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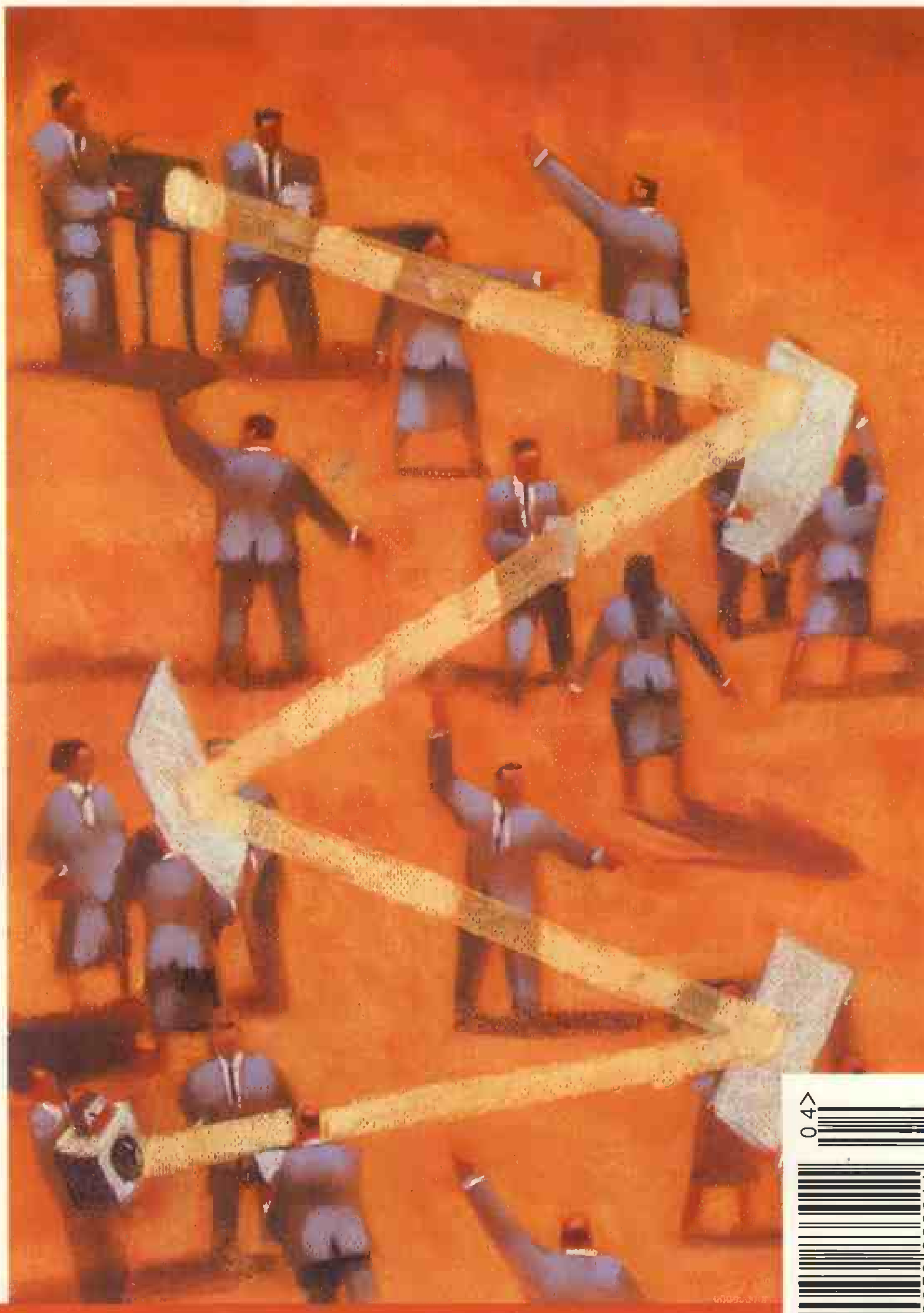

OPTICAL STORAGE – present and pending

Slew rates become much more important at high levels – Ben Duncan provides evidence on page 303.

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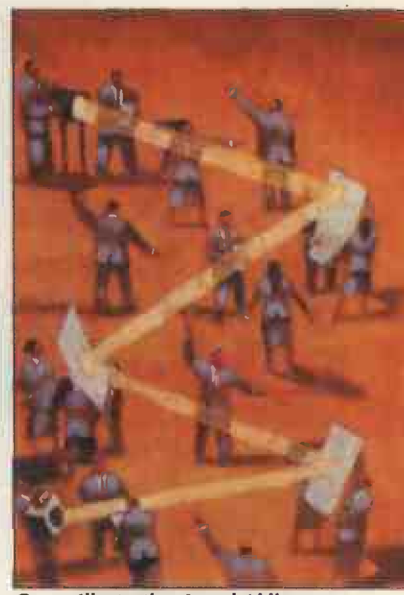
REGULARS

Special offer – see page 296



Cover Illustration Jamel Akib

Optical storage technologies – present and future – explained, page 280.

New laser technique could result in fewer visits to the dentist – page 272.

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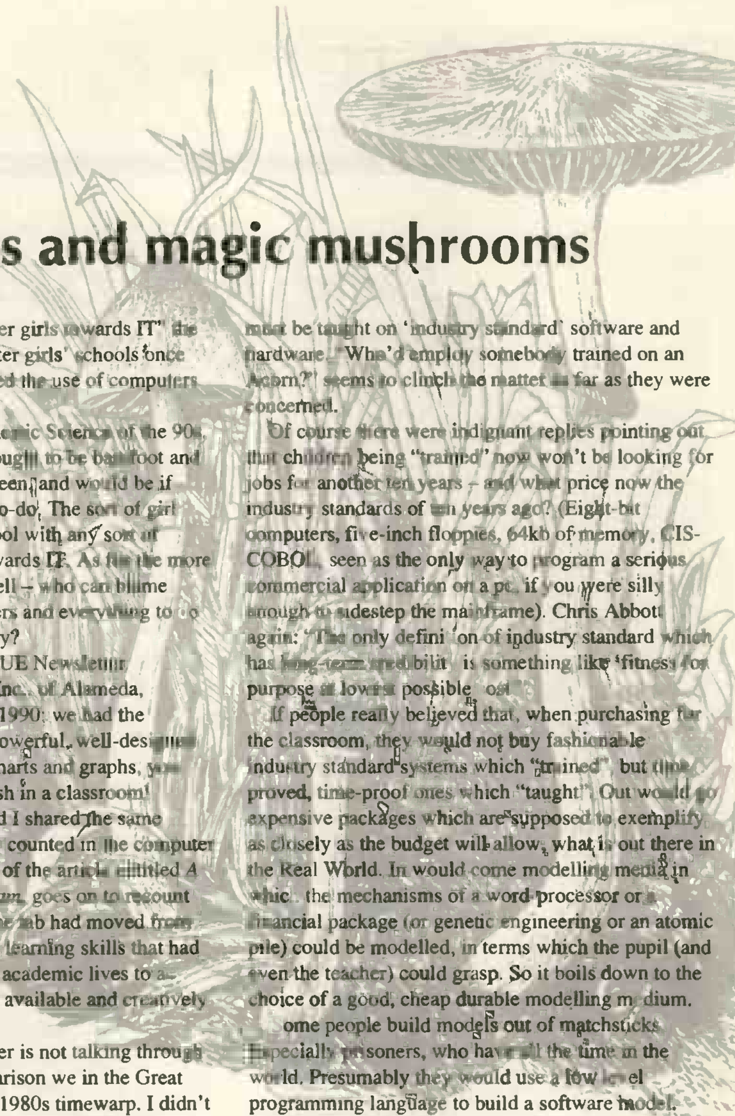

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Matchsticks and magic mushrooms

"We do tend to push the weaker girls towards IT" the headmistress of one of the better girls' schools once confided in me, as we discussed the use of computers in education.

So there it is. IT is the Domestic Science of the 90s, fit only for cotton-heads that ought to be barefoot and babbit by the time they're sixteen, and would be if their parents weren't so well-to-do. The sort of girl that's lucky to be leaving school with any sort of qualification. So push 'em towards IT. As for the more academically-minded girls, well – who can blame them if they consider computers and everything to do with them beneath their dignity?

I read in a recent issue of CUE Newsletter (Computer-Using Educators, Inc. of Alameda, California): "The dilemma in 1990: we had the technology, we could create powerful, well-designed word-processed documents, charts and graphs, you name it. What power to unleash in a classroom! Unfortunately my students and I shared the same secret – all of these skills only counted in the computer classroom." Schandler, writer of the article entitled *A Goal Without a Plan is a Dream*, goes on to recount how things have changed. "The lab had moved from the place where students were learning skills that had little relevance to their real or academic lives to a studio where tools were made available and creatively used."

Assuming that Ms Schandler is not talking through her sweatband, then by comparison we in the Great Britain of 1995 are stuck in a 1980s timewarp. I didn't say 1990s because in the eighties we were ahead of the Californians in the constructive use of computers in the classroom.

But the world moves on – and it seems Britain doesn't. Chris Abbott, writing in *Educational Computing and Technology*, November 1994, recounts his embarrassment at having to tell erstwhile overseas visitors, who had come to this country to see what had been achieved by the network of LEA centres, that most of them have closed. "The 1993 Education Act suggests that private sector centres will develop overnight, like so many mushrooms, where LEA centres close. No such magical events have taken place." He judges that, "there are only two kinds of organisation which now have the funding, the resourcing and the legal right to develop new structures: the universities and the IT industry."

Both of course have their own agenda. Industry will argue, as its running-dogs have been doing in the correspondence column of *Computing*, that children must be taught on 'industry standard' software and hardware. "Who'd employ somebody trained on an Acorn?" seems to clinch the matter as far as they were concerned.

Of course there were indignant replies pointing out that children being "trained" now won't be looking for jobs for another ten years – and what price now the industry standards of ten years ago? (Eight-bit computers, five-inch floppies, 64kb of memory, CIS-COBOL seen as the only way to program a serious commercial application on a pc, if you were silly enough to sidestep the mainframe). Chris Abbott again: "The only definition of industry standard which has long-term credibility is something like 'fitness for purpose at lowest possible cost'."

If people really believed that, when purchasing for the classroom, they would not buy fashionable industry standard systems which "trained", but time-proved, time-proof ones which "taught". Out would go expensive packages which are supposed to exemplify, as closely as the budget will allow, what is out there in the Real World. In would come modelling media in which the mechanisms of a word-processor or a financial package (or genetic engineering or an atomic pile) could be modelled, in terms which the pupil (and even the teacher) could grasp. So it boils down to the choice of a good, cheap durable modelling medium.

Some people build models out of matchsticks. Especially prisoners, who have all the time in the world. Presumably they would use a low level programming language to build a software model. Those of us for whom time (and patience) is in short supply need to model with larger components and subassemblies we could in principle build ourselves – or at least take apart and understand. More like Lego than Lucifers.

Who can manufacture these goodies for us? Universities? When I worked in a university it was academic suicide to be caught making things easy for people with an IQ of less than 100. And as for industry – well! Who's paying? What are they buying?

I'm not being cynical. Both parties play the game by rules which are handed down to them. It's up to our rulers to make rules which are productive and beneficial, supposing they feel sufficiently motivated to do so. Education of the next generation – isn't that sufficiently motivating? Not if your mentality is straight out of "Chitty-Chitty-Bang-Bang".

Ian Clark, Educational Interfaces, Bishop Auckland

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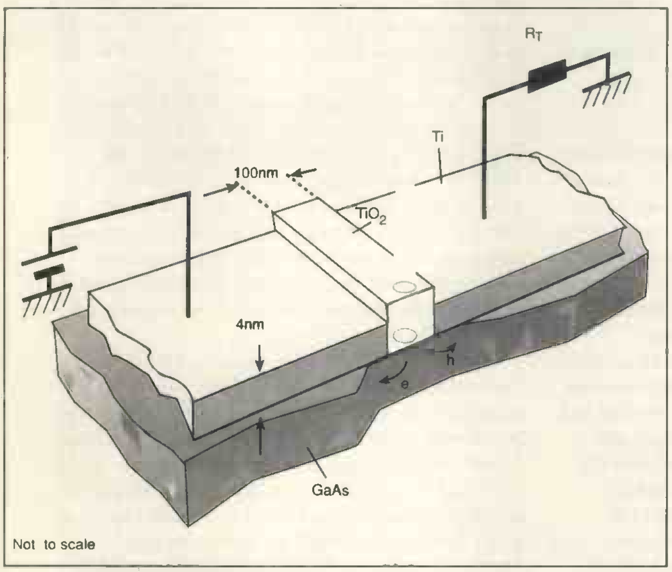
UPDATE

Switch for 500Gbit/s data

Researchers at the Electro Technical Laboratory in Japan have recently developed an 'Auston' switch that can produce electrical pulses only 570fs in duration. This very fast electro-optic transducer opens up the prospect of communication at data rates above 500Gbit/s.

The switching element is a gap only 100nm wide between the ends of two titanium strips laid down on the surfaces of a special gallium arsenide substrate. The switch is triggered by a 40fs laser pulse focussed onto the GaAs surface through the gap. The incident energy causes electron hole pairs to be formed in the substrate, briefly connecting the ends of the strips together. The narrow gap and the special cold grown substrate, which ensures that residual pairs recombine rapidly, mean that the switch turns off again in less than 600fs.

A 100nm wide gap between two titanium strips forms the active element of this experimental high-speed switch, developed in Japan.



Actual measurement of the switch closure time is performed using a lithium-tantulate crystal connected to the switch by a transmission line. The electrical pulse from the switch passes under the crystal, changing its refractive index. This change is detected using light pulses from the same laser.

The switch cannot be used as a receiver in current optical communications systems because light from existing optical fibres has insufficient energy (the wave length is too long) to trigger the switch. Development of an effective optical 'up converter' is needed before the full potential of the switch can be realised. The only current practical use is as a detector in nuclear accelerators where the electron-hole pairs are formed by particles passing through the substrate.

Steve Bush, *Electronics Weekly*

● Silicon switching speeds are expected to be pushed close to 100GHz with the launch this Autumn of the first working devices built using 0.1μm lithography. The devices are expected to result from collaboration between Sandia National Laboratories and AT&T Bell Labs. A fast field effect transistor, running at 90GHz, should be the first device produced with a 0.1μm gate (see picture caption story on page 271).

Suppliers pressure BT on high ISDN prices

ISDN equipment suppliers ganged up on BT at last week's London ISDN user show, forming a pressure group to force reductions in ISDN charges. They are angry at the high price BT charges for installing basic-rate ISDN lines, claiming it is stifling the ISDN market.

Although BT is running a special offer of a £300 installation fee, down from £400, this is still much higher than France's £80 and Germany's £50. Some suppliers say Mercury should enter the basic rate market (Mercury only supplies primary rate ISDN) and force BT into a price war.

At the show, suppliers held the inaugural meeting of Agis (Action Group of ISDN suppliers), aiming to pressurise BT and exchanging information to ensure interoperability of equipment. Mark Heath, ISDN marketing manager for Chase Research, said: Our intention is not to be a beat-up BT group, although everyone would like to beat them up."

BT agreed to send a representative to an Agis meeting to answer questions. More queries will no doubt be voiced by Dataflex Design, one of the group's founding members, which went into liquidation in mid-February but has now been bought out by Amstrad.

Chip growth beats expectations

Chip boom expectations for the third successive year are causing semiconductor industry analysts to revise their market forecasts for 1995 in a hurry.

Motorola has jacked up its capital expenditure plans this year to $4.5bn, compared to $3.3bn last year, forecasting chip market growth between 17 and 21 percent this year.

Jerry Junkins, president of Texas instruments, reckons that the world chip market will grow 21 percent this year to reach $124bn compared to $100bn in 1994.

Mike Glennon of US market analysts Dataquest, which had forecast 14 percent worldwide growth for semiconductors this year said, "My personal opinion is that 14 percent was too low; in Europe we're looking at a 15 to 20 percent rise and world-wide we could be seeing 20 to 25 percent increase this year."

If the forecasts turn out to be correct, this will be an unprecedented third year in a row for 20 percent plus growth in the semiconductor industry. In 1993 the industry grew 31 percent, says Dataquest, and last year it grew 28 percent.

TI believes the European market will grow 22 percent, the US market 22 percent and the Japanese market 17 percent. But the star of the show will be non-Japanese Asia ('Asia-Pac') with growth of 32 percent.

SiC grows into power player

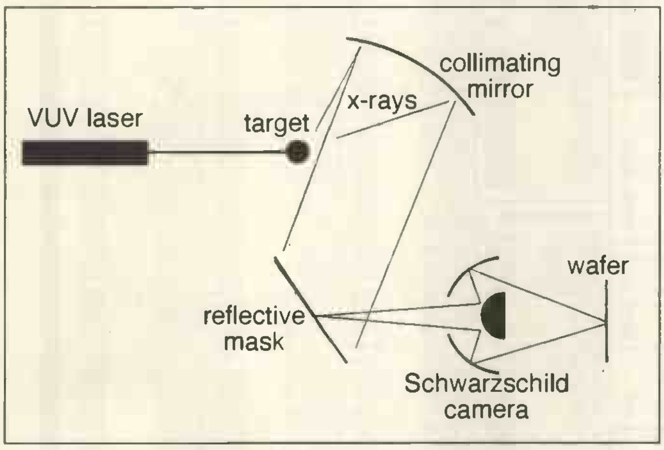
Faster power semiconductors, handling substantially higher voltages and temperatures, could be appearing in commercial applications within three years, according to latest Swedish research. The performance breakthrough will come through projected advances in silicon carbide (SiC) technology, from which the new devices will be fabricated.

Theoretical advantages of silicon carbide have been known by electronics engineers for some years. But its high melting point and extreme hardness – the very properties that have given it such popularity in the tooling industry – have made practical electronics application difficult.

Now, findings from a joint power semiconductor research programme conducted by ABB, Industrial Microelectronics Centre and Linköping University, predicts that commercial production of simple devices using SiC will be possible within three to five years. Resultant devices should need less space, have much lower power losses and generate fewer harmonic currents than conventional power semiconductors.

The attraction for power specialist ABB is that the technology could be applied to high-voltage ac (hvac) and possibly high-voltage dc (hvdc) power transmission. But SiC could also enable electronics to be used in environments that have previously proved too extreme, such as car engine and automotive applications and melting furnace sensing.

All three research organisations have now moved into the second phase of their development programme, aimed at commercialising the process.
Jonathan Campbell



Above – a production facility for making 0.1µm geometry chips, built by Sandia National Labs, uses ultra-violet laser focused on a target material. A small lump of the material is vapourised turning into ionised gas. As this gas absorbs more laser energy, it heats and the plasma shines X-ray light. The rays are collimated by a mirror and shone on a reflecting mask made of a substrate coated with layered synthetic material. Circuit pattern is drawn on this with a non-reflecting organic material. A 'Schwarzschild camera', using spherical reflecting surfaces instead of lenses, makes a shrunk X-ray image of the mask pattern on the wafer. Special photoresists capable of working with soft X-rays have been developed.

Cellular battle

Mobile communications suppliers Motorola and Interdigital Communications have begun a court battle which could have significant implications for the whole cellular telephone industry. Motorola is the first to contest Interdigital's patent claim over the tdma channel coding protocol forming the basis of most digital mobile 'phone networks, including GSM.

Interdigital is reported to have asked Motorola for a $200m royalty patent. Motorola claimed the patents were invalid and Interdigital is pressing for increased damages and an injunction to stop Motorola making TDMA phone systems.

AT&T, Siemens and Matsushita have agreed licences with Interdigital, but the world's largest suppliers are awaiting the outcome of Motorola's court case, which is expected to last until April.

Competition for Intel's P6

Cyrix says its *M1* microprocessor can take on the Intel *P6* and win according to its architect, Mark Bluhm. The firm is readying the microprocessor for a June introduction with samples now being delivered to key customers.

"The figures being put out by Intel show the *P6* has about a 33 percent performance advantage over *Pentium*, given equal clock rates," said Bluhm. "What we are seeing with the *M1* is a two times performance increase over the *Pentium* in some benchmarks and overall a 50 percent advantage."

The current *M1* samples run at 100MHz and Cyrix plans a 133MHz version to be available early next year. By the third quarter this year Cyrix anticipates it will be shipping about 100,000 units a month. The majority of those chips will be manufactured by IBM and delivered to about five key customers.

"We've got a *P6* class machine that will be delivering revenues months before Intel," added a Cyrix spokesman. "At least one of the customers we are currently sampling plans to make the *M1* their high-end microprocessor."

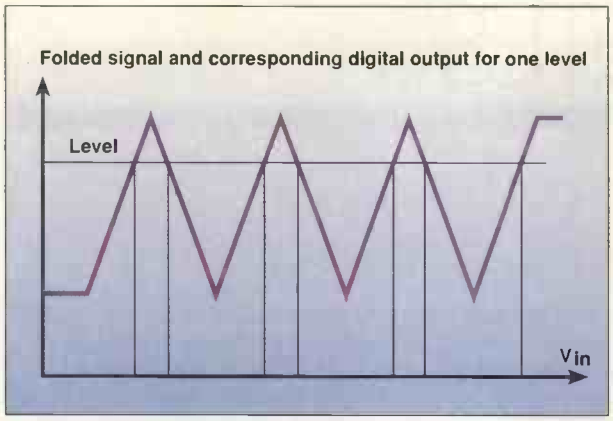
Cyrix originally planned to introduce the *M1* late last year but problems forced it to delay the chip's introduction. The *M1* is similar to the *P6* in that it uses a super-scalar, super-pipelined design.
Tom Foremski and Simon Parry, *Electronics Weekly*.

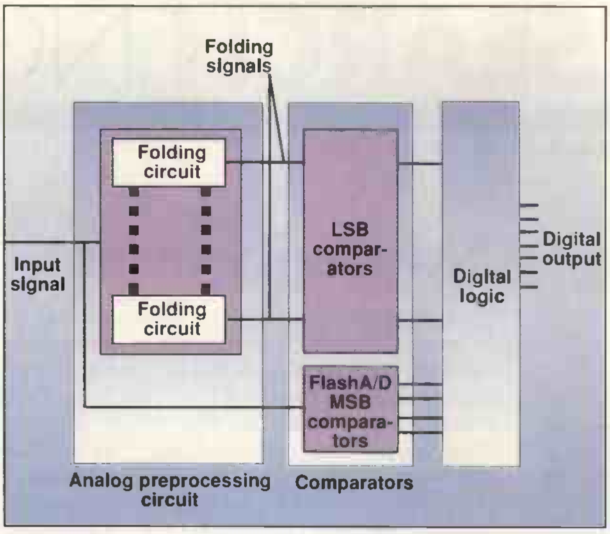
Videoconference standard for analogue lines

Telecommunications manufacturers and operators plan to finalise a videoconferencing standard by the end of the year. It will support videophone and data services over existing analogue telephone lines. The ITU standards committee has approved a first draft of the H324 videoconferencing standard which is the analogue line equivalent of the H320 ISDN-based standard. The intention is to have a final draft ready by November with the first analogue pc videophone cards expected to appear early in 1996. Effectively this will kill off attempts by BT and AT&T to impose proprietary protocols.

The standard will lean heavily on existing data transmission protocols and silicon. H324 uses the V.34 28.8kbit/s data modem protocol, and the channel is divided into a 5.3 or 6.3kbit/s audio stream using a new audio compression algorithm. This leaves around 22kbit/s for compressed video. Core video compression is the discrete cosine transform based H.261 algorithm used in H.320. To achieve the level of compression needed to squeeze 30 frames/s video picture into a 20kbit/s stream, an interpolated motion estimation on P and B frames, similar to that used in the MPEG standard, is implemented.

According to Mike Whybray of BT Research's video group at Martlesham, picture quality is still limited by the 20kbit/s bit rate. "It is better than existing analogue videophones but not as good as two channel ISDN videophones", he added.



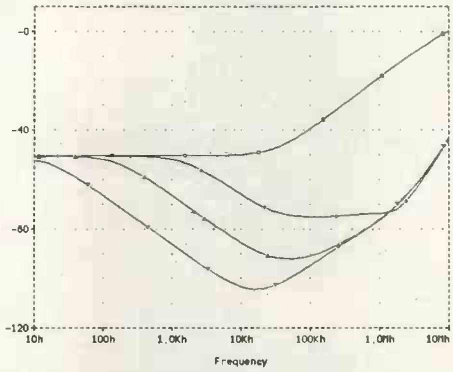


Placing Europe among the leaders in ultra-fast a-to-d converter design, folding interpolating technology uses analogue preprocessing to produce a re-entrant transfer function (above). This reduces the number of comparators needed, which in turn lowers power consumption relative to full flash designs. The re-entrant technique reduces the number of input stages needed since quantisation levels between comparator stages are interpolated. In addition, analogue preprocessing of the input means each comparator detects more than one level of input.

UK company Phoenix Design, in collaboration with Thomson, is currently working on a folding interpolating device capable of converting 8 bits to 1GHz. This type of device is needed for example in spectrum-surveillance radar counter measures. The technique is already being used to produce commercially available byte-wide converters operating to 650MHz. ■

Well off the rails...

Douglas Self's otherwise excellent article on power amplifier distortion derived from the power supply was marred slightly due to the erroneous replacement of Fig. 12 by a duplicate of Fig. 6. Here is the correct Fig. 12. Apologies to you, the readers, and to Douglas of course – ed.



RC *filtering of the negative rail is effective at medium frequencies, even with 1000µF of filtering. Resistance* R *is 10Ω. Top to bottom, curves show 0, 10, 100 and 1000µF decoupling on V–.*

Building the Tesla coil?

Malcom Wells, author of the article on Like Lightning in the March edition, has sent us these further notes that will be of interest to anyone thinking of constructing Tesla's coil.

It is interesting to note that the main secondary of Tesla's very large Colorado Springs coil has a height/diameter ratio of 1.25:1. This, according to a table of values compiled by Medhurst, gives an optimally low value for 'H' of 0.46 in the secondary self-capacitance equation of my article. Also, the secondary was mounted well off the ground, which further reduced its capacitance. The very useful formula developed by Medhurst allows a coil to be designed for a predictable resonant frequency, to avoid clashing with radio beacons.

In my article, in the box entitled *Essential Equations*, there was an error in the equation relating to the minimum height per turn for the primary coil. In addition, the equation should have been separated from the text. The equation should have read,

$$\text{Minimum height per turn} = \left[\frac{3V_c}{N_p} + D_{wire}\right]\text{mm}$$

where V_c=peak primary capacitor voltage in kilovolts, N_p is the number of turns and D_{wire} is the diameter of the pipe – which should be as large as possible – used for the primary coil in millimetres. This gives a minimum clearance of 3mm/kV between turns.

The sphere should be mounted *d*mm above the secondary, and finally, the toroidal terminal should be mounted d_1/2mm above the coil. ■

RESEARCH NOTES

Jonathan Campbell

No-cavity laser

As the laser melts tooth enamel to improve cavity resistance, a computer monitor plots the temperature of the tooth during heating up and cooling. (Picture James Montanus)

Few people, outside the most fanatical of curry eaters, will have wondered what it it might be like to have their teeth melted. But if scientists from the University of Rochester and Eastman Dental Center reach their goal we could all one day share in that experience – and have healthier mouths into the bargain.

The trick is, say the researchers, to use a specially-tuned CO_2 laser to raise the outermost 5μm of the tooth to 1000°C, instantaneously melting, then fusing, the enamel coating. Enamel that is more chemically resistant to the acids that cause cavities should be the effect, with fewer fillings needed.

The laser is tuned to 9.3 or 9.6μm, rather than the conventional 10.6μm, as at these wavelengths the light is absorbed almost completely by the enamel. This, and using 25 100μs pulses at a time, enables the surface of the tooth to be melted while its core is unaffected. When the enamel fuses after treatment, it is claimed to be 70-85% more resistant to attack – a figure reached by dunking treated teeth in acid for 7h then in a saliva-like solution for 17h to simulate conditions in the mouth.

So far the researchers have used only extracted teeth in the laboratory, and more studies are needed before tooth melting is a useable technique for dentists.

But why wait. Just insult the waiter before you order your next chicken madras and try it out for yourself.

And not a laser in sight.

Dope hope for drams

Large doping concentrations required for some elements of high-density drams make *in situ* arsenic doping of polycrystalline silicon an attractive option for vias or substrates. With arsenic, autodoping effects on access devices are lower than with phosphorus-doped polycrystalline silicon.

But transferring arsenic doping development technology into manufacturing reality has been difficult because of slow deposition rates and radial non-uniformity across the wafer caused by addition of the dopant gas.

RPS Thakur and C Turner of Micron Semiconductor look to have found a straight-forward solution using conventional low pressure chemical vapour deposition (*Appl Phys Lett*, Vol 65, (22), pp.2809-2811).

The two researchers have simply used a standard vertical thermal reactor to deposit a stack of doped and undoped layers up to a target thickness. Redistribution of the dopant is then achieved by post-annealing.

The method could be easily integrated for high volume production of thicker polycrystalline silicon films used for dram access memory cell capacitor plates in cmos semiconductor technology.

Following the road to driverless cars

Fleets of autonomous vehicles effortlessly steering their way between our towns and cities may be the stuff of science fiction. But work being carried out at the Robotics Institute, Carnegie Mellon University, and the National Institute of Standards and Technology (Nist), is bringing that day ever closer. Nist has already linked together a perception system and steering/control on a robotic vehicle.

Now, using a new algorithm (Henry Schneiderman and Marilyn Nashman, *A discriminating feature tracker for vision-based autonomous driving, IEEE Transactions on robotics and Automation*, Vol 10, No 6) the vehicle has been kept centred in its lane, under a variety of conditions, at speeds of up to 100km/h. It was even able to keep on track in the rain, at dusk and at night with headlights.

The researchers say that their new algorithm is different because it explicitly addresses the uncertainty concerning how quickly the road changes with time, and also takes into account the uncertainty of the visibility of lane markers in each individual image. As a result the vehicle is able to cope with 7m gaps in lane markers (pavements edges, white lines etc) and momentary loss in their visibility.

Though the system is reported to have performed well, the researchers say they must now develop algorithms of increasing reliability and robustness under all driving conditions.

Splinter in the eye could be a chip

Successful bench-testing of a prototype microchip retina, designed to be surgically implanted in the eye, is being hailed as real advance in development of a bionic vision system. Such a system could help overcome one of the world's most common forms of blindness.

In a complete system, the ultra-thin microchip will work with a miniature camera and laser fitted on a pair of spectacles. Its purpose is to by-pass defective rods and cones by stimulating healthy nerve cells in the eye directly with tiny electrical currents. If successful, the project could mean a breakthrough for people suffering from retinal diseases where the rod and cone cells – the cells in the eye that receive light – have been destroyed.

Retinitis pigmentosa is the leading inherited form of blindness, affecting about 1.2 million people worldwide. The condition causes a slowly progressive loss that first affects peripheral vision but eventually consumes all vision. Similarly, macular degeneration impairs central vision and removes the ability to read, though peripheral vision is maintained. In both, the healthy retinal nerve cells that would have passed on the visual signals from the rods and cones cannot transmit that information to the brain. Blindness is the result.

Now, in a project led by Professor John L Wyatt of Massachusetts Institute of Technology's Department of Electrical Engineering and Computer Science and the Research Laboratory of Electronics, and by Dr Joseph F Rizzo of the Massachusetts Eye and Ear Infirmary and Harvard Medical School – a wide variety of scientists from different fields is making progress towards a working technology.

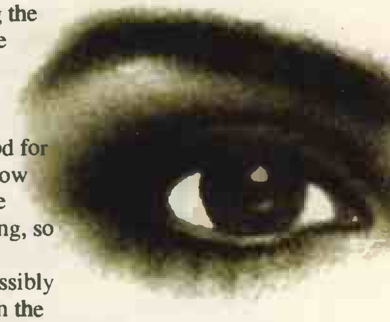
So far the team has designed, and successfully bench-tested, a prototype of the microchip, using an external laser. The laser powers the chip via an invisible infrared beam that will also convey the visual information sensed by a tiny electronic camera (the researchers have not yet tested the laser with the camera). Both camera and laser will fit on a pair of spectacles.

As part of the programme, researchers have also developed new techniques for implantation and have completed a number of tests to determine the electrical stimulation thresholds of cells in the eye.

Many challenges still lie ahead, with perhaps the greatest being the potential for damage to delicate retinal tissue at the interface between retina and implant.

But the team's immediate objective is to refine the method for applying the silicone coating now used on the implant. Tests have revealed tiny leaks in the coating, so a more reliable encapsulation method must be developed, possibly employing new materials. Even the smallest leak of salt from the eye into the implant would destroy the function of the chip.

So far the researchers have successfully recorded signals from the visual part of animal brains following electrical stimulation to an area of the retina roughly as large as the implant will stimulate. The next major goal will be surgical implant of the completed prosthesis and verification of the brain's response to the implant.



Toning up hearing aid control

Clever design of a small, simple and low power wireless receiver promises to make life a little easier for the hard of hearing. Building the receiver into a hearing aid will allow users to vary their aid volume via simple remote control, while the dual-tone multi-frequency technology exploited is only a small step away from wireless programming of aids to suit individual ear characteristics.

The volume-control receiver measures 1918μm by 1109μm and is being fabricated in low-threshold-voltage cmos by AMS International of Austria. It has been designed by Alexander Reyes and Edgar Sanchez-Sinencio at Texas A&M University, and J Francisco Duque-Carrillo at the University of Extremadura in Spain (*A Wireless Volume Control Receiver for Hearing Aids, IEEE Trans on Circuits and SystemS II: Analog and Digital Signal Processing*, Vol 42, No 1, 1995). The design has three main blocks: a detector to select the correct incoming dtmf signals; a decoder to process the frequencies and decide if a valid command is present; and a gain stage which changes the volume of the hearing aid to a new value.

Frequency range for most hearing aids is 100Hz to 8kHz, so audio frequencies are used to activate the receiver, with the dtmf control frequencies selected to avoid harmonics and distortion.

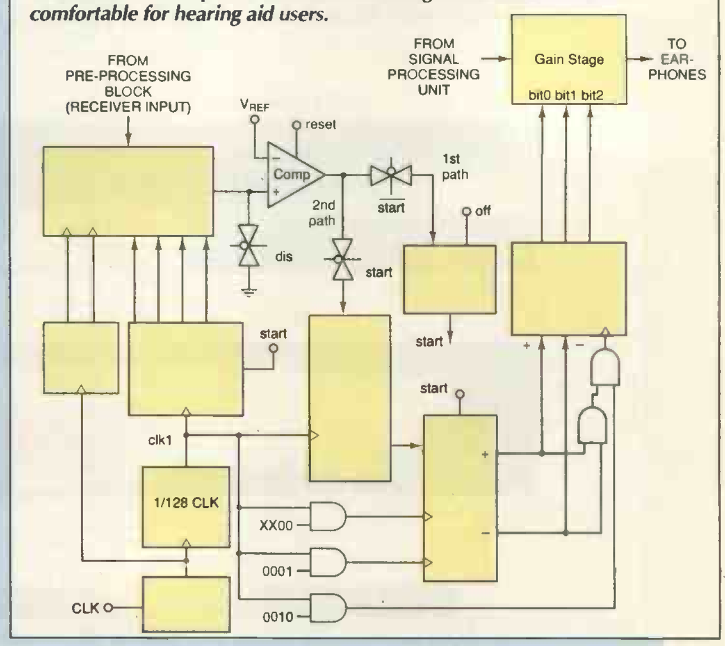
Previously, dtmf receivers have called for at least two filters and usually amplitude detectors, digital logic, voltage references, zero crossing detectors etc. Typically they are implemented in a layout at least 2.4 by 3.2mm and consuming 1.25mW.

But high performance of the new receiver – it detects and decodes audio frequencies within 0.41% of their nominal values – has been obtained by squeezing the most out of each of the various sub-circuits. For example the operational transconductance amplifier (ota) yields a high voltage-gain of 87dB, and dissipates only 9.3μW, while the single switched capacitor bandpass filter provides high Q and a variable centre frequency control from minimum capacitance area.

Finally, static flip-flops replace other common memory cells, reducing the space needed to implement the control logic.

Using a similar design to decode and store the configuration for hearing aid signal-processing-unit-equalisers could allow the next generation of programmable units to break free of the physical connections now necessary to customise a hearing aid for individual hearing characteristics.

Small-size and low-power dtmf receiver designed to make life more comfortable for hearing aid users.



Squeaks of pleasure

For serious audiophiles already feeling nervously inadequate about how the upper limits of their equipment may be affecting enjoyment (see Ben Duncan's article this issue, and *Letters*, *passim*), there is good news: cds may no longer be limited to their miserably deficient (??) bandwidths.

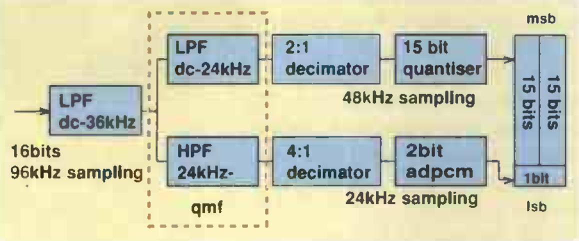
Any technique to increase the bandwidth and dynamic range of cds and dat has the problem that it has to be compatible with current products. Unfortunately, as Mituya Komamura of Pioneer Electronics Corporation reminds us (*Wide-band and wide dynamic-range recording and reproduction of digital audio, J Audio Eng Soc*, Vol 43, No 1/2, 1995), previous work has shown that high-frequency components in music above 20kHz induce the activation of α-eeg rhythms and can affect the perception of sound quality. He also points to a wide-band dat recorder, able to record frequency bandwidths up to 44kHz, that has been gaining a good reputation with audio engineers.

Komamura's solution is that input digital-audio data could be quantised by 16bits at 96kHz sampling frequency, and band-limited up to 36kHz by a low-pass filter. Low-pass output would be split into two sub-bands (dc to 24kHz and 24-36kHz) by a quadrature mirror filter bank. The 96kHz sampling frequency of the lower band signal could then be divided by two, and the higher one by four, so that the sampling frequencies become 48 and 24kHz.

The higher band signal would then be coded by two-bit adpcm and embedded in the least significant bits of 16bit data slots. Such a system would have a bandwidth 1.5 times that of conventional technology. The lower band, coded by 15bit noise shaping quantisation with subtractive dither, would have a dynamic range wider than of cd and dat but would be compatible.

It makes my α-eeg rhythms syncopate just to think about it.

Encoder in a new dat/cd audio system capable of wringing 50% more bandwidth from conventional cd technology.



Satellite tomography maps out ionospheric disturbances

Joint US and Russian trial of a monitoring method that allows electron densities in the upper atmosphere to be plotted as a 'map' has opened the door to better prediction of the ionospheric storms that disturb radio signals and wreck satellites.

The technique – ionospheric radio tomography – has been around for some years. But it is only as a result of the US/Russian study, directly comparing satellite radio tomography with conventional approaches, that the technique has been shown to give good results (*International Journal of Imaging Systems and Technology*, Vol 5, pp.148-159).

The ionosphere is a highly variable part of the atmosphere between 100-1000km. In radio tomography, a satellite sends radio signals through the ionosphere to receivers located at intervals on the ground. Analysis of the radio signals once they reach Earth indicates variations in the density of the electrically charged gas that makes up the ionosphere. The variations can be plotted as contour maps that indicate the general structure of the ionosphere, including small-scale phenomena.

Conventionally, the large radar facilities used to produce images of the ionosphere – there are currently six in the world – are expensive to build and operate, precluding a large world-wide network. But radio tomography opens up the real possibility of global maps of the ionosphere because the receivers involved are small and portable, and can be widely distributed.

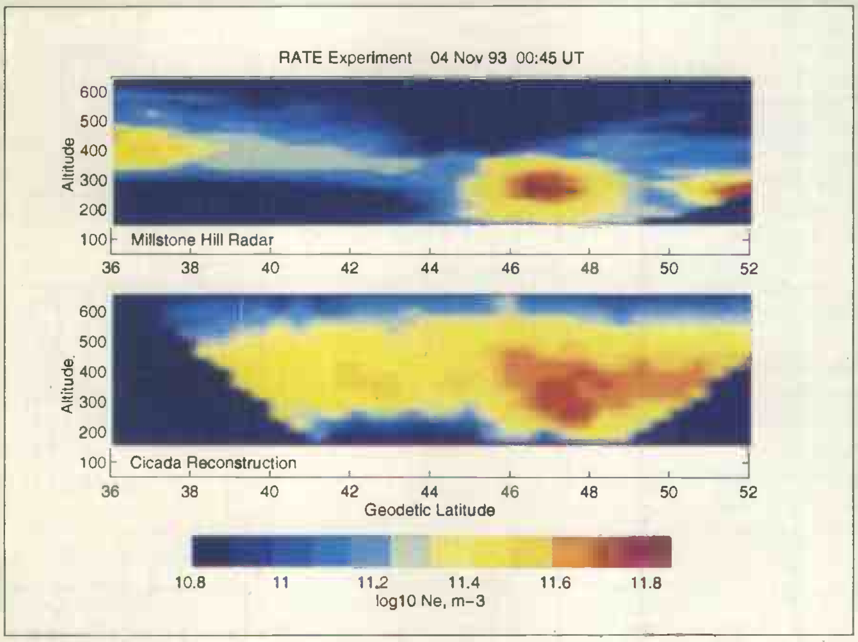
In the US/Russian experiment, the scientists placed four receivers provided by the Russians in a north-south line along the north-eastern US and eastern Canada. Russian navigation satellites flew over these sites every hour, sending down radio signals to all four receivers simultaneously. The resulting data were then analysed to produce an image of the ionosphere using mathematical algorithms developed by the Russians.

Air Force scientists also placed US receivers at the same four sites and recorded signals from US satellites, analysing the data with their own set of algorithms.

Images produced by the US and Russian experimental tomographic techniques were then compared to actual images of the ionosphere made over the same period from the Millstone Hill radar facility in the America.

The result? Both the tomographic images "compared very well to the Millstone Hill results," according to principal investigator for the work John Foster of the Atmospheric Sciences Group, MIT Haystack Observatory.

A bonus to the experiment is that it coincided with a severe ionospheric storm. The large amount of data on the storm, coupled with the severity of the event itself, means that scientists "will publish many more papers on the geophysics of what took place," concluded Dr Foster. ■



Radio tomography plot obtained with the Russian Cicada navigation satellite shows good agreement with the conventional plot from the radar station at Millstone Hill. The plots show the situation shortly after onset of the severe storm.

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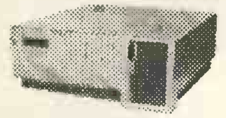

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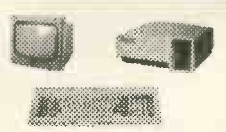

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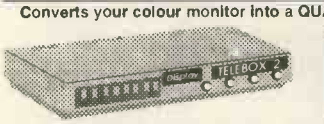

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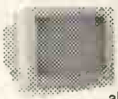

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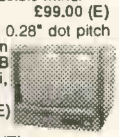

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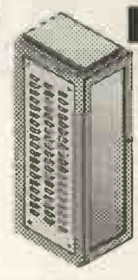

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Front-end FILTERING

Interference from the mains can degrade performance in sensitive instrumentation. A simple notch filter will not always remove hum due to frequency and component drifts, but Radhakrishna Rao has designed a self-tuning filter that overcomes the usual problems.

Active notch filters have become indispensable in many applications where a signal is corrupted by a dominant, single-frequency interference signal. Such a signal is the 50Hz power supply hum in bio-medical systems.

The analogue front-end proposed here is for cancelling the 50Hz power line interference and its harmonics from ecg-emg signals in bio-medical applications. The configuration uses only operational amplifiers and matched mosfets. It requires no precision components and is based on a simple frequency-correction scheme using both phase and magnitude comparison. The scheme could be implemented as a monolithic device.

Common problems

When the frequency of the interference signal is fixed and known, a symmetrical narrow-band notch filter can be used to remove it. For this, the pole ω_p and zero ω_z frequencies of the filter must be made equal to the interference frequency using precision, low-tolerance passive components.

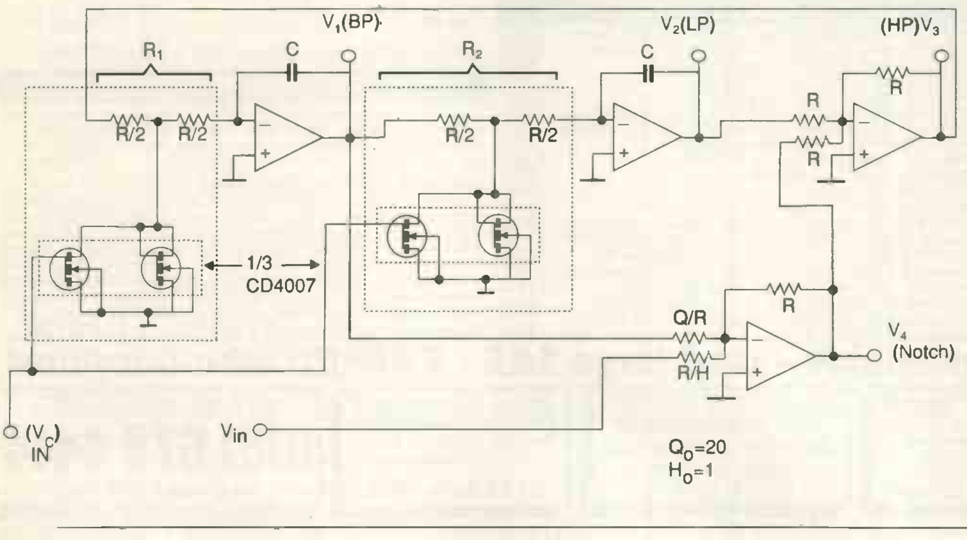
In practice, either there is an uncertainty in the frequency of interference or there is a drift in the values of passive components that determine ω_p and ω_z. Furthermore, the tolerances of passive components cause ω_p to be different from ω_z.

The solution to this problem involves a self-tuning notch filter. Here, the filter is automatically tuned to the incoming interference frequency by making both ω_p and ω_z voltage-controlled.

Power line interference consists of the dominant 50Hz component and its harmonics at 100Hz and 150Hz. The four op-amp modified Kerwin-Huelsman-Newcomb biquad[1] is used for the basic notch filter owing to its low passive and active parameter sensitivities, especially at low frequencies. For this filter, **Fig. 1a**, both ω_p and ω_z are determined by the same set of passive components. Independent outputs for bandpass, lowpass, highpass and notch are available simultaneously. These are voltages V_1, V_2, V_3 and V_4 respectively.

The filter is tuned to the incoming interference frequency by replacing the frequency-determining resistors with voltage-controlled resistors. The control voltage for these is derived by a frequency-correction scheme. As Fig. 1a shows, the frequency-determining resistors, R_1 and R_2, have been replaced by voltage-controlled equivalents using matched pairs of linearised *CD4007* mosfets[2]. The T-network is used to increase the effective variation in resistance offered by the fet. Since a notch filter needs to be tuned, a scheme for filter zero tuning is more appropriate. This is because zeroes control the significant characteristics for the notch filter. Such a scheme is

Fig. 1. A Kerwin-Huelsman-Newcomb (KHN) biquad notch filter is modified by adding linearised mosfets to form voltage controlled resistors. Error signal to control the mosfets (V_C) is derived by magnitude and phase comparison of the highpass and notch outputs (Fig. 2).



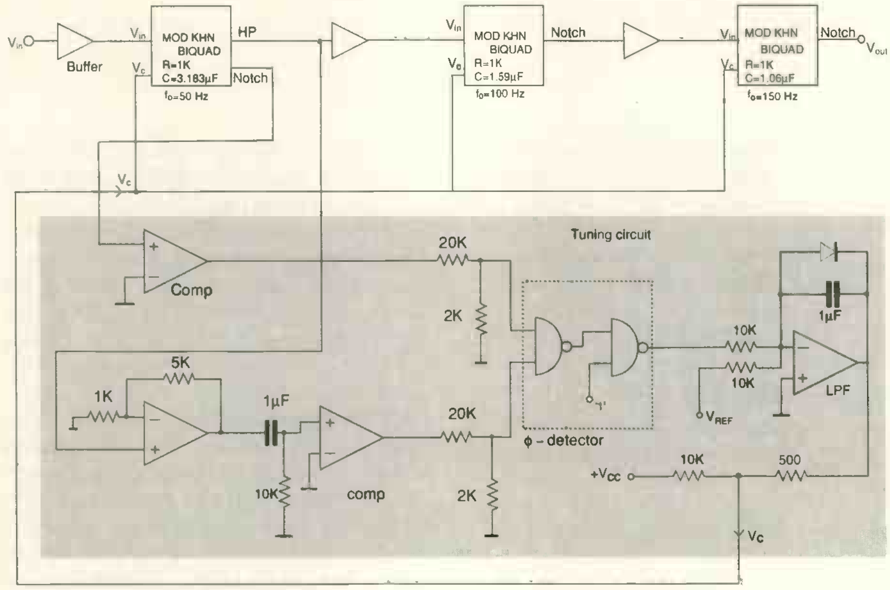


Fig. 2. An error signal is obtained from the phase detector to control the direction of tuning. The magnitude of this error is decided by the difference between the input frequency and the zero frequency. A dc level is then generated by comparing the error signal with a reference voltage.

shown in the lower half of **Fig. 2**.

Frequency correction in this method is based on both magnitude and phase comparison using notch and highpass or lowpass outputs for deriving the error signal. The magnitude of this error signal is proportional to the difference between the input frequency and the zero frequency to which the filter is to be tuned and is obtained using the notch output.

The direction of tuning is obtained by detecting the phase difference between notch and highpass outputs. The magnitude of this error signal is then compared with a reference voltage (V_{ref}) in order to generate a dc control voltage (V_0) which is used for varying the voltage controlled resistances in the filter. The above arrangement, thus, forms a stable negative-feedback frequency-correction loop.

The analogue front-end is then developed as shown in Fig. 2. A cascade of three modified KHN biquads with notch frequencies at 50Hz, 100Hz and 150Hz is employed. The first-stage

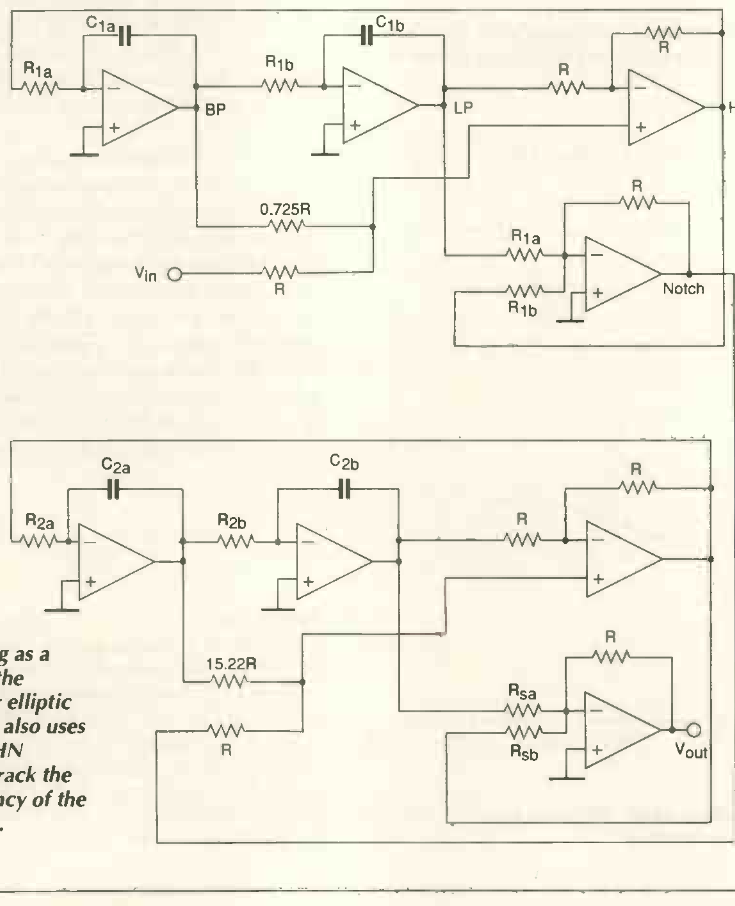


Fig. 3. Acting as a slave filter, the fourth-order elliptic bypass filter also uses cascaded KHN biquads to track the zero frequency of the master filter.

The voltage-controlled resistor

When mosfets are operated in the non-saturating region with small values of V_{DS}, ($V_{DS} \le V_{GS}-V_T$) the drain-to-source resistance is almost linear and bi-directional. By varying the gate voltage, the drain-to-source resistance can be altered and the device acts as a voltage controlled resistor.

Feeding back half the drain voltage to the gate with two large-value resistors, extends the linear operating range.

The advantage of using matched mosfets is that precisely matched resistors can be avoided.

Mos theory demonstrates how the drain-source resistance varies.

In the current saturation region where $V_{DS} \ge V_{GS}-V_T$,

$$I_{DS} = K(V_{GS}-V_T)^2$$

In the non-saturating region where $V_{DS} \le V_{GS}-V_T$,

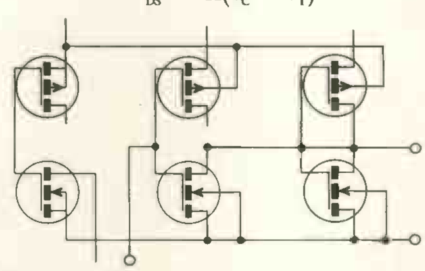
$$I_{DS} = 2K\left[(V_{GS}-V_T)V_{DS} - \frac{V_{DS}^2}{2}\right]$$

$$I_{DS1} = 2K\left[(V_C-V_T)V_{DS} - \frac{V_{DS}^2}{2}\right]$$

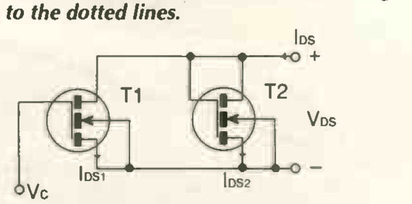
$$I_{DS2} = 2K\left[(V_{DS}-V_T)V_{DS} - \frac{V_{DS}^2}{2}\right]$$

$$I_{DS} = I_{DS1} + I_{DS2}$$

$$r_{DS} = \frac{V_{DS}}{I_{DS}} = \frac{1}{2K(V_C-2V_T)}$$



Part of a 4007 cmos dual complementary pair is used for its matched mosfets. External connections are made according to the dotted lines.



The resulting circuit is a pair of linearised matched mosfets which are used as a voltage controlled resistor.

notch filter is self-tuned to the incoming interference frequency - the dominant 50Hz component - using the above scheme.

The second and third stage notch filters are self-tuned to their respective notch frequencies (100Hz and 150Hz, these being the harmonics of the 50Hz component) using the master-slave approach[3]. To this effect, the first-stage filter is taken as the 'master' and the second and third stage filters as the 'slaves'. The frequency determining resistors in the slaves are replaced by voltage controlled resistors of the same value as for the master's. Matched pairs of mosfets are used for this. The control voltage for these resistors is obtained from the same tuning circuit for the master, Fig. 2. With this mechanism the zeroes of the slave filters, which are filters with notches at 100Hz and 150Hz, are made to track the zeroes of the master; a filter with a notch at 50Hz. Notch frequencies for the slaves are now determined only by the ratios of capacitor values of the slaves to the master's.

Fets rather than multipliers are recommended for voltage controlled resistors. This is because the tuning range obtained using fets can be made just sufficient to cancel the varying frequency components. These are typically 48 to 52Hz for a 50Hz component.

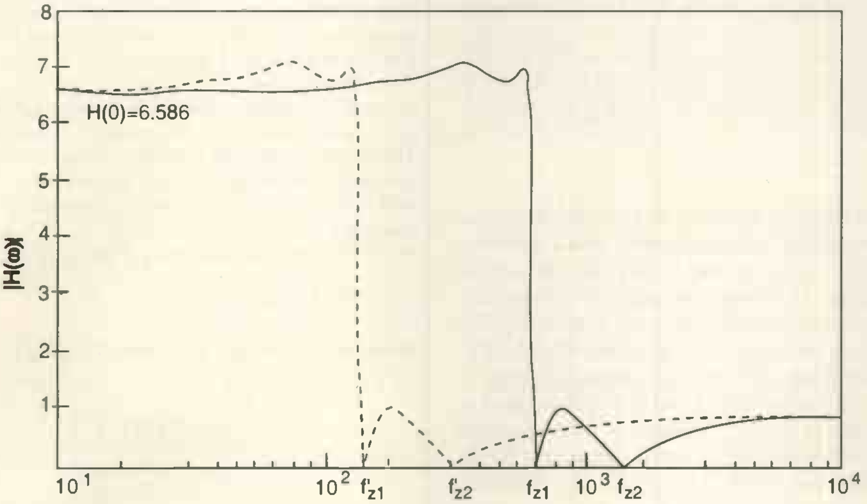


Fig. 4. Tuning ranges for the slave filter are shown in this magnitude response plot. The zeros, f_{z1} and f_{z2} of the slave were made to track the zero frequency f_0 of the master over the entire tuning range of the master.
The master tuning range, TR_m=1.8kHz
Slave ranges are at f_{z1}=735.6Hz and f_{z2}=1348.7Hz
Fot the band f'_{z1} to f_{z1} this gives $f_{z1}=TR_m \times f_{z1}/f_0$ and for f'_{z2} to f_{z2}, $f_{z2}=TR_m \times f_{z2}/f_0$

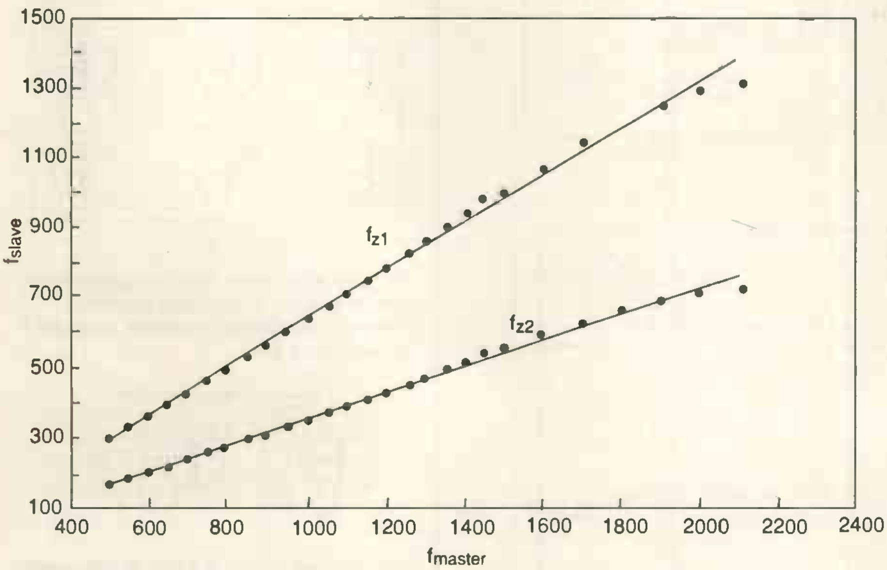


Fig. 5. The tracking accuracy can be determined from these plots; the slope given by the ratio between the the zero frequencies of the slave and the master filters.

Further applications

The above configuration, therefore, forms an analogue front-end for cancellation of power-line interference in bio-medical systems. The scheme can also be extended to the tuning of other monolithic filters such as the inverse chebyschev and the elliptic. The zeroes of the master, which is a self-tuned notch filter, can be made to tune the zeroes as well as poles of such monolithic filters, which now function as slaves.

As an illustrative example, a fourth-order elliptic lowpass filter using cascaded KHN biquads (for which the pole and zero frequencies can be made different) was used as the slave filter, **Fig. 3**. The self-tuned modified KHN biquad functioned as the master. The zeroes, f_{z1} and f_{z2} of the slave were made to track the zero frequency f_0 of the master over the entire tuning range of the master, as observed in **Fig. 4**. Tracking accuracy can be deduced from **Fig. 5**, The slopes of the plots are given by the ratios of zero frequencies, f_{z1} and f_{z2} of the slave to the master zero frequency, f_0.

The given configuration is thus shown to be applicable as an analogue front-end as well as for realisation of monolithic filters. The basic filter section employs only single-ended op-amps in inverting mode and all resistor values in it can be made equal. The frequencies of interest are governed only by the ratios of capacitor values. This allows frequency scaling of the filter's response. Tuning of the filter relies on commonly available matched pairs of mosfets. Such an isotopic nature of the filter topology is suitable for its implementation at vlsi level. ■

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The optical drive

Just as you cannot be too thin or too rich, you can never have too much data storage space. The increasing size of application programs, operating systems, and data files has quickly rendered every generation of hard-drive capacity inadequate.

Furthermore, the trend towards multimedia and image-based files has strained the storage capabilities of hard-disk drives. An alternative to continuously buying new and bigger hard drives is provided by optical disk-drive technology.

An optical disk-drive can provide infinite storage capabilities since extra storage space is easily obtained by using additional disk cartridges. These are relatively inexpensive.

Optical drive technology has progressed rapidly from curiosity to commodity, but according to Praveen Asthana, the evolution has only just begun...

Tape is also capable of providing infinite storage in this fashion, but it is far too slow to be a real alternative, especially when the data needs to be accessed randomly.

Optical disks provide the ideal combination of robustness, low cost and performance. The random access capabilities of optical drives are now approaching those of low end magnetic hard drives.

Optical drives are available in four basic types – compact disk (cd) based, magneto-optic based, phase-change based, and ablative-worm based. While these technologies are all different, the drives do share some similarities in their opto-mechanical technology. I will examine in detail a magneto-optical (MO) drive as it is the most complicated. Designs for the other classes of drives are essentially subsets of the magneto-optical drive design. The recording technologies that define the different optical drives will then be explained.

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Optical head technology

The purposes of the optical head are to transmit the laser beam to the optical disk, focus the laser beam to a diffraction limited spot, and to transmit readout signal information from the optical disk to the data and servo detectors.

Whether the recording technology is magneto-optic, ablative 'write-once, read-many', or phase change, the laser diode is the key component in optical storage. The first two generations of optical drives used infrared lasers emitting in the 780nm or 830nm wavelengths. The next generation of drives will use red laser wavelengths emitting at around 690nm. Continuous laser output power is typically around 40mW. In order to ensure good wavefront quality, the lasers need to be index guided.

A schematic of the optical head in a magneto-optic drive is shown in **Fig. 2**. Output of the laser diode is collimated – i.e. made into a plane wave – by a lens and then passed through beam shaping optics. These adjust the elliptic profile and astigmatism of the beam. The beam then passes through a polarising beam splitter, which reflects some 30% of the beam towards a detector and transmits the rest towards the disk.

The light that is reflected is incident on a light detector. This detector is part of a power servo loop designed to keep the laser at a constant power; output of the detector is connected to the laser driver circuitry. This is very important. Without a power servo loop, the laser power will fluctuate as the laser junction heats up, which can adversely affect the read performance.

The beam transmitted by the beam splitter travels to a turning 90° mirror, called a beam bender, mounted on a movable actuator. During track-seeking operations, this actuator can move radially across the disk. The beam reflected by the turning mirror is incident on an objective lens – also mounted on the actuator – which focuses the light on the disk.

This type of optical head design in which the laser, the detectors, and most of the optical components are stationary while the objective lens and beam bender are movable, is called a split optics design.

In early optical drive designs, the entire optical head was mounted on an actuator and moved during seeking operations. This led to slow seek times of around 200ms because of the mass on the actuator. A split optics design, which is possible because coherent light can be made highly collimated, lowers the mass on the actuator, allowing much faster seek times of around 30ms.

The objective lens also acts as a collector for the light reflected from the disk. This light, used for the servo systems and during reading, contains the readout information. Reflected light follows the incident path up to the fixed optical element. The portion of the light transmitted by the beam splitter is, unfortunately, focused by the collimating lens back into the facet of the laser. Optical feedback affects the laser by causing it to mode-hop randomly, which results in amplitude noise in the laser output beam.

This amplitude noise affects the data signal sufficiently to be considered a major problem. To control the laser noise, injection of high frequency current, also called hfm, into the laser is used in practice. The hfm is usually around 350-500MHz and is of sufficient amplitude to drive the laser below threshold at that frequency. This technique prevents the laser from becoming single mode, because it turns the laser off before it can settle into a single mode. In this way the effects of optical feedback are greatly reduced and the coherence length of the laser is less.

In general, increasing hfm current can decrease the amount of noise. But as a practical matter, the hfm injection current cannot be made too large as it may violate government limits, such as FCC, on allowable electromagnetic radiation from computer accessories.

The light reflected by the first beam-splitter is further split by a second beam-splitter into two components: one for the servo and the other for the data. In this head design, two detectors are shown for data detection. This is specific to magneto-optic read back in which two detectors are needed to implement what is known as differential detection, i.e. the difference in the signals incident on the two detectors is taken.

Phase-change, cd and WORM drives only need one data detector. Light transmitted through the second beam-splitter is incident onto a special multi-element detector which is used to generate the servo signals. The servo system is discussed in more detail later. It is by no means an overstatement to say that the development of high quality servo systems has played a vital role in making high capacity optical disk drives a reality.


Source: IBM Corporation

Fig. 1. This 5.25in Powerbox optical drive, designed as an external SCSI peripheral for pcs and workstations, holds up to 1.3Gbyte per disk. Its 3.5in counterpart holds up to 230Mbyte.

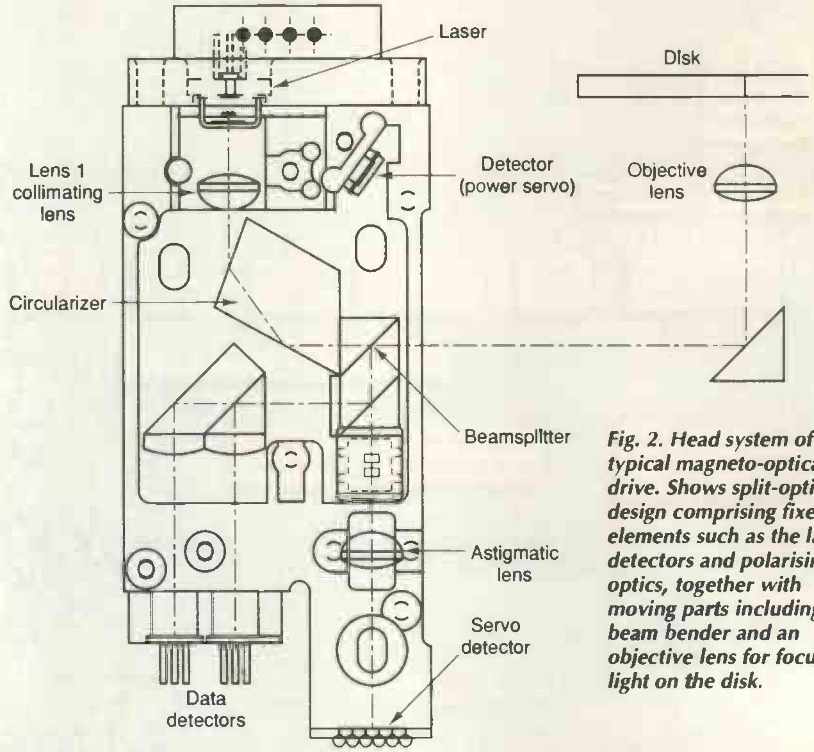


Fig. 2. Head system of a typical magneto-optical drive. Shows split-optics design comprising fixed elements such as the laser, detectors and polarising optics, together with moving parts including the beam bender and an objective lens for focussing light on the disk.

The servo system

The servo system is what enables the focused laser spot to be positioned with accuracy on to any of the tracks on the disk. The extremely high track densities of optical disks – in the region of 18 000 tracks per inch – require that the laser spot position is controlled to within a fraction of a micrometer.

To be able to move across the entire disk surface requires a large actuator. But the larger the actuator, the higher the mass that needs to be adjusted to the rapid changes in the track position – due to run-out in the disk – as the disk spins. Consequently, a compound actuator consisting of a coarse actuator and a fine actuator is used to control the radial position of the laser beam on the disk.

The fine actuator, with its very low mass, can change the spot position rapidly over a limited range. The coarse actuator has a slower response, but has a much wider range of motion and is used for long seek operations. Writable optical disks have a continuous spiral groove, as in a phonograph record, to provide information on the relative track location.

In addition to tracking and seeking, the laser spot in an optical drive must be kept in perfect focus on the disk regardless of the disk's motion. There can be quite a lot of vertical motion if the disk has tilt or is slightly warped. To achieve focus, the objective lens must be constantly adjusted to correct for the axial motion of the disk surface as the media spins.

Lens position is controlled by a focus servo mechanism. A typical focus actuator consists of an objective lens positioned by a small linear voice coil motor. The coils are preferably mounted with the lens to reduce moving mass, while the permanent magnets are stationary. The lens can be supported by either a bobbin

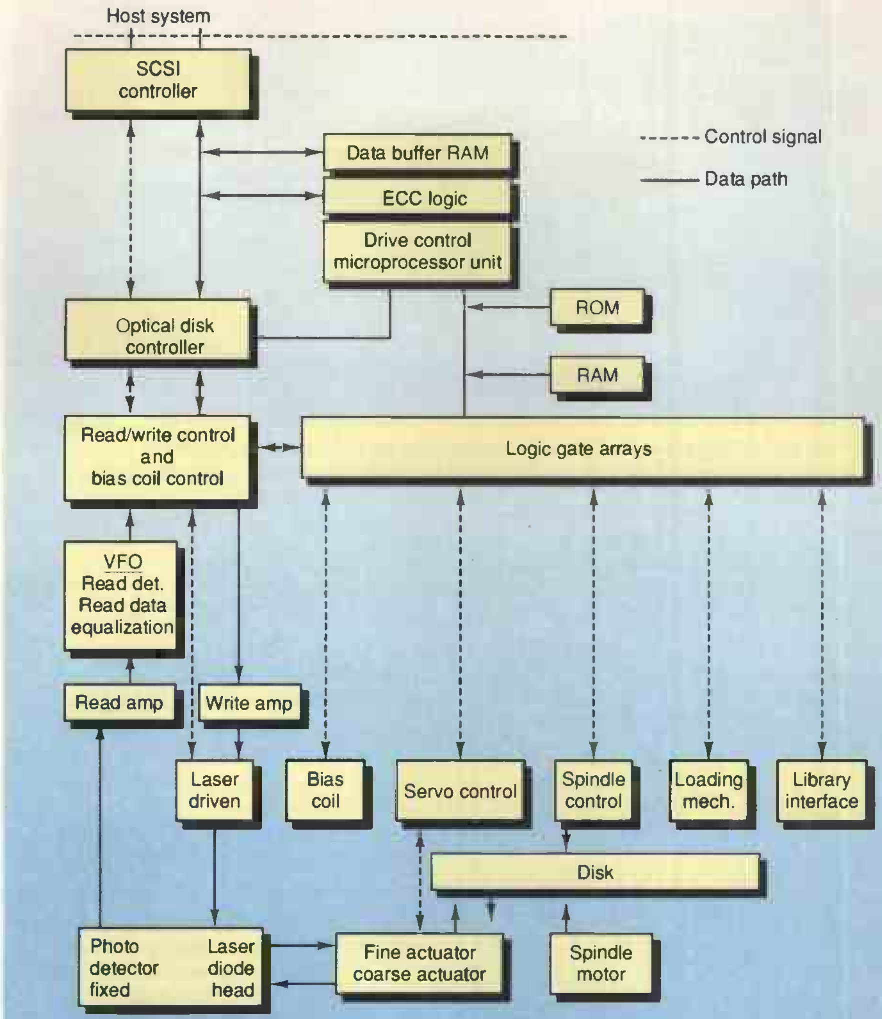


Fig. 3. Key electrical functions and interfaces of an optical drive. The VFO is the variable-field oscillator used to synchronise data.

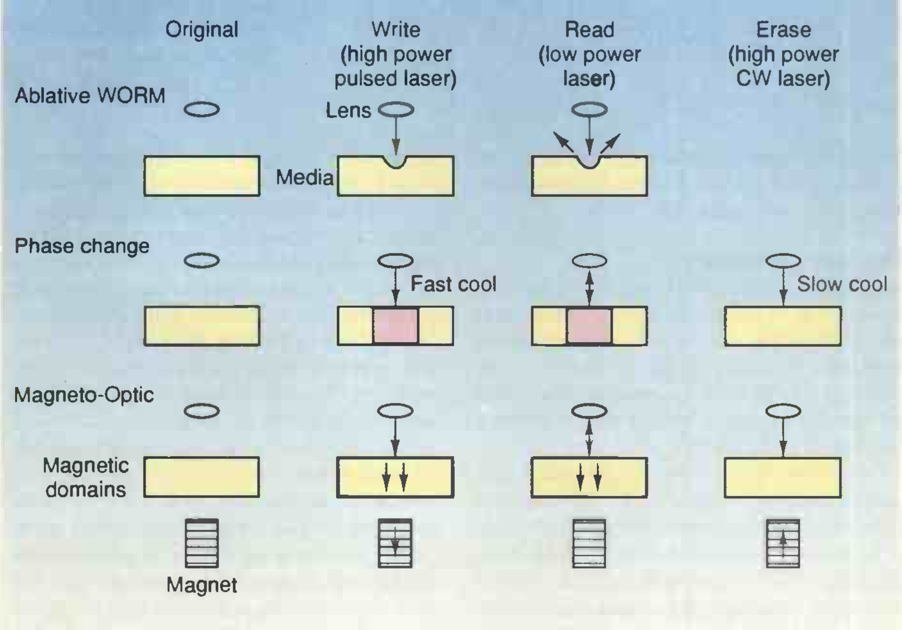


Fig. 4. Optical recording techniques compared. The primary difference is the material used. A magneto-optical drive could record on phase-change or WORM media too.

on a sliding pin or elastic flexures. Critical factors in the design are range of motion, acceleration, freedom from resonances, and thermal considerations.

As mentioned earlier, the tracking mechanism consists of a coarse and a fine actuator. In high performance drives, the coarse actuator consists of a linear voice coil motor driving a rail mounted carriage while the fine actuator acts to produce small radial displacements of the laser spot on the disk. The compound tracking actuator configuration has advantages over a single actuator not only in track following performance, but also when seeking between tracks.

I have discussed the servo and actuator technology required to ensure proper focus as well as the track following and seeking operations of the laser stylus. The optical head, servo, and actuators cover the essential opto-mechanical part of the drive. Next I will consider the formatting, recording, and reading out of data.

The data channel and SCSI

Most optical drives connect to host computer systems using the Small Computer Systems Interface, or SCSI. A schematic block diagram of the functions in an optical drive, based on the IBM 0632 CHA model. This 1.3Gbyte half-high optical drive is shown in **Fig. 3**.

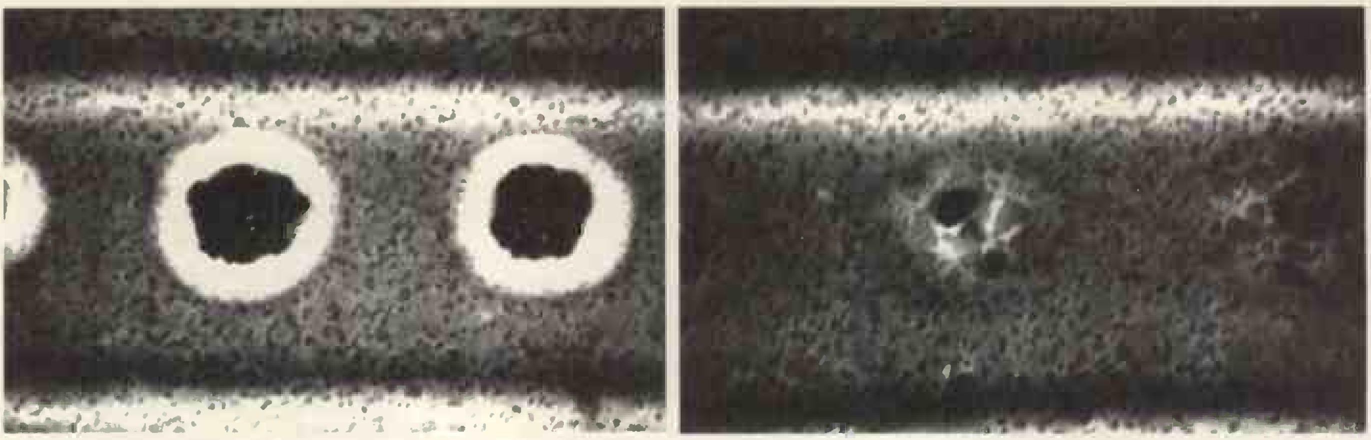
The SCSI controller handles the flow of information to and from the host – including commands. It also provides arbitration and disconnect/reconnect functions. Through the logic gate arrays, the drive control microprocessor unit controls all the functions of the optical drive. These include,

- servo control
- spindle motor spin up/down
- actuators
- laser driver
- magnetic bias coil – an electromagnet used in magneto-optical recording
- loading mechanism for disk load/unload
- library interface providing control lines useful in a jukebox environment.

The rom is the control storage for the microprocessor while the ram provides the microprocessor working storage. The optical disk controller, odc, is a key controller of the data path. It transfers commands from the SCSI controller to the microprocessor for interpretation. In addition it provides handshake lines to channel data appropriately through the buffer ram to the write channel, or from the read channel to the SCSI output (through the ram buffer).

Buffer ram varies in size from 1Mbyte to 4Mbyte and provides temporary storage of data read out or data to be written. Used appropriately, the buffer ram can enhance performance of the drive by providing read-ahead cache or segmented write-cache capabilities.

Data input to the drive over the SCSI for recording is first broken up into fixed block sizes of, for example, 512Kbyte or 1024Kbyte length. It is then stored in the data buffer ram.



Source: IBM Corporation

Fig. 5. A good optical recording medium must have a sharp threshold for the onset of writing. This ensures that a lower power beam can read out the information without affecting the written information. Shown is the threshold property of a commercial ablative WORM medium. At 8mW, above, perfect marks are formed, but at 7mW of writing power, right, there is hardly any marking. This means that a read power of 1mW should be perfectly safe.

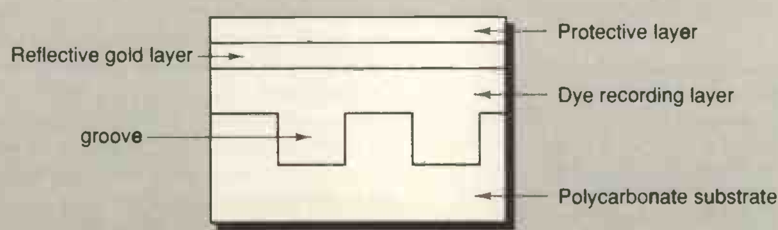


Fig. 6. Basic structure of a recordable CD-R disk. The groove is for tracking and timing information. A cd-rom drive reading this disk cannot sense the groove.

Magneto-optic, phase-change and worm drives can be classified as fixed block architecture technologies in which data blocks are recorded much like in hard drives (Marchant, 1990). Blocks of data can be placed anywhere on the disk in any sequence. The current cd recordable, or CD-R, drive is, on the other hand, an example of a non-fixed block architecture (because its roots are in cd-audio). In current recordable cd drives, input data is recorded sequentially, like a tape player, and can be of any continuous length.

Error correction and control, ecc, bytes are added to each block of data. Optical drives use Reed-Solomon codes which are able to reduce the error rate from 1 in 10^5 to about 1 in 10^{13}. Error-correction encoding and decoding is managed by the optical disk controller.

To extract ones and zeros from the noisy analogue signal from the photodetectors, optical drives use a number of techniques such as equalisation which boosts the high frequencies and thus provides greater discrimination between spots. Using an analogue-to-digital converter, the analogue data signal is converted into channel bits. These channel bits are converted back into customer data bytes using basically the reverse of the encoding process.

Clocking of data coming off the disk is provided by the variable field oscillator, or vfo. Data is clocked into the decoder, which removes the modulation code. Remaining special characters are removed from the data which is then fed into the forty-byte ecc alignment buffer to correct any errors. Once data has been read from the disk, it is stored in a ram buffer and then output to whatever readout device is hooked by SCSI to the drive.

Having outlined how data is recorded on a spinning disk, I will now turn to specific topics such as recording physics that delineate the various recording technologies, **Fig. 4**.

Phase-change recording

Phase-change recording takes advantage of the fact that certain materials can exist in multiple metastable – i.e. normally stable – crystalline phases. Each of these phases has differing optical properties, such as reflectivity. Thermal energy, as supplied by the focused beam of a high power laser, above some threshold can be used to switch from one metastable state to another.

Energy below the switching threshold should have no effect. In this way a low power focused spot can be used to read out the recorded information without affecting it. In any optical recording system, it is critical to have a sharp threshold for the onset of writing in any recording technology to ensure that readout can be performed without degradation of recorded marks. **Figure 5** is an example of the sharp threshold for writing.

To achieve this kind of multiple metastable states, phase change materials typically are a mixture of several elements such as germanium, tellurium and antimony ($Ge_2Sb_2Te_5$). In an erasable material, recording is affected by melting the material under the focused spot and then cooling it quickly enough to freeze it in an amorphous phase. Rapid cooling is critical, so the design of the heat sinking capability of the material is important. Erasing of phase change material is achieved by an annealing process. This involves heating the material to just below the melting point for a long enough period to recrystallise the material and erase any amorphous marks.

Phase-change drives are simpler than magneto-optical drives. They need less complicated optical heads and do not need a bias magnet. However, most rewritable optical drives are based on magneto-optical technology. This is largely because early phase change disks had very limited number of overwrite cycles, of the order of a thousand, while magneto-optic disks were shown to have a million overwrite cycles.

Phase change technology has come a long way since then, even achieving on the order of 100,000 overwrite cycles.

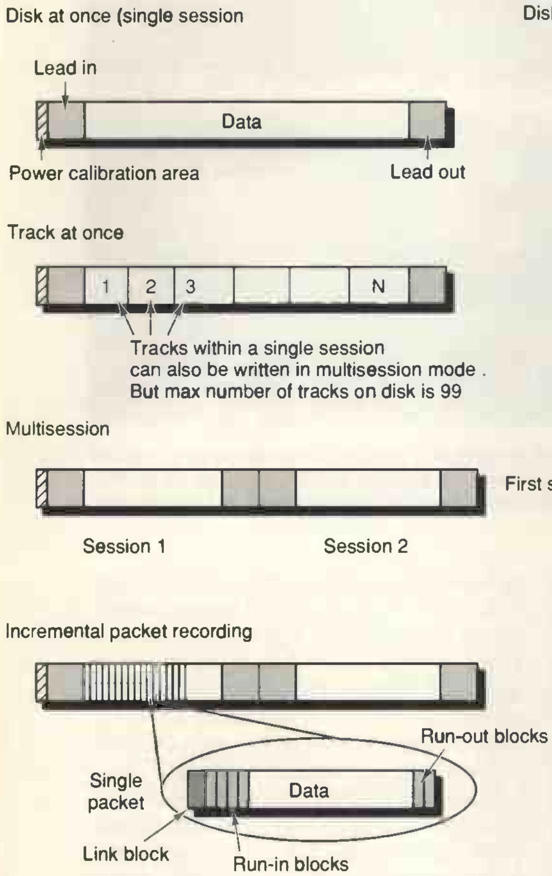
WORM technology

Write-once-read-many technology has a clear place in data storage because it allows permanent archiving capability. Neither magnetic disk storage, nor magnetic tape storage can provide similar write-once capabilities. There are at least four different types of optical write once technologies that are found in commercial products: ablative, moths-eye, phase-change, and dye-polymer.

Ablative WORM disks consist of tellurium based alloys. Data is written using a high power laser to burn a hole in the material. IBM offers 5.25in drives with current capacities of 1.3Gbyte that use this type of WORM technology.

A second type of WORM material is what is known as textured material, such as in a moth's eye pattern. The material is usually a platinum film. Writing is accomplished by melting the textured film to a smooth film, which changes the reflectivity.

Phase-change technology provides a third type of WORM technology using materials such as tellurium oxide. In the writing process, amorphous (dark) material is converted to crystalline (light) material by applying heat



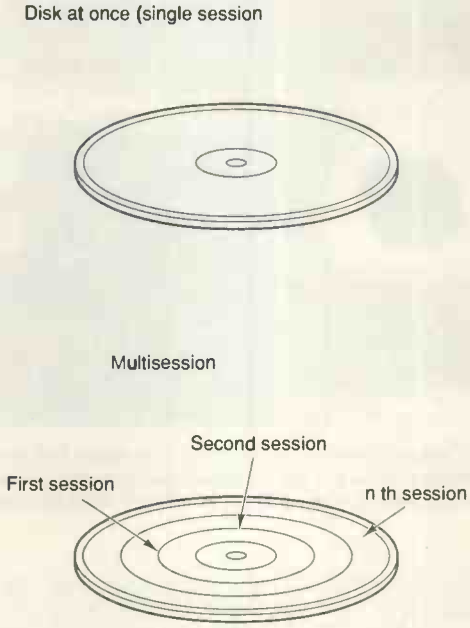


Fig. 7. Recording modes available for recordable cd technology. Incremental packet recording increases flexibility of recordable cd and makes it more suitable as a mass storage device for the desk-top. In packet recording, unlike other modes, there is no limit to the number of recordings – provided that there is space on the disk.

from a focused laser beam. The change cannot be reversed. Dye-polymer media, the fourth type of WORM media, are used in recordable cd drives and is discussed in more detail later.

Magneto-optic technology

In a magneto-optical drive, data recording is achieved through a thermo-magnetic process. This process relies on the threshold properties of the Curie temperature of magnetic materials. Energy within the focused optical spot heats the recording material past its Curie point of about 200°C, a threshold above which the magnetic domains of the material are susceptible to external magnetic fields of the order of 300 gauss.

The external magnetic field is used to set the state of the magnetisation vector in the heated region to either 'up' (a one bit) or 'down' (a zero bit). This vector represents the polarisation of the magnetic domains. When the material cools to below the Curie point this orientation of the magnetic domains is fixed.

This recording cycle has been shown to be highly repeatable – over more than a million cycles – in any given region without degradation of the material.

Magneto-optical disks can be safely read by low power laser beams of about 2mW at the disk surface. This is because the coercivity of a magneto-optic material remains high until very close to the Curie temperature. Near the Curie temperature, at about 200°C, coercivity rapidly drops by two or three orders of magnitude as the magnetic domain structure becomes disordered. Until the Curie point, it is not affected by magnetic fields or laser light.

During readout, the recorded ones and zeros are sensed by a low power linearly polarised readout beam and by utilising the polar Kerr effect. In this effect the plane of polarisation of the light beam is rotated by 0.5° or so by the magnetic vector. The direction of rotation, which defines whether the bit is a one or a zero is converted by the polarisation optics into an intensity change which is sensed by the readout detectors and channel.

Although the tiny amount of Kerr rotation results in a very small amount of signal modulation riding on a large dc bias, the technique of differential detection permits acceptable signal-to-noise ratio (snr) to be achieved.

Recordable compact disk – CD-R

The writable - i.e. write once – version of the popular cd-rom is known as CD-R and was introduced about four years ago. A CD-R disk can store about 650Mbyte of data. A recorded disk looks very much like a stamped cd-rom, and is playable in most cd-rom players.

Early CD-R drives were very expensive – of the order of $50,000 each – and were used only by professionals mastering cd-rom disks. Over the next three years, CD-R drive prices fell quickly to around $10,000 in 1993.

By the end of 1994, the drive price had fallen to less than $2000, with blank disks costing about $12 per disk in the shops. The OEM price for CD-R drives has already fallen to less than $1000 for small quantities. The dramatic drop in prices, combined with the fact that recordable disks were compatible with cd-rom players has created a great deal of interest in this technology.

A recordable cd is coated with an organic polymer that can change its local reflectivity permanently upon sufficient heating by a laser spot. Structure of a CD-R disk is shown in **Fig. 6**. When the organic dye polymer is locally heated by the focused spot of a laser beam, polymeric bonds are broken or altered resulting in a change in the complex refractive index within the region. This refractive index change results in a change in the material reflectivity. There are half a dozen organic dye polymers that are commercially being used. Two examples are phthalocyanine and polymethane cyanine.

Like cd-rom drives, CD-R drives have relatively low performance compared with optical or hard drives. Just as in a cd-rom drive, the seek times are on the order of a few hundred milliseconds while the maximum data rate for a quad-speed drive is about 600Kbyte/s.

Seek time is slow because recordable drives spin the disks in constant linear velocity (clv) mode as defined in the Red Book standards. Constant linear velocity means that the disk rotation speed varies with the radius at which the read head is positioned in such a way as to ensure that the linear velocity is constant with radius. In contrast, constant angular velocity devices like optical WORM disks have seek times on the order of 40ms.

Recording CD-Rs

To understand the attributes and limitations of recordable cd, it is important to understand the various recording modes that it can operate in. For fixed block architecture devices such as magneto-optical drives, the question of recording modes never comes up as there is only one mode, but in recordable cd, there are four modes, as in **Fig. 7**.

The four recording methods in CD-R drives are: disk-at-once, or single session, track-at-once, multisession, and incremental packet recording. In disk-at-once recording, one recording session is allowed on the disk, whether it fills up the whole disk or just a fraction of the disk. The data area in a single session disk consists of a lead-in track, the data field, and a lead out track. Information such as the table of contents is within the track lead-in.

In single-session writing, once the lead-in and lead-out areas are written, the disk is considered 'finalised'. Even if there are blank areas on the disk, further recording cannot take place. After the disk is finalised – and only then – it can be played back on a cd-rom player, which needs the lead-in and lead-out tracks present just to read the disk.

Having only the capability of recording a single session can be a quite a limitation for obvious reasons, so the concept of multisession recording was introduced. An early proponent of multisession recording was Kodak which wanted multisession capability for its PhotoCD products . In multisession recording, each session is recorded with its own lead-in and lead-out areas.

Multisession recorded disks can be played back in cd-rom drives that are marked multisession compatible, assuming that each session on the disk has been finalised with lead-in and lead-out areas. Unfortunately, the lead-in and lead-out areas for each session take up lots of overhead; about 15Mbyte. With this kind of overhead, the ultimate maximum number of sessions that can be recorded on a 650Mbyte disk is 45 sessions.

Rather than do multisession recording, the user may opt for track-at-once recording. With this technique, one or more tracks can be written in each session. The maximum number of tracks that can be written on the disk is 99. However the disk or session must be finalised before it can be read on a cd-rom drive.

Because of the way input data is encoded and spread out, it is imperative to maintain a constant stream of information when recording. If there is an interruption in the data stream, it affects the whole file being recorded – not just a sector as in magneto-optical or WORM drives. If the interruption is long enough, it will usually lead to the disk being rendered useless – a 'golden coaster' as it is referred to in the industry. For this reason, it is important to have a fast hard drive capable of feeding data to the drive buffer continuously.

It is inconvenient to use the hard drive of the personal computer to directly feed the CD-R drive, because it ties up the hard drive and basically the whole computer. To address this issue, some CD-R drives come in a package that includes a dedicated 1Gbyte hard drive. This solution, however, raises the cost of using recordable cd.

Many of the above problems or inconveniences can be alleviated through a new recording method just being introduced. Called incremental packet recording, this method involves breaking the input data up into packets of specified size, for example 128Kbyte or 1Mbyte.

Each packet consists of a link block, four run-in blocks, the data area, and two run-out blocks. Run-in and run-out blocks help delineate packets and allow some room for 'stitching' – i.e., providing overlap if perfect synchronisation is not achieved when recording an adjacent packet in a different recordable cd drive.

Packet recording has several advantages. To begin with there is no limit to the number of packets that can be recorded, up to the space available on the disk of course. In this way, limitations imposed by track-at-once, multisession or disk-at-once can be avoided.

Secondly, if the packet size is smaller than the drive-buffer size – as is likely to be the case – a dedicated hard drive is not needed while recording. Once the packet of information has been transferred to the drive buffer, the computer can do other tasks while the CD-R drive carries out the recording.

With the advent of packet recording, recordable cd technology becomes much more flexible than in the past. As a result, the technology is much more attractive as a general purpose removable data storage device. It can be used for back up purposes as well as for storing smaller files.

Table 1. Optical technologies compared.

	Multifunction Optical 5.25in MO/WORM	3.5in Optical	CD-R/CD-E	Phasewriter Dual (PD)
Disk diameter	130 mm	90 mm	120 mm	120 mm
On-line capacity	650Mbyte[1]	230Mbyte	650Mbyte	650Mbyte
Function	Rewritable and write-once	Rewritable	Write once but CD-E will be rewritable	Rewritable
Performance				
read data rate	1200-2300KB/s[2]	600-1300KB/s	700KB/s	870 KB/s
write data rate	400-800KB/s (mag-opt) 660-1200KB/s (WORM)	200-400KB/s	700KB/s	870 KB/s
seek time	30ms	35ms	300ms	200ms (average)
Mean time between failure (MTBF)	180 000 hours	40 000 hours	30 000 hours	–
Current drive price	$2000	$700	$1700	$1000
Current media price	$100	$30	$12	$100
Expected drive price (May '96)	$1500[3]	$600	$700	$700
Key attribute	High performance	Portability (shirt-pocket size disks)	Can read cd-rom Readable in cd drive	Can read cd-roms
Migration path	≥1300MB (early '96)	≥650MB (early '96)	≥3GByte (late '97)	not clear
Upward compatible with next generation drives?	Yes (read/write compatible)	Yes (read/write compatible)	not clear (current CD-R disks not readable in HDCD drives)	not clear

Notes:
1) On-line capacity means capacity available without human intervention – i.e. capacity on a single side.
2) In constant angular velocity (CAV) drives, linear velocity varies as a function of radius. Thus, data rate varies as a function of the radius of the read/write head because the linear data density is constant as a function of radius. The higher data rate is possible when the head is at he outer radius of the disk, while the lower data rate occurs when the head is at the inner radius.
3) These are projections based on industry consultant forecasts.

There is an interchangeability problem with some cd-rom players as they post a hard error when they encounter the link block at the beginning of each packet. To address the problem of interchange with packet written disks, the Optical Storage Technology Association is investigating whether an appropriate device-driver utility will enable older cd-rom drives to accept packet written disks.

An alternative is to use variable sized packets in which the packet size is equal to the file size. This technique is attractive because it reduces the incompatibility with cd-rom drives, but it also has its own problems – further illustrating the point that converting a recordable cd into a mass storage technology will require a certain amount of compromise and reduced expectations.

Cd-rom compatibility issues

One of the key attributes of recordable cd drives is that in principle the disks they write can be read by cd-rom drives. Given the phenomenal success of cd-rom and the rapidly growing installed base of titles and drives, it is clearly advantageous for an optical drive to be cd-rom compatible in some way.

Recordable cd drives can read cd-rom disks, and write recordable cd disks that can be read on cd-rom drives. However these disks are write-once only.

Several companies are working on a new technology called CD-E or compact-disk-erasable which is a recordable cd based on phase-change material. The CD-E drives will supersede CD-R drives since they will have not only all the features of the recordable cd drives but also those of erasable cds. CD-E drives have the potential to become a mass storage medium and to replace the floppy drive on desk top computers.

Recognising the importance of cd-rom compatibility, Matsushita Corporation (Panasonic) recently announced a new rewritable optical drive called phasewriter dual or 'PD'. This drive has the capability to read cd roms. The phasewriter-dual drive writes to a phase change rewritable media. However, unlike CD-E drives, the disks written by the pd drive cannot be read back by cd-rom drives, and thus cannot take advantage of the installed base of cd-rom drives.

A bewildering array of choice in 1995

How do optical drives available now, or being introduced in 1995, compare with each other? There seems to be a bewildering array of choices for the consumer. Which should the consumer choose? Of course this depends on what is important to the user – capacity and performance or low cost or cd-rom compatibility. A comparison of the various technologies are given in **Table 1**.

The primary delineators are performance and cost. Magneto-optical/WORM and phase change drives offer much higher performance than recordable cd drives. On the other hand, recordable cd offers cd-rom compatibility and has a greater possibility of being low cost because of the cd-rom base.

Over the next three years, the market will decide in which application segments each of these features is most important.

The next generation: HDCD and DVD

The 650Mbyte capacity of current cd-rom and recordable CD-R drives has remained unchanged since the introduction of cd-rom in 1985. However, the growing interest in putting video on cd is forcing the need to increase the capacity – video is very storage intensive.

The motivation for developing video cds is that they will replace the video tape, just as audio cds replaced the phonograph record and audio tape in the mid eighties. Furthermore the high-capacity cd is also needed in computer applications to replace cd-rom disks.

In december 1994, Philips and Sony proposed a new compact disk standard called HDCD, high-density cd, which can hold

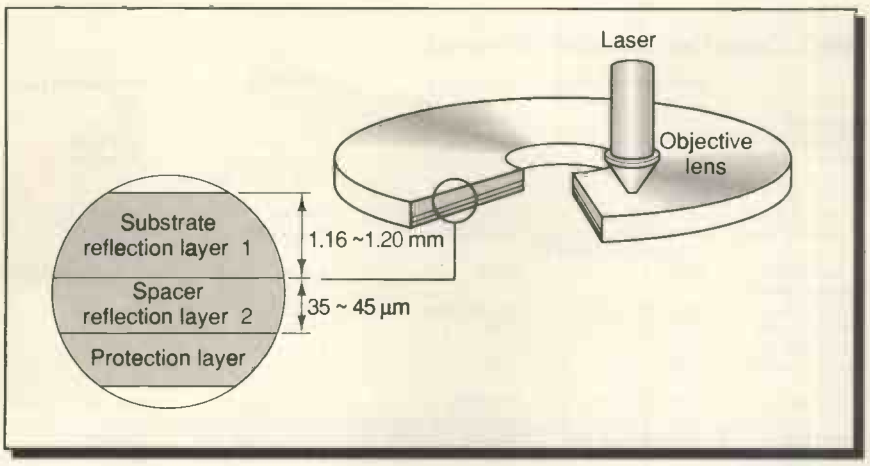


Fig. 8. Announced earlier this year, Sony's new 3.7Gbyte high-density compact disk is claimed to be capable of recording up to 135 minutes of MPEG-2 compressed video.

3.7Gbyte on a single disk. A second version of the HDCD is likely to have two data layers, as shown in **Fig. 8**, thus doubling the capacity on a single platter to about 7Gbyte.

A two layer HDCD should cost only marginally more than a single layer HDCD. Not to be outdone, Toshiba and Time Warner with support from Matsushita released an alternative standard for a videoCD in January 1995. Toshiba calls it the DVD or digital video disk but it has also been referred to as SD or super disk. This disk is double sided and can store about 5Gbyte per side. A comparison of the two mutually incompatible standards is given in **Table 2**.

How will the increases to HDCD or DVD capacity be achieved? In any disk based system the main way of increasing capacity is to make the marks smaller and put them closer together since there is no chance of increasing the size of the disk. To go from current cd-rom capacities to HDCD or DVD capacities – a factor of at least five jump in storage density – requires shorter wavelength lasers, higher numerical aperture lenses, tighter track pitches and higher linear densities.

The laser chosen for the next generation cds will be a red emitting type operating at 635nm, as opposed to the 780nm of currently used for cds. However the switch to a red laser leads to a potential problem in compatibility. This is because current dye-polymer media formulations for recordable cds are not compatible at red laser wavelengths. As a result, the recordable disks you are using today will not be read back by the next generation of higher density compact disk drives. Lack of upward compatibility, though not widely advertised, will not be warmly greeted by any users who are currently archiving data or photos on recordable cd media.

Several US and Japanese media companies are working on recordable cd media that is compatible at both the infra-red laser wavelengths, found in current drives, and the red laser wavelengths. An introduction of such media into the commercial market is imminent and will solve the current upwards-compatibility problem.

The fact that two competing proposals have been introduced for the videoCD standard has caused quite a stir, bringing back memories of the VHS versus Betamax videocassette standard wars. The videoCD standards could well be a replay of the videocassette war unless the two sides get together and compromise on a single standard.

Currently in Toshiba/Time Warner's favour is the fact that several of the Hollywood studios – the 'content providers' – seem to be supporting DVD. Sony and Philips have been lining up the major computer and software makers to support HDCD as the next generation replacement to the immensely popular and widespread cd-rom.Realising the importance of cd rom, Toshiba has agreed to modify its standard so that it will support cd rom, and has begun talks with computer and software companies, Which of the two competing standards the industry will adopt is still up ion the air, but hopefully, Sony/Philips and Toshiba can sort out their differences and converge to a single standard.

Optical technology trends

Whichever specific format is adopted, optical drives will continue to see both incremental and radical improvements from a technology perspective. The incremental improvement process will concentrate on the four core technology elements: the laser, the media, the recording channel, and the opto-mechanics.

The laser will see continual improvements in its operating life and power, which in turn will allow disks to spin faster. Media will see improvements in substrates and active layers. Optics and actuator systems will have improved servo systems to allow finer positioning. In addition there will be reductions in noise, smaller optical components, and lighter actuators for faster seek operations.

The recording channel and electronics will see an increase in the level of electronics integration, a reduction in electronic noise and power consumption, and better signal processing and error correction.

There is a considerable amount of technical growth potential for optical drives, in both performance and capacity. After all, optical drives are a relatively new commercial technology. Remember that the first rewritable optical drives only started being marketed about six years ago.

Acknowledgments

Thanks to Lee Jesinowski of IBM Tucson for the electronic block diagram of the optical drive, and to Blair Finkelstein of IBM Tucson for the photograph on WORM threshold mark formation. ■

Probing further

There is a number of excellent books that provide an overview of optical disk systems. A classic is *Optical Recording* by Alan Marchant (Addison Wesley, Reading, MA, 1990) which provides an overview of the various types of recording as well as the basic functioning of an optical drive. A more detailed study of optical disk drives and their opto-mechanical aspects is provided in *Principles of Optical Disc Systems* by G. Bouwhuis, J. Braat, A. Huijser, J. Pasman, G. van Rosmalen, and K. Schouhamer Immink (Adam Hilger Ltd, Bristol 1985).

For recent developments in the field of optical storage, you are advised to attend the meetings of the International Symposium on Optical Memory (ISOM) or the Optical Data Storage (ODS) Conferences (held under the auspices of the IEEE or the Optical Society of America).

Table 2. Comparison of proposed high density compact video disk standards – Sony/Philips' IDCD versus Toshiba/Panasonic's DVD.

	Sony/Philips HDCD	Toshiba/Panasonic DVD
Number of sides	Single sided (but up to 2 data layers)	Double sided
Disk capacity.	3.7GB (7.4GB for 2 layer)	5GB/side (10GB total)
Playing time	135min/layer	142min/side
Track pitch	0.84µm	0.725µm
Laser Wavelength	635nm	635nm
Numerical aperture	0.52	0.6
Plays current cd-rom titles	Yes	not clear

POWER dissipation in capacitors

Perfect capacitors dissipate no power, but in the real world, many factors combine to reduce efficiency and increase failure rate – as capacitor consultant Cyril Bateman explains.

Capacitors don't take power. So began my lecturer when introducing the topic of capacitor phase angles. I remembered these words only too clearly some years later when investigating capacitor failures occurring in the line time-base circuit of the first all solid state 110° colour televisions.

My experiences as a capacitor designer and applications engineer have clearly demonstrated that all capacitors have a limited power handling capability, similar to the safe operating area of semiconductors.

Directly or indirectly, overstressed capacitors are involved in most circuit failures. Obviously, all components wear out eventually. But overstressed capacitors used with pulse waveforms in power switching circuits can fail very quickly. Worse still, before the capacitor ultimately fails, it can directly contribute to the failure of switching semiconductors, and in doing so, mask the prime failure mechanism.

Manufacturers sometimes determine the power rating of a capacitor by subjecting samples to sinewave stress while monitoring the temperature rise. To confirm long term reliability, this is often supported by stressing at elevated temperature. These results can be related to end-use conditions provided that the capacitor's rms power level in circuit can be measured or calculated.

Why should capacitors cause such problems used with pulse waveforms? Power dissipated in a capacitor is dependant on $I^2 esr$, alternatively $VI.\tan\delta$. While capacitive reactance is inversely proportional to frequency, equivalent series resistance, or esr, is not. Depending on frequency and capacitor type, while esr generally reduces with rising frequency, this is not always the case. In some combinations esr can exceed the capacitive reactance, and can also increase with frequency.

Since the esr of a capacitor is frequency dependent, the capacitor's power dissipation can be measured in circuit only when using sinewave stimulation. Given a mathematically defined waveform, however, power level can be calculated following the classical methods.

This article proposes a method of calculating capacitor power dissipation for any waveform, – demonstrated by two recent applications reported in *Electronics World* – together with a prototype 'snubber' circuit[1,2,3].

In the first of these applications, an article on Cuk converters, Finnegan found that using a standard electrolytic component for the 3.3μF capacitor of **Fig. 1** was unsuitable. Equivalent series resistance caused overheating. In the second article, physically small capacitors are needed for C_1 and C_2 of **Fig. 2**. The choice of ceramic type matters here since some exhibit more losses than others and an incorrect type could easily result in overheating.

According to the Fourier theorem, time and frequency domains interrelate and can be transformed with no loss of information – provided complete waves are used and sufficient harmonics are computed. In theory an infinite number of harmonics is required. In practice fifty harmonics have proved to suffice.

Converting from the time domain into the

Resistance in capacitors

For a capacitor, equivalent series resistance, esr, is a single lumped resistive value representing all real losses. This loss comprises three main sources:

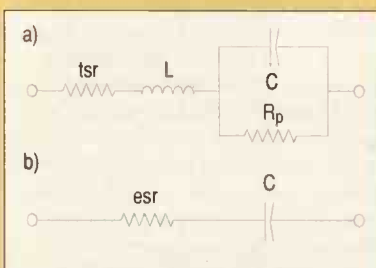
- True series resistance, tsr, comprising the actual metallic resistances.
- R_p comprising:
  a) Dielectric loss due to molecular and interfacial polarisation.
  b) Leakage resistance, measured at dc volts.

$$esr = tsr + \frac{R_p}{1+\omega^2 (R_p)^2 C^2}$$

$$esr = \tan\delta \times X_c$$

$$esr = \cos\theta \times |Z|$$

Equivalent series resistance tends to reduce with increasing frequency, but by considerably less than the theoretical halving for each doubling of frequency. Ultimately attaining a minimum value when $X_c = X_L$, i.e. at the series resonance of the capacitor as a series *LCR* system.[8,9,10]



Capacitor equivalent circuit - (a)
Lumped value series esr - (b)

In the capacitor equivalent circuit a), true series resistance is caused by actual metallic resistances inherent in the component make-up. Solving this equivalent circuit mathematically into real and imaginary terms results in the series equivalent model b). This is an equivalent series resistance representing all real capacitor losses.

frequency domain for calculations offers many benefits. Not least of these is the ease with which frequency dependent parameters can be accommodated, since all calculations are also simplified.

With capacitors subjected to non-sinusoidal waveforms, the resulting capacitor current by harmonic depends on the harmonic amplitude multiplied by the harmonic number. Given an ideal capacitor and ideal square wave, the current resulting from each harmonic equals that of the fundamental frequency. For other waveforms it is possible that harmonic currents can exceed that of the fundamental. Since the equivalent series resistance change by harmonic is always less than ideal, the power contributed by a harmonic can considerably exceed that of the fundamental, **Table 1**.

Having established the need to avoid early capacitor failures, the sequence for calculating capacitor power dissipation is:

- Fourier transformation of the stimulus waveform observed across the capacitor terminals, into the frequency domain, by amplitude and phase.
- Determination of esr for each harmonic frequency, by measurement or from published characteristics.
- Determination of relative capacitor current and phase for each harmonic frequency.
- Complex multiplication of stimulus waveform harmonic amplitude and phase with capacitor current and phase for each harmonic frequency.
- Reverse Fourier transformation (synthesis) back into time domain.
- Calculation of capacitor power by each harmonic and relevant esr, over one period of the fundamental waveform.
- Calculation of rms power dissipated in the capacitor.

A closer look

Many suitable fast Fourier calculation routines have recently been published. The essential requirements are sufficient data points for accuracy, say 256 minimum, the provision of $k(0)$ (mean) data, with sufficient harmonic data by amplitude and phase, such that reverse transformation accurately recovers the original time-domain voltage waveform's shape and amplitude.

I have successfully used an enhanced version of the Larsen-Dyrik *BBC Computer* program adapted to run on the *Archimedes*[4]

Table 1. These results, derived from the 22nF snubber, show that power loss due to harmonics can exceed that of the fundamental.

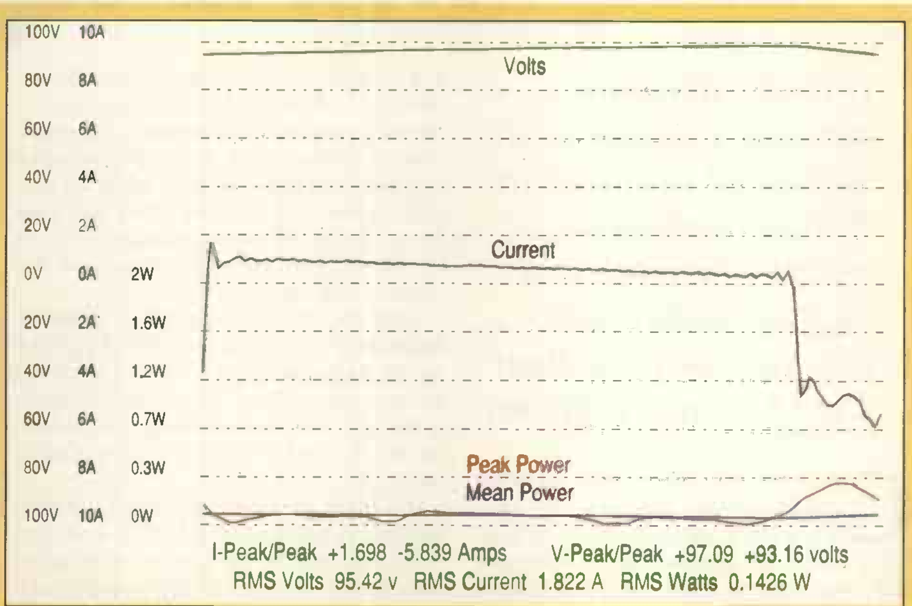
Harmonic number	Amplitude (V)	Phase (°)	Current (A)	Power (W)
$k(0)$ (mean)	113.13	0	0	0
$k(1)$	118.5	236.9	1.64	3.75
$k(2)$	89.72	202.5	2.48	4.87
$k(3)$	53.32	164.5	2.21	2.79
$k(4)$	21.5	112.7	1.19	0.64
$k(5)$	9.72	354.3	0.67	0.18
$k(6)$	14.66	272.4	1.22	0.49
$k(7)$	14.05	215.1	1.36	0.56
$k(8)$	10.7	151.2	1.18	0.38
$k(9)$	8.28	79.05	1.03	0.27
$k(10)$	6.8	6.96	0.94	0.21

Capacitor performance simulations

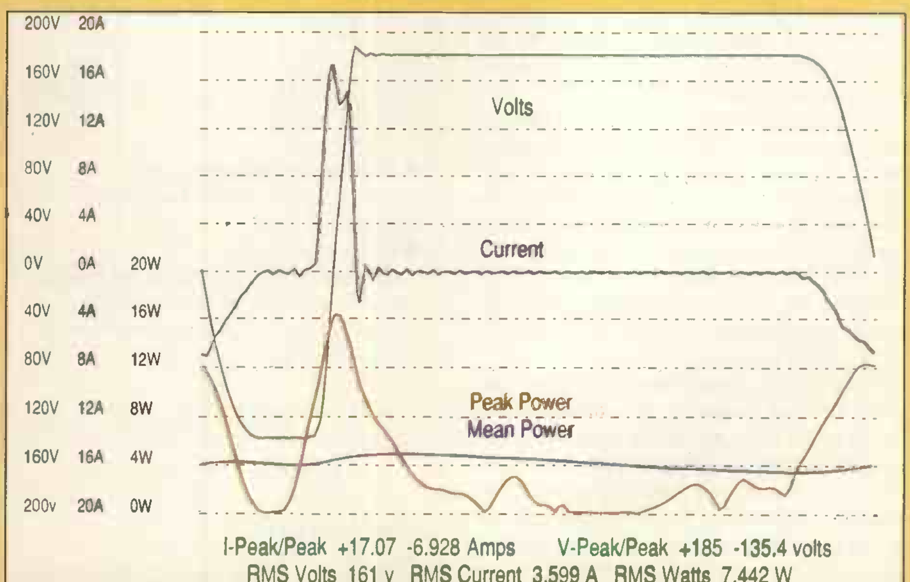
The procedure used to generate these results graphs commenced with time domain simulation of the circuit using *Pspice*. This produced the voltage waveform which would be measured across the capacitor if displayed on an oscilloscope. The 22nF snubber was in fact derived from an oscillogram.

This waveform was digitised into 256 *X Y* co-ordinates and stored on disk using a custom program. The data file became input to calculation programs running on my Archimedes. These are FFT conversion, followed by frequency domain analysis, complex multiply and reverse FFT to restore to time domain.

The results are displayed on screen exactly as shown here. To provide the best resolution, I wrote a dedicated routine to output these curves as a vector datafile in Archimedes Draw format. This permits easy conversion to formats compatible with other machines.



Plot 1. Energy transfer for capacitor C_3 of Cuk converter Fig. 1. This 160V rated component is 3.3μF. A Wima MKS4 metallised polyester gives acceptable power loss for long life.



Plot 2. Experimental snubber using 22nF chip capacitors made from X7R ceramic was unsatisfactory. Capacitor seriously overheated and failed – quickly.

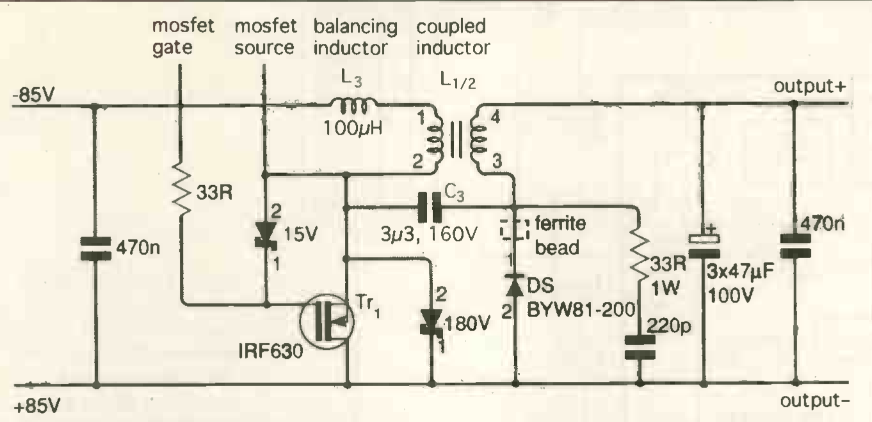
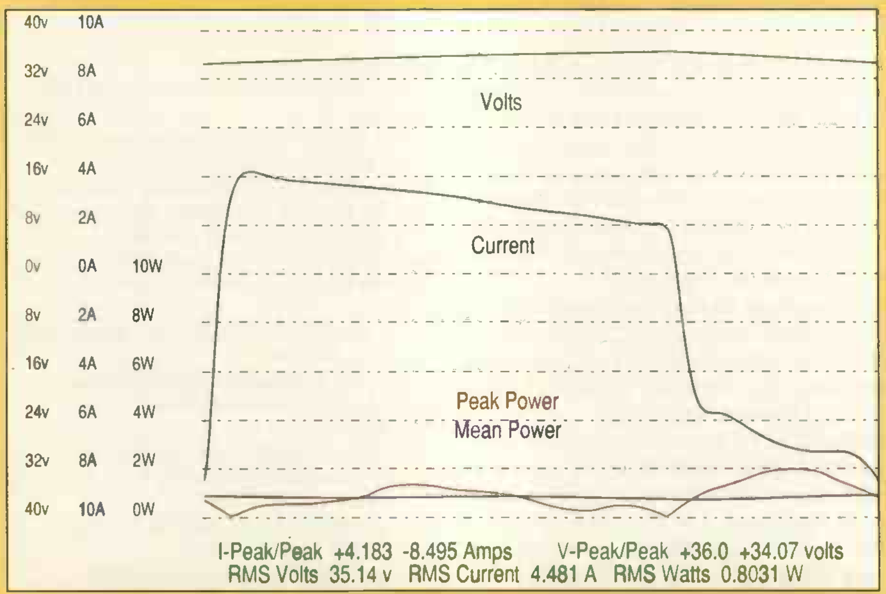
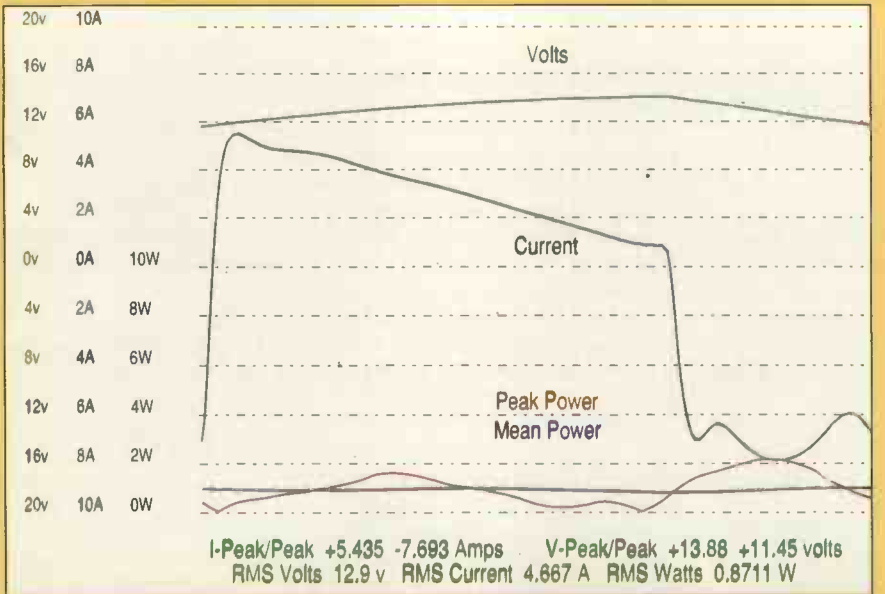


Fig. 1. An important capacitor in this Cuk converter is the 3.3μF device C_3, coupling pins 2 and 3 of the transformer. Finnegan found that ordinary electrolytic capacitors overheated and low-esr types were not available for such high voltages.



Plot 3. Energy transfer for 1.1μF capacitor C_1 of the Harris/IBM converter, Fig. 2. Photograph in Jan '93 EW+WW, p58, shows parallel capacitors make up these values. With careful choice of ceramic dielectric power spread over 7 surface-mount capacitors should be satisfactory.



Plot 4. Performance of 0.88μF reservoir capacitor C_2 in Harris/IBM converter, Fig. 2. Parallel capacitors make up the value, as for plot three, but here, only six devices are needed.

Before any fast Fourier transform can be run, one period of the waveform must be described, mathematically if feasible, or more generally as X and Y co-ordinates, matching the input needs of the chosen FFT calculation.

Fundamental to this method is the assignment of esr and capacitance values for each harmonic frequency used. Whenever possible these should be interpolated from the nearest practical measured frequency.

Modern *LCR* bridges can measure at many frequencies to at least 1MHz. At these higher frequencies the best four-terminal measurement techniques must be used, and preferably with short component lead lengths.

At frequencies greater than are possible with *LCR* bridges, network analysers measuring by $|Z|\angle\theta$ are most appropriate. Since most capacitors of interest will resonate below 10MHz, as an unreducible minimum, a measurement of impedance with frequency can determine the minimal value of equivalent series resistance, at the resonant frequency.

Simulating capacitor performance

Analogue circuit simulators are based either on frequency-domain or time-domain simulation methods. Time-domain based simulators can allow for amplitude dependant anomalies that are typical of semiconductor junctions.

At the University of California, Berkeley, a development grant from Sprague Electric in 1970-71 funded the preliminary development of the Cancer program – from which Spice, subsequently Spice2, developed. This work was specifically targeted to the needs of the integrated-circuits group at the University. Indeed the word Spice derives from 'Simulation Program with Integrated Circuit Emphasis'.

Full details of the development of Spice2 are contained in Laurence Nagel's doctoral thesis, memorandum No. ERL-M520, 9 May 1975.

Spice2-derived simulators comprise small-signal frequency-domain analysis together with time domain transient response and dc transfer functions.

Consequently, although Spice2 based simulators are able to calculate power dissipation, use of the time-domain calculations, inhibits frequency dependency.[11,12]

While many nodal frequency domain simulators do not support frequency dependency, provided matrix reduction techniques are not used, this enhancement is possible.

With non-sinusoidal waveforms, substantial errors of power result from the capacitor's frequency dependent equivalent series resistance. Eliminate these errors by simulating in the frequency domain and using the appropriate esr values. I have built this capability built into my simulators for the Archimedes.

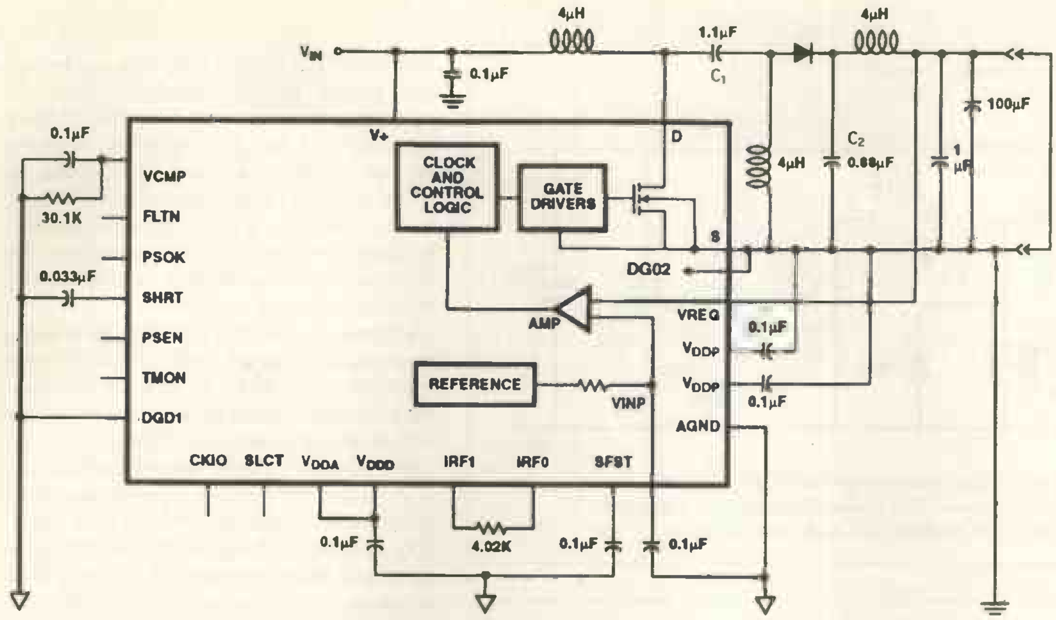


Fig. 2. This single-ended primary-inductance converter, based on a chip produced by Harris and IBM, involves small capacitors which are susceptible to overheating if the wrong ceramic is chosen. In the article here, 36V input and 5V output are assumed.

Provided that the frequency of the highest significant harmonic is below the capacitor self resonance, current and phase by harmonic relative to unity input stimulus can be calculated by many means – including pocket calculator. Should there be any harmonics having significant power above this resonant frequency, then a full capacitor model together with simulation software is preferred. This model has to include inductance and equivalent series resistance.

Complex multiplication of each harmonic component of the stimulus waveform with the relative capacitor current and phase provides the final frequency-domain result. These are output when required. Complex multiplication is simplified if data by magnitude and phase angle is used. In this case multiply the respective magnitudes but add the angles.[5]

Capacitor current in rms can be calculated directly from the above complex multiplied frequency domain results.[6,7]

The reverse fourier transform or synthesis, provides the time domain capacitor current waveform and the rms power dissipated by the capacitor. Additionally an estimate of the peak and mean power levels throughout the waveform can be deduced. This can be used to reduce power levels by fine tuning the waveshape. These capacitor power levels should be calculated by harmonic frequency, amplitude, phase and relevant esr for each of the data points through one waveform.[6]

This proposed method applies not only to switched mode supply simulation. The above sequence is simple and quick, if performed using dedicated computer routines. I have developed some of these myself.

Preparing *X*, *Y* data points defining the stimulus waveform is simplified by using a curve fitting routine. All subsequent data transfers can be automated. For any stimulus waveform, data from the fft can be stored as a library file, and reused as needed, until a different stimulus is desired.

Given capacitor data-file libraries, the traditional method of choosing capacitors by trial and error becomes obsolete. ■

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Fourier transforms – forward and reverse

The forward Fourier transform $F(y)$ of a function of the real variable $f(x)$ (where $f(x)$ may be real or complex) is defined by the integral.[13]

$$F(y)=\int_{-\infty}^{\infty} f(x)\exp(-j2\pi xy)dx$$

In the context of this article, the above expression states that a time-domain measurement can be transformed into the frequency domain with no loss of data.

Displaying the sinusoidal component frequencies which comprised the original time domain waveform by their respective amplitudes and relative phases. An example of a time-domain measurement is an oscilloscope displaying amplitude by time.

This transform is subject to certain conditions. The waveform described must be periodic – e.g. one complete cycle of a repetitive waveform. In addition, the waveform must have a finite average value. The waveform must have a finite number of maxima and minima in a period.

Provided that the time-domain waveform sampling rate satisfies Nyquist and the number of samples is a multiple of two, the transform can be mathematically simplified as defined by the fast fourier transform.[14]

Note that since each harmonic data pair contains two dimensions, the number of data pairs is restricted to one half of the number of samples used for the time domain.

Having performed an fft on the waveform, the resulting data is a number of data pairs describing amplitude and phase by harmonic, Table 1.

Having Fourier-transformed the waveform, subsequent calculations can follow the rules of conventional sinewave analysis.

Results from the reverse Fourier transform can be obtained simply by evaluating, harmonic by harmonic from time t=0 to 2π radians and by the number of samples the following expressions,

where, k is the harmonic number, and ψ is relative phase angle in degrees. Expressions $|Mv(k)|$, $|Pv(k)|$, $|Mc(k)|$ and $|Pc(k)|$ are the magnitude and phase results of the FFT and complex multiply respectively.

$I(k)=(|Mv(k)|\omega kC)\sin(K\omega t+\psi\pi/180+\pi/2)$ (Ref . 4)

$I(k)=(|Mc(k)|\sin(tk+|Pc(k)|\pi/180))$ (Ref. 4-6)

$V(k)=(|Mv(k)|\sin(tk+|Pv(k)|\pi/180))$ (Ref. 4-6)

$Current,\ rms=((|Mc(k1)|/\sqrt{2})^2+(|Mc(k2)|/\sqrt{2})^2\ldots)^{0.5}$ (Ref. 4,10)

$Voltage,\ rms=((|Mv(k1)|/\sqrt{2})^2+(|Mv(k2)|/\sqrt{2})^2\ldots)^{0.5}$ (Ref. 4,10)

$Watts,\ rms=((|Mc(k1)|/\sqrt{2})^2 esr(k1))+((|Mc(k2)|/\sqrt{2})^2 esr(k2))+\ldots$ (Ref. 4)

Phasor diagrams and standing waves

Geoffrey Billington looks at how currents and voltages are distributed in a standing wave and how these can be represented with phasor diagrams.

Phasor diagrams are extremely useful in combining voltages present at a point on a transmission line from forward and reflected waves. Finding the resulting, measureable voltage at a point is made very much simpler with a phasor parallelogram. This is taken a step further with a novel combined voltage and current diagram which allows currents and impedances to be dealt with.

Power considerations

Once steady state has been reached, power is delivered to the line at a steady rate and dissipated in the load at a steady rate. If the line is loss free, these two rates are equal.

Energy injected into the line before steady state is achieved remains there in the form of the forward and reflected waves. In the steady state, energy stored on the line is continually leaking out at one end and is being continually topped up at the other by the transmitter. The energy returned by the reflected wave helps to maintain the forward wave.

When the generator is switched off, the energy stored on the line will leak rapidly away as the waves echo back and forth.

As both the forward and backward waves are energy carriers, the terms 'forward power' and 'reflected power' are frequently used. Reflected power does not mean wasted power so there is no reason why an ideal line should not transmit energy efficiently even if there is appreciable reflection; a high standing wave ratio. On the other hand, using practical transmission lines, the existence of voltage and current maxima is likely to result in increased energy losses. These losses increase with the square of the current in the lines and the square of the voltage between them, so reduced losses at minima will not compensate for increased losses at maxima.

Travelling wave voltages

Imagine a continuous train of periodic waves moving down a long line from left to right. Everyone invariably forms a mental picture of a series of ripples of constant wavelength travelling down the line at a constant speed, but it is important now to think more carefully about what this picture really represents.

To monitor the variation of voltage along the line, let everything happen in slow motion and let centre zero voltmeters be connected at various points, shown in **Fig. 1**. Each meter swings from side to side over the same range as the wave passes, but there is a phase lag which increases steadily with distance moving down the line. For instance, voltmeter V_2 is playing follow-the-leader with V_1 but lags, reaching its maximum a little later. Similarly, V_3 lags V_2, and so on. By the time we reach $V_{\lambda/2}$ the lag has increased to half a cycle, and the lag of V_λ is one full cycle; it has come back into step with V_1. This pattern repeats all along the line. The distance between V_1 and V_λ is defined as one wavelength (λ) or the 'repetition distance'.

"For a wave travelling from left to right, the phase lag produced by moving a distance 'L' *to the right is* L/λ *cycles. Moving a distance to the left produces a lead of* L/λ *cycles."* In other words, a movement of one tenth of a wavelength causes a phase shift of one tenth of a cycle. For waves moving in the opposite direction, the above statements apply by simply interchanging the words 'lag' and 'lead'.

Instantaneous and rms measurements

From now on, phrases such as 'the voltage' or 'the current' will normally be referring to rms values, not instantaneous ones.

Standing-wave maxima and minima

At any point on a mismatched line there are two alternating voltage components. One is due to the forward wave and one is due to the reflected wave. What we actually measure with a voltmeter is the resultant of these two components. This depends on the phase difference between the component voltages at that point which varies from point to point.

Providing the line is long enough there will be at least one point where the component voltages are in phase or, strictly speaking, where the phase difference is a whole number of cycles. Suppose P is such a point. If the for-

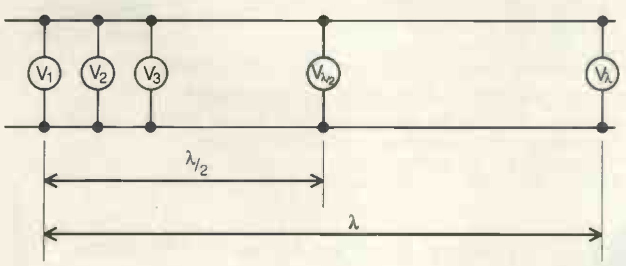


Fig. 1. If the system could be run in slow motion, voltages measured along a transmission line would reveal a follow-the-leader pattern. Travelling along the line to the right, the phase lags that of the previous voltmeter.

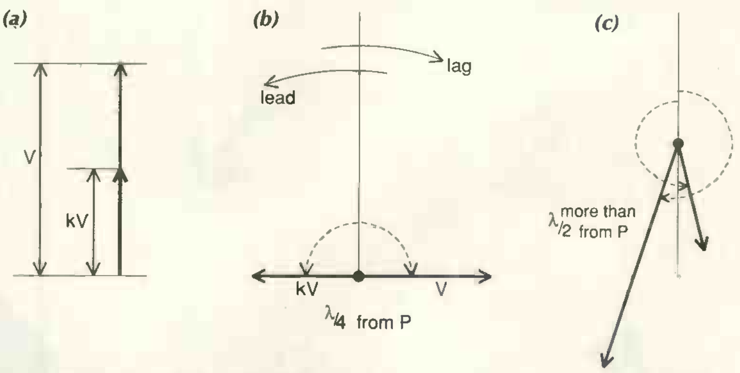


Fig. 2. The relationship between maxima and minima is represented by phasor lengths, (a), and phase by an angle of rotation, (b). Phasors rotate around 360° and cross over as the measuring position on the line moves by more than a full wavelength, (c).

ward-wave voltage component is v and the reflected component is kv, the resultant voltage will have its greatest possible value of $v+kv=v(1+k)$.

Now imagine we move down the line towards the antenna. As the distance increases, the forward wave is delayed an increasing amount while the reflected wave arrives earlier. This means a greater and greater phase difference between the components.

After moving a quarter wavelength, the forward wave will lag on its phase at P by one quarter of a cycle, while the reflected wave will lead by a similar amount. The result will be that the component voltages are half a cycle out of phase, and the resultant voltage will be $v(1-k)$. This is the minimum possible value of the voltage.

A further movement of one quarter wavelength will bring the components into phase and yet another quarter wavelength will bring them into opposition. This stationary pattern will repeat continuously, if the line is long enough, and is termed a 'standing wave'.

All this becomes quite obvious when the component voltages are represented by phasors. Resultant voltage at any point on the line is found by drawing a phasor parallelogram.

Using phasors

The two component voltages are represented by phasors drawn like the hands of a clock. The long hand is v units long and the short hand is kv units. In practice it is often convenient to assume that v=1 and that the lengths are 1 and k units

At a point like *P*, a voltage maximum, the two hands are both pointing in the same direction. Say both point to 12 o'clock. As we move down the line the hands are rotated through equal angles in opposite directions, the angle representing change in phase; for instance a movement of one quarter of a wavelength is represented by rotating each phasor through 90°. This brings the hands pointing to three o'clock and nine o'clock. Phasor lengths must be subtracted to give $v(1-k)$, a voltage minimum in this case.

A further quarter of a wavelength will bring both hands pointing to six o'clock, another voltage maximum. However they are pointing in the opposite direction to when they were at 12 o'clock, so the resultant voltage at the new maximum is in antiphase with the original case, **Fig. 2a,b,c.**

Further movement down the line causes the phasors to cross over and arrive at another voltage minimum after another quarter wavelength, then arrive at their starting position. After a total movement of one full wavelength down the line, the voltage maximum at this point is in phase with the voltage at P.

Rotation convention

In moving down the line towards the antenna the forward wave is delayed, ie lag is introduced, and the reflected wave arrives earlier, ie it leads. Conventionally, lag is represented by a clockwise rotation and lead is represented by an anticlockwise rotation.

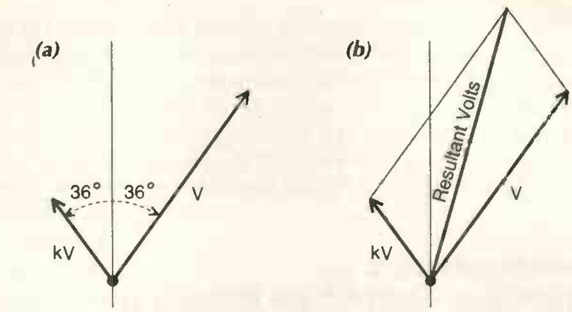


Fig. 3. A point λ/10 from a maximum is represented by two phasors rotated 36° in opposite directions from the zero line, (a). The resultant voltage at this point on the line is found using a phasor parallelogram which combines two components; the voltage due to the forward wave and the voltage due to the reflected wave, (b).

Finding resultant voltage at a point

Suppose that the point of interest is one tenth of a wavelength from a voltage maximum. With the clock hands both at 12 o'clock, rotate each hand through one tenth of a cycle, 36°, in opposite directions, **Fig. 3a.**

Now construct a parallelogram, Fig. **3b**, with the clock hands as sides. The resultant voltage is given by a diagonal drawn from the junction of the clock hands. It should be clear that when the components are in phase or in antiphase, the parallelogram collapses into a straight line. When this happens, the figure still works, giving the same answers as obtained by a straightforward addition or subtraction.

The phasor voltage parallelogram (clock hand diagram) shows how the numerical value and phase of the line voltage varies from point to point. Surprisingly enough it can be easily modified to also give current.

Voltage and current parallelograms

The currents may also be represented by phasors and it is a fairly simple matter to draw current and voltage phasor diagrams side by side with the correct phase relationships.

One thing which holds for both forward and reflected waves is that at all times and all points in the circuit, voltage/current=Z_0. Note that this is not true for the resultant values of voltage and current.

First think about the forward wave. As $V/I=Z_0$, V and I must be in step and peak at the same instant. We therefore draw the forward-wave current and voltage phasors in parallel, ie the phase difference is zero. The length of this current phasor must be equal to v/Z_0, where v is the rms voltage.

Now think about the reflected wave. Once again the current must peak at the same instant as the voltage but it has already been pointed out that reversing the direction of the wave reverses the sign of the current relative to the voltage. For the reflected wave, the current phasor must point in the opposite direction to the voltage phasor.

Mathematically speaking, reversing the sign of the current without reversing the sign of the voltage means that $V{\times}I$ is negative, ie the flow of power is reversed, which is of course true for the backward wave.

The length of the reflected wave current phasor is kv/Z_0. The resultant current is given by the diagonal drawn from the junction of the current phasors, **Fig. 4**.

"If the current diagram is scaled up by a factor Z_0, it becomes identical to the voltage parallelogram, except that the other diagonal gives the resultant."

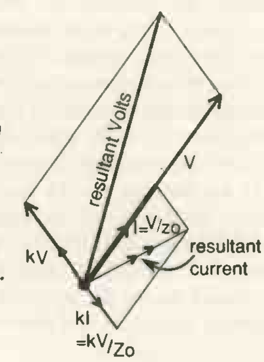


Fig. 4. Current is easily represented on the same phasor diagram as voltage with a small calculation to take into account the line impedance.

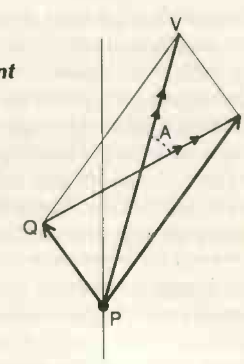


Fig. 5. Placing voltage and current phasors on one parallelogram enables the phase angle, A, between them to be seen. PV=resultant voltage, PI=resultant current×Z_0.

Combined voltage-current parallelogram

The voltage parallelogram and the scaled up current parallelogram may be superimposed to give a combined voltage-current parallelogram, **Fig. 5.** In this the length of the diagonal drawn from the tip of the short clock hand to the tip of the long clock hand represents the current, scaled up by a factor Z_0. Angle *A* between the diagonals is equal to the phase difference between voltage and current.

The combined diagram is a very powerful tool. It gives a complete description of the voltage and current at any point on the line, and enables this to be compared with the voltage and current at any other. This is illustrated very well when it is applied to a line terminated in a perfect reflector, ie when k=1.

Investigating the case when k=1

It is worth while drawing a few diagrams for the case when k=1, and thinking about what they mean. For instance, as the sides of the parallelogram are all equal the diagonals are always at right angles no matter what size the figure is. Voltage and current are in quadrature at all points. The phase of the current also remains constant apart from a sudden reversal on passing a current zero: the voltage behaves in a similar way. You will notice that the electrons in neighbouring half wavelength segments are surging back and forth in opposition, causing maximum voltage fluctuation at the points of zero current.

Line impedance, Z

In general, reflection coefficient, k, have any value between zero and unity. The ratio of voltage to current is termed the line impedance Z, and on an unmatched line Z will vary from point to point. At the output end of the line, reflection must take place so that Z is identical to the terminating impedance. At the transmitter end of the line, Z is the impedance the line presents to the transmitter.

As the diagram allows you to measure the relative magnitudes of line voltage and current and their phase difference, Z may be found at the point in question. ('Line' values of voltage or current means the resultant values). On a matched line, $Z=Z_0$ and does not vary.

Simplifying phasor diagrams

To simplify matters slightly, instead of rotating both clock hands, it is easier if the long hand is kept vertical and the small one is rotated through the double angle. If you do this, the diagrams no longer enable you to compare the phases of voltages and currents at different points on the line. But this is immaterial in finding the line impedance at a given point.

Secondly, the diagram itself may be simplified. **Figure 6** shows the voltage and scaled up current parallelograms for a point on the line. They are close together and side by side. **Figure 7** shows how unwanted lines can be omitted to give a single diagram. The scale is immaterial. Vector length **PO**=1unit and **OV=OI**=k units. **PV** represents the resultant voltage while **PI** represents the resultant current, scaled up by $Z_0 L$. Angle *A* gives the phase difference between the line voltage and current. Vector **VOI** acts as an indicator which can be rotated to reveal conditions at any point on the line.

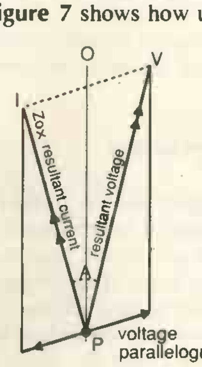


Fig. 6. With current scaled by a factor Z_0 onto the voltage parallelogram, PO represents both the forward wave voltage and the forward wave current.

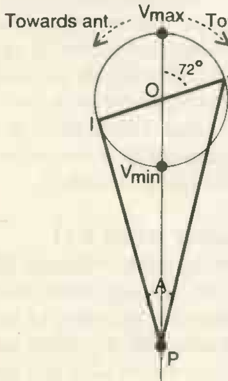


Fig. 7. Angle A gives the phase difference between line voltage and current. Removing unwanted lines, VOI can be rotated to show the line conditions at any point from generator to antenna.

An arrowhead is placed at *V*. It is helpful to imagine a clock face behind the indicator with the 12 o'clock and 6 o'clock points marked V_{max} and V_{min} respectively. Figure 7 shows the indicator set for a point one tenth of a wavelength from a voltage maximum. This requires a rotation of 2×36=72° from the 12 o'clock position, Fig. 7. A clockwise rotation is shown in Fig. 7 so the point in question must be on the transmitter side of the voltage maximum. If there is sufficient information to draw the diagram for a given point, then the line impedance, Z at that point can be found.

Z=resultant volts/resultant amps
Resultant volts=**PV**, Resultant amps=**PI**/Z_0.
$Z=Z_0$(**PV/PI**) or (Z/Z_0)=(**PV/PI**).

Now imagine that the indicator is set at 12 o'clock and then rotated to show the effect of moving along the line. As the indicator is rotated, the diagram changes shape, **PV** and **PI** change in length, and the angle between them, *A*, opens and closes.

Maxima and minima

At 12 o'clock, **PV** has its maximum length and **PI** has its minimum length. This point is a voltage maximum and a current minimum. **PV** and **PI** lie one on top of the other, so here angle *A* is zero. Voltage and current are in phase so *Z* is purely resistive and at its maximum value which is greater than Z_0. At 6 o'clock, current is at a maximum and voltage at a minimum, so *Z* has its minimum value which is less than Z_0, therefore purely resistive. At all other points *A* is not zero, voltage and current are not in phase, so *Z* is complex.

Complex values of *Z* may be represented either by a resistance R_s and a reactance X_s connected in series, or alternatively by a parallel combination of R_p and X_p.
Standard ac theory gives:

$R_s=Z\cos A$, $R_p=Z/\cos A$
$X_s=Z\sin A$, $X_p=Z/\sin A$

Provided a suitable rotation convention is used the diagram also shows whether *X* is inductive or capacitive. If movement along the line away from the transmitter is represented by an anticlockwise rotation and movement towards the transmitter by a clockwise rotation, *X* will be capacitive if **PV** is on the right of **IV**, and inductive if on the left.

A numerical example

One tenth of a wavelength of 50Ω coax is terminated with a 100Ω resistor. The termination is a pure resistance which is greater than Z_0, so the termination will be a voltage maximum. Standing wave ratio=100/50=2. This gives:

k=(2–1)/(2+1)=0.33

One tenth of a wavelength is represented by a rotation of 2×(360/10)=72°.

The indicator is initially set to 12 o'clock to represent conditions at the termination, and then rotated clockwise through 72°. Figure 7 should be drawn with k=0.33 and could then be used to solve this problem. Vectors **PV** and **PI** and angle A are measured and **PV/PI** is evaluated. Then,

PV/PI=1.2 and *A*=35°.
$Z=Z_0$(**PV/PI**)=50×1.2=60Ω.

The series components of *Z* are:
$R_s=60\cos35°=49\Omega$, $X_s=60\sin35°=34\Omega$.
The equivalent parallel components are:
$R_p=60/\cos35°=73\Omega$, $X_p=60/\sin35°=105\Omega$.

As **PV** is on the right of **PI**, *X* is capacitive. This applies to both series and parallel cases.

Alternatively, using the rotation convention, **PI** leads **PV** ; the current leads the voltage so 'Z' is capacitive.

If the termination of the line had been a resistance of 25Ω (ie $Z_0/2$ instead of $2Z_0$) 'k' and the swr would have been the same numerically, but the termination would have been a voltage minimum so the initial position of the indicator would have been at six o'clock.

The diagram as a visual aid

The real beauty of the diagram is that it reveals and clarifies so many aspects of transmission line behaviour. You only need a few rough sketches wihout any maths to get a good idea of how impedance, voltage and current and their relative phase all vary along an unmatched line. Try investigating the result of altering 'k', including making it very nearly zero and very nearly unity. Remember that the greater the angle, *A*, the larger the reactive component compared to the resistive component. When k=1, *A* is a right angle and so *Z* is a pure reactance.

Comparing phases at different points

I have already pointed out that because one of the phasors is kept fixed, the diagram cannot be used to compare the phase of voltage or current at one point with that of another. If you are interested in making such comparisons, go back to the combined voltage-current parallelogram.

The diagram and the Smith chart

Those of you familiar with the Smith chart may notice that if the diagram had been drawn upside down with *P* vertically *above* 'O', giving V_{min} at 12 o'clock and V_{max} at six o'clock, the operating rules become similar to those of the smith chart: 'X' will be inductive when the indicator points to the right, and capacitive when it points to the left. Otherwise the diagram is used exactly as before. ■

Further reading

'An impedance diagram for transmission lines', *Radio Communication*, January, February 1992.

In *Visualising electron disturbances*, March 1995, the paragraph beginning 'This is because the termination of a line . . .', pp 235, should be preceded by the heading **'Why reflection occurs'** - Ed.

SpiceAge competition results

Here are solutions to the competition held in the February 1995 edition of *EW+WW*.

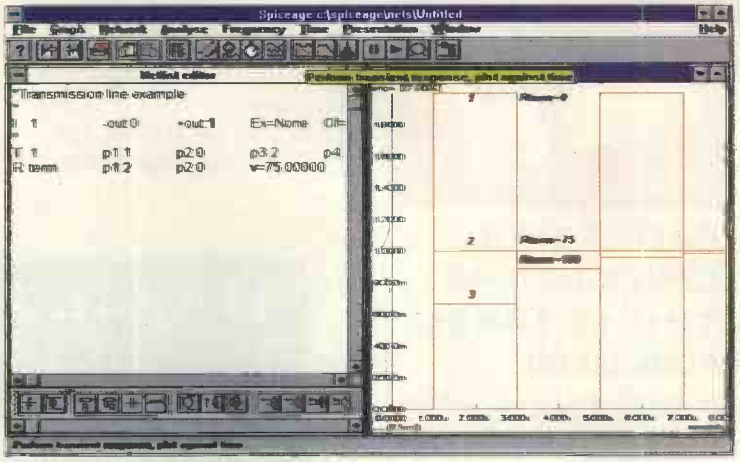
The two best entries answered all the questions correctly and completed the tie breaker with interesting and viable applications for *SpiceAge*.

Congratulations to winners, Andrew Pate of Keighley and R. Smedley of London who will each be receiving a copy of *SpiceAge for Windows Level 3*.

Q1. SpiceAge for Windows is written in which country: USA, Germany, UK, Israel, Australia, Japan?
A1. *In the UK by Graham Baxter.*

Q2. In a lossless circuit, resonance is given by $\omega=1/(LC)^{0.5}$ rad/s. Write down the analogous expression for a mechanical spring + mass system. Please define completely the terms within your expression.
A2. *$\omega=(k/m)^{0.5}$ where ω is angular velocity (for example in rad/s), k is spring stiffness (for example in N/m) and m is mass (for example in kg). Frequent wrong answers included some dependency on the gravitational constant and hence orientation of the system. (May be that's why upright pianos sound different from grands!)*

Q3. Which of the following traces which show the current in the terminating resistor is the correct response of a 75Ω transmission line terminated with a 150Ω resistor to which a current of 1A is suddenly applied.
A3. *Working man's answer is that it cannot be trace 2 as that corresponds to a matched terminating resistor. It cannot be trace 1 which is showing no dissipation in the system (the output is in fact short circuited). It must be trace 3.*



Which of the following traces showing current in a terminating resistor is the correct response of a 75Ω transmission line terminated with a 150Ω resistor to which a current of 1A is suddenly applied? See Q3.

Q4. You are on site repairing a board for which you need a 20kΩ 5% resistor. You have just dropped your ohmmeter in a puddle and your toolkit contains only one each of 5% values fitting the 10, 12, 15, 18, 22, 33 and 47 decade multipliers between 1kΩ and 47kΩ. What do you do to keep within the original value to three significant figures?
A4. *The tolerances are of course a red herring because it is the same for both the target and element values. The problem may thus be reduced to finding the simplest (series and/or parallel) combination of nominal values. There are several solutions but the exact answer is given by $18k+((2.2k)^{-1}+(22k)^{-1})^{-1}$. Only one person found this solution.*

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Accuracy	±1 digit + timebase error

Period

Frequency range	5-25MHz
Resolution	10^{-7}ns to 1µs
Accuracy	±1 digit + timebase error

Note: resolution depends on measurement time and input frequency.

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SUMMING UP CURRENT

Using variable integration, the ACF2101 switched integrator has a wide dynamic range, measuring currents down to a few nanoamps. Douglas Clarkson discusses using the device for detecting currents at photodiode levels.

As the number of applications involving currents generated by sensing devices increases, so also are refinements being sought in measurement circuits. With photodiodes, for example, where the induced current is proportional to received signal levels, current measurement may be over a broad range. It may also vary significantly with time.

Where photodiodes are measuring light levels directly or indirectly in the case of ionising radiation there is often the requirement to measure the total amount of a pulse of light or a pulse of radiation. Manufactured by Burr-Brown, this precision dual integrator can be used to undertake precision integration in a variety of modes to cater for such measurements.

The advantages of such a device are compactness and low droop voltage with time. Such units can be controlled with standard ttl logic levels.

General applications include current to voltage conversion, photodiode signal integration, current measurement, charge measurement, ct scanner front end and general medical, scientific and industrial instrumentation.

Device basics

Figure 1 shows the block diagram of the *ACF2101*, which is designed to operate on +5V and –15V supplies and draws typically 3mA on the negative supply and 12mA on the positive supply. Control of the device is basically undertaken by means of hold and reset logic pins, shown in **Fig. 2.** When hold is high and reset low (*on* condition), input current is

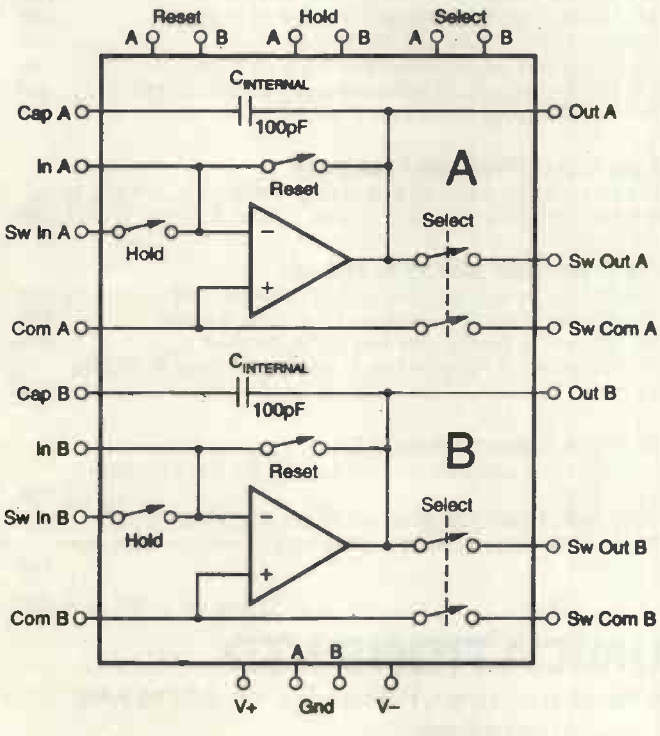


Fig. 1. Functional diagram of the ACF2101, precision integrator from Burr-Brown. Features include droop characteristics of 1nV/µs and integrate slew rates of 3V/µs.

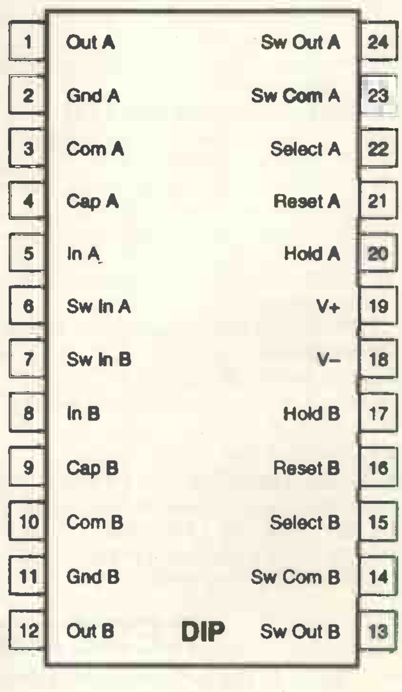
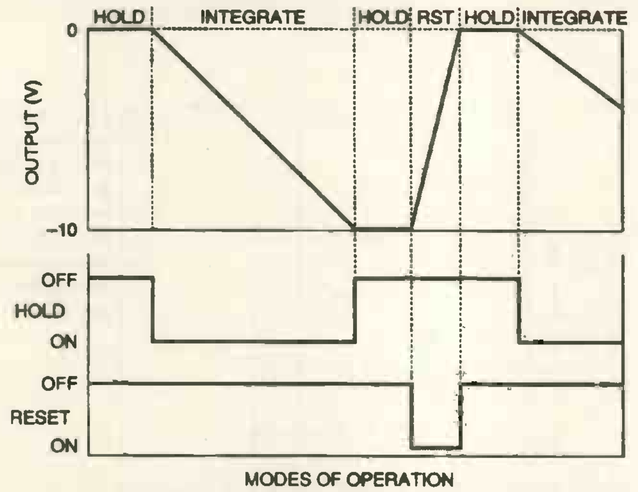


Fig. 2. Pinout of the ACF2101 dip. The pins are arranged this way to allow for the stray-current guard rings.



SWITCH	MODE OF OPERATION Integrate	Hold	Reset
Hold Switch	ON	OFF	ON/OFF
Reset Switch	OFF	OFF	ON

ON: Switch shorted; Logic 0 input. OFF: Switch open; Logic 1 input.

Fig. 3. Input current is integrated when the hold line is high and reset line is low.

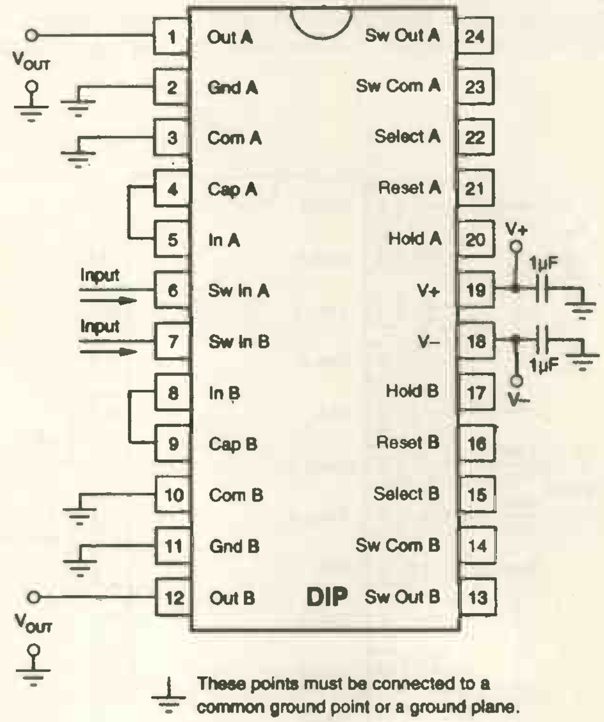


Fig. 4. Connections to the IC when using the internal precision capacitances is very straightforward.

integrated and the output voltage falls below 0V. The limit to negative integration voltage is –10V. Logic control is outlined in **Fig. 3**.

The device has an internal precision 100pF capacitance for each channel selectable using the circuit connections of **Fig. 4**. Where larger currents need to be integrated, a separate external capacitor can be included in each circuit as shown in **Fig. 5**.

Using the relationship,

$$V_{out} = -\frac{I_{in} \times \mathrm{dt}}{C}$$

where I_{in} is the input current, dt is the time of integration and C is the integrating capacitor, values of I_{in} dt and C can be calculated for a full scale value of V_{out} of –10V.

In applications in optical measurement, data can be required to be captured rapidly in micro seconds or over longer time periods, up to several 50Hz cycles. For longer periods, if a signal is being sampled over 50 half cycles at 10ms per half cycle then the sample time is 0.5s. With a typical photodiode current of 5µA the required capacitance will be 0.5µF.

It is important that the capacitance used is of sufficiently good quality. A high performance polypropylene type will minimise leakage losses, for example. Such integrators are valuable for example in measuring levels of light from sources which are varying rapidly, such as fluorescent tubes. The integrator can be switched on for a set period in order to capture an integral number of cycles of visible or ultra-violet light output.

In addition each output amplifier has an output select switch which allows for multiplexing of devices using an instrumentation amplifier as shown in **Fig. 6**. For all devices unselected, the integrated charge is held in each device and the output is not connected. For selection of a device, the output is communicated to the instrumentation amplifier. In this way a series of channels can be controlled by a common set of logic signals and individual integrated channels can then be selected.

Functional use

Normally the cycle of operation will be: *reset* – clear residual charge, *hold, integrate, hold* – maintain final voltage and read value. Data would normally be read at some point during the hold cycle. Voltage droop taking place during the *hold* cycle is given by

$$droop = \frac{200 fA}{C}$$

where C is integration capacitance in farads and fA is femto amps.

For a 100pF capacitance and with no additional leakage currents this is equivalent to 2mV/s or 2nV/µs.

The logic switching of *hold* and *reset* will cause charge transfer to take place. It will be of a sign which is a function of the sign of the transition – positive going or negative going. The magnitude of this switch is typically 0.1pC and this corresponds to a voltage offset of 1mV for a 100pF capacitance. Where this effect becomes significant, its impact can be minimised by ensuring that the *reset* and *hold* logic transitions cancel out.

These effects of voltage droop and charge transfer are reduced for larger values of capacitance. Thus for 10,000pF the voltage droop will be 0.2mV/sec and the charge transfer voltage 10µV.

Table 1. Values of input current integration time and integration capacitance to achieve full scale output of –10V.

I_{in} (µA)	dt(s)	C(pf)
0.01	0.1	100
0.1	10m	100
1	1m	100
10	100µ	100
100	10µ	100
10	1m	1000
100	1m	10000

Timing control

Control of the *ACF2101*'s integration time is a key element of successful use of the device. Timing needs to be accurate for a set configuration, but options should be available to select various integration times. Thus reproducibility of integration period needs to be good. There are many ways in which such consistency and control of timing can be achieved. Some of them are considered here. Assuming a logic level transition is used to initiate a timing sequence, a device such as the *4538* dual retriggerable monostable multivibrator provides a convenient way of producing an reproducible integrating pulse width, as shown in **Fig. 7**. With an input low to high transition on pin 4, the inverse output on pin 6 gives the required logic transition for time *t* where t=0.7RC. **Table 1** indicates how a range of values of *t* can be configured.

Table 1. Value of pulse duration with cmos 4538 monostable.

Value *R*	Value *C*	Time *t*
10k	0.1μF	0.7ms
100k	0.1μF	7ms
1M	0.1μF	70ms

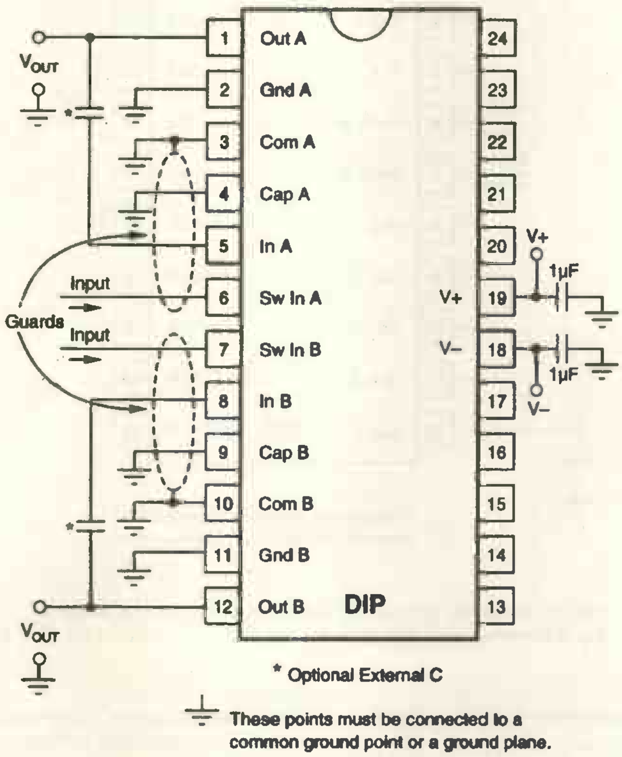


Fig. 5. Precision capacitances with low leakage current are used externally from the ic to alter the time constant of the integrator.

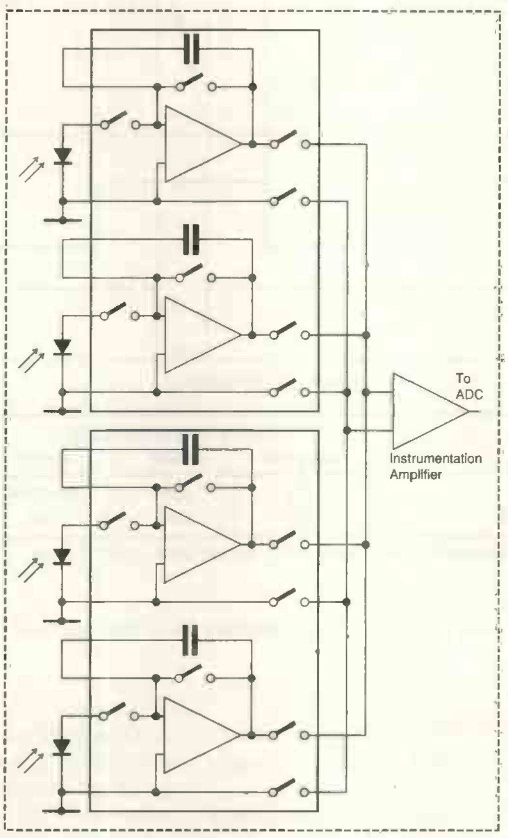
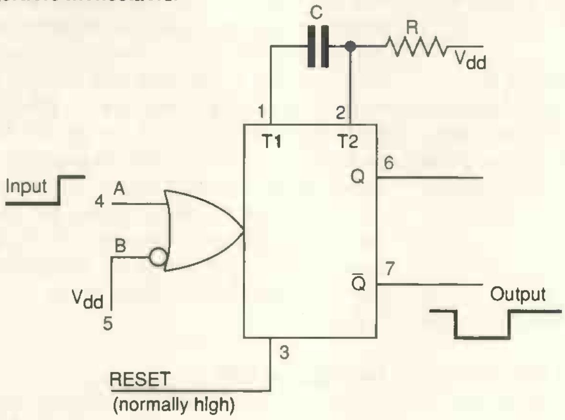


Fig. 6. Signals may be multiplexed using the output switches of the ACF2101 and an instrumentation amplifier.

Fig. 7. An integrating pulse is provided by a simple stand alone retriggerable monostable.



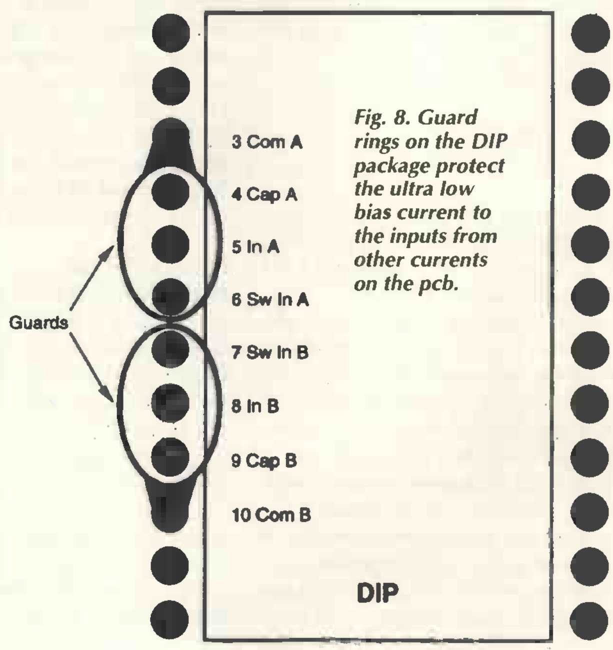


Fig. 8. Guard rings on the DIP package protect the ultra low bias current to the inputs from other currents on the pcb.

In this configuration the timing is a function of the specific values of R and C. There is also the short term problem of temperature drift and long term problem of device aging. For a resistor with a temperature coefficient of +500ppm (parts per million) a 10° rise in temperature will result in a change of value of 0.5%. For a few extra pence per component, resistors with a temperature coefficient of +50ppm can be obtained – reducing the percentage change in value for a 10° rise in temperature to 0.05%.

Stable capacitors such as polypropylene have a negative coefficient of around –200ppm while types such as polyester have values of around +300ppm. Where possible, temperature coefficients should be of equal magnitude but of opposite sign. Often, however, more control is required over timing – both for accuracy and range of values. The circuit of Fig. 7. can be triggered with a value of RC large enough to complete long pulses of several seconds, but with an external timing transition to reset the output. This output can be, for example, a timing line derived from a crystal of value 32.642kHz and divided down by a *4020* 14-stage binary counter. The positive going start pulse could initialise the counter and set the inverted output of the *4538* monostable low as a long pulse (several seconds is triggered.

Choice of integration period can be influenced by the nature of the signal being captured. Where, for example visible or ultra violet light levels from fluorescent sources are being measured, then the light output takes the form of rectified sine waves. At 50Hz cycles, the period of each positive cycle is 10ms. It would be appropriate to use a 250ms or 500ms integration pulse width with such signals to minimise problems of signal aliasing.

Circuit design using *ACF2101* devices

The ultra low operational amplifier bias current of around 100fA, requires careful pcb design. **Figure 8** indicates how so called guards are required to protect the inputs. Current which could flow from other tracks to the input track is instead trapped at the guard track. Handling boards can also increase voltage droop. Cleaning boards using solvents and de-ionised water minimises this effect.

Summary

The *ACF2101* device has wide application in circuits measuring currents over a wide dynamic range. Care is needed with circuit layout in order to prevent leakage currents being picked up by the device as 'current signal'. Obtain a current data sheet on the device and not a preliminary one. ■

Further reading

AC2101 data sheet, Burr Brown International, 1 Millfield House, Woodshots Meadow, Croxley Centre, Watford, Herts, WD1 8YD.

Simulated attack on slew rates

Late last year, Douglas Self discussed some practical limits for slew rates. But were his arguments relevant to modern music? "No," says **Ben Duncan**, *and sets out his simulations to prove it.*

Slew rate limits based on Baxandall's 2.2kHz criterion may be true of classical music recorded with dynamic mics, analogue taped and cut onto vinyl with twenty year old technology. But for most music reproduced by power amplifiers today that limit is patently false.

Take live performances: a glance at the spectrum analyser during an Iron Maiden gig – engineered by my colleague Doug Hall – would show the 20kHz led lit almost solidly throughout numerous concerts over the past decade or so.

Consistently high hf levels, with amplitudes as large as the loudest bass notes, are unexceptional with certain genres of music. Iron Maiden achieves them with nothing more than traditional (if heavily thumped) percussion instruments.

Low mass modern capacitor microphone capsules and local buffering to reduce hf attenuation and loading dips in stage-to-mixer cabling have boosted the acquisition of percussive edges with quasi-fundamentals of 15kHz and above[1]. Live sound and recording consoles have equalisers on every input channel, and most are in use. But I have measured none that does not also increase ultrasonic frequencies when any kind of boost is dialled up on the hf control(s), pushing frequencies above 4kHz. This unadvertised – and not readily avoidable – ultrasonic boosting can at least counter, if not overwhelm, rf filtration and band-limiting in the system. In this way, unexpectedly high level ultrasonic signals can appear at the power amplifiers' inputs.

As for recorded replay in studios and homes, cd and dat can handle full level at 20kHz. While for direct-cut vinyl recordings, the bandwidth above which no additional musically significant cues are heard can range, for some listeners, to 200kHz.

Unfortunately, old data ignores the break-neck development in the past seven years, such as fm synthesis of electronically generated, manipulated and sample-based music. In effect new 'virtual instruments' have been created to add to the family employed by mainstream music, untrammelled by

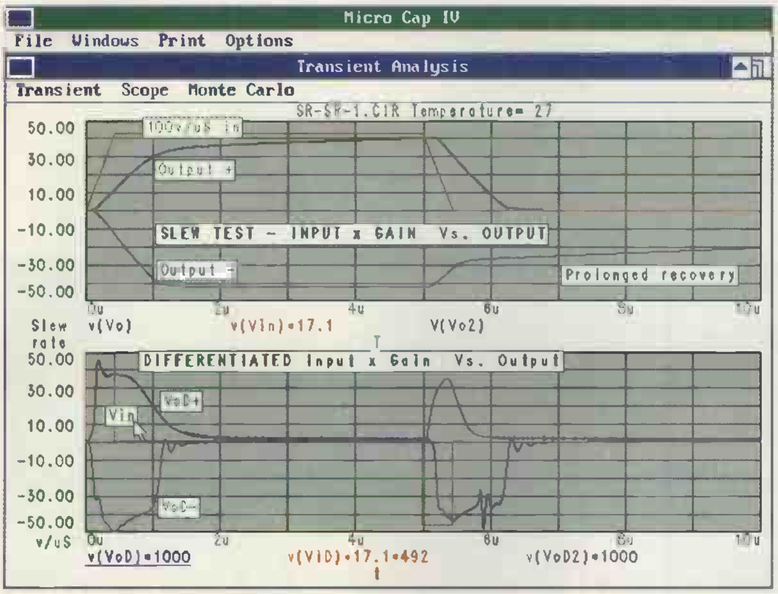


Fig. 1. Benchmarking Self's slew analysis with MicroCAP IV *begins with the generic circuit, using ideal current sources. The upper panel shows (uppermost) the input test step magnitude, which rises at 100V/µs, multiplied by the gain, followed by the amplifier's output responses. The lower panel is scaled in V/µs. Unlike physical test circuits, testing can be simultaneously carried out with both +V and –V signals. Note the negative output signal (lowermost) is worryingly prolonged.*

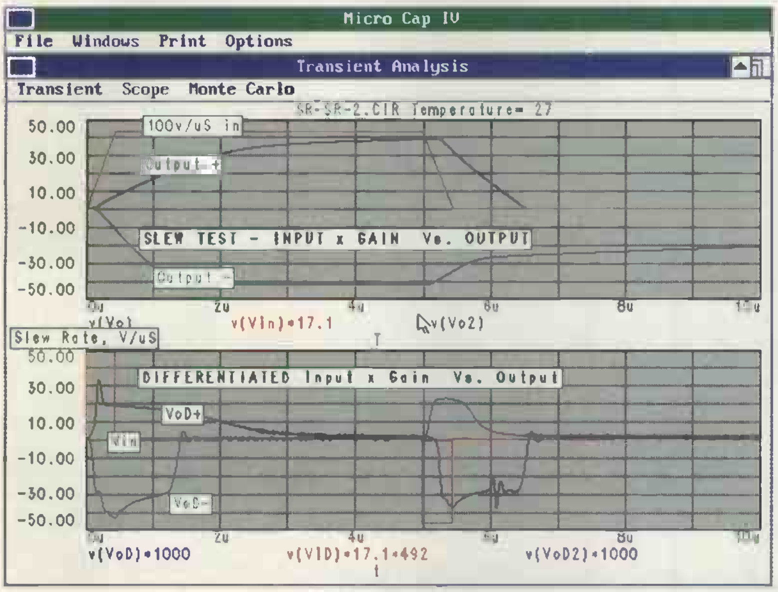


Fig. 2. MicroCAP*'s ideal current sources are exchanged for the current source circuitry in Self's Fig. 1 (EW+WW, Sept, 1994). The negative recovery problem persists. Both output polarity amplitudes are near to clip but still safely short of it.*

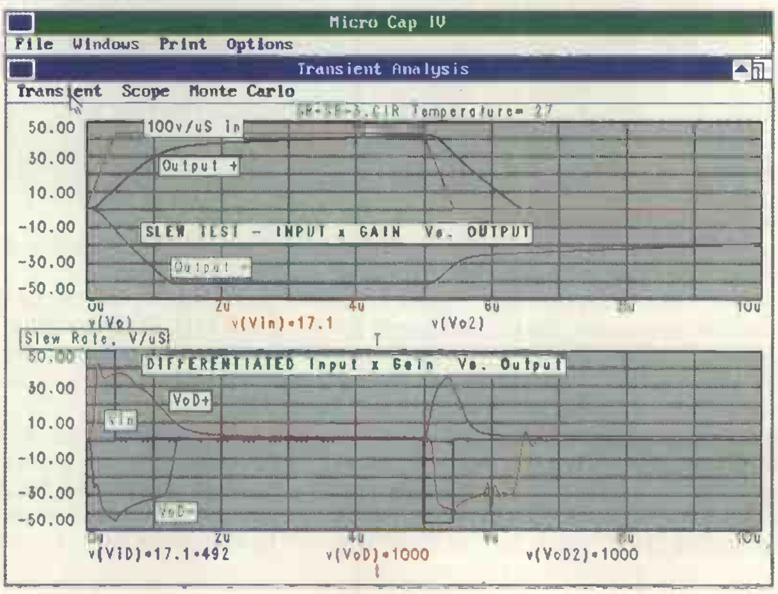


Fig. 3. Self's current source improvements are installed. This gets positive slew performance back to Fig. 1 but does little to fix the asymmetry and does nothing for the messy peaking, and chewing-gum negative recovery.

the limits of wood and metal.

Many of these new sounds and 'instruments' can include or produce full (0dB) levels at 20kHz. Even the breathy sounds of close-miked vocals can become unacceptably sibilant when handled live by some bipolar power amplifiers – a problem widely experienced as disappearing when faster mosfet amplifiers are substituted.

RF filtration

Must an amplifier have rf filtration on its input port?

The ideal per-stage bandwidth for high quality music reinforcement, monitoring and reproduction should be at least 100kHz and ideally no more than –3dB at 200kHz[3]. Above this, input filtration is positively desirable. Without it, whether the incoming program can slew at a rate that taxes the amplifier is not an issue. Radio frequency can (and regularly does) make egress in the cabling preceding most power amplifiers. Self's design (his Fig. 1 in ref. 2) omits input filtration, yet a 900MHz mobile phone signal need only peak at 100mV to slew at 500V/µs. Just 11mV peak ingress would out-slew even a 50V/µs capability.

Radio frequency filtration does not slow fast edges; it just reduces the amplitude portion of the 'rate' or amplitude-frequency product. So input (and other portal) rf filtering (cf interior band-limiting) only protects an amplifier with a marginal slew rate if rf levels are below a certain threshold.

In the real world, the ubiquitous rf debris that is part of urban living may be less obliging. Fast-edged 'clicks' caused by dust particles on vinyl striking a stylus can also tax amplifiers harder than the toughest programme.

Rateable values

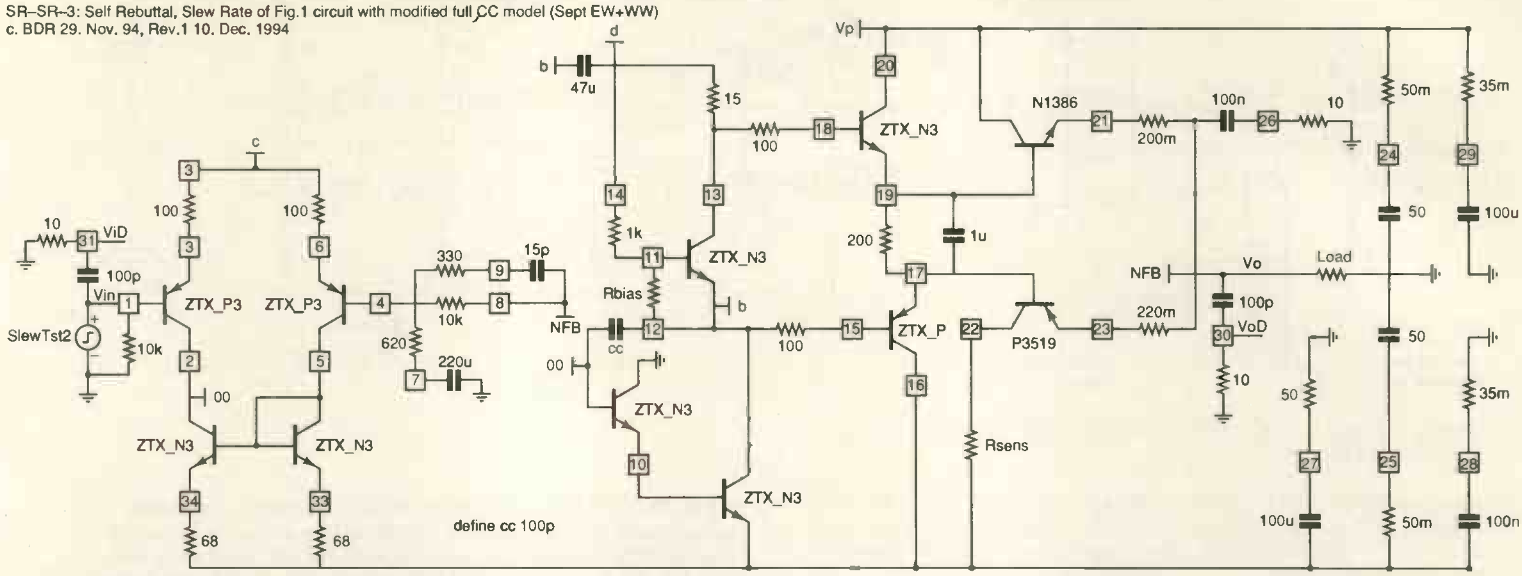
The foregoing shows why higher slew limits than those needed to reproduce a 20kHz – let alone 2kHz – sine wave at 0dBr, can be justifiable. For any given degree of hf filtration, more-than-adequate slew limits alone can distance the ears from the very unpleasant and ear-fatiguing distortion that can begin when an amplifiers' slew limit is approached by a factor of two, or even a ten. At the point where visible slewing begins, when signal slew is equal to the amplifier's limit, thd is already about 1% and the damage – with grating high harmonics – has already been done.

These events may happen rarely in many systems and situations, and in others not at all. But no-one should not dismiss the validity of higher-than-expected slew rates as being beneficial to sonic quality because a difference could not be heard with casual listening. No-one would remove the rear fog lamps from a car on the grounds that other vehicles had never run into it.

Walt Jung's original and in-depth work on slewing[4] covered this ground 18 years ago and should be compulsory reading for anyone writing on this topic.

Slewing can trigger prolonged indigestion in amplifiers. Even when it doesn't, as little as 1-2s of gross distortion due to slew-limitation during a performance can effect listener enjoyment for several minutes afterwards – an effect analogous to gross video corruption when watching a spell-binding film.

The upshot is that prudent minimum slew limits for quality audio should be at

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.MODEL Slewtst2 PUL (VONE=2.50 P1=10n P2=438n  P3=5u P4=5.45u
P5=10.01u)

.MODEL ZT_XP3 PNP (BF=250 BR=50 XTB=1.5 IS=382.06F CJC=45.5P
CJE=278P RB=40M RC=65M VAF=154 TF=780P TR=30N MJC=453.4M VJC=577.4M
MJE=347.71M VJE=281.65M CJS=1F VAR=38 NF=1.0025 NR=1.0012 ISE=103.5F
ISC=700F IKF=1.15 IKR=420M NE=1.3642 NC=1.19 RE=87.5M IRB=1 RBM=100M
VTF=1K MJS=500M)

.MODEL ZTX_N3 NPN (BF=200 BR=33 XTB=1.5 IS=320.07F CJC=80P CJE=350P
RB=87M RC=70M VAF=76 TF=860P TR=24N MJC=489.6M VJC=767.6M MJE=376.61M
VJE=440.67M CJS=1F VAR=51 NF=1.0041 NR=1.0008 ISE=80F ISC=60F IKF=1.6
IKR=450M NE=1.57 NC=1.079 RE=80M IRB=1 RBM=100M VTF=1K MJS=500M)

.MODEL N1386 NPN (BF=43 XTB=1.5 IS=33.69F CJC=352P CJE=13.4N RB=400M
RC=13M VAF=75 TF=4.54N TR=1000P MJC=336.47M MJE=500M VJE=800M CJS=1F
VAR=100 ISE=6.279P ISC=0.1F IKF=21.3 IKR=1K RE=100M IRB=500M RBM=100M
VTF=1K MJS=500M)

.MODEL P3519 PNP (BF=169.339 BR=10 XTB=1.5 IS=956.328F CJC=1.06282N
CJE=151.61P RB=370.37M RC=359.275M VAF=100 TF=3.90097N TR=76.5695N
MJC=468.399M VJC=775.989M MJE=330.622M VJE=781.588M CJS=1F VAR=100
ISE=4.36566P ISC=0.0000121848F IKF=6.06776 IKR=1K NE=1.33389
NC=2.77489 RE=44.2091M IRB=1 VTF=10 ITF=111.342M XTF=499.973M
MJS=500M)

.define load 8
.define Rbias 320
.define Rsens 1m
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. define load 8
. define Rbias 320
. define Rsens 1m

Self enhanced BJT DCS:

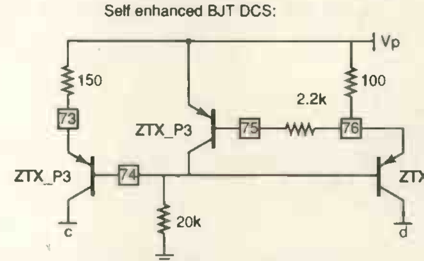


Fig. 4. Simulation circuit used for Fig. 3 precisely follows Self's own Fig. 1[1] schematic. British semiconductor models are used for the small signal bjts; Zetex (ex-Ferranti) is the only general semiconductor maker in the world so far to issue Level 3 (Gummel-Poon) bjt data. Output devices are modern, fast (20MHz), T03P. Models were created with extra data supplied by the Japanese maker. This was entered into MicroCap's refined parametric extraction program (PEP), to derive the Level 3 parameters.
The sister circuits used to plot Figs. 1 and 2 differ as follows:
In Fig. 1, all the current source circuitry is replaced by ideal, independent sources. In Fig. 2, current source circuitry connections are as shown in Self's Fig. 1.
In all three circuits, differentiator RC values have been scaled to permit accurate results up to 500V/μs.

least Walt Jung's 18 year old recommendation of 0.5V/μs per peak output volt (pk V_o). This allows an 80kHz power bandwidth (referred to 1% distortion) and remains almost safe for a great deal of reproduced sound: by a factor of four for cd and dat, and at least five for live BBC (vhf, fm) broadcasts.

The 55V/μs per 40V rms cited of Pass amounts to 1.4V/μs per pk V_o and corresponds to 200kHz bandwidth. That figure, independently established by Rupert Neve, is ideal both for live music reproduction and a first rate recording chain.

Taking Pass's criteria as the reference for today's highest quality design, then a 75W into 8Ω (i.e. 35V peak swing)

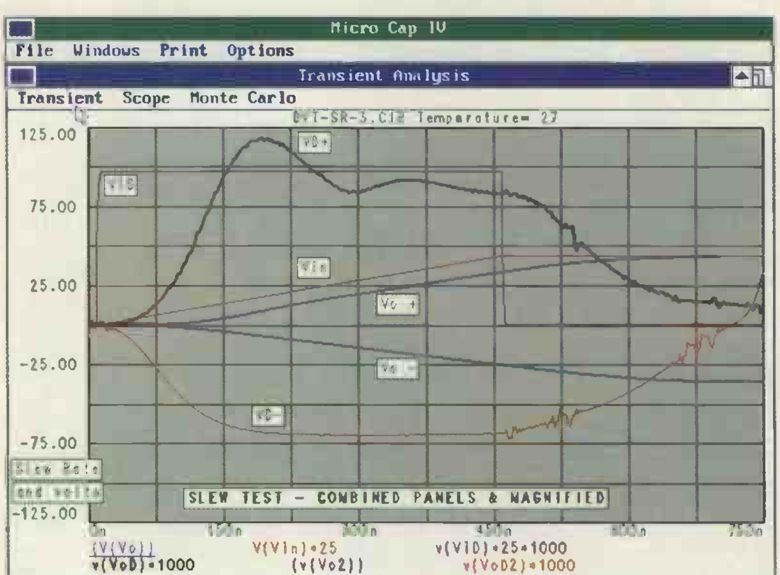
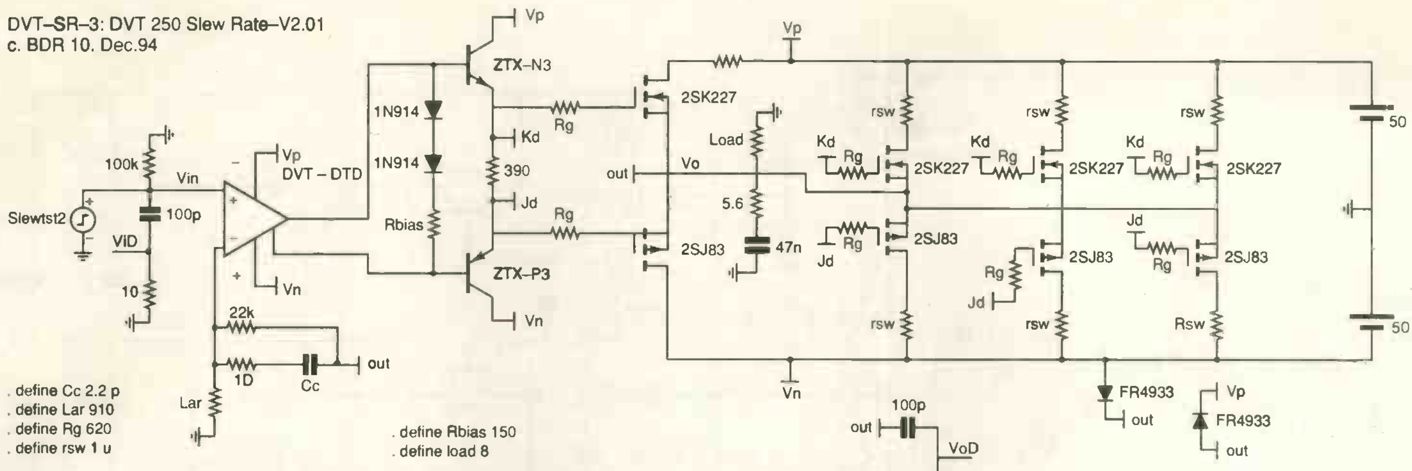


Fig. 5. When launched in 1984, the Rauch DVT250s was the fastest slewing professional power amplifier. The plot demonstrates at least +80 and –70V/μs, this before any speed-up emitter degeneration resistors are added. For magnified detail, upper and lower panels have been combined, so the Y scale reads both magnitude of the inner three plots in volts; and slew rates in V/μs of the three outermost plots (Vid & VoD+, VoD-).

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.MODEL Slewtst2 PUL (VONE=1.75 P1=10N P2=460n P3=5U
P4=5.45U P5=10.1U)

.MODEL 2SK227 NMOS (VTO=-284M KP=20U L=2U W=36.6M
GAMMA=0 PHI=600M LAMBDA=8.47M RD=338M RS=52M
CGSO=358.7P CGDO=358.7P TOX=100N NSUB=0 NSS=0 TPG=1
UO=600 KAPPA=200M)

.MODEL 2SJ83 PMOS (VTO=168.5M KP=20U L=2U W=38.5M
GAMMA=0 PHI=600M LAMBDA=50.4M RD=26U RS=396M
CGSO=358.7P CGDO=358.7P TOX=100N NSUB=0 NSS=0 TPG=1
UO=600 KAPPA=200M)

.define Cc 2.2p
.define Lar 910
.define Rg 620
.define rsw 1u
.define Rbias 150
.define load 8
```

Fig. 6. The DVT has simpler drive circuitry than Self's, coupled to complementary power mosfets. The drive IC is a macro of a fast op-amp built from discrete components (DT). For fair comparison with Figs 1-3, the 8Ω load, gain and differentiator values (10Ω and 100pF) are identical. Switch resistors 'rsw' allow ³/₄ of the output mosfets to be disconnected to view the performance of one mosfet pair.

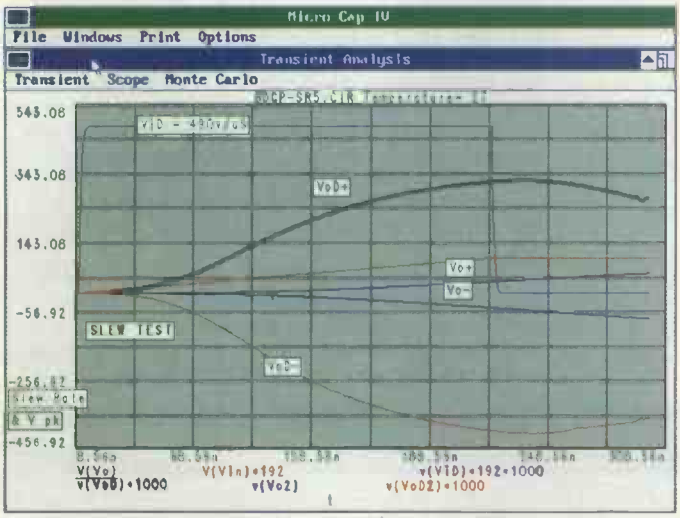


Fig. 7. Testing the high speed topology with a 485V/µs step yields about +315, –405V/µs. The mild asymmetry is academic, with nearly 2V/µs/V_o pk available for 160V swings (800W into 16 Ω). Speed is readily increased further. Physical (as opposed to simulated) slew testing of this order needs care, as aside from the usual high speed traps [7] the low resistance differentiator networks shown dissipate heavily.

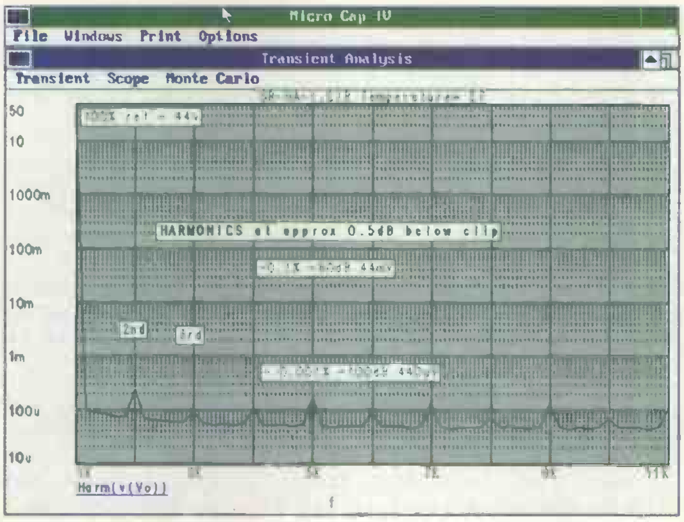


Fig. 9. Simulated harmonics in Self's Fig. 1 circuit, 0.5dB below clip. The top left marker tells us that the 1kHz fundamental is at 44V (discovered by using MicroCap's cursor function). So the ugly ninth harmonic, at 115µV, is about 0.0003%. This is audibly more significant than it appears.

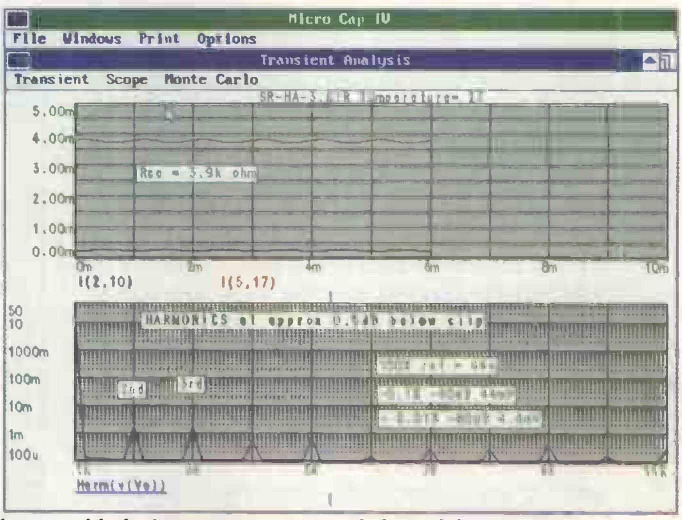
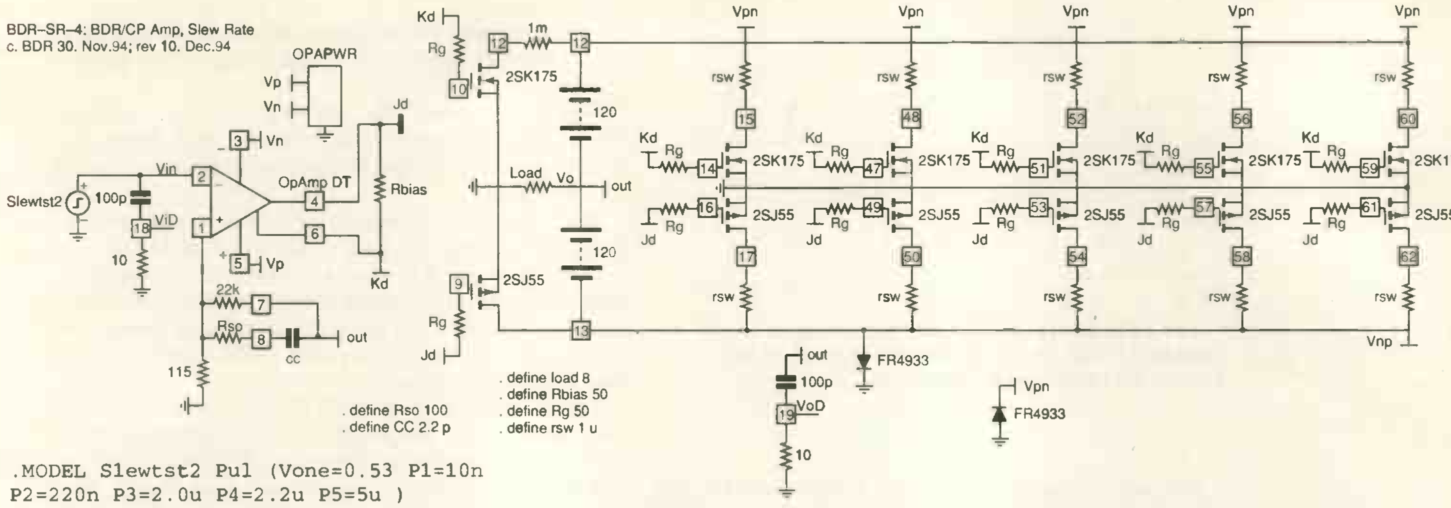


Fig. 10. With the input stage current unbalanced 'by uninspired guesswork', here increasing R_c to 3.9kΩ, the harmonic structure changes. Here, the evens are strengthened over Fig. 9, but not enough.

Virtual spectrum analysis

Using *MicroCap IV* as a spectrum analyser is straightforward, requiring just the Harm operator (`Harm[v[nn]]` where nn is usually the output) an analysis period that is an integer of the stimulus period, and adequate memory. Here, just over 1MB (of 4MB system memory) was free for use. A *486-DX* with Dos-6 is a sensible minimum. Even with *MicroCap IV* – the fastest pc software simulator – each high-accuracy run will offer you a minute or two to spare.

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.MODEL Slewtst2 Pul (Vone=0.53 P1=10n
P2=220n P3=2.0u P4=2.2u P5=5u )

.MODEL 2SK175 NMOS (VTO=-284M KP=20U
L=2U W=36.6M GAMMA=0 PHI=600M
LAMBDA=8.47M RD=338M RS=52M
CGSO=358.7P CGDO=358.7P TOX=100N

NSUB=0 NSS=0 TPG=1 UO=600 KAPPA=200M)

.MODEL 2SJ55 PMOS (VTO=168.5M KP=20U L=2U
W=38.5M GAMMA=0 PHI=600M LAMBDA=50.4M RD=26U
RS=396M CGSO=358.7P CGDO=358.7P TOX=100N NSUB=0
NSS=0 TPG=1 UO=600 KAPPA=200M)

.MODEL FR4933 D (IS=139.577N RS=14.1313M
TT=100P CJO=1.10199N VJ=217.281M M=380.094M
EG=850M XTI=2 BV=270 RL=1000G)

.define Rbias 50
.define load 8
.define rsw 1u
.define Rso 100
.define Rg 50
.define CC 2.2p
```

Fig. 8. The high speed topology again employs lateral mosfets. Also like the DVT, it has a discrete op-amp drive stage. The number of mosfets does have an effect on slew limit, and besides, six are needed to comfortably drive 8Ω speakers at 113V rms. In real use, the fast recovery diodes are important for spike protection. The 'OPAPWR' box is a dual rail psu macro for the driver op-amp stage.

amplifier, even modified to ±50V/μs, is just about acceptable. But it is certainly unsuited for the higher swings often required into inefficient speakers that are the domestic norm. And also for today's weight- and volume-challenged live music monitoring and reinforcement amplification.

In practice, with advanced hf drive units (both dome tweeters and compression drivers) used in professional live music monitoring able to handle music transients above 165V without tearing, serious listeners can safely use amplifiers with swings much higher than the 35V on offer.

Like a powerful car in experienced hands, the headroom is demonstrably safer for drive units and ears alike – no matter how counter-intuitive this seems.

The present record of about 160V_o pk output per channel is set by one of my own designs – made by a UK company – and two similarly rated units from US competitors. Their seemingly outrageous one-and-a-bit kilowatts into 8Ω/channel provides about 2dB of headroom when full range monitor speakers are called on to replay at live-performance spls. Using the Pass criterion, and assuming these goliath amplifiers are intended to handle full range music signals, then their slew limit needs to be over of 220V/μs.

In his article, Self comments: "There was an isolated claim of 200V/μs, but I doubt it". This is undoubtedly a reference to an unpublished circuit drawing I supplied to aid Doulas's research. It did indeed show an amplifier slewing at 200V/μs.

Using simulation, I will now put the slew limitations of the Self amplifier in perspective.

Out-slewn

Figure 1 shows the slewing performance of Self's Fig. 1[1] based – with one crucial simplification – on an earlier circuit[5]. Upper panel of the transient analysis plot shows the output signal – which should eventually reach 42V – and the input test signal multiplied by the 1kHz (midband) gain.

At 5μs, where the positive output has nearly settled, hf roll-off caused by C_f (C_7) and C_{dom} (C_3) accounts for the difference in actual 40V and predicted 42V amplitudes.

Lower panel shows input and output signals as above, but with the voltage differentiated so the Y scale is in *V*/μs. In *MicroCap*, these rate-of-change graphs can be achieved by simply entering `D(v(nn))` where `nn` is either the node name (in this instance `ViD` and `VoD`, meaning differentiated V_{in} or V_{out}), or the automatic node number. This is *MicroCAP*'s equivalent of Self's "Agricultural *Spice*" plots in his Figs 4 and 5[1]. But to include the effect of loading the output with the differentiation network needed for real physical measurements, the differentiated signals are all derived from simulated *RC* differentiators.

Simulated differences

The circuitry entered in *MicroCap IV* to produce Figs.1 - 3 differs from Self's explicit circuit in only minor respects which will not greatly affect the predictions.

The supply comprises perfect dc from 50V batteries, albeit with 30mΩ series resistance, so the rails do drop slightly when loaded.

All transistor parameters are perfectly matched between identically named individuals.

Protection diode D_1 and the input dc blocking cap C_1 are excluded, and electrolytic capacitors C_2 and C_4 are modelled as ideal non-polar components. All the small signal bipolar transisotrs are Level 3 (Gummel-Poon) models.

All simulations were carried out at 27°C, the default temperature. Stepping over more realistic operating temperatures – up to 80°C in some cases – is straightforward but requires individual graphs for clarity.

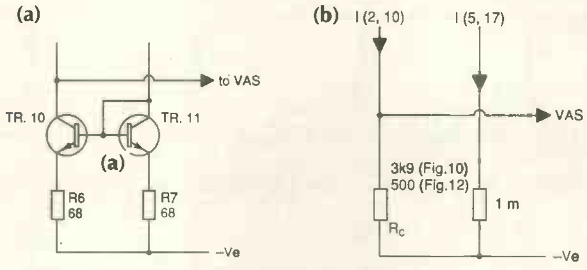

Fig. 11a. Sub-circuit simulated in Fig. 9. Network b) is employed in Figs 10 and 12. Putting a single R_c in the forward leg causes a major imbalance.

The output referred 100V/μs test signal would be well above the scale maximum of 50V/μs. Multiplication by 492 rather than 1000 – the factor needed to get the differentiator outputs converted so each volt = 1V/μs – brings it into visibility.

According to the model, the topology used by Self is slewing at +37V/μs in the main flattish region, though peak slew is +45V/μs. Negative slew rate has no flat region, but peaks at over –50V/μs, and asymptotes about –37V/μs.

Only one exception to a full, frank simulation was used to achieve these results: ideal current sources were used in place of $Tr_{1,6}$ and Tr_{14} and surrounding parts (see *Simulated differences* panel).

In **Fig. 2** the current-sources were re-entered exactly as in Self's circuit. Here, the flat portion of the positive slew limit reduces to +18V/μs, and the peak to about +32V/μs. Negative slew (V_o–) peaks at –43V/μs, and if there were a flat region, it would be about –30V/μs.

After rewiring with Self's Fig. 7 current-source improvements (**Figs. 3 and 4**) we are back to +36V/μs or a peak of +42V/μs, and as much as –45V/μs. This is not quite the ±50V/μs claimed by Self, but it is near enough, and follows the right pattern. The feedthrough 'braking' effect on positive slewing is real and undisputed.

What is disputed is the ability of other designs to outpace the Lin topology employed by Self, which dates back forty years. Quality-conscious designers have long ago moved on, having discovered that as soon as one aspect of this cantankerous topology is perfected, another collapses into disorder or asymmetry.

Fig. 12. With the input stage almost perfectly re-balanced without the current mirror, the harmonic structure is once again different, and the seventh is thankfully reduced. Note inter-collector current waveform discrepancies in the upper panel. In the top panel, the lighter curve is 5, 17 while the darker is 2, 10.

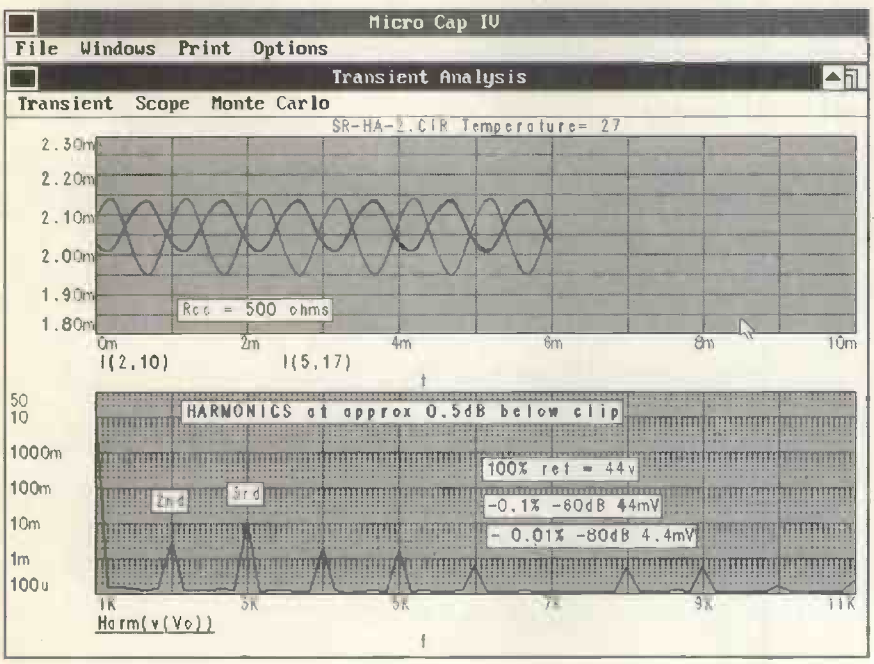

Figure 5 shows the slewing of the *DVT250s* 440W/4Ω/channel professional mosfet amplifier, **Fig. 6**, first produced in 1984 by Rauch Precision, a company founded by Jerry Mead.

When launched, its slew performance was exceptional, though the circuitry was fairly conventional. In the later version simulated here it employs six transistors – excepting the output mosfets, current source and mirror. Distortion is higher than Self's. But it still meets the essential raw criterion of well below 0.1% thd into any rated load at any level below clip, at any audio frequency.

The plots confirm that the *DVT*'s slew rate is about +85 and –70V/μs in the flat portion, ie. at least 0.88V/μs per pk V_o for the 62V peak swing. Peak slew is 120V/μs.

Tellingly, second-hand samples are much sought after today and reach high resale prices. An oft cited reason is effortless treble quality that sound engineers have realised they are not experiencing in more recent amplifiers.

In the past decade, leading designers charged with creating quality high-power amplifiers for music have had to develop topologies to ease provision of the higher slew rates needed for high swings, above 90V.

Figure 7 demonstrates graceful and only mildly asymmetrical slewing in excess of ±300V/μs, in response to the 485V/μs test signal – a performance achieved with my 160Vpeak-capable design as well as **Fig. 8**. Speed can readily be pushed higher, yet percentage thd is not being traded: at over 1300W into 8Ω, it can be as good as 0.02% at 20kHz. In all cases, mosfets are used. Bipolar junction devices would be vapourised without added anti-saturation circuitry.

When amplifiers with such a high swing and high slew are bare-tested with rf filtration removed, even heavy-gauge pvc cabling made for 50Hz mains can melt, demonstrating the gross energy abstraction of pvc as dielectric. Naturally, this does not occur with 'audiophile' cables of ptfe construction.

Harmonic argument: Self rebuttal

In *Distortion in Power Amplifiers*[5], Self describes the thd-reducing effect of perfectly balancing the collector currents in the input differential pair transistors. He makes no mention of a classic *EW+WW* article[6] on this topic however. Instead, he infers that some practices in this area are 'misguided': for example when the collector resistor (i.e. R_2 in his Fig. 7a) is set for pair current imbalance. But simulation can demonstrate that there is method in the apparently wayward component values used in some highly-rated amplifier designs.

Figure 9 shows *MicroCAP-IV*'s powerful spectral analysis being used to give a second opinion on the same very high order (below three parts per million) spectral resolution carried out with the Audio Precision System One test set. As before (Figs. 3, 4), the circuit being simulated is Self's, with full current source modifications.

The amplifier is being driven about 0.5dB below true clip, with a near perfect pure dc power supply (batteries), and a 1kHz stimulus, into an 8Ω load. This is one of the most benign conditions a real power amplifier can expect.

The amplifier does indeed demonstrate low harmonics in this static domain. But notice that while the second is the greatest, the next largest are the scarcely benign fifth and the positively metallic sounding seventh harmonics.

Even harmonics are depressed by comparison, and the second is not good at masking this. Worse, the real order – to our ears – greatly emphasises the high harmonics, including some distinctly grating components above the tenth, not plotted here for clarity. While small, these are not necessarily masked by much higher, but paradoxically less audible lower order distortions created by loudspeakers.

In **Fig.10**, the current mirror (**Fig. 11a**) used to enforce close current balancing has been removed and replaced with Self's original arrangement[5], a single R_c in the forward leg. Here (**Fig.11b**) it is 3.9kΩ, which causes a major imbalance. The upper plot shows how the forward leg current ($I(2,10)$ seen lowermost, is about 250μA, starving the voltage amplification stage of current drive.

Although thd has increased (try summing the heights of the triangles) the sonic qualities will be different and, to many ears, more pleasant and more rounded. The reason is that the odd and even harmonics are almost evenly paired, with the exception of the recessed sixth and tenth. These same considerations have been used for centuries by musical instrument makers to adjust timbre.

Lastly, **Fig. 12** shows what happens when R_c is readjusted for near perfect balance without the current mirror. Note how the much-magnified current plots in the upper panel show how the current waveforms, while dc matched, are dissimilar in amplitude and harmonic content, making a nonsense of perfect quiescent matching.

Once again, the harmonics are different. The seventh is below the noise floor, the tenth is nearly masked, while the second is recessed. Overall, a hard, nasal sound would be predicted by the harmonically conversant.

So much for balancing.

Douglas Self's series on amplifiers provides a lucid insight into the nth degree static linearity of low frequency amplifiers for industrial loads, test equipment satisfaction and low common denominator 'consumer audio'.

But in the context of true high fidelity audio and what analogue electronics can offer to the unfettered reproduction of all kinds of music, it falls far short. ■

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2µ2 160V rad 22mm, 2µ2 100V rad 15mm .... 100/£10
10n/33n/47n 250V AC x rated 15mm .... 10/£1
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1µ0 100V rad 15mm, 1µ0 22mm rad .... 100/£6
0.22µ 250V AC X2 RATING .... 4/£1
0.22µ 900V .... 4/£1

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ISDN the worldwide network

Integrated services digital network – isdn – makes it possible for anyone with a telephone line to communicate digitised speech and data world-wide at up to 144kbit/s. Mike Button explains how.

The demand for data traffic in today's high technology world has fuelled an incredible growth in the requirement for data-communications equipment, which includes fax, telex, X25 packet switching, lans and modems.

Data communication requires either its own separate digital network or modems that convert digital information for transmission over analogue telephone lines. Modems give access via the world-wide telephone network, but are slow and inefficient. Separate digital networks are more efficient but expensive to install. In addition they are generally proprietary offering little or no equipment portability.

Digitization of the telephone network began in the 1960s on the trunk lines between public exchange switches. Around a decade later the first digital telephone exchanges appeared. Currently, the only part of the network remaining exclusively analogue is the subscriber line connecting the telephone to the public or private telephone exchange. The majority of today's telephone lines are still analogue, optimized for transmission in the voice spectrum of 300Hz to 3.4kHz, but this is changing.

The digital future

Tomorrow's telephone network, available now, will be digitized from end to end. It will provide integration of voice and data communications giving an efficient and compatible world wide network. Transmission rates will increase dramatically while errors decrease to a negligible level, making communications faster and more efficient.

More than one signal may be sent simultaneously on a single telephone line. Both voice and a wide variety of data services will be available to users over one network with standard interfaces and recognised set-up procedures without the expense of additional or special lines. Ultimately this will bring cheaper equipment and a wider range of services. This network is called isdn.

Standards for inter-connectability

The international telegraph and telephone consultative committee, CCITT – a United Nations organization began working on standards for isdn in 1978. It published a series of recommendations that have become a world-wide industry standard.

Fundamentals of the isdn structure are specified

I — **Integrated**
Telecommunications and data communications services together on one system.

S — **Services**
Voice, data, text, pictures, video, telex, facsimile, telemetry and alarms.

D — **Digital**
Transmission from end to end using digital techniques with voice digitization handled at the terminal.

N — **Network**
One world-wide network, based on existing public telephone lines, providing standard interfaces and call procedures.

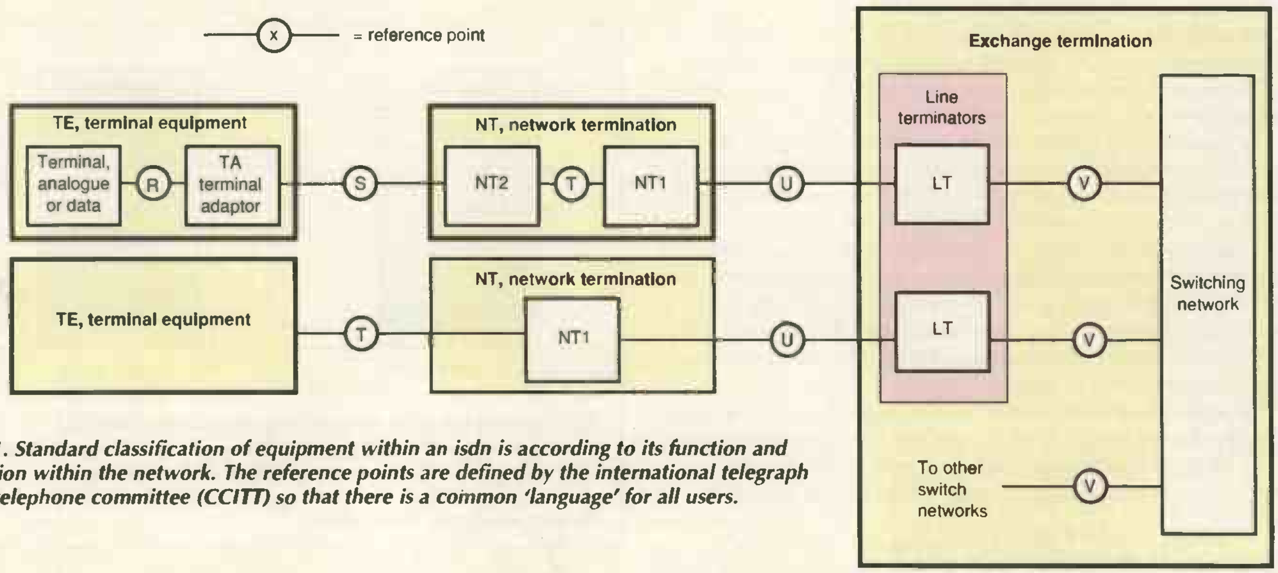


Fig. 1. Standard classification of equipment within an isdn is according to its function and position within the network. The reference points are defined by the international telegraph and telephone committee (CCITT) so that there is a common 'language' for all users.

in the CCITT I.411 recommendation which assumed a network architecture using the existing analogue telephone network. It is based on a 64kbit/s channel where the analogue voice signals are converted to an 8bit byte transmitted at 8kbytes/s. One of the features of isdn is that it retains, in general, this transmission rate and allows either voice or data to be carried on the channel.

The isdn definitions specify two classes of service each with a different transmission rate. See the panel on services for a description.

Isdn features

One of the main features of isdn is its flexibility. Once a connection has been made, two or more compatible terminals can process the data in any chosen manner. Analogue voice signals can be mixed with data in any combination but communicating terminals must know what each other is doing.

Both the basic rate and the primary rate transfer channels of data at an effective rate of 8kbytes/s. The basic rate handles an 18-bit word (18 × 8000 = 144K) and the primary rate a 248bit word (256 × 8000 = 2.048M). For example, the basic rate can be partitioned in ways other than '2B plus D'. Both the 'B' channels could be combined to give 16kbit data at 8kHz which would be adequate for a high definition slow scan television picture. Alternatively the full bandwidth of the primary rate can be employed to give a 1.92Mbit/s 'H12' channel. The permutations are endless.

Isdn architecture according to CCITT

In order that users may have a common language to express and define their requirements the CCITT has a set of definitions which all users can 'understand'. The hardware is divided into several segments called 'termination equipment' and the functionality of the network is defined by 'reference points'. Isdn configuration does not depend on whether the basic or primary rate is employed.

Figure 1 shows the topography of the network identifying specific classes of equipment at CCITT defined reference points. The isdn equipment, listed below, is classified by its function and location within the network.

Termination equipment. The exchange termination is the interface between the telephone switching network and other parts of the exchange. This includes the interface to the line termination and to other parts of the switching network. The line termination is located at the telephone exchange and performs osi physical layer functions for the 'B' and 'D' channels plus the osi layer two and three for the 'D' channel. See panel below for details on the osi reference model.

Network termination, *NT*, is divided into two types; type *NT1* performs such functions as line length extension and two-to-four wire conversion (U to S interface).

Termination *NT1* deals only with layer one of the osi model and as such has no intelligent logic. *NT2* types are intelligent devices that actively participate in call routing and control functions. They can be connected, simultaneously, to multiple isdn line types. Network terminations often form the boundary between

Isdn services

Basic rate. The basic rate runs at 144kbit/s and provides two 64kbit/s bearer channels for either voice or data and one 16kbit/s data channel for signalling or data. This '2B plus D' arrangement is the standard service provided to the user. This basic rate service is now available from BT as 'Isdn 2'.

Primary rate, Europe. The primary rate runs at 2.048Mbit/s and provides 30 bearer channels, one data channel and one synchronization/control channel – all at 64kbit/s. This '30B plus D' arrangement is provided when more traffic or data handling capacity is required. In North America and Japan the primary rate runs at 1.544Mbit/s and is configured as '23B plus D'. The primary rate is used to connect private branch exchanges to public exchanges or to interconnect basic rate services within and between exchanges. Private branch exchanges can be interlinked to form large private networks and computers can be interconnected via telephone networks. Local area networks have been developed using the primary rate to interface with other lans. This primary rate service is now available from BT as 'Isdn 30'.

The open system interconnect (osi) model for networks

In the early 1970s data communications had no recognised standard and hence there was no compatibility between product vendors. It soon became apparent, as the market for data communications grew, that it was in everyone's interest to have the capability to interconnect equipment from different manufacturers.

In 1978 the International Standards Organization (ISO) commenced work on the open system interconnect reference model to provide the framework for orderly communication across different data networks. The work of this committee was supported by the CCITT in their recommendation, X.200.

The model chosen by the ISO is a seven layer structure in which each layer provides particular logic services that belong to each other.

1 The physical layer, the lowest layer in the CCITT X.200 architecture, provides mechanical and electrical functions and procedures for the installation and maintenance of power supply, hardware clock and timing, provision of connectors and so on.

2 The data link layer residing immediately above the physical layer provides for the transfer of units of information between the two ends of the physical link such as framing, flow control and error detection. In isdn the CCITT Q.921, lapd (link access procedure d), is the key layer 2 protocol for signalling on the 'D channel'. Examples of data link layer protocols are – bisynchronous, and synchronous data link control (sdlc) of IBM's sna, digital communications message protocol (ddcmp) of DEC's Decnet, Ethernet's local area network (lan) IEEE standard. ISO's high level data link control (hdlc), a subset of which is known as lapb (link access procedure balance).

3 The network layer is responsible for addressing, switching and routing functions that are needed to set up a path for transparent transmission.

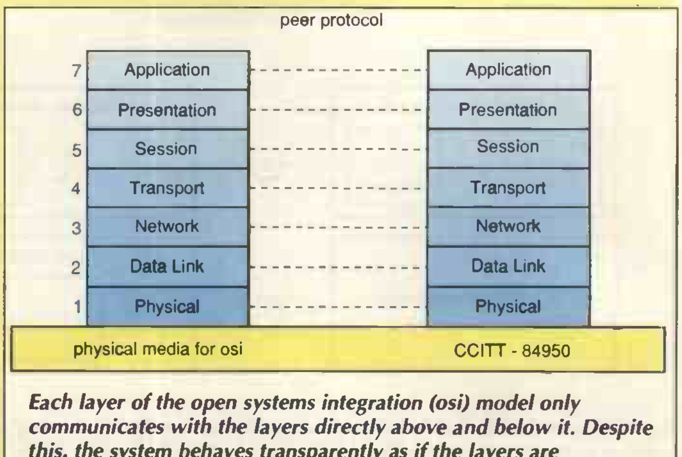
4 The transport layer is responsible for providing the required performance at a minimum cost based on the current state of the network.

5 The session layer co-ordinates the interaction of the application processes at each end. It controls session establishment, management and release together with error reporting.

6 The presentation layer provides for code conversion data between different types of termination.

7 The application layer is the layer that is application specific and communicates directly with its opposite number.

The main objective of the osi definitions is to simplify the communication between different layers. Each layer is defined to provide services to the next higher osi layer. Adjacent layers interact with predetermined requests and responses called primitives. At each layer there is a 'peer' protocol to govern the layer interaction.



Each layer of the open systems integration (osi) model only communicates with the layers directly above and below it. Despite this, the system behaves transparently as if the layers are connected directly to their counterpart at the receiving end, shown here by the dotted lines.

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the subscriber equipment and public exchange.

Terminal equipment can be telephone instruments, computers and faxes that are directly compatible with isdn. Terminal adapters are provided to connect non isdn compatible equipment to the network. They convert the communications protocol used by non isdn equipment to either the basic or primary rate. The *DP2000* isdn telephone, available from BT, is piece of terminal equipment with a combined terminal adapter for additional data or voice equipment.

Reference Points

Reference points identify the connection points between equipment classes only. They do not specify any implementation or protocol of this inter-connection. Reference point 'R' is the boundary between non isdn compatible equipment and the network. Reference point 'S' is the boundary between the *NT2* equipment and the terminal equipment/adapter. If there is no *NT2* equipment then the reference point 'S' will not exist. Reference point 'T' is the boundary between the *NT2* equipment and either the *NT1* or terminal equipment.

Reference point 'U' is the boundary between line termination and network termination.

Reference point 'V' is the boundary between the exchange switch and the line termination (LT). The 'R' reference point is defined as an interface between the isdn and a non isdn world. Various standard data interfaces are possible at the 'R' point, for example RS232, X.25 and X.21. A non-isdn subscriber terminal needs a terminal adapter (TA) to connect to the isdn world. The S-interface is the user interface into the isdn network.

Termination of the public network, *NT1*, ensures the conversion between the U and the T interfaces. These functions are bit rate conversion, clock and frame re-synchronization, frame conversion and power supply. *NT2* is an intelligent module which carries out the conversions between the T-interface and the S-user interface. It also provides the interconnection between the subscriber terminals both at the physical layer and the software level. When the S and T interface correspond to the same port, *NT2* is not needed. *NT1* and *NT2* may also be combined to ensure a direct conversion from the public network (U-interface) to the user interface (S-interface).

Integrated circuits for isdn

There is a host of integrated circuits available from different manufacturers. To give a list and function summary of all types is beyond the scope of this article. For more information, interested parties should contact either the manufacturer direct or one of the many distributors. Mitel, AMD, Philips and GEC have

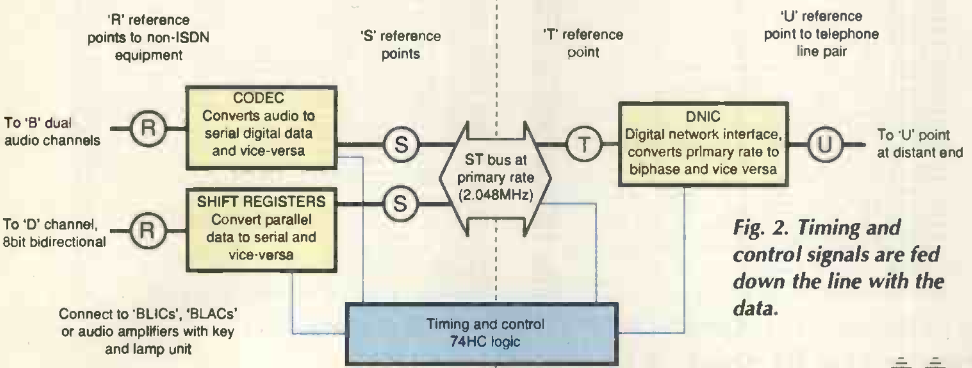

Fig. 2. Timing and control signals are fed down the line with the data.

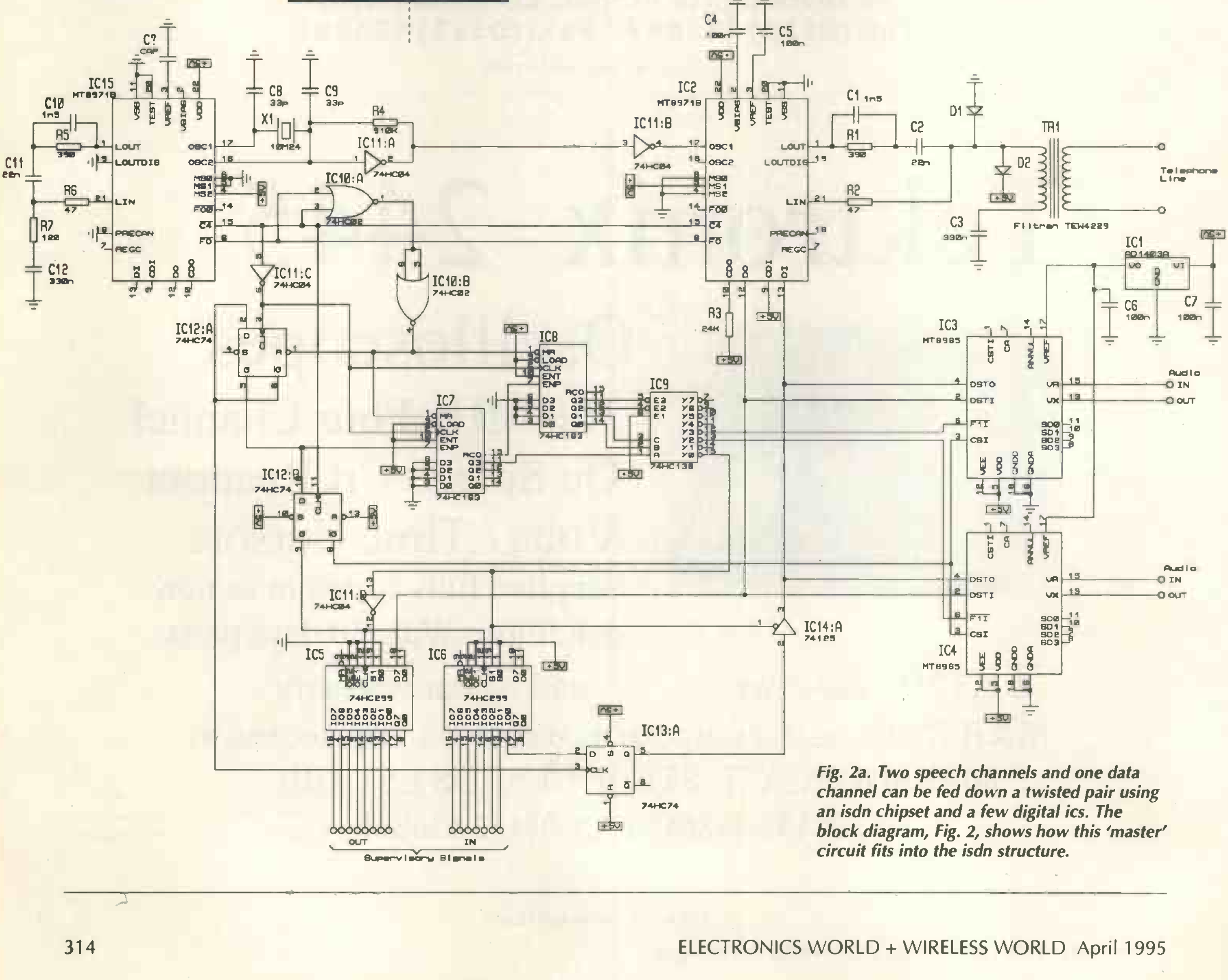

Fig. 2a. Two speech channels and one data channel can be fed down a twisted pair using an isdn chipset and a few digital ics. The block diagram, Fig. 2, shows how this 'master' circuit fits into the isdn structure.

data books on their range of isdn integrated circuits and I have found the distributor, Electronics 2000, very helpful. The more common circuits are given below.

Coder decoder. A codec provides the conversion between the voice-band analogue signals from the telephone instrument and the digital signals required by a pcm (pulse code modulation) signal as used by isdn. It is manufactured as a single integrated circuit incorporating digital to analogue and analogue to digital converters operating at a sample rate of 8kHz to accommodate the 300Hz to 3.4kHz audio bandwidth. Codecs from different manufacturers vary slightly in function but they all have the following features in common:

- Operation at basic and/or primary rate.
- Analogue input with corresponding digital output.
- Digital input with corresponding analogue output.
- Chip select or synchronization input.
- A non-linear analogue to digital companding, A-law or µ-law.

In order to maintain an acceptable dynamic range the analogue signal is converted to or from an eight-bit digital signal in a non linear manner. There are two companding (conversion) laws presently *en vogue*. Generally the A-law is used in Europe and the µ-law in North America, Table 1.

Subscriber line interface circuit, slic

This device, normally a hybrid circuit, provides all the 'borsch' functions (battery feed,

Table 1. European and North American codec companding laws convert the analogue signal to an eight-bit digital signal, made up of a sign bit, (S), three cord bits, (C) and four step bits (D).

Companded digital signal									European A-law		North American µ-law	
7	6	5	4	3	2	1	0		Chord Value (mV)	Step Value (mV)	Chord Value (mV)	Step Value (mV)
S	C	C	C	D	D	D	D					
	0	0	0					0	0.0	1.221	0.00	0.613
	0	0	1					1	20.1	1.221	10.11	1.226
	0	1	0					2	40.3	2.44	30.3	2.45
	0	1	1					3	80.6	4.88	70.8	4.90
	1	0	0					4	161.1	9.77	151.7	9.81
	1	0	1					5	332.0	19.53	313.0	19.61
	1	1	0					6	645.0	39.1	637.0	39.2
	1	1	1					7	1289.0	78.1	1284.0	78.4
Examples									Cord = 3			
S	C	C	C	D	D	D	D					
1	0	1	1	0	1	0	0		80.6+(4X4.88)		70.8+(4X4.9)	
+	3				4				=+100.12mV		+90.4mV	
									Cord = 2			
0	0	1	0	0	1	1	1		40.3+(7X2.44)		30.3+(7X2.45)	
–	2				7				–57.38mV		=–47.45mV	

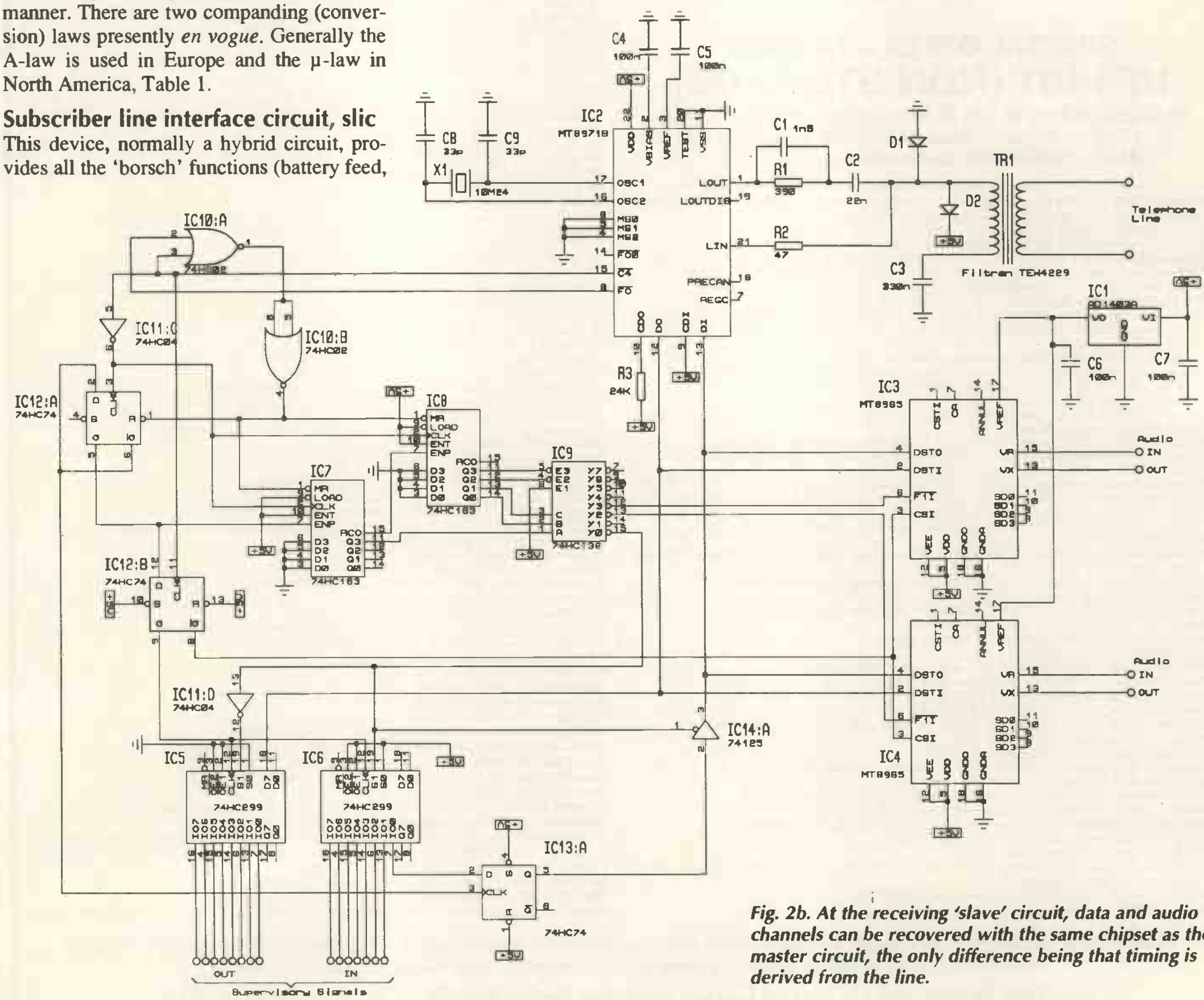

Fig. 2b. At the receiving 'slave' circuit, data and audio channels can be recovered with the same chipset as the master circuit, the only difference being that timing is derived from the line.

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HP8445B Tracking Pre-selector DC – 18GHz – £400-£600 or HP8445A – £250.

HP8444A Tracking Generator – £750 – 1300Mc/s.

HP8444A Opt 059 Tracking Generator – £1000 – 1500Mc/s.

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Marconi TF2008 – AM-FM signal generator – also sweeper – 10Kc/s – 510Mc/s – from £250 – tested to £400 as new with manual – probe kit in wooden carrying box.
HP Frequency comb generator type 8406 – £400.
HP Vector Voltmeter type 8405A – £400 to £600 – old or new colour.
HP Sweep Oscillators type 8690 A & B + plug-ins from 10Mc/s to 18GHz also 18-40GHz. P.O.R..
HP Network Analyzer type 8407A + 8412A + 8501A – 100Kc/s – 110Mc/s – £500 – £1000.
HP Amplifier type 8447A – 1-400Mc/s £200 – HP8447F. 1-1300Mc/s £400.
HP Frequency Counter type 5340A – 18GHz £1000 – rear output £800.
HP 8410 – A – B – C Network Analyzer 110Mc/s to 12GHz or 18GHz – plus most other units and displays used in this set-up – 8411a – 8412 – 8413 – 8414 – 8418 – 8740 – 8741 – 8742 – 8743 – 8746 – 8650. From £1000.
Racal/Dana 9301A – 9302 RF Millivoltmeter – 1.5-2GHz – £250-£400.
Racal/Dana Counters 9915M – 9916 – 9917 – 9921 – £150 to £450. Fitted FX standards.
Racal/Dana Modulation Meter type 9009 – 8Mc/s – 1.5GHz – £250.
Marconi RCL Bridge type TF2700 – £150.
Marconi/Saunders Signal Sources type – 6058B – 6070A – 6055A – 6059A – 6057A – 6056 – £250-£350. 400Mc/s to 18GHz.
Marconi TF1245 Circuit Magnification meter + 1246 & 1247 Oscillators – £100-£300.
Marconi microwave 6600A sweep osc., mainframe with 6650 PI – 18-26.5GHz or 6651 PI – 26.5-40GHz – £1000 or PI only £600. MF only £250.
Marconi distortion meter type TF2331 – £150. TF2331A – 200.
Tektronix Plug-Ins 7A13 – 7A14 – 7A18 – 7A24 – 7A26 – 7A11 – 7M11 – 7S11 – 7D10 – 7S12 – S1 – S2 – S6 – S52 – PG506 – SC504 – SG502 – SG503 – SG504 – DC503 – DC508 – DD501 – WR501 – DM501A – FG501A – TG501 – PG502 – DC505A – FG504 – 7B80 + 85–7B92A
Gould J3B test oscillator + manual – £200.
Tektronix Mainframes – 7603 – 7623A – 7613 – 7704A – 7844 – 7904 – TM501 – TM503 – TM506 – 7904 – 7834 – 7104 – 7623 – 7633.
Alltech 757 Spectrum Analyser – 001 22GHz – Digital storage + readout – £2000.
Marconi 6155A Signal Source – 1 to 2GHz – LED readout – £400.
Barr & Stroud Variable filter EF3 0.1Hz – 100kc/s + high pass + low pass – £150.
Marconi TF2163S attenuator – 1GHz. £200.
Farnell power unit H60/50 – £400 tested. H60/25 – £250.
Racal/Dana 9300 RMS voltmeter – £250.
HP 8750A storage normalizer – £400 with lead + S.A or N,A Interface.
Marconi TF2330 – or TF2330A wave analysers – £100-£150.
Racal/Dana signal generator 9082 – 1.5-520Mc/s – £500.
Racal/Dana signal generator 9082H – 1.5-520Mc/s – £600.
Tektronix – 7S14 – 7T11 – 7S11 – 7S12 – S1 – S2 – S39 – S47 – S51 – S52 – S53 – 7M11.
Marconi mod meters type TF2304 – £250.
HP 5065A rubidrum vapour FX standard – £2.5k.
Systron Donner counter type 6054B – 20Mc/s – 24GHz – LED readout – £1k.
Racal/Dana 9083 signal source – two tone – £250.
Systron Donner – signal generator 1702 – synthesized to 1GHz – AM/FM – £600.
Tektronix TM515 mainframe + TM5006 mainframe – £450 – £850.
Rhodes & Schwartz power signal generator SLRD-280 – 2750Mc/s – £250-£600.
Ball Efratom rubidrum standard PT256B-FRKL – £1000.
Farnell electronic load type RB1030-35 – £350.
Racal/Dana counters – 9904 – 9905 – 9906 – 9915 – 9916 – 9917 – 9921 – 50Mc/s – 3GHz – £100-£450 – all fitted with FX standards.
HP4815A RF vector impedance meter c/w probe – £500-£600.
Marconi TF2092 noise receiver. A, B or C plus filters – £100-£350.

Marconi TF2091 noise generator. A, B or C plus filters – £100-£350.
HP180TR, HP182T mainframes £300-£500.
Fluke 8506A thermal RMS digital multimeter. £400.
Philips panoramic receiver type PM7900 – 1 to 20GHz – £400.
Marconi 6700A sweep oscillator + 6730A – 1 to 2GHz – £500.
HP8505A network ANZ + 8503A S parameter test set + 8501A normalizer – £4k.
Racal/Dana VLF frequency standard equipment. Tracer receiver type 900A + difference meter type 527E + rubidium standard type 9475 – £2750.
HP signal generators type 626 – 628 – frequency 10GHz – 21GHz.
HP 432A – 435A or B – 436A – power meters + powerheads – Mc/s – 40GHz – £200-£1000.
Bradley oscilloscope calibrator type 192 – £600.
Barr & Stroud variable filter EF3 0.1Hz – 100Kc/s + high pass + low pass – £150.
Marconi TF2370 spectrum ANZ – 110Mc/s – £900.
Marconi TF2370 spectrum ANZ + TK2375 FX extender 1250Mc/s + 1st gen – £1.5k.
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HP 8640B S/G AM-FM 512Mc/s or 1024Mc/s. Opt 001 or 002 or 003 – £800-£1250.
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HP 4437A Attenuator – £150.
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Philips PM5390 RF syn – 0.1 – 1GHz – AM + FM – £1250.
Philips PM5519 Colour T.V. pattern generator – £250.
S.A. Spectral Dynamics SD345 spectrascope 111 – LF ANZ – £2500.
Tektronix R7912 Transient waveform digitizer – programmable – £400.
Tektronix 496 Analyzer 1kHz – 1.8GHz – £3.5k.
Tektronix TR503 + TM503 tracking generator 0.1 – 1.8GHz – £1k – or TR502.
Tektronix 576 Curve tracer + adaptors – £900.
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Tektronix 1502/1503 TDR cable test set – £1000.
Tektronix 7L5 LF analyser – 0 – 5Mc/s – £800. OPT 25 – £1000.
Tektronix AM503 Current probe + TM501 m/frame – £1000.
Tektronix SC501 – SC502 – SC503 – SC504 oscilloscopes – £75-£350.
Tektronix 465 – 465B – 475 – 2213A – 2215 – 2225 – 2235 – 2245 – 2246 – £250-£1000.
Kikusui 100Mc/s Oscilloscope COS6100M – £350.
Farnell PSG520 Signal generator – £400.
Nicolet 3091 LF oscilloscope – £1000.
Racal 1991 – 1992 – 1988 – 1300Mc/s counters – £500-£900.
Tek 2445 150Mc/s oscilloscope – £1400.
Fluke 80K-40 High voltage probe in case – BN – £100.
Racal Recorders – Store 4 – 4D – 7 – 14 channels in stock – £250 – £500.
Racal Store Horse Recorder & control – £400-£750 Tested.
EIP 545 microwave 18GHz counter – £1200.
Fluke 510A AC ref standard – 400Hz – £200.
Fluke 355A DC voltage standard – £300.
Schlumberger 5229 Oscilloscope – 500Mc/s – £500.
Solartron 1170 FX response ANZ – LED dislay – £280.
Wiltron 610D Sweep Generator + 6124C PI – 4 – 8GHz – £400.
Wiltron 610D Sweep Generator + 61084D PI – 1Mc/s – 1500Mc/s – £500.
Time Electronics 9814 Voltage calibrator – £750.
Time Electronics 9811 Programmable resistance – £600.
Time Electronics 2004 D.C. voltage standard – £1000.
HP 8699B Sweep PI YIG oscillator .01 – 4GHz – £300. 8690B MF – £250. Both £500.
Schlumberger 1250 Frequency response ANZ – £2500.
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overvoltage protection, ringing feed, line supervision and 2 to 4 wire conversion) needed to connect a telephone instrument to a telephone line. It receives the audio signals from the codec and is controlled by the 'D' channel.

Subscriber line audio circuit, slac

This device, normally a hybrid circuit, is called a central office interface by the Americans. It provides a complete audio and signalling link between a telephone exchange line circuit and a telephone line. It looks and behaves like a telephone instrument providing loop seize and ringing detector circuits. It takes and receives the audio signals from the codec and is controlled by the 'D' channel.

Digital network line circuits (dnic, dsic, dsc, isac-s)

The connection between the telephone exchange and an isdn terminal is normally a twisted pair line. To transmit isdn data over this medium some form of electrical conversion is needed to enable both data detection and synchronization. A number of interface circuits have been devised to provide a variety of signalling codes over both two-wire and four-wire circuits. Some of the more common signalling codes used by these devices are described below.

2B1Q – Two binary to one quaternary. This takes two bits of isdn code and converts them to a four level signal. It halves the transmitted baud rate.

4B3T – Four binary to three ternary. This takes four bits of isdn code and converts to a three level signal. It reduces the transmitted baud rate by three quarters.

Biphase. Converts the isdn signal into level transitions. (Binary '0'= falling edge and binary '1'= rising edge, similar to 'Manchester code'). It does not reduce the transmitted baud rate but does have the feature of concentrating the power spectrum in a relatively narrow bandwidth.

Data rate adapters. These provide the means whereby low speed signals such as RS232 can be converted to the basic rate. They have similar functions to a codec except that they handle serial/parallel data signals instead of analogue signals.

Digital Switches. Within a telephone exchange digital switches are used to cross connect basic and primary rate signals.

Circuit example

Most integrated circuit manufacturers now produce chip sets to provide isdn facilities. It is now relatively easy to use this technology to produce equipment to access isdn.

The circuit example given in **Figs 2(a)** and 2(b), taken from an existing design, show how a single twisted pair line can be used to carry two speech channels and one data channel at distances up to one kilometre. Figure 2 gives details of how this circuit fits into the isdn scheme. Figures 2(a) and 2(b) give actual circuit details. This circuit pair could be used, for example, to enhance the facilities of a field telephone at fêtes and galas.

For simplicity, circuits for the termination equipment are not given. The analogue signal inputs and outputs could be connected to buffer amplifiers to drive operator's headsets or two slic standard telephone instruments may be connected at each end.

The data inputs and outputs may be connected, via suitable buffers, to a key and lamp unit or directly to a slic control input.

Integrated circuits used to provide isdn facil-

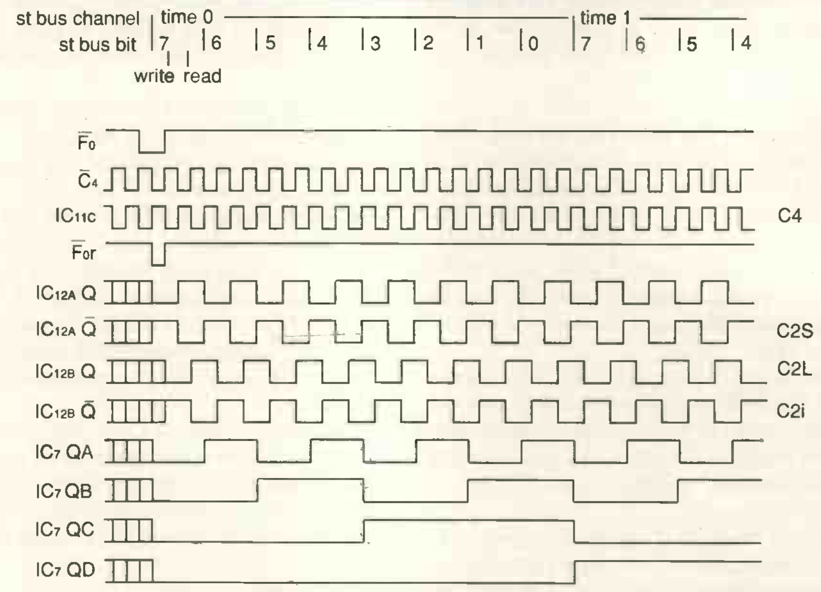

Fig. 3. Timing is synchronised with the frame-start signal on the Mitel st bus. A 2.048MHz quadrature signal is generated by IC12 to control the codec.

Design overview

The dnic operates in either a 'master' or 'slave' mode. When used as a 'slave' all timing and framing information is derived from the line signals and these pins are configured as outputs. The distant end 'master' must generate these timing signals and therefore needs the clock and framing signals as inputs. At the 'master' end a second dnic provides a low cost source for these signals. Apart from these differences the two circuits have the same logic and control circuit.

The dnic assumes the data input (DI pin) and output (DO pin) are at the primary rate and uses the well defined Mitel st bus. The channels used for the speech and signalling data are predefined by the dnic. Channel '0' is used for the signalling data and channels '2' and '3' for the analogue data. (Channel '1' is allocated to dnic control functions not employed in this example.)

The complex analogue signals on the line pair are fed via the transformer TR_1 into the L_{IN} and L_{OUT} of the dnic (IC_2). These are in the frequency range of 10kHz to 500kHz. The network comprising R_1, R_2, C_1 and C_2 provides the necessary feedback to enable the on-board digital signal processor to control the internal phase locked loop circuit and obtain timing and framing signals. The 10.24MHz signal is either generated by a phase locked loop controlling a crystal oscillator at the 'slave' end or via IC_{15}, double buffered by IC_{11A} and IC_{11B}, at the 'master'end.

The frame start signal on the st bus, $\overline{F_0}$ is gated via IC_{10A} and IC_{10B} with the C_4 clock to produce a frame start reset. The connection on the dnic C_4 pin is fed, via an inverter, into a d-type flip-flop IC_{12A} which, together with IC_{12B}, generates a quadrature 2.048MHz signal. The output of IC_{12B} is connected to *74163* counters IC_7 and IC_8 which, with the *74138* decode IC_9, generate the Mitel st bus timing signals. A timing diagram is given in **Fig. 3**.

Timing uncertainty only applicable during the period when there is no synchronization between the master and slave end. Output $\overline{Q}$ of IC_{12B} is connected to the codec C2I input and controls the codec (*MT8965*). A rising edge causes the codec to present data to the DI bus and the falling edge causes the Codec to read signals from the DO bus. Output Q of IC_{12B} clocks the signals in and out of the shift registers (*74299*). Output $\overline{Q}$ of IC_{12A} clocks the *7474* buffers at the start of each st bus bit. Outputs of the 4 bit counters IC_7 and IC_8 (*74138*) are connected to a three-to-eight bit decoder, IC_9 (*74138*), which generates the channel time signals for the st bus. Channel time 0 (IC_9 Y_0) is the data channel. ('D' port). Channel time 2 (IC_9 Y_2) is the first audio channel (B_1 port). Channel time 3 (IC_9 Y_3) is the second audio channel (B_2 port).

ities are supplied by Mitel; a brief description of these circuits is given below. Full details of the dnic and codec are abailable in data sheets of the devices, available from Mitel or their distributors.

Digital network interface circuit, dnic

The *MT8971B* device is intended for use as an interface for the integrated services digital network. It may be used in practically any application requiring high speed basic rate duplex data transmission over two wires. It is a multifunction device capable of providing transfer of speech or data over a distance of a kilometre or more.

With the more expensive *MT8972B* and a loop extender circuit distances of up to seven kilometres can be achieved. An adaptive echoing technique is used by the on-circuit digital signal processor to transfer data in a '2B' + 'D' format.

Several modes of operation are available. In this example, modes two and six are used for transferring speech and data in master and slave mode. In order to provide framing and control signals, extra bits are added to the serial data steam and the data plus synchronisation signals are transmitted at 160kbit/s.

The *MT8965* coder decoder

This device provides the conversion interface between the voiceband analogue signals and the digital signals required in a digital pcm-switching system. An 'A-law' encoding and decoding is provided to CCITT specifications.

Line signalling

The line signalling 'D' channel is controlled by two universal shift registers IC_5 and IC_6. Output Y_0 from IC_9 performs the following functions.

The tri-state buffer is enabled which presents the output of IC_{13A} to the di bus.

IC_6 is changed from the parallel load mode to the 'shift right' mode.

Next, IC_5 is changed from the 'hold' mode to the 'shift left' mode.

At the rising edge of output Q of IC_{12A} (which occurs at the start of each bit period) data from IC_6 is clocked into the 'd-type latch' IC_{13A} (*7474*), the output of which is then presented to the DI bus via the enabled tri-state buffer IC_{14A}.

At the rising edge of Q of IC_{12A} which occurs three quarters through each bit period, $IC_{5\&6}$ are clocked and the data present is shifted left once. This process continues during the period that output Y_0 from IC_9 is active (0V).

Availability

BT is currently offering an isdn service to its customers. Both basic rate and primary rate services are available, together with compatible terminal equipment. Akin with the early days of the Post Office telephone service, the installation and rental charges will prohibit installation of isdn to all those but the most affluent or those with a real need.

The installation charge for a single line basic rate service is £400 with a quarterly rental of £84. The provision of a primary rate service starts at an installation charge of £594 with a quarterly rental of £508. A telephone type terminal suitable for these services is £699.

All these charges are excluding vat. Full information is available from BT on their freephone number 0800 181514.

Prospective users with a high volume of data traffic will see savings on call connection charges as data can be sent at a much higher rate than currently available from modems. This, of course, assumes the distant end also has the isdn service connected. ■

Mike Button is a consultant design engineer with TDR Ltd. He can be contacted on 01666 577 464.

Further reading

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Isdn Protocol Reference Model, CCITT Recommendation I.320.
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Low distortion attenuator for hi-fi

With a distortion figure of just 0.005% over most of the audio range, this switch-mode attenuator designed by John Linsley Hood removes the transistor matching problems normally associated with high-performance attenuators.

A number of circuit arrangements have been proposed to allow the magnitude of an audio signal to be controlled by an externally applied dc voltage. Most of them, however, introduce a relatively high level of harmonic distortion into the signal being controlled.

A typical layout, using a junction fet as a voltage controlled resistor in one limb of a resistive attenuator, is shown in **Fig. 1**. Unfortunately the channel resistance is modulated by changes in the gate-drain potential. This leads to substantial signal distortion.

The classic method of minimising this is by introducing some negative feedback from the drain circuit. A typical layout for this, from Siliconix application note *AN73-1*, is shown in **Fig. 2**. The resistor ratio suggested by Siliconix is $R_2 = R_3 \geq (10R_1 // r_{ds(max)} // R_L)$. This reduces the distortion at 1V rms and 1kHz from 10-15% to 1-1.5%. Although this value decreases as the signal level is reduced, it is not good enough for hi-fi applications, where ideally a thd figure of less than 0.01% is sought.

Figure 3, from Precision Monolithics application note *AN-105, 1/86* is a much better arrangement. In this design, input signal voltage is caused to modulate the stage current of two matched long-tailed pairs comprising a matched quad-transistor array. Even-order harmonic distortion, caused by the curvature of the transistor V_b/I_c characteristics, is minimised by driving the two halves of the circuit in push-pull.

A thd of less than 0.03% is claimed, at an unspecified signal level or operating frequency. Experimentation indicates that achieving this performance demands a very high degree of matching of all four transistors.

Using two separate monolithic matched pairs is not satisfactory. The Precision Monolithics circuit philosophy is also applied, in simplified form, in the Motorola *MC3340*. At one time, this device was widely used as the basis for remote volume controls in televisions. It was claimed that the IC had a thd figure of 0.6% at less than 0.5V rms. Unfortunately, this performance is only achievable at low attenuation levels. At 50dB attenuation the quoted thd has worsened to greater than 3%.

A better and more modern circuit is the National Semiconductors *LM1040* for which a thd figure, at 300mV and 1kHz, of 0.06-0.03% is claimed. This device is intended for use as a voltage controlled gain, tone and balance sys-

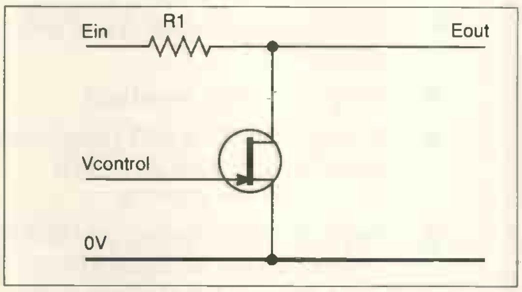


Fig. 1. Simple attenuators made from fets suffer from channel resistance modulation due to changes in gate-drain potential. The result is substantial distortion.

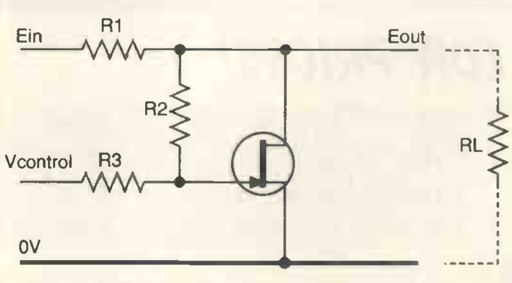
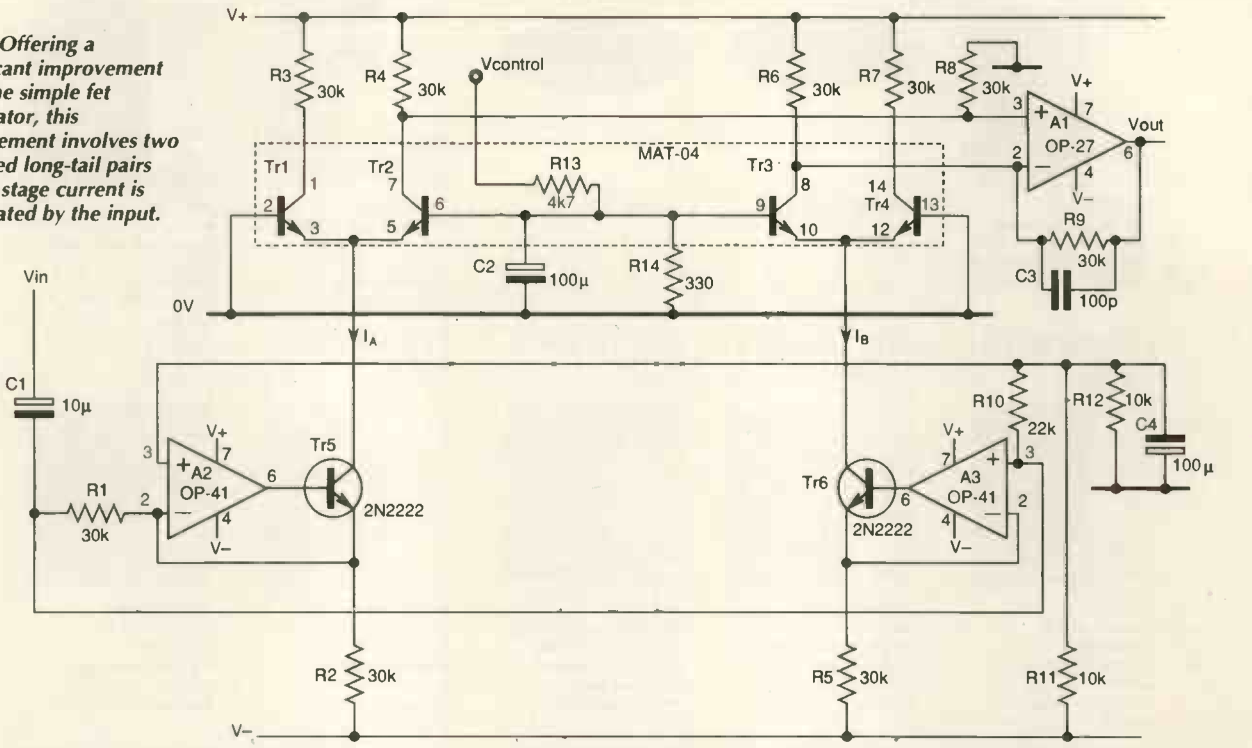


Fig. 2. Introducing negative feedback from the drain helps linearise the fet attenuator, but this does not reduce distortion to levels acceptable for hi-fi.

Fig. 3. Offering a significant improvement over the simple fet attenuator, this arrangement involves two matched long-tail pairs whose stage current is modulated by the input.



tem for an audio pre-amplifier. It incorporates a very similar balanced long-tailed pair gain control circuit to the PMI *MAT-04*.

Problems raised by the necessity for matched components can be avoided by using the switch-mode attenuator system shown in **Fig. 4**. Here, the output signal level after removing the switching waveform depends on the ratio of the 'open' to 'closed' durations of the switch.

If the switching frequency is sufficiently high, the removal of the switching artifacts can be done without the need for excessively steep-cut low-pass filtering. In the circuit proposed the chopping frequency is 130kHz. This is adequately separated from either the 44.1kHz cd sampling frequency or the 176kHz four-times over-sampling frequency used in cd players to avoid possible trouble on this account. On the other hand, the chopping frequency is high enough to permit the full 20Hz to 20kHz audio pass-band to be transmitted.

A rectangular-wave signal with adjustable mark-to-space ratio can be generated by a hex inverter IC. **Figure 5** shows a configuration with a *CD4060* hex inverter running from a +12V supply line. In this example, IC_1 and IC_2 form a bistable latch. This latch is caused to flip backwards and forwards between its two stable modes when the phase-inverted triangular waveform generated by IC_3 overrides the voltage derived from IC_2 via R_1.

The resulting triangular voltage waveform and an input dc control voltage are summed at the input of a chain of inverters, $IC_{4\text{-}6}$. This produces a rectangular waveform output whose mark-to-space ratio can be made to lie within the range 5%:95% to 95%:5% depending on the dc input voltage and the relative values of R_4 and R_5. Output waveform is then fed to a switching fet, such as the *NS J111*, to give the final voltage controlled attenuator layout shown schematically in **Fig. 6**.

At 1V rms, over the range 100Hz to 10kHz, the thd was of the order of 0.005%, and not

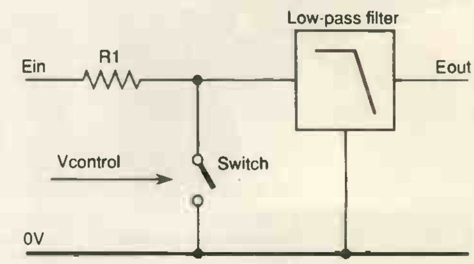


Fig. 4. The switch-mode attenuator offers high performance without the need for matched components .

greatly increased at higher attenuation levels. The effect of the circuit on a 1kHz square wave input was only that to be expected from the pre- and post-chopping low-pass filtration. In the prototype, this was provided by a cascaded pair of third order unity-gain Sallen and Key low-pass filters with an f_T of 23kHz. ■

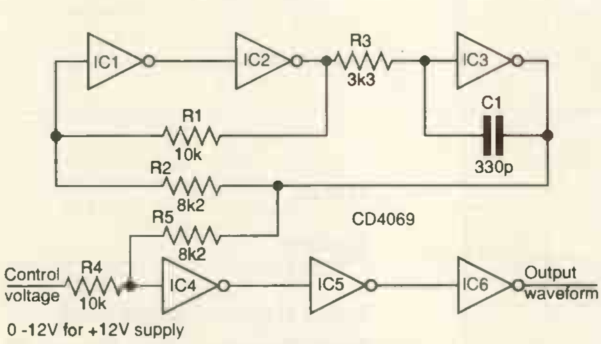


Fig. 5. A cmos hex inverter chip is ideal for producing the rectangular waveform needed to drive the chopping attenuator.

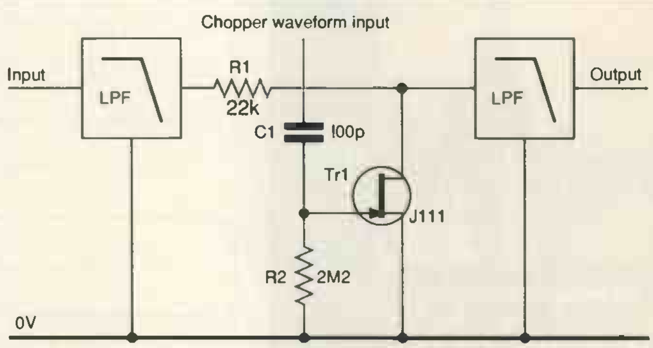


Fig. 6. Over the range 100Hz to 10kHz, this high-performance chopping attenuator for audio produces a thd figure in the region of 0.005% for a 1V rms input.

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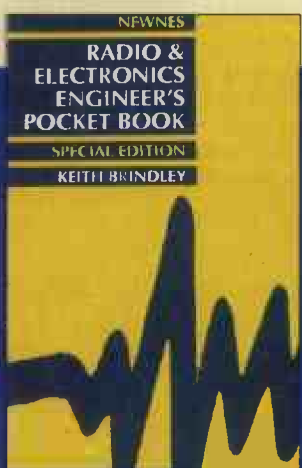

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LETTERS

Letters to *"Electronics World + Wireless World"* Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.

Statistically significant

In his reply to my letter (February, *EW + WW*) Douglas Self questioned the "statistical evidence" that mosfets in general have lower distortion than bipolars, in reference to audio circuitry: I thought he'd never ask.

Take first the *Hitachi power mosfet data book 1-15a* in which two 100W amplifiers are described, one of hybrid design, the other, all fet.

The hybrid offers 0.01% distortion at 100kHz, but within the audio range offers 0.002% at 1kHz, and 0.003% at 20kHz.

The all fet version offers 0.01% distortion at 50kHz and within the audio band falls to under 0.002% at 1kHz and about 0.0035% at 20kHz. But these distortion figures are largely unchanged at full power, at the rated output of the amplifiers: 100W, 8Ω load.

There is also the series of amplifiers by J P Rimmer, formerly of Pantechnic. These used Hitachi type circuitry but with selected transistors in the driver stage in which the overall distortion was slightly better than the Hitachi version. Results were 0.002% at 1kHz and better than 0.005% anywhere in the audio band (20Hz - 20kHz) and any power up to 150W for the *PFA200* version.

A confirmation of this kind of performance was given in a review by Gordon J King (*ETI*, June 1978) of the Hitachi commercial amplifier *HMA7500* (80W/channel). It was severely tested by King using 15kHz/16kHz tones into 5Ω modules of impedance and 60° phase angle at 16kHz (for intermodulation tests). The two tone signal at the output of the amplifier had a peak value of 28V across the complex load.

Second order product of the difference frequency (1kHz) was –78dB or 0.012%, while distortion measured under normal testing was typically –90dB, 0.003% at 80W.

Moving on to the *ETI5000* design by Dave Tilbrook (*ETI*, June, 1982), distortion is almost non-existent at 0.0007% at 1kHz and 100W and 0.003% at 10kHz and 100W.

John Linsley Hood's mosfet design (*ETI*, July, 1984) showed a distortion at 80W (full power) of –94dB at 1kHz, about 0.002%, and at 10kHz 0.021% at 80W. Hood's amplifier had a greater input sensitivity than others quoted and would probably give better distortion figures, gain for gain. Also, the description implies that the low levels of distortion were mainly from the signal source itself.

Excellent linearity was demonstrated by the David Hafler *DH200*, *DH220*, *DH500* series of amplifiers. The oldest, *DH200*, offered 0.0015% at 1kHz, 0.005% at 10kHz and 0.01% at 20kHz, 100W, 8Ω load. The design is now some 20 years old, and figures would be improved with a more substantial power supply.

The American Company, Musical Design – where circuits bore a remarkable similarity to the David Hafler, except for fet drivers – offered low distortion from its *D140* and *D180* amplifiers: typically 0.006% at 1kHz and 140W.

Moving on to Sage Audio and the Supermos series, these amplifiers were exceptional at 0.0002%, 1kHz and rated power.

Of course some designs have given poorer distortion figures. But the statistical evidence is overwhelmingly in favour of the lower distortion variety.

Distortion in bipolar amplifiers is much more markedly distinguished with increased power and frequency.

One amplifier that should be considered is that by Edward Cherry whose 60W nested differential feedback loop amplifier (*ETI*, May, 1983) produced 0.002% at rated output and 1kHz, falling to 0.015%, 2nd *and* 3rd harmonic at 6kHz. But the circuit was complicated and required special earthing arrangements on the board. By contrast the simple Hitachi mosfet configuration must be a clear winner in every respect, barring the cost of mosfets.

All the circuits I have described, except the Sage modules, have used Hitachi mosfets and as such have been set for 100mA bias current per pair of output devices. The Hafler circuits used slightly more quiescent at 275mA for two pairs. In my earlier letter I suggested that the low 45mA quiescent value in Self's mosfet circuit would be significantly contributing to the crossover distortion artefacts observed.

I also can not agree with Self's comment regarding the independence of slew-rate limit and bandwidth. In absolute terms it is difficult to separate one from the other. Any amplifying device fed with a signal of changing rate will suffer from an upper limit at which it can reproduce that rate of change, thus encompassing both bandwidth and slew-rate. There is surely no mystery in mosfet amplifiers having vastly superior slew rates simply because they can be used at much higher frequencies (implying wider bandwidths).

Returning to distortion and early rate of roll off of frequency in bipolars, the major source of distortion comes from the relationship between current gain and collector current. In power bipolars this is a graph shaped like a parabola – ie very non-linear, showing the gain varying according to the collector current. Variation is marked, offering quite a gain spread, with the lowest at low and high currents and highest at intermediate currents. Added to this is the relationship between gain and frequency in which the power bipolar rapidly loses gain as the frequency increases. Using typical figures of F_T = 2MHz and gain of 100 this device will only offer a gain of 1 at 20,000Hz. Clearly, as the frequency rises the normal overall feedback decreases, and so distortion rises.

The situation is quite different with mosfets. Here the V_g/I_d characteristic is very linear and since the devices are much faster they have less trouble coping with high frequencies – provided they are driven by low impedances. Distortion tends to fall as the drain current (I_d) increases (run at 100mA+ please). Bipolars also suffer 'notch' distortion due to storage of minority charge carriers.

I have said before that in my experience, going back 25 years of reading test reports from most of the major journals, I have found with few exceptions that amplifiers containing bipolar output devices regularly roll off at around 15kHz at full power, many starting before this frequency. This experience is a matter of public record and the evidence is there for anyone who cares to sift through all of the reports in the same way I have done.

The reasons for this behaviour are, as I have shown, that output devices are being selected which are simply too slow: that is also a statistical fact. Faster devices are less rugged and more expensive.

I have most of the articles and literature referred to for mosfet amps if further doubt is expressed.

V J Hawtin
Middlesex

Shifted search

Can any reader help me? I have been looking, with little success, for a number of months for a frequency shifter circuit diagram.

My aim is to construct a frequency shifter with a microphone input, having the capacity to shift the frequency down to 1Hz and up to 25,000Hz. The output would be to a socket so that a loudspeaker or recording equipment could be plugged in. Stepping of the shifted frequency should be smooth not doubled or in octaves etc, and I need the circuit diagram to be fairly idiot proof – I am a fairly competent constructor but no expert.

PH Cope
6 Marlborough House
Langton Close
Lincoln
LN2 2HJ

Defence or attack

If it were true, as Colin Long (*Letters*, January) seems to imply, that high military spending makes good economic sense then we might reasonably expect to see the major industrialised military economies outperforming their non-military competitors.

In fact we see the reverse, because such activity is bad for any economy, whether an autocratic command system of the (old) USSR or the democratic demand economy of the USA. Even with rich oil and mineral reserves, they are so obviously outperformed by Germany and Japan, as indeed is the UK.

Mr Long questions why the MoD has come in for such criticism. It is because it has compounded the

problem through damaging interference. Using a variety of devices, including its 'massive purchasing power', the MoD has taken a vigorous and broad-based electronics industry and reduced it to a single, essentially military, supplier. An industry that once supplied a rich variety of products to its domestic market is now reduced to a critical dependence on the tax-payer.

Like Mr Long, I also want to see Britain at the top, with British products in the high street again, designed and manufactured by British industry using British components.

I want to see our industry competing in foreign markets, with self respect restored. But the remedy is not to spend larger amounts of tax-payers' money on more armament manufacture.

Dwight D Eisenhower, possibly this century's greatest warrior, summarised the situation: "...we must guard against the acquisition of unwarranted influence, whether sought or unsought, by the military-industrial complex, we must never let the weight of this combination endanger our liberties or our democratic progress".

Pacifism Mr Long. No, pragmatism.

R M Burfoot
Avon

Follow the leader

Research Notes has always received my best attention, being an inspirational source of new ideas. But the "Voltage follower that gives the lead" (January, p.12) immediately brought up memories of a similar circuit, titled "Feed-forward floating power supply (High-response-speed equalizer circuit)" by Eiichi Funasaka and Hikaru Kondou, *Journal of the Audio Engineering Society*, Vol 30, No.5, May 1982, p.324-329.

With due respect to the remarkable work of Mr Lidgey, it is surprising not to see the early reference mentioned.

Erik Margan
Slovenia

Percentage player

As a fascinated bystander in the debate on audio amplifier distortion, may I request a stronger distinction be made between harmonic and non-harmonic distortions?

Most musical sounds, including the voice, are generated by physical mechanisms which produce a harmonic series, the relative amplitudes of which determine the characteristic 'sound' of the instrument. A variation of a few percent in the absolute level of these harmonics is generally perceived (if at all) as a subtle change in timbre or brightness, of the kind also produced by a great many non-electronic effects including room acoustic. So provided that an amplifier produces harmonic distortions which drop sufficiently quickly with increasing order *n* to avoid masking the genuine harmonics of typical signal sources, can we not virtually ignore its thd figure?

In my experience the overall 'sound' of a good amplifier is far more affected by slight frequency response deviations from flatness than purely harmonic distortions – or, for that matter, phase distortions.

In contrast, non-harmonic distortions are audible even in very small amounts. Intermodulation distortion in particular multiplies with the complexity of the musical signal to produce a thick muddy or tinny background which obscures musical detail and can make listening quite unpleasant.

If, as I would suggest, the non-harmonic distortion of an amplifier is much more audible than the harmonic, then it follows that the preoccupation of your recent contributors with harmonic distortion is misplaced, even if an entirely (harmonically) distortionless amplifier would in principle produce no intermodulation distortion. Furthermore, should not distortions produced by non-linearity be directly dependent on signal level (as in rf mixers), and percentage distortion figures therefore strictly meaningless?

Perhaps we should abandon the increasingly irrelevant % thd figure and standardise on one of the various multiple-tone intermodulation measures available, preferably with a graph of measured signal to noise+distortion ratio, in decibels, versus signal level?

A New
Bristol

Worse performance...

I was intrigued to read the item in *Research Notes* (February, *EW + WW*) concerning electronics making us worse drivers. From bitter experience I can report that although "we all expect shorter stopping distances", under adverse conditions the ABS system will reduce the stopping distance from what it might have been without ABS, but will not reduce it to what it would have been under normal or dry surface conditions.

My experience only cost me an increased insurance premium. What is more worrying is the phenomenon of the side-road speeder who can be seen approaching a stop sign at an almost insupportable rate of deceleration, because he does it every morning under normal conditions. I just hope the morning there is a film of black ice on the road is the morning I go a different way to work.

Incidentally, recent experience in Montreal indicates that drivers quickly lose their caution after the first snowfall and become careless on adverse surfaces. This appears to parallel the learning mechanism for drivers of ABS equipped cars.

That is until you find out the hard way.

Nic Houslip
Birmingham

...from electronic systems

Research Notes (February, *EW + WW*) posed the question: "Does electronics make us worse drivers?"

I would like to take the point further. Not only does electronics make us worse drivers but it also has the capability to make us worse pilots, worse managers, worse designers and even worse mothers and fathers!

For example how many parents can say with hand on heart that we have not used the technology of the TV and video recorder just a little too much to give us that occasional peace and quiet. Indeed, how many of us can cope with the modern technology of the video recorder at all. A four year old child may be able to deal with the situation but isn't the technology supposed to be designed for the users irrespective of age?

Anything, used to excess can be damaging, and the same is true of technology. Whether it be video recorders that have too many glossy features, or industrial manufacturing systems that are supposed to be the 'solution' to all business ills.

The answer to these far-reaching problems lies with engineers. It is high time we stopped doing things right (just because it is technologically possible) and started doing the right things!

Andrew Ainger
Berkshire

Cables and cars

I was interested to read in Barry Gillebrand's article, 'Interfacing piezoelecric cable', January, pp. 21-23, that the type of cable we used some 40 years ago in an application for traffic speed measurement, has now become a fully fledged product.

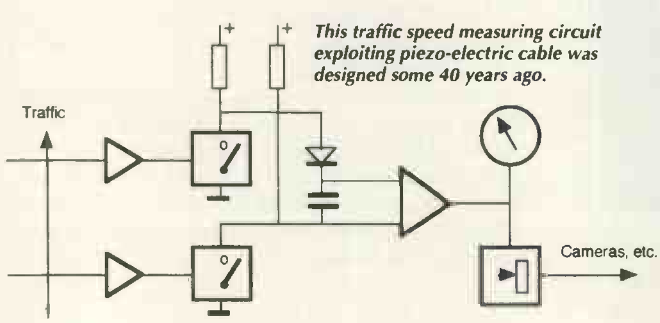
My colleague Joop van der Kam had discovered that if insulated wire is squeezed in a vice, a voltage is generated between its core and the metal of the vice. In fact any cable insulated with modern plastic materials generates electric noise when flexed[1] and special measures have to be taken to get rid of unwanted flexural noise in cables used for low-level electrical signal sources, such as microphones, etc.

Cause of the noise is separation of electrical charge that takes place when two different materials are brought in close contact. This can easily be demonstrated by pressing a coin onto a plastic bag. When the coin is removed, the deposited electrical charge can be made visible by dusting the plastic with fine particles (flour, pencil lead filings, etc), revealing the coin's features.

The discovery led to development of a traffic speed measuring apparatus in 1957[2]. It became rather popular because its cost was only some 10% of the competing radar device (in the era of the radio valve and klystron) and the prosecuting authorities liked the easy proof of whodunit' because of the direct physical contact between car and cables. A radar device relies of course on an invisible beam and unless the conditions were ideal, you could never be completely sure whether the return signal had been influenced by it reflecting off buildings or other traffic participants.

Today I would design the traffic speed meter with digital electronics, but in 1957 we did it as depicted in the figure.

Two, toughened coaxial cables



This traffic speed measuring circuit exploiting piezo-electric cable was designed some 40 years ago.

were laid at 2m distance across the road. Given the high input impedance of radio vales there was no particular cable interfacing problem to solve. The input amplifiers triggered two thyratrons, one of which started, while the other stopped, the charging of a low-loss capacitor when the front wheels of a vehicle crossed the cables. Thus the voltage across the capacitor was directly related to the time interval between the cable signals and hence to the vehicle's speed.

Voltage was measured by a high input impedance amplifier connected to a meter and a level discriminator. The scale of the meter in km/h (or mile/h) could be calculated from a simple exponential function and calibration was very straightforward. Adjustment of the level discriminator allowed the apparatus to be reset automatically for vehicles below a set speed limit, while trespassers would initiate automatic action such as triggering a camera or alerting a police officer down the road.

On completion of an operational cycle, the capacitor would be discharged, the thyratrons extinguished and the apparatus would be ready for the next measurement. Counters could be connected for automatic traffic analysis and the unit could be powered from a car battery.

Joop van Montfoort
Somerset

References
1. R A Rasmussen. "Flexural noise in cables", *Bell Lab Rec*, 37, 8, p.305, August 1959
2. J C van der Kam and J E van Montfoort. *Een Electronische Verkeerssnelheidmeter*, 13.8.1958 (unpublished internal report Philips, Eindhoven).

Bye-bidirectional I²C

Jean-Paul Brodier (*Letters*, February) is incorrect in stating that the SCL line in an I²C system needs to be bidirectional to handshake properly. I²C handshaking is done via acknowledge bits on the SDA line. The only time a bidirectional SCL line is needed is in a multi-master system, which also calls for additional hardware for arbitration.

Correction: in editing my *Circuit Idea* the word *programmed* was left out of the following phrase: "...writes 01 to a location programmed as 00 in the eprom...".

Mike Harrison
White Wing Logic
Essex

Figured out

In Fig. 6 of the article 'Analogue design with a 5V supply' (February, *EW + WW*, pp.162-165) by Walt Jung and James Wong the equations should read:

for G=100,

$(R_1+R_2)/R_3=(R_5+R_6)\ /R_4=99$

for G=10,

$R_1/\ (R_2+R_3)=R_6/\ (R_4+R_5)=9$

These respective resistor ratios should match to 0.01% or better.

Laurence Marchini
Oxon

Cathode ray conundrum

I am intrigued by the possibility that a cathode ray tube is capable of generating a reactionless force.

Consider what happens in an oscilloscope when the trace is directed towards the top of the screen. First, a beam of electrons is deflected upwards by deflector plates. These electrons then will have a small upwards momentum which is balanced by a downwards force on the deflector plates in accordance with Newton's equations of motion.

It occurs to me that what happens next is unusual and we cannot safely use our experience of mechanical systems to predict the precise effects. The beam of electrons is accelerated forwards towards the screen through a high potential, usually over 10,000V. Such voltages are sufficient to impart a velocity to the electrons which is a significant proportion of the velocity of light, *c*. According to Relativity theory, the electrons will undergo an increase in mass. Assuming their upwards velocity is not affected by this sideways acceleration, they will have acquired increased upwards momentum without any increase in the downwards reaction force. This will be realised as a net reactionless upwards force on the entire device when they hit the cathode.

I understand that when this particular problem is considered in text books, the answer given is that the vertical velocity of the electrons is somehow reduced (in the absence of any downwards force and contrary to Newton's second law) so that Newton's third law should be preserved in some form. This is an answer I had already considered and rejected. The trouble is that the amount that the vertical velocity of the electrons would have to change is a function of the vertical velocity of the reference frame from which the problem is considered – even when that velocity is an insignificant proportion of *c*. This would be absurd. I may have made a mistake but I am unaware that the conventional view has been tested by experiment, and unlike some other ideas that have been put forward on these pages, this idea is definitely testable. In any case, I don't see why Newton's second law should be so readily discarded to save his third law.

I do not suggest it might be possible to measure this force in an oscilloscope. I calculate that for a typical oscilloscope (Tektronix *2235*), the magnitude of the reactionless force (if it existed) would be sufficient to produce an apparent weight reduction of only 0.3µg for **1mA** of anode current. I have made enquiries about getting a purpose-designed thermionic device built where the level of predicted weight loss would rise to milligrams and so unambiguous measurements could be made. Building of a suitable device to do this seems to be possible, but it is unfortunately beyond my resources.

Has the idea been experimentally tested before or would anyone be interested in helping test it?

Interestingly, one of the few other places where this force might be expected to arise is in a force precessed gyroscope, although once again the level would only amount to a weight change less than one µg for any practical system, and it would not be practical even to attempt to measure it this way.

The possibility of this force has nothing to do with precession, which is in fact a nuisance.

Unfortunately, even if this force exists, I do not see many practical applications for it. However, if it were implemented, the amount of kinetic energy that would have to be stored in a device in order for it to be able to lift its own weight would be enormous and certainly beyond any technologies we have at present. Nevertheless, it has a certain theoretical interest.

R Lerwill
Clwyd ■

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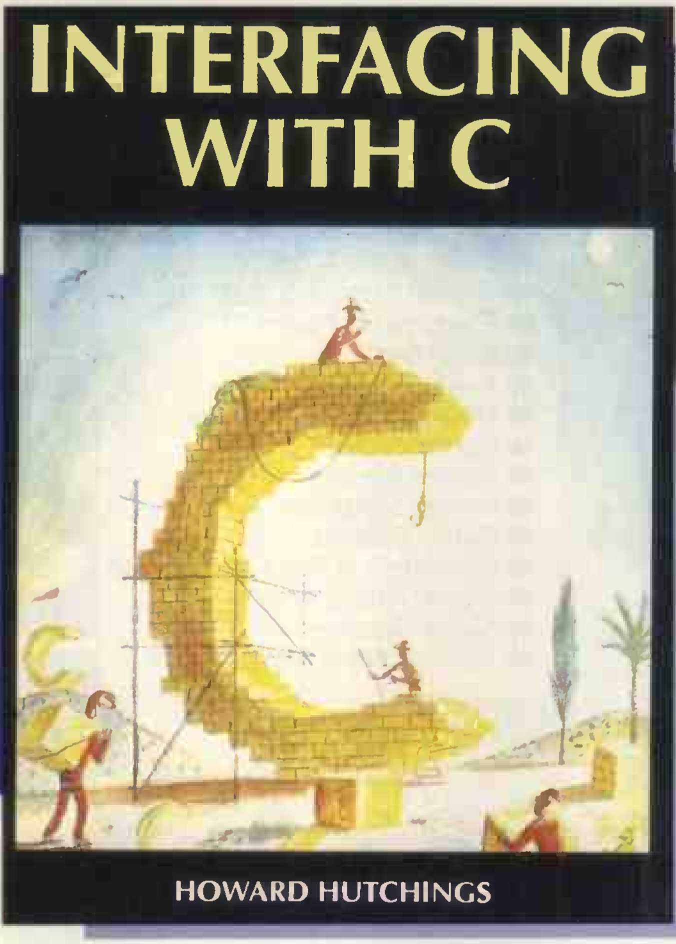

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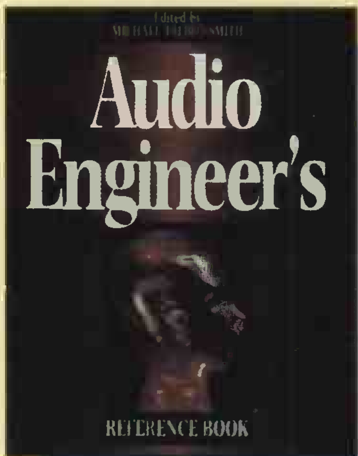

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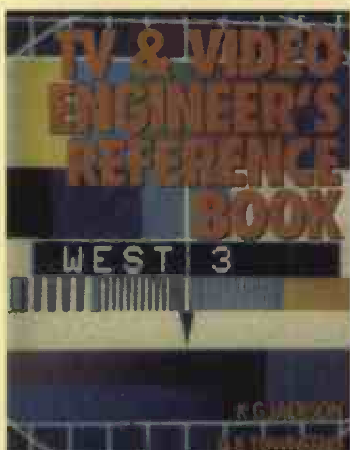

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Filters improve using cf op-amps

Historically, analogue designers have relied on passive filters for applications involving frequencies greater than a megahertz. But with current feedback op-amps designers can now make active filters capable of working at a megahertz and above, as Doug Smith explains.

This article first appeared in EDN.

Until recently, designing viable active filters with cut-off frequencies at 1MHz or greater was difficult because voltage feedback amplifiers with sufficient gain-bandwidth products and short propagation delays were simply too expensive.

The emergence of current feedback or transimpedance amplifiers has significantly changed this picture. Using these amplifiers, together with a conscientious design and pcb layout, you can design active filters that operate at high frequencies. Active-filter applications are no longer restricted to the audio-frequency range.

Active *RC* filters have many advantages over passive filters, these becoming increasingly important as frequency increases. For example, there is no insertion-loss penalty and you can even have power gain if needed.

A doubly-terminated passive filter would attenuate the signal by at least 50%. The elimination of inductors is the biggest advantage offered by active filters. This advantage doesn't involve size considerations alone.

Passive inductors are only linear for low power levels, much like transistors with no negative feedback. As you pump more current through the inductor, the magnetic core material begins to saturate and the inductor generates its own harmonic-distortion terms. The filter's transfer response will not necessarily suppress these signals. In an active *RC* filter, the amplifier quality and design sophistication set the dynamic range. Theoretically, designers have a good deal of control over both of these parameters.

As a case study, consider three situations – a low-pass antialiasing filter, a band-pass filter, and a high-Q notch filter – in which active filters that incorporate current-feedback amplifiers provide a viable alternative to passive filters. All three filters will be designed around the Burr-Brown *OPA603*.

You could implement the three designs using carefully selected, video-speed conventional op-amps. However, current feedback amplifiers more readily satisfy the low transit time and large bandwidth at high gain requirements for the example circuits. Let's start with the design of an antialiasing filter, **Fig. 1**, to drive the input of an *ADC603* – a 12-bit, 10MHz a-to-d converter.

When dealing with a-to-d converters in filter work, the Nyquist theorem states that if any converter input harmonic frequency is greater than half the sampling rate, those frequencies must alias, or fold back, into the passband. Normally, this condition is not desirable. To skirt the issue, you must suppress any input frequencies that exceed the Nyquist rate before the converter sees them.

The result of this manoeuvre is that the required attenuation becomes a function of converter resolution. It is also important for the filter to roll off as fast as possible. An elliptic response is the best choice because the addition of transmission zeros in the stop band creates the sharpest roll-off theoretically possible for a particular number of poles without having to rely on mutual inductance.

The first step in designing the filter is calculating the attenuation requirements. You can do so by estimating the theoretical signal-to-noise ratio, snr, using the expression

$$snr=6.02N+1.8\text{dB},$$

where N is the number of bits. For the *ADC603*, the expression yields,

$$snr=6.02\times12+1.8=74.04\text{dB}.$$

The calculation shows that the guaranteed stop-band attenuation must be greater than 74dB. A search of standard design tables shows that a fifth-order elliptic lowpass response is a reasonable compromise between the transition width and the filter order. The general transfer function for this filter is,

$$T(s)=\left(H_a\frac{s^2+b_{0a}}{s^2+a_{1a}s+a_{0a}}\right)\left(H_a\frac{s^2+b_{0b}}{s^2+a_{1b}s+a_{0b}}\right)\left(\frac{a_0}{s+a_0}\right)$$

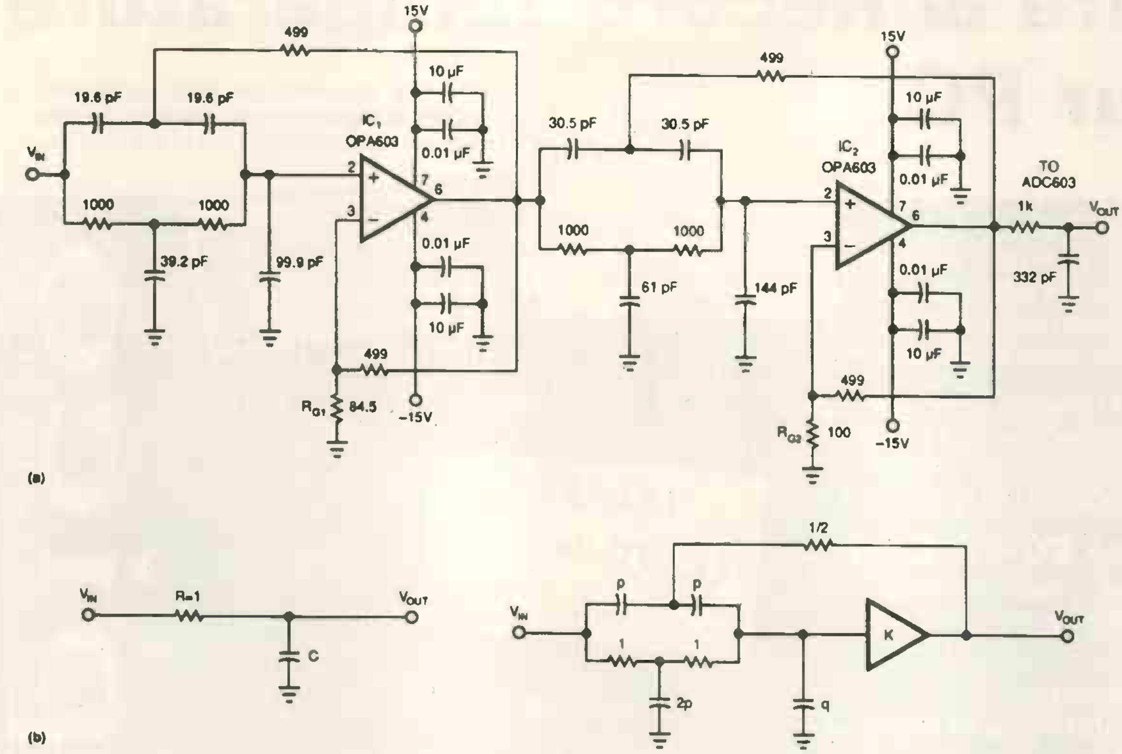


Fig. 1. You need an antialiasing lowpass filter when you're driving an A/D converter. A fifth-order elliptic design (a) proves to be the best choice in such an application. You can use two second-order sections and one first-order section (b) to form the necessary filter.

You can now form the filter by cascading two second-order sections and one first-order section, Fig. **1b**. The essential equations for the second-order sections are,

$$T(s) = H(s^2 + b_0) / s^2 + a_1 s + a_0$$

$$p = \frac{1}{\sqrt{b_0}}$$

$$q = \frac{(b_0 / a_0) - 1}{2\sqrt{b_0}}$$

$$K = 2 + \frac{(b_0 / a_0) - 1}{2} - a_1\sqrt{b_0}$$

The essential equations for the first-order section are,

$T(s)=a_0/s+a_0$

$a_0=1/RC$.

The task is to design a fifth-order elliptic antialiasing filter Fig. **1a** with a guaranteed stop-band attenuation of 75dB and no more than 3dB of passband ripple. In addition, the maximum attenuation should begin at 5MHz, which is half the sampling rate.

Transfer coefficients[1] for this case are

$a_{1a}=0.096035$
$a_{0a}=-0.945044$
$b_{0a}=10.47185$
$a_{1b}=0.285481$
$a_{0b}=0.413907$
$b_{0b}=4.328514$
$a_{0c}=0.191095$.

Corresponding component values are,

$p_a=0.309021$
$q_a=1.557592$
$K_a=6.875982$
$p_b=0\text{-}480652$
$q_b=2.272930$
$K_b=6.011362$
$R=1$
$C=5.232999$.

This filter prototype has an f_3 bandwidth of 0.15912 (1rad/s), and its maximum attenuation begins at a stop-band frequency of 0.3171. In this case, you have to scale the frequency to the stop-band frequency, rather than to f_3. In addition, you can arbitrarily scale the impedance to 1kΩ. Multiply each resistor by this impedance value; divide every capacitor value (p, q and C) by the frequency-impedance scaling factor, K_f,

$K_f=1\text{k}(5\times10^6\text{Hz})/0.3171\text{Hz}=1.577\times10^{10}$.

Final component values, rounded to three significant figures, are,

$p_a=19.6\text{pF}$
$q_a=98.8\text{pF}$
$K_a=6.88$
$p_b=30.5\text{pF}$
$q_b=144\text{pF}$
$K_b=6.01$
$C=332\text{pF}$.

Using a feedback resistance of 499Ω you can choose the closest 1% values for gain resistors, $R_{G1}=84.5\Omega$ and $R_{G2}=100\Omega$.

High-Q bandpass filters have many uses. One is isolating a particular harmonic of a distorted sine wave before amplifying the signal to more easily measure the magnitude. Many common active filter configurations run into problems in such applications because the value of Q is highly sensitive to changes in the gain – and thus the frequency response – of the amplifier. One of the best filter topologies in this situation is an extension of the basic Sallen-Key circuit **Fig. 2a**[2]. Adding a second amplifier can raise the potential value of Q by two orders of magnitude.

For stable operation, K_1 should be greater than zero and K_2 should be less than zero. The transfer function is,

$T(s)=K_1\times K_2 s/((1-K_1K_2)s^2+(4-K_1)s+2)$.

From this expression, you can determine that,

$$Q = \sqrt{\frac{2(1-K_1K_2)}{4-K_1}}$$

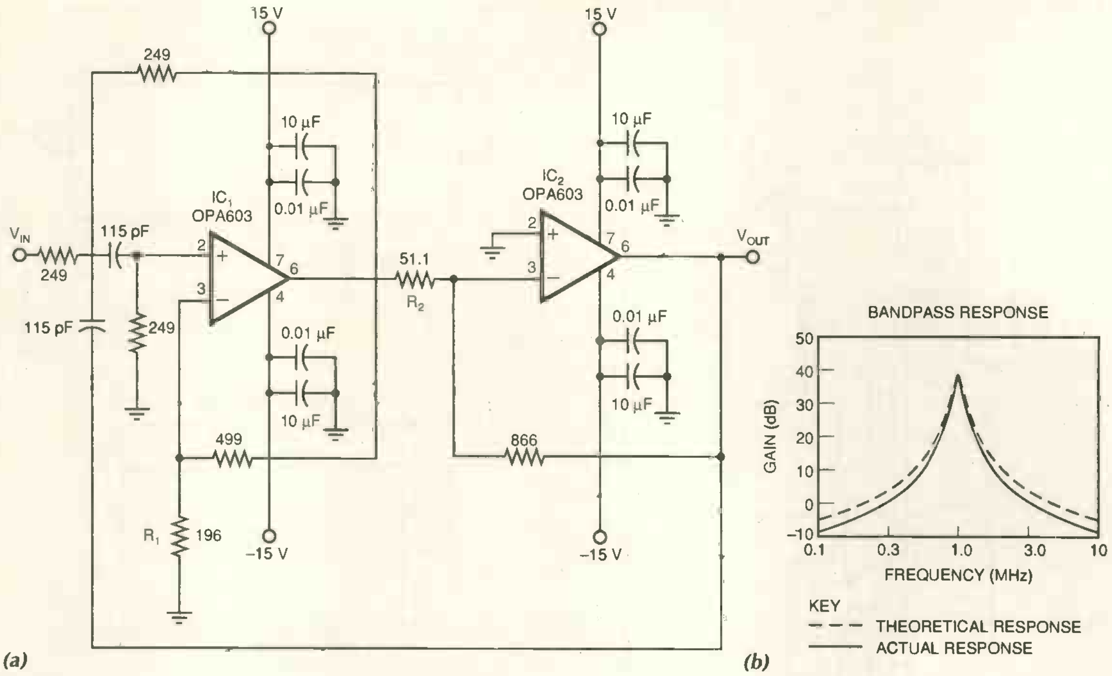
$$\omega_0 = \sqrt{\frac{2}{1-K_1K_2}}$$

Sensitivities of most concern involve the variations of Q when the gain of either amplifier changes. Analysis shows that,

$S_{K1}{}^{Q}=K_1(1-4K_2)/(4-K_1)(1-K_1K_2)$
$S_{K2}{}^{Q}=K_1K_2/1-K_1K_2$.

You can neglect $S_{K2}{}^{Q}$ because it is approximately equal to 1 and is not a serious limitation. Although its probably not obvious, there's a tradeoff between K_2 and $S_{K1}{}^{Q}$. The higher the gain of K_2, the lower the value of $S_{K1}{}^{Q}$. In a voltage type op-amp, higher gain inherently means lower bandwidth. However, a transimpedance amplifier has the ability to maintain its bandwidth at high gains. This characteristic gives current feedback amplifiers a clear advantage in this situation.

Fig. 2. To develop high-Q bandpass filters, you can add a second amplifier to the the basic KRC circuit (a) to raise the potential Q by orders of magnitude. The actual bandpass response of the filter is very close to the theoretical value (b), although the response shows a slightly lower gain.



(a) *(b)*

Putting theory into practice

Choosing a bandpass filter with a centre frequency of 1MHz and a –3dB bandwidth of 40kHz, the sensitivity to variations in gain should be no greater than 9.

First, the required value of Q is,

$$Q = f_0/BW = 1\text{MHz}/40\text{kHz} = 25.$$

Simultaneously solving the equations for Q and $S_{KI}{}^{Q}$ gives K_1=3.556357 and K_2=–17.01347. The corresponding centre frequency for this prototype is then f_0=0.287996. This centre frequency needs to be scaled to 1MHz. Arbitrarily choosing a value of 1kΩ for the resistors gives a final value of C=115.3pF. The required K_1 gain can be realised using a 499Ω resistor for the feedback and a 196Ω resistor for R_1.

Gain, K_2 is a different situation because the feedforward resistor of the second amplifier is the load resistance of the first amplifier. As a result, the feedforward resistor value needs to stay reasonably large. If you limit the second feedforward resistance to 50Ω, the second stage feedback resistor will be 866Ω and R_2 will equal 51.1Ω.

The dynamic range of high-frequency, moderately priced spectrum analysers is often less than 80dB. However, you can effectively increase the measurement range by suppressing the fundamental frequency of the input signal by a known amount without affecting the rest of the frequency spectrum. This application doesn't require a high-order, band-reject filter – a low order high Q notch filter will work well.

The classic twin-T network, **Fig. 3a** is a promising candidate for the job. The transfer function of this network is,

$$T(s) = s^2+\omega_0{}^2/s^2+4\omega_0 s+\omega_0{}^2.$$

This circuit has two drawbacks – it is somewhat sensitive to passive component tolerances, and it has an intrinsic Q value of 0.25. The first drawback creates no problem but the second drawback must be overcome. You can substantially increase circuit Q value by adding a second amplifier to the network[3]. The new transfer function is now,

$$T(s) = s^2+\omega_0{}^2/s^2+4\omega_0(1-K)s+\omega_0{}^2$$

and the Q value is now a function of K,

$$Q = 1/4(1-K).$$

As K approaches 1 from below, Q increases in an unlimited fashion. If K is greater than 1, however, the circuit is unstable. Although wide bandwidth at high gain is not as important here as it was in example **Fig. 2**, the comparatively lower transit time of a current-feedback amplifier should yield superior performance in this application.

A specific example will prove the point. The task is to design a 1.5MHz notch filter that has a –3dB bandwidth of 225kHz. The first step is to calculate Q using the expression,

$$Q = f_0/BW_{-3dB} = 1.5\text{MHz}/225\text{kHz} = 6.66.$$

You can use this value to calculate,

$$K = 1-(1/4Q) = 0.9625.$$

If R_1 is set equal to 1kΩ then,

$$C = 1/2\pi f_0 R_1.$$

If you let R_2 also equal 1kΩ then $(1-K)R_2$=37.5 and KR_2=962.5. Figure **3b** shows the final notch filter design. Both amplifiers are configured as unity gain buffers, and the feedback resistance is set at 499Ω. The actual response, Fig. **3c** shows a slight excess attenuation beyond the notch frequency, but the performance is still good.

Making the case for current feedback

Don't get the idea that something is inherently wrong with voltage feedback, even at high speed. In fact, voltage-feedback amplifiers generally have a lower noise-floor specification than current feedback amplifiers. However, when comparing voltage and current feedback amplifiers, you must take the application into consideration. Current feedback, or transimpedance, amplifiers have some distinct performance advantages as waveform speed gets higher and higher. These advantages can translate into higher-performance active filters.

The most striking difference between voltage feedback and transimpedance op-amps is that with a fixed feedback resistor, the current feedback amplifier has very low gain bandwidth tradeoff. Transimpedance amplifiers maintain bandwidth at high gain settings – an advantage in active filter topologies because a large gain is needed to minimise sensitivity.

In addition, transimpedance amplifiers have very high slew rates compared with those of conventional voltage op-amps. A typical slew rate for a video-speed voltage feedback amplifier is in the 200 to 300V/µs range. A comparable current feedback amplifier might slew as fast as 2500V/µs. This slew-rate disparity is easy to explain. In a conventional amplifier, the slew rate is the ratio of the bias current flowing through the slewing node to the capacitance that can be referred back to that node.

In a transimpedance amplifier, the feedback current mirrors and adds to the bias current flowing through the slew-rate limiting node. Because more current is available to charge the capacitance, the slew rate increases. The feedback current is proportional to V^{OUT}, which is proportional to V^{IN}. So the harder you drive a current feedback amplifier, the faster it slews. In practice, this effectively eliminates slew rate as a limiting factor in high speed, active filter design.

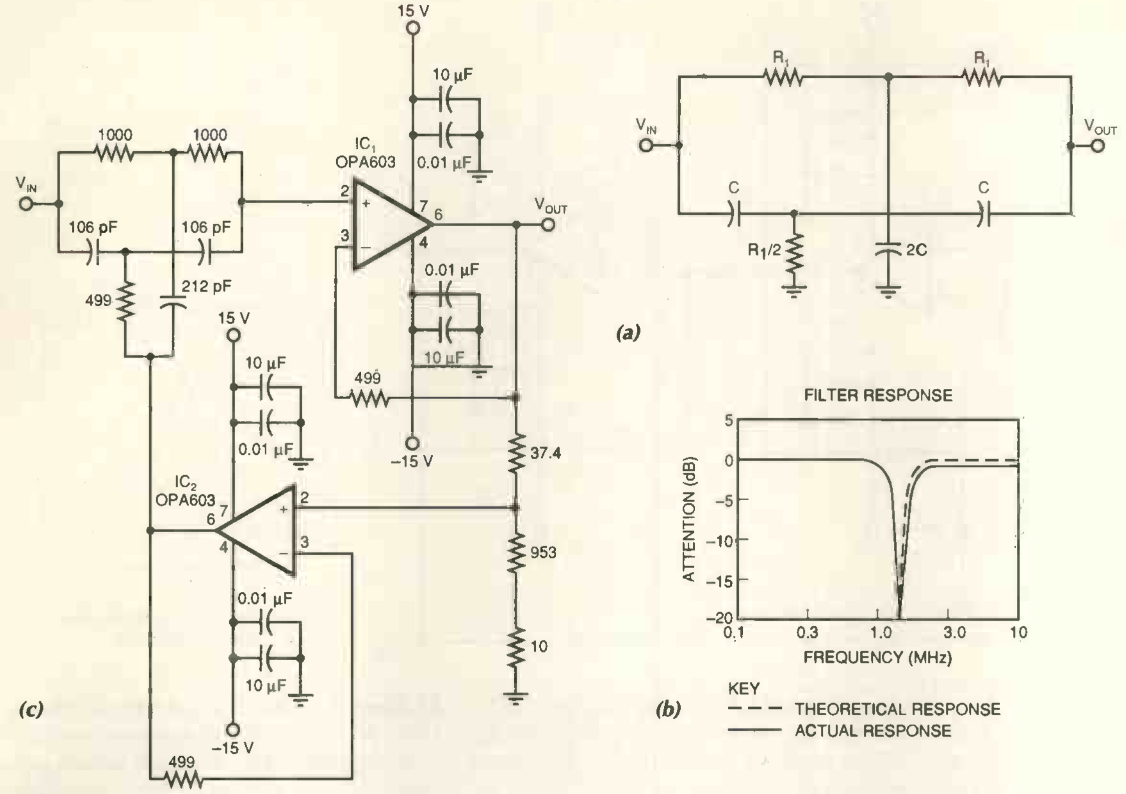


Fig. 3. When you need a high-Q notch filter, the classic twin-T network (a) is a good starting point. By adding a second amplifier (b), you can substantially raise circuit Q. The actual response of the filter (c) shows a slight excess attenuation beyond the notch frequency.

One final factor that favours the transimpedance amplifier is settling time. Designers often choose a filter transfer function for best time-domain response. Therefore, ensuring that the amplifier settles to the required level substantially faster than the filter has to settle is crucial. High speed amplifiers are complicated devices and acceptable ac response does not necessarily ensure an acceptable settling time. Many conventional voltage feedback op-amps use internal pole-zero cancellation to increase their bandwidths. Analysis shows that a small mismatch in the pole-zero cancellation has a negligible effect on frequency response, but the scheme can dramatically boost settling time. Transimpedance amplifiers have settling time problems too. Although transimpedance amplifiers settle to 0.1% (10 bits) or 0.02% (12 bits) in as little as 15ns, the settling time to 0.01% can be relatively long. The same current flow that increases the slew rate of a transimpedance amplifier also upsets the amplifier's bias point slightly, and a finite amount of time is required for the bias point to return to equilibrium. This effect is small, but it can often extend the 0.01% settling time to several microseconds. ■

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1. Kerwin, W J, 'An Active RC Elliptic Function Filter', *IEEE Region Six Conference Record, Volume 2*, April 1966, p647.
2. Kerwin, W J, *Linear Active Circuit Design, Part 4, Course Notes for ECE-553*, University of Arizona, Spring 1987, p. 51.
3. Williams, Arthur B, *Electronic Meter Design Handbook*, McGraw-Hill, New York, 1981, p6.

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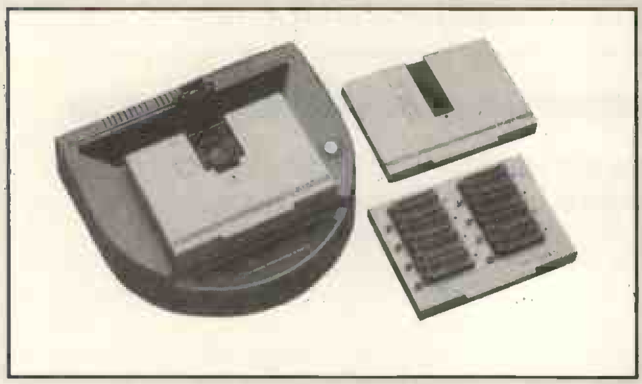

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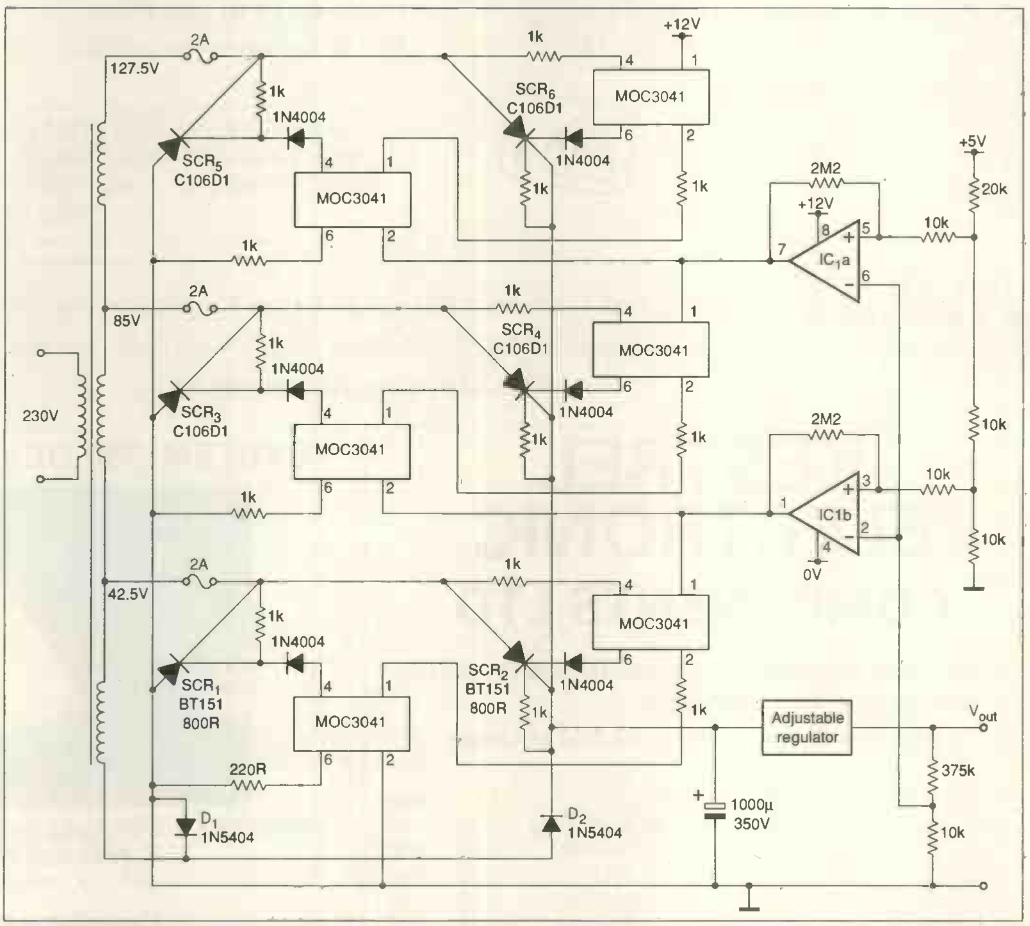
Power supply preregulator with no moving parts

Using no relays, this circuit switches the ac input to a series of full-wave bridge rectifiers, providing smoothed and regulated dc output.

At switch-on, the op-amp outputs are both high and $SCR_{1,2}$ fire alternately to form a full-wave bridge driven from the 42.5V transformer tap. When V_{out} increases above about 48V, IC_{1b} output goes low, allowing $SCR_{3,4}$ to fire alternately, the 85V tap now driving the bridge. Similarly, as V_{out} reaches 96V, both op-amp outputs are low and $SCR_{5,6}$ fire to form a bridge driven by the 127.5V tap. Optocouplers *MOC3041* enable zero crossing triggering of the scrs.

If the output is short-circuited, only $SCR_{1,2}$ fire to minimise dissipation.

Gregory Freeman
Nairne
South Australia



Series of scr bridges replaces relays on a tapped transformer secondary.

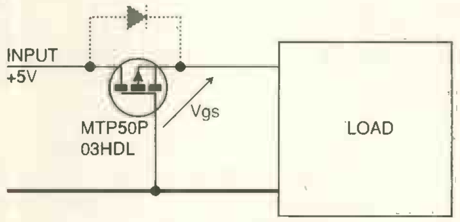
Polarity-dependent switch

Depending on the polarity of the input voltage, this low-loss mosfet connects or disconnects the supply to the load; it is used as a polarity discriminator or simply as a self-synchronising switch.

With a positive input, the parasitic diode across the mosfet conducts, the positive feedback causing the mosfet to saturate to an $R_{ds(on)}$ of 30mΩ. When the input is negative, no current flows and V_{gs} remains at zero, the input voltage possibly reaching the mosfet rated breakdown, in the case of the *MTP50P03HDL* 30V.

Do not allow the positive excursion of input voltage to exceed the 10V gate/source voltage of the mosfet.

Kristen Ellegard
Oslo, Norway



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Testing sampling rate

For correct sampling of a sinusoid, samples must be taken at least twice per cycle – the Nyquist rate. This circuit gives a positive indication as to whether this is taking place.

Transistor Tr_1 behaves as an emitter follower when Tr_2 is off and as a diode when Tr_2 is on, it being controlled by V_s in **Fig. 2c,** itself derived from the sampling voltage.

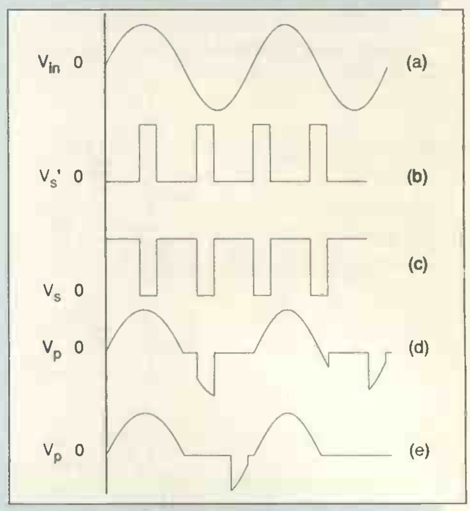
When a sampling pulse arrives during a negative half-cycle, Tr_2 is off and V_{in} appears at V_p, which is shown in **Fig. 2d** for a correctly sampled wave and at **Fig. 2e** for an incorrect one.

Voltage V_p goes to comparators $OA_{1,2}$, $D_{1,2}$ clamping the negative outputs to ground; OA_1 output is high during positive half-cycles, but OA_2 goes high only when a negative-going pulse appears. Bistable device FF_1 is reset at the start of each half-cycle and set at the first negative pulse. Bistable FF_2 is set by $OA_{3,4}$ at the start of negative half-cycles.

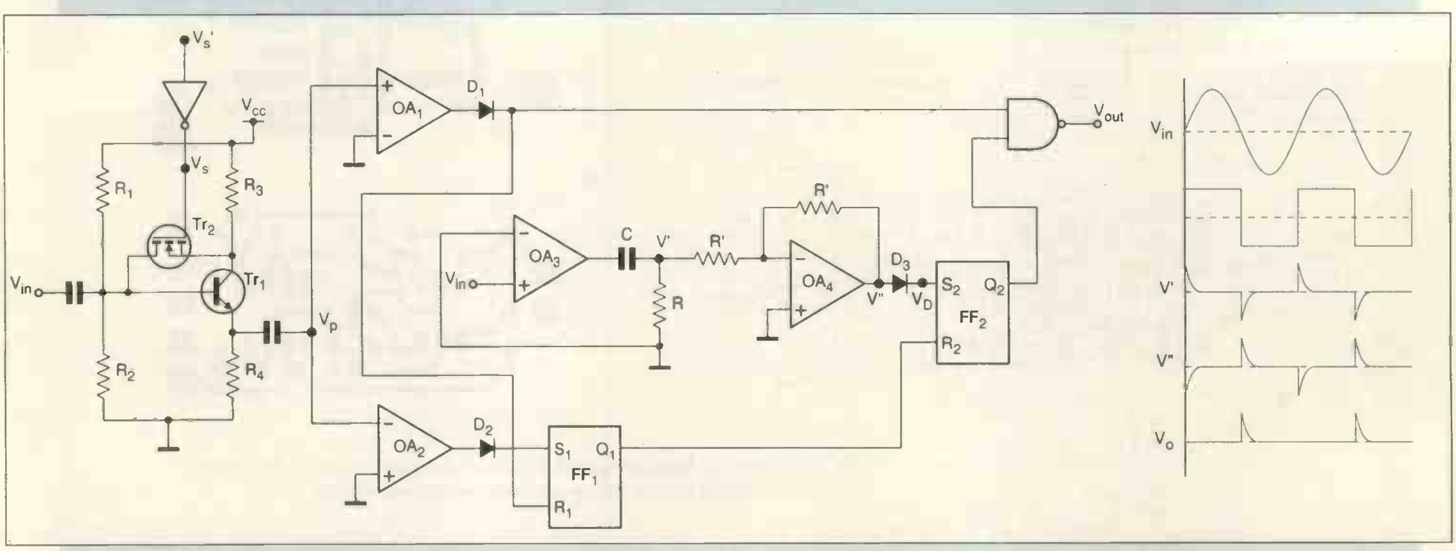
Since FF_2 depends for its reset pulse on the setting of FF_1, which itself is set by the appearance of a negative pulse, Q_2 will remain high and V_{out} low if the pulse does not arrive, indicating that incorrect sampling is taking place.

If sampling rate equals input frequency exactly, the circuit fails since the negative half-cycle may be sampled each time.

K N Sunil Kumar
Visakhapatnam
India



If sampling rate of a sinusoid is below Nyquist rate ($2f_{in}$), this arrangement, below and right, provides a constant low at the output.

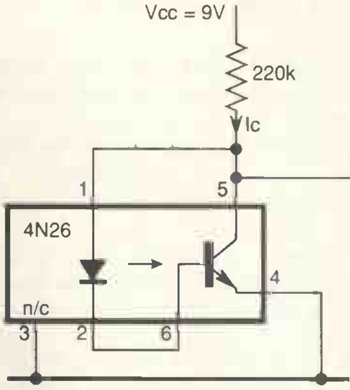


Noise source from an optocoupler

Output from this photoelectric noise source is typically 12μV, which is about 28dB above thermal noise; some samples of *4N26* opto-isolators give a much greater output. Noise output was measured over a 10kHz bandwidth, in which *1/f* or flicker noise is likely to be large.

Connected as shown, the led controls the transistor working point accurately. Output impedance is $1/(20I_c)$ or about 1.4kΩ in this case and the output can be matched to loads of 50-600Ω or more. Grounding pin 3 reduces hum pickup.

W Gray
Farnborough
Hampshire



Noise source, after Hickman (EW+WW, November 1993) uses opto-isolator which also controls transistor working point.

High-speed buffer features low input-capacitance

An input capacitance of 1.2pF or 2.7pF, depending on devices, and a bandwidth of 50MHz with 10mV offset come from the use of a bootstrapped fet input buffer, as described by Horowitz and Hill*.

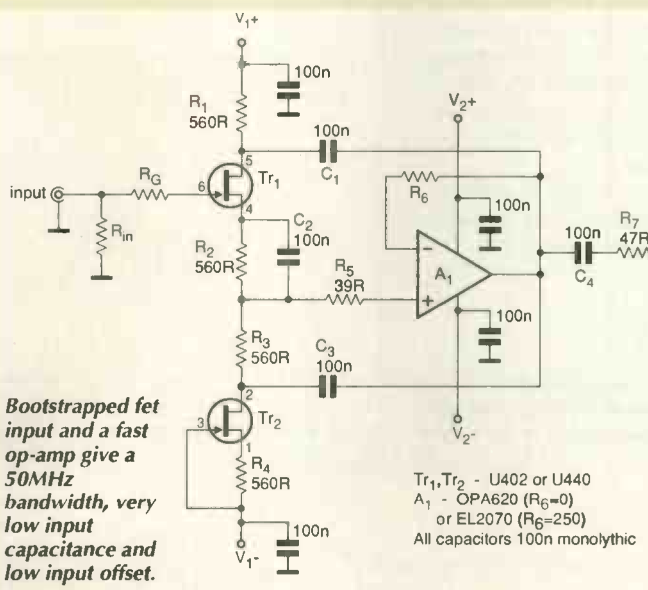
Both the *OPA620* or the *EL2070* op-amps have been tried as bootstrap driver, with no discernible difference in performance, although the *EL2070* is a current-feedback type and needs at least 220Ω in the feedback path to ensure stability. Since the *OPA620* is a voltage-feedback op-amp, the feedback resistor could be shorted.

Output impedance is 50Ω in the circuit shown, but the *RC* output circuit could be dispensed with to avoid the 6dB loss in gain due to the resistor and to obtain a dc response; at dc, resistor R_4 should be adjustable to maintain a reasonable input offset. No output snubbing is needed to stop oscillation with a reactive load if load impedance is 50-200Ω and resistive, otherwise the resistance should be over 20Ω.

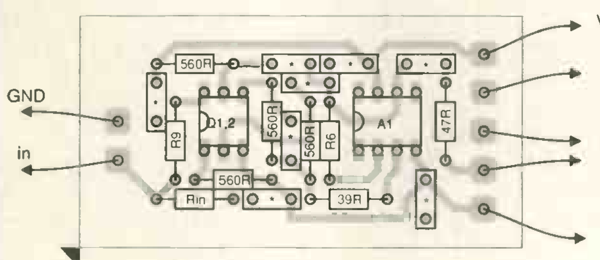
Both *U402* and *U440* dual fets work well in the circuit. Using the *U402*, bandwidth is more than 50MHz and input capacitance with R_G at 10kΩ is 2.7pF. The *U440* gave an input *C* of 1.3pF, but with a greater input offset.

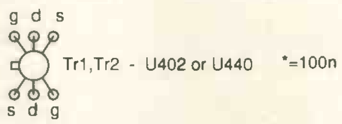
Phil Denniss
University of Sydney
NSW
Australia

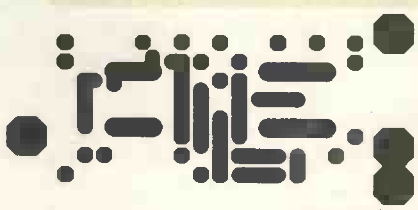
Reference
Horowitz and Hill. *The Art of Electronics*, second edition, p.135.

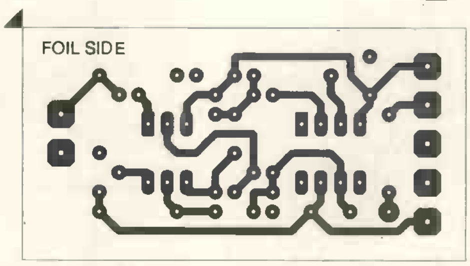


Bootstrapped fet input and a fast op-amp give a 50MHz bandwidth, very low input capacitance and low input offset.









Due to the high speeds involved, the pcb for the buffer needs careful layout. These patterns are approximately 1:1.

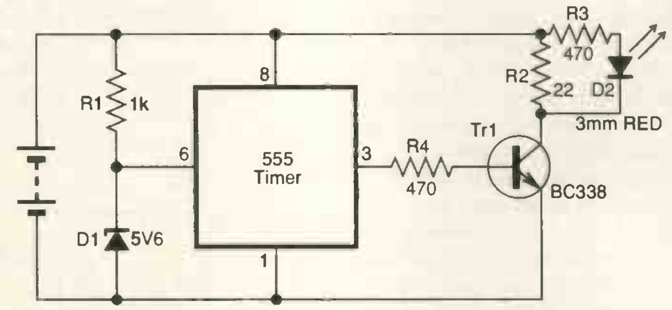
Safe NiCd battery-pack discharger

Although nickel-cadmium batteries need regular charge/discharge cycles, the discharge must be limited to avoid reverse charging of weak or partially discharged cells in the pack. This circuit limits the discharge of an eight-cell pack to a 1V terminal voltage per cell.

Connecting the battery takes pin 6 to the 5.6V zener voltage, the transistor conducts and the led indicates discharge. When battery voltage reaches $1.5V_{ZD1}$, pin 3 goes low, the transistor turns off and battery discharge stops: the led turns off.

To take any number of cells up to a maximum of 12, the zener voltage should be $^2/_3$ the final terminal voltage and zener current adjusted by R_2 to 0.5A.

Bill Hume
Newmilns
Ayrshire



NiCd battery-pack discharger, for up to 12 cells, limits discharge to avoid damage to less than perfect cells.

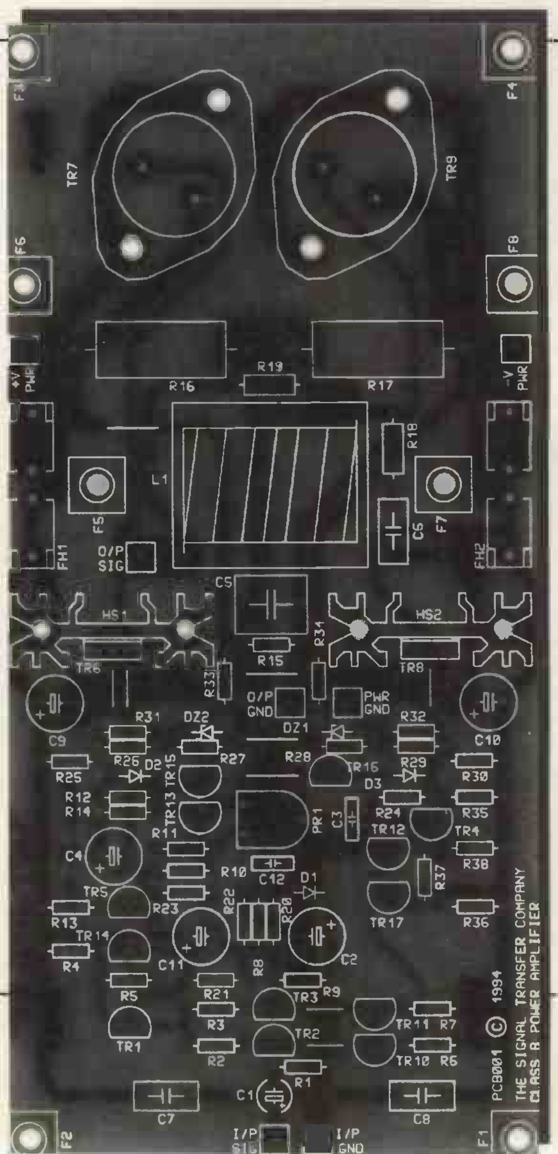
PCBs for Douglas Self's power amplifier series

Circuit boards for Douglas Self's high-performance power amplifier are now available via *EW+WW*.

Detailed on page 139 of the February 1994 issue, Douglas Self's state-of-the-art power amplifier is the culmination of ideas from one of the most detailed studies of power amplifier design ever published in a monthly magazine. Capable of delivering up to 100W into 8Ω, the amplifier features a distortion of 0.0015% at 50W and follows a new design methodology.

Designed by Douglas himself, the fibreglass boards have silk-screened component IDs and solder masking to minimise the possibility of shorts. Sold in pairs, the boards are supplied with additional detailed constructional notes.

Each board pair costs £45, which includes VAT and postage, UK and overseas. Credit card orders can be placed 24 hours on 0181-652 3614. Alternatively, send a postal order or cheque made payable to Reed Business Publishing to EW+WW, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.



APPLICATIONS

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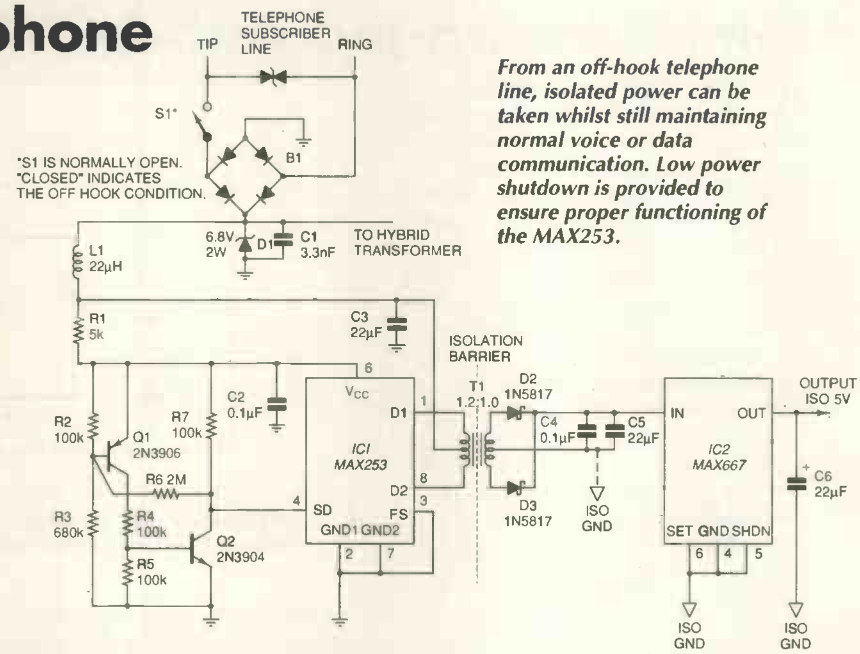
Power from the 'phone

This power supply, from Maxim's *Engineering Journal*, volume 17, is useful in portable systems that connect to domestic telephone lines. Such systems are modems and telephone test sets. For apparatus needing 150mW or less, the circuit eliminates the need for batteries and mains adapters. It draws power from the telephone jack without affectting voice signals.

Built into peripheral equipment such as pcmcia modem cards, it can spare the battery in a host computer. Line current available to a telephone in the off-hook state is limited not by regulations or code, but by the sum of impedances in the central-office battery and intervening phone lines. These line impedances vary greatly in proportion to distance from the central office, so the customary practice of matching impedances for maximum power transfer is impractical. But, the zener-clamp termination, D_1 works well for line impedances to 1kΩ and for worst-case conditions. It also meets the one condition imposed on line current by the phone system: off-hook current must exceed 20mA to ensure activation of a network-access relay in the central office.



From an off-hook telephone line, isolated power can be taken whilst still maintaining normal voice or data communication. Low power shutdown is provided to ensure proper functioning of the MAX253.

5V isolated supply

Diode D_1 provides approximately 6.8V to the center tap of T_1 and 5V to the Vcc terminal of IC_1. A 400kHz oscillator driving a flip-flop, inside IC_1, generates two push-pull, 50% duty-cycle, 200kHz square waves that drive internal, ground-referenced switches. , in turn, these connect to the primary of *T1*. Isolated power on the secondary side is first rectified by schottky diodes, *D2* and *D3*, and then regulated to 5V by the low-dropout linear regulator IC_2.

Transformer, T_1 has a center-tapped winding whose *ET* product (a voltage-time product of 25Vµs) is sufficient to prevent saturation under worst-case conditions. Similarly, the turns ratio should provide the minimum-required output voltage for maximum load and minimum input voltage. This calculation must assume worst-case losses in D_2 and D_3. This turns ratio produces a much higher secondary voltage for best-case conditions and, for some applications, that is acceptable. Otherwise, add the linear regulator, IC_2, as shown.

For isolated 5V outputs, the ideal turns ratio is 1.2 ct:1.0 ct (ct is center tapped). The transformer should be wound on Magnetics Incorporated 'W', Fair-Rite "76" or other high-permeability magnetic material. To minimize radiated noise, choose a pot core, E/I/U type core, toroid, or other geometry with closed magnetic paths.

Consider a typical toroid such as the 40603-TC from Magnetics, Inc. (0.125" thick with a 0.230" outside diameter). For 6.8V inputs this core should have a 48 turn primary (24 turns on either side of the centre tap), which yields a nominal, end-to-end primary inductance of 8mH. The secondary can be scaled for any reasonable output voltage required. Forty turns, for example, (20 either side of the centre tap) delivers 5.2V minimum as required by the linear regulator to maintain a regulated 5V supply.

Transistors *Q1*, *Q2*, and the associated resistors assure a low-power shutdown mode for *IC1* until its supply voltage can sustain a full power-up. *IC1*'s supply current is fairly constant, so light filtering, provided by *L1* and *C3*, is sufficient to prevent noise from entering the hybrid transformer.

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Tel 01734 845 255, fax 01734 843 863.

In-circuit programmer from LPT1

With the capability of programming a *PIC16C84* microcontroller without removing the device from the target circuit, this low-cost serial programmer is controlled using a pc parallel port. Microchip's application note *AN589* describes a circuit which can also read back internal *PIC* data. This feature is very useful where changes in program code or constants are necessary to compensate for other system features.

For example, an embedded control system may have to compensate for variances in a mechanical actuator performance or loading. The basic program can be programmed and tested in design. The final program and control constants can be easily added later in the production phase without removing the microcontroller from the circuit.

Automatic software and performance upgrades can also be implemented via in-system programming. Upon receiving new system software via disk or modem, a control processor with the included programming code could perform an in circuit reprogramming of other microcontrollers in the system.

This programmer can load program code, part configuration, and eeprom data into the *PIC16C84*. In read back mode, it can verify all data entries.

Parallel interface

The *PIC16C84* microcontroller is put into programming mode by forcing a logic low level on RB_7, pin 13 and RB_6, pin 12. Pin 4, Line *MCLR*, is first brought low to reset the device and then brought to the program/verify voltage of 12V to 14V, where it remains for the rest of the programming or verification time. After entering programming mode, RB_7 is used to serially enter programming modes and data into the device. A high-to-low transition on RB_6, the clock input, qualifies each bit of the data applied on RB_7. The first six bits form the command field and the last 16 bits form the data field. The latter is composed of one zero starting bit, 14 data bits, and one zero stop bit. The incremental address command is the command field only.

The read mode is similar to programming mode except that the data direction of RB_7 is reversed after the six bit command to allow the requested data to be returned to the programmer. After the read command is issued, the programmer tri-states its buffer to allow the device to serially shift its internal data back to the programmer.

The rising edge of clock input RB_6 controls data flow by sequentially shifting previously programmed data bits from the part. The programmer qualifies this data on the falling edge of RB_6. Note that 16 clock cycles are needed to shift out 14 data bits.

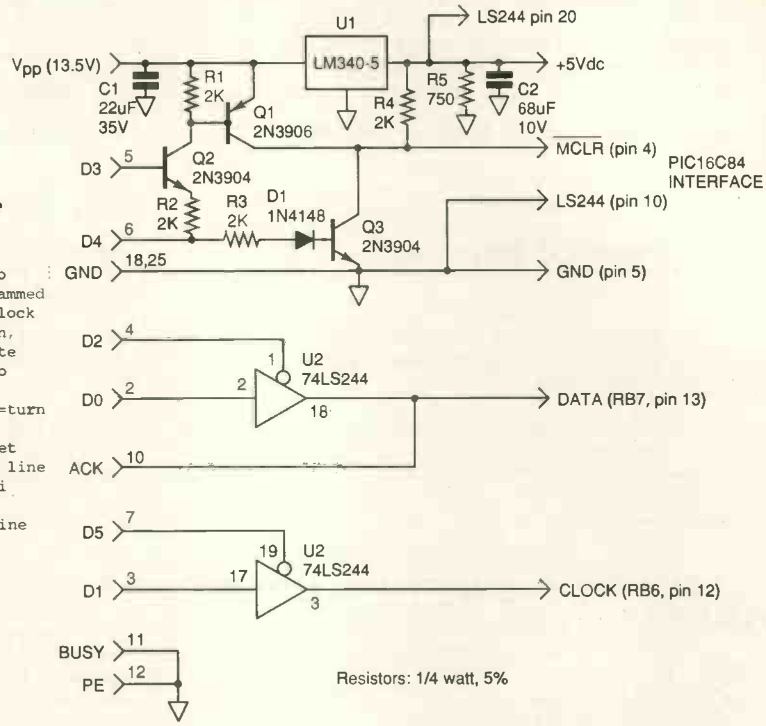
Accidental in-circuit reprogramming is prevented during normal operation by the *MCLR* voltage which should never exceed the maximum circuit supply voltage of 6Vdc and the logic levels of port bits $RB_{7,8}$. After programming or verification, the *MCLR* pin is brought low to reset the target microcontroller and electrically release it. The target circuit is then free to activate the *MCLR* signal.

If *MCLR* is not forced by the target circuit, a 2kΩ pull up resistor in the programmer, R_4, provides a high logic level on the target microcontroller. This enables execution of its program, independent of the programmer connection. Provision should be made to prevent the target circuit from resetting the target microcontroller with *MCLR* or affecting RB_6 and RB_6 during the programming process. In most cases this can be done without jumpers.

A logic high on parallel interface latch bit, D_4, turns on Q_3 causing *MCLR* to go low and place the target device in reset mode.

Parallel interface bit functions.

```
d0 data output to
part to be programmed
d1 programming clock
d2 data direction,
0=enable tri-state
buf, send data to
part
d3 Vpp control 1=turn
on Vpp
d4 MCLR=0, 1=reset
device with MCLR line
d5 clock line tri
state control,
0=enable clock line
```



The PIC16C84 can be programmed with the aid of a single tri-state buffer ic, a regulator and a handful of discrete components without removing it from circuit.

Reset is then removed and the program or verify voltage is applied by a logic high on D_3 and a logic low on D_4. This turns off Q_3 and turns on Q_2 and Q_1. Simultaneous reset and program mode is prevented by connecting the emitter of Q_2 to latch bit D_4. Data and clock are connected to the device via a tri-state buffer U_2. Pc parallel port interface bit D_0 is used for data and port bit, D_1 is used for clocking.

During programmng mode both clock and data buffers are enabled by port bits D_2 and D_5. During read mode, the data buffer is tri-state activated via D_2 and the data acknowledge signal line on the printer port is used to receive verification data. After programming or verifcation, both the data and clock lines are put into the high impedance state via D_2 and D_5, allowing the these lines to be used by the target circuit. This allows the programmer to remain physically, but not electrically connected to the target system. ■

Arizona Microchip Technology Ltd,
Unit 6, The Courtyard, Meadowbank,
Furlong Road, Bourne End,
Bucks SL8 5AJ
Tel 01628 851 077, fax 01628 850 259

***This control-program listing, designed to aid in-circuit programming of PIC microcontrollers, is one of a number of code fragments from Microchip's* Embedded Control Handbook.**

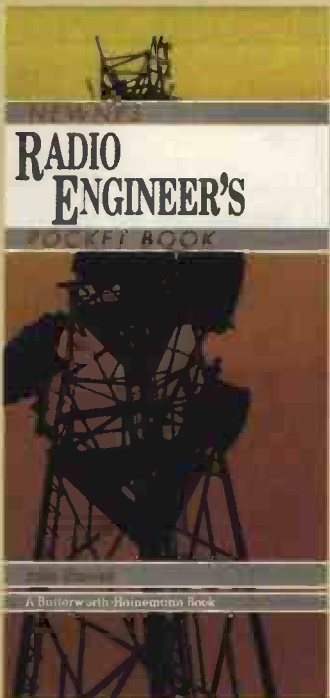
```
//FUNCTION PROTOTYPE:  int ser_picl6c84(int cmd, int data)
// cmd:  LOAD_CONFIG      - part configuraton bits
//       LOAD_DATA        - program data, write
//       READ DATA        - program data,  read
//       INC_ADDR         - increment to the next address (not automatic)
//       BEGIN_PROG       - program a previously loaded program code or data
//       LOAD_DATA        - load EEPROM data registers (BEGIN_PROG must follow)
//       READ DATA        - read EEPROM data
// data :          1) 14 bits of program data or
//                 2) 8 bits of EEPROM data (least significant 8 bits of int)
// cmd : RESET            - provides 1 ms reset pulse to target system
//       PROGRAM_MODE     - initializes PIC16C84 for programming
//       RUN              - disconnects programmer from target system
//function returns:     1) 14 or 8 bits read back data for read commands    2) zero for write data commands
//                      3) PIC_PROG_EROR=-1 for programming errors          4) PROGMR_ERROR=-2 for programmer errors

#include <bios.h>
#define LOAD_CONFIG       0
#define LOAD_DATA         2
#define READ_DATA         4
#define INC_ADDR          6
#define BEGIN PROG        8
#define PARALLEL_MODE     10                        // not used
#define LOAD_DATA_DM      3
#define READ_DATA_DM      5
#define MAX_PIC_CMD       63                        // division between pic and programmer commands
#define RESET             64                        // external reset command, not needed for programming
```

Continued over page

```
#define PROGRAM_MODE      65                        // initialize program mode
#define RUN               66                        // electricaliy disconnect programmer
#define PIC_PROG_EROR     -1
#define PROGMR_ERROR      -2
#define PTR               0                         // use device #0

int ser_ pic16c84(int cmd, int data)                // custom interface for pic 16c84
  {
   int i, s_cmd;
   if(cmd <=MAX_PIC_CMD)                            // all programming modes
     {
     biosprint(0,8,PTR);                            // set bits 001000, output mode, clock & data low
     s_cmd = cmd;                                   // retain command "cmd"
     for (i=0;i<6;i++)                              // output 6 bits of command
       {
       biosprint(0, (s_cmd&0x1) +2+8,PTR);          // set bits 001010, clock hi
       biosprint(0, (s_cmd&0x1)   +8,PTR);          // set bits 001000, clock low
       s_cmd >>=1;
       }
     if ((cmd ==INC_ADDR)||(cmd ==PARALLEL_MODE)    // command only, no data cycle
       return 0;
     else if (cmd ==BEGIN_PROG)                     // program command only, no data cycle
       {
       delay(10);                                   // 10ms  PIC programming time
       return 0;
       }
     else if((cmd ==LOAD_DATA)||(cmd ==LOAD_DATA_DM)||(cmd ==LOAD_CONFIG))   // output 14 bits of data
      {
      for (i=200;i;i-)                              // delay beteen command & data
      biosprint(0, 2+8, PTR);                       // set bits 001010, clock hi; leading bit
      biosprint(0,   8,PTR);                        // set bits 001000, clock low
      for (i=0;i<14;i++) ;                          // 14 data bits lsb first
        {
        biosprint(0, (data&0x1) +2+8,PTR);          // set bits 001010, clock hi
        biosprint(0, (data&0x1)   +8,PTR);          // set bits 001000, clock low
        data >>=1;
        }
      biosprint(0,2+8,PTR);                         //  set bit s001010, clock hi; trailing bit
      biosprint(0,  8,PTR);                         // set bits 001000, clock low
      return 0;
      }
    else if((cmd ==READ_DATA)||(cmd ==READ_DATA_DM))        // read 14 bits from part, lsb first
      {
      biosprint(0,4+8,PTR);                         // set bits 001100, clock low, tri state data buffer
      for (i=200;i;i-);                             // delay between command & data
        {
        biosprint(0,2+4+8,PTR);                     // set bits 001110, clock hi, leading bit
        biosprint(0,  4+8,PTR);                     // set bit 001100, clock low
        data =0;
      for (i=0;i<14;i++)                            // input 14 bits of data, lsb first
        {
        data >>=1;                                  // shift data for next input bit
        biosprint(0,2+4+8,PTR);                     // set bits 001110, clock hi
        biosprint(0,  4+8,PTR);                     // set bits 001100, clock low
        if(!(biosprint(2,0,0)&0x40)) data += 0x2000;        // use ack line for input, data lsb first
        }
      biosprint(0,2+4+8,PTR);                       // set bits 001110, clock hi,trailing bit
      biosprint(0,  4+8,PTR);                       // set bits 001100, clock low
      return data;
      }
    else return PIC_PROG_EROR;                      // programmer error
    }
  else if(cmd == RESET)                             // reset device
    {
    biosprint(0,32+16+4,PTR);                       // set bits 110100, MCLR=low (reset PIC, programmer disconnected)
    delay(1);                                       // 1ms delay
    biosprint(0,32   +4,PTR);                       // set bits 100100, MCLR=high
    return 0;
    }
  else if (cmd ==PROGRAM_MODE)                      // enter program mode
    {
    biosprint(0,32+16+4,PTR);                       // set bits 110100, Vpp off, MCLR= low (reset PIC16C84)
    delay(10);                                      // 10ms allowing programming voltage to stabilise
    biosprint(0,8,PTR);                             // set bits 001000, Vpp on, MCLR=13.5V, clock and data connected
    delay(10);                                      // 10ms allowing programming voltage to stabilise
    return 0;
    }
  else if (cmd ==RUN)                               // disconnects programmer from device
    {
  biosprint(0,  32+4,PTR);                          // set bits 100100
  return 0;
    }
  else return PROGMR_ERROR;                         // command error
}
```



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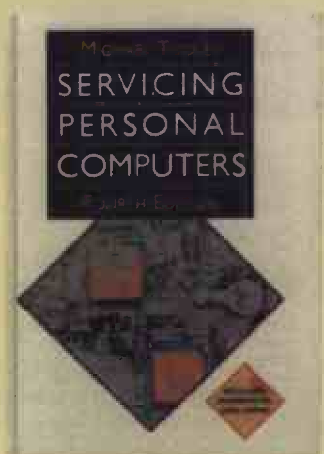

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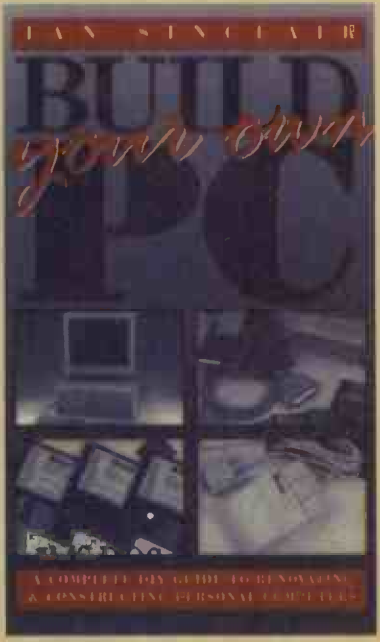

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Winning designs II

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Fourth prize – 50Hz/60Hz inverter uses a single IC

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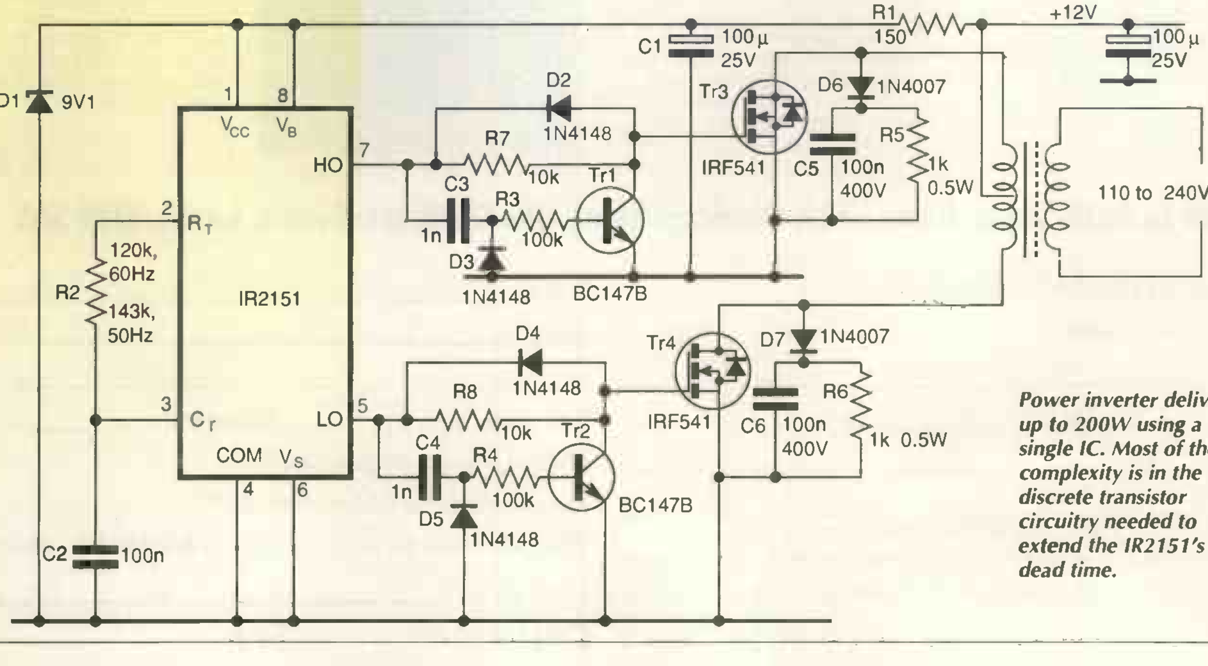
Although the dead time of about 1μs generated by the IC is sufficient to prevent the cross conduction of the mosfets, it is not sufficient to allow the transformer ringing signals to be dissipated in the appropriate snubber circuits. Fortunately, dead time can be increased to the desired value by adding a few inexpensive



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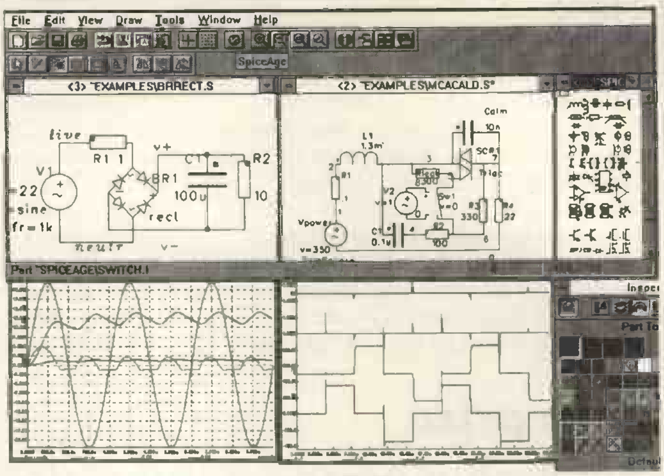

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Farnell P.S.U. TSV70 Mk11 0-35V/0-10A+70V/0-5A	£250
Wayne Kerr auto balance universal B642	£200
Wayne Kerr universal bridge B224	£200
Wayne Kerr component meter B424/N resistance/capacitance/inductance	£195
Wayne Kerr automatic LCR meter model 4225	£300
Norma precision wattmeter D4155	£350
Norma multi function meter D4135A	£300
Norma Model D5155 AC-Power analyser	£800
EIP 575 source locking microwave counter opt. 02.04 10Hz-18GHz	£1.650
Systron Donner 6054B microwave counter 20Hz-18GHz	£650
Thandar TA2160 20MHz logic state analyzer	£275
Schlumberger 7055 microprocessor volt meter	£250
Schlumberger A200 Digital voltmeter	£50
Schlumberger A220 Digital voltmeter	£50
Schlumberger 7055 Microprocessor voltmeter	£250
Wavetek 20MHz sweep/modulation/generator type 193	£250
Wavetek Model 3000 Sig. generator 1-520MHz	£300
Wavetek Sweep generator Model 164 30MHz	£300
Redifon synthesized R1001 15KHz-30MHz all modes	£500
Redifon R500 Brand New (back lit) HF	£750
Redifusion VLF/LF (in current use with Royal Navy Submarines)	£750

Racal HF Receiver RA329B consisting of RA217D receiver and MA323 FSK loudspeaker terminating unit. Frequency range 1-30MHz All mode reception	£275
Harris automatic antenna coupler 1.6-30MHz	£500
Bird 43 Thruline watt meter	£100
Rhode & Schwatz 0-2000MHz UHF attenuator 50ohm 0-140dB	£125
Tektronix 454 150MHz Dual trace scope	£150
Tektronix 453 100MHz Dual trae scope	£150
Tektronix 475 200MHz Delay sweep/dual trace	£450
Tektronix 465 100MHz Delay swep/dual trace	£375
Tektronix 2215A 60MHz delay sweep dual trace	£375
Tektronix 7854 c/w 7511 plug-in/7111 fitted with S1 sampling heads – 1GHz	£1600
Tektronix 492 spectrum analyser 50KHz-18GHz	£4500
Tektronix curve tracer 577	£750
Crotech 3131 15MHz Dual trace scope	£125
Gould OS250A 15MHz Dual trace scope	£120
Gould OS255 15MHz Dual trace scope	£120
Gould OS300 20MHz Dual trace scope	£175
Gould OS3500 60MHz Dual trace scope	£200
Gould OS3600 100MHz Dual trace scope	£275
Gould J3B 10Hz-100KHz LF Oscillator	£150
SE LABS 111 18MHz Dual trace scope	£85
Telequipment D1011R 10MHz Dual trace scope	£75
Telequipment DM63 50MHz 4 trace storage scope	£150
Telequipment D67 25MHz Dual trace scope	£75
Telequipment DM64 10MHz Dual trace scope	£75
Trio CS1566A 20MHz Dual trace scope	£150
COU150 35MHz delay timebase scope	£95
Kemo phase meter DP1	£95
Siemens 7KB4303 Wattmeter	£300
Radiometer AFM38 mod. meter	£300
Aplab 8559 500VA variable frequency convertor	£800
Avo RM215-F/1 insulation resistance and breakdown tester	£250
Pegelmesser 200Hz-620KHz model D2155	£500
Fluke 8000A Digital multimeter	£45
Time CD Millivolt pot source Model 404N c/w DC current source Model 505	£250
Nagra T instrumentation tape recorder	£1200
Datron 1061A Autocal digital multi meter True RMS AC/Current	£700
Datron 1051 Multifunction meter	£65
Datron 1055 DC Voltmeter	£65
Datron 1030A RMS Voltmeter	£65
Datron 1030 RMS Voltmeter	£65
Leader LCR Bridge LCR resistance/capacitance/inductance D, Q measuring range 9V operation	£150
Haven temperature calibrator OTB5 oil/water bath	£500
Haven thermocal i.s. thermocouple simulator/calibrator	£600
Avo 8 Mk 5 + 6 c/w case, leads, prods etc., (c/w cal cert NPL)	From £85
Avo CT160 valve tester c/w valve data book	£75

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components to the basic inverter circuit.

Supply voltages V_{cc} and V_B, and hence the outputs HO and LO, are limited to 9.1V by components R_1, C_1 and D_1. Oscillator frequency is set by components R_2 and C_2. When high side output HO goes high, transistor Tr_1 holds the gate of power mosfet Tr_3 low for a dead period of about 300μs, determined by components R_3 and C_3.

Diode D_2 discharges the gate capacitance of Tr_3 instantaneously when the HO output goes low. Although the turn-on time of Tr_1 with this arrangement increases to about 30μs, it is a very small part of the period of this inverter output.

Components Tr_2, C_4, D_5, R_4, D_4 and R_8 provide an equal dead period to the low side output LO. Components R_5, C_5, D_6 and R_6, C_6, D_7 provide the snubber action for $Tr_{1,2}$.

This circuit configuration is not just limited to 50/60Hz inverter. High-frequency inverters with ferrite-core transformers, and having proper dead times, could form elements of electronic ignition systems and dc-to-dc converters.

By changing R_1, the inverter can be powered by batteries of other voltages.

M. S. Nagaraj
ISRO Satellite Centre
Bangalore

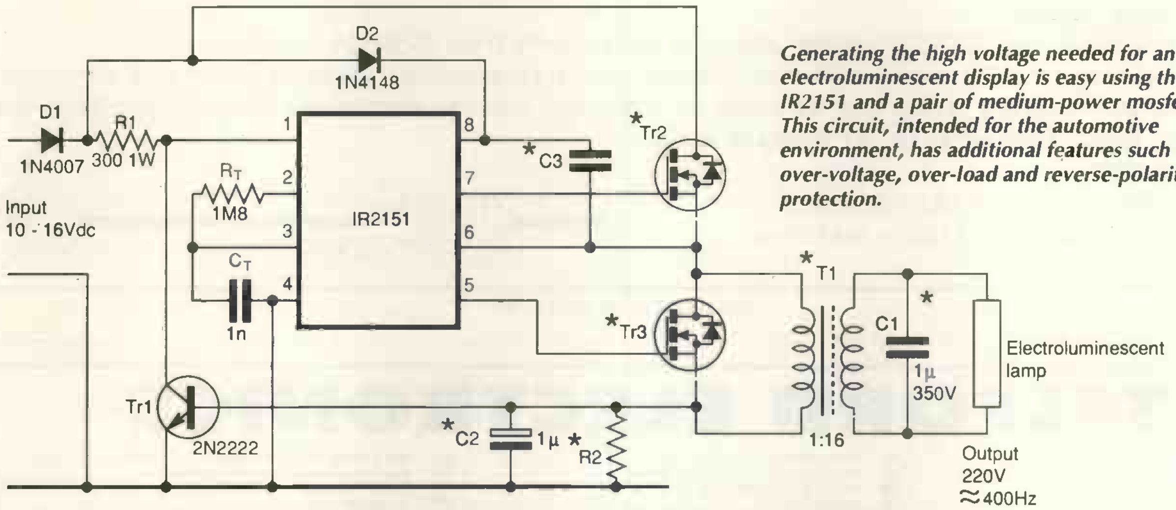
Fifth prize

Electroluminescent lamp driver for automobiles

This driver for electroluminescent displays incorporates all the features normally needed for automotive electronic circuits. Diode D_1 serves as protection against reverse polarity. In the event of an overload on the output transformer, Tr_1 robs V_{cc} of voltage and disables the circuit. Capacitor C_1 takes care of radio-frequency interference.

Values quoted for C_T and R_T cause astable operation at 400Hz, which is suitable for this type of lamp. Over-voltage protection is taken care of by the IR2151's internal zener diode.

Clyve J. Caines
Nairobi
Kenya



Generating the high voltage needed for an electroluminescent display is easy using the IR2151 and a pair of medium-power mosfets. This circuit, intended for the automotive environment, has additional features such as over-voltage, over-load and reverse-polarity protection.

R1 is 5% carbon film, C_T is polyester, C2 is 1μ or more, items marked * are determined by the size of lamp used

Sixth prize

1-to-3 phase converter

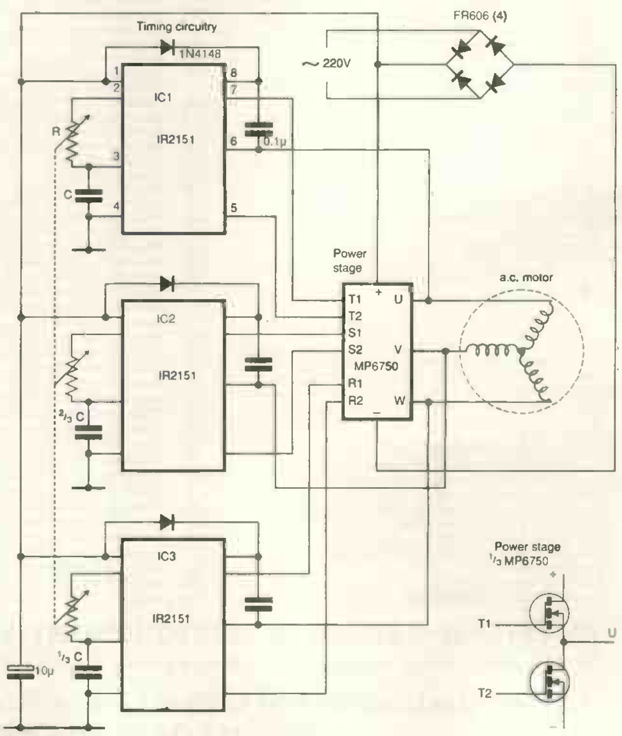
Rotary speed of a three-phase motor depends on the frequency of the applied voltage. The motor is linked to the converter by six power transistors contained in a type *MP6750* module from Toshiba, a third of which is shown bottom right in the diagram.

Inductance of the motor windings acts as an integrator that converts the pulses of varying widths into a sinusoidal signal.

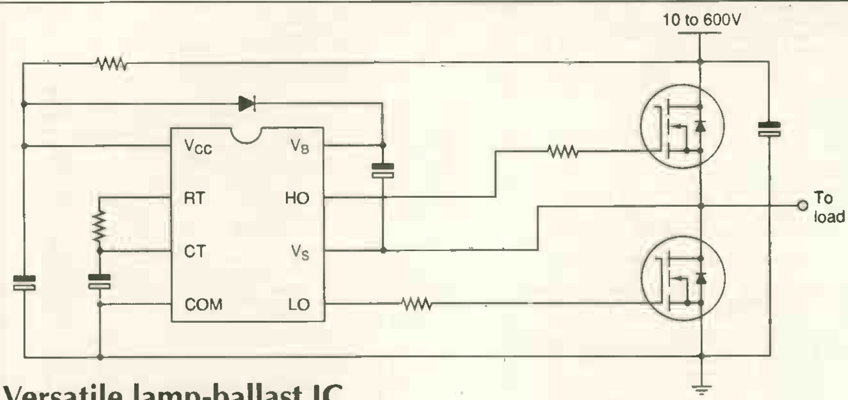
The converter is based on three *IR2151* timer circuits. Each voltage supplied to the three pairs of power transistors is phase shifted by 120°. This is done by reducing the timing capacitors of each *IR2151* progressively by a third, i.e. C, $^2/_3C$, $^1/_3C$.

Frequency of the timing circuit, and hence the speed of the motor, is set to a desired value by altering the resistance of three equal resistors, R, only. This method of converting single-phase mains to three-phase can be used to control small three-phase motors up to approximately 700W, irrespective of whether they are synchronous or asynchronous types.

Kamil Kraus
Rokycany
Czech Republic



This 1-to-3-phase mains converter is suitable for driving small motors up to 700W, regardless of whether they are synchronous or asynchronous.



Typical connections for the IR2151 in self-oscillating mode show that the device needs few external components. Power for the high-side switch gate comes from a 1μF bootstrap capacitor. This is charged to around 14V whenever V_s is pulled low during low-side power switch conduction. The fast-recovery bootstrap diode blocks dc bus voltage when the high-side switch conducts.

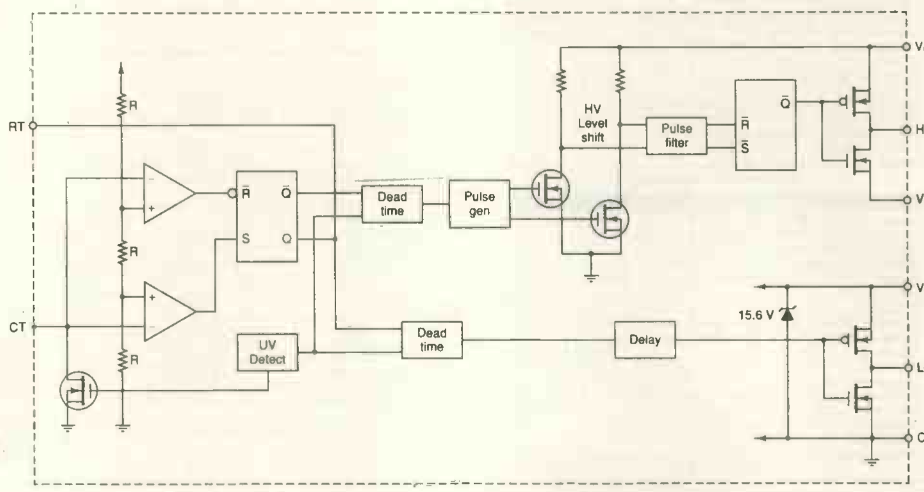
At the front end of the IR2151 is a timing circuit that is very similar to the established 555. Two timing pins are available externally, opening up the possibility for numerous applications other than lamp ballasting. Dead-time generators are incorporated to ensure that the two power mosfets being driven by the device do not conduct simultaneously.

Versatile lamp-ballast IC

The *IR2151* is a fluorescent lamp ballast but, as you can see from the three designs presented here and those shown last month, the device has many potential uses. As the top diagram illustrates, the *2151* is essentially a *555* timer with integral level shifting and power-mosfet drive circuitry. A typical application circuit is shown in the lower diagram.

Features of the device are,

- Floating channel bootstrappable
- Operates to 600V
- Tolerant to negative transients
- dV/dt immune
- Undervoltage lockout
- Programmable oscillator frequency
- Matched channel propagation delay
- Low side in phase with RT pin ■



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ACTIVE

A-to-d and d-to-a converters

10MHz sampling a-to-d. Datel's *ADS-119* is a 12-bit, 10MHz sampling converter with guaranteed no missing codes over the military range of temperatures. Signal-to-noise is 69dB and thd –68dB, both around 6dB better than other devices available. It is usable in military and industrial areas and contains a sample/hold amplifier, control and timing logic and error correction. Datel (UK) Ltd. Tel. 01256 880444; fax 01256 880706.

Discrete active devices

Power rectifiers. Surface-mounted bridge rectifiers by Shindengen handle reverse voltage up to 800V. *S1Z* types are contained within a 5.9 by 2.6mm footprint, 3mm thick, and are believed to be the smallest and lightest available. Somewhat larger and 30% cheaper are the *S1N* series with ratings from 100V to 800V. Flint Distribution. Tel., 01530 510333; fax 01530 510275.

DSP modules. First in LSI's new range of digital signal-processing modules based on the Texas *TMS320C44* are the *MDC44T* and *MDC44S*. LSI's hardware provides more functions on the single-width TIM-40 standard, so that the *C44* is used to the full. Both modules possess increased memory – in the case of the *MDC44S*, 8Mbyte of sram, more than five times that of the earlier *MDC40S*. In the *44T*, two 50MHz *C44* processors allowing improved access to the global memory bus increase flexibility and the true Harvard architecture, which enables simultaneous data and program fetches, maximises the exploitation of the *C44*'s processing capabilities. *MDC44S* has one processor and is expected to find application in graphics and imaging, where large zero-wait-state memory is essential. Both 40MHz and 50MHz grades are to be available to provide maximum operating speeds of 50Mflop/s. Loughborough Sound Images Ltd. Tel., 01509 634300; fax, 01509 634333.

Digital signal processors

PC104 DSP board. *OROS-SP104* is a digital signal-processing board using the TI *TMS 320LC31* processor, the low-power version of the '*C31* taking 120mA at 33Mflops or 20mA when idle. It is in the *PC104* format, with 128Kbyte-1Mbyte of ram and can be programmed in C for which a set of development tools is provided. An interface in the host pc is also used for initial program loading. Oros. Tel. (France), 00 33 76 90 62 36; fax 00 33 76 90 51 37.

Linear integrated circuits

GaAs MMIC amplifier. Samsung's *SMP-11206* microwave IC is for use in applications up to 5.5GHz in the industrial, scientific and medical bands. It provides 12dB of gain at 2.4GHz, working from one 4V-9V rail, at a low voltage-standing-wave ratio. Dc blocking is provided on the radio-frequency output. Anglia Microwaves Ltd. Tel., 01277 630000; fax 01277 631111.

Logic

Caller ID chip. Mitel has introduced the *MT8843* caller line identification circuit, intended for the caller line display service announced by BT and for similar services elsewhere. It provides all alerting tone detection required by BT and in caller line ID on call-waiting systems. Guard time is programmable and the device meets BT's requirement for loop-reversal detection, which is used in applications other than BT's caller ID to provide a ringing detector. Mitel Semiconductor. Tel., 01291 430000; fax 01291 430400.

Memory chips

'Densest' serial memory. At 128Kbit, Xicor's *X25128* is claimed to be the world's densest serial eeprom. It is meant to support the *Serial Peripheral Interface* and has a 2MHz bus frequency, 2.7-5.5V working and a current requirement of less than 1μA. Memory can be partitioned into blocks, with a feature called block lock, with levels of write protection, so that access is allowed to some portions while data in others is protected. Micro Call Ltd. Tel., 01844 261939; fax 01844 261678.

Thin sram. EDI has a number of 4Mbit static rams in the Thinpack ceramic package, which is only 1.9mm high, for use in military or high-reliability applications. Leads of the package are trimmed and formed and compatible with the plastic TSOP Type II pack. Micro Call Ltd. Tel., 01844 261939; fax 01844 261678.

Mixed-signal ICs

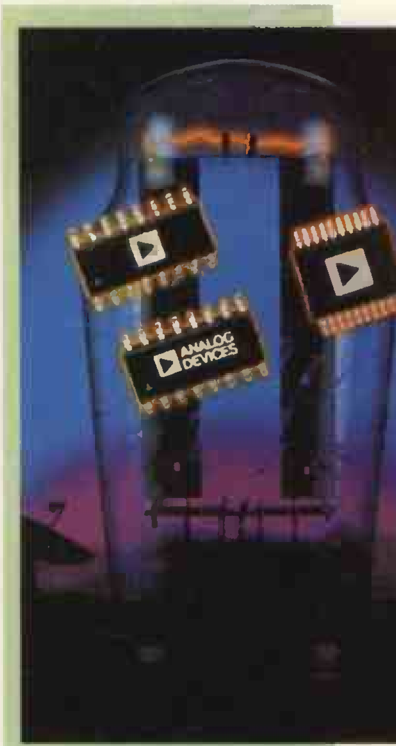
Battery capacity monitor. Meeting the requirements of the Intel/Duracell System Management Bus and Smart Battery Data, Benchmarq's *bq2040* capacity monitor is for use with NiCd, NiMH and lithium-ion batteries, sending information on mAh capacity on the SMBus or indicating capacity directly by leds. The device is in a 16-pin SOIC. Sequoia Technology Ltd. Tel., 01734 258000; fax 01734 258020.

Avalanche photodiode. InGaAs/InP avalanche diodes in the *EG&G C30644* and *C30645* series are available in the UK. Featuring a gain of 10, they are intended for use in optical-fibre communications operating at 1300-1550nm. Fibre pigtails can be supplied with both devices, the *C30644* with a monomode pigtail and low back reflection or with a grin lens and multimode pigtail. Quantum efficiency is 85% and responsivity 8.9A/W at 1300nm. Pacer Components Ltd. Tel., 01734 845280; fax 01734 845425.

Laser diodes. AlGaLnP red laser diodes by Sanyo are now obtainable here. The 635nm diode has a threshold current of 50mA, an optical power output of 5mW and operating temperature 50°C, while the 670nm type works at 30mA and 60°C and has a high-speed response. Jayex Components Ltd. Tel., 01734 810799; fax 01734 810844.

Programmable logic arrays

135MHz PLDs. Lattice Semiconductor's *ispLSI* and *pLSI 2032* are 32-macrocell E^2CMOS, high-density programmable logic arrays, the 'isp' meaning in-system programmability. They are both 80MHz, 110MHz or 135MHz, 7.5ns devices, supporting the *Pentium* and *Power PC* processors. Each has 32 registers and universal i/os, two dedicated inputs, three dedicated clock inputs and a dedicated global output enable, all being connected by a global routing pool. Micro Call Ltd. Tel., 01844 261939; fax 01844 261678.



Single-chip solutions

Rf receiver. Integrating all the components of a receiver's rf/if strip,the *AD607* and *AD608* from Analogue Devices feature an ultra low-power architecture. The two ics are designed for wireless systems using a minimum supply voltage of just 2.7V down to –25°C and consume less than 25mW. They are suitable for applications using protocols such as GSM, cdma or tdma.
The *AD607* has a linear IF amplifier with 100dB range whereas the *AD608* has a logarithmic amplifier with 90dB of RSSI range and limited output. Both devices have low noise mixers a with 500MHz bandwidth as the first stage with an internal preamplifier which requires only –16dBm of LO drive. The mixer output will drive an industry standard 10.7MHz 330Ω filter. Am, fm, cw and ssb are all demodulated by the *AD607* with the *AD608* providing fm and pm demodulation capability. Analogue Devices. Tel 01932 232222

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PASSIVE

Passive components

Modem transformer. Integrity Technology introduces the *T14Z* telephone line-matching transformer, for a transmission speed to 28.8Kb/s in V.34. Using an EI, 14mm alloy core and a phenolic bobbin, the transformer is contained in a package about 0.5in cube with hard copper pins. It is designed to connect directly to 600Ω lines with zero bias. Primary winding is rated at 80mA continuous and 125mA in the ringing cycle. Integrity Technology Corporation. Tel., (USA) 00 408 262-8640; fax 00 408 262-1680.

Pulse transformer. Occupying a mere 0.625in square of board space, the same as an earlier single type, DDC's dual pulse transformer is meant for use in MIL-STD-1553 dual redundant data bus systems. It is in diallyl phthalate encapsulation and is available in through-hole, surface-mount or flat-pack versions. All types have centre-tapped primaries and multiple taps on secondaries to cope with existing systems. Data Device Corporation. Tel., 01635 40158; fax 01635 32264.

Metal-film chip resistors. Gothic Crellon has a series of precision metal-film resistors that offer good pulse stability in single pulses up to 200W for 1µs. Resistance range is 100Ω-100kΩ at ±0.1% in the E24 or E96 series of values. Temperature coefficient is less than 2.5×10^{-5}K, power dissipation 0.125W maximum, voltage rating 100V dc or rms maximum and thermal resistance 170K/W. Gothic Crellon Ltd. Tel., 01734 788878; fax 01734 776095.

Connectors and cabling

SM wire-to-board connector. Claimed to be the smallest available, the *Molex 53261* connector has a mounted height of 8.5mm on a 1.25mm pitch, with a 250V, 1.2A rating per contact. Connectors are supplied in tapes and reels and there is a cable assembly service on offer. Flint Distribution. Tel., 01530 510333; fax 01530 510275.

Network outlets. MOD-TAP's range of very low profile wall and floor outlets are for use in boxes of 16mm depth and are compatible with Euromod and Modsnap accessories. They can be floor-mounted to either flat metal or plastic faceplates with an 18mm clearance. MOD-TAP Ltd. Tel., 01703 701919; fax 01703 704063.

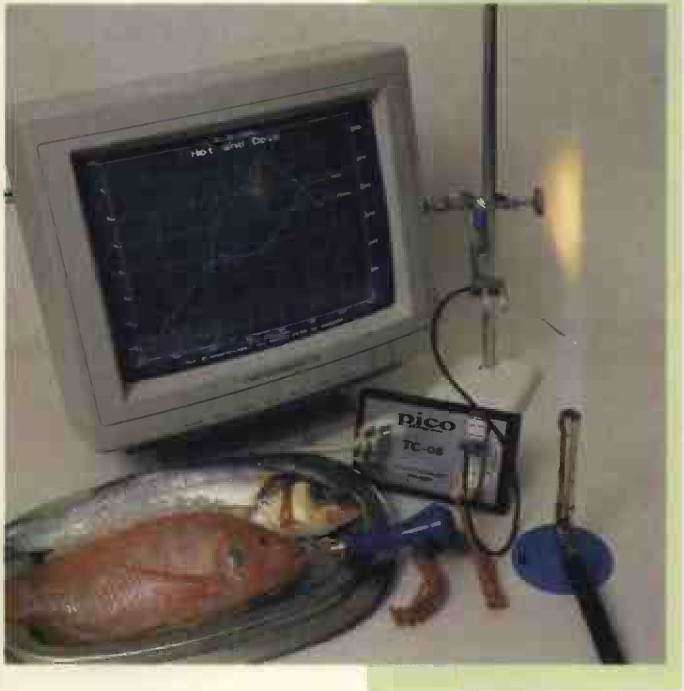
Pcb terminals. Pcb terminals pitched at 3.5mm and 3.81mm from *Wieland* come in standard and pluggable versions, with the standard type as a vertical or horizontal connection, this having a fixing cam. Pluggable units are of pin-strip or edge-card connection and plug-and-socket. Ratings cover 6A/125V to 12A/125V. Wieland Electric Ltd. Tel., 01483 31213; fax 01483 505029.

Static control connectors. For establishing connection between different connector types, 4mm and 10mm for example, in static control applications, TBA has a range of kits. Where no ground point is present, a hand tool enables the user to fit studs to make the connection to accept 4, 7 or 10mm studs and banana plugs. TBA Industrial Products Ltd. Tel., 01706 47422; fax 01706 46170.

DIN 41612 connectors. *Apfel* DIN 41612 connectors in the *U/L* range are now obtainable from Westfield. Types available include *B, C, Q* and *R* (including half types), pcb-mounting solder, wire-wrap, press-fit and idc ribbon models in 16-96 ways. Contacts are of the dual-beam type in a number of forms. Westfield Distribution Ltd. Tel., 01488 685183; fax 01488 685430.

Sensors

Thermocouples. A package that includes the *PicoLog* datalogging software, Pico's *TC-08* is a thermocouple to pc interface. The unit requires no power supply and connects to the pc via the serial port. Eight different thermocouples can be accomodated (B, E, J, K, R, S A and T types) and the software allows samples as fast a once a second and as slow as one per hour. Advanced temperature processing functions include filetring, min/max detection and alarm setting. A real time display is available in either graphical or text format. Pico Technology Ltd. Tel. 01954 211716



Displays

Contrasty LCDs. Hitachi's *LMG7380* graphic/alphanumeric liquid-crystal display has a contrast ratio of 18:1, a six-times improvement over the earlier *LMG6380*, achieved by the use of a film-retardation layer and a fluorescent backlight. Size is 160 by 68 by 11mm. Other units in the range include the *LMG5738XUFC-OOT* VGA display and the *LM9520RPCC* 320 by 240 colour and IC controller. Eiger Technologies Ltd. Tel. 01928 579009; fax 01928 579123.

Filters

Piezo IF filter. Narrow-passband filters from Murata, the *SFE10.7MV5* and *SFE10.7MT* for fm radio and am up-conversion, use the second overtone vibration mode to achieve ±13kHz bandwidth and 35dB spurious response suppression. Murata Electronics (UK) Ltd. Tel. 01252 811666; fax 01252 811777.

SAW filters. Surface transversal acoustic wave filters by GPS are on quartz and offer a –20°C to 80°C temperature range, compared with the 0-40°C range in other materials; group delay is less than 150ns. First available is the *DW9249*, intended for use in the IF in DECT digital cordless telephones, operating at a centre frequency of 112.32MHz with a –3dB passband of 1.152MHz. Adjacent-channel rejection is 20dB. GEC Plessey Semiconductors Ltd. Tel. 01793 518510; fax 01793 518582.

Hardware

Sealed cases. Meeting the IP54 rating, instrument cases in ABS by *Serpac* are suited to use outside and in hostile industrial applications. They come in sizes from 57 by 92 by 38mm to 83 by 143 by 64mm and are sealed by a gasket between the box and lid, which is secured by four or six self-tapping screws, and O-rings for the screw holes. Options include the I series with a recessed top for a membrane keypad, the standard type with a flat top and a slanted model for wall mounting, all with pcb mounting pillars. OKW Enclosures Ltd. Tel. 01489 583858; fax 01489 583836.

Personal DSOs. Tektronix has produced a range of digitising storage oscilloscopes at a price low enough that they can be considered personal instruments while retaining lab. instrument accuracy and performance. The range of *TDS 400A* instruments encompasses bandwidths from 200MHz to 400MHz with sampling to 100Msample/s and offers, when coupled with a range of accessories, an array of features for all kinds of electrophysical measurement. Both new models have a graphical user interface and offer automatic measurement of 25 parameters; there is also an FFT/maths option and a 3.5in floppy drive to allow results to be saved and imported to Windows and Macintosh applications. One of the accessories is the *P5200* high-voltage differential probe, which allows the 'floating' measurement of voltages up to 1300V when no ground point is available, the *P5200* converting the floating voltage to a ground-referred one with no capacitance penalty. Record length is 120K to allow a complete sight of a long signal while retaining the ability to see detail. Tektronix UK Ltd. Tel., 01628 486000; fax, 01628 474799.

Instrumentation

Sound-intensity probe. Improvements to *Bruel & Kjaer's* sound intensity probes increase physical robustness and extend

Snap-in bezels. *RMF* bezel/filter assemblies snap into a suitable panel aperture without the assistance of screws or fasteners of any kind. The ABS bezels are available in a range of sizes to match most types of display, black being standard and colours available to order. Combined coloured filters have a non-glare surface and come in red and clear for leds and lcds, colour again being available to order. Panel thickness required is 0.04-0.125in. and the range of sizes is from 1.343 by 1.906in to 1.656 by 8.531in. UV-Tec Ltd. Tel., 01252 844880; fax, 01252 844885.

frequency range. A stainless steel alloy brace replaces all the existing mechanical parts, doing away with adaptors and angle pieces and placing the probes in the IEC 1043 Class I. They come with a pair of microphones matched in phase and amplitude, a spacer allowing a 7.1kHz frequency range. Upgrade kits for types *3545*, *3547* and *3548* are available. Bruel & Kjaer (UK) Ltd. Tel. 0181 954 2366; fax 0181 954 9504.

350MHz DSO. Gould's new *DataSYS 940* is a low-noise digital storage oscilloscope with a 350MHz bandwidth. This figure is guaranteed on all ranges from 5V/div to 2mV/div. Display modes include refresh, persistence, roll, X/Y and pre-trigger and live zoom. Each of four channels has a 50,000-word memory. There are also measurement and analysis functions, configurable to user requirements. Gould Instrument Systems Ltd. Tel. 0181 500 1000; fax 0181 501 0116.

Power analyser. *Dranetz PP1*, now for hire from Livingston, is a trouble-shooter for power-line problems such as those causing motors to burn out and computers to crash when, according to chart recorders, all seems well. The *PP1* reports on individual cycle problems with voltage and current, monitoring trends in voltage, current, power, volt-amps, power factor, harmonics, kilowatt-hours and demand, being programmed from any of these or from external events such as the switching of equipment. Livingston Hire Ltd. Tel. 0181 943 5151; fax 0181 977 6431.

Stepper drive. Digiplan's *PDHX* is a 4000step/rev ministepper drive, combining all motion-control facilities and power supply in one unit, intended to supply the requirements of point-to-point applications needing good dynamics, smooth, low-speed rotation down to 0.001rev/min and fast response; it reacts to a registration signal from a sensor in under 15μs. An optically isolated 24V i/o interface allows input from thumb-wheel switches, remote controls and plcs. The self-adaptive, switched-mode psu handles 100-250V input and supplies a 70V, 5A bus for high-speed torque at speeds up to 50rev/min. Indexing is powered by the built-in supply and is programmable over an RS232C serial link from any computer in Parker's X-code control language, which has over 160 high-level commands, maths functions, subs and constructs such as If-THEN-ELSE and REPEAT loops. Parker Hannifin plc, Digiplan Division. Tel., 01202 699000; fax, 01202 695750.

Programmable function generator. Providing full digital control by way of a GPIB interface, TTi's *TG1304* analogue function generator's output frequency is accurate to within 0.01%. Two separate function generators in the instrument cover 0.01Hz-13MHz from 50Ω and 0.005Hz-50kHz from 600Ω, the lower-frequency circuit being mainly intended to sweep and modulate the other, although it is suitable for use on its own for sine, square and triangle waves. Both generators output 2mV-20V pk-pk. Thurlby Thandar Instruments Ltd. Tel. 01480 412451; fax 01480 450409.

Audio analyser. Trio-Kenwood has introduced the *VA 2230*, a microprocessor-controlled combination instrument including a 0.005% distortion, 5Hz-100kHz audio generator; voltmeter for ac and dc reading in volts, dBu and watts; a distortion meter giving total distortion, harmonic distortion and its analysis; and a frequency counter. S:n, channel ratio and sinad are also given. The 100-point programmable memory and GPIB port are provided. Trio-Kenwood UK Ltd. Tel. 01923 816444; fax 01923 819131

Video test generator. *VG-812* is an rgb programmable video test generator by Ginsbury that has a 40-program memory. It is intended for testing computer monitors and can be programmed for timing, colour and pattern by the user, a high-speed clock allowing the evaluation of high-resolution monitors. It has an RS-232 port for remote control by a PC. Ginsbury (UK) Ltd. Tel. 01634 290903; fax 01634 290904.

Clamp meters. Clamp meters by *Yokogawa* have both digital and analogue displays, the analogue readout being in the form of a fan-shaped bar graph to simulate a moving-coil meter. As well as the usual current measurements found on clamp instruments, these provide digital multimeter functions, a peak-hold function being available and analogue output for recording. Martron Instruments Ltd. Tel. 01494 459200; fax 01494 535002.

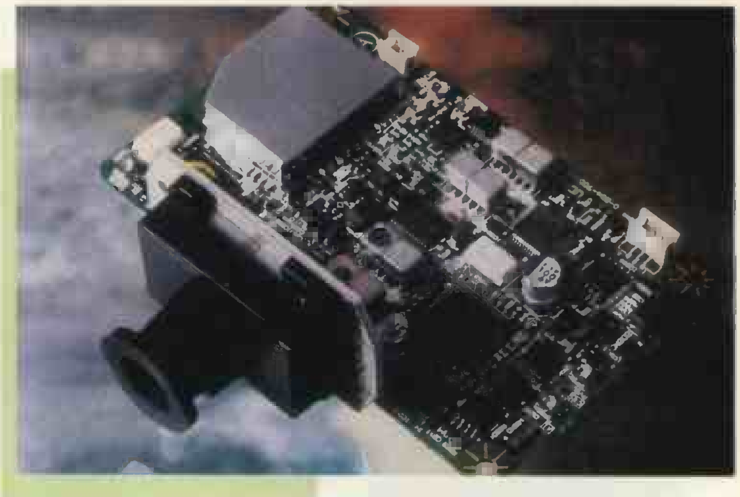
Electric-field meter. Holaday has the *HI-3638* to measure ELF/VLF electric fields in the 5Hz-400kHz range. Its sensor is cable connected to the digital readout to give complete isolation and fields from 0.4V/m to 40kV/m can be measured. Holaday Industries. Tel. 01628 478155; fax 01628 476871.

Microwave leakage monitor. For continuous monitoring of microwave fields, the *HI-2602 Microwave Interlock Monitor* takes input from a single remote sensing probe. Relay contacts close and a led illuminates if microwave level exceeds a preset limit up to 1mW/cm^2. Holaday Industries. Tel. 01628 478155; fax 01628 476871.

Interfaces

Analogue-input VME. *DVME-614* analogue-input VME boards by Datel have dual, high-speed a-to-d converters to avoid phase skew in synchronous channels. In the *614F*, two channels are sampled at up to 2MHz each at 12-bit resolution, while the *614G* variant has two 1MHz, 14-bit channels. Both have fifo memory options to 16Ksample to prevent data loss by sending bursts of samples to the host while conversion proceeds. Datel (UK) Ltd. Tel. 01256 880444; fax 01256 880706.

16-channel analogue i/o. LSI's *PC/16IO8* simultaneously samples 16 12-bit analogue inputs at up to 25kHz and up to 48kHz with fewer channels, each channel having its own a-to-d converter for speed and phase alignment. Eight analogue outputs come from two 12-bit d-to-a converters at update rates of up to 100kHz. There is a *DSPLink* parallel interface for control and to enable the *PV/16IO8* to be used with peripherals and other i/o boards. Eight buffered ttl-level digital i/o channels allow integration with the host dsp. Loughborough Sound Images Ltd. Tel. 01509 634300; fax 01509 634333



Vision systems

Camera on a card. Sony's *CCB-GC5/P* series of cameras are card mounted (54 by 86mm) providing a 330-line resolution from a $^1/_3$in imager in both NTSC and PAL formats, automatic exposure control and auto tracing white balance. Options include a choice of lens, a Y/C output and cables to allow the ccd head to be up to 150mm away from the board. Similar in other respects, the *CCB-GC7YC/P* gives a 470-line resolution through the Y/C output. Supply is 7-13V. Sony Computer Peripherals & Components. Tel., 01932 816000; fax, 01932 817001.

Literature

Op-amps. Harris's *High-performance Op-amps and Buffers* brochure gives full details of the voltage and current feedback op-amps and buffers with bandwidths in the 45-858MHz range introduced since the company's 1993-4 databook. As well as performance figures, there are details of the special features of op-amp design. Harris Semiconductor UK. Tel. 01276 686886; fax 01276 682323.

Connector catalogues. Four catalogues from *Robinson Nugent* describe interconnection products. There is a pga socket brochure on sockets for the *Pentium;* a catalogue on *PAK 5* and *PAK 8* smt fine-pitch board-to-board connectors; a third on *MEMPAK* PCMCIA connectors; and one on 2mm products. Robinson Nugent (Europe) Ltd. Tel. 0031 4990 75755; fax 0031 4990 77155.

Noise suppression. *Panasonic* offers a catalogue of noise suppression and filtering components for power line, signal line and surge pulse protection. Components described include line filters, ceramic-disc and chip capacitors, chokes, emi filters and bead cores and inductors. There is also an introduction to the law and regulations on noise and the principles of its suppression. Panasonic Industrial (Europe) Ltd. Tel. 01344 853827; fax 01344 853803.

Gas plasma displays. A colour brochure from *Cherry* describes the company's latest range of gas plasma displays, including the *Plasmadot* family of full-field dot-matrix types and a series of interface controllers and dc-to-dc converters. Technical details and some application information are included. Cherry Electrical Products Ltd. Tel. 01582 763100; fax 01582 768883.

Light measurement. Instruments for light measurement, applications and the basics of radiometry and photometry are all described in a new catalogue from *International Light* of Massachusetts, which also includes tutorial information. International Light Inc. Tel. (USA), 00 508 465 5923; fax 00 508 462 0759.

Flexible circuit material. Rogers Corporation offers a colour brochure describing, in general terms, its capabilities in the manufacture of flexible pcb substrates in polyimide-based laminates or all-polyimide material and adhesive types in acrylic, butyral phenolic and epoxy. Copper

Please quote *"Electronics World + Wireless World"* when seeking further information

Energy management controller. Microchip says its new *MTE1122* IC, developed in partnership with Coast Energy Management, will reduce power consumption in consumer and industrial equipment using ac motors by up to 30% by the use of the company's *PIC16/17* 8-bit, risc-based microcontroller. Motor load is digitally monitored several thousand times per second and power consumption controlled as required by varying the ac signal to allow a constant speed at reduced power. Arizona Microchip Technology Ltd. Tel., 01628 851077; fax, 01628 850259.

cover is rolled annealed or electro-deposited. Rogers Corporation. Tel. (USA), 00 203 774 9605; fax 00 203 774 9630.

Sensors. *The Sensor Technology Sourcebook* lists and describes companies and organisations involved with sensors of all kinds by application and alphabetically, with names and telephone/fax numbers, lists databases, technology transfer specialists and commercial products. Technical Insights Inc. (USA). Tel. 00 201 568 4744; fax 00 201 568 8247.

Materials

EMI/RF seals. At prices not much more than those for ordinary dust and moisture seals, James Walker's *Shieldseal 107* conductive elastomer in standard or custom profiles achieves IP65 rating and provides broadband shielding with an attenuation of 120dB E-field at 1MHz, 90dB at 100MHz and 50dB plane wave at 10GHz. Volume resistivity is 6Ω/cm. James Walker & Co, Ltd. Tel. 01483 757575; fax 01483 755711.

CFC-free foam. *Jiffycel* foam packaging protection is now made without chlorofluorocarbons or other harmful chemicals, while retaining its physical properties of reusable resilience and capability of being recycled. Jiffy Packaging Co. Ltd. Tel. 01606 551221; fax 01606 592634.

Power supplies

Fixed-*f* smps. Using fixed-frequency switching to meet EN55022 level B EMI needs, XP's *NFN 25* and *40* universal-input units are available in 25W and 40W versions and various output voltages including 5, 12, 15 and 24V singles and 5V with ±12V, 5V with 12V and –5V, and 5V with ±15V. Universal input handles 85-264V, 47-440Hz. XP plc. Tel. 01734 845515; fax 01734 843423.

PCMCIA power controllers. Micrel's range of controllers now includes the *Micrel 2561*, a low-cost device in 14 or 16 pin SOIC packaging. It will switch between the three V_{cc} voltages off, 3.3V and 5V and the five V_{pp} voltages off, 1V, 3,.3V, 5V and 12V, selection being by means of two digital inputs for each output. Output current is up to 750mA for V_{cc} and 200mA for V_{pp}. There is full protection for equipment and power supply. Hawke Components Ltd. Tel. 01256 880800; fax 01256 880325.

'Smallest' ldo regulator. Described by National Semiconductor as the smallest and highest-performing low dropout regulator family, the 50mA *LP2980* is one of the company's TinyPak series in the 8.2 mm^2 SOT-23 package. Dropout voltage is 120mV at 50mA and 7mV at 1mA; quiescent current 375μA/80μA. Input voltage is –0.3V to 16V and output voltage, 3V, 3.3V or 5V, accurate to within ±0.5%. National Semiconductor GmbH. Tel. 01049 814110382; fax 01049 814103515.

Efficient regulator. Having a 4-40V operating range, Linear's *LTC1159* high-frequency, synchronous, switching regulator is 90%-95% efficient with loads in the 0.02-2A range while providing 5V from a 10V input. Two external mosfets are driven at frequencies to 250kHz, the unit automatically switching between continuous and burst operation for higher efficiency. Quiescent current is 250μA and 20μA when shut down and dropout is 200mV at 1A and 100% duty cycle. Micro Call Ltd. Tel. 01844 261939; fax 01844 261678.

Radio communications products

Satellite receiver. R L Drake of Ohio has introduced the *ESR410*, a miniature, rack-mounted satellite receiver using synthesised tuning and block conversion. Frequency range is 950-2050MHz and the unit is meant for master-antenna television, commercial or industrial audio, video and data. For audio, there is noise reduction and Wegener stereo compatibility, audio subcarriers in the 5-9MHz range being tuned at the front panel and the unit has three selectable audio IF bandwidths. R L Drake Company. Tel. 00 513 866 2421; fax 00 513 866 0806.

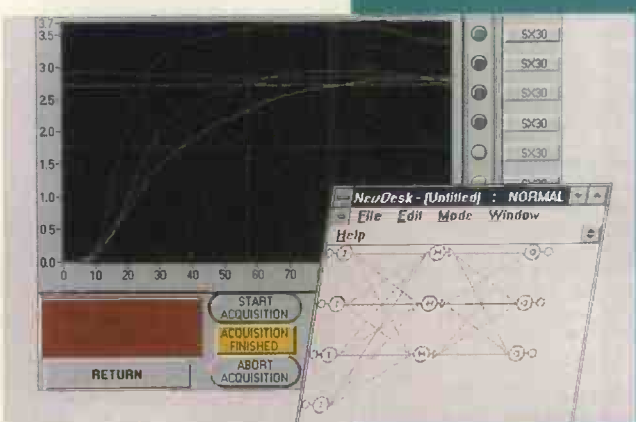
Switches and relays

Photovoltaic relays. Relays in IR's *PVT412* series are designed to meet telecom and electrical safety requirements of the major countries without the need for adaptation. They are single-pole, normally-open solid-state relays using an IC photovoltaic generator and IR's Hexfet power mosfets as output switch. Switching performance is ±400V ac peak or dc at up to 140mA ac and 210mA dc; input/output isolation is 4000V rms. Active internal current limiting in the *PVT412L* meets the FCC Part 68 lightning-surge requirement up to 200A. International Rectifier. Tel. 01883 713215; fax 01883 714234.

Fused isolators. Expanding its range of *Slimline NH* fused isolators, Rittal can now achieve fused protected power distribution to 630A by the use of fuse links NH00, 1, 2 and 3 in widths of 100mm. The range uses double interruption per phase and double arcing chambers; line feed is either from top or bottom and the units fit directly to busbars on 185mm centres. Rittal Ltd. Tel. 01709 704000; fax 01709 701217.

Surface-mounted switches. Jeil offers a range of surface-mounted switches for 12V working, having a contact resistance at 50mA of 100mΩ. Travel is 0.25mm and operating force is chosen from the 100, 160 or 260g range. Height from the board is between 13mm and 1.6mm and overall size is 5.7 by 4mm to 10 by 6mm. A dust-proof type, the *JTP 1290*, is also available. Pedoka Ltd. Tel. 01462 422433; fax 01462 422233.

Neural network for smart noses. Windows-based neural network by NCS, the *NeuRun*, is in use by the French firm AlphaMOS to add automatic decision making to the *Fox 2000* electronic nose, little effort having been needed to embed artificial intelligence in Fox's LabView-based software. *NeuRun* is compatible with pc and windows dos software and hardware, so that it can be used to construct and upgrade applications in a modular manner without the need to program or redesign existing software. In this case, *NeuRun* is embedded as a background task behind *LabView*, odour samples being analysed by *NeuRun*, a decision on the sample made and the decision transferred back for display in less than one second. Neural Computer Sciences Ltd. Tel., 01703 667775; fax, 01703 663730.



Cashpoint keyboards. Programmable keyboards for shops, made by DED are protected from liquid spills and are of the membrane type with feel. *POS-Page 854/865* have 120 keys, covered by a transparent sheet which forms a menu page to indicate each key's function, the page being produced using software supplied and the keyboard itself or simply marked up by pen. Each key can be programmed to generate a macro of up to 16 characters on one level or on each of four concurrent levels. Several options for interfacing to computers, scales and printers are provided. DED Ltd. Tel. 01797 320636; fax 01797 320273.

Transducers and sensors

Piezoelectric film sensors. Strain sensors from Pro-Wave, based on polyvinylidene fluoride film, offer improved sensitivity over ceramic types. A two-pin, pcb-mounted unit, the *FS-2513P*, measures 13 by 25mm, is encased in a moisture-resistant coating and has a sensitivity of 0.5mV/g with a 25-70Hz frequency range. Capacitance and output impedance at 1kHz are 1.5nF and 100kΩ. Quantelec Ltd. Tel. 01993 776488; fax 01993 705415.

Electret microphones. Future Components handles the complete range of Panasonic's electret capacitor microphone cartridges in the *WM-034* and *WM-54* series, intended for use in both industrial and consumer application. WM-54B is a new, 4.5mm deep model with a choice of sensitivity in the –46dB to –40dB ±2dB or ±3dB, working in the 20-16000Hz range at 2.5V. *WM-034B/C* are commonly used types operating at 4.5V with –46dB to –38dB ±3dB sensitivity, while *WM-034D* is a 1.5-10V model and *WM-034F* a 1.5V type for 100-5000Hz working with high and low frequency roll-off for telephone use. Future Components Ltd. Tel. 01279 758999; fax 01279 757676.

COMPUTER

Data communications

Plug-and-play IEEE 488.2 controller. National Instruments's IEEE 488.2 controller board is now available in a ready-to-wear, jumperless version. *AT-GPIB/TNT(PnP)* fits 16-bit ISA plug-in slots and is designed for true plug-and play operation in compatible systems, in which hardware settings are automatically in place at switch-on. In a non-compatible system, the board provides configuration by means of the NI-488.2 software configuration facility that is supplied with the board, in addition to dos and Windows drivers. It is compatible with LabVIEW, LabWindows/CVI and LabWindows applications packages. National Instruments UK. Tel. 01635 523545; fax 01635 523154.

RS232 voltages from transceiver. Linear's *LTC1348* RS232 transceiver IC delivers true RS232 voltages from one 3.3V supply. It is a three-driver, five-receiver DTE unit drawing 500μA and needing only three 0.1μF capacitors for the RS232 voltages. It has four current-saving modes of operation, including a 10μA 'receiver-keep-alive' mode and full protection from esd and overvoltage. *LTC1348* supports data rates up to 120kbaud. Linear Technology (UK) Ltd. Tel. 01276 677676; fax 01276 64851.

Development and evaluation

8051 emulator. *TX51* is a low-cost emulator supporting all rom-less variants of the 8051, including the Dallas *DS320*, up to an oscillator frequency of 30MHz and can be configured for 3V working. The single-cable ROMlink can is also available for the unit. *HiTOP* development environment is used, in which one can debug code in C, PLM or Pascal at source level, and view and modify variables while the program runs. On offer is a free information pack and demo disk. Hitex (UK) Ltd. Tel. 01203 692066; fax 01203 692131.

Programming hardware

Dual programmer. *SMS Sprint Dual* is a twin programmer for both development and production, its most popular version being capable of handling 48-pin dips PLCCs up to 84 pins and a JTAG connector for in-circuit programming. For production, the instrument will program devices in parallel, speed of operation being enhanced by the use of the host pc's cpu and ram. Concentrated Programming Ltd. Tel. 01279 600313; fax 01279 600322.

Software

Windows XRAY Monitor. Microtec offers a Windows version of the *XRAY Monitor* debugger for the Motorola 68000 family, using all the short-cut facilities of windows for speed, including a button bar for common commands and icon dragging. Assembler and high-level source code are visible together and there is context-sensitive help. Configuration to target hardware is easy and even the faster target processors run at full speed. Microtec Research Ltd. Tel. 01256 57551; fax 01256 57553.

Mixed-mode circuit simulation. Windows-based *IsSpice4* by Intusoft is an advance on Spice 3 in that it will now simulate both analogue and digital circuitry in the same .EXE. The event-driven *IsSpice4* algorithm supports 12-state digital data, and also real, integer and user-defined data, so that it is capable of simulating, say, dsp functions and sampled data filters in an analogue environment. Real data is handled by the event-driven simulator, so that the sampled-data filter can be simulated quicker than in an analogue model. Intusoft. Tel. (USA) 010 310 833-0710; fax 010 310 833-9658.

Graphics. Numerical Algorithms Group now distributes template Graphics Software packages, from two-dimensional presentations to high-level three-dimensional visualisation. *FIGraph*, a 2D/3D charting system for Fortran and C generates line, bar, pie and contour graphs on graphics terminals, IBM mainframes and the *X Window* system. *FIGARO+* ANSI/ISO PHIGS+ is for graphics in cad/cam on workstations, mainframes and *Windows NT*. TGS is a licensee of the *OpenGL* software interface for 3D applications, and *Open Inventor*, a C++ authoring system based on OpenGL. These products complement NAG's *IRIS Explorer*. Numerical Algorithms Group. Tel. 01865 511245; fax 01865 310139. ■

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