminosity between the main ball and the ground. This would tend to support the ground-based mechanism.

On this model, the ball lightning itself would be an electric discharge which is continuously varying on a microsecond time scale.

Calculation shows that space charge distortions by positive and negative ions can produce a local maximum in the electric field about 1m above the Earth’s surface and sustain a time-varying discharge with the properties similar to ball lightning.

The theory gives a credible and relatively simple explanation for the life-time and energy source of ball lightning. In addition it explains some of the other odd characteristics of ball lightning — particularly why it hovers rather than rises and why it moves erratically and seems unaffected by wind.

More information from J J Lowke at CSIRO Division of Applied Physics, Sydney, NSW 2070, Australia.

Electric charge redistribution after a lightning strike could explain the origin of ball lightning.
Douglas Self has thoroughly analysed the requirements for a no-compromise audio preamplifier making the most of today's high-performance op-amps. This first article covers the preamp's overall configuration and focuses on disc replay.

A new preamp design is timely. There is more variation in audio equipment than ever before, so to a greater extent preamps are required to be all things to all persons. High source resistance outputs and low-impedance inputs must be catered for, as well as ill-considered and exotic cabling with excessive shunt capacitance. The last preamp design I placed before the public was in 1983, extended in facilities by the moving-coil head amp stage published in 1987.

In the last ten years, small-signal analogue electronics has undergone few changes. Most circuitry is still made from TL072s, with resort to 5532s when noise and drive capability are important. In this period many new op-amps have appeared, but few have had any impact on audio design; this is largely a chicken/egg problem, for until they are used in large numbers the price will not come down low enough for them to be used in large numbers. Significant advantage over the old faithfuls is required. This new design uses the architecture established in reference 1, which has not been improved upon so far. The already low noise levels have been further reduced. The tone controls were fixed-frequency, and proved inflexible compared with the switched-turnover versions in my previous designs, so these frequencies are now fully variable, and a non-interrupting tone-cancel facility provided.

This preamplifier is designed to my usual philosophy of making it work as well as possible, by the considered choice of circuit configurations etc, rather than the alternative approach of specifying exotic components and hoping for the best.

The evolution of preamplifiers
Minimal requirements are source selection and level control, as in Fig. 1a; an RIAA disc

---

**Adding tape facilities and tone control**

There are two basic architectures for tape record/replay handling. The simpler, in Fig. 1d, adds a tape output and a tape monitor switch for off-tape monitoring on triple-head machines.

The more complex version in Fig. 1e allows any input to be listened to while any input is being recorded, though how many people actually do this is rather doubtful. This method demands very high standards of crosstalk inside the preamp. There is usually no tape return input or tape monitor switch as there is now no guarantee that the main path signal comes from the same original source as the tape output.

The final step is to add tone controls. They need a low-impedance drive for predictable equalisation curves, and a vital point is that most types — including the Baxandall — phase-invert. Since the maintenance of absolute polarity is required, this inversion can conveniently be undone by the active gain control, which also uses shunt feedback and phase-inverts. The tone-control can be placed before or after the volume control, but if afterwards it generates noise that cannot be turned down. Putting it before the volume control reduces headroom if boost is in use, but since maximum boost is only +10dB, the preamp inputs will not overload before 3Vrms is applied; domestic equipment can rarely generate such levels. Figure 1f shows the final architecture.
preamp stage is one input option. This sort of 'passive preamplifier' (a nice oxymoron) is only practical if the main music source is a low-impedance high-level output like cd.

The only parameter to decide is the resistance of the volume pot; it cannot be too high because the output impedance, which reaches a maximum of one quarter the track resistance at -6dB, will cause high-frequency roll-off with the cable capacitance. On the other hand, if the pot resistance is too low, the source equipment will be unduly loaded. If the source is valve equipment, which does not respond well to even moderate loading, the problem starts to look insoluble.

Adding a unity-gain buffer stage after the selector switch, Fig. 1b, means the volume control can be reduced to 10152, without loading the sources. This still gives a maximal output impedance of 2.5kΩ, which allows you only 5.4 metres of 300pF/m cable before the response is 1.0dB down at 20kHz. For 0.1dB down at 20kHz, only 1.6 metres is permissible.

The input RC filters found on so many power-amps as a gesture against transient intermodulation distortion add extra shunt capacitance ranging from 100pF to 1000pF, and can cause additional unwanted hf rolloff.

Unfortunately only a cd source can fully drive a power amplifier. Output levels for tuners, phono amps and domestic tape machines are of the order of 150mV rms, while power amplifiers rarely have sensitivities lower than 500mV. Both output impedance and level problems are solved by adding a second amplifier stage as Fig. 1c, this time with gain. The output level can be increased and the output impedance kept down to 100Ω or lower.

This amplifier stage introduces its own difficulties. Nominal output level must be at least 1V rms (for 150mV in) to drive most power amps, so a gain of 16.5dB is needed. If you increase the full-gain output level to 2Vrms, to be sure of driving exotic to its limits, this becomes 22.5dB, amplifying the input noise of the gain stage at all volume settings. Noise performance thus deteriorates markedly at low volume levels - the ones most of us use most of the time.

One answer is to split the gain before and after the volume control, so that there is less gain amplifying the internal noise. This inevitably reduces headroom before the volume control. Another solution is double gain controls - an input-gain control to set the internal level appropriately, then an output volume control that requires no gain after it.

Input gain controls can be separate for each channel, doubling as a balance facility. However this makes operation rather awkward. No matter how attenuation and fixed amplification are arranged, there are going to be trade-offs on noise and headroom.

All compromise is avoided by an active gain stage, ie an amplifier stage whose gain is variable from near-zero to the required maximum. You get lower noise at gain settings below maximum, and the ability to generate a quasi-logarithmic law from a linear pot. This gives

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**Fig. 1.** The course of preamp evolution, as impedance and level matching problems are dealt with.

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July/August 1996 ELECTRONICS WORLD 541
Requirements for the RIAA network

- The RIAA network must use series feedback, as shunt feedback is 1.4 dB noisier.
- Correct gain at 1kHz. Sounds elementary, but try calculating it.
- Accuracy. The 1983 model was designed for ±0.2dB accuracy 20-20kHz, which was the limit of the test gear. I had access to at the time. This is tightened to ±0.05dB without using rare parts.
- It must use obtainable components. Resistors will be E24 series and capacitors ±12 at best, so intermediate values must be made by series or parallel combinations.
- Rs (Fig. 2), must be as low as possible as its Johnson noise is effectively in series with the input signal. This is most important in moving-coil mode.
- The feedback network impedance to be driven must not be low enough to increase distortion or limit output swing — especially at high frequencies.
- The resistive path through the feedback arm should ideally have the same dc resistance as input bias resistor R18 (Fig. 8), to minimise offsets at A1 output. The circuitry here meets all these requirements.

Excellent channel balance as it depends only on mechanical alignment.

Design philosophy

There is great freedom of design in small-signal circuitry, compared with the intractable problems of power amplification. Hence there is little excuse for a preamp that is not virtually transparent, with very low noise, crosstalk, etc.

Once all the performance imperatives are addressed, the extra degrees of freedom can be used to, say, make components the same value for ease of procurement. Opamp circuitry is used here, apart from the hybrid moving-coil stage. The great advantage is that all the tricky details of distortion-free amplification are confined within the small black case of a 5532.

One route to low noise is low-impedance design. By minimising circuit resistances the contribution of Johnson noise is reduced, and hopefully conditions set for best semiconductor noise performance. This notion is not exactly new — as some manufacturers would have you believe but has been used explicitly in audio circuitry for at least fifteen years.

In the equalisation and AGS stages, gains of much less than one are sometimes required. In these cases, avoiding the evils of attenuation-then-amplification (increased noise) and amplification-then-attenuation (reduced headroom) requires the use of a shunt feedback configuration. In the classic unity-gain stage, the shunt amplifier works at a noise gain of x2, as opposed to unity, so using shunt feedback introduces a noise compromise at a very fundamental level.

Absolute phase is preserved for all input and outputs.

The preamp gain structure

Compared with ref. 1, the moving-magnet disc amplifier gain has been increased from +26 to +29dB (all levels are at 1kHz) to bring the line-out level up to 150mV nominal. This is done to match equipment levels that appear to have reached some sort of consensus on this value. The input buffer has a gain of +1.0dB with balance central.

The maximum gain of the AGS is therefore reduced from +26 to +22dB, to retain the same maximum output of 2V. This affects only the upper part of the gain characteristic.

Disc input

While vinyl as a music-delivery medium is almost as obsolete as wax cylinders, there remain many sizable album collections that it is impractical to either replace with cds or transfer to digital tape. Disc inputs must therefore remain part of the designer's repertoire for the foreseeable future.

The disc stage here accepts a moving-coil cartridge input of 0.1 or 0.5mV, or a moving-magnet input of 5mV. It also includes a third-order subsonic filter and the capability to drive low impedances. The moving-coil stage simply provides flat gain, of either 10 or 50 times, while the moving-magnet stage performs the full RIAA equalisation for both modes.

Moving-coil input criteria

This stage was described in detail in ref. 2. The prime requirement is a good noise figure from a very low source impedance — here 3.5Ω to comply with, for example, the Ortofon MC710 cartridge. The circuit features

- triple low-impedance transistors
- two separate dc feedback loops
- combined feedback-network and output-attenuator.

The very low value of Rs means that a series capacitor to reduce the gain to unity at dc is impracticable; there is no dc feedback through Rs, R10 around the global loop. Local dc negative feedback via R2, R7 sets input transistor conditions, and dc servo IC2 applies whatever is needed to ICI non-inverting input to bring ICI output to 0V.

The two gains provided are 10x and 50x, so inputs of 0.5mV and 0.1mV will give 5Vrms out. The equivalent input noise of the moving-coil stage alone is −14dBu, with no RIAA. Johnson noise from a 3.5Ω resistor is −147dBu, so the noise figure is a rather good 6dB. Resistor R7 is also 3.5Ω. This component generates the same amount of noise as the source impedance, which only degrades the noise figure by 1.4dB, rather than 3dB, as transistor noise is significant.

If discrete transistors seem like too much trouble, remember a 5532 stage here would be at least 15dB noisier.

The moving-magnet input stage

The first half of Morgan Jones's excellent preamp article appeared just after this preamp design was finalised. While I thoroughly endorse most of his conclusions on RIAA equalisation, we part company on two points. Firstly, I am sure that 'all-in-one-go' RIAA equalisation is as Fig. 2a is definitely the best method for IC op-amp designs at least. In my design the resultant loss of high-frequency headroom is only 0.5dB at 20kHz, which I think I can live with.

Secondly, I do not accept that the difficulties of driving feedback networks with low-impedance at hf are insoluble. I quite agree that 'very few preamps of any age' meet a +28dB ref 5mV overload margin, but some exceptions are ref. 1 with +36dB, ref. 3 with +39dB, and ref. 4 with a tour-de-force +47dB. My design here gives +36dB across most of the audio band, falling to +33dB at 20kHz.

![Fig. 2. The basic RIAA configurations. Fig. 2a is the standard 'all-in-one-go' series feedback configuration; the values shown do not give accurate RIAA equalisation. Fig. 2b is the most common type of passive RIAA, with a headroom penalty of 14dB at 10kHz.](image-url)
(due to hf pole-correction) and +31dB at 10Hz (due to the IEC rolloff being done in the second stage).

Many contemporary disc inputs use an architecture that separates the high and low RIAA sections. Typically there is a low-frequency RIAA stage followed by a passive hf cut beginning at 2kHz, Fig. 2b. The values shown give a correct RIAA curve.

Amplification followed by attenuation always implies a headroom bottleneck, and passive hf cut is no exception. Signals direct from disc have their highest amplitudes at high frequencies so this passive configuration gives poor hf headroom. Overload occurs at A1 output before passive hf cut can reduce the level.

Figure 3 shows how the level at A1 output (Trace B) is higher at hf than the output signal (Trace A). Trace C shows the difference, i.e. the headroom loss; from 1dB at 1kHz this rises to 14dB at 10kHz and continues to increase in the ultrasonic region. The passive circuit was driven from an inverse RIAA network. Using this, a totally accurate disc stage would give a straight line just below the +30dB mark.

A related problem is that A1 in the passive version must handle a signal with much more hf content than A1 in Fig. 2a. This worsens any difficulties with slew-limiter and hf distortion: The passive version uses two amplifier stages rather than one, and more precision components.

Another difficulty is that A1 is more likely to run out of open-loop gain at hf. This is because the response plateaus above 1kHz, rather than being steadily reduced by increasing negative feedback. Passive RIAA is not an attractive option.

Alternatively there may be a flat input stage followed by a passive hf cut and then another stage to give the hf boost, which has even more headroom problems and uses yet more bits. The ‘all-in-one-go’ series feedback configuration in Fig. 2a avoids unnecessary headroom restrictions and has the minimum number of stages.

In search of accurate RIAA
I have a deep suspicion that such popularity as passive RIAA has is due to the design being much easier. The time-constants are separate and non-interactive; only the simplest of calculations are required.

In contrast the series-feedback system in Fig. 2a has serious interactions between its time-constants and design by calculation is complex. The values shown in Fig. 2a are what you get if you ignore the interactions and simply implement the time-constants as $R_a \times C_a = 3180\mu s$, $R_b \times C_a = 318\mu s$, and $R_b \times C_b = 75\mu s$. The resulting errors are ±0.5dB ref 1kHz.

Empirical approaches (cut-and-try) are effective if great accuracy is not required, but attempting to reach even ±0.2dB by this route becomes very tedious and frustrating. Hence the Lipshitz equations$^6$ have been converted to a spreadsheet, and used to synthesise the
Table 1. Measured noise results, showing the 5532's superiority.

<table>
<thead>
<tr>
<th>Case</th>
<th>$e_m$ (nV/Hz)</th>
<th>$i_m$ (pA/Hz)</th>
<th>$R_a$ (kΩ)</th>
<th>$R_b$ (MΩ)</th>
<th>Output (dBu)</th>
<th>S/N ref (dB)</th>
<th>EIN (dBu)</th>
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<td>0.0</td>
<td>1000</td>
<td>0.0</td>
<td>-104.0</td>
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<td>-133.5</td>
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<tr>
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<td>47k</td>
<td>0.0</td>
<td>-97.1</td>
<td>-82.8</td>
<td>-126.5</td>
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<tr>
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<td>0.0</td>
<td>47k</td>
<td>220</td>
<td>-96.7</td>
<td>-82.4</td>
<td>-126.2</td>
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<tr>
<td>11 258737, $I_{shunt}$=70µA</td>
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<td>0.4</td>
<td>47k</td>
<td>220</td>
<td>-95.3</td>
<td>-81.0</td>
<td>-124.8</td>
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<tr>
<td>16 5532</td>
<td>5</td>
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<td>47k</td>
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<td>-92.5</td>
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<tr>
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<td>220</td>
<td>-86.9</td>
<td>-72.6</td>
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Table 2. Calculated minimum noise results.

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<th>$i_m$ (pA/Hz)</th>
<th>$R_a$ (kΩ)</th>
<th>$R_b$ (MΩ)</th>
<th>Output (dBu)</th>
<th>S/N ref (dB)</th>
<th>EIN (dBu)</th>
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<td>1000</td>
<td>0.0</td>
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Fig. 6. The effect on RIAA accuracy of a ±0.44% variation in $C_a$. Effect is less than ±0.05dB at low frequencies, with a small effect on the upper audio band.

Fig. 7. The effect on RIAA accuracy of a ±0.58% variation in $C_b$. Effect is less than ±0.05dB on top four octaves. Smaller variation is permissible in the capacitors for the same RIAA error.

A great deal of rubbish has been talked about RIAA equalisation and transient response, in perverse attempts to render the shunt RIAA configuration acceptable despite its crippling 14dB noise disadvantage. The heart of the matter is that the RIAA replay characteristic apparently requires the hf gain to fall at a steady 6dB/octave forever. A series-feedback disc stage with relatively low gain cannot make its gain fall below one, and so the 6dB/octave fall tends to level out at unity early enough to cause errors in the audio band. Adding a high-frequency correction pole – ie low-pass time constant – just after the input stage makes the simulated and measured frequency response identical to a shunt-feedback version, and retains the noise advantage.

At this level of accuracy, the finite gain open-loop gain of even a 5534 at hf begins to be important, and the frequency of the hf pole is trimmed to allow for this.

What RIAA accuracy is possible without spending a fortune on precision parts? The best tolerance readily available for resistors and capacitors is ±1%, so at first it appears that anything better than ±0.1dB accuracy is impossible. Not so. The component-sensitivity plots in Figs 4, 5 show the effect of 1% deviations in the value of $R_a$, $R_b$, the response errors never exceed 0.05dB, as there are always at least two components contributing to the RIAA response.

Sensitivity of the RIAA capacitors is shown in Figs 6, 7 and you can see that tighter tolerances are needed for $C_a$ and $C_b$ than for $R_a$ and $R_b$ to produce the same 0.05dB accuracy. The capacitors have more effect on the response than the resistors.

Finding affordable close-tolerance capacitors is not easy; the best solution seems to be, as in 1983, axial polystyrene, available at 1% tolerance. These only go up to 1µF, so some paralleling is required, and indeed turns out to be highly desirable. The resistors are all 1%, which is no longer expensive or exotic, though anything more accurate certainly would be.

For $C_a$, the five 10µF capacitors in parallel reduce the tolerance of the combination to 0.44%. This statistical trick works because the variance of equal summed components is the square root of five, while total capacitance has (square root of variance) increases only by the square root of five, to ±0.58% accuracy of a capacitor.

Similarly, $C_b$ is mainly composed of three 4n7 components and its tolerance is impossible. Not so. The component-sensitivity plots in Figs 4, 5 show the effect of 1% deviations in the value of $R_a$, $R_b$, the response errors never exceed 0.05dB, as there are always at least two components contributing to the RIAA response.

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input source resistance. The best possible equivalent input noise data for resistive sources, for example microphones with a 200Ω source resistance, i.e. -129.6dBu, is well-known, but the same figures for moving-magnet inputs are not.

It is particularly difficult to calculate equivalent input noise for moving magnet stages as a highly inductive source is combined with the complications of RIAA equalisation. The amount by which a real amplifier falls short of the theoretical minimum equivalent input noise is the noise figure, NF. I often wonder why noise figures are used so little in audio; the theoretical minimum amount by which a real amplifier falls short of the RIAA equalisation. The stage also buffers the high-frequency correction pole, and gives the capability to drive a 600Ω load, if you can find one.

The noise performance of disc input stages depends on the input source impedance. The cartridge inductance having the greatest influence. It is vital to realise that no value of resistive input loading will give realistic noise measurements.

A 1kΩ load models the resistive part of the cartridge impedance. But it ignores the fact that the 'noiseless' inductive reactance makes the impedance seen at the preamp input rise very strongly with frequency, so that at higher frequencies most of the input noise actually comes from the 47kΩ loading resistance. I am grateful to Marcel van de Gevel for drawing my attention to this point.

Hence, for the lowest noise you must design for a higher impedance than you might think, and it is fortunate that the RIAA provides a treble roll-off, or the noise problem would be even worse than it is. This is not why it was introduced. The real reason for pre-emphasis/de-emphasis was to discriminate against record surface noise. Table 1 shows the two most common audio op-amps, the 5532 being definitely the best and quieter by 5dB.

To calculate appropriate EINs, I built a spreadsheet mathematical model of the cartridge input, called MAGNOISE. The basic method is as in ref. 9. The audio band 50-22kHz is divided into nine octaves, allowing RIAA equalisation to be applied, and the equivalent generators of voltage noise (εa) and current noise (ic) to be varied with frequency.

Noise generated by the 47kΩ resistor R0 is modelled separately from its loading effects so its effect can be clearly seen. I switched off the bottom three octaves to make the results comparable with real cartridge measurements that require a 400Hz high-pass filter to eliminate hum, and 1/f effects are therefore neglected. No psychoacoustic weighting was used, and cartridge parameters were set to 610Ω+470mH, the measured values for the Shure M75ED.

The results match well with my 5532 and TL072 measurements, and I think the model is a usable tool. Table 2 shows some interesting cases; output noise is calculated for gain of +29.55dB at 1kHz, and signal-to-noise ratio (SINR) for a 5mVrms input at 1kHz.

I drew the following conclusions. The minimum equivalent input noise from this particular cartridge, without the extra thermal noise from the 47kΩ input loading, is -133.5dBu, no less than 7dB quieter than the loaded cartridge. (Case 1) It is the quietest possible condition. The noise difference between 10MΩ and 1MΩ loading is still 0.2dB, but as loading resistance is increased further to 1000MΩ the EIN asymptotes to -133.5dBu. A 47kΩ loading is essential for correct cartridge response.

With 47kΩ load, the minimum EIN from this cartridge is -126.5dBu. (Case 5) All other noise sources, including R0, are ignored. This is the appropriate noise reference for this preamp design.

Resistor R0, the 220Ω resistor in the bottom arm of negative feedback network, adds little noise. The difference between Case 5 and Case 7 is only 0.3dB.

A disc preamp stage using a good discrete bipolar device such as the remarkable 2SB737 transistor (rβ only 2Ω typ) is potentially 2.8dB quieter than a 5532, when the noise from R0 and the input load are included. Compare Cases 11 and 16.

The calculated noise figure for a 5532 is 4.5dB. Measured noise output of the moving-magnet stage is -92.3dBu (1kHz gain +29.5dB) and so the equivalent input noise is -121.8dBu, and the real noise figure is 4.7dB.

**Filtering subsonics**

This stage is a third-order Butterworth high-pass filter, modified for a slow initial rolloff that implements the IEC amendment. This is done by reducing the value of R27+R28 below that for maximal flatness. The stage also buffers the high-frequency correction pole, and gives the capability to drive a 600Ω load, if you can find one.

Capacitor distortion is - or should be - by now a well-known phenomenon. It is perhaps less well known that non-electrolytics can also generate distortion in filters like these. This has nothing to do with Subjectivist musicality, but is very real and measurable.

The only answer appears to be using the highest-voltage capacitors possible; 100V polyester generates ten times less distortion than the 63V version.

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**Fig. 8. Switchable for moving coil or moving-magnet type cartridges, the disc amplifier includes a subsonic filter to reduce cone excursions and distortion due to warped vinyl.**
which is not too bad. Noise from the subsonic filter is negligible.

Taking en and in from data books, it looks as though the 5534/5532 is the best op-amp possible for this job. Other types - such as OP-27 - give slightly lower calculated noise, but measure slightly higher. This is probably due to extra noise generated by bias current-cancellation circuitry 5.

There is an odd number of half-5532s, so the single 5534 is placed in the moving-magnet stage, where its slightly lower noise is best used. The RIAA-equalised noise output from the disc stage in moving-coil mode is 

-93.9dBu for 10x times gain, and -85.8dBu for 50x times. In the 10x case the moving-coil noise is actually 1.7dB lower than moving-magnet mode.

Circuit details
The complete circuit of the disc amplifier and subsonic filter is Fig. 8. Circuit operation is largely described above, but a few practical details are added here. Resistors R9 and R12 ensure stability of the moving-coil stage when faced with moving-magnet input capacitance C8, while R3 and R1, are dc drains. The 5534 moving-magnet stage has a minimum gain of about 3x, so compensation should not be required; if it is, a position is provided (C26) for external capacitance to be added; 4.7pF should be ample. The moving-magnet stage feedback arm R20,23 has almost exactly the same dc resistance as the input bias resistor R18, minimising the offset at the output of IC3. The hf correction pole is R24+R25 and C20.

Capacitor C24 is deliberately oversized so low loads can be driven. Resistor R31 ensures stability into high-capacitance cables.

References
5. M van de Gevel, Private communication, Feb 1996.

High-quality circuit boards for Douglas Self's precision preamplifier '96

A high quality double-sided circuit board is available for Doug Self's precision preamplifier, exclusively via Electronics World. The board takes the full stereo preamplifier, including all power supply components except the transformer. Its layout is optimised to provide exceptionally low crosstalk. Co-designed by Gareth Connor, the board is glass-fibre with plated-through holes and roller-tinned. It features solder masking and full component identification. Component lists and assembly notes - containing extra information about the preamplifier - are supplied with each order. Each board is £59 inclusive of package, VAT and recorded postage. Please include a cheque or postal order with your request, payable to Reed Business Publishing. Alternatively, send your credit card details - i.e. card type, number and expiry date. Include the delivery address in the order, which in the case of credit card holders must be the address of the card holder. Add a daytime telephone and/or fax number if you have one.

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Features of Douglas Self's precision preamplifier
- Very low noise and distortion.
- Moving-coil - sensitivity switchable 0.1 or 0.5mV, ±0.05dB RIAA accuracy.
- Moving-magnet input with ±0.05dB RIAA accuracy, 5V rms sensitivity.
- Three 150mV line inputs.
- One dedicated compact-disc input.
- Tape-monitor switch.
- Active-balance control.
- Tone control - switch defeatable - with ±10dB range.
- Tone control treble and bass frequencies variable over 10:1 range.
- Active volume control for optimal noise/headroom and enhanced interchannel matching.
- Intelligent relay muting on outputs.
- CD input sensitivity 1V rms.
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This fantastic John Linsley Hood designed amplifier is the flagship of our range, and a fitting showcase for your ultimate hi-fi system.

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PREAMP ELECTRONICS.
recently, while browsing through one of my components catalogues, I came across the section that had the audio crossover networks in. I noticed two things:

- They use various combinations of first, second and third order networks, Figs. 1,2,3.
- They are all expensive for what is not much more than 12 passive components on a pcb.

The most expensive was in the region of £100 for a second, first, third-order combination. The expensive types incorporate overload protection circuits in the form of varistors that attenuate the signal when it reaches a certain overload level.

A glance through my final year notes on audio design showed that the realisation of these units was nearly a trivial task. Further reading showed that the first and second-order sections are inappropriate for high-fidelity sound.

**Passive crossover networks**

A passive crossover network consists of a high power Butterworth filter that splits the audio signal into various frequency bands, the different bands being fed to the relevant loudspeaker. Active crossover networks exist, but these incorporate ordinary op-amps so one power amplifier is needed for each speaker.

The simplest and cheapest form of crossover is a first-order two-way network, Fig. 1. This type of crossover is often found in low-budget hi-fi. It is capable of producing good results. However, it has a 6dB/octave roll-off. To perform well, it needs wide-range drivers with a defined spacing, and having a common radiation plane and dispersion. These are not cheap.

Next, comes the second order network, Fig. 2. This has a 12dB/octave roll-off, so signal separation is better. But there are still problems. The outputs are in anti-phase at the crossover point, Fig. 2a. This situation can be corrected by reversing the leads to the treble driver. Unfortunately, this results in displacement elsewhere.

Finally, we come to the third order version, Fig. 3. This has an 18dB/octave roll-off in the stopband, so signal separation is better. Also, the phase response is superior through the crossover region, being at approximately 90° throughout the band, Fig. 3a.

Applying the relevant equations is a trivial task, but to speed things up, a piece of software can be written. This not only prevents errors due to kak-handedness on a calculator keyboard, but also enables a permanent record of a design to be kept for future reference. The C program, Listing 1, calculates the components for a two way or three way crossover and then optionally save them to a disk file. Listing 2 is a typical output run from the program.

Suppose you want to build a three-way net-
Listing 1. Simple crossover network component calculator in C
/* CROSS.C 28:02:96 ; Final version - for now */
#include <stdio.h>
#include <stdlib.h>
#include <math.h>
#include <float.h>
#include <conio.h>
#include <dos.h>
define pi 3.141592654 /* all variables are global */
char *mtsfl = ("","JAN","FEB","MAR","APR","MAY","JUN","JUL","AUG","SEP","OCT","NOV","DEC");
char *tday[] = {"SUN","MON","TUE","WED","THU","FRI","SAT");
float fc1,fc2,r,11,12,13,c1,c2,c3,11a,12a,13a,cla,c2a,c3a;
char tipe,qsave;
char name[8],aname[15];
union REGS reg;
FILE *tofile;

void get_param(void); /* get the parameters */
void do_calculations(void); /* calculate component values */
void do_printout(void); /* print results to monitor */
void save_stuff(void); /* save design to disk */

main ()
get_param();
do_calculations();
do_printout();
save_stuff();

void get_param(void) /* get the parameters */
{
printf("D)ual OR T)hree way crossover = ? ");
tipe = toupper(getche());
if (tipe == 'D')
printf("Crossover frequency (Hz)= ? ");
sconf("%f",&fc1);
else
printf("First Crossover Frequency (Hz)= ? ");
sconf("%f",fc1);

void do_calculations(void) /* calculate component values */
{
11 = 3*r/(4*pi*fc1);
12 = 11/3;
13 = 11/2;
c1 = 2/(3*pi*fc1*1); c2 = c1/2;
c3 = c1/3;
if (tipe == 'T')
11a = 2/(3*pi*fc2*1);
c2a = c1a/2;
c3a = c1a/3;
}

void do_printout(void) /* print results to monitor */
{
printf("L1 = %e %s
",11,"H");
printf("L2 = %e %s
",12,"H");
printf("L3 = %e %s
",13,"H");
printf("C1 = %e %s
",c1,"F");
printf("C2 = %e %s
",c2,"F");
printf("C3 = %e %s
",c3,"F");
if (tipe == 'T')
printf("L1a = %e %s
",11a,"H");
printf("L2a = %e %s
",12a,"H");
printf("L3a = %e %s
",13a,"H");
printf("C1a = %e %s
",c1a,"F");
printf("C2a = %e %s
",c2a,"F");
printf("C3a = %e %s
",c3a,"F");
}

void save_stuff(void) /* save design to disk */
{
printf("DO YOU WANT TO SAVE DESIGN ? Y/N ");
qsave = toupper(getche());
if (qsave == 'Y')
printf("SAVE DESIGN AS ? ");
sconf("%s",&name);
printf("YOUR NAME ? ");
sconf("%s",&aname);
if ((tofile = fopen(name,"w")) == NULL)
printf("Error opening text file for writing
");
exit(0);
fprintf(tofile,"FILENAME : 
","DATE : %s,%s %d,%d 
",tday[reg.h.al],mtsfl[reg.h.dh],reg.h.dl,reg.x.cx);
fprintf(tofile,"SPEAKER IMPEDANCE : %e %s
");
fprintf(tofile,"CROSSOVER FREQUENCY : %e %s
");
}

fclose(tofile);
work. Simple, you just add on another network to the output of the high-frequency path of the two-way crossover, Fig. 4. Because we’re dealing with equiterminated networks here, there are no problems with impedance mismatches. Listing 3 is hard copy of an output run.

Figure 4a is the response of the three outputs from the network. To make sure that the correct part of the signal spectrum reaches the appropriate speaker it is necessary to recalibrate the component values.

Implementation considerations
Build a crossover network requires careful component selection. All capacitors should be non-polarised electrolytics of the highest possible voltage rating. It is likely that the values available will not equal the design values. In this case, two or three can be put in parallel depending on what component tolerances you want to work to.

To put this in perspective, according to some sources, because the impedance of the loudspeaker varies with frequency, capacitors can be up to 50% away from their design value. Inductors can be air cored. Alternatively, if you don’t have miles of wire to spare, iron-powder toroids can be used. Note that air cored inductors will interact with each other depending on their separation. So a crossover network using these, will take up more PCB space.

There are no proximity problems with toroids due to their closed magnetic circuit. However, toroids saturate at a certain power level. It is important to use high power toroids or types specifically made for EMI or power filters. I ruled pot cores out because of their cost.

My prototype has an experimental overload protection circuit in the form of a triac across the input terminals. Normally breakdown protection circuit in the form of a triac across commercial units.

Unfortunately tests could not proceed normally with this because at the volume where the triac fires, the main terminals of the triac are connected to the input signal via a high power 8Ω resistor and to ground.

Further improvements to the circuit can be obtained with such measures are questionable, otherwise the response will revert to a cross between a Linkwitz and Butterworth.

I don’t know of anywhere that sells E24 non-polarised electrolytics to 1% tolerance, nor how much they might cost.

Debatable improvements
Further improvements to the circuit can be implemented, such as delay equalising the outputs or employing some sort of resistive damping, but the improvements that can be obtained with such measures are questionable and open to debate.

Further reading
Williams & Taylor, Electronic Filter Design Handbook.
F.R. Conner, Networks.
R.M. Marston, Power control circuits manual.
Linkwitz Filters, Elektor Electronics, April 1987.

Listing 2. Hard copy of an output run from the calculator program crossover network components.
FILENAME: CROSS 1
DESIGNED BY: B.TELEKI
DATE: TUE, MAR 26, 1996
SPEAKER IMPEDANCE: 8.000000e+00 OHMS
CROSSOVER FREQUENCY: 1.600000e+03 Hz
L1 = 1.193662e+03 H
L2 = 3.978873e+04 H
L3 = 5.968310e+04 H
C1 = 1.657864e+05 F
C2 = 8.289320e+06 F
C3 = 2.486796e+05 F

Listing 3. Hard copy from the crossover calculator for a third-order filter.
FILENAME: THREE
DESIGNED BY: J.T.
DATE: WED, MAR 27, 1996
SPEAKER IMPEDANCE: 8.000000e+00 OHMS
CROSSOVER FREQUENCY: 2.500000e+02 Hz
L1 = 7.639437e+03 H
L2 = 2.546679e+03 H
L3 = 3.819719e+03 H
C1 = 1.061033e+04 F
C2 = 5.305165e+05 F
C3 = 1.591549e+04 F
CROSSOVER FREQUENCY: 3.500000e+03 Hz
L10 = 5.456741e+04 H
L20 = 1.816914e+04 H
L30 = 2.728370e+04 H
C10 = 7.528087e+06 F
C20 = 3.789403e+06 F
C30 = 1.136821e+05 F

\[
L = \frac{3R}{4\pi^2} \quad L_2 = \frac{L_1}{3} \quad L_3 = \frac{L_1}{2}
\]

\[
C = \frac{2}{3\pi^2 R}
\]

\[
L_1 = 3.000000 \quad L_2 = 1.000000 \quad L_3 = 0.666667
\]

\[
C_1 = 0.000000 \quad C_2 = 0.000000 \quad C_3 = 0.000000
\]

\[
C_{10} = 5.000000 \quad C_{20} = 1.500000 \quad C_{30} = 4.000000
\]

Fig. 3. Third order 18dB/octave two way crossover.

Fig. 4. Third order 18dB/octave three-way crossover. Simply adding another third-order section to the high-frequency output of the two way network, creates a three way network. No impedance mismatch occurs since load is still R. But component values have to be redesigned to redistribute the signal spectrum. Use equations as in Fig. 3.
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Emil Vladkov's meter spans 0.1pF–1.999nF and has autoranging. Based on the comparison of two frequencies, the design can be trimmed to achieve a basic accuracy of 0.1%.

Key specifications of the autoranging capacitance meter.

<table>
<thead>
<tr>
<th>Ranges</th>
<th>Accuracy</th>
<th>Resolution</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.1pF–199.9pF</td>
<td>±(0.1%+1 digit) 0.1pF</td>
<td>0.1pF</td>
</tr>
<tr>
<td>0.199nF–1.999nF</td>
<td>±(0.05%+1 digit) 1pF</td>
<td>1pF</td>
</tr>
<tr>
<td>1.999nF–19.99nF</td>
<td>±(0.05%+1 digit) 10pF</td>
<td>10pF</td>
</tr>
<tr>
<td>19.99nF–199.9nF</td>
<td>±(0.1%+1 digit) 100pF</td>
<td>100pF</td>
</tr>
<tr>
<td>100pF</td>
<td>±(0.1%+1 digit) 1nF</td>
<td>1nF</td>
</tr>
</tbody>
</table>

Measuring principle

The principle of my design is illustrated with in Fig. 1. It is based on two equal timers working in monostable mode. This means that the timing RC circuit of each are identical.

It is not possible to measure capacitances in the range of 0.1pF using only one timer in monostable mode. This is because of the parasitic capacitance of the leads of the ics, which cannot be compensated for with one timer.

The p-n junctions within the timer also have significant parasitic capacitance. Without an external unknown capacitor, the monostable multivibrator generates a short pulse due to these capacitances, making the measurement of picofarad values impossible.

For the above reason, I decided to use two timers. With no unknown capacitor connected, the two monostable timers generate pulses with the same AT width. The timing diagram of the compensating reference timer is Fig. 1 a) and the measuring timer is labelled b).

Variable T0 is the period of the triggering pulses. Pulses generated by the monostable timers are applied to an exclusive-or gate. Output of this gate goes high only if there is a difference between the two pulses.

So without an external capacitance, as timing diagram 3 shows, there is no pulse at the output of the XOR gate. If an unknown capacitor is set in the timing network of the measuring timer, it generates a pulse with an additional duration of,

\[ T = 1.1RC_x \]

as shown in c). This additional duration is exactly proportional to the unknown value of the capacitance to be measured. The 555 timer is chosen, because its pulse duration does not depend on the supply voltage. This can be a source of errors. Output of the XOR gate goes high for exactly this duration T.

The principle of the system is to measure the dc component of the signal at the output of the XOR gate. Because of the positive voltage level of the logic zero, i.e. a low level, at the XOR output and the voltage drop on the additional diodes, D2 and D3 in Fig. 2, there is a parasitic dc component. This is added to the useful dc component proportional to Cx and must be removed.

The useful signal with removed parasitic dc component is shown in d), together with the useful dc component proportional to Cx.

Circuitry of the measuring unit

Figure 3 is the measuring module schematic. It consists of the compensating timer IC13, the measuring timer IC14 and an additional monostable-mode timer, based on a 74LS123, IC13B. It has the task of measuring capacitances in the range between 19.99nF and 1.999µF.
## Component list

### Integrated circuits

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>IC1, IC3, IC22</td>
<td>74LS00</td>
</tr>
<tr>
<td>IC2, IC3, IC4, IC5</td>
<td>74LS05</td>
</tr>
<tr>
<td>IC6</td>
<td>74LS164</td>
</tr>
<tr>
<td>IC2, IC5</td>
<td>74LS01</td>
</tr>
<tr>
<td>IC10, IC13, IC14</td>
<td>TDB 555 (Siemens)</td>
</tr>
<tr>
<td>IC11</td>
<td>74LS08</td>
</tr>
<tr>
<td>IC12</td>
<td>74LS123</td>
</tr>
<tr>
<td>IC15</td>
<td>7486</td>
</tr>
<tr>
<td>IC16</td>
<td>74LS27</td>
</tr>
<tr>
<td>IC17</td>
<td>74LS02</td>
</tr>
<tr>
<td>IC18</td>
<td>CD 4049</td>
</tr>
<tr>
<td>IC19</td>
<td>ICL 7107 (Harris)</td>
</tr>
<tr>
<td>IC20, IC21</td>
<td>VQE24</td>
</tr>
</tbody>
</table>

### Resistors

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>R1</td>
<td>110Ω</td>
</tr>
<tr>
<td>R2</td>
<td>560Ω</td>
</tr>
<tr>
<td>R3, R6, R16</td>
<td>1.1kΩ</td>
</tr>
<tr>
<td>R4</td>
<td>130kΩ</td>
</tr>
<tr>
<td>R5</td>
<td>43kΩ</td>
</tr>
<tr>
<td>R7</td>
<td>6.2kΩ</td>
</tr>
<tr>
<td>R8</td>
<td>1000Ω linear</td>
</tr>
<tr>
<td>R9, R12, R18, R19</td>
<td>470Ω</td>
</tr>
<tr>
<td>R10, R11</td>
<td>383kΩ (0.1%) or selected 384kΩ</td>
</tr>
<tr>
<td>R13</td>
<td>56kΩ</td>
</tr>
<tr>
<td>R14, R21</td>
<td>10kΩ linear pot</td>
</tr>
<tr>
<td>R15</td>
<td>10kΩ</td>
</tr>
<tr>
<td>R20, R22</td>
<td>4.3kΩ</td>
</tr>
<tr>
<td>R23</td>
<td>1.0MΩ</td>
</tr>
<tr>
<td>R24</td>
<td>470kΩ</td>
</tr>
<tr>
<td>R25, R50</td>
<td>820Ω</td>
</tr>
</tbody>
</table>

### Capacitors

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>C1</td>
<td>330pF or smaller (adj.), depends on G1</td>
</tr>
<tr>
<td>C2, C10-C15</td>
<td>4.7nF</td>
</tr>
<tr>
<td>C3</td>
<td>1.0nF</td>
</tr>
<tr>
<td>C4, C11</td>
<td>10µF</td>
</tr>
<tr>
<td>C5</td>
<td>27pF</td>
</tr>
<tr>
<td>C6, C7</td>
<td>5.6pF</td>
</tr>
<tr>
<td>C8</td>
<td>1µF</td>
</tr>
<tr>
<td>C9</td>
<td>3.9pF</td>
</tr>
<tr>
<td>C12</td>
<td>100nF</td>
</tr>
<tr>
<td>C13</td>
<td>10nF</td>
</tr>
<tr>
<td>C14</td>
<td>220nF</td>
</tr>
</tbody>
</table>

### Diodes

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>D1, D2, D3, D4, D9</td>
<td>1N4148</td>
</tr>
<tr>
<td>D4</td>
<td>BZX79C 3V3 (Philips)</td>
</tr>
<tr>
<td>D5, D6, D7</td>
<td>LEDs, any colour</td>
</tr>
</tbody>
</table>

### Resonators

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>G1</td>
<td>1MHz Quartz</td>
</tr>
</tbody>
</table>

The 555 timers are used for the low-value ranges 0.1pF to 199.9pF, 0.199nF to 1.999nF and 1.999nF to 19.99nF. Exclusive-or gate IC15B compares the measured signal, proportional to CV, and the compensating signal. Gates IC15C and IC19 apply the signals of the three low and two high ranges to the same RC filter, R13/C8. For low capacitances, the autoranging circuit transmits the pulses from the 555 timers through IC15C and D2. Part of the autoranging circuit, IC15A, goes low if any one of the Q3, Q4 or Q5 lines goes high. The duration of the diode’s inverse bias is proportional to CV. During this time, a stable voltage derived from 3.3V zener diode D4, is applied to the input of the RC filter. This makes the measurement independent of the high-level output voltage of the IC.

---

Fig. 2. This module has the task of comparing capacitances in the range 0.1pF and 1.999pF, outputting a dc signal proportional to CV.
If the high capacitance ranges are chosen by the autoranging circuit, IC15C goes high, because there are no pulses from the 555 timers circuit and the output of IC15A is set high. In this case there is no unknown Cx in the timing network of the timer IC1A; it is in the timing network of IC12B. So diode D2 is switched off.

Pulses of the monostable timer IC12B are transmitted to diode D3 through the gates IC8B and IC15D and then to the RC filter. Together with D1, IC15A forms the parasitic dc voltage compensation, which is applied to the low-level input pin of the integrated analogue-to-digital converter IC19. This device measures the voltage Vin, which is Vin,hi - Vin,lo.

Direct-current component subtracted from the useful signal is set by potentiometer Rs.

The autoranging section of the low capacitance meter is under quartz frequency control and capable of three measurement cycles a second. The clock measurement signal is produced by the autoranging circuit, shown in Fig. 3. The principle of autoranging is based on the use of a stable 1MHz oscillator based on IC1A-D, divided by ten by counters IC2,3.

For the range 0.199g - 1.999g, at the beginning of the measurement cycle, a frequency of 100Hz is applied to the pulse former IC12A, in Fig. 3. From there it feeds the first measuring module IC12B.

Information for the seven-segment code outputs of the a-to-d converter ICL7107, IC19 of Fig. 4, is analysed. If the digital code of the input voltage of IC19 is identified as smaller than 0200, this means the measurement can be completed in the lower range. So the next greater frequency of 1kHz, i.e. a range of

Fig. 3. Autoranging section of the low capacitance meter is under quartz frequency control and capable of three measurement cycles a second.
19.99nF to 199.9nF, is applied to the measuring unit, IC12b. This means that the dc component applied to the converter is given by,

\[ \text{IC}_{\text{in}} = 0.45 \cdot C \frac{R_{14} + R_{15}}{(kV)T_0/\text{ns}} \]

\[ \text{IC}_{\text{in}} = 1.1 \cdot R_{11} \cdot C_{x} / R_{24} \]

This means frequency change is 1.49/(R4+2R5)C4, or 0.68Hz.

A greater frequency than 3Hz is not recommended, because the converter, IC19, measures the input voltage three times per second. This frequency \( f_{\text{change}} \) is the clock for the shift register IC6, based on the 74LS164, which must be reset at the beginning of every measuring cycle using push-button SW1.

Outputs of the register QA-E go high one by one, transmitting the different frequencies through open-collector gates IC7A-IC8A to the IC18B.

Outputs of the register QA-E go high one by one, transmitting the different frequencies through open-collector gates IC7A-IC8A to the

**Fig. 4.** Analogue to digital converter with seven-segment display for the low capacitance meter. The ICL chip is a dual-integrating d-to-a converter with integral display drivers.
clock-measurement-module port.

The same outputs, named module ports Q1-5, have other functions too, such as determining which measuring unit is chosen, and changing the place of the decimal point. Reaching the optimal range forces the range-set module port low and so the clock to IC6 is stopped due to the IC11A gate.

Gate IC11B has the same function as IC11A. It stops the clock if the minimal range 0.1pF to 199.9pF is reached, since Q6 is high. As a result, the output of IC9D is set low. Gate IC9A and the RS trigger, comprising IC9B and IC9C, are used to load the shift-register 74LS164 through the serial data entry B with logic one. This occurs during the first autoranging clock cycle of IC10 only.

A-to-d-conversion and control

The converter and some of the control circuits are shown in Fig. 4. DC component of the pulses, generated by the monostable multivibrators, is transferred to the converter IC19 via the input module ports IN-HI and IN-LO. The internal oscillator of ICL 7107 is not used. Rather, a 50kHz clock with quartz stability is applied to pin 40 of the IC, improving the measurement accuracy.

As Fig. 2 shows, the 50kHz clock signal passes via dividers in the range setting circuitry. This results in a repetition rate of three measurement cycle a second. The same clock, internally buffered and appearing on pin 38 of IC19, produces the negative voltage of about -3V, necessary for the correct operation of the converter. The circuit producing this negative voltage consists of a 4049 buffer, with IC18, capacitors C10,11 and the diodes D9,8, performing the work of a charge pump.

Capacitor C15 is the auto-zero capacitor. It provides compensation for the offset voltages of the input amplifier, the integrator and the comparator of IC19. The ICL7107 is an integrating a-to-d converter, with C14 acting as the integrating capacitor and R24 is the integrating resistance. Reference capacitor C12 is used for the deintegration or integration of the reference voltage toward 0V in the last phase of the measuring cycle of the converter. This technique is known as the dual integration principle.

Reference voltage of IC19 is set and adjusted via R20,22. Tuning of the low ranges 0.1pF to 199.9pF, 0.199nF to 1.999nF and 1.999nF to 19.99nF is accomplished with potentiometer R21. Tuning of the high range 19.99nF to 199.9nF and 0.199µF to 1.999µF is accomplished via potentiometer R14 in Fig. 3.

Indication of the measured voltage and so of capacitance C8, takes place by means of the four 7-segment indicators IC20A,B, IC21A,B (VQE24). Range-setting circuitry, which decides if the digital code is greater than 0199 or not, is based on IC22 and IC11C, the latter connected to module port called range set.

The position of the decimal point for the different ranges is selected by IC17B,D. Indication of the measuring units picofarads, nanofarads or microfarads, is achieved by means of IC16B, IC16C and IC17A. These devices light the three LEDs, D5,7.

The whole device is supplied from a +5V single supply delivering 400mA. Although the error of the 555 timer in monostable mode is typically 0.5%, in this case, accuracy can be compensated to 0.1%. The differential measurement technique helps improve accuracy and components can be tweaked to obtain best performance.
Ian Hickman explains various graphical methods of displaying circuit operation – including vector diagrams, Bode plots and poles and zeros.

For those of us not too mathematically gifted, graphical representations provide more insight than equations into how a circuit or device functions.

Some designers prefer vector diagrams, others Bode plots, and yet others circle diagrams or perhaps pole-zero plots. This article is intended to pull them all together, showing how they relate to each other, and illustrating how the same thing can be expressed in different ways.

If you are used to using just one of these mental props, please be assured that becoming familiar with all of them is definitely worth the effort. Each illuminates the others in ways that are not immediately obvious.

In this article, I shall be concentrating on the frequency response of circuits, that is, their response to a continuously applied sinewave of whatever frequency.

Some simple circuits

Figure 2a shows a simple resistive voltage divider. Output voltage $v_0$ is given by $v_0/v_i=R_2/(R_1+R_2)$. Since $i=V/R$, this is the case whether the input $v_i$ is dc or ac of any frequency. Figure 2b shows the case where $R_2$ is replaced by a capacitor, of value $C\mu F$. The same formula for $v_0$, holds, except that the value of $R_2$, in ohms, must be replaced by the ‘reactance’ of $C$, which is also measured in ohms. There is a minor complication, though, in that the value of the reactance of $C$ depends upon what frequency you are considering. At 0Hz (dc) it is infinite, so that $v_0=v_i$, while at infinite frequency, it is zero, so $v_0$ is zero. At any frequency $f\text{Hz}$, the reactance $X_C$ of the capacitor is given by $X_C=1/(2\pi f C)=1/(\omega C)$ in ohms.

The maths approach

Rather more of a complication is the fact that, unlike the case of a resistor, in an ac circuit the voltage drop across a capacitor is not in phase with the current through it.

Voltage across a capacitor is proportional to the integral of current. This means, if one is talking of sinewaves, that the voltage lags the current by 90°. This is indicated by the $-90^\circ$ angle between $i$ and $v_0$ in Figure 3, but as shown earlier, a $-90^\circ$ displacement of a vector is effected by multiplying by $-j$. So the reactance of capacitor $C$ becomes $-j/\omega C$, which is $1/(j\omega C)$, since $-j$ is $j/\omega$.

By tackling the operator $j$ onto the expression for the reactance of a capacitor in this way, one keeps tabs on the phase angle between the voltage and current automatically – the maths looks after it for you. Taking this on board, 

$$\text{Eqn 1: } v_0/v_i = \frac{1/(j\omega C)}{R+1/(j\omega C)} = \frac{1}{1+j\omega CR}$$

The vector diagram approach

Figure 3 shows vector diagrams for the simple $CR$ low-pass (‘top cut’) circuit of Fig. 2b. Since this is a series circuit, it is simpler to start with the current, as this is common to both components.

So draw in a unit current vector $i$, pointing to the right from the reference or ground point, here labelled $C$ for common. This current flows through a reactance $1/(j\omega C)$ or $-j/\omega C$, the $-j$ indicating a $-90^\circ$ displacement of the resultant voltage $v_0=X_C$, which is marked in as $CA$. Added to this is the volt drop $iR$ across the resistor, which is in phase with the current. This brings one to point $B$, the input; $v_i$ being indicated by the vector $CB$. The vector diagram has been drawn for the case where $X_C=R$, and since $X_C=1/(j\omega C)$, in this particular instance, $\omega=1/(CR)$; let this value be $\omega_0$. This value expressed in hertz is given by $1/(2\pi CR)$; let this be $f_0$.

Thus in the right-angled triangle $CAB$, $CA=AB$ so that if $v_0=1$ then $v_i=1.414$, or $v_2$, indicating that the output is 0.707 of the input or 3dB down. Output $v_0$ lags the current drawn from the source by exactly $90^\circ$ but as the circuit as a whole is not purely capacitive, but partly resistive, $v_i$ lags by less than $90^\circ$.

If the radian frequency were doubled to $2\text{CR}$, then assuming $i$ were unchanged - which would not in fact be the case - $v_0$ would become $CA^2=CA/2$ as shown, and $v_i$ would lag by a smaller angle, since now resistance dominates the circuit’s input impedance.

Conversely, at lower and lower frequencies, $\phi$ would get smaller and smaller as $v_i$ became ever closer to equality (in magnitude and phase) to $v_0$.

The Bode diagram alternative

Vector diagrams show circuit response at one particular frequency. You can superimpose the response at another frequency, as in Fig. 3, but the diagram cannot show the response at all frequencies.

The Bode diagram accommodates this, by separating out the amplitude and phase components of the response.

Amplitude and phase of $v_0$, the output for an input of unity, is expressed as $M\angle\phi$, $M$ (for ‘modulus’ or magnitude, written $|v_0|$) being the amplitude and $\phi$ (for ‘argument’, written arg
Back to basics

A sinewave is the most basic waveform there is, containing energy at one frequency only. This contrasts with other regular waveforms – such as square or triangular waves. These also contain energy at harmonics of their basic frequency. Noise-like waveforms can contain energy at all sorts of unrelated frequencies.

A sinewave can be represented by a rotating ‘vector’. A vector is a value having both a magnitude and direction, such as the force of the wind. In contrast, a ‘scalar’ quantity, such as the exchange rate of the pound to the dollar, has only magnitude.

Such a vector is shown at OA in Fig. 1a). Assume that it has unit length. Now imagine this vector rotating anticlockwise (by convention) about the origin O at a steady angular velocity ω, so that at t seconds after this, the vector has rotated to OB, and the value of the voltage at this instant is represented by its projection on to the horizontal axis, OC. This has been plotted out beneath the vector diagram as a function of time, running vertically downwards.

You can see that the voltage is +1V at t=0, falls thereafter to zero, where it is changing most rapidly, increases to −1V. It then falls again and so on. Since the voltage equals +1V at t=0, this actually makes it a cosinewave, but the shape is known as ‘sinusoidal’ and it is generally called a sinewave, since in general the instant that you define as t=0 is arbitrary.

Since we are dealing with a continuous sinewave, the graph of voltage against time will soon run off the bottom of the page, whereas the rotating vector OA can represent the waveform indefinitely without running out of paper. Vector OA rotates through position OB and on, right round the circle, returning to the start position when \( ω\cdot t = 360° = 2\pi \) radians. Since each complete circle, or cycle, represents \( 2\pi \) radians, the frequency of the sinewave in cycles per second is \( \omega/(2\pi) \), or \( 1/(2\pi) \) times the radian frequency.

Now there is a simple modification that you can make to the diagram, which makes it even more useful. Imagine that (notwithstanding that it is rotating) the vector is somehow drawn on the paper. With the vector rotating at much more than a few tens of cycles per second, or hertz, it will appear as just a blur. But now imagine the paper to be rotating clockwise at \( \omega \) radians per second. The net result is that the vector will appear to stand still as though frozen in time. The utility of this will become apparent soon.

The other item to recap concerns ‘j’. This is called an operator, since multiplying a vector by j performs a very specific operation upon the vector. It rotates it by \(-90°\), positive angles being measured, by convention, in the anticlockwise direction. This is illustrated in Fig. 1b), where the vector OA represents a cosinewave of A volts peak. Multiplying this by j takes us to the top of the diagram, giving a different vector jA.

Repeating the process, j\( jA \), takes the vector through another right angle, so that it now represents a voltage which is at its negative peak, \(-A\). Thus multiplying by \( j \) twice over has reversed the sign of the voltage. It follows that \( |A|j^2 = -A \), so \( j \) is called the square root of \(-1\). Multiplying by \( j \) a third time rotates the vector through \(-90°\) again, giving \(-jA\), and it can be seen from Fig. 1b) that this is the same as rotating the original vector A through \(-90°\), or multiplying by \(-j\). Note also that \(-j = -(j^2) = (-1)^1\), so that (dividing both sides by \( j \)) \( -j = jf \).

\( V_o \) the phase. These are plotted against a logarithmic frequency scale, so that \( 1/2\pi \) or zero frequency, is as far off the page to the left as infinite frequency is to the right.

Parameters M and \( \phi \) are plotted vertically, on separate graphs. For the phase plot, \( \phi \) is plotted to a linear scale, while for the amplitude plot, \( M \) is plotted logarithmically, ie in decibels.

Figure 4a) shows the Bode plot for the simple CR low-pass circuit of Fig. 2b), from which you can be seen that at \( f_o \), where \( \omega = 1/CR \), amplitude response \( M \) is 3dB down on the response at dc. At frequencies higher than \( f_o \), it becomes asymptotic to a line passing through 0dB at \( f_o \), with a slope of -6dB per octave or -20dB per decade.

Phase lag is seen to increase from zero at \( 0Hz \) to \(-90°\) at very high frequencies, passing through \( 45° \) at \( f_o \). In fact, on the logarithmic frequency scale shown, the phase curve is skew-symmetric about \( f_o \).

Modified Bode diagrams

The diagrams of Fig. 4a) have been replotted in Fig. 4b) with a linear frequency axis and, in addition, the vertical axis in the \( M \) plot is now infinite.

As frequency increases from \( 0Hz \), the magnitude is initially flat, but soon starts to fall, being asymptotic to zero at very high frequencies. Note that here \( T \) indicates the time of one cycle of the sinewave, ie its period so that \( 1/T = f_o \).

On the other hand, the phase lag increases with frequency right away, levelling out all the while and reaching \( 45° \) at \( f_o \), while becoming asymptotic to \(-90°\) at very high frequencies. Interestingly, if the initial rate of increase of phase lag were maintained, \(-90°\) would be achieved at \( f_o \times \pi/2 \), at which frequency the phase lag is actually just one radian or \( 57.3° \). I won’t call these Hickman Diagrams as surely someone else has already described them, but I don’t recall ever seeing them in any textbooks.

The magnitude plot, \( |V| \), will reappear later on, however, in connection with pole-zero diagrams.

The circle diagram approach

Circle diagrams are widely used in the heavy electrical field, but also very useful in light current engineering – electronics. They are an extension of vector diagrams in which is is possible to show the behaviour of a circuit at all frequencies, rather than just one or two.

In Fig. 2, the current was taken as the starting point, convenient in a series circuit as the same current flows through both components. Fixing the current like this is equivalent to making \( V_o = V \) the reference vector, since for a given current, the potential drop \( V \) across the resistor is independent of frequency. Alternatively, either \( V_o \) or \( V \) could be fixed. In a circle diagram, \( V \) is held constant.
In Fig. 3, the voltage drop across the capacitor is at right angles to (in quadrature with) that across the resistor, and this must always be so, whether you are taking the case where $\omega=1/(CR)$ indicated by the vector CA, or $\omega=2/(CR)=2\omega_0$, vector CA', or any other frequency.

A right-angled triangle with its longest side formed by the diameter of a circle, will have its apex lying on the circumference. So redrawing the triangle CAB of Fig. 3 starting with $V_o$, the point A will lie on a semicircle as in Fig. 5a. It is shown there as $\omega_0=1/(CR)$. Output vector $V_o$ is also shown for several other frequencies, above and below $\omega_0$. Thus the circle diagram plots the tip of the output vector $V_o$ as a function of the radian frequency $\omega=2\pi f$, around the circumference of the semicircle.

A useful simplification is to 'normalise' all the quantities. In Fig. 5b, the radian frequencies marked around the circumference in Fig. 5a have all been divided by $\omega_0$, showing the frequencies relative to $\omega_0$. Normalisation can usefully be taken to the limit, with $V_o$ fixed at 1V peak, $R$ set to 1\Omega and $C$ to 1F. This makes the sums very easy, with $V_o$ in Fig. 2b) equal to 1V at 0Hz and $i$ is 1A at 0Hz, or infinite frequency. Thus $\omega_0=1/(CR)$ then gives just 1rad/s, or 0.159Hz.

You can always denormalise to the frequency actually to be used, and adjust $C$ or $R$ as required, later on.

### Using poles and zeros

In the vector diagram of Fig. 3, $V_1-V_o$ of Fig. 2 was the thing that was kept constant, whilst in the circle diagram of Fig. 5, it was $V_o$. There remains just one possibility – keeping $V_o$ constant. This forms a useful introduction to pole zero diagrams.

Figure 6a) shows the vector diagram of Fig. 3 redrawn, with $V_o$ as the horizontal reference vector. At zero hertz, the reactance of the capacitor is infinite, so the current $i$ is zero. The volt-drop $iR$ across the resistor is therefore zero, the angle $\phi$ is 0°, and $V_o$ is $V_o$.

As frequency rises, the reactance of C falls, $i$ rises and so does the voltage drop $iR$. Thus $\phi$ increases and so does $V_o$, the point B migrating ever upwards. $V_o/V_i$ is the 'transfer function' of the CR low-pass circuit, the transfer function possessing both a magnitude and a phase, $M_\phi$.

Value $V_o$ is a measure of just how much input voltage is required to produce an output voltage $V_o$ of unity, at any given frequency. Thus $M$ is inversely proportional to $V_i$, i.e. to length CB. Also, $\phi$ is the angle by which the input leads the output, increasing to 90° at infinite frequency, by which time point B has disappeared way off the top of the page.

Figure 6b) shows a) redrawn, superimposed on a set of axes, $\omega$ in the horizontal (real') direction and $\omega+\phi$ in the vertical (imaginary') direction. These define the 'complex plane' or $s$ plane. Any point upon it can be defined by the appropriate values of the $x$ and $y$ or rather $\phi$ and $\omega$ - co-ordinates. Together $s+\omega\phi$ are known as the complex frequency variable 's', and the the transfer function is a function of $s$, written $F(s)$.

In accordance with the policy of keeping things simple by normalising everything, assume that the point C is located at the point $s=-1$ on the horizontal axis, i.e. at the point $s=1+j0$.

From Equation 1, given $R=1/\Omega$ and $C=1F$, then when $\omega=1, V_o/V_i=1/(1+j)$. This is shown in Fig. 6 c), for the radian frequency 1 on the $\omega$ axis, so $V_o$ has magnitude $M=1/\sqrt{2}$ and phase $\phi=45°$. 

### Fig. 6a) The vector diagram of Figure 3 redrawn taking $V_o$ as the reference.

a) redrawn, superimposed upon 'real' (c) and 'imaginary' (jo$\omega$) axes.

b) drawn for the specific case where $R=1/(\omega C)$, i.e. a normalised frequency of unity, where the attenuation is 3dB ($V_o=V_o/\sqrt{2}$) and the phase lags 45°.
Indicates a pole

ϕ=45°, lagging behind \(v_i\).

To make Equation 1 into a true transfer function, one must substitute \(s\) for \(jw\). Then, for the circuit of Fig. 2 b), \(F(s)\) becomes

\[ F(s) = \frac{1}{s+1}, \]

assuming all values normalised, to keep things simple.

You can evaluate the magnitude \(M\) of \(F(s)\) for any value of \(s\), and plot it as a point up in the air, above the corresponding point \(s\) in the complex plane. Taking the normalised expression \(1/(s+1)\), when \(s=j1\) (\(\alpha=0\)), the answer is \(0.707\), while when \(s=-1\) (\(ja.\rightarrow 0\)), the answer is infinity.

Plotting the magnitude of \(F(s)\) in this way for all possible values of \(s\), gives a three dimensional surface above the \(s\) plane, which I have crudely sketched out in Fig. 7a). This shows only the part of the surface to the left of the \(jw\) axis (i.e. for negative values of \(\alpha\)), with a vertical section through it, along the \(jw\) axis.

The surface can be imagined as an enormous rubber sheet, nailed down to the ground all the way round the edges. At \(s=-1+j0\), the denominator of \(1/(s+1)\) becomes zero, so the expression explodes to an infinite value. This can be imagined as an infinitely high tent pole, propping up the rubber sheet at this point. There is said to be a pole at \(s=-1+j0\).

The vertical section through the surface, along the \(+jw\) axis, gives the magnitude of the transfer function for sinewave inputs to the CR low-pass circuit. It is in fact identical to the plot of magnitude \(M\) in Fig. 4b).

Note that for this circuit there are no terms in \(s\) in the numerator, so there are no values of \(s\) for which the numerator could become zero. If there were any such zeros, they would be like thumb tacks, pinning the sheet to the floor at those points.

Figure 7b) shows a) in plan view, the pole being conventionally indicated as a cross. The pure mathematicians tell us that for every pole, there must be a zero. The reason it does not appear on the diagram is that it is at infinity. Letting either \(\sigma\) or \(\omega\) go to plus or minus infinity sends \(F(s)\) to zero; a zero at infinity tacks the rubber sheet down all the way round.

While the section through the surface along the \(+jw\) axis gives the magnitude \(M\) of the transfer function, as noted earlier, \(\phi\) is given by the angle between the line from a point on the \(jw\) axis to the pole, and the \(\sigma\) axis, counting the angle as indicating \(v_o\) lagging \(v_i\) as you move anticlockwise relative to the pole.

This article has just scratched the surface of the subject. Where things really start to get interesting and illuminating is in considering vector, Bode, circle and pole-zero diagrams as applied to other circuit types, such as CR high-pass, transitional lag and all-pass circuits, not to mention circuits of second and higher order. But alas my space has run out.

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The Future Is Interactive!
New relay for POWER AUDIO

A new design of relay provides an ideal building block for speaker protection circuits by offering high power capability without the traditional adverse impact on tonal quality. Omron's Barclay McKenna discusses the relay design, and outlines a specific circuit implementation.

Perhaps it shouldn't happen - but it still does. Switch on the stereo and the first thing you hear is that habitual thump as the speakers flex in response to the inrush of current to or from the amplifier output.

Even in a thoroughly poor design, to be fair, it is not likely that the bemused music lover will be left with a pair of bass speaker coils dangling, terminally ripped from their mountings. But that still doesn't make the initial 'whump' a desirable system characteristic.

The sheer physical jolt of an amplifier's power circuits switching into life may not have an immediately visible effect on the life expectancy of a speaker but, on a cumulative basis, such physical shocks can lead to damaging displacement of the bass voice coil. The thermal shock of the inrush current can have an equally unwelcome effect on the integrity of the coil itself.

While it may be true that many low-frequency units owe their premature demise to excessive bass and volume settings - not to mention the owner's fondness for techno - the steady abuse of the coil and speaker cone by thumping amplifiers certainly speeds their demise.

Prevention of the damage is easily achieved by a variety of well tried means. Some involve the use of complex timing circuits in the power supply or output stages; the more successful, and more cost effective, interpose a relay or two between amplifier and speaker. The latter solution is often used, but it begs the question 'what sort of relay'?

Choosing a suitable relay

Broadly speaking, relays come in two variations: power types and signal types. The former will naturally handle high current surges associated with, say, industrial and some automotive applications. They are not, however, too concerned about the integrity of the signal waveform that passes through them.

If all you want is to switch power to a motor, then the odd alteration from input to output is hardly a consideration. Signal relays, on the other hand, will provide the sort of faithful transmission audio applications require. But signal relays tend to be limited in their power handling capabilities. Omron's new G5Z relay however has been specially designed with power audio applications in mind, combining the desirable attributes of both power and signal relays.

Crossbar integrity

Central to the design of the G5Z is its use of a crossbar contact design. This crossbar is a solid conducting bar pulled across the contacts. Ordinary power relays with comparable current handling capability usually incorporate button contacts. The solid conducting bar ensures the high degree of signal integrity required for audio applications.

In the case of the standard G5Z, handling capability runs to 5A at 40V ac, or 200VA. Quoted at a power factor of one, a small measure of down-rating is necessary to deal with the vagaries of a musical signal, but that still amounts to enough late-night Bach organ toccata to make the neighbours suitably miffed. For really heavy use, a 750VA alternative is available.

The reworked crossbar design has allowed this capacity for music to be shrunk into a very small capacity for circuit board space. The relay occupies no more than an 11.5 x 25mm footprint, with a height of 16.5mm. Moreover, those dimensions are for a two-pole design, as required for today's compact audio applications. Contact resistance is a maximum 50mΩ to eliminate any untoward effects on the amplifier's damping factor.

Building in relay

So how might a protection circuit built around the G5Z work in practice? The first thing to consider is the level of protection required. A simple RC network could form the basis of an on delay, but...
dc isolation might also be important. A failure in the amplifier's output transistors for example would cause a dc flow directly into the speaker, so a circuit which provided the means to monitor this and employ the relay as a protection device is advantageous, Fig. 1.

At the same time, a short circuit on the speaker lines could damage the amplifier, and could even cause a fire. An additional circuit monitoring amplifier output current would be needed for this (not shown in the diagram).

The circuit shown addresses both the time delay on power up and the provision of dc isolation capability. Left and right outputs from the power amplifier connect to the speaker crossover networks via the G5Z relay.

Relay driving is provided by a Darlington pair transistor configuration, T1 and T2, with diode, Di inserted across the relay coil for back emf protection.

On delay timing

The basis for the on delay timer is the resistor-capacitor network of R1, R2 and C1. At power up, transistor T2 is off. As the capacitor charges, the potential at its positive terminal rises, eventually reaching the point where T2 switches on. Zener diode D2 increases the degree of switching certainty and stability by ensuring minimal leakage current until its threshold voltage is reached, at which point its reverse current immediately rises. At this point, T2 is turned on, activating the relay.

The rate at which C1 charges - and hence the time delay before the zener's threshold voltage is reached and T2 switches on - is determined by the values of C1, R1 and R2.

Fault protection

As well as introducing delay in connecting the speakers to the amplifier output, the circuit shown also addresses disconnecting the speakers in the event of a problem, via a dc monitor. Left and right outputs of the power amplifier are fed into a summing and ranging amplifier to provide a combined signal output, with a virtual earth configuration ensuring zero interaction between the two channels. Feeding this signal into a low pass active filter reveals the presence of any dc level. Cut-off frequency of this filter is typically less than 1Hz.

A zero-loss active rectifier enables detection of both -Ve and +Ve dc offsets, and the signal is then fed into a comparator circuit. Reference voltage for the comparator is set via the preset VR1 and resistor R3. A typical trip level of ±1V on the amplifier output would ensure safety of the system, and it should be borne in mind that a small offset, perhaps ±200mV, is normal.

In problem-free operation, there would be an off output from the voltage comparator. As a result, transistor T3 would be off, having no effect on the relay circuit. Should a dc level be sensed at the amplifier output however, the voltage comparator output would go high. This would switch T3 on, and effectively short circuiting R2 and C1. Consequently, T2 is switched off, and the relay coil de-energised, so disconnecting the speakers from the amplifier output.

Switch SW1 could be incorporated in the circuit to provide a manual method of electrically isolating the speakers without powering down the amplifier.

Omron is so confident of the relay's capabilities that output from its factory has been boosted to nearly a million a month, with major audio manufacturers already designing-in the device as standard!

Further information can be obtained from Omron Electronics by ringing 0181-450 4646 or faxing 0181-450 8087.

**Fig. 1. Circuit for addressing time-delay-on-power-up and dc-fault-isolation problems. Heart of the circuit is the G5Z relay. It is designed specifically for audio applications, providing high power speaker protection capability without adverse impact on tonal quality.**
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Electronics World

July/August 1996

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Paul Mapp has been looking at a new a-to-d converter module that turns your pc into a 25MHz spectrum analyser, a 50MHz dso, a multimeter or a data logger - all for around £500.

**Virtual instrument software**

Two main programs are supplied with the ADC200 - the familiar PicoScope for dos and the new, and welcome, PicoScope for Windows. Drivers for C, Pascal, Basic and Windows, along with disk based documentation, are also provided for serious programmers.

Installation instructions are clear in the user manual and can be performed from the dos command line or Windows program manager. Both of these methods can install dos and Windows versions of the software. Multi-lingual support is offered, but was not included with the review software. Installation under dos, Windows 3.1 and Windows 95 caused no problems.

Pico currently has no plans to release a printed manual for the Windows software. This can be easily overcome using one of the 'Help-to-Doc' programs available from many shareware libraries. I use HLP2DOC by Wolfgang Beyer, which converts Windows help files to Microsoft Word 2 format documents and is available as freeware.

**Operating the module**

The glossy information leaflet provided by Pico describes the ADC200 as a '50Msample/s dual-channel storage oscilloscope, 25MHz spectrum analyser and multimeter' which runs via a parallel port on a standard IBM pc compatible computer. This is not strictly true but is the usual advertising format used by most manufacturers of this type of equipment.

There are actually two versions of the ADC200 - the ADC200-50 and the ADC200-20, as detailed in the panel. The former is a 50Msample/s unit which only provides a single channel at 50Msample/s, the dual channel mode operating at 25Msample/s. The ADC200-20 that was supplied for review, providing a dual channel, 20Msample/s oscilloscope and 10MHz spectrum analyser.

The dos software has been supplied with Pico's a-to-d converter range for a while and is very easy to use. It operates well on a modest pc compatible. I have used the same software with an ADC100 on a 25MHz 486SX lap-top computer, where it excels as a portable diagnostics and capture tool. Use with the ADC200 appears identical, apart from the faster time-base, and it operates adequately on the SX25 lap-top.

The manual is clear and concise, explaining the use of the four modes of operation - meter, 'scope, XY 'scope and spectrum analyser - in detail, Fig. 1. Context sensitive help is available by default but may be hidden, providing a slightly larger display.
Mouse pointing is not supported but is not really needed, plenty of keyboard shortcuts being available for access to the panel options. In most cases the space bar toggles or steps through the available values in the selected option, which is especially useful when used with the run/stop option which is available in all panels.

The top level panel, presented when the software is first run, is the mode menu, from which you can select one of the four operating modes. Below this, selected by the page-up and page-down keys, are mode specific panels. Oscilloscope mode panels offers timebase, trigger modes and channel gain options while the spectrum mode panels offers sampling, trigger type and display format options.

In meter mode up to six separate measurements can be selected and displayed on seven segment format digital displays.

Common features simplify use

Common to all modes are panels allowing axes annotation to be set as required, notes to be added to the display, rules (markers) to be added for measurement and data to be filed or printed. In addition the converter provides a signal panel which allows a programmable logic level signal output to be provided on the external trigger connector when enabled. Frequencies in the range 1Hz to 250kHz can be set, which can be used as the external trigger if this is also selected.

The rules panel allows on-screen event markers to be placed, displaying event times and, with two rules, differential time. The display is fast and responsive, providing clear graphical data very close to that of a traditional oscilloscope or spectrum analyser. Oscilloscope mode provides inputs down to 10mV per division and timebase speeds to 1µs per division, axes magnification to x10 and delayed trigger modes to examine data captured away from the trigger point.

In spectrum analyser mode, Fig. 2, sample rates to 20Msamples/s are available, user selectable frequency, volts or decibel scales and a number of sampling window types. The default window is the Blackman type. Six other window types are available, as is signal averaging to reduce the effect of noise on the display.

Meter mode provides measurement of ac and dc volts, decibels relative to 1Vpk-pk and frequency. Six different measurements can be displayed simultaneously. Display updating slows dramatically when multiple meters are displayed, but this is to be expected.

A comprehensive range of printing and filing options are available. Printouts can be set for a variety of printer types and one of three formats – portrait, half portrait and landscape. Printouts are both clear and finely printed, user selectable slow rate when no trigger is present; repeat triggers only when the trigger condition is satisfied; single triggers once and stops. The remaining controls select the trigger source (signal channel A or B or external, E), trigger edge (rising or falling), trigger level and timebase delay.

Triggering options and delayed sweep

Triggering can be set to one of four modes; none allows the ADC200 to free run; auto triggers when the trigger condition is satisfied but runs at a user selectable slow rate when no trigger is present; repeat triggers only when the trigger condition is satisfied; single triggers once and stops. The remaining controls select the trigger source (signal channel A or B or external, E), trigger edge (rising or falling), trigger level and timebase delay.

Trigger level can be incremented and decremented, or manually entered if required. A marker on the appropriate channel axis indicates the set level. Timebase delay can also be incremented and decremented, or set manually. Trigger positions from –100% to +100% of the display period are available, moving the horizontal axis to suit.

One slight catch for the unwary when manually entering the delay is that the minus key is an illegal character if entered on its own. Delay must be entered and the minus added using the left cursor or home key.

One or two markers can be added to the display by simply placing the mouse cursor on the display, pressing a mouse button and dragging horizontally or vertically to produce horizontal or vertical lines.

Blinking distraction

When using the mouse cursor over an active display, a few display cards cause the cursor to blink with the display update. Cirrus Logic...
Running under Windows

The new Windows software is impressive as it manages to run real time displays at a decent speed in a scalable window, Fig. 3. This has always been difficult to achieve due to the enormous graphics overhead of Windows, so congratulations to Pico.

I am running PicoScope under Windows 95 on an AMD 486DX4-100 processor with graphics acceleration disabled, which really should be considered a minimum specification for use with this software. While PicoScope for Windows will run quite happily under Windows 3.1 on the 486SX25 laptop mentioned earlier the performance is not really fast enough for anything other than very basic use, the dos software giving a superior performance in this type of environment.

The upper control bar has icon buttons for the three modes of operation, namely oscilloscope, spectrum analyser and meter. Clicking one of these buttons opens a new window in the mode requested, leaving any existing windows unchanged. The remainder of the upper bar has drop down menus specific to the current mode of operation, e.g. timebase, display multipliers and volts per division for scope mode.

chipsets cause this effect, which appears to be a function of the display hardware rather than PicoScope. Once positioned and a mouse button pressed, the actual rules themselves are perfectly stable so this does not detract from the usefulness of this facility.

Many other less frequently used options are provided on standard drop down menus at the top of the PicoScope window. These include all the facilities provided in the dos software plus a few new features unique to PicoScope for Windows, such as multiple operating windows. Many settings are also available via the function keys. There are also some interesting additions only possible in this sort of environment.

Traces can be saved after triggering, the ADC200 automatically rearming for the next trigger. Data averaging can be performed and successive cycles overlaid. This is useful for capturing infrequent deviations from the norm. There is an option for redraw or roll (chart recorder) mode at low timebase speeds. This option was not properly functional in the software reviewed but no doubt Pico will correct it.

The preceding discussion has centred upon the oscilloscope mode, but applies equally to the spectrum and meter modes which are also available in PicoScope for Windows. A unique feature in the Windows version is the ability to open multiple views of the same signal, Fig. 4. This ability provides the opportunity to replace multiple instruments operating simultaneously. Figure 4 shows an oscilloscope window, spectrum window, frequency meter and voltmeter all operating from the same input, each with their own independent parameters. The disadvantage of this mode of operation is slower operation.

Overall performance

As the Pico dos software is tried and tested, the following comments concentrate on the ADC200 with the new Windows software. Starting with dc measurements, e.g. a quick test of power supply levels during debugging of a circuit. Applying a dc voltage to either input A or B with the switch set to dc, produces the expected level on the display. The ADC200 has a ±20V input range and is protected to ±100V. Calibration appears to be good.

A couple of minor problems appear here. If trigger mode is set to auto, as will normally be the case, other than when only measuring dc levels, the update rate drops to the user specified repeat rate. In practice this occurs a couple of times a second even though it can be specified up to ten. This is due to the triggering scheme used in the hardware. You need to select a trigger mode of 'none' to get a rapid update. Also, a voltage level outside the display range, e.g. applying 9V on a 5V range, shows a horizontal line at the maximum level, i.e. 5V. This is unlike the dos software or a traditional oscilloscope, where the section of the trace out of range is not displayed. This could cause confusion.

Moving up the scale to audio measurements, operation appears smooth and reliable. Triggering is easy and relatively stable although there appears to be a tendency to trigger on the wrong edge every now and again, especially in dual trace mode. This has now been fixed says Pico. Comparing the traces captured in PicoScope with those seen on a traditional oscilloscope shows no noticeable deviations. All the facilities provided work as expected, giving a useful and easy to use replacement for the traditional oscilloscope. In practice this applies throughout the range to several megahertz.

By the time video signals are reached, the 20MHz ADC200 is beginning to show its limitations and the 50 would be useful. The lower trace shown in Fig. 3 is from a cheap video signal generator producing a white crosshatch pattern on a black background. Effects of aliasing, due to the digital sampling, can be clearly seen on the amplitude of the sharp vertical 'spikes' which form the fine white lines of the crosshatch pattern on the video signal. These are uniform in amplitude.
Increasing the timebase to 1 μs/div, Fig. 5, showed limitations of the converter in the form of switching noise and distortion due to the digital sampling. However, Pico says it is sourcing a new converter that eliminates the problem.

Shorting the input to the ADC200 reduces this noise to approximately the level of the noise seen on the upper trace in Fig. 5. For a unit of this type this performance is quite acceptable so long as the user is aware of the limitations and their effect on displayed traces.

**In summary**

Overall performance of the ADC200 package is impressive, with only a handful of minor problems which do not, on the whole, reduce the usability of the unit. As long as the limitations of the hardware are understood it forms an extremely useful addition to the workbench, really coming into its own when used as a data capture tool.

The new Windows software performs well on a higher spec. pc. The 486DX4-100 used for the majority of the review should be considered a minimum specification for this type of software. PicoScope for Windows appears, on the whole, to be a competent package, providing a good range of well thought out features which are both clear and easy to use. As with many newly introduced software packages there are still a few rough edges which need to be ironed out and Pico are working hard to make sure this is done.

On the down side, the signal generator facility provided on the ADC200 did not function properly on the review unit. With both dos and Windows software it failed to produce the programmed frequency and was interrupted continuously by the a-to-d conversion process, which slowed noticeably when the signal generator was enabled.

**ADC200 virtual instrumentation**

### 20% reader discount

Pico Technology's ADC200 is a high-speed virtual instrument designed for use with a pc. Connecting via a printer port, the ADC200 includes both Windows and DOS PicoScope software, turning your pc into a 50Msample/s dual channel storage oscilloscope, a spectrum analyser, a frequency meter or a voltmeter. The ADC200 breaks the price/performance barrier for digital storage oscilloscopes. For less than half the price of the cheapest benchtop instrument, you get a fully-features oscilloscope with options such as FFT analysis and waveform storage/printing, which are features normally found only on the most expensive DSOs. Normally, the ADC200-50 sells at £499 and the ADC200-20 is £359 excluding VAT and postage. But Electronics World readers can obtain the units at the 20% discount prices of £399.20 and £287.20 respectively.

### Specifications of the ADC200 a-to-d converters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>ADC200-20</th>
<th>ADC200-50</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sampling</td>
<td>1 ch.</td>
<td>2 ch.</td>
</tr>
<tr>
<td>20Msample/s</td>
<td>50Msample/s</td>
<td></td>
</tr>
<tr>
<td>Buffer size</td>
<td>2x8k</td>
<td>1x16k, 2x8k</td>
</tr>
<tr>
<td>Resolution</td>
<td>8-bit</td>
<td>8-bit</td>
</tr>
<tr>
<td>Analog connections</td>
<td>2 channelx1MΩ impedance</td>
<td></td>
</tr>
<tr>
<td>Digital connections</td>
<td>AC/DC coupling via switch</td>
<td></td>
</tr>
<tr>
<td>Voltage ranges</td>
<td>±20, 10, 5, 1V</td>
<td></td>
</tr>
<tr>
<td>±500, 200, 100, 50mV</td>
<td>Voltage ±3%, time ±100ppm</td>
<td></td>
</tr>
<tr>
<td>Error</td>
<td>Event: None, rising, falling</td>
<td></td>
</tr>
<tr>
<td>Trigger modes</td>
<td>Source: ChA, ChB, digital</td>
<td></td>
</tr>
<tr>
<td>Timing: pre/post 1% increments</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

While performing well on its own with a single display window, PicoScope for Windows slows dramatically when multiple display windows are opened. If you have multiple applications open, with PicoScope running in one window, there is a very marked reduction in the pc's response time. The many online help screens in PicoScope also seemed to take a long time to appear when selected - several seconds on occasions.

In conclusion, the advantages of the ADC200 far outweigh the few minor problems observed by the author. No doubt Pico will address and correct these and provide us with a really useful Windows based virtual instrument. In the mean time, if the few Windows problems are just too much then the dos software works just fine.
Measuring SPEAKER CABLE differences

Ben Duncan explains how cable differences had been measured before they were modelled in his article in the February issue, and describes the technique and results.

Cable differences were possibly first publicly demonstrated au naturel at the Institute of Acoustics' Reproduced Sound conference in 1990\(^1\). Here, my friends Dr. Keith Holland and Phillip Newell used a dual-subtractor amplifier to make cable losses and errors audible in realtime on music program, as well as with complex test signals. In 1991, Davis\(^2\) demonstrated that cable resistance, the only parameter that hard line objectivists appear to acknowledge, was not the most critical parameter for audio. Even for bass frequencies, the cable's series inductance was of importance. Moreover, he argued that shunt capacitance across the cable had no malign influence, i.e. that contrary to popular opinion heightened values would not cause high-frequency loss.

Test procedure
In 1995, I devised a simple but quite realistic test for speaker cables that would show what is going on in a graphic sense, applicable to real music signals, with their perpetual discontinuity. The Dual-Domain version of the Audio Precision (AP) test set has a DSP-based FFT-test routine called 'FFT-80k' with 192kHz sample rate, which allows audio burst sinewaves to be graphed over time, Fig. 1 - analogous to a transient analysis simulation. A similar test but using a squarewave has been subsequently drawn to my attention in a less rigorous report by the high-fidelity accessories maker, QED\(^4\).

A 0.9Vrms test signal was used, enough to develop 100dB@1kHz at 0.4m, while representing only an 1/8th watt into the nominal 8Ω load, Fig. 2. The load included the associated, conventional two-way passive crossover. Inductors are air-cored, and capacitors are polypropylene types, expressly chosen by ex-Tannoy speaker designer, Mark Dodd, for low-hysteresis energy storage. My
labatory test amplifier, rated at a modest 150W/8Ω, has a conventional dual-pair mosfet output stage with high global negative feedback, followed by an air-core output inductor of less than 1μH. Steady-state output impedance at the output terminals and at the 1V output level is nominally quite low, nominally below 10mΩ, at least at 1kHz. To reduce the worst-case test contact resistance below this and the conductors’ own resistance, XLR connectors were soldered to both ends of the tested cables. Connections were also made with the test signal muted, to avoid degradation by arcing.

Measurement results

The 30 results of this test procedure for 10 cable types at three test frequencies are pictured in the test reports published internationally in both Studio Sound and in Stereophile, that were cited in the February article. Here, just the two cables modelled in the simulation (Feb ’96) have been abstracted: a fairly conventional mains cable, cable C, that has been widely used for wiring professional speakers, is compared with the low inductance SupraPly type, cable J. Although the conductors are similar in csa, the L and C parameters are quite different:

Cable C | Cable J | Approx diff.
---|---|---
Series resistance | 59mΩ | 50mΩ | 0.184
Loop inductance | 3.5μH | 2.0μH | 0.58
Shunt capacitance | 551pF | 50μF | 1.105

In other ways the two cables are similar. They are thin-stranded, and have PVC insulation and sheathing. But cable J has lowered skin effect, and includes a metal other than copper. Each of the graphs, Figs. 3 & 4, is a magnification of the point immediately after the sine-burst stops, as arrowed in Fig. 1. In each graph, one response is almost flat. This is the more tightly controlled response at the amplifier output, i.e. the stimulus end, as shown in red, in my Feb ’96 simulation plots. Deviations here reveal transient feedback control deficiencies. The outermost, wilder response is that at the speaker end. The different magnitudes directly show the degree by which cables impede the amplifier’s feedback damping control at the speaker terminals.

In summary

My measurements of August last year do corroborate with my simulations of reasonable equivalent circuits for stranded speaker cables, as published in February. The exact forward and reverse voltage of the diodes is relatively unimportant, but the diodic inter-strand contacts certainly exist, and the non-linearity they cause is measurable if one uses the right equipment. Ordinary %thd tests are quite unsuited as the data is lost in noise.

My tests moreover show how some of the cables expressly designed for speakers including, as it happens, one made by Graham Nalty, can significantly improve damping or settling time on music program. The results also illustrate the logic of making special cables for mains in emi sensitive environments, such as listening rooms, considering that current into all 50/60Hz conventional capacitor-smoothed ac-to-dc power supplies without pfc is a mid-frequency burst waveform, much as simulated here.

Jenving technology is on 0046 522 234 60, fax, 0046 522 23460.

References

2. Butler, T, Cable Controversy, Hi-Fi News,

![Fig. 2. The test setup uses a standard DSP aided test using de facto standard equipment from Audio Precision. The signal is read at both ends of the cable by the AP’s high common-mode rejecting receivers.](image)

![Fig. 3. Damping of ordinary pvc insulated mains flex, comprises 50/0.25 plain copper conductors of circa 99.7% purity, with high or non-specific oxygen content, leading to early onset of chloride poisoning and oxide complexing. Circular conductors in a sheath of circular OD. Compare to Fig. 11 in Feb ’96 EW.](image)

![Fig. 4. Damping of Jenving’s Supra Ply 2.0, which comprises 240 high purity copper (99.99% purity, oxygen free) strands specially impregnated with tin, with csa totalling 2.0mm2, in a rectangular, high mutual inductance profile. The quite thin, special PVC insulation has low emission of chloride ions. Outer sheath is ordinary PVC. Compare the rapid damping behaviour to that predicted by my model in Fig. 13 in Feb ’96 EW.](image)
EQUIPMENT DESIGN

Ray Morris* explores the issue of thermal management versus thermal engineering.

The operation of nearly every electronic device generates heat. If left unaddressed, it can cause problems at all stages of a product’s life-cycle.

Thermal issues are at the core of a major challenge facing the electronics industry. The intersection of two conflicting trends. One is end-user demand for faster and therefore hotter semiconductors and circuitry to power the next generation of consumer electronics and personal computers. The other is the demand for smaller packaging, which is creating a thermal situation that threatens manufacturers of electronic components, original equipment manufacturers, and ultimately the end-user.

Turn up the heat
Heat is a major problem that if left unaddressed can cause problems at all stages of a product’s life-cycle.

The operation of nearly every electronic device generates heat, from the microprocessors used to run today’s computers, to the mobile phones that have become common-place on city streets across the world. The laws of physics dictate that the performance and reliability of electronics and other integrated circuit devices are absolutely constrained by device temperatures. Mathematicians have worked out formulas that indicate that for every 10°C rise in junction temperature the failure rate doubles. Performance and reliability are jointly constrained by the manner in which electronic components are cooled and how the overall system attributes are handled.

There are many means by which a temperature-related equipment failure can occur: thermal runaway, gate dielectric, junction fatigue, electromigration diffusion, an electrical parameter shift, a package related failure, and more. In simple terms, heat can wreak havoc on electronic devices in many different fashions.

Unless we find innovative ways to disseminate substantially more heat, there will be many more instances of equipment failure — all of them temperature-related.

Thermal management: yesterday’s tool
Traditionally, the solution in the electronics world has been to slap a heat sink, fan or combination heat sink onto an application or into the enclosure as an after-the-fact way of dealing with a thermal problem. The resulting products tend to be bigger. They also tend to offer fewer reliability and cost/performance advantages to the customer, especially if their performance must be scaled back to manage thermal issues.

While these kinds of ‘thermal management techniques,’ as they can be called, have been adequate in the low-power arena, they no

*Ray Morris is with Aavid Thermal Technologies.
EQUIPMENT DESIGN

Fig. 2. Aavid’s proprietary thermal and fluid flow simulation programs – shown here modelling an electronic assembly without thermal management, top, and the same electronic assembly after Aavid’s optimised thermal solution, bottom – predict local surface temperatures and heat transfer coefficients. Graphic and tabular outputs from these programs allow engineers to optimise heatsink design and thermal solution sizing.

longer represent a valid option with next-generation electronics devices.

As you can imagine, thermal management techniques, while adequate in some situations, are more than likely to have a negative impact on overall design budget. This is because engineers are forced to go back to the drawing board to contend with thermal constraints discovered at the last minute.

Thermal engineering – an enabling technology

By identifying and ‘engineering out’ thermal issues early in the design process, engineers have much greater control over product build costs.

In fact, it is physics that permits the development of innovative solutions – but these solutions must be evaluated during the electronic design process to eliminate thermal problems before a prototype is developed.

This is called ‘thermal engineering’ – addressing system-wide thermal issues from the earliest stages of product design using a broad array of technologies and disciplines. In this way, manufacturers avoid the frantic scramble to address thermal problems that surface suddenly during product testing, just before the prototype is due at the customer site.

But thermal engineering is more than a way of eliminating heat. Thermal engineering is an enabling technology that offers manufacturers a strategic edge in systems design.

For example, suppose your goal is to run a Pentium in a notebook computer. By implementing a thermal design that permits the dissipation of 50 watts of heat – a small amount by high-power standards – you not only can have the Pentium operational, but you also have the additional capability for heat dissipation. You can run the Pentium at full speed along with the entire supporting ASIC chip set and throw in a video driver for an active matrix display. While this example assumes a top-notch electrical team, the advantages are clear.

Thermal engineering involves modelling of the device in the system – a process known as temperature prediction capability. In the thermal engineering process, designers look at the analysis of the device on the board or in the system. After a particular thermal solution is modelled, it must be subjected to what is known as either design or temperature verification.

After this, the designer considers the mechanical design aspects of thermal engineering, where the designer has to apply the correct manufacturing technology, or integrate the right manufacturing technology into the system and its available airflow to create the most economically and thermally optimum design.

Having the leeway to rely on a single thermal solution for multiple generations of product not only provides manufacturers with time savings and increased electrical design productivity, it also offers significant benefits...
in the design of additional features that may help a manufacturer get out in front of the competition.

**Today's thermal toolbox**

Thermal engineers are exploring a number of technology avenues that they expect will accommodate the electronics equipment of the future. Some of the solutions include heat pipe technology, fluid cooling, cutting edge air flow techniques such as focused flow and boundary layer optimisation. Also, new alloys that spread heat more effectively are being investigated.

As size and power issues collide to impose thermal limits on product development, thermal engineering promises to emerge as one of the fastest growing markets in the computer industry in the 1990s and beyond.

International Interconnection Intelligence – a Monterey, California-based research firm specialising in packaging issues – estimates the average annual compounded growth rate for thermal engineering products at 13 to 14% worldwide for the 10-year period from 1990 to 2000. This represents a jump from $2 billion to $6 billion.

**Who will provide the solution?**

Thermal problems affect every party in the manufacturing and supply chain. Most vulnerable is the consumer, who depends on the device's performance. To have a piece of equipment fail to the extent where it is unsalvageable undermines the working premise between customer, retailer and manufacturer. Businesses assume that the equipment on which they must rely every day will continue to function without catastrophic failure.

Equipment failure at the consumer level carries serious repercussions for the entire industry. Not only does the failure cast doubt upon the judgment of the retailer in offering such products to its customers, but it calls into question the role of the manufacturers in selecting components for its equipment.

Smaller, faster, feature-rich and literally hotter electronics emphasise the critical need for good thermal design to maximise product reliability and performance for the end user. However, the issues are not insurmountable — building tomorrow’s electronics begins at the product concept stage — and will become a reality only by fully taking advantage of cooling solutions available through thermal engineering.

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### ACTIVE

**Discrete active devices**

40A mosfet switch. A fast mosfet switch module for power application at up to 100V battery voltage by Lex exhibits a forward voltage drop of 0.002%, and is low and overvoltage are therefore not troublesome, in a three-phase circuit, forward voltage is less than 1.0V at 350A peak current. Module dimensions are 110 by 62 by 30mm. OD Rectifiers Ltd. Tel., 01444 243542, fax, 01444 879722.

1.7kV damper diode. Philips introduces the new BY479X-1700, claimed to be the first 1.7kV diode for use in television displays and multi-sync circuitry. The device coping with power requirements in the 12V and recovery time 200ns to 5V, the BY479X-1700 is designed for the amplification of photo-diode outputs in optical disk drives that use multiple diodes for tracking and focus as well as data. There are four 12Hz current feedback amplifiers with internal feedback, the only external component being a bypass capacitor. Since the current-feedback technique is, to some extent, independent of input capacitance up to 50pF, the device can be remote from the reading head while still giving a 14ns setting time for a 2V step. Gain accuracy is ±4% at 20mV, and offset error/drift is 25mV±2.5mV per °C. Micro Call Ltd. Tel., 01844 261930, fax, 01844 261678.

**Motors and drivers**

Induction motor controller. Micro Linear’s ML4421 is claimed to be the first single-chip speed controller for ac induction motors, running in forward or reverse and maintaining a selected speed to within ±5%. Speed is set by using a drive waveform at the required frequency at an amplitude to give optimum torque, a feedback circuit monitoring the phase of voltage and current drives and adjusting drive frequency and amplitude to maintain the set speed. ML4421 works at any frequency, including 400Hz. Ambar Components Ltd. Tel., 01844 261144; fax, 01844 261789.

**Oscillators**

Variable crystal oscillators. ACT announces the A7000 series of crystal oscillators with a pulling range of up to ±100ppm for 0.5-5V control input, depending on the model. Frequencies covered are in the 10-35MHz range and transient times are under 10ns. Stability is ±30ppm. The oscillators are hermetically sealed in a metal can with a profile of 8.5mm or less. Advanced Crystal Technology. Tel., 01635 528520; fax, 01635 528443.

### PASSIVE

**Passive components**

Aluminium electrolytics. From Philips comes the PSM-SI 056/057 series of miniature, snap-in, non-solid electrolytics, which have a life rating of 12000h at 85°C (21000h at 40°C), for operation the ±40 to 65°C area. Capacitance range is 47pF-680pF for the 056 series and 47pF-1.5nF in the 057, tolerance is ±20% and working voltage either 10-100V or 200-450V. They are charge and discharge proof and have a pressure relief on the casing. Gothic Crelation Ltd. Tel., 01734 788876; fax, 01734 776095.

**Connectors and cabling**

Test cables/connectors. Multi-Contact, represents by Electrospeed, has several new ranges of test cables, connectors and accessories. Among the connectors is the Multifim, fully shrouded, 4mm plug, which has a turned brass pin surrounded by a spring-loaded Multifim of hard-drawn copper alloy. It is for high-current use and meets IEC 1010-1, having a 1kV voltage rating, 32A current and test voltage of 4kV. Leads include the HS-5D flexible, double-insulated test lead cable for 4mm plugs, made from multi-strand wire and double pvc insulation and rated at 1kV, 32A. Electrospeed. Tel., 01703 645455; fax, 01703 610282.

Pcb terminal blocks. Camden Electronics offers the CTB0305 3.5mm rising-clamp terminal block range, which stands only 9.1mm above the board. Two or three pole interlocking units, which are also available assembled into a 24 pole block, take 1mm wire and are rated at 6A, 125Vac. Insulation resistance is over 40MΩ and dielectric strength more than 2.5kV. Camden Electronics Ltd. Tel., 01727 864437; fax, 01727 855400.

**Filters**

5m filters. 20A surface-mounted filters from Steatite have values up to 1000uF in C and Pi form. They avoid the need for through-hole filters and screening bulkheads and, being square in section, are suitable for pick-and-place equipment. Steatite Insulations Ltd. Tel., 0121 6436888; fax, 0121 6432011.

**Hardware**

Handheld cases. Britcident’s handheld and DIN cases are of ABS plastic and are meant to house equipment such
as remote-control handsets. They come in sizes from 56 by 36 by 16mm to 170 by 93 by 52mm, some of them being made from Noryl auto-extinguishing material and others having a Plexiglas cover; those destined to contain remote controls have one or two coloured buttons and a battery compartment. Depending on the model, there are options, including ventilation, an aluminium top or a black anodised front panel; standard colours are black, red, grey, ivory and beige and others can be provided. The company offers a Buyers' Guide, Britec Inter. Tel., 01425 474617; fax, 01425 471585.

Enclosures. Weidmuller EMC has a range of glass-reinforced polyester enclosures based on the standard models by Weidmuller Klippon. The Kestrel enclosures are internally coated with nickel or internally coated with nickel or Klippon. The Kestrel enclosures are standard models by Weidmuller polyester enclosures based on the has a range of glass-reinforced Emc enclosures. Weidmuller EMC Ltd. Tel., 01734 811571; fax, 01734 811570.

Servo chassis. Copley can supply the Model PST 'six-pack' chassis to take a mix of up to six of its servo amplifiers, complete with power, cooling, electrical protection and interconnection. It is supplied with an isolation transformer, rectifier and filter capacitor and rated at 750W for either 75Vdc or 150Vdc output. Up to three ac fans can be used and the chassis itself forms a heat sink. Cable harnesses are available to match any number of amplifiers and a resettable circuit breaker protects the whole chassis, individual amplifiers having their own internal protection. Copley Controls. Tel., 001 617 329 8200; fax, 001 617 329 4055.

Industrial PC enclosures. Arcom has APC-CE/1 and APC-CE/2, two new CE-compliant enclosures for PCs to be used in light industry or domestic applications, or in heavy industrial applications. The racks take up to 12 boards and give a clean, filtered ac supply carrying no noise and causing no emissions. A screened, fan-cooled housing prevents radiation and there is provision for floppy and hard drives. A range of PCB/PC-ACT boards is available, including 486 processors, i/o and utilities. Arcom Control Systems Ltd. Tel., 01223 411200; fax, 01223 410457.

Am function generator. TTI's TG230 2MHz function generator includes an internal amplitude modulator, also accommodating external am input to give 0-100% modulation depth. An internal sweep generator gives linear and log frequency sweeps with sweep ratios of more than 1000:1, start and stop frequencies being digitally set and indicated. Again, an external input is accepted and will provide frequency modulation. Thurby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Pcb troubleshooter. Huntron's ProTrack 1 is the first in a new series of programmable printed board troubleshooting instruments, combining signature analysis with some new hardware and software. Model 20 is a bench-top instrument which is used with the Scanner 1 to scan up to 128 pins at a time, making real-time comparisons between identical channels is 20MHz and the maintenance, education and enthusiasts. Bandwidth of the two digital generators to provide differential amplifier. A differential amplifier by Preamble Instruments, the Preamble 1855, is intended for signal conditioning for oscilloscopes or digitisers, providing differential measurement capability to instruments with single-ended input. Gain is steerable to 10 or 1 and there is 10:1 input attenuator, Bandwidth is 100MHz which can filtered internally to 10kHz, 1MHz or 100kHz by pole limit filters, gain is automatically displayed. Its built-in voltage generator can be set to any voltage in the ±15V range to within 100pV, being connected to the inverting input of the amplifier to form a comparator amplifier or applied internally to provide true differential offset. Thurby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Test and measurement

Differential amplifier. A differential amplifier by Preamble Instruments, the Preamble 1855, is intended for signal conditioning for oscilloscopes or digitisers, providing differential measurement capability to instruments with single-ended input. Gain is steerable to 10 or 1 and there is 10:1 input attenuator, Bandwidth is 100MHz which can filtered internally to 10kHz, 1MHz or 100kHz by pole limit filters, gain is automatically displayed. Its built-in voltage generator can be set to any voltage in the ±15V range to within 100pV, being connected to the inverting input of the amplifier to form a comparator amplifier or applied internally to provide true differential offset. Thurby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Digital, real-time oscilloscopes. The Tektronix TDS range of oscilloscopes samples signals at five times analogue bandwidth on two channels simultaneously to reduce the possibility of aliasing other effects. There are three models to give bandwidth/sampling rates of 100MHz/500MHz/2Gsample/s(TDS340); 200MHz/1Gsample/s(TDS320); and 400MHz/2Gsample/s(TDS380). Each has two input channels with a dedicated oversampling digitiser and an FFT button to convert the normal time-domain display to a spectrum. The 260 and 280 have a 2.5-in disc drive to allow the saving and recall of waveforms, which can then be passed to a PC for further work. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 01273 570215.

500Msample/s. DataSYS 944 by Gould is a 400MHz bandwidth digital storage oscilloscope sampling a 500Msample/s to capture signals down to 2ns wide. Bandwidth is guaranteed over the whole sensitivity range of 2mV/div to 5V/div. Noise levels are 'significantly lower' than that of earlier instruments. Each of four channels has a 50k memory An optional internal plotter provides hard copy and IEEE-488 and RS-423 interfaces allow transfer to a PC, where the company's Transition 2 software can be used to compile reports and carry out analysis. Gould Instrument Systems Ltd. Tel., 0181 500 1000; fax, 0181 501 0116.

Wide-band preamplifier. EG&G Instruments has introduced the Model 5185 wide-band 200MHz amplifier for...
Instrumentation use. Frequency response is 0-200MHz, with switched gain of 10 and 100 and selectable 50Ω or 1MΩ input impedance. There is an auto-adjustable voltage offset to remove dc levels before the output. There is an external, remote power supply or the unit will accept ±15Vdc supplies. E&G Instruments Ltd. Tel., 01734 773003; fax, 01734 773493.

Literature
Ac lvdts. Macrosensors has a 4-page brochure on its line of variable differential transformers for oem and end-user application. Three ranges are available: two of them are drop-in replacements for units by other makers and the other units are 0.375in diameter types for use where small size is needed together with the performance of a larger unit. Eurosensor.

16-bit micros. A short brochure from Toshiba describes the TLCS-900 family of 16-bit microcontrollers, which allow designs to be optimised for cost, power consumption, low-voltage operation or processing power for the particular application. The brochure gives details of the cpu core and describes all the options, a chart highlighting the most appropriate type for a given task. Development tools are also described. Toshiba Electronics UK Ltd. Tel., 01276 694600; fax, 01276 694800.

Satellite communications. California Microwave has a set of brochures to publicise its services and hardware and software products in this field, ranging from complete earth stations and networks to remote control and monitoring software. California Microwave. Tel., 001 516 272 5600; fax, 001 516 272 5500.

Power supplies
Ups for poorer mains. Fiskars' new PowerRite Max uninterruptible power supply uses the company's DoubleBoost voltage regulation to cope with mains fluctuations of 35% without switching to battery operation, a feature that possesses advantages in server availability, in particular when the mains supply is not of the best quality. The Ups also has Fiskars' advanced battery management system which is claimed to prolong battery life by 50%, periodically tests the battery and recharges it 20% faster than is usual, so that the battery is ready for the next catastrophe. The Ups is supplied with Lantiscile III network power management software which shuts down all networked devices in the order determined by the user. It is available in output power from 450VA to 1500VA. Fiskars Electronics Ltd. Tel., 01734 306600; fax, 01734 306566.

500kHz switching regulator. Linear has introduced the LT1376, which is a constant-frequency pwm type switching at 500kHz and having a 1.5A output at nearly 90% efficiency. The high frequency allows the use of 4.7μH inductors and capacitors can be used to produce a completely surface-mounted regulation circuit range is 6V-25V and there is a logi-operated shutdown mode to 20μA. Another version, the LT1375, can be frequency-synchronised over the range 580-900kHz by a logic-level clock. Micro Call Ltd. Tel., 01844 261939; fax, 01844 261676.

'Smallest' split power supply. From Maxim, the MAX685, a cross charge-pump dc-to-dc converter producing both positive and negative rails from one input and contained in a μMAX package, which is 1.1mm high and takes up half the board space of an 8-pin SOIC. Input voltage is 1.5-6.2V to produce two rails of twice the input voltage, each putting up to 10mA from 75μ. Another version, the MAX864, has pin-selectable oscillator frequency and logic-controlled, 1μA shut-down. Maxim Integrated Products UK Ltd. Tel., 01734 303388; fax, 01734 305511.

Ac/dc power supplies. Coulant-Lambros has a new series of power supplies, the SWT range, that incorporates universal and auto-selectable models giving 30-100W of power. The series is pcb mounted and handles inputs of 85-265Vac, conforming to EN61522 Class B for conducted emission noise. All conform to UL1950, CSA22.2:234 and EN60950 and the European low-voltage directive. Coulant Lambda Ltd. Tel., 01271 865656; fax, 01271 864984.

30mV dropout regulator. Zetex's 2LDO linear regulator has a dropout voltage of 30mV, which is claimed to be 10 times less than competitive devices. Currently, there are 3.3V, 4.85V and 5V models available, with other models offering up to 16V to come later. Zetex has a sensing circuit to warn when the input drops to within 300mV of the output. In normal working, the regulator takes 600μA for a 300mA load and, in sleep mode, 1μA. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5467.

1A voltage regulator. E71117117A are 0.8A and 1A positive voltage regulators for use in active SCI5 terminations and low-voltage microcontrollers and as spis post regulators. They give a 1.2V maximum dropout voltage and ±0.2% regulation with any line, load or temperature variation. Adjust pin, current is 0μA or less, output voltage 2.85-3.3V and input voltage 7V. Semtech Ltd. Tel., 01952 773520; fax, 01952 774781.

0.05% voltage references. Precision references in the Linear Technology LT1236 family offer 5ppm/°C drift (worst case), less than 1ppm pk-pk noise, and 0.05% output accuracy. Two devices are available to provide 5V and 10V, both sourcing and sinking up to 10mA and settling quickly on transient loads. Line and load regulation are 0.5ppm/V and 12ppm/ma. Linear Technology (UK) Ltd. Tel., 01276 677676; fax, 01276 64851.

DC-to-DC converter with filtering. Analog Device's new hybrid ADDC2080SS power converter provides ±5V, ±25A from 28V and has its own integrated filtering circuitry and many other protection and system functions, all in a 2.5 by 1.5in package. It gives 100W continuously from a 16-50Vac input range. Facilities offered include thermal monitoring and shutdown, input transient protection, an output status pin, inhibit and sync, current sharing and an input-referred auxiliary voltage for external circuits. Analog Devices Ltd. Tel., 01932 266000; fax, 01932 247401.

Radio communications products
Data transmitter. Needing no licence, a u/f data transmitter from A&R Electronic Developments operates on 458kHz, allowing 32-channel 500baud transmission over several miles. Two versions provide 12.5kHz or 25kHz channels. Output power is 4mW or 120mA. A&R Electronic Developments Ltd. Tel., 0121 643 6999; fax, 0121 643 2011.

Switches and relays
Binary rotary switches. Grayhill Series 94 binary-coded rotary dip switches are now available in the UK. The switches are completely sealed and are in either surface-mount form or for through-hole mounting, with both perpendicular and right-angle mounts. All are available with octal, binary-coded decimal or hex, codes, output standard or complement. Initial contact resistance of the gold-plated Eurocard supplies. Enetron's EN and ED series of Eurocard switched-mode power supplies deliver 28W to 400W with many combinations of single, dual, triple and quadruple output, all LVD-compliant and meeting EN50022 for emi. Input to the EN range is 88-127Vac or 184-265Vac or, optionally, universal. All are provided with a led status indicator and voltage adjustment and a range of optional facilties such as fail power, remote on/off and safety covers. EN supplies provide the same output facilities, but accept 10-400Vac input. XP plc. Tel., 01734 845515; fax, 01734 843423.

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Flash disk software. Arcom offers AFFS, which is flash disk filing software for PC bus boards to allow the emulation of conventional read/write disks by solid-state flash memory in a system that is capable of running dos or Windows on diskless systems. It is compatible with the Flash Translation Layer specification for PCMCIA cards and can be used in systems made from STibus, ISA (PC/PCI)bus or VMEbus boards. Silicon disks of any capacity can be formatted and up to 23 disks can be installed on an embedded system. Arcom Control Systems Ltd. Tel., 01223 41057.

**COMPUTER**

Industrial PC. The MCC-2000 series of industrial PCs from IBS is based on a standard ISA bus passive backplane, taking front-loading plug-in cards with fixing screws and handles. The plug-in modules give PLC-style front access, having plug-in terminal blocks for I/O. It comes with a 486 CPU module that has all standard PC I/O, and can be used with flat-panel interfaces. Standard PC cards for the PC bus up to 23 length can be accommodated, Integrated Measurement Systems Ltd. Tel., 01703 771143; fax, 01703 704301.

486 cpu card. Iosis announces a companion to its 455 half-size single-board computer – the – which takes a selection of processors, including the Pentium Onboard, and provides flat-panel and syga drive on board. An internal PCI bus is used for the IDE and VGA and a PCI/104 expansion bus and ISA for peripherals. Memory of 2700 simms can be 1-64Mbyte of dram. Iosis. Tel., 0117 9734035; fax, 0117 9237295.

**Data acquisition**

I/O boards. Amplion Liveline has five new 200 Series data acquisition boards for Windows DLI, and dos, all being mountable on DIN rails and supplied with screened cables. PC214E is a low-cost, 48-line I/O board with three independent 16-bit counter timers, PC218E/2E5E have between 6 and 18 programmable counter timers with digital I/O, and the PC227E is a tri-compatible digital I/O type giving 72 lines. Example libraries and drivers are Delphi, Borland and Visual C++, and are given for Windows DLL, in Borland and Microsoft C++ for dos. Amplion Liveline Ltd. Tel., 0800 525 335 (free); fax, 01273 570215.

**Data communications**

Data transceiver. Maxim’s MAX3221 1-drive/1-receiver RS-232 transceiver uses the company’s AutoShutdown technique to reduce supply current to 1µA when it does not sense a valid signal level at the input, turning on when a valid signal is present at any input. A dual charge pump power supply and low-dropout transmitter deliver fys RS-232 performance with supplies of 3-5.5V at a 120Kbps data rate. Four 0.1µF capacitors are needed externally. Maxim Integrated Products UK Ltd. Tel., 01734 303388; fax, 01734 305511.

**Development and evaluation**

PCI fpga design kit. Actel announces the ACT3, which is a set of software and hardware for the design of zero-wait-state PCI interfaces using field-programmable arrays. There is a full suite of design tools, including Actel Designer Series 3.0, a VHDL or Verilog test bench for verification, the first zero-wait-state VHDL or Verilog-HDL Core PCI master, slave and bridge models and the the new ACT3 family of fpgas. Actel Europe Ltd. Tel., 01256 292909; fax, 01256 55420.

**Data logging**

Em field logger. In view of a possible link between exposure to electromagnetic fields and certain illnesses, it is necessary to gather data on the strength of fields. Delta T Devices has produced the ELF Logger, which measures both electric and magnetic fields, having large diameter magnetic field sensing coils. Frequency response is 45-300Hz and sensitivity 0.1-1000nt for magnetic fields and 2.5-5000V/m for electric fields. Logging rate is programmable from 1/s to 1/day and the memory handles 16000 readings, expandable. Delta-T Devices Ltd. Tel., 01638 742922; fax, 01638 743155.

**Mass storage systems**

Hard disks for mobiles. From Integral Peripherals come the Viper 8510PA 510Mb, 1.8in PCMCIA type III drive and the Platinum1200 1.3Gb, 2.5in type, both intended for use in portable computers. Both drives use the company’s dynamic head-loading technique, which holds the heads away from the disk surface when not in operation to prevent ‘head slap’. Both also use the MicroGlide technique, in which the heads fly very close to a non-textured, polished disk. Integral Peripherals. Tel., 0113 303 449 8009; fax, 0113 303 449 8089.

**Programming hardware**

Updates from Data I/O. Data I/O has introduced new versions of its programmers, which now handle new devices from Altera, AMD, Latisio, Motorola, NEC, QuickLogic, WS1 and Xilinx. The 2900 v.3.8 takes devices with up to 44 pins and the 3900 v.2.8 88-pin devices, taking 250-pin types with interface adaptors. UniSite v5.0, the top-end instrument, will handle virtually every available device, regardless of technology or package. Data I/O Ltd. Tel., 01734 440011; fax, 01734 448700.
**IEEE Electromagnetic Compatibility Society Meeting 1996**

**Session 6: Magnetic Field Testing and Measurements**

*Presentation Title: Performance Assessment of Magnetic Field Sensors*

*Presenter: Dr. Jane Smith*

*Abstract:*

The paper presents a comprehensive evaluation of various magnetic field sensors used in electromagnetic compatibility testing. The analysis includes a comparison of sensor characteristics such as sensitivity, frequency response, and accuracy. The study is based on experimental data collected from multiple sources and considers the operational environment and application requirements.

*Keywords: Magnetic field sensors, accuracy, frequency response, application.

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**Session 7: EMI Analysis and Mitigation Strategies**

*Presentation Title: Advanced Techniques for Reducing EMI Interference*

*Presenter: Dr. John Doe*

*Abstract:*

This presentation introduces advanced methodologies for minimizing electromagnetic interference (EMI) in complex systems. The focus is on practical strategies for system design and implementation that can significantly reduce EMI-related issues. The strategies address both passive and active mitigation techniques.

*Keywords: EMI, interference, system design, mitigation.

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**Session 8: Testing and Evaluation of EMI-Resistant Components**

*Presentation Title: Evaluation of EMI-Resistant Materials and Components*

*Presenter: Dr. Mary Johnson*

*Abstract:*

The paper reviews the testing and evaluation techniques used to assess the performance of EMI-resistant materials and components. It discusses the importance of material selection and design in achieving effective EMI protection and presents case studies from various industries.

*Keywords: EMI-resistant materials, components, testing, evaluation.

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**Session 9: Standards and Regulations for EMI Compliance**

*Presentation Title: Compliance with International EMI Standards*

*Presenter: Dr. Robert Brown*

*Abstract:*

This presentation examines the latest international standards for electromagnetic compatibility (EMC) and their implications for compliance testing. It covers regulatory requirements and provides insights into practical implementation strategies for various sectors.

*Keywords: EMI standards, regulations, compliance, implementation.

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**Session 10: Future Trends in EMI Research**

*Presentation Title: Emerging Technologies in EMI Mitigation*

*Presenter: Dr. Richard Green*

*Abstract:*

The paper discusses the emerging technologies and trends that are shaping the future of electromagnetic compatibility research. It highlights the potential impact of these technologies on the design and implementation of EMI solutions.

*Keywords: Emerging technologies, EMI mitigation, future trends.

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**Session 11: Interactive Q&A Session**

*Presenter: All presenters*

*Abstract:*

This session provides an opportunity for audience members to ask questions and engage in discussions with the presenters. It is designed to facilitate knowledge exchange and address specific concerns related to EMI research and application.

*Keywords: Interactive session, Q&A, audience engagement.

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Encoding the IF signal in a heterodyne receiver can improve discrimination, reduce spurious carrier component and remove flicker noise from the demodulated signal. Andy Thurston* describes an encoding method using sigma-delta conversion.

*Dr AM Thurston is with GEC Marconi Research.

The availability of powerful digital processing engines at acceptable prices has allowed the inclusion of various new features in modern radios. Many communicate with digital modulation schemes, such as GSM or TETRA. These radios have many features uniquely implemented by dsp or microprocessor techniques: channel coding and decoding; dealing with time-division, multiple-access, tdma, frame structures; processing of protocol for call set up and maintenance; and correlation techniques to detect training sequences.

Additionally, many features formerly implemented by analogue circuitry can be handled in the digital domain, such as demodulation, channel filtering, both carrier and timing recovery, and finally equalisation.

Clearly an analogue to digital converter is required, but at which point in the architecture? One option is to encode the baseband I and Q signals separately using a suitable pair of a-to-d converters and appropriate IF and RF circuitry, Fig. 1.

Converting at a low IF

Alternatively you may choose to perform the a-to-d conversion at a low IF, Fig. 2, and thus gain several benefits. Firstly, the final mix to baseband I and Q signals may be performed digitally, which is particularly simple if the sample rate is four times the intermediate frequency since the orthogonal local oscillators required for this mix become the repeated digital sequences (1, 0, -1, 0) and (0, 1, 0, -1), or cos Πn/2 and sin Πn/2. Generation of these local oscillators is extremely simple, and the mixing process can be performed with addition and subtraction only.

Secondly, since the orthogonal baseband paths are now of a digital rather than analogue nature, perfect matching of the phase and amplitude characteristics of these paths becomes a matter of course. In contrast, in an analogue implementation of these paths, to achieve, say, a 40dB discrimination between upper and lower sidebands, it is necessary to match the gains to better than 0.1dB and to match the phase to better than 0.01rad, or 0.57°.

Finally, the issue of dc offsets arising from the mix to baseband or from the baseband a-to-d conversion process does not arise, and similarly the problem of local oscillator breakthrough in the final mixing stages does not occur. Both the dc offsets and local oscillator breakthrough on mixing to baseband have the effect of masking any signals at the carrier frequency. Bandpass conversion effectively prevents this problem.

The final IF

The choice of the final IF at which the a-to-d conversion will take place must be considered. If the chosen IF is too low then the filtering requirements at the preceding IF are made more demanding, whereas if the IF is too high then the cost of the a-to-d converter may become prohibitive. This is especially true if a standard multi-bit converter is used.

However, consideration of the nature of the IF signal reveals that a conventional multi-bit converter is perhaps not the best means of performing this conversion. After all, the actual signal bandwidth will typically be fairly narrow in comparison to the IF itself, and the full Nyquist bandwidth of the converter is not really required. It is seen that the ratio of the sampling frequency to the channel bandwidth is very high, and that a great deal of oversampling has been employed.

This fact may be exploited by using a bandpass sigma delta a-to-d converter, which is an
adaptation of the more familiar baseband sigma delta d-to-a converters found in bit-stream compact-disc players (see panel).

Oversampling for performance
The basic principle of a sigma delta converter is the closed loop suppression of quantising noise. A simple model of such a converter is shown in Fig. 3. Bandpass filter A is designed to have high gain at the IF. Closed loop analysis then shows that noise added by the quantiser is heavily suppressed at the output, typically by a magnitude similar to the open loop gain. The spectrum of a typical output of such a converter is shown in Fig. 4. Quantising noise is heavily attenuated around the IF, giving far improved resolution of the encoded signal. Provided that a multi-bit quantiser is employed then the system is easily analysed, modelling the quantiser as a linear gain with the addition of quantising noise.

The situation is far more difficult when a single bit quantiser, or comparator, is used, since the open-loop gain is not defined in any meaningful sense. The open-loop responses of such converters are highly non-linear and conventional control theory is of limited or no use. The basic principle of suppression of the noise at those frequencies of high open-loop gain still applies, but keeping the closed loop stable becomes more of a challenge.

The design of stable single bit converters may be approached in a number of ways, such as search algorithms which aim to minimise the measured inband noise, or by use of empirical rules regarding the filter design. The filter A would typically consist of a chain of high-Q resonators, the outputs of which are summed in the required ratio at the input to the quantiser. Neglecting the contribution of the earlier filter stages to this summed total would simply result in instability. Also, to more evenly suppress the quantising noise across the passband, the individual resonators would typically be tuned to separate frequencies across the passband rather than together at the centre, giving an equi-ripple response to the passband. Finally, note that since the quantiser also samples the output of the loop filter at the main sampling frequency, it is the sampled pulse response of the filter. This is important, and the response of the filter between sampling instants has no effect on the noise shaping or stability.

Second-order conversion
A second-order bandpass sigma-delta converter is shown in Fig. 5. This converter is implemented using continuous time LC resonators, two current-steering d-to-a converters, a sampled comparator and some digital delay. The 2xcos(τπ/2) component is synthesised in the first resonator as the simple ringing of

Sigma-delta conversion
Sigma delta conversion is intrinsically a narrow-band technique employing a high degree of oversampling with a closed loop quantisation-noise shaping circuit to give enhanced performance over a small fraction of the Nyquist bandwidth. The technique is generally used with a single bit quantiser, since the resulting circuit is very simple and may be made extremely linear. The linearity stems from this use of a single bit d-to-a converter.

In a multi-bit d-to-a converter, each level must be replicated accurately with respect to the other levels, and also consistently, without variation. In a single-bit d-to-a converter, each level must be replicated consistently, but the absolute level of the signal has no effect on the linearity, just on the overall gain, since two repeatable points, however positioned, always lie in a straight line.

Although originally developed as a baseband conversion technique, several institutions have developed bandpass variants over the last few years which are now finding application in various radio products. The design of such bandpass a-to-d converters is considered in this article.
the filter in response to the pulse from the d-to-a converter. The \((n/2)\times\cos(\pi n/2)\) component is synthesised in the second resonator as the growing response to its sinusoidal input, the output of the first resonator. The two components are then added, the first as the voltage across \(R_1\), and the second component as the voltage across the second resonator.

The composite signal is then reproduced across \(R_2\), where the corrective pulse from a second d-to-a converter is added to give the composite pulse response shown in Fig. 6. The correction d-to-a converter is required to supplement the main response for the sample at \(n=0\), since the main converter is still active at this stage and has only imparted one half-unit of charge into the filter, leaving it deficient in amplitude and requiring correction.

The overall result is a continuous-time filter response with the required sampled pulse response. The output spectrum of this converter is shown in Fig. 4.

Delay around the loop is set to be two sample periods, measured from the initial sampling of a pulse, through the digital delays in the feedback path, to the centre of the d-to-a converter pulse and consequently to the \(n=0\) position in the synthesised filters pulse response. This delay corresponds to exactly one half of a wavelength at the IF, since the IF is one quarter of the sampling frequency.

Half-wavelength phase inversion around the loop is used to constitute the negative feedback, rather than the use of the inversion usually found at the input summing node of a baseband negative feedback loop. Correct maintenance of this loop-delay is required to ensure best performance.

An incorrect delay would give the converter an unwanted natural resonance, degrading the inband performance and amplifying the quantising noise at the resonant frequency. If the loop delay is too far out, instability can result.

**Frequency and bandwidth**

The performance of such second-order a-to-d converters is a function of the sampling frequency and of the bandwidth of the converter. Simulation results show that the inband noise power density at overload is approximately given by

\[
NPD = 30 - 50 \log(F_s) + 40 \log(BW) \text{ dBO/Hz},
\]

where \(F_s\) and \(BW\) are the sampling frequency and bandwidth in megahertz, and the overload point of the converter is 0dBO. Performance at lower signal levels is some 2dB better.

This equation is slightly subjective in that the overload point is not well defined, as it is for a conventional multi-bit a-to-d converter, and the actual choice of the overload point may be made differently for different applications. Also, the equation assumes that the Q-factors of the filters are extremely high, and in practice some loss of performance will always result from the use of damped resonators.

One method to enhance the performance of a given sampling frequency and bandwidth is to raise the order of the converter. However, this cannot be achieved without providing additional protection against overload, since converters of order three or above cannot be guaranteed to automatically recover from an overload condition. Enhanced suppression of noise derived from the extra filter stages is obtained at the expense of lowering the overload point of the converter, until eventually the converter is unconditionally unstable.

One method of providing the overload protection is to include overload detect and reset circuitry, and this method is often used in audio converters. The audio signal is arranged to naturally limit before reaching the overloading amplitude. If the a-to-d were allowed to overload then the result would be an audible click as the converter is reset.

A more elegant means of protection is shown fitted to a third order bandpass converter in Fig. 7. In this arrangement two discrete signal paths are provided through from the main d-to-a converter to the input of the comparator. One unlimited path for first and second-order filtered components which correspond exactly to the total filter path of the previous second order converter, and one limited third order path where the signal is passed through a limiting amplifier before being recombined with the first and second order components.

Overload protection is given as follows. When signal levels in the filters are such that the limiter is not activated, the converter operates as a purely third order mode, giving the additional noise suppression of the third resonator. When the converter overloads, the signal levels in the filter rise, and the limiter is activated. The contribution of the third resonator is now reduced in comparison to the unlimited first and second order components, and the converter reverts towards an unconditionally stable second order mode.

Behaviour of the second-order unlimited modulation process is to naturally suppress the signal which has built up in the third resonator, until the converter is no longer in an overloaded condition. Performance of the converter during this process lies somewhere between that of a pure second and pure third order converter, and so initially would seem to compromise the performance that can be achieved with an unlimited third order configuration. This is not in fact the case.

**Exceptional overload performance**

The performance obtainable exceeds that of an overload detect/reset converter because of the position of the limiter within the architecture. Output of the limiter is applied directly to the input of the comparator, and hence any noise generated by the limiting process is shaped by exactly the same sigma delta process that shapes the quantising noise. Consequently, even if the limiter is frequently activated, the initial rise in the inband noise level is trivial. For this reason the level of the third order component in the filtered signal can be raised to such an extent that the basic third order sigma delta process would be unstable, and the limiter is relied upon completely to retain stable behaviour.

The additional noise suppression achievable

---

**Fig. 6. Synthesised pulse response in second order sigma delta. Desired impulse response is \([2, -3, 0, 4, 0, -5, 0, \ldots]\). This is successfully synthesised in the resonators of the converter shown in Fig. 5. Discontinuity seen around \(t=0\) is caused by the correction d-to-a converter switching on and off, necessary to boost the first sample to its required value of 2.0.**

**Fig. 5. Second order bandpass sigma delta a-to-d converter. Resonant characteristic of the pulse response is synthesised in the tuned circuits and across \(R_1\) in response to current pulses from the d-to-a converter. The correction d-to-a converter is necessary to supplement the pulse response at the first subsequent sampling instant, as shown in Fig. 6.**
by raising the third order filtered component this high outweighs the additional noise generated by the limiting action. Hence, a net improvement in the performance is obtained.

Simulations have shown that an extra 3dB of noise suppression can be achieved using this technique compared to overload detect/reset circuits, though the latter retain their performance for longer. It does, however, yield performances which may be difficult to achieve in an integrated circuit, and at intermediate frequencies certainly as high as 70MHz and possibly higher still. Integrated versions of these can be made using switched capacitor circuits, or by implementing the resonators with state variable filters based on op-amps. Switched capacitor circuits and op-amps tend to have relatively high noise figures. As a result, additional gain may be required when dealing with low level signals, requiring more power. The gain may additionally cause distortion and degrade the blocking specification of the receiver.

An example of a commercial application of a bandpass sigma-delta converter is the GEC-Marconi Communications H2550 digital hf receiver, which employs two third-order bandpass sigma delta converters in a noise cancellation configuration. The sampling frequency is 10MHz and the converters operate on an intermediate frequency of 2.5MHz. The overload point of the converters is -13dBm, and they achieve a noise figure better than 16dB and a third order intercept point better than 28dBm.

The converters are implemented with discrete components and achieve their performance with high-linearity low-noise rf circuitry. In a fully integrated form, though, there are many available baseband sigma-delta a-to-d converters.

I am not aware of any bandpass equivalents currently on the market. Several IC design companies are, however, known to be engaged in their design, and we look forward with interest to their release.

\[ NPD = -70 \log(F_s) + 60 \log(Bw) \ \text{dB/Hz} \]

at overload, and approximately 6dB lower at lower signal levels. The advantage over a second-order system thus depends on the bandwidth and sampling frequency used. Note also that the penalty for using damped resonators is greater in the third-order case, and thus the full advantage will seldom be achieved. However, the benefits of the higher order architecture are generally found to be worth the extra circuitry.

A second way to enhance performance is to perform noise cancellation, Fig. 8. The filtered error signal of a first a-to-d converter is encoded in a second converter. Subsequently, the second signal is filtered with the inverse of the error filter then added to the output of the first converter. The net effect is to cancel the error present in the output of the first converter, leaving only the error introduced by the second converter. By placing gain in the error filter it is ensured that the errors of the second converter are at a lower level than those of the first, and an improvement of typically 30dB can be obtained in a practical system. This method can be used to obtain extremely high performance from a converter, but there is a significant processing overhead required to implement and sustain the digital inverse of the error filter.

If a fixed equalising filter is used then the degree of cancellation which can be obtained is limited, whereas to achieve the figure of 30dB requires that the characteristics of the error filter are measured. This is done by applying a calibration signal to the main d-to-a converter of the first converter, and the inverse filter derived from the calibration measurements.

All of the above converters have single-bit outputs requiring decimation filters and digital mix to baseband to obtain a more useful signal. Typically, after mixing to baseband, the signals would require decimation by between 32 and 128 to reduce the sampling rate to the Nyquist rate of the baseband signal. Such filters are not considered here.

Discretely biased

The techniques described are biased towards discrete implementation, which may not be the best approach for all applications. It does, however, yield performances which may be difficult to achieve in an integrated circuit, and at intermediate frequencies certainly as...
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Dedicated ICs can correct power factor to 98% but since they output dc, they are not much use for improving existing supplies. Irving Gottlieb describes the passive alternative.

The power-factor concept involves the phase relationship between two waveforms. Normally, these are the sinusoidal-voltage wave from the ac power line and the resultant sinusoidal-current wave consumed by the load, Figs 1a, b.

Figure 1a shows the relationship with a purely resistive load. Voltage and current waves are in phase and there is no angular displacement between them.

In this ideal mode of operation, unity power-factor is realised. This derives from the definition of power-factor which is cosine θ, where θ is simply the angular displacement — i.e. phase — of the zero-crossings of the waves in the manner illustrated in Fig. 1b. Note that in Fig. 1a, θ is zero, so cosine θ is 1.

In Fig. 1b, the current wave lags the referenced voltage wave by the angle θ, which now is appreciable, at around 60°. Power-factor, that is the cosine of 60°, is 0.50. Such a low power-factor would be caused by an inductive load, such as a lightly-loaded induction motor.

Because of the motor's reactance and attendant poor power-factor, utility companies are obliged to supply about twice the line current that would be needed for unity power-factor operation. This decreases the efficiency of power transmission, increases costs and plays havoc with voltage regulation. It also causes temperature rise in transformers, and can upset the performance of circuit breakers and fuses. In addition, the customer finds that the amount of power obtainable from the wall socket is reduced relative that available at unity power-factor operation.

Although this information is old-hat to most electronics designers, confusion often sets in when one or both of the waves are non-sinusoidal. This is not surprising, for the consequences are easily misinterpreted. Consider the following situations.

Non-sinusoidal waveform problems

In Fig. 2a, it would be natural enough to see a unity power-factor by virtue of the simultaneous zero-axis crossing of the voltage and current waveforms. However, the many odd-harmonics comprising the square current-wave only partially satisfy this criterion. As a result, the ac line would see a lower than unity power-factor and extra current would have to be supplied in order to accommodate the harmonic energy.

As far as power-factor is concerned, it is as if physical reactance were present in the load. Having digested the nature of the waveform combination in Fig. 2a, confusion may still occur. Consider Fig. 2b, involving triac waveforms with a resistive load. It turns out that the power-factor at the load is always unity because the waveforms of load voltage and load current — although non-sinusoidal —
always have the same shape. As a result, they also have identical percentages and phases of harmonic content. This is not so from the ac line’s point of view. Here, the comparison is between the sinusoidal line-voltage and the non-sinusoidal line-current. When adjusted for fractional load power, the line power-factor of triacs can be painfully low. And the situation is even worse for single scr control circuits, where even-harmonics are present in the line-current.

**Power factor in psus**

In Fig. 3, waveforms 1 and 2 are characteristic of the full-wave or bridge rectifier circuits generally found in power supplies. Note the narrow and peaked haversine current-wave. This corresponds to a low power-factor, usually about 0.63, or 63%.

Considerable third-harmonic energy is injected back into the ac line and various equipments can malfunction from the electromagnetic interference. The peak current can cause excessive errors in certain measuring instruments too.

If the excessive third-harmonic current in the peaked wave can be prevented from flowing, trapezoidal waveform, no. 3 in Fig. 3 results. This is not an ideal modification, but it can yield a power-factor in the vicinity of 85% - a significant improvement over the ‘natural’ 63% power-factor of power supplies.

The advantage of the trapezoid stems largely from the reduction of the dead-angles clearly evident with the narrow haversine wave. This waveshape modification is brought about by insertion of a parallel-resonant tank-circuit in one of the leads between the ac voltage-source and the rectifier bridge, Fig. 4. Note that this tank must resonate at the third-harmonic of the line-frequency.

**Implementing the technique**

Optimisation of this technique will probably require empirical effort. Certain inherent contradictions must be balanced. For example, the tank circuit should have both, high resonant-impedance and high Q. These parameters are mutually antagonistic.

I have determined that a good start is to target the resonant impedance for 300Ω. This works well for off-line supplies in the 20-40W range. Use is made of the relationship,

\[ Z_r = \sqrt{\frac{L}{C}} \]

where \( Z_r \) is the resonant impedance in ohms, \( L \) is the inductance in henries, and \( C \) is the capacitance in farads. The trick is to simultaneously satisfy the resonance equation,

\[ f_r = \frac{1}{2\pi\sqrt{LC}} \]

where \( f_r \) is 150Hz for 50Hz lines, and 180Hz for 60Hz lines. Doing this is not difficult if you start with an inductance of approximately 300mH. Note that \( f_r \) must be exactly the third-harmonic of the power-line frequency.

With \( L \) at about 300mH, choose \( C \) arbitrarily. Thereafter, the value of \( C \) can be adjusted to obtain best results. More about adapting for different power-supply ratings later.

The inductor’s Q is governed by the \( CL \) ratio and by ohmic, eddy-current and hysteresis losses. In light of this, best performance may not be obtained from ordinary 50/60Hz silicon-steel cores with E-I laminations. Tape and powdered cores - especially toroids - tend to be better prospects.

Admittedly, this simple technique falls short of the performance attainable from active harmonic-suppression. The active alternative causes the current wave to follow the sinusoidal voltage-wave much more closely, improving power-factor to 98% or more. However, dedicated power-factor correction ICs usually deliver several hundred volts of dc. As a result, they are not applicable for improving the power factor performance of most existing power supplies.

**An alternative correction scheme**

Another comparison can be made with the scheme shown in Fig. 5 which has been suggested in other literature. In principle, the energy storage of a large, high-Q shunt connected tank circuit can present an essentially resistive load to the ac line, producing a good power-factor. In practice, such a resonant circuit must operate at line frequency and certainly must store much more than the consumed energy of the power supply. As a result, unreasonable demands are imposed on the size of the core, and on the resonating capacitor. Besides, such a scheme could be vulnerable to damage from third-harmonic energy already on the power line.

So, all things considered, the simple passive harmonic-suppression technique described here can fill the niche where substantial improvement - rather than near perfection - is acceptable. Whereas the approximate LC values depicted in Fig. 4 apply to a nominally 30W supply operating from 120V, appropriate scaling factors can be used for other arrangements. If, for example, the supply operates from 12V, a 30mH inductor would be resonated by a capacitance of around 30µF.

![Fig. 3. Power supply waveshapes. 1) Line voltage sinewave. 2) Typical line current due to full-wave rectifier. 3) Line current after attenuation of the third harmonic by resonant tank. The near trapezoidal current results in a higher power-factor than the peaked wave of 2.](image)

![Fig. 4. Basic technique for power-factor improvement of a 30W supply. Passive method makes use of a third-harmonic wavetrap. Approximate LC values are shown. Exact third-harmonic resonance is required for optimum effectiveness. This will be 150Hz for 50Hz power lines and 180Hz for 60Hz lines. For a 30W supply, the resonant impedance, \( \sqrt{L/C} \) is about 300Ω. Scaling factors may be used for supplies of other ratings.](image)

![Fig. 5. Theoretically valid, but impractical method for power-factor improvement. This technique also makes use of the frequency selective energy storage of a resonant tank. However, in this case, the LC circuit must be tuned to \( 'f' \), the frequency of the power line. Moreover, energy storage must be great enough so that the resonant Q is not appreciably depleted by the rectifier load. Because of these requirements, outsized core dimensions and capacitor values are needed for all but the smallest of power supplies.](image)
**LETTERS**

Letters to "Electronics World"
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Golden ear tarnished

Some people may observe that in the past I have been harsh in my criticisms of 'Golden Ear' persons, who claim that they can hear magical differences between different types of speaker cables and audio interconnect cables.* However, I have ammended to criticise equally sharply, those who unfairly criticise 'high-end-audio' enthusiasts.

In the *EW+WW* of December 1995, pp. 1012-1013, the Rev. Williamson states that "...exotic components - whether passive or active - are a waste of money," in his simple audio preamplifier circuit. (In this case, one simple preamp/tone control circuit was constructed with inexpensive components, and another preamp was built with the same nominal circuit, but with "exotic" (highly expensive) components.)

In 224 out of 248 listening tests, the listeners said they could not hear any difference. The writer said, therefore, that "...exotic components... are a waste of money."

Firstly, let me state that I have tin ear, and while I am accused of having bad hearing primarily by my wife, I can still hear everything I need to. Yet I cannot easily distinguish between different audio processes, by listening. I probably did too much lawn-mowing when young. But I am still capable of thinking.

It may be that for most people, it is a generally poor investment to buy audio equipment built with 'exotic components'. But it is quite unfair to argue that nobody should ever spend money on the best quality components.

Of the 24 listeners who could tell the difference between the fancy-parts preamp and the cheap-parts preamp, there might be eight people who could hear a difference, but could not tell any preference. There might be eight others who could hear a little difference, but did not think the cost worthwhile. But there might be eight others with sophisticated tastes in audio equipment, and really good ears -


and they might consider it an extremely wise investment to buy a piece of audio equipment that sounds better. I can't say they are wrong, and Mr. Williamson is wrong to say that they are wrong, to prefer a more costly preamp.

It may be true that for most people, costly components are a waste. But for the author to assert that it is wrong for everybody or for any particular person to buy costly components, is bad thinking. And I insist that such blanket condemnations be rejected by any thinking person - and especially by readers of your pretty-good magazine. First point made.

Second point: the author implies that he made two preamps, one with expensive parts and one with fancy parts, and a few people could discriminate between the two. But the author does not indicate what differences were made in the parts list. He implies that some active components were different, and some passive components were different. He implies that blind or double-blind tests were made, and some listeners could hear a difference.

But the original circuit used inexpensive TL071 and TL072 operational amplifiers. I have no doubt that a trained ear might be able to hear some small but significant difference between a good, low-noise audio op-amp and the inexpensive ones as specified in the original circuit.

I am willing to believe that some components may make a real difference in the audio quality. I believe, for example, that the active components - the op-amps - may make a discernable difference. Maybe not by me - and maybe not by 90% of the people - but by some people. For example, two different op-amps might have the same amount of harmonic distortion, but completely different transient intermodulation distortion performance. It's not reasonable to expect different op-amps to sound the same. Point two made.

Point 3: I am also prepared to argue that when other expensive passive components were substituted, probably nobody could tell the difference. Let's say that if you change one capacitor from Mylar to polypropylene or to Teflon, nobody can tell the difference.

Let's say there is a possibility that if all the capacitors were changed from Mylar to Teflon, nobody could tell the difference, in double-blind testing. If the resistors were all changed from cheap carbon resistors to fancy metal films, in most circuits, nobody can hear a whit's worth of difference. But if the op-amps are changed, perhaps 9% of listeners could tell a difference. So, maybe some exotic components are a waste of money. And maybe some are not.

If the op-amps are installed in sockets, it might get really interesting to take the cheaply-built preamp, and substitute high-quality op-amps. And it might be interesting to take the 'exotic' preamp and swap the TL072s. Now which one sounds good, which one sounds like the straight-through wire, and which one sounds odd?

So I argue that it was quite unfair for the author to condemn the use of 'exotic components' on a blanket basis. It is just as unfair to condemn ultra-hi-fi audio equipment, with no scientific or subjective evidence, and without any critical thinking, as it is for the 'Golden Ear types' to praise some components with no scientific or subjective evidence and without any critical thinking.

Robert Pease
California

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Funny you should mention it

Your Update page headline 'EMC all over again', p. 444, June '96, brought a wry smile to my face. Yes, I took notice of the DTI during the EMC Awareness Campaign, heeding the warnings like 'If your product is not compliant, you will be out of business'; yes, I approached my MD and explained what we would have to do; yes, my MD allocated me an extra budget with which to purchase test equipment. And, after all that I worked my butt off in order to make 501 products compliant in four years.

Now, can someone tell me, when will all of these non-compliant companies be out of business? Many of my colleagues have laughed themselves silly at me, telling me that nothing will ever come of it all.

Who is laughing now, six months on? Certainly not me. Yes, I know that the Directive is complaint driven. I am sick of telling TSO's about non-compliant products. CE marking for the LVD becomes mandatory on 1 January 1997. I see a lot of products coming into our service department - even products from big name manufacturers - which would not stand a chance of meeting the requirements of the LVD. Indeed, I was present at the main trade show for my particular trade this year at Frankfurt. I saw many thousands of none CE marked products there, and I saw a lot which should not have been marked, but were. Not once in the seven days I was present did I see a Bundesamt für Post und Telekommunikation (BAPT) official (the German equivalent of a Trading Standards Officer), or a Customs official, checking product compliance.

One product which sticks in my mind - a high powered stereo audio amplifier made by a well respected manufacturer - was branded with the CE mark letting the world know it was compliant. On the rear panel, next to the CE mark were a pair of gold-plated, uninsulated 4mm binding posts - a direct contravention of Clause 15.1.2: EN60065: 1994. There is no way that such a unit could be compliant with the existing electrical safety regulations, or the LVD.

So, will I worry about the LVD for 1997? No. All of my products have been compliant for years. Will my competitors worry about it? Very unlikely.

K. C. Aston
Leeds

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Agree to disagree about valve sound

Referring to Letters in the April issue, I hope that Nick Wheeler and I can agree to disagree about 'valve sound'. But if he happens to have a vintage Bentley, then perhaps I should offer to swap him a Ford Mondeo.

McFadden's intuitive explanation of the operation of the concertina phase splitter is elegant, but I was unable to immediately recognise it with the Thévenin model. Rather than theorising, the conflict seemed to be most easily resolved by...
experiment. If the equation he quotes is correct, then \( r_{out} \) should tend towards \( 1/\beta \), resulting in extremely wide bandwidth into a representative load of 100pF. Therefore deliberately selected a valve with low \( \beta \) (Mullard ECC83) in order to make the measurement more manageable.

The first test was to determine whether the bandwidth at cathode was equal to that at the anode, so a 500kHz square wave was applied, and both outputs monitored. As you see from the diagram, the outputs are virtually identical, proving that the output resistances are equal. Channel 2 was inverted to make comparison easier. The slight noise on the anode (lower) trace is a consequence of power supply hum which was incompletely removed by the averaging function of the oscilloscope.

The next test was to determine the output resistance by measuring the bandwidth, so a 1Vpk-1kHz sine wave was set at the cathode output, whose frequency was increased until the level fell to 70mV. The -3dB cut-off was at 2.32MHz as measured by an external frequency counter. The -3dB -t
t

resistances are changed. Letting \( \beta \) (\( R_\beta = R_{\beta} \)), and substituting in the triode gain equation gives a feedback factor of \( r_{\beta} = \frac{r_\beta}{R_\beta + r_\beta} \).

Output resistance, \( r_{OUT} \), of a common cathode triode amplifier with no feedback is \( r_\beta \).

If the feedback is actually working to reduce the output resistance at the anode, then this value must be divided by the feedback factor:

\[
\begin{align*}
\text{Feedback gain} & = \frac{1}{r_{\beta} + r_\beta} \\
\text{Output resistance} & = \frac{r_\beta}{r_{\beta} + r_\beta}
\end{align*}
\]

This is almost identical to the Langford-Smith equation, but for a factor of \( \beta +1 \) rather than \( \beta +2 \).

A can of worms has been opened here, since we now have a phase splitter whose output resistance changes with loading. If the concertina is loaded with a Class B stage, then at any instant only one valve will be switched on, and since the input capacitance is made up largely of Miller capacitance, only one output of the concertina will be loaded capacitively at a time, the output resistances revert to those I gave previously (Letters, Mar '96), and the build-out resistor is required. Although the 'Bevois Valley amplifier' operates substantially in Class A, and should therefore be degraded by a build-out resistor, once output transformer losses at high frequencies are taken into account, feedback will drive the output stage into Class B at high frequencies, thus explaining the improvement in square wave performance seen in the prototype. If a concertina is driven into a low level driver stage, then a build-out resistor is undesirable, but if driven into a higher level stage (which may enter Class B, and whose gain and Miller capacitance change with level), then a build-out resistor may be required. The value of this resistor might need to be tailored for a specific output transformer, but the theoretical value offers an upper limit.

As to the question of balancing signal currents, the mistake is entirely mine, and although I will not be modifying the amplifier shown in the illustrations (because it sounds so good), a fourth version currently under construction will be modified. To round off, the somewhat extensive component changes would be as shown in the table below.

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>CFBK</td>
<td>100pF</td>
</tr>
<tr>
<td>RORE</td>
<td>68kΩ</td>
</tr>
<tr>
<td>CFBK</td>
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<td>CFBK</td>
<td>100pF</td>
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<tr>
<td>RORE</td>
<td>68kΩ</td>
</tr>
</tbody>
</table>

Morgan Jones
Southampton
Testing, testing 123

There has been a very large response to my article on a high-performance microphone preamplifier and, as many of us know by now, the SSM2016 has made an untimely exit from the market. I must apologise for not ensuring supplies of the device; I obtained some without difficulty quite recently and didn't imagine there would be this level of demand, or lack of supply. So, the only sensible option seems to be redesigning for the 2017. The noise and distortion performance is not very much worse than the 2016, I was not making use of the 2016's output capabilities, and the other aspects of the circuit — input circuitry, protection, metering, filtering and power supply — remain valid.

I have therefore redesigned the preamplifier board for the 2017, attending to the vital details of keeping input and outputs separated and ensuring separation of signal and power grounds, and it seems to work very well.

I have also tried out the INA103 which was most helpfully pointed out as an available alternative. It has extremely low distortion — far lower than any available microphone — but it is slightly noisier. Unfortunately, like the 2017 and unlike the 2016, it does not give direct access to the input transistors. As a result, they cannot be exactly trimmed for dc balance. This means that with both the 2013 and the 2017 you either have to use an electrolytic in series with the gain-setting resistors, upset the input impedances with an external bias circuit or put up with a slight click as the gain is changed.

I have opted to place a very large 1000µF electrolytic bypassed with polyester in the gain-setting section and have to confess I can't reliably hear a difference between this and direct coupling. It would be interesting to know if anyone else can. Offset voltage across the capacitor is less than a millivolt so we would expect 'shelf life' from it.

I am having the boards made, single-sided, drilled and toned. If you would like a set, please send a cheque or money order for £28 (all inclusive) payable to MicroPower Measurements at 4, Elwick Terrace, Hutton Rudby, North Yorkshire TS15 0DH.

For interest I append the diagram of the preamplifier section incorporating the 2017, all the other parts of the project remain unchanged.

Simon Bateson
Hutton Rudby
North Yorkshire

Better windows

In the April issue, Phil Denniss makes the mistake of confusing the multitasking WIMP based paradigm of computing, with Microsoft's rather poor implementation of it in the form of Windows. The British company Acorn has produced a far

This circuit replaces the one on p.420 of the May issue. Performance is a tad lower, but at least the chip is available.
more successful implementation of this method of interacting with computers. As a user the test is how much work I can get through in a session, not how fast the clock speed of my machine is, and compatibility means being able to transfer my data between different machines using different processors, not dancing to the tune of Microsoft's cash register.

Mistaken identity

Mr Hopwood (Letters, May '96) has jumped to the conclusion that I was accusing him of belief in anti-gravity and perpetual motion (free energy). As far as I know, Mr Hopwood's only connection with the latter is his suggestion that 'cold fusion' might need to be primed by sunlight (EW + WW Letters, Dec. '93). I was, in fact, thinking of Dr Aspden who, readers will recall, believes in this nonsense and has also proposed a 'cyclotron' theory in order to 'explain' the so-called link between electromagnetic fields and disease.

Dr David Fisher
Cardiff

Cancer and power

Mr Brown's letter in the May issue of EW raises a fair question. Dealing with Denver first, the surveys I quoted were based on a relationship between power line routing and so-called radiation cancers. Over the past 90 years it has been established that these are different to those induced by uranium mining. The markers for uranium ingestion are lung cancer and other disorders caused by internal alpha particle irradiation from dust and radon, and are separately classified in the ICD statistics I quoted. On the matter of statistical quantities, I entirely agree with Mr Brown. The numbers are so small, that like CJD, the only certainty will be a buildup of evidence from large scale surveys. What is interesting is that areas of the country with high background radiation levels like those on 'young' granite do not have high levels of radiation induced cancers of the types noted in power line surveys. They do have appreciable radon induced cancer levels - hence the boil, so we must wait a few years to prove the effect with new and better equipment.

Anthony Hopwood
Upton-on-Severn
Worcestershire

Play nice

Dr Fisher clearly does not like Mr. Hopwood. From his reply Hopwood shows that he couldn't care less about that, so good for him. I might not believe his views but I go along with Voltaire.

In my 71 years on this earth I have long ago learned that if one side in a debate or discussion resorts to personal abuse then it is clear that that person cannot make a case for his views especially when he resorts to lies as well. We see it every day in Parliament.

Who does Dr Fisher think he is to set himself up as superior to Prof. Eric Laithwaite (anti-gravity)? Why doesn't he become an MP? He appears more suited to that occupation.

There is too much 'thought policing' in the scientific community already - not least in Nature - and to some extent, I am sorry to say, in Electronics World. Finally, no physicist nor engineer should use the words 'never' or 'impossible' in a predictive sense as it is an open invitation to Murphy's Laws to slap egg on your face in the future.

G.E.Miller
Hastings
East Sussex

On autopilot

Almost exactly three years ago I was one of about twenty-five subjects in an investigation to examine the effects of magnetic fields on human navigation. To be more precise it was an experiment to test what is known to biologists as 'compass' ability.

The researcher has been publishing peer reviewed papers on the subject since 1987 and has written at least one book on it. His work is not universally accepted, but he can be certain that his fellow scientists will not descend upon his laboratory and insist that he repeat the effects he claims forthwith.

That appears to be exactly what happened to Anthony Hopwood, and Dr Fisher wants to use the outcome to discredit any further effects that Mr Hopwood claims to have detected.

Les May
Rochdale, Lancashire

Too noisy

I'm surprised that recent correspondence on the Sallen & Key filter topology has not mentioned another drawback namely its poor noise performance. Although the topology is often chosen for its design simplicity and operating stability it possesses a high gain to internally generated noise.

In the low-pass form pictured, where typically $R_1$$R_2$ and $C_1$$C_2$ form an attenuator which near the cut-off frequency passes a high proportion of the output signal back to the positive input. This results in a noise output much higher than the amplifier's input-referred internally generated noise $V_{in}$, and may surprise those who take the maximum circuit gain to be unity. It is of course only the gain to ground-referenced signals that is unity - the amplifier still has a very high gain to differential inputs!

The problem is worst in the final stage of high-order filters where the ratio of $C_1$ to $C_2$ is generally highest, but can be significant also in second or third order designs with general-purpose op-amps, as I found when building a low-noise photodiode amplifier. The problem can be reduced by replacing the op-amp with a single transistor (in common collector mode) having a much lower noise figure, or by using a different filter topology.

Anthony New
Bristol

Down the garden path

The discussion in EW about transistor linearity is - to my mind - somewhat misleading. Douglas Self has repeatedly claimed that the bjt is at least ten times more linear than a mosfet.

But the fact is that the bipolar transistor has the highest non-linearity possible in a silicon device. The current change is 2.72 ($\pm$) times for any 25mV base voltage change, which is so high that it can hardly be seen in a linear scale. The bjt is anything but linear.

What Self is referring to is the linear characteristic of the complementary stage. This is an effect of its large negative feedback. This invisible feedback is nearly 60dB in Class A, an effect of the high gain. Had it been open loop, the gain would be 1000s, which is 800. Now the gain is one. The rest is feedback. Small wonder that the stage is linear. But that is not the merit of the bjt. The linearity is borrowed purely.

"So what?", you may remark. The linearity is there, at least in Class A. However, one has to be clear, because there is a tremendous difference between bjt and mosfet stages.

Bengt Olsson
Saltby-Boo
Sweden

Historical units

I was interested to read the 'Gems and Oddities' described by Ian Hickman in his article 'Circuit Reflections' in the May 1996 issue. It refers to the Jar as the unit of capacitance once used by the Navy and described in the 1925 edition of the Admiralty Handbook. Apparently 1 Jar was equal to 1.1 nF.

I came across a similar oddity when working in the laboratories of Standard Telephones & Cables Ltd in the early 1930s. The company made repeater valves for the Post Office Telephone Network and the network had its own unit of attenuation, it was the 'mse'. All the valve test sets, decade attenuator boxes and similar equipment that I used were calibrated in units of mse.

The mse, I discovered, was the attenuation produced by one mile of standard cable. Here was a very practical unit, since an amplifier with a gain of 20msc would send the conversation a further 20 miles down the line.

If I remember correctly, 1 msc was equal to 0.088 dB.

OH Davie
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Hands-on Internet

Cyril Bateman discusses the benefits and limitations of new applications on the World Wide Web.

The World Wide Web has recently become more commercialised with the introduction of many glossy corporate Web pages and several new professional search tools. One of these well worth trying is located at Netscape.com.

While the last article featured Java – a staple source of sustenance to programmers and designers alike – this month I suggest a bite of the ‘chilli’ pepper, being the logo adopted by Netscape for their Infoseek information system.

If you use Windows ’95, you can download iSeek for use on your local machine. If not, you can log on to 'http://cgi.netscape.com’ to run their searches. This site provides easy access using the Infoseek program or one of the twenty alternatives offered. Infoseek was the search tool used exclusively in preparing this article, Fig. 1.

A commercial aspect of Infoseek is the ‘ISN Internet Shopping Network’, which by UK standards offers some keen prices. It is accessed via the ‘Computers and Internet’ sub-topic at this site, Fig. 2.

Fig. 2. The Infoseek Internet Shopping Network home page. This has been accessed via the Infoseek Computers and Internet ‘other’ topic sub-menu. This on-line shopping mall has much to offer, but check prices, currency and taxes first.

Fig. 1. The Infoseek Web Search home page from http://cgi.netscape.com. Windows 95 users can download iSeek Beta client, for local use. Others can use Infoseek or choose from twenty alternative search tools on this page. You get the most relevant matches, related topics and a brief resume with each response.

Fundamental elements of Internet, namely the ‘Archie’ and ‘Gopher’ search engines, were covered in the April ’96 issue. However, as outlined in June ’95, other cataloguing systems have slowly developed.

The original site at ‘Cern’ and the well established ‘Yahoo’, ‘Lycos’ and ‘Web Crawler’ searchers, have now been supplemented. The ‘Alta Vista’ system claims to have indexed 16 million Web pages, the ‘Electric Vista’ database of 1000 full text newspapers, magazines and journals, and the infoseek guide. Infoseek, recently rated number one search service by PC Computing, provides stock quotations and company profiles in addition to access of the Internet databases, Fig. 1.

For surfers outside North America, one aspect of this commercialisation, while resulting in very pretty presentations, is less user friendly. Many recent Web pages have begun using graphics intensive pages resulting in very slow updates and costly downloads by modem. Previously this could be overcome by turning off graphics, relying on the text page menu selections only. But some new pages incorporate menu selections within the graphic to encourage one to view their effort.

The more established page designers however still manage to produce equally attractive pages and, by using smaller GIFs or JPEGs, ensure much faster access, Fig. 3.

In addition to the Web browser applications discussed in the May
issue, Java as a development language is proceeding apace for all hardware and operating systems. 'JFactory' already offer a production quality interface design environment for building Java applications, but they too are guilty of using large and slow loading graphics. The sample shown is a GIF file of some 62K in size, Fig. 4.

Some have claimed that 'Java' could develop to supplant Visual Basic, Visual C++ and Delphi, as the leading Windows development package. Licences have now been taken by almost all major software houses including Microsoft. While most development is taking place in North America, IBM is developing Java support for OS/2 and AIX at their UK Hursley Park site near Winchester, Fig. 5.

A brief explanation of Java, together with IBM's view of the importance of 'Java', can be downloaded from the company in an 80 page technical overview of the Java language and the HotJava browser.

Continuing with the 'Spice' macromodel and Semiconductor houses topic, the Maxim Integrated Products home page is a model of simplicity and rapid download. This company offers Spice macro-models, regularly issued design guides and product samples on request to registered users, Fig. 6. A visit to Philips Semiconductors on two recent occasions resulted in some pretty pictures but at the slowest download rate ever. I hope that feedback to their web manager will soon correct this. Until then this site is best accessed with graphics turned off.

Unfortunately it is not possible to judge the size of a graphics file until after it has been downloaded. Some of the better web pages, for large graphics, show a box stating the file size, giving one the choice to download the graphic. But this desirable option has not been made available on the Philips site, Fig. 7.

When accessing Texas Instruments, remember the address is TI.com. Using the address Texas reveals some interesting pages not relevant to electronic design. The TI home page at http://www.TI.com also has a pretty, but painfully slow loading, graphic. Corporations approving these pages might be less happy if forced to view using a modem. Perhaps this should be de rigueur, until we can all enjoy better bandwidth access, Fig. 8.

On the topic of bandwidth I am still using my original 14.4kbaud Zoom modem running under
COMMUNICATIONS

Fig. 8. The Texas Instruments Home Page at http://www.TI.com. Another site with a slow downloading image, again best visited with graphics turned off. While TI does have macromodels available, this was unclear when the page was visited. Ask for the 'Linear Info - Access Selection Disk with Macromodels'.

Fig. 7. The Philips Semiconductors home page at http://www.semiconductors.philips.com. This is a small part of this page showing the 50kbyte GIF graphic — which downloaded very slowly. Unless the Internet or your system is working very quickly, turn off graphics first. Perhaps if sufficient viewers feedback their download times, this page will be changed.

OS/2 Warp with ibm.net as provider. If starting from scratch, I would now purchase a 28.8kbaud modem. Recently ibm.net and many others have moved to the fastest currently available modems at 28.8bps. The increased speed of downloads on a good day is noticeable even when using a local 14.4k modem. Previously the best typical speed achieved was around 100Kbyte per minute. Since my providers' upgrade, on occasion transmissions have peaked around 3140bps, or around 170Kbyte a minute. Perhaps my upgrade costs to 28.8 bps are now justifiable.

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Listening for clues

Although simple, the squawk box can be invaluable – especially when looking for a signal buried in noise. Ian Hickman has taken the squawk box to its ultimate, combining a fully protected laboratory audio power amplifier with an af millivoltmeter.

Every electronics laboratory in which I have ever worked has had a general-purpose laboratory amplifier. Such a device is invaluable for "listening to" a wide variety of circuit types. You might think that the more obvious option is to see what is going on with the aid of an oscilloscope. But if there is a mixture of noise, hum and possibly other signals as well, it may be difficult or impossible to interpret the display, or even to trigger the oscilloscope so as to obtain a coherent picture.

With a mixture of signals present in a signal, an audio-frequency spectrum analyser might be more appropriate. Few laboratories possess such an instrument. Luckily though, most electronic engineers have not one but two af spectrum analysers – one on each side of their heads. All that is needed to make use of these is the universal labamp.

Labamps are usually knocked up in a hurry to help solve a particular problem. Typically, they consist of an amplifier and a small loudspeaker, housed in a die-cast box and powered by an internal dry battery. With its tinny low-fidelity reproduction, such a device is generally known as a 'squawk-box' – and very useful it can be too. In fact, its advantages are as numerous as its disadvantages.

For example, running from an internal battery means that the amplifier is immune to the hum problems which might, in a mains-powered version, be caused by earth loops. But unfortunately, just when you need the amplifier in a hurry, it usually transpires that the batteries are flat because the last user left the unit switched on. Then again, it would be useful to be able to hear whether the signals in the circuit under observation are corrupted by hum not caused by an earth loop, but with its small loudspeaker, the typical squawk-box remains silent on this topic.

A new squawk-box approach

A while ago, I resolved to replace my squawk-box – the last in a long line, mostly converted from superannuated radios – with a version having a decent frequency response and a generous output of a few watts. This indicated mains operation, but with precautions to avoid the possibility of hum due to earth loops.

For building audio amplifier systems, a calibrated input step attenuator and meter circuit are incorporated, making the unit double as an af millivoltmeter. In addition, a 600Ω unbalanced signal output is provided, permitting use as a hum-loop free, general-purpose pre-amplifier.

For good measure, access is provided to the loudspeaker's voice coil, to permit the unit to be used also as an extension speaker. Optionally, the amplifier's output can be made available to drive an external speaker.

The input stage

While the unit is provided only with an unbalanced high impedance input, the actual input stage is balanced. This permits the rejection of any hum present on the "earthy" input low line, on which
the wanted signal may be riding. The input stage therefore uses a conventional three op-amp instrumentation amplifier configuration, as shown in Fig. 1, using three quarters of a TL081 quad op-amp. Gain of the input pair to balanced or ‘push-pull’ signals is equal to \((2R_1+R_g)/R_g\), while their gain to common-mode or ‘push-pull’ signals is unity, i.e. these appear unaltered at the output of the input pair.

While the input pair provides no common-mode rejection as such, the balanced-to-unbalanced signal ratio is improved by the ratio \((2R_1+R_g)/R_g\), which could be large. Output of the input pair is applied to the input of the third amplifier, whose gain to balanced signals is the ratio \(R_3/R_2\).

However, assuming the two \(R_g\) and two \(R_s\) are exactly matched, they form a bridge circuit, so that the common mode component appearing at the inverting input of the third op-amp exactly cancels that appearing at the non-inverting input. Thus the overall common mode rejection ratio, or CMRR, is that provided by the associated \(R_3/R_2\) mentioned earlier.

In the present application, the wanted signal appears between the input terminals in unbalanced form, but the circuit still responds to the difference voltage between the two terminals. However, any hum due to an earth loop on the input low terminal, i.e. the outer of the BNC input socket, will appear also on the input high lead or centre pin of the socket. In this way, a common-mode component and will be rejected as described above.

**Measuring millivolts**

The af millivoltmeter stage is a simple one, using a full-wave rectifier circuit. It is scaled to read rms when the input is a sinewave. Obtaining a linear scale was at one time difficult, due to the forward volt drop of the necessary diodes.

Various schemes were formerly used. These ranged from individually calibrating the meter scale to allow for the diode nonlinearity, to using a high impedance, such as the collector output of a transistor to approximate a constant current source. The circuit chosen appeared in *Wireless World* many years ago. It encloses the meter and a bridge rectifier in the feedback loop of an op-amp, Fig. 2.

Assuming the open-loop gain of the op-amp remains high up to the highest frequency of interest – 20kHz in this case – it will force the voltage at the inverting terminal of the op-amp to follow that at the non-inverting input. In the process it will force a current defined by the lower resistor through the meter. This will happen regardless of the volt-drop across the diodes, which will in any case vary slightly with temperature.

As the input voltage passes through zero, the op-amp becomes momentarily open loop. Just a small voltage difference between its input terminals forces it to slew as rapidly as it is able until the other side of the diode bridge turns on, restoring closed-loop operation.

**The complete instrument**

This is shown in Figs 3 and 4. Figure 3 shows the input at a BNC socket applied via a dc blocking capacitor and a 4.7kΩ safety resistor to a range switch \(S_2\). This, in conjunction with the gain of the following stages, provides nine input ranges giving full-scale deflection, f.s.d., factors for the millivoltmeter function of \(3\text{mV f.s.d. to 30V f.s.d.}\).

The input of the op-amp connected to the wiper of \(S_2\) is protected by back-to-back diodes. These are rated at 75mA peak current, which corresponds to a peak input voltage of about 350V. But as the peak dissipation in the 4.7kΩ safety resistor under these circumstances would be over 25W, this should be regarded as only a momentary withstand voltage, or a 4.7kΩ resistor of the fusible variety could be used.

The input appears between the non-inverting inputs of the first stage of the instrumentation amplifier, which provides a gain of \(x20\). When monitoring an earth-free source, e.g. a piece of battery operated kit, \(S_1\) can be closed, providing an earth for the item under test. Where a hum loop problem is encountered, \(S_1\) should be opened, breaking the loop.

The associated 15kΩ resistor provides a ‘static drain’, to keep the input amplifier earth-referenced, even if the input socket is left open-circuit. When the unit is connected to other equipment, any ground-line float is limited by the associated zeners to just over \(\pm3\text{V}\).
The third section of the quad op-amp provides a gain of around 6x, the 2.2kΩ potentiometer permitting an adjustment for maximum common-mode rejection. This is obtained when the ratio of input resistor to feedback resistor on the inverting side equals the potentiometer ratio on the non-inverting side. Output of the instrumentation amplifier stage is made available at a BNC output socket, labelled 'Monitor', at an impedance of 600Ω unbalanced.

The outer of this socket connects to the circuit's OV line, and hence is referenced to the input-socket outer, or the instrument's mains supply earth, according to whether S1 is open or closed. The low frequency -3dB point of the input amplifier is 3.2kHz. This is much lower than the loudspeaker and enclosure are capable of, but it was chosen to provide a wider-than-audio frequency response at the monitor output.

The third section of the quad op-amp also drives the meter stage, which uses the final section of IC1. Here again, the 100nF coupling capacitor and 1.5MΩ resistor provide a frequency response extending below the bottom of the audio range. Germanium gold bonded diodes were used in the meter circuit, for their low forward volt-drop.

Overall sensitivity is set up with the 1kΩ potentiometer. In normal operation, the 10kΩ resistor driving the bridge plays no useful part, but limits the current applied to the meter when a large input overload is applied. While inclusion of this resistor is not good for the frequency response, the instrument is nevertheless flat from 20kHz to 10kHz, and less than 1dB down at 20kHz.

Figure 4 shows the output amplifier and power supply stages. The power amplifier used is the TDA2030. For convenience, this device was mounted with its associated components on the matching ready made pcb, RS434-576: Power supplies, input stage and millivoltmeter stage, on the other hand, were all constructed on a piece of 0.1in matrix copper-strip board.

Loudspeaker volume is controlled both by the setting of the input attenuator S2 and by the 100kΩ logarithmic volume control at the input of the TDA2030. The on/off switch S4 is ganged with the volume control. The TDA2030 amplifier drives a wide-range twin cone loudspeaker type RS249-031, mounted in its matching cabinet RS249-801.

Alternatively, the loudspeaker may be switched to a three-pin socket, allowing it to be used as an extension speaker. For further versatility, the amplifier may be used to drive an external loudspeaker connected to pins A and C of the three way socket, either by itself or in parallel with the internal speaker, according to the setting of S3.

The dual-rail power supply is conventional, providing about ±17V to the TDA2030 under quiescent conditions. This is dropped to a stabilised ±10V for the preamplifier and millivoltmeter stages. A load resistor is fitted to ensure rapid discharge of the smoothing capacitors on switch-off, preventing a possible nasty surprise if the unit is opened up.

Practical considerations
With its fully enclosed cabinet, the loudspeaker creates considerable pressure inside the enclosure when reproducing low frequencies at volume. Care is therefore needed with construction to avoid rattles. Both BNC sockets, the meter and S1 were mounted on a small Formica panel covering the lower third of the grille cloth, clear of the loudspeaker cut-out. The panel was firmly screwed into place, with the rear of the components projecting back through the front panel of the cabinet, in holes just large enough to accommodate them.

Switch S1 and the volume control were mounted on an aluminium subpanel the same size as the Formica panel, but mounted behind the enclosure's front panel, which had holes just large enough to accommodate the shafts of the controls. A rebate is formed on the inside of the front panel to clear the nuts and bosses. The aluminium panel also had a hole to clear the rear of the meter, and small holes to pass the leads to/from the BNC sockets and S1.

Care is needed to avoid hum pickup — especially on the more sensitive ranges. Inside the cabinet, aluminium foil was fitted to cover half of the bottom of the case and the whole of the side where the stripboard input amplifier/af millivoltmeter was mounted. Like the aluminium subpanel, this foil was connected to the power supply OV rail.

The mains transformer was bolted firmly to the base of the cabinet, on the opposite side from the input board.

In summary
The amplifier described here has been in use in my laboratory for months now and has proved entirely reliable. With its wide frequency response, it is much more informative than a small diecast-box-housed battery-operated squawk-box when used to monitor activity in an audio circuit. I have also made a simple diode probe, permitting signal tracing in radio and intermediate-frequency circuits.
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Circle No. 122 on reply card

Circle No. 123 on reply card
An addressable digital voice recorder chip can replace a panel meter in many applications, and is particularly useful for people with impaired vision. Here, Heikki Kalliola explains how such a device is used as a speaking thermometer.

The heart of this talking thermometer is an analogue memory chip, in which a binary addressable voice can be stored. In this case the meter scale is first spoken to the chip. Temperature related voltage is produced by a sensor. An analogue-to-digital converter chip converts this voltage to binary data for addressing the voice chip memory. When a reading is triggered a loudspeaker outputs the message in current address containing the temperature information.

Circuit operation
The sensor connects to an 8-bit parallel output a-to-d converter whose seven most significant bits are used for memory addressing. Trimmers R2, R12 are required for calibration.

Pushing S1 initiates a pulse from Tr1's collector, triggering conversion by pulling up the chip select pin, /CS. The pulse starts conversion, the result of which remains at the output pins until next triggering.

Playback starts by taking pin /CE momentarily low and continues until memory overflow is reached. A logic high at the power-down pin, PD, keeps power consumption low. This pin must be pulled down when playing or recording. The pin also acts as reset switch should a memory overflow occur, in which case it must be pulled up then down again.

Playback starts from the address presented to the speech chip by momentarily pushing S1. It continues until end-of-memory, or memory overflow. The button pulls down pin /CE and creates a reset pulse to pin PD via Tr1. The same pulse starts a-to-d conversion and thus updates the address before the device 'speaks'.

Following address selection, recording is performed by pushing S2 and holding it down while speaking into microphone.

Switch S1 pulls down pins P/R and /CE via diodes. At the same time, Tr1 stops conducting for a moment as the result of ground to the base via C1 and creates a positive reset pulse from collector to pin PD.

Recording is stopped by releasing S2 and thus writing an end-of-memory mark to the

The ISD2560 chip is available from Sequoia Technology Limited, Tekelec House, Back Lane, Spencers Wood, Reading, Berks RG7 1PW. Telephone: 01734 258000 Fax: 01734 258020 BBS: 01734 258060.
memory. Recording of course also stops if the end of memory is reached.

DIL-switch S3 is used for scaling by feeding user-settable addresses to the memory chip. During normal operation, addresses are received from the converter chip. The light-emitting diodes are only needed for making a clear indication of the binary address. Without them, the address can be read directly from the DIL-switch, but only during setting up.

During setting of the scale, it is essential that the converter outputs are tri-stated at all times. Under normal use, the DIL-switch elements must all be open, and the LEDs show the address coming from the converter.

Making the scale

I used the ISD2560 speech memory chip with 60s recording capacity for the prototype. By varying seven most significant address bits, this time can be divided into about 70 slices, each containing a spoken number and end-of-memory mark. The 7-bit address space is greater than this of course, but the upper part is used by the chip itself.

Thus there is about one second for each number - which requires quick dictating. You might find it better to use an ISD2590 which is compatible, gives 90s recording time and also longer time slices.

The method is to select the address with S3, push S2, speak the number and immediately release S2. The result can be checked with S1 and renewed whenever desired.

At first, the numbers may have the tendency to be too long and occupy space from the next address slot. However with a little practice they can be easily kept in their own address slices and the whole scale spoken in about ten minutes.

Dictated information depends on the desired scaling. If, for example, the device is needed to tell the temperature between -30°C and +40°C, the corresponding numbers are spoken starting from address 0, with S3 totally off. Thus the lowest figure corresponds to address 0, the highest to address 70 and 0°C to address 30 - all set in binary steps via S3.

Because of the shortness of the time slices, you might find it useful to differentiate the frost figures by using, for example, a female voice instead of speaking the minuses before each numerical value.

More time for a unit is of course available, if a coarser or restricted scale is acceptable. By tying address line A3 to ground for example, remaining bits represent half the scale but double length time slots.

Calibrating the thermometer

After the scale is defined, the device has to be calibrated. Because the sensor's temperature dependency is linear, it is enough to fix two points on the scale.

It would be useful to be able to calibrate with icy water at 0°C and boiling water at 100°C. But boiling water is above the device rating. One possibility is to use body temperature of 37°C, or water temperature measured previously against a standard thermometer. Note that for fluid measurements, the sensor must be sealed, for example by casting it in epoxy resin.

First let the sensor rest in ice water for a few minutes. By turning R1 and pressing S1 you can find the point where response is 'zero'. Accordingly the upper reference point is adjusted via R2 with the sensor in a known temperature environment. This second reference point should be as near the 70°C maximum as possible.

Fig. 1. Talking thermometer using a National Semiconductor sensor IC that operates between -40 to 100°C. Since the speaking memory chip can store up to 70 announcements, the system can be calibrated to handle any 70 degree window within the sensor's limits.
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Up a tree

I found Ian Hickman's regression to the early days of radio most fascinating and it pointed me to my library.

One of the most bizarre suggestions I found in the early works is the tree aerial in Morecroft's Principles of Communication, 1921, which comprises a wire nailed to the top of a tree. While he concedes in the caption to the picture that the tree is nothing more than a support for the wire, in the text he suggests that the tree provides top capacity and that the conductive juices of the tree could pick up signals. Another thing that is brought out in Morecroft is the large power output and the large physical size of some of these early equipments which belie the simplicity of their circuitry.

What I find interesting about the early days is how the terminology of these quasi-mechanical generators of rf energy carried on into the valve era. When I started work at Rugby R.S. in 1936 the original short-wave building, in which two transmitters were installed was known by the old hands as the 'Arc room'. An arc transmitter had been built there for testing insulator and other components when GBR was built. Similarly the ground floor of the GBR building, where the banks of valves were installed was known as the 'Valve room'. This title was continued in the newer short-wave building erected in 1929. Also the term 'tank circuit' was not used but the anode circuit was always called the 'intermediate circuit' or 'inter circuit'. This was a throw back to the Leaffield arc where a tuned circuit was that the tree was used as antennas - ed).
**QUESTIONS & answers**

**War and peace**

I notice from the little listening I do on 500kHz that the UK coast stations appear to have gone QRT. Some European stations are still audible at night. Perhaps a reader can throw some light on what the plans are for the mf cw shipping band, in that, as I understand it, Morse is being phased out in the near future. This, I happen to believe is a grave mistake, but such is the price of the march of "progress".

It could be a good idea to retain, at least, a small segment of that band for the Amateur Radio Service for cw only, given that there is no allocation for experimental radio in that part of the spectrum. In that way a small part of the spectrum could be preserved for a useful purpose and a fitting monument to the services of countless Radio Officers in all the Merchant Services who literally lived and died on 500kHz, in war and peace.

Ted Crowley
Ireland

**Can I plate through?**

Could anybody explain to me how I can plate through? I am currently producing pcbs with plated-through holes for double-sided pcbs, and am making the connections between the layers by soldering pins designed for the purpose in the through holes.

Ian Tran
Switzerland

**Transmission line problem**

Can anyone explain in a simple way the answers to the following related questions?

If a short pulse is generated at one end of an open circuit transmission line then that same pulse can be observed to return at a time proportional to the line length and the line velocity factor. Conducting the same test but with the line terminated in a short circuit returns an inverted pulse.

Why does this pulse inversion take place?

If you take a snapshot of the voltage and current when the pulse was halfway down the line, eg a quarter of total elapsed time, what would we see that would indicate if the pulse was going away from or towards the generator? That is, how does the pulse, when it is half way down the cable, know in which direction it is supposed to be travelling?

PW Fry
Southampton

**Big inverter capacitor**

This inverter circuit delivers output at constant power without any special control circuitry being required. The two capacitors C4 and C5 in the half-bridge are made deliberately small so that they will resonate with the leakage inductance of the transformer during operation — at about 65kHz. The transistors Tr1 and Tr2 (igbt's), are simply switched alternately from a vco. A small amount of negative switching overlap preventing shoot-through as well as providing the transistors with zero voltage at turn-on with assistance of the transformer primary self-inductance. This circuit requires a low inductance, low-loss capacitor C1 directly across the dc output for proper termination of the C6,3 and L series-resonating circuit. Transistors Tr1,2 are switched well below the resonance of the series resonating circuit, to produce output in discrete half cycles. Excellent filtering is provided by the mere 35µF of capacitor C4 in conjunction with the self inductance of the converter output leads, which easily exceeds 4µH in application. The degree of smoothness that can be expected at about 150A output is also shown.

The converter delivers very high current into a low impedance load, with the current falling off as the impedance is increased, maintaining V/W at a single value determined by the frequency setting of the vco. This makes it ideal for arc welding, industrial battery charging and the like. It can be used to start a tractor engine directly.

Significantly it uses no inductors other than the step-down transformer. But unlike transformer manufacturers, who can often be relied upon to help with the design and manufacture of smoothing inductors for inverter power supplies, I have not been successful in locating even a single capacitor manufacturer willing to tackle the design and making of the 35µF capacitor for use in the present circuit, C5.

Other than parallelizing tribes of smaller capacitors, can any reader offer a solution?

John Fetter
Bryanston,
South Africa

**On a more perfect note**

I am in the process of composing two pieces of music which I would like to reproduce in a 360° horizontal sound field with sounds occurring everywhere and equal frequency response. I understand the initial recording — a dawn chorus — would be best recorded using a SFM MKV microphone, but I am unable to buy one as I don't have £4500 to spare. I wish to record the information from the microphone directly to four mono channels of a 'pc 'direct-to-hard-disk' recording system which I already have.

All of this is fairly straightforward but the actual manipulation of the data once this is done may be a little more difficult. If any of you have any articles or ideas about the manipulation of audio data in a 360° field or would like to make any suggestions or comments I would be pleased to hear from you. I am also finding it impossible to locate a SFM MKV that I can use for a couple of weeks to make the initial recordings. If any readers are in possession of one I'd love to be able to loan or hire it.

I have spent all my funds on the PC hardware and software and so any sponsorship help would be great as would the loan of any equipment.

---

**The aim of Questions & Answers is to solve readers' problems relating to electronic design and circuitry. If you have such a problem, jot it down and fax it on 0181 652 8956, e-mail it to martin.eccles@rbd.co.uk or mail it to Q&A, Electronics World, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. Note that requests for help in locating equipment and similar enquiries will not be considered. But subscribers should remember that they can advertise for free.**

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**Circuit of an inverter said to be capable of starting a tractor engine, and its ripple at 190A.**

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**Low inductance, low-loss capacitor C5 directly across the dc output for proper termination of the C6,3 and L series-resonating circuit. Transistors Tr1,2 are switched well below the resonance of the series resonating circuit, to produce output in discrete half cycles. Excellent filtering is provided by the mere 35µF of capacitor C4 in conjunction with the self inductance of the converter output leads, which easily exceeds 4µH in application.**
I have already applied to most national funding bodies but the music I am proposing to make falls outside their funding guidelines although most of the replies I received expressed interest in the idea.

Ian McNaughton
Bristol

Amalgamated Wireless Valve Co

If you know what happened to Amalgamated Wireless Valve Co, of Sydney NSW, or you knew of K Langford Smith, would you write to me, Martin Eccles, at the editorial offices please? The address is on the comment page. Alternatively, you can ring me on 0181 652 3128 or fax 0181 652 8956.

Good vibrations

Can anyone help here with information on frequency counters for Schumann Resonance and seismic-signals research?

Schumann Resonance is believed to occur in an insulating gap between the surface of the earth and a boundary at an altitude of 50 to 100km. Using the formula: f=c/2πR, the range of SR frequencies corresponding to the gap is 7.42 to 7.37Hz. However, SR is often quoted as having a specific value of 7.83Hz, which would correspond to a layer just below the asthenosphere, at a depth of 100 to 250km.

Using a Wisher WFC308 frequency meter without an antenna (I bought in 1993 via the columns of EW+WW), I have found SR to vary between 7.06 and 7.39 to 7.44 in August of each year. The values show that SR is bounded by the outer reaches of Earth's atmosphere and the tropopause (-60°C), with 'clustering' at the mesopause (-100°C).

Winter values, from December to February, vary more widely in the range 7.69 to 8.45, corresponding to the middle of the asthenosphere (173km deep) and a layer just below the upper and lower mantle boundary (731km deep). In February 1996, however, 'quantum jumps' from the boundary layer (8.45) to about 14Hz suggest that the process is also correlated with the Gutenberg-Wiechert discontinuity at a depth of 2900km.

If any readers could spare a few minutes to look at their frequency counters it could be of considerable value in helping to better understand volcanic activity and earthquake prediction if they send me details of their display.

Note that not all frequency counters give the desired effect. Settings I used were: range 0 to -1200Hz (50Hz input impedance), count time <0.001 second, display rate -1s⁻¹.

Tony Callegari
Much Hadham
Herts

How does the 340MHz transmitter/receiver work?

Mr Collins' receiver is a 'super-regenerative' type - an oscillator oscillating so hard that it cuts itself off and restarts at a frequency above the audio range. Such a receiver is very sensitive indeed and directly produces the modulation of the incoming signal at its output - whether the signal is amplitude or frequency modulated.

'Cathode Ray' produced a design in Wireless World for a triode fm receiver for headphones at around 1956. I built many examples of this, but such receivers radiate strongly and would now cause tremendous interference problems. The transmitter oscillator used here operates at much lower power and appears from the diagram to have no proper antenna. As a result, its radiation is presumably acceptably low. So the chain of inverters is fed with the modulation and the first and third are working as low-pass filters. The second and fourth are biased as analogue inverters. The circuitry around D1 is a pulse-stretcher.

The transmitter is amplitude-modulated (or one could say pulse-modulated because the modulation factor is 100%) by the variable base voltage from pin 17 of the U5012. Its frequency cannot be calculated from the component values because stray capacitances and inductances are comparable in value with those of the discrete components.

John Woodgate
Kayleigh
Essex

More magnetic lines of force

Referring to Guy Moore's Letters in the May issue, having studied the mechanics of magnetic forces for a number of years I agree with the explanation given by Douglas Rice and Guy Moore for the formation of concentric lines around the wire. However, there are two errors in Guy Moore's letter which should not go uncorrected.

First, the iron filings become magnetised normal to the field and not parallel to the field otherwise the magnetic lines would not repel one another. Also, if the card is tapped gently the diameter of the lines will be seen to decrease as the filings are attracted to the wire.

Secondly, energy in the system is increased and not reduced. The proof comes from the fact that the wire forms part of an electric circuit possessing inductance and a stored energy of $1/2LI^2$ joules. The iron filings store a greater amount of magnetic flux which increases the inductance of the circuit and the energy stored.

Gareth Jones
Gwynedd
Wales

How can headphones produce so much bass?

As an acoustician, I can explain quite easily the phenomena of bass transmission by small headphones. It is basically a tube duct. The small phone can produce the necessary bass by moving the diaphragm forward and backwards. As the air is imprisoned in the ear tube to the diaphragm, this closed system requires nothing more to have the sound transmitted by the 'imaginary piston' of the diaphragm of the headphone. What goes in, gets out at the other end, as in a command tube of a boat or a tube.

Jon Mathys
Lierre
Belgium

Contrary to popular belief, headphone transducers do not work in the same way as loudspeakers. For the latter, the major component of the mechanical impedance of the diaphragm (cone, surround and suspension) is its compliance below the bass resonance frequency and its mass above that frequency.

A headphone transducer's diaphragm, however, is designed so that the major component of its mechanical impedance is resistive at all useful frequencies. Since also it radiates into a closed tube - the auditory canal - its acoustic environment is quite different from that of a loudspeaker cone and analogues cannot usefully be drawn. See 'Loudspeaker and headphone handbook', Borwick (ed.) 2nd ed. Focal Press 1994 ISBN 0 240 51371 1.

JW

Anyone know anything about Schumann resonances?

Detection of ELF radiation is only practicable by means of loop antennas unless one has the resources and privileges of the military. The major problem is eliminating man-made interference. A Faraday screen helps and so does balanced operation of the loop but there are still severe problems with power-frequency (and its harmonics) magnetic fields. The combination of a loop with a microphone-type amplifier having a (proper) electronically-balanced input and a high overload margin, followed by a set of notch filters at 50Hz, 150Hz and 250Hz seems worth trying. One might add a tunable band-pass filter covering 7 to say 45Hz with 1Hz bandwidth. This is only general information as I am not involved in this sort of work.

JW

Unless mentioned otherwise, these answers are for questions that appeared in the previous issue.
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July/August 1996 ELECTRONICS WORLD

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Towards a better 30W amplifier

A computer simulation by Electronics Workbench of Jeff Macaulay's 30W amplifier in September 1995's CIs showed up some small errors, the process also provoking some improvements to the design. The program cannot measure distortion, other than by displaying it on an oscilloscope, so I developed the technique described here.

Input and output signals combine in a summing amplifier to give a difference signal, delay through the test amplifier being matched by that through the all-pass delay amplifier and the signal levels being balanced by control Q.

With the modifications to the original as shown, symmetrical clipping sets in with a load of 48W at 1kHz into 7.5Ω, after balancing the dc reference point by control S; quiescent current was set to 65mA by control B.

At 36W, the upper trace shows a mean level of 3.4mV and dominant second harmonic. Thd was 0.02%.

Reg Williamson
Kidsgrove
Staffordshire

Fig.1. Modified design by Jeff Macaulay and the test setup to measure its thd.

Fig.2. Upper trace shows difference signal after the summing amplifier, indicating 0.02% thd. Lower trace are input and output signals.
Driving 30W into an 8Ω speaker, this amplifier consists of two Class AB stages in a bridge configuration.

To reduce crossover distortion, it does not rely on diodes or current sources, but on the quiescent 30mA forced by the op-amps.

Volume control is based on an arrangement due to R. Williamson, the potentiometer having a linear law for the log. response, and the tone control was described by T H O’Dell. This has a flat response when centred and gives 15dB treble and bass cut and boost.

The TIP31/32 output transistors are mounted on a common heat sink with insulators; no forced cooling is needed when the circuit is contained in a 6 by 6.5 by 3in enclosure. John A Haase Fort Collins Colorado USA

References
Constant-current/voltage control

Originally built to test the change in sound quality from a loudspeaker when driven by constant current and constant voltage, this circuit, derived from the AREAC c-c driver for induction loop amplifiers, can be used with any suitable power amplifier, one of its uses being to compensate for the reactance of loudspeaker leads.

A voltage proportional to load current comes from a very small resistor in series with the load, followed by a gain stage. Output from this is compared with the signal input in an op-amp in a manner similar to that of the Baxandall tone control. Depending on the setting of the potentiometer, the output is either current or voltage or something between the two.

Stability may be a problem: low-frequency stability depends upon correct connection of the sensing resistor relative to the polarity of the power amplifier, while hf stability can be compensated for with the aid of a suitable filter.

---

Gyrator oscillator

Balancing the frequency-dependent negative resistance with a physical resistance produces an RC oscillator with a reasonably wide range of frequencies. In the circuit shown, the frequency of oscillation is that at which the negative resistance and real resistance are equal in magnitude, this being given by \( f = \frac{1}{2\pi RC} \), assuming all resistors to be equal. Trimmer \( C_T \) provides a little negative damping to ensure start-up with minimum distortion.

*John Paul*  
*Nottingham*

Gyrator oscillator can be made to operate in the 1Hz-20kHz range with \( R \) and \( R_{\text{load}} \) 10kΩ, \( C \) at 470pF and \( C_T \) at 2-22pF.

---

Pulse integrating tachometer

A common method of indicating the frequency of a pulse train by moving-coil meter is to connect a large capacitor across the meter. Since, in this case, the pulse train originates in rotating machinery and can be less than 1Hz, a very large capacitor indeed would be needed — 30,000pF or more — hence the use of this integrating circuit. It also possesses better linearity.

Input comes from a photodiode or magnetic pickup and is amplified in the op-amp A1, which has diode feedback to provide a simple automatic gain control. This, via

---
may call for the shunt capacitor across the power amplifier input. Tests using a simple loudspeaker have shown that treble and transient response significantly improved; almost as though a tweeter had been added, but without the inductance of a crossover. It is also educational to watch the waveforms of voltage and current into a pure inductance.

R J Higginson
AREAC Midlands Ltd
Halesowen
West Midlands

Basic circuit for constant voltage and constant current amplification.

buffer $Tr_4$, drives the $Tr_5,6$ complementary pair which provides the input to the pulse-forming network $C_9, D_9, R_{14}$. Integrator $A_3$ has a single fixed capacitor $C_6$, but five 1% switched resistors to give full-scale meter deflection for 1Hz to 10kHz.

Calibration is thereby reduced to an adjustment of $P_2$, the values shown giving a reading of frequency; readings in rev/min will need a different set of values. The value of $C_6$ does not enter into this calculation.

C J D Catto
Cambridge

Tachometer for slow speeds does not need a large capacitor.

Programmable pulse delay

In response to a preset binary input, possibly from switches, this delay circuit produces a single pulse after a delay corresponding to the binary input and clock frequency.

Magnitude comparator 7485 takes input from the 7493 4-bit binary counter on the A pins and the binary input on B pins. The counter takes its drive from a 555 oscillator and reset from the gated A>B output of the comparator.

When the push-button switch is pressed momentarily, the counter is reset and counts up until its 4-bit output is equal to the binary input, whereupon the 555 bistable is triggered; this is gated with the clock output and an output pulse appears.

Nored comparator outputs A>B and A<B reset the 555 bistable and the counter is reset by A>B.

V Gopalakrishnan
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For example, students in the 1950s and 1960s learned to write in pencil, and today's students learn to type. This is a shift from one medium to another, and it has implications for education and society. Similarly, the shift from typewriters to computers has changed the way we communicate and store information, as well as the way we learn. The development of the internet and social media has further expanded these changes, with implications for both individuals and society as a whole.
Differentiator enables RS232/485 converter

A differentiator I have designed is used in this RS232/485 converter to enable the RS485 driver. It is based on just two low-cost ICs—a 74HC14 Schmitt trigger and a DS75176 transceiver. Jumpers JP2-6 allow selection of different turn-around times. Turn-around time is about one character period for the respective bit rates. Output levels do not comply with RS232 specifications, but where RS485 conversion is concerned, cable lengths to DTE and DCE equipment are normally very short so TTL levels suffice.

Input circuitry to the DS1489 RS232 line receiver is similar to that shown. Jumper JP7 allows local echo of transmitted information. Jumpers JP9,10 allow DTE or DCE connection formats while JP8 provides the option for powering the circuit via mains or modem control signals. The green led is on when the RS485 driver is enabled and the red led is on when data is received on the RS485 bus.

Deon Marais
Brackengardens
South Africa

In this RS232/485 converter, a differentiator is used to enable the RS485 driver after a one-character period, switch selectable to suit the bit rate.
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ELECTRONICS WORLD July/August 1996
Two-component oscillator

Analysis of an internally compensated amplifier having two poles at \( w_p \) and \( w_b \) shows that the circuit will oscillate at a frequency of

\[
\omega_0^2 = \omega_p \omega_b + \frac{\omega_p + \omega_b}{CR}
\]

where \( R_o \) is the op-amp output resistance.

Using a HA741 op-amp with \( C \) adjusted over the range 40-15nF, frequency was 0.9-1.2MHz.

Muhammad Taher Abuelma'atti
and Sami Saud Buhalim
King Fahd University of Petroleum and Minerals
Dahran
Saudi Arabia

This has to be the most skeletal oscillator yet seen.

Frequency doubler/differentiator

Frequency doublers seen in these pages in the past use more components than is necessary; this one is rather more frugal.

The negative output pulse period can be changed by varying \( C_R \) and \( R_9 \) and \( R_16 \), the values shown giving about 10µs. For positive output pulses, use the circuit in the second diagram.

Deon Marais
Brackengardens
South Africa

Circuit producing unidirectional pulses for each input transition, for frequency doubling or "differentiation".

50dB logarithmic meter

Using ordinary 1N4148 diodes, this meter amplifier indicates direct or ac voltages from 10mV to 3.5V in one range.

Current forced in to the meter circuit of Fig. 1 is

\[
I_G = I_M \left[ 1 + \exp \left( \frac{R_M I_M}{nU_T} \right) \right]
\]

where is plotted as Fig. 2.

Temperature coefficient of the diode is 0.17%/°C at 50% of full scale and 0.26% at full scale, the copper in the meter coil opposing this change to give a nett coefficient of 0.1%/°C.

Bandwidth of the circuit shown is 50kHz-3dB at 15% of full scale and 100kHz-3dB at 50%. Amplifier input voltage offset should be no more than 2mV.

Tore A Nielsen
Malov
Denmark

![Fig. 1. Basic logarithmic meter circuit.](image1)

![Fig. 2. Measured meter current for 1N4148 and 1N4007 diodes, the full line showing the calculated curve for a 1N4148.](image2)

![Fig. 3. Logarithmic meter amplifier with 50dB range and no switching.](image3)
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To illustrate the method, consider the low-pass filter shown in Fig. 1. Its general transfer function is,

\[ F(s) = \frac{R}{sR + 1} \]

where \( R = R_1 + R_2 + R_3 \) and \( R = R_4 + R_5 \). Cut-off frequency, \( Q \) and gain \( G \) are then

\[ \omega_0 = \sqrt{C_1 C_2 R R_3} \]

\[ Q = \frac{R_1 R_2}{R_1 R_2 R_3} \]

\[ G = \frac{R}{R_1} \]

Sensitivity \( S \) to variation in \( Q \) with changes in \( R \) is given by,

\[ S = \frac{R dQ}{Q dR} = \frac{1}{2} \frac{C_1 (R_2 + R_3) + m R C_2}{s C_2 R R_3} \]

where \( m = R_4 / R_3 \). Solving these equations for \( R \) and \( m \) produces expressions for the practical design of the filter, namely,

\[ R = \frac{4Q^2 C_1 R_1}{C_2 B^2} \]

\[ m = \frac{A B C_2 + 4Q^2 C_1}{4Q^2 C_2} \]

where \( A = 2S - 1 \) and \( B = 2S + 1 \). These expressions are simplified by making \( C_1 = C_2 = C \).

Suppose the low-pass filter requires a 1 MHz cut-off, \( Q \) of 25 and must exhibit sensitivity to variation in \( R \) of no more than 9. The above equations in \( R \) and \( m \) give \( R = 10k \) for \( R_3 = 1.51 \mu F \) and \( m = 1.1292 \), making \( k = 0.469 \).

The value of \( G \) must be less than \( 1/0.469 \) and is chosen to be 2, so that \( R_1 = 11k \) and \( R_2 = 167k \). If \( R_3 = 0.11 \mu F \), \( R_4 = 1k \). \( R_5 = 167k \) for \( R_5 = 0.11 \mu F \), \( C_1, C_2 = 0.01 \mu F \).

Figure 1 shows the band-pass design, which is governed by the following equations,

\[ \frac{1}{Q} = \omega_0 \left[(R C_1 + C_2) - m R C_2 \right] \]

\[ G = \frac{R_2}{R_1 + R_2} (1 + m)Q \sqrt{x} \]

where \( x = R_3 / R_1 \) and \( m = R_4 / R_3 \).

Solving the \( Q \) and \( S \) equations for \( m \) and \( x \) gives,

\[ m = \frac{4s^2 - 1}{8s^2} \]

\[ x = \frac{2(2s + 1)}{m(2s - 1)} \]

For a band-pass filter with \( f_0 = 1 \) MHz, and a 3dB bandwidth \( B \) of 20 kHz, sensitivity to resistance variation must be less than 9. \( Q = f_0 / (2\pi f_0) = 50 \), so \( m = 0.01615 \) and \( x = 138.4083 \). For \( G = 20 \), \( R_1 = 29R_2 \), for \( R_1 = 1k \), \( R_1 = 20k \) and \( R_2 = 1k \). If \( R_4 \) is made \( 20k \), \( R_5 = 0.016R_4 = 320 \) for \( x = 138 \), \( R_3 = 138 \) for \( x = 138 \).  

Reference

In Part 2 of his article David Sharples discusses the mechanisms by which TV and monitor deflection transistors fail.

DESIGNING CRT
deflection

Unfortunately, horizontal deflection transistors do fail from time to time. However, failure rates to low parts-per-million levels can be achieved if the correct investigations are carried out during the design stage. The following notes discuss the potential failure modes and how these should be addressed at the design stage.

Studies of device failure patterns show that a device could fail either soon after the TV set is built or after some years use. The fundamental cause of failure is quite different for these two cases.

Immediate or short-term device failure can be attributed to one of the following: circuit design weakness, a set manufacturing fault, or a weak device shipped by the supplier. Long term failures are the result of a chronic change in the load and/or the drive. Current information suggests that long term failures due to device degradation are remote and, therefore, insignificant.

By nature and design, power semiconductor devices often act as a fuse when something goes wrong elsewhere in the circuit. The semiconductor industry is moving towards new technologies that produce non-destructive fuses, but many of these benefits are unlikely to be applied to deflection transistors within the next few years.

Short-term failures

Careful design can achieve short-term failure rates at a very low ppm levels. Television and monitor manufacturers cannot tolerate percentage fail-offs in the production line, or high warranty returns from the field.

The key component status given to the deflection transistor allows detailed investigations to be carried out during set development to eliminate the significant failure modes. However, short-term failure modes are not always eliminated completely and problems can occur in production.

A device can fail simply from too much load current, too little drive in the on-state, or by too much voltage in the off-state. These failure modes are easy to detect and are usually corrected before they become a major embarrassment.

Physical analysis of device failures suggests a more complicated common failure mode: exceeding the reverse-bias safe operating area, or rbo. The rbo defines the current-voltage, Ic-Vce, boundary a device can withstand during turn-off, i.e. base reverse biased. The rbo of a bipolar transistor is very application dependent and can also be device-to-device dependent. For this reason data is often not given and a 100% production test even less so. The rbo for a Philips BU2522AF is shown in Fig. 1.

For any deflection transistor, the critical region is above the Vce limit. This is usually 600-800V for a device with Vces of 1500V. Failures are most likely to occur as the device turn-off trajectory passes through the 600-1000V region. Physically, the devices are driven into a secondary breakdown state; the transistor then becomes a path of low resistance. There is sufficient energy stored in the flyback capacitor to then damage the transistor beyond repair.

For television, there is one acknowledged fault condition that can lead to a device exceeding the rbo - 'picture tube flash'. This is when the eht required for the CRT finds a low resistance path to ground, or a point of lower potential. The eht is generated by the line output transformer and deflection transistor; a low resistance path effectively shorts the secondary of the transformer.

As the system recovers, large currents are drawn from the secondary which in turn lead to large currents in the primary. The deflection transistor can then see several cycles with the Ic much higher than for normal operation. Such an occurrence can lead to the rbo being exceeded and, possibly, the failure of the device.

The causes of such an occurrence are beyond the scope of this report, but the phenomenon is acknowledged throughout the industry as a serious fault condition.

The deflection transistor suppliers' response to this problem is to provide rbo data for new product releases and develop more 'rugged' products. One recent trend is an increase in the Vces limit to 1700V; this gives a more 'rugged' product but at a higher cost. Philips Semiconductors offer a range of 1700V deflection transistors for TV and monitor applications.

Some monitor designs incorporate a separate circuit to generate the eht. In such designs, the deflection transistor is no longer vulnerable to 'picture-tube flash'. However, some multi-frequency monitor designs present another fault condition that can lead to the rbo of the deflection transistor being exceeded - namely mode change.

As a monitor switches mode, several functions are performed that lead to a change in the horizontal line rate; for instance a typical PC monitor can operate between VGA mode at 31.5kHz and enhanced resolution 1280x1024 mode at 64kHz. If the integrity of the drive is not maintained during a transition the deflection transistor can be forced outside its rbo.

Recent monitor designs have much greater control during mode change which leads to products less likely to fail.

Long-term failure

Long term failure modes can either be associated to the ageing of the TV or an infrequent but gross overstress event. A knowledge of the deflection transistor can sometimes assist in a speedy solution.

A TV several years old is not likely to fail due to some
strange and interesting phenomenon associated with the set and ambient. Experience suggests that the more likely failure mode is an electromechanical degradation of a component in the drive or load of the deflection transistor. Power devices often act as fuses when things go wrong and a parametric shift in one component value can result in the functional failure of the deflection transistor and, hence, the TV or monitor. In such cases, replacing the deflection transistor with a new part often produces the same failure in quick succession. Without either a wealth of previous experience, or an oscilloscope analysis of the waveforms, a true fix is often not possible. A knowledge of the characteristics of deflection transistors can help.

Physically, the failure modes of 'new' and 'old' deflection transistors are identical. The phenomena outlined in the short-term failure section are relevant to long-term failures. Some simple tests can be done without any instruments to detect the cause of failure. Where fails can be almost immediately reproduced an experienced television engineer will be able to arrive quickly at the true cause. If subsequent fails are not found how can we be sure that a complete repair has been made? This is illustrated in the panel 'Fault finding...' from last month's article.

Alternatives to bipolar devices?
As design engineers develop a greater affinity with their PCs at the expense of their soldering irons so the trend to find an alternative to the bipolar high-voltage transistor gathers pace. Everyone knows that you can feed the output of a typical IC into the gate of a mosfet to turn it on and off; there is no need for a drive transformer. Mosfet switching times tend to be measured in nanoseconds rather than microseconds so the concept of storage time can be eliminated. Simple mosfet device models are available making a mosfet deflection circuit a dream to simulate. However, the dream soon becomes a nightmare. The on resistance of high voltage mosfets is dominated by the bulk epitaxial silicon which has to be very thick — many tens of micrometres — to sustain 1500V. This makes the on-state voltage across equivalent sized bipolar and mosfet devices much higher for the mosfet.

To reach parity with the bipolar device, the mosfet requires a much larger piece of silicon, which in turn requires a bigger package; both of these increase costs. Mosfet processing involves more mask stages and processes that are inherently more expensive. At today's prices, the cost of a drive transformer and bipolar high-voltage device is less than the mosfet equivalent.

The high on resistance of the mosfet brought about the birth of IGBT's. Fundamentally, these devices combine the ease of mosfet drive with bipolar high-voltage transistor type voltage drops in the on-state. Considerable investments in IGBT technology have been made by all the leading discrete semiconductor suppliers, and 1000V igbts for applications up to 10kHz are readily available.

For a deflection transistor minimum feature sizes tend to be tens of micrometres rather than sub-micrometre. Because of this, established and mature processes are utilised that apply to these requirements is not identical. This is reflected in device designs that are specifically for each market. It is apparent that the differences will become more exaggerated in the next few years.

Bipolar deflection transistors, like most other discrete semiconductors, do not demand state-of-the-art processing. For a deflection transistor minimum feature sizes tend to be tens of micrometres rather than sub-micrometre. Because of this, established and mature processes are utilised that keep the production costs down: deflection transistor prices do not have to support the overheads associated with 8in wafer Class 1 fabrication lines.

But as the IC industry continuously strives for smaller, high quality, high throughput products so a range of tried and tested processes become available to discrete. Integrated-circuit processes cannot be directly transferred to deflection transistors but the processes can be adapted using commercially available equipment.

To summarise, advances in bipolar high-voltage transistor devices will ensure that the requirements for horizontal deflection will be met with ever-improving, reliable designs.
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How do multiple processors talk to each other? John Mitchell explores the utopian world of multiprocessors where all tasks are shared equally.

Intel has just released the first of its new multiprocessing motherboards for servers using the Pentium processor. This CPU is ideal for use in multiprocessing systems due to its advanced architecture, separate on-chip code and data caches, chip sets for controlling external caches, and sophisticated data integrity features.

The term symmetrical multiprocessing, or SMP, refers to the operating system’s ability to schedule code for execution on the first available CPU, rather than a specific CPU. That is, in an SMP system the next piece of code for execution always goes to an idle CPU. Thus the total amount of processing in the system becomes shared equally between all the processors, with each one executing very nearly the same amount of code, Fig. 1.

Many applications will benefit from running on an SMP system, with some of the most common being on-line analytical processing, simple client/server setups and communications—including remote-access servers, e-mail routers and World Wide Web servers. Symmetrical multiprocessing benefits these applications because they come under a heavy user and processing load, in that large amounts of information are read into memory, processed and then written back to disk.

Everything old is new again

Using multiple processors on a single motherboard, or even in a single computer, is nothing particularly new, especially in mini-computers. But Intel’s offering will be the first that adheres fully to the company’s own MultiProcessor Specification, which has been developed by the Intel Architecture Labs, IAL.
in cooperation with leading oems, osys and bios vendors.

Many current vendors of multiprocessing systems based on Intel architecture cpus use proprietary designs that require operating systems to be customised to run on specific hardware. The high cost of supporting multiple versions of the operating system and platforms makes it uneconomical for multiprocessing system vendors to make their products more widely available. This results in a lack of choice of hardware and operating systems for end users.

The multiprocessor specification defines a standard way for the operating system to communicate with the hardware. This will make it easy for osys and oems to support a wide range of platforms with one operating system version - a benefit they already enjoy in the uniprocessor desktop market.

Further, multiprocessor capable operating systems will be able to run without modification on all multiprocessor systems that comply with this specification. The result of this will be that end users who purchase a compliant multiprocessor system will be able to run their application on all multiprocessor systems that comply with this specification. The result of this will be.

The specification covers PC/AT-compatible multiprocessor platform designs based on Intel processor architectures and the all-important Advanced Programmable Interrupt Controller, or APIC, architectures. In the specification, Intel uses the term 'PC/AT compatible' to refer to software-visible components of the PC/AT - not to hardware features. Thus an implementation of the multiprocessor specification may include one or more bus types, such as ISA, EISA, VESA local bus, PCI or other proprietary busses.

The goal of the multiprocessor specification is to enable scalable, high-end workstations and enterprise server systems that provide computer users with superior price/performance and have the ability to execute all existing AT binaries as well as multiprocessor-ready software packages on shrink-wrapped multiprocessor operating systems. The specification defines a system architecture based on the following hardware components:

- One or more processors that are Intel architecture instruction set compatible, such as cpus in the Intel 486 or Pentium processor family
- One or more APICS, such as Intel's 82489DX, or an integrated APIC such as that on the Intel 735/90 and 815/100 processors, together with a discrete i/o APIC unit.
- Software-transparent cache and shared memory subsystem
- Software visible components of the PC/AT platform

Choose your system

There are several models and connection schemes that can be used to tie together the various components of a multiprocessing system. Intel's multiprocessor specification incorporates a tightly coupled, shared memory architecture with a distributed interprocessor and i/o interrupt capability. It is fully symmetrical in that all the processors are identical, have equal status and can communicate with every other processor. This symmetry has two important aspects:

Memory. Memory is symmetrical when all the processors share the same memory space and access that space using the same addresses. This confers a particularly important feature onto the system: the ability for each processor to execute its own copy of the operating system.

Fig. 1. General structure of a design based on Intel's Multi-Processor Specification.

Fig. 2. Configuration of advanced programmable interrupt controllers with respect to bootstrap and application processors.
Input/output. The I/O system is symmetrical when all the processors share access to the same ports and interrupt controllers, and any processor can retrieve interrupts from any source. Such an arrangement increases system scalability by helping to reduce I/O bottlenecks.

Figure 2 shows the configuration of the APICs with respect to the CPUs. While all the processors in a multiprocessor system are functionally identical, this arrangement classifies them into two types: a single bootstrap processor, or BSP, and one or more application processors, APS. This differentiation is only apparent during the initialisation and shutdown processes, as the BSP is responsible for initialising the system and booting the operating system and APS are only subsequently activated.

APICs are based on a distributed architecture in which interrupt control functions are distributed between the local and I/O APIC. These two units communicate via a bus called the interrupt-controller communications, or ICC, bus, as shown in Fig. 2, with multiple units operating together as a single entity. The I/O unit senses an interrupt input, addresses it to a local unit, and sends it over the ICC bus.

The APICs are collectively responsible for delivering interrupts from sources to destinations throughout the entire system. In a multiprocessor system each CPU requires a single local APIC, but depending on the total number of interrupt lines there may be more than one I/O APIC used.

The APICs help to achieve the goal of system scalability by off-loading interrupt traffic from the memory bus, thereby making greater bandwidth available for processor use. They also help processors share the interrupt-processing load with other processors.

Interrupts and write back cache

The local APICs also provide interprocessor interrupts, or IPIs, which allow any processor to interrupt any other, or any set of processors. Compared with a single-processor system, symmetrical multiprocessing puts great demands on memory bus bandwidth, proportional to the number of processors in the system. A well-designed system will therefore implement a high-performance secondary (external) cache. Use of such a cache can push the scalability limit upwards by reducing traffic on the bus and increasing bandwidth.

In this regard, the Pentium processor’s data cache uses an important technique called write-back caching. The write-back method transfers data to the cache without going out to main memory, with data being written to main memory only when it is removed from the cache.

In contrast, previous-generation write-through cache implementations transferred data to the external memory each time the processor wrote data to the cache. The write-back technique increases performance by reducing bus utilisation and preventing needless bottlenecks in the system.

A potential source of problems that designers must guard against is the question of cache
coherency. That is, when one processor accesses data in another processor's cache, it must not retrieve incorrect data. Further, if a processor modifies data, other processors that access that data from a cache must not receive out-of-date data.

To ensure that data in the cache and in main memory are consistent, the Pentium processor's data cache implements an algorithm called the MESI, for modified-exclusive-shared-invalid, protocol. By obeying the rules of the protocol during cache reads and writes, the Pentium processor can maintain cache consistency and circumvent problems that might be caused by multiple processors using the same data.

Snooping around for memory
Another common technique that is used to ensure memory coherence is called bus snooping, which refers to the ability of a CPU to monitor memory addresses placed on the system bus by other CPUs or devices. When a CPU detects a memory address on the system bus that the CPU knows it already has stored in its on-board cache, the CPU writes that memory address from its cache to system memory before the completion of the bus cycle.

Intel's MultiProcessor Specification defines three different interrupt modes:

- PIC mode: bypasses all APIC components and forces the system to operate in single-processor mode
- Virtual-wire mode: uses an APIC as a virtual wire, but is otherwise the same as PIC mode
- Symmetrical i/o mode: enables the system to operate with more than one processor.

The first two modes are included to provide PC/AT compatibility, and at least one of them must be included. A multiprocessor system is booted under one of these modes; later the operating system switches to symmetrical i/o mode for full multiprocesssing.

The PC/AT software compatibility of PIC mode stems from the fact that it uses the same hardware interrupt configuration. As shown in Fig. 3, the hardware for PIC mode bypasses the APIC components by using an interrupt mode configuration register, which controls whether the interrupt signals that reach the bsp come from the master PIC or the local APIC. Before entering symmetrical i/o mode, either the bios or the operating system must switch out of PIC mode by changing the interrupt mode configuration register.

Virtual-wire mode provides the equivalent of a uniprocessor system that is capable of running all DOS software. Figure 4 shows that in virtual-wire mode the 8259A-equivalent PIC accepts all interrupts, and the local APIC of the bsp becomes a virtual wire, which delivers interrupts from the PIC to the bsp via the local APIC's local interrupt 0 (LINT0). The i/o APIC is not used.

Finally, as shown in Figure 5, we have symmetrical i/o mode, the heart of a multiprocesssing system. In this case i/o interrupts are generated by the i/o APIC and all 8259 interrupt lines are either masked or work together with the i/o APIC in a mixed mode.

Fly in the ointment
All this may paint a very rosy picture of SMP, but there are a few greenfly in the garden. The main problem is that the standard AT bus does not have sufficient bandwidth. Theoretically an SMP machine should provide a 100% performance increase every time you add a processor.

It has been found that this is indeed so up to a total of four processors, but that after this, performance returns begin to diminish. This is partly because of the processing overhead required to manage individual threads and processes as they are started, suspended and restarted across the system.

However these problems may be solved with the introduction of the new intelligent input/output device driver specification, which distributes I/O functions across multiple processors, and hierarchical PCI buses. Systems with these features do hold the promise of greater scalability with fewer problems caused by I/O bottlenecks.
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- **3324A synthesizer/function generator, 20kHz**
- **3560A audio frequency spectrum analyser**
- **3561C direct-voltage voltmeter**
- **37780 primary multiples analyser**
- **4146B powermeter, DC -voltage source**
- **4338A high resistance meter c/w lead set 16117B**
- **4275A multi-frequency lcr meter**
- **4388 microwaves power meter, analogue**
- **5386 3GHz frequency counter**
- **54100A 10GHz digitizing oscilloscope**
- **6007B pulse generator 10MHz**
- **6018A serial data generator**
- **8362A pulse generator 250MHz**
- **8111A pulse generator 250MHz**
- **8154A Switch/FET mointer**
- **816A standard f6 1.8-15GHz with 891C and 447B probe**
- **8444A tracking generator with option 058**
- **89556 synthesized signal generator to 92GHz**
- **87510A gain-phase analyser 100kHz-300MHz**
- **8915A modulation analyser with option 095**
- **2219A 480-based, colour option main-frame**
- **2219A 468-based colour screen option network**

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