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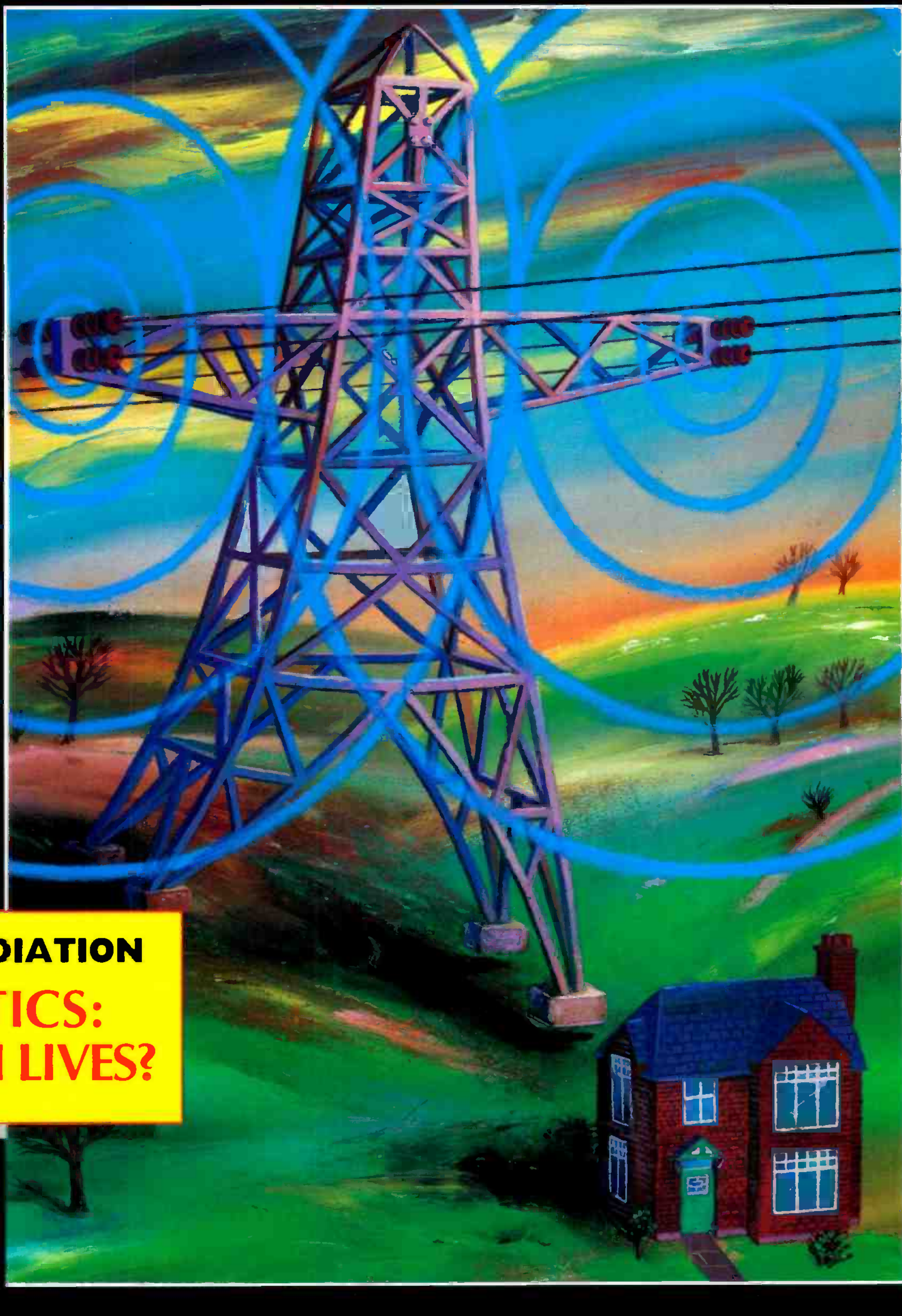

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In next month's issue. Very occasionally in electronics, a designer comes up with a totally original idea. This analogue filter system will implement as many poles as you like – and with true linear phase. With the main characteristics determined by just a single integrator, David Grundy's system is a chip manufacturer's dream.

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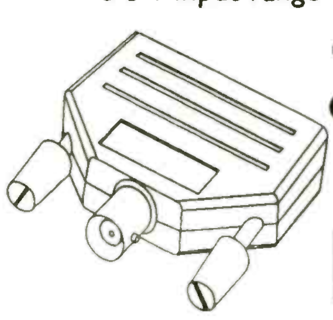

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CIRCLE NO. 107 ON REPLY CARD

The power behind killing fields

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In the Spring of 1974, a 47 year old experimental psychologist, Dr Nancy Wertheimer, began spending a couple of days each week driving through the residential suburbs of Denver, Colorado. Occasionally she would stop the car at one of the addresses on her list, get out and start making notes on what she found. Which was rarely, if ever, out of the ordinary.

The list from which Dr Wertheimer drew her itinerary comprised the home addresses at birth of every child in the Greater Denver area who had died of leukaemia between the years 1950 and 1969. Dr Wertheimer was an epidemiologist looking for environmental links to this predominantly childhood cancer.

Addresses on the list were unremarkable, not revealing clusters of anything in particular at first. After a while, she did begin to notice pole mounted transformers, the last leg of the local power distribution system taking the 7600 volts area feed down to 115 volts for the house feeds. The associated wiring straggle, so common to the roof lines of American towns and cities, also began to leap out at her along with the cylindrical transformers. More addresses brought more pole-mounted transformers and heavy service feed wiring stitched from pole to pole.

But there was nothing out of the ordinary here, she decided, having consulted her physicist friend, Ed Leeper. Leeper agreed that, while all houses carry AC mains, stray magnetic fields from the steel clad transformers dwindle to nothing after a few feet.

Being open minded people, the two decided to check stray magnetic power line fields for certain elimination. The physicist built the psychologist a simple inductive loop detector coupled to a sensitive amplifier. So armed, Dr Wertheimer revisited some of the addresses on the list. The device hummed loudly when located at the base of a transformer pole, just as expected. More surprisingly, it continued to hum loudly when the psychologist walked away from the pole, along the block, under the wires. The hum only reduced after several service drops to nearby houses.

Dr Wertheimer repeated the observation at other addresses from the childhood leukaemia list. She also noted on further examination that leukaemia excesses tended to occur in clusters along the wiring span between transformer poles and up to their third service drop: this corresponded to the highest current levels flowing in the line associated with the greatest low frequency AC line fields.

Such were the origins of suspicion that power lines can kill in more ways than by simple electrocution.

How significant is the risk... Indeed, is there any risk? It is as difficult to state positively that 'low frequency fields will kill you' any more than it is to say 'smoking will kill you' although there can be few people who now consider smoking to be without risk. I remain sceptical when the National Grid company defends its assertion that power line fields are harmless by referring to its own studies which show no excess disease; by any inspection they are deeply flawed. In one particular survey 95% of cases and controls lived under calculated fields of less than 10nT. Just 0.5% of cases and controls were habitually exposed to more than 100nT. There isn't anyone on either side of the argument who has observed pathogenic effects below 200nT; it is extremely unlikely that such a survey would show a disease excess.

According to National Grid's Dr David Jeffers, it is not company policy to route new power lines over population centres. It takes such avoidance "purely on aesthetic grounds" and not for any reasons of health risk. Publicly it remains unconvinced of potential danger. But those who live under them might well feel uncomfortable: US utilities follow a policy of prudent avoidance because they are not able to discount entirely a potential hazard.

It would therefore seem prudent to discount any reassuring noises from the National Grid until it publishes the results of flawless research, a study which looks for disease in high risk groups as designated and predicted by the Grid's opponents.

Frank Ogden

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REGULARS

UPDATE

EURO FIRST FOR UK VIDEOPHONE

British defence technology is being turned into Europe's first consumer priced videophone by one of the UK's largest defence electronics companies.

GEC Marconi is preparing to licence its low cost videophone technology to a number of international companies including BT and Deutsche Bundesposte Telecom. The aim is to create a European standard for consumer videophones which will work on existing networks.

Consumer electronics company Amstrad has already licensed the system for a £500 videophone to be launched in September. Marconi's priority is to create an international market for the product through further licensing deals.

The video phone will be one of a number of products in a new commercial electronics enterprise set up as part of Marconi Defence Systems. The company is looking for markets for electronics systems developed over the years for the Ministry of Defence.

A proprietary video and audio compression system squeezes a colour video picture into a 14.4kbit/s data stream which can be transmitted over standard analogue telephone lines. Voice compression techniques developed for military radio systems allow the system to use a 5.6kbit/s speech channel.

The videophone's camera uses Marconi's proprietary CCD array originally developed for night sights. The received picture will be displayed on a four inch liquid crystal screen provided by the Japanese company Seiko.

***Richard Wilson**, Electronics Weekly.*

16Mb static rams to enter production

In a development which could change the way that computer makers use semiconductor memory, Japanese chip maker Fujitsu says that it is now sampling a commercial 16Mbyte static ram. The device has the equivalent complexity of a 64Mbyte dram part.

Although the new chip incorporates some new design ideas, it essentially relies on conventional and tested production methods which should enable volume production by the end of the year. Quantum effects place a limit on the scalability of the charge storage element in a dynamic ram, but there is no such restriction on the scaling of the associated transistors. As memory chips become denser, there is a push in favour of using transistor bistables rather than capacitors as the storage element. So srams could eventually displace drams in main computer memory.

Non volatile gate arrays will reprogram

Field programmable gate arrays based on non-volatile, electrically erasable technology will soon make an impact on systems development. Produced by Lattice Semiconductor, these devices don't require a configuration program to be loaded into an integral static ram at boot-up as with the Xilinx devices. Neither do they rely on Actel one-time fuse technology.

Lattice has started out with a 6000 gate FPGA available in both 50 and 80MHz options. Devices cost roughly $1/MHz in volume. Other part densities are expected throughout the coming year.

With news from NASA that the ozone layer is disappearing over the Northern hemisphere as well as the South, there is now some evidence that the electronics industry is starting to remove CFCs from its production processes. This wave soldering machine operated by the Hughes Aircraft Company has excess flux removed from its workpieces by a water based process – a job traditionally done by halogenated hydrocarbons.

Ace chip makers agree standard risc

IDT, Siemens and Toshiba have announced that they can offer customers "multiple sourced standard derivative products" based on the Mips risc microprocessor architecture.

The threesome hope to stimulate the market for R3000 and R4000 microprocessors by giving customers the security of more than one source for derivations of Mips micros. The Mips camp, which has formed a consortium called ACE (advanced computing environment) received a blow recently when one of its members, DEC, surprised the world with a very high performance proprietary microprocessor architecture – the Alpha – which would be available for third party licence.

The new ACE agreement allows each of the participating companies to develop their own versions of the Mips micro based on common technical standards and parameters. It seems unlikely that DEC's Alpha chip will make much of an impact on the Mips powered workstation business in the short term; individual chips cost over $3000 in small qualities and consumes 30W of power, but for this, it can crunch 400mips at peak.

At technology's cutting edge – the chips really fly

High point of the microchip year is the International Solid State Circuits Conference, San Francisco. David Manners reports

Can silicon replicate the functions of the human nervous system and brain? How soon will micromechanisms – tiny machines capable of running inside an artery for instance – be part of normal hospital treatment, and how close are we to "nomadic" computing – the newest trend in portable, wireless communications between computers and people? 1992 ISSCC addressed all these questions – and more – providing some startling answers, and an insight into advances into the hottest areas of silicon technology.

Micro stars hit 1000 mips

As for the individual chips, the stars of the 1992 ISSCC were the microprocessors which are exploding in capability. Last year the top rated micro clocked at 100MHz. This year's fastest was 250MHz. And although these are ISSCC specs and therefore best case/best condition figures and usually some three years from the market – at least two of this year's ISSCC micros are already on the market or very close to it.

One such is DEC's 150MHz Alpha microprocessor now on sale for some $1500 a shot in quantities over 1000. That makes it 50 per cent faster than the fastest experimental part of last year's ISSCC.

The Alpha as described at the conference clocked at 200MHz which, with a two-instructions-per-clock capability, gives a peak performance of 400mips. The 64-bit processor, made in 0.75μm three level metal cmos, has an 8Kbyte data cache and two associated translation buffers, an 8Kbyte instruction cache, a four entry 32 byte per entry write buffer, a pipelined 64-bit integer execution unit with 32 entry register file and a pipelined floating point unit with an additional 32 registers. It uses 1.68 million transistors to do this. Alpha dissipates as much heat as the average large power transistor – 30W. The chip measures 16.8mm x 13.9mm and is presumably expensive and difficult to manufacture.

Viking advance

The other ISSCC microprocessor on the brink of commercial application was the SuperSparc or, as it was known during development at Texas Instruments – Viking. The ISSCC paper describing progress was given jointly by its architects, Sun Microsystems, and developers Texas Instruments.

SuperSparc/Viking is another 64-bit processor. It has a pipelined 32-bit integer unit, memory management unit, 20kbyte instruction cache, 16kbyte data cache, double precision floating point unit and bus interface unit.

The chip runs at 40MHz and, like Alpha, has a superscalar architecture allowing the execution of multiple instructions per clock cycle – up to three in this case. It is built in 0.8μm, three level metal, bi-cmos technology, and its 3.1 million transistors in the 15.98 x 15.98mm die dissipate 8W.

Alpha and SuperSparc are meant for today's market but another ISSCC microprocessor, a Hitachi 1000mips machine, is still some way from that. Made in a somewhat futuristic 0.3μm process it contains two superscalar processors each clocking at 250MHz and executing two instructions per clock so delivering 500mips each or one giga instruction in concert. The chip measures 8.1mm x 8mm.

Other microprocessor trends are: more on-chip cache memory – the largest 1992 ISSCC cache was 36kbytes compared to 1991's largest at 8kbytes; diversity of architectures with cisc processors and vector processors in among the predominant risc superscalars; a move to 3.3V for four out of the six ISSCC microprocessors – and the Hitachi chip showed, with some aplomb, that even bi-cmos can survive below 5V.

GaAs won't replace silicon

Chips used for communications showed no

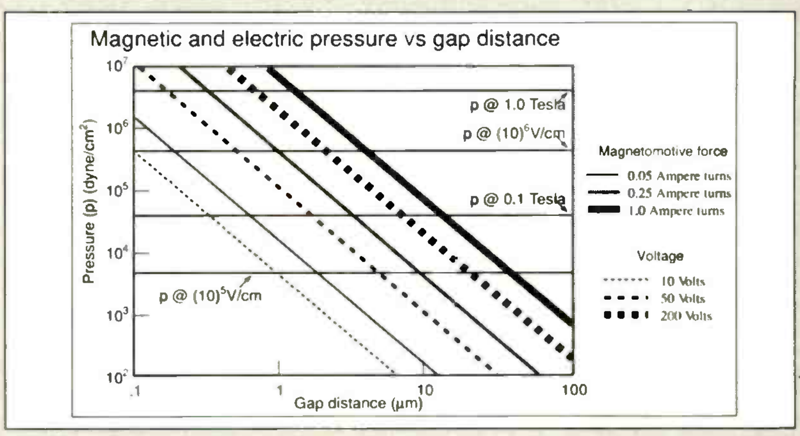
Machines in the blood

A dose of intravenous technology can be delivered with micromechanisms.

One projected use for these tiny machines is to run along inside veins cutting away cholesterol; others are in magnetic read heads or atomic force positioners.

Sensors and actuators are the main types of micromechanism said Professor Henry Guckel of the University of Wisconsin in an ISSCC Plenary Address, and the largest market is for the pressure transducer. A force sensor has been fabricated 200μm long, 40μm wide and 2μm thick with a fundamental resonance of 500kHz. He went on to describe how a bearing for a gear shaft of 25μm diameter can be fabricated requiring tolerances to ±25 angstroms.

The process fashions a mould using a thick photoresist technology with exposure by a highly collimated X-ray source and developing with perfect exposed-to-unexposed selectivity. The mould is then filled with electro-plated metal. According to Prof Guckel, quite complicated micro-machines can be fabricated by this method.



less dramatic advances than the computer chips. Whereas gallium arsenide is the usual technology for chips used in microwave and fibre-optic systems operating between 1GHz and 10GHz, the ISSCC showed the continual stretch in the capabilities of silicon is allowing it to replace GaAs in these areas.

Four out of six papers on gigahertz communication chips used silicon achieving transition frequencies of 40GHz compared to the 55GHz f_t of the two gallium arsenide devices described. A Fujitsu paper on a 1GHz low-power silicon prescaler demonstrated that bipolar speed-power products can be competitive with gallium arsenide: an NEC paper on a 40GHz bipolar silicon process with on-chip spiral inductors implemented a 28GHz dynamic frequency divider demonstrating the high-frequency capability of silicon.

A paper from UCLA on a 1.1GHz decision and clock recovery circuits in a one micron nmos process illustrated the performance achievable by scaled mos and indicated that sub-micron cmos technology may challenge bipolar silicon and gallium arsenide in the 3GHz region. Another NEC team used 40GHz bipolar silicon technology to implement a chip-set comprised of analogue circuits with a bandwidth exceeding 10GHz and digital circuits operating at 10Gb/s for a 10Gb/s direct-detection optical receiver.

The conclusion is that, in high frequency communications applications, the technology choice is widening rather than narrowing with cmos, nmos, bipolar and gallium arsenide all playing a part.

Limitations of dram

Memory chips, always the technology drivers, appeared to be running into a host of technological barriers according to ISSCC speakers. So large is the physical area of dram becoming that some at the conference predicted that dram die would be bigger than sram die by the end of the decade.

Already there is little to choose between the die size of srams and drams at the same level of process technology (though not density). Among ISSCC rams, 16Mbit sram die were measuring in the 220mm^2 region while the only 64Mbit dram (from NEC) measured 190mm^2; the capacitors in drams cannot hold the necessary charge if they are shrunk further.

ISSCC discussions expected dram sizes to sharply increase: Hisashi Shichijo of Texas Instruments predicted 256Mbit drams in the 270-300mm^2 region and 1Gbit drams in the 390-450mm^2 bracket. "drams are looking like microprocessors" he told a discussion session. The big issue with increasing die size is the cost of manufacturing. If dram sizes grow at this rate the technology may cease to be the cheapest way of delivering mass semiconductor storage.

One alternative is sram which could take over from dram as the most cost-effective memory chip if its size shrinks below the size of dram . However another possibility is flash eeprom technology with all its inherent advantages of non-volatility. The size of an NEC 16Mbit flash at ISSCC was 116mm^2 and the size of a Toshiba 4Mbit about half that. Unfortunately flash chips are slow – the NEC chip accessed in 58ns – and they can only be erased in sectors rather than bytes. However the potential of flash lies in its superior scalability over dram.

Russians make it small

Of course there is always the possibility of non-conventional technology taking over and one candidate, thought to be the most likely replacement technology, was discussed at the ISSCC – nanoelectronics. Here the Russians appear to be leading the field with a paper to the conference from NPO Mikroelektronica of Moscow describing how its researchers had fabricated quantum devices with feature sizes down to 3nm.

The Russian technique envisages atomic-scale memory devices with a storage capacity of 10 terabits per cm^2 of silicon. The Russian researchers had also used the technology to make nanometre fets with propagation delays down to 1ps. By using nanoelectronic techniques in recording functions, it was said that it was possible to achieve a recording density 1000 times greater than that achieved by laser recording.

An ISSCC discussion session on nanoelectronics concluded that a major breakthrough was needed if chip densities were to progress beyond the one gigabit stage and that nanoelectronics based on silicon was the most likely route to achieving it. Other routes such as cryogenic temperature devices, III-IV compounds and superconductors are "far from realistic application" said one speaker from IBM. ■

Verve in replacing nerves?

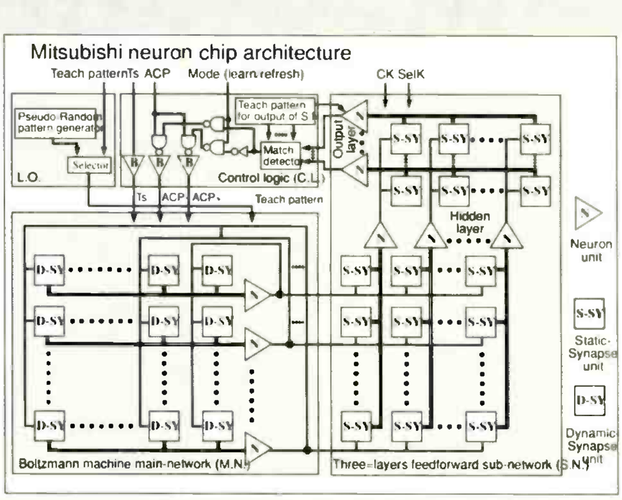
Far from realistic application, although emerging as a practical proposition for the first time, were neural networks. These systems replicate the architecture of human brain neurons in that they have vast amounts of i/o interconnecting with each other. At the conference it was said they are now becoming important because, after many years of being implemented only in software, advanced fabrication techniques were now making them a practical possibility in hardware. And it is only as hardware that they have any useful application in realtime computing.

Neural nets become useful in making self-learning machines which can update and amend their data bases automatically. Mitsubishi described an analogue self-learning neural network chip containing 400 neurons and 40,000 synapses. Synaptic density is obtained through dynamic storage on capacitors with refresh every 100ms. NTT described a digital neural network that achieves 8bn connections-per-second computational speed. It uses a local representation model for the neuron to reduce area and interconnections.

UCLA reported a 64-neuron self-organising analog neural net capable of unsupervised learning. The net concurrently evaluates several dot products of the state and weight vectors and passes the results through a winner-take-all circuit resulting in response times of 600ns.

Toshiba presented a two chip analogue chip-set used to implement back-propagation and learning. One chip has 24 by 24 neurons and does local weight control and update. Learning is realised on the second chip that has 24 neurons.

As well as the human brain, ISSCC considered the human nervous system with chips that can be implanted in the body to stimulate nerves which have lost their function eg deafness, or to monitor internal conditions such as arterial blood pressure and temperature. The Fraunhofer Institute of Duisburg, Germany described the latter



device – a cmos sensor and signal converter chip including signal conditioning and amplification using switch-capacitor techniques inserted into the body via a catheter.

The University of Michigan presented an active probe for stimulating the nervous system. The paper's authors pointed out that the normal use of electrodes implanted in tissue to stimulate the nervous system via small electric currents was inefficient as the wires tended to splay out. The university's chip is bonded onto four 15µm thick pointed shanks containing 16 electrodes which are sunk into the tissue. On the chip are digital data decode and channel select circuitry and 16 D-to-A converters.

REGULARS

RESEARCH NOTES

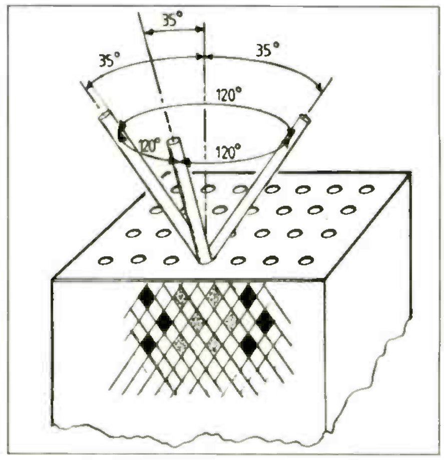
Semiconductor mirrors are frequency sensitive

Conductors, insulators and semiconductors derive their properties from the arrangement of electrons in the valence and conduction bands of the crystal lattice. Since this property derives from the relationship between the atomic structure of the lattice and the wave-like nature of the electron, it should be possible – in theory – to apply the same band-gap principles to the photonic waves that comprise light. In particular, it ought to be possible to create the photonic equivalents of insulators and semiconductors, and recent work suggests we might be closer than we think to such a break-through.

A photonic insulator would be a fascinating material, quite different from something that was merely opaque. In everyday terms it would function, over a certain range of wavelengths, as a near-perfect mirror. That in itself would be useful, but it is the devices created from it that would have extremely interesting properties.

For example if an excited atom were surrounded by a photonic insulator, it could not easily emit a photon and return to the ground state. This relatively lengthy excited state could be harnessed, for example, in the design of a whole new range of versatile and efficient lasers.

Judging by a recent paper by Yablonovitch *et al* at Bell Communications Research, NJ, and Brooklyn Polytechnic University (*Phys Rev Lett*, Vol 67 No 17), such photonic insulators may well soon be with us. Following calculations on the vector nature of photon fields, Yablonovitch's team have concluded that photonic insulators are not only a theoretical



Each hole is drilled through a mask at angle of around 35° producing a three dimensional structure.

Radio pulses point to new planets

One of the more intriguing scientific debates over the last few months concerns the interpretation of radio pulses from the depths of space, or to be more precise, from pulsars PSR1829-10 and PSR1257+12. Excitement has been high because scientists are scenting a definitive answer to the question of whether planets exist outside our solar system.

Pulsars are rapidly rotating neutron stars, believed to be the remnants of massive supernova explosions in far distant galaxies.

As the celestial cinders rotate they emit beams of radio energy, sweeping the heavens like giant lighthouses rotating at anything up to 1000rpm. So regular is the pulse repetition rate that, when pulsars were first discovered, astronomers thought they had found clear evidence for existence of extra-terrestrial life rather than emissions from this most accurate of all natural oscillators.

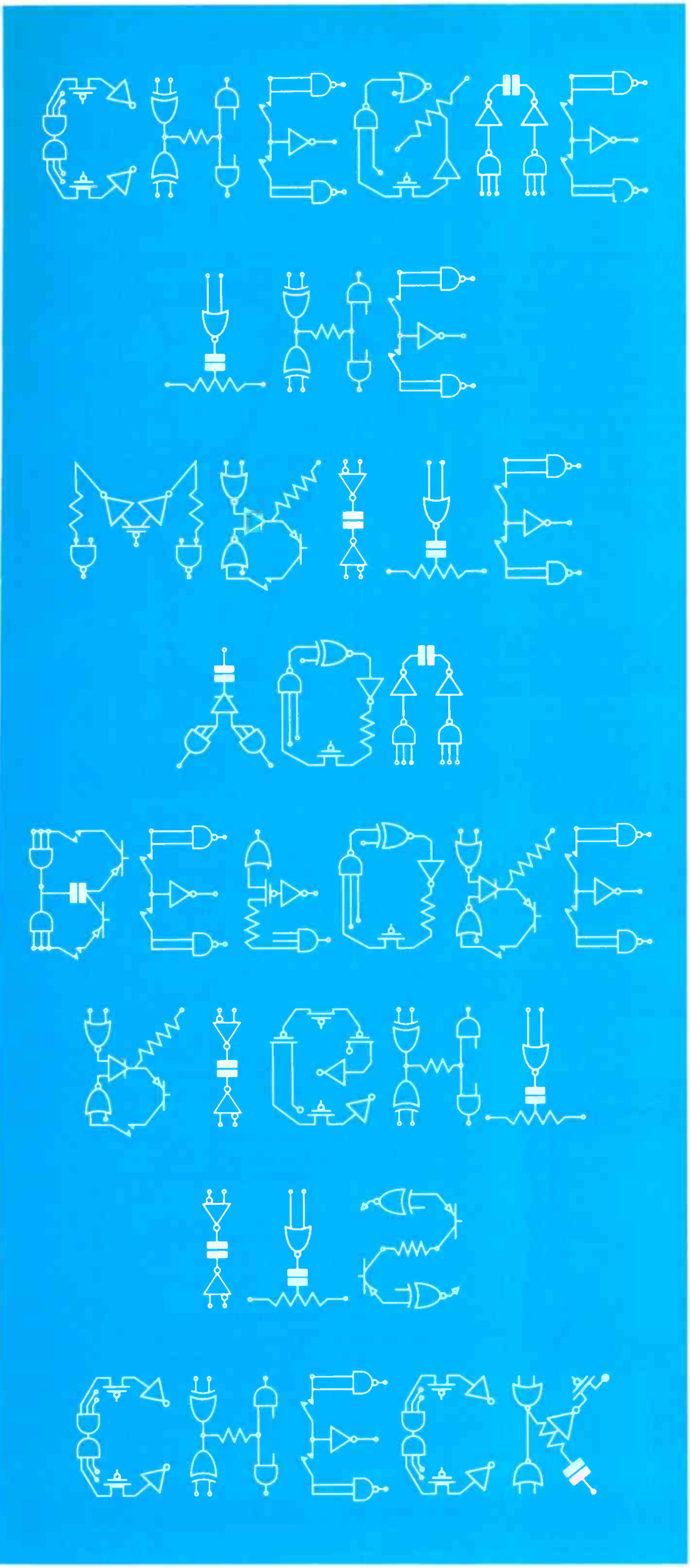
Last year a team from Jodrell Bank caused great excitement when it announced the finding of a periodic variation in the otherwise-very stable behaviour of PSR1829-10. They claimed the irregularity could only be explained on the basis of a Doppler shift due to the movement back-and-forth of the pulsar in relation to the Earth. That in turn had to imply the existence of some gravitational attraction from a co-rotating body – presumably a planet. Such a planet, a most unlikely companion to a neutron star, was calculated from the data to be at least ten Earth masses and to be revolving around the pulsar with a period of six months.

Another sensation came in January when the somewhat contrite team, led by Professor Andrew Lyne, wrote again to *Nature*, (Vol 355, No 6357) saying it was all a big mistake. Their computer program, in calculating the variations in the pulsar-to-earth distance, had failed toallow for the Earth's own motion around the Sun. When this was taken into account, the periodic variation disappeared, and with it the evidence for the planet.

But just as Lyne *et al* were licking their wounds, astronomers were beginning to assess the significance of another report (*Nature*, Vol 355 No 6356) from a group using the Arecibo radio telescope in Puerto Rico. Their calculations on signals from a different pulsar, PSR1257+12, revealed not one, but *two* separate periodicities of 98 and 67 days. This time there was no obvious mathematical error, nor was it possible to explain away the seemingly obvious conclusion.

So does this mean that there definitely are planets outside our Solar System? The answer, to a reasonably high degree of certainty, could soon emerge from a test suggested recently by a group at the Cornell Center for Radiophysics and Space Research in New York (*Nature* ,Vol 355 No 6358). They point out that if PSR1257+12 has two planets, such planets ought to interact with each other, producing a characteristic gravitational signature superimposed on the two fundamental periodicities. The Cornell group reckon that the observation of such signatures would amount to virtual proof of the planets' existence.

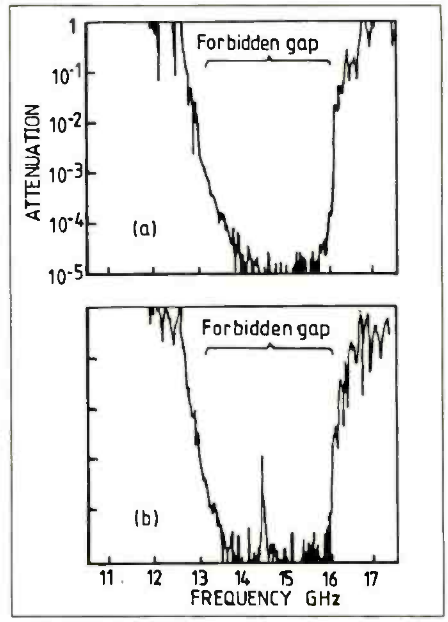
Sadly, even if such planets do exist, their chances of being inhabited are nil. Radiation from the nearby pulsar would ensure the immediate destruction of any emergent life form.



possibility, but should be capable of practical realisation and to that end they have devised a dielectric structure which has forbidden bands for photons of particular wavelengths. In effect that makes the structure an insulator at those wavelengths.

To confirm the theory the team constructed a scaled-up model using a dielectric substrate, drilled out with millimetre-sized mechanical tools, and subjected it to microwave radiation. The group then created a cavity in the material – analogous to a dopant atom in an electronic semiconductor – creating a resonant tunnelling effect so allowing radiation at a specific wavelength to escape through the otherwise-insulating material. The Q of the acceptance peak was found to be about 1000: high, but considerably less than the theoretical maximum.

Having established that photonic insulators work at microwave frequencies, the next step will be fabricate them at optical wavelengths. Horrendously difficult as this might seem, Yablonovitch believes it should be possible using reactive ion etching. The result should be a fascinating material, and not just for its potential applications in lasers, optical filters and so on. A photonic insulator or semiconductor with its near-perfect mirror qualities and high refractive index should have a sparkle superior to that of diamond. Add to that the frequency-selective nature of the reflectivity and the extra colour filtration made possible by doping and the cosmetic possibilities would seem unparalleled.



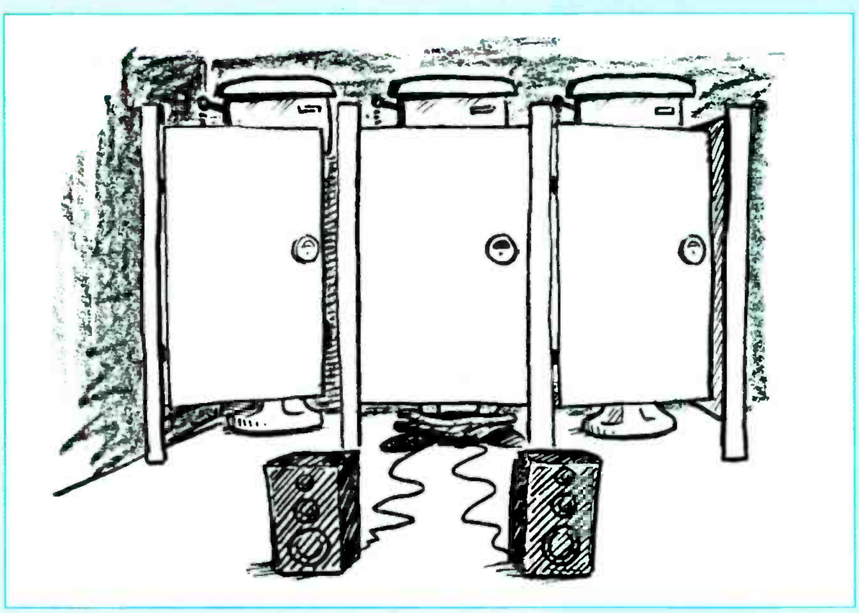
Attenuation at microwave frequency through a defect-free photonic crystal model with the forbidden gap at 13-16GHz, and through a photonic crystal with a single acceptor in the centre.

Are you sitting comfortably...?

Modern technology can have both good and bad effects on the human body. That fact is generally unquestioned. But by far the most interesting examples of how the ubiquitous microchip can kill or cure come in two recent issues of different medical journals

Dr Michael Spencer, in a letter to the doctors' magazine *GP* (24.01.92) admits to great difficulties in using his cordless phone in the loo. He confesses he finds it quite stressful counselling his clients while worrying that any uncontrolled noise may lead his patient to guess that he is not sitting at his desk surrounded by learned tomes. As a result he is worried about becoming "anal retentive" (constipated).

But it seems he is worrying unduly. An editorial in *The Lancet* (25.01.92) explains that indiscretions of this sort are normally inhibited by a spinal reflex (negative feedback to the likes of us). The reflex ensures that any physical activity – even talking – automatically tightens up the muscles of the pelvic floor, preventing...well modesty forbids.



But for those that need a little help in this delicate area, electronics yet again comes to the rescue. *The Lancet* reports successful research into the use of biofeedback. A team at Edinburgh (GUT 1991; 32: 1175) has designed an electrode that can be connected to an electromyograph display or to a loudspeaker, giving an instant indication to a patient of how tense the muscles are in the nether regions. While hard-wired in this way, patients were encouraged consciously to reduce their muscle tenseness while straining every day for at least a fortnight. One heroic soul kept it up for six weeks!

Results of this DIY biofeedback are described by *The Lancet* as "useful and cost-effective". Perhaps some enterprising company will patent a fully integrated home loo, specially designed to bring express relief...and all for the cost of a plug-in probe and a few chips.

There's only one snag. You'll have to forget the cordless phone.

Scientists warm to Josephson junctions

How much faster can computers go? Many researchers are beginning to flag possible limits based on the problems of chip fabrication or other fundamental barriers of quantum mechanics.

But, according to a number of papers recently published, the medium-term need for faster circuitry may well be satisfied neither by gallium arsenide, nor by some electro-biotic slime – sorry, Langmuir-Blodgett film, but by our old friend the Josephson junction. What has helped re-ignite interest in Josephson junctions is development of much simpler cryostat cooling technology by Fujitsu Laboratories (*Science*, Vol 255:301).

Josephson junctions handsomely beat every other logic family in the two most important factors, namely speed and power consumption. Faster switching operations generally mean a greater power dissipation and if existing switching technologies were made faster or denser by, say, another order of magnitude, designers would be up against the overwhelming barrier of heat.

Josephson junctions switch in picoseconds and operate with milliamps and millivolts; their only real snag is the need for cryogenics. A practical Josephson junction consists of a pair of niobium metal electrodes separated by a thin insulating barrier and needs to be immersed in liquid helium at 4.2K. Not ideal for your average lap-top!

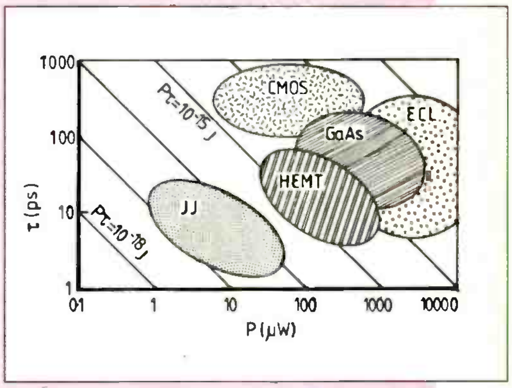
The Josephson effect is produced by superconducting electrons which, in the absence of an applied voltage, can tunnel through the insulating barrier creating a "supercurrent". If this supercurrent exceeds about 1mA, the junction voltage suddenly jumps from zero to about 3mV, a sort of bistable latch action that can be exploited for high-speed digital switching.

Now Hitachi, NEC, ETL and Fujitsu have fabricated experimental devices that will perform the same functions as a standard silicon processor but with clock speed of around 800MHz and a power consumption of a mere 5mW.

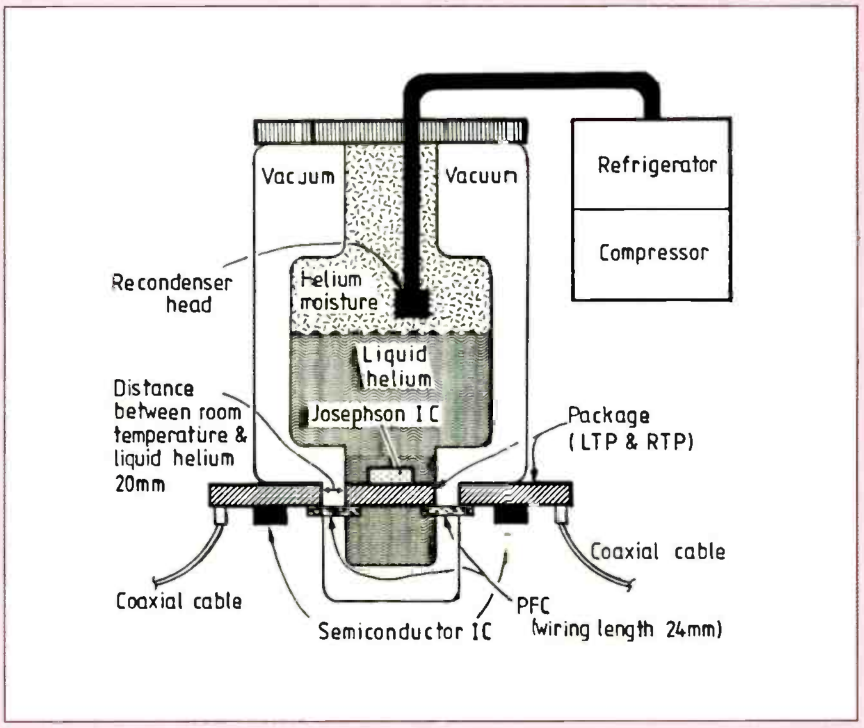
The real problems in developing Josephson circuitry lie not in the low logic levels which can be overcome electronically, but in the practical problems associated with the extremely low temperatures. You can't just stick one of these devices inside a flask of bubbling helium and solder it via a standard PCB to silicon cmos.

But S Kotani of Fujitsu Laboratories has developed, in conjunction with other Japanese companies, a novel cryostat in which the arrangement of the Josephson IC and the other components of the circuit are cleverly integrated so as to minimise connector lengths and also to minimise loss of "cold." Special thin polyimide film cable connections reduce the heat input to little more than 1W. The system is not only thermally efficient, but operates successfully as a fully integrated microprocessor clocking at a frequency of 1.1GHz.

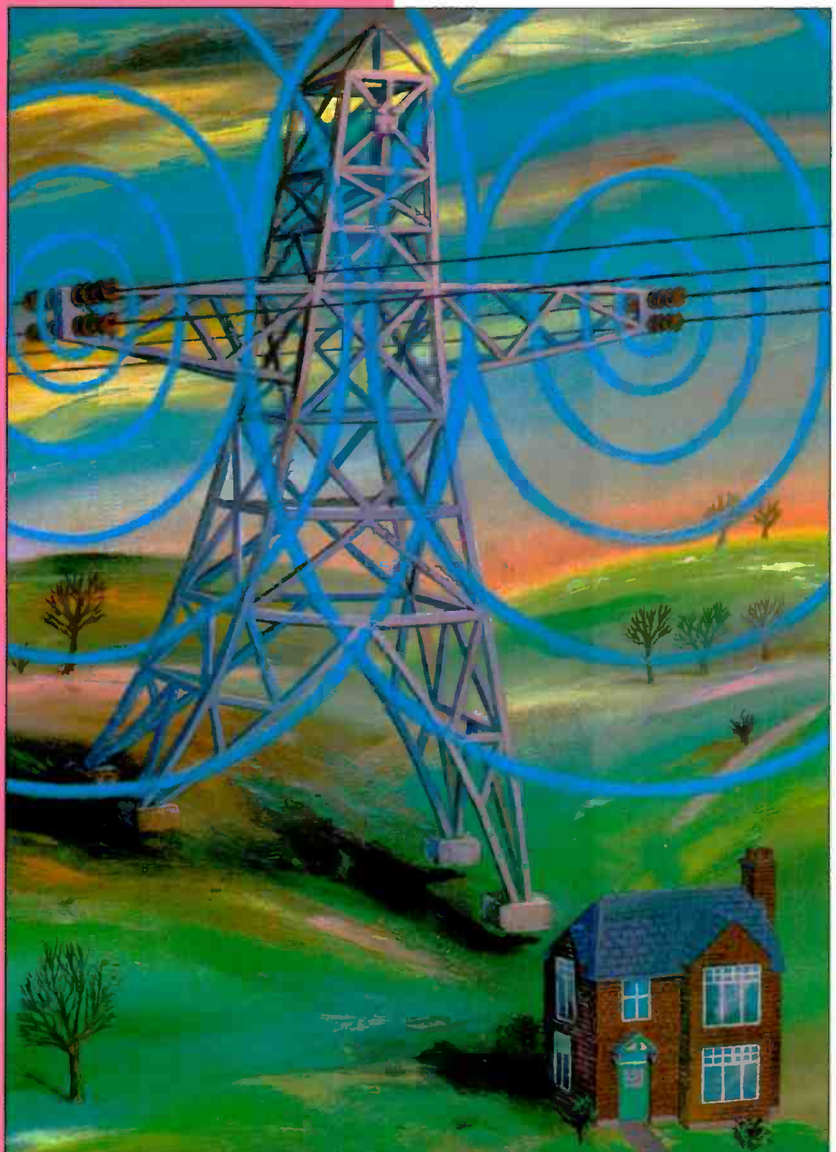
Liquid helium, it appears, will remain essential for systems of this kind, since high temperature superconducting Josephson junctions are far too prodigal in their use of power. Nevertheless hybrid systems of the sort developed by Fujitsu would seem to have a useful role to play in keeping up the relentless pursuit of computing speeds, especially in the interim before radically new technologies are developed. ■



Josephson junctions deliver impressive combinations of switching time and power consumption. The diagram compares Josephson junction, HEMT high electron-mobility transistor, GaAs fet, Si ECL and cmos logic.



Signals transmitted through a polyimide flexible cable penetrate the vacuum wall between the low-temperature and room-temperature packages.



POWER POLITICS: playing with children's lives?

How dangerous are the power lines around our homes? Alasdair Philips reports new evidence which shows that for some families the question may be one of life or death.

Public concern over the risk to health of electromagnetic fields is likely to become "one of the important environmental public health issues of the nineties", according to the National Regulatory Research Institute in the USA[1].

But in the UK, developers are still being allowed to build next to and underneath electrical power-lines. Some Local Authorities are now starting to introduce restrictions, but these are really as a result of the pressure of concerned individuals and so lack the consistency and backing that Central Government advice would provide.

Concern has been triggered by evidence pointing the finger at long-term chronic exposure to low-level power frequency magnetic fields as a danger to health.

Unfortunately scientific acceptance of the ELF (extremely low frequency) interaction has been hampered by lack of a credible mechanism. This has tended to mean that many scientists have dismissed the evidence out of hand, and has made the methodology and planning of experimental work difficult. Even the way that biological and physical sciences are separately taught has had a hindering effect on understanding.

But now several likely resonance mechanisms have been put forward to explain how low levels of low-frequency electric and magnetic fields could affect biological processes.

Biological interactions

Just as a radio receiver can be used to detect and decode a specific coherent signal at an intensity below the overall background noise, there is considerable evidence that coherent time-varying fields which coincide with natural ion magnetic resonances can have biological effects at very low levels – for example effects on the pineal gland.

The pineal gland is a tiny, cone-shaped structure within the brain, whose main function is secretion of the hormones melatonin and serotonin.

Melatonin plays a major part in controlling human circadian rhythms: its production peaks at night and is suppressed by bright lights and low ELF magnetic fields. It also affects levels of other important hormones – including serotonin. Segal *et al* (1989) found considerably lowered levels of serotonin and dopamine in mercat monkeys chronically exposed to low levels (up to 10μT) of ELF magnetic fields.

Prudent avoidance–the UK reaction

The term "Prudent Avoidance" was published in a report by US Office of Technology Assessment for the US Congress. It covers not only transmission lines and distribution lines, but house wiring and other electrical devices, and implies taking prudent action to minimise fields at modest cost.

Professor Ross Adey explains[8] that prudent avoidance as a part of public policy was "a very necessary way to tackle the problem". It implied not putting power lines over people's heads, restricting the access of builders to vacant land next to high-voltage power lines, and specific avoidance of school sites as a matter of public responsibility.

"British authorities' attitudes absolutely avoid confronting the evidence as it now exists," he says, although the NRPB (National Radiological Protection Board) is doing work of the "highest merit in its scientific content, and it indicates that there is reason to be concerned."

Whose problem is health?

The electricity supply industry was restructured with effect from March 31 1990. Prior to that the Central Electricity Generating Board (CEGB) produced the bulk of electricity in England and Wales and the 12 area boards purchased this and distributed it to their customers.

Fossil fuelled power stations were divided between two new companies, National Power PLC and PowerGen PLC, and the nuclear stations given to Nuclear Electric.

The network of 400kV and 275kV transmission lines, and two pumped storage systems, were given to another new enterprise, the National Grid company.

Distribution lines of 132kV and below were given to the 12 regional electricity companies, who, through a holding company, also own the National Grid Company.

As electricity generation is now in the open market place there has been a flood of planning applications for building new power stations from commercial enterprises – and this will have a knock on effect for power distribution.

In the case of the new 1725MW gas-fired power station on Teesside, building permission was given directly by the Secretary of State for Energy in December 1990. The National Grid Company now has the task of connecting it into the Grid and has faced thousands of objections to its new power lines. As a result a Public Inquiry has been granted, starting on May 19, which will include consideration of possible adverse health effects.

It is very costly to minimise magnetic fields from lines so hundreds of millions of pounds are at stake. For the electricity industry, spending vast amounts of money to have the health effects case dismissed is still a cheaper option.

In contrast, contesting Councils cannot reclaim money from the government and will have to find it from their own coffers.

Both chemicals are known to affect mood changes and are linked with depressive illness: a study of clinically depressed people published in *The Lancet* in 1987 showed lowered levels of serotonin in every case.

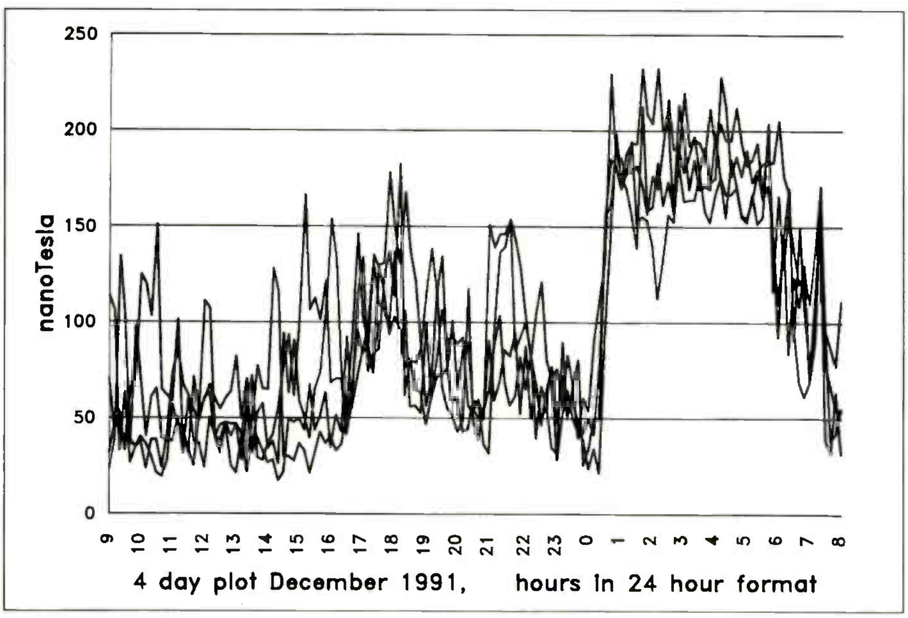
Serotonin also has an important function in many bodily processes including effectiveness of the immune system.

Mechanisms for concern

One probable mechanism, based on an ion cyclotron resonance model, is gaining considerable support[3, 4]. It relates the motion of electrically charged particles to the magnitude of the steady magnetic field surrounding them.

The importance of this for humans is that blood contains ions in the critical mass range for resonance in the Earth's geomagnetic field and an alternating field between 1 and 500Hz. If a time-varying magnetic field parallel to the steady field (or an electric field orthogonal to it) is applied near the resonance frequency, then energy is transferred to the charged particle. The theory seems to be well corroborated by experimental results on calcium, lithium, sodium, and potassium ions, all of which play key roles in living systems[5].

50Hz magnetic field in a domestic bedroom. RMS magnetic flux, from 415V + neutral supply at 8m distance. As most people are asleep and relatively motionless during the night, these coherent magnetic fields provide the optimum conditions for ion-resonance interaction to occur.



Unfortunately use of the word "cyclotron" in the original paper has caused some problems, as what is described is not actually classical cyclotron resonance. More recently, similar models have been proposed by Lednev[21] and Dr John Male of the National Grid Research and Development Centre[22]. Both propose that a magnetic field, amplitude modulated at the cyclotron resonance frequency, can affect ion vibrational patterns, though not through actual cyclotron resonance.

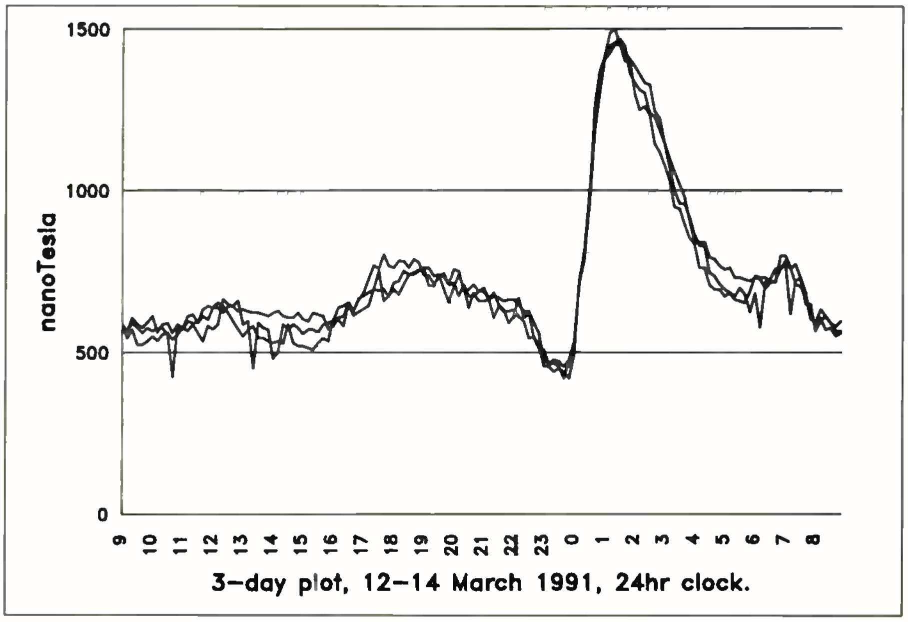
Drs Jafary-Asl and Smith of Salford University have a quantum magnetic resonance model which, for the Earth's geomagnetic field, relates the precession of molecules to a permanent magnetic moment[6, 7].

Both mechanisms may be involved with resonances of hydroxyl (OH) and hydronium (OH_3) which resonate in the 40 to 65 Hz band in the geomagnetic field[4].

Another mechanism now under investigation shows how changed spin states in haemoglobin molecules can produce a significant net magnetic moment. This attracts and traps the lymphocyte cells, greatly impairing the immune system response, and may well explain many cases of lymphatic leukaemia and other diseases[19].

Debating point

Professor Ross Adey of the Loma Linda University in Florida, a respected researcher in this field for over 30 years, recently stated on Radio Scotland[8] that from the scientific point of view, there are four major areas about



Main bedroom, Glebe Cottage, RMS magnetic flux from 33kV 50Hz power line. The field in this bedroom hardly falls below twice the level at which a three-fold increase in childhood cancer has been repeatedly demonstrated

which "there can be very little doubt as to the significance of the findings".

He listed the areas as: effects on the immune system, with a reduction in the ability of the circulating white blood cells to kill tumour cells; effects on foetal development; control and regulation of cell growth – including tumour formation, and the effects on very powerful hormonal mechanisms in the central nervous system and the brain, which in turn have connections to cancer and cancer-related problems.

Professor Adey pointed out that work is being conducted in many laboratories worldwide, so that the old argument that this research describes uncorroborated experiments is no longer true. He says the most significant finding is that effects appear to be strongly synergic with chemical factors – although many of the effects could be seen from the fields alone.

Some known chemical promoters are significantly enhanced in their action by the presence of power-line frequency magnetic fields. Adey suggested that this may be a pointer to the clustering of cases where there may be a common chemical factor as well as the magnetic fields.

Looking at other possible effects, privately funded UK research work has also associated power-line magnetic fields with clinical depression and suicide[9], and headaches and depression[10] though this does not seem to have been followed up by the authorities.

Different levels

The threshold at 60Hz above which published studies have repeatedly shown a three-fold increase in childhood cancers is 250nT. In 1990, a statement submitted to a US Congress committee by Dr Robert Becker, one of the pioneer researchers in this field, proposed that all new construction, transmission lines, substations and distribution lines be required to produce "no more than 100nT" in any adjacent dwelling, school or public building. The proposal suggested the utility companies should produce a plan to bring existing installations into compliance by the year 2000.

Becker's view is probably fairly extreme, but at present there are no regulatory levels for the UK.

The current *NRPB Guidance GS11(1989)* – for electric shock and heating effects only – puts the level 20,000 times higher at 2mT, with the electric field maximum set at 12.3kV/m.

In its report R239 (July 1991) *Biological Effects of Exposure to Non-Ionising Electromagnetic Fields and Radiation*, the NRPB acknowledges that: "there are several possible areas of biological interaction which have health implications and about which our knowledge is limited."

"Aesthetics" determine power line policy

As part of their continuing studies perhaps the NRPB should investigate specific clusters already known about.

A good example is the case of the village of Dalmally in Perthshire[11]. Here at least eight people have died of cancer and three from motor-neurone disease in the last five years. All lived in houses lying near to the 275kV transmission line connecting Cruachan – the UK's largest pumped-storage hydro-electric scheme – to the Grid, and will probably have

Statutory limitations

Sweden's limit for ELF magnetic fields is 250nT at 0.5m in front of a VDU. New York has just set the limit for VDUs in schools to 200nT at 0.3m.

In the UK we can still sleep in 50Hz ELF fields of 10,000 times these level and Dr Dennis (NRPB) is on record as saying that it is likely that it will be many years, if ever, that statutory limits on exposure will be applied in the UK[14].

In the meantime the NRPB and the Secretary of State for Health should come up with clear interim guidelines for new power-lines and building work near existing lines. The guidelines should be specified in magnetic and electric field strength levels and not just distance – though statutory rights-of-way as are used in the US would be a useful start.

Local authority planners surely have the right to be given this important information.

What price research

What is the electricity companies' approach to research? In fact both National Power and PowerGen are closing research facilities, with the loss of several hundred jobs. Since the retirement of Peter Chesters last year, nobody on the main board of National Power has full time responsibility for environmental issues. According to reports, Chesters' entire department is now being disbanded. (*Electrical Review*, Vol 224, No 23, Dec 1991).

National Grid Company's research projects associated with electric and magnetic fields are:

Personal exposures study: 200 volunteers from the supply industry have been wearing exposure meters. Full analysis will be available by the end of 1992.

Magnetic environment research vehicles: Three vans have been fitted out to enable them to measure the magnetic environment. Initially they measured magnetic fields inside volunteers' homes. However there seems to have been only a small output of data considering *three* vans active over *two and a half* years. Results are said to show average background levels around 50-60nT in houses.

Biological studies: Basic research is being supported in universities. Contracts were granted in 1990 for six separate studies.

Epidemiological studies: The UK Coordinating Committee for Cancer Research is planning a new large scale study of the incidence of childhood cancer in the UK. Although the UKCCR was not intending to include magnetic fields in this study, it now plans to do so.

Ross Adey

Professor Ross Adey was previously Director of the Nasa Space Biology Institute, Professor of Anatomy & Physiology UCLA, Founder UCLA Brain Research Institute, and a member of the National Academy of Science Committee studying the bio-physical effects of the Seafarer ELF Communication System.

He has been in the forefront of research in this field for about 30 years and has played a leading role in development of the "whispering cell" theory by which intercellular interactions are controlled.

Over the last 20 years much of the pioneering research in cellular communication and control field has been done at Ross Adey's laboratories.

With regard to cancer, Adey now states that the chemicals which seem to be the problem are those which work on cell membranes, and not on cell nuclei.

"The concept of cancer arising from damage to DNA is now being supplanted by a different view of what the tumour problem really is: that it is a problem at cell membranes where the electromagnetic fields and these chemicals act together synergistically".

been subject to quite high fields, especially at night when they were resting.

It should be quite easy to compare this small community with a similar one without power-lines but with all aspects of health included. Unfortunately that kind of study does not seem to get funding.

David Jeffers, National Grid's Spokesman on these matters recently put the Grid's position quite clearly, on Radio Scotland: "Our view is that there is certainly not an established link between any of the electric and magnetic fields and health effects".

There maybe a case on "prudent" grounds for keeping lines away from houses... "But we have done so on aesthetic grounds". National Grid

National Grid was looking into the issue, said Jeffers, but that was only because it recognised that its consumers were concerned about the issue, and so it felt it should mount an effort to put their fears to rest. There may be a case on "prudent" grounds for keeping lines away from houses and National Grid had always followed that policy: "But we have done so on aesthetic grounds," he said.

As a footnote, Scottish Power says that planning is a function of local government and it sees no justification on health grounds for changing present practice with regard to the routing of power-lines.

Jeffers pointed out that the NRPB was charged with advising the Government on exposures and "there was a very clear Statement from the Secretary of State for Health last year that the NRPB has that responsibility[12]."

Replying to articles in *EW + WW* ("The Killing Fields"[17], Feb 1990), Dr John Dennis, then an Assistant Director of the NRPB said it would be "premature" to specify limits based on the possibilities for long term effects on health. Nevertheless: "there is obviously a case for avoiding unnecessary exposure and reducing exposure levels where this can be done easily" (*EW + WW*, April 1990, Letters).

Two years later the electricity companies still have not accepted even that level of concern, though they are now private companies whose combined profits last year were £1355 million. A statement by National Grid recently said that "Future research will be assessed on a commercial basis"[13].

Mistaken or misleading?

The CEGB sponsored UK studies assume a very low background 50Hz magnetic field level of 10nT, and only take into account calculated levels due to overhead electrical lines at the birth addresses.

From my measurements I believe that a significant proportion of the UK population lives in 50Hz fields approaching or exceeding 100nT. People living near high power feeds could be subject to fields up to 25μT (25,000nT) – for example in older city areas, on the lower floors in blocks of flats, and in rural areas where power is still distributed along streets on poles.

My front bedroom is about 7m from a 415V three-phase and neutral pole supply to which only 14 houses are connected.

The figure shows the field levels experienced during one week, with the levels from four days plotted on the same graph. The resulting patterns were very similar from day to day, with noticeable peaks at 0700, 1800 and 2100hrs, and a very pronounced peak when the "Economy-7" off-peak electric storage heater demand starts just after midnight. As most people are asleep and relatively motionless during the night, these coherent magnetic fields provide the optimum conditions for ion-resonance interaction to occur. Concern about field levels caused me to get rid of my off-peak electric storage heaters a couple of years ago. Completely switching off my house supply makes negligible change in the magnetic field levels other than very close to electrical appliances.

The second figure shows the field in a bedroom of a house near to a 33kV distribution line. Again the close agreement from day to day is interesting, as is the very pronounced peak between 0100 and 0400hrs. The field in this bedroom hardly falls below twice the level at which a three-fold increase in childhood cancer has been repeatedly demonstrated, and peaks at six times that level. These graphs only show total magnitudes.

To investigate the ion-resonance theories we must know the magnetic vector details, as well as those of the local geomagnetic field. No epidemiological work published to date has used this essential extra data. (More details of the requirements are given in the section dealing with AC magnetic field measurement.)

In the introduction to its first study[15] the CEGB quotes the work by Wertheimer and Leeper, and Tomenius – both of which implicate fields above 250nT as being associated with a threefold increased risk of childhood cancer. Despite this CEGB has only chosen about 1.5% of cases and controls that were born in calculated fields of above 100nT for its own study. I have challenged the purpose and value of this study[17].

Recently[20] Dr Nancy Wertheimer, the epidemiologist working in this field since 1974, has made a number of comments about survey work. She suggests that because most studies have not checked that their control groups are actually experiencing low exposure, they are like those that compare people who smoke 2.5 packs of cigarettes a day with those who only smoke two packs. She goes on to say that evidence suggests that the critical period is between one and two years before diagnosis. Just using the birth address, as is done in the UK study, will therefore lead to invalid data.

An extended and re-analysed[16] study allows that over 95% of cases and controls lived in calculated fields of less than 10nT, and only 0.5% in fields over 100nT and admits that the study stood "no realistic chance of detecting any raised relative risk associated with a field of more than 100nT". A more recent adult study also suffers a similar problem[18].

In January 1992, Dr David Jeffers, of National Grid, publicly stated:

"We have a number of strong players in this programme. We have had what you call epidemiological studies, carried out in Yorkshire, of how childhood cancer correlates with the fields in the houses and how close the houses are to power lines, and the answer was no, we did not find a correlation in Yorkshire. A similar study on adults and leukaemia, carried out in Yorkshire and Lancashire, also did not find an association[12]."

I hope Dr Jeffers will explain to all of us his justification for describing these studies as "strong players", other than as part of "an effort to put their fears to rest". ■

Measuring magnetic fields in your own home

An AC magnetometer suitable for serious epidemiological work can be constructed. Alasdair Philips shows how.

A magnetic field varying with time is always produced by an alternating current. Magnetic fields from man-made sources generally have higher intensities than background varying fields. Flux densities of up to about 40μT are produced at 0.3m from many domestic appliances and also directly under large power-lines.

Due to the low frequencies involved, the electric and magnetic fields should be analysed separately and not as true electromagnetic radiation, and equations may be derived which predict the fields at any point in space at some distance from wires[16].

What the equations show – verified in practice – is that the field due to a three-phase power-line is a magnetic vector, rotating in a plane normal to the direction of the line, the horizontal and vertical components of which are usually of different magnitude. The vector describes an ellipse whose semi-major axis represents the magnitude and direction of the maximum value of the field, and whose semi-minor axis represents it a quarter of a cycle later. This may been seen in practice by connecting the amplified outputs of two orthogonal magnetic sensor coils to the X and Y inputs of an oscilloscope.

Measurements on power-lines show the existence of odd-harmonic currents of significant magnitude – especially with local distribution lines. As the distance of the measurement point to the line increases, the fraction of the field ascribed to harmonics tends to rise. The effect is due to the fact that while the fundamental fields tend to cancel, provided the currents are nearly balanced and power-factor is near-unity, the unbalanced harmonic current fields do not.

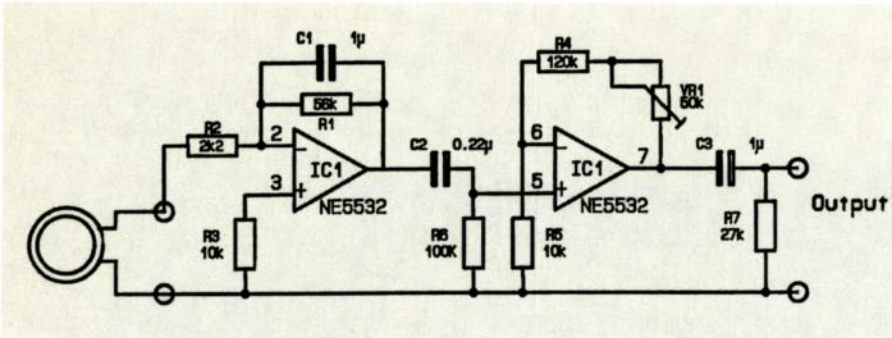
Quantifying the risk.

AC magnetic field sensing

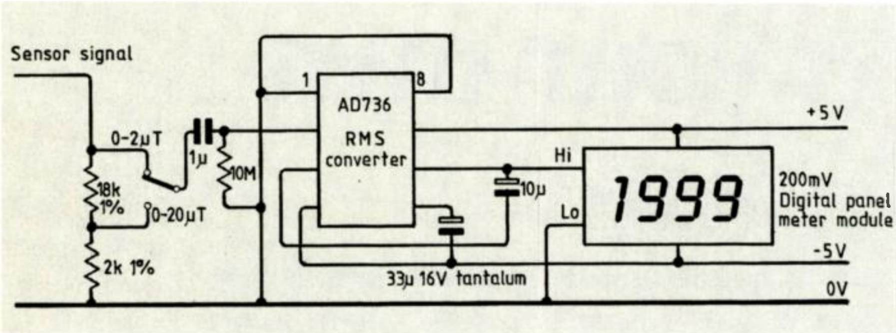
For our purposes we are interested in measuring power-line frequency fields with a spatial resolution appropriate to the size of human organs, and with a useful range up to about 50μT with a resolution of 1nT.

The standard generally accepted as defining the measurement requirements is the ansi/IEEE standard 644-1987, *IEE Standard Procedures for Measurement of Power Frequency Electric and Magnetic Fields from AC Power Lines*. Recommendations are that an air-cored magnetic search coil of diameter about 100mm is used: details of calibration procedures are given in the standard.

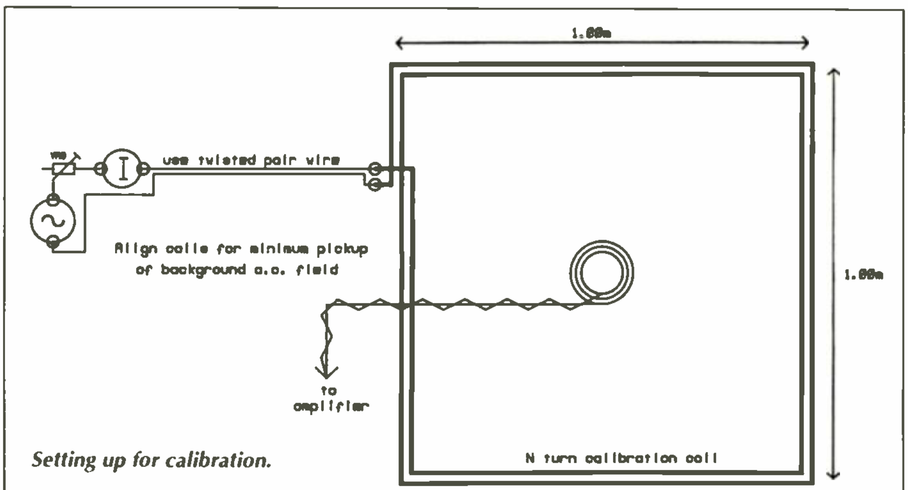
Using a coil as the sensor element has the advantage that the measurement signal has a direct and simple relationship with the quantity to be measured. But a single coil only senses the field normal to the plane it lies in, so to measure the total AC field, three orthogonal coils must be used and the total field vector calculated. The approach provides the spatial information needed to relate to the local geomagnetic field. For a coil sensor, the electromotive force is equal to the rate of change of flux with which it is linked. If the magnetic flux density is given by B where $B=B_0\sin(\omega t)$, then the magnetic flux ϕ through the enclosed



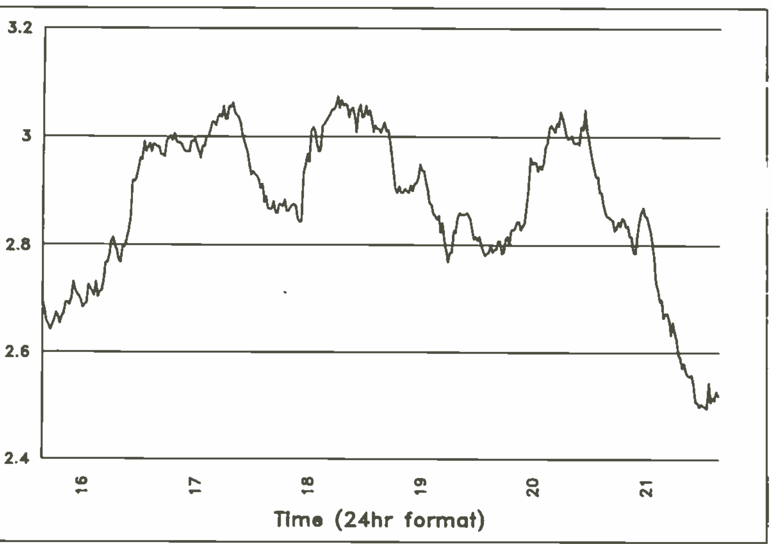
Converter circuit based on the AD736 RMS/DC converter circuit..Output is 100mV/µT.



Use an amplifier with little low frequency noise



Setting up for calibration.



Example logged magnetic field from a 440kV line. RMS magnetic flux in µT at 30m.

area A is given by $\phi=AB$.

For a loop of N turns:

$$\begin{aligned} EMF &= -N\,AdB/dt \\ &= -NAd(AB_0\sin(\omega t))/dt \\ &= \omega NAB_0\cos(\omega t) \end{aligned}$$

showing that the induced EMF provides a measure of the B-field strength. Using scaled SI units with N turns, B in µT, A in m^2, EMF in volts, and $\omega=2\pi\phi$ where ϕ is in Hz, then (in µV):

$$EMF = -6.284NAfB_0\cos(\omega t)$$

So the EMF will rise at 6dB per octave (ie will double when the frequency doubles) as long as the coil and associated circuitry does not approach resonance and is measured with a high impedance meter. If the EMF is measured as an RMS value, then $B_0\cos(\omega t)$ may be replaced with B_{RMS}.

For the suggested air-cored coil of 215 turns on a 100mm diameter former, self resonance occurs at about 100kHz and the coil has an inductance of about 10mH and a resistance of about 100Ω.

Construction of a sensor coil

Several small hand-held commercial meters use a high inductance ferromagnetic cored sensor coil with the advantage that if the coil resistance/inductance ratio is made low enough then it may be connected into a current amplifier – useful for wide-band measurements. This type of sensor also tends to have a comparatively high output.

But to comply with the ansi suggestion that we do not disturb the field by using a ferromagnetic sensor, and that we use a sensor which averages the field over 50 to 150cm^2, it is not practicable to make a suitable sensor with a high enough R/L ratio and a self-resonance outside of the measurement band for satisfactory use with a current amplifier.

One convenient source of coil former is a single corrugated section of ribbed plastic field drainage pipe. When a 215-turn coil is wound in the corrugation it has a nominal diameter of about 95mm and a convenient output of about 0.5µV/nT at 50Hz.

The coil, after winding, should be electrically shielded – especially if it is to be used near high-voltage power-lines. Shielding is easily accomplished by folding aluminium cooking foil into a tape and wrapping it around the coil assembly. A 10mm gap should be left next to the lead terminations, and one end (only) of the tape connected to the screen of sensor lead to stop eddy currents circulating in the shield affecting the magnetic field signal. As the signals are small, low-noise twin-twisted screened cable should be used.

Amplifier

When the signal from the coil is amplified by a factor of 200 at 50Hz it provides a convenient output of 100µV/nT, giving a scale of 0-1999nT on a 200mV DVM and 0-19.99µT on the 2V range.

Many low noise operational amplifiers are suitable. It is best to use an amplifier with little low-frequency (1/f) noise. As the signal is AC coupled, DC. stability is not a problem.

The amplifier needs a voltage gain of about 200 at 50Hz and a response falling at 6dB/octave. A very low turnover frequency is required and the high gain below 50Hz which ensues from this will cause spurious error signals if the coils are moved in the Earth's magnetic field while readings are taken.

The values given in the circuit diagram are a reasonable compromise, and cause negligible errors in the harmonics.

Measurement

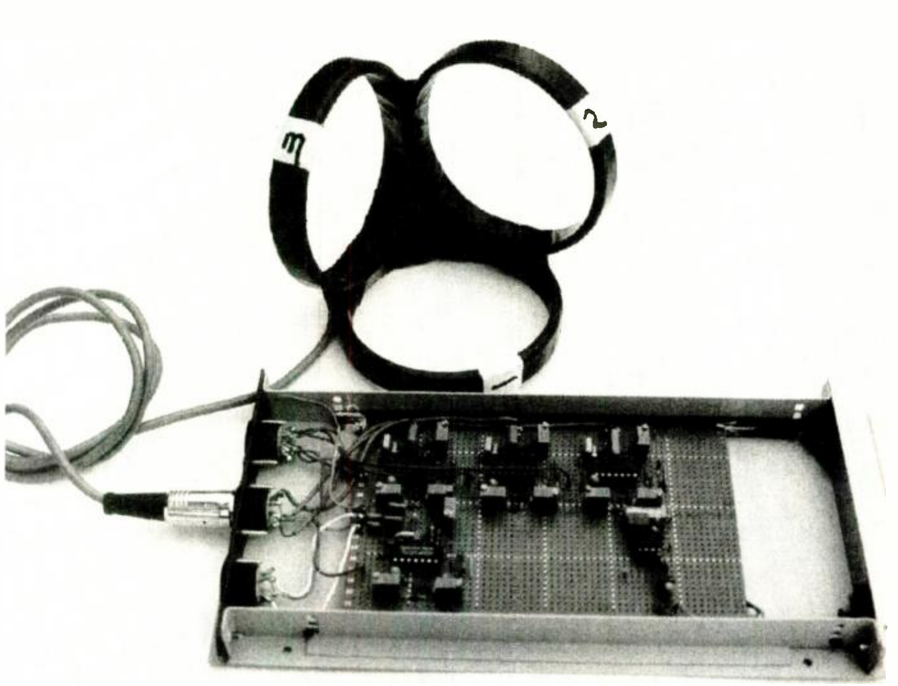
Output from the amplifier may be taken straight to a suitable AC reading voltmeter as described above. For a complete hand-held instrument, a suitable RMS/DC converter should be used, followed by a commercial digital panel meter module.

Due to the possible non-sinusoidal signal waveform, a true RMS conversion process should be used in the metering circuitry (see diagram) such as the popular *AD736* RMS to DC converter device. Most hand-held DVMs are average-measuring, RMS calibrated.

Calibration

Calibration is best carried out by placing the field sensor in a nearly-uniform magnetic field. Although Helmholtz coils have often been used, for most purposes a single coil of suitable size is acceptable. Such a set-up is described in ansi standard (644-1987) which specifies a coil with 1m sides. The conductors are assumed to be of small cross-section compared to the overall size, and may easily be retained in a frame made from 15mm square plastic electrical trunking.

For a coil with 100 turns the magnetic field at the centre of the coil is given by B=1131μT, or more usefully, 8.85mA/μT. For the sensor coil, the average field over the coil area stays within about 1% for approximately ±50mm movement from the centre.



Three orthogonal coils allow the total field vector to be calculated.

Magnetic units

The SI system of units has many derived magnetic units, but flux density and field strength are sufficient for our purposes.

SI defines the magnetic intensity or field strength (*H* in A/m) by Ampere's theorem for the intensity due to a current element:

$H = I.d_s \sin f/(4\pi r^2)$

In the older CGS system the unit was the Oersted where 1A/m=4πx10^{-3}.

Flux density is the usual quantity quoted in work connecting magnetic field with health effects. In SI, flux density (*B*) is defined as:

$B=\mu_0\mu_r H$ where μ_0 is the relative permeability, μ_r is permeability of free space and *H* is the magnetic field strength in A/m. Many references still use the old CGS unit of magnetic flux density where 1 Tesla=1000 Gauss.

The range usually quoted for power-frequency health effects is between 10nT and 100μT, or 0.1mG and 1G. The level often quoted in the childhood cancer work is 2.5mG or 250nT.

Epidemiological measurements

For epidemiological work it is imperative to record the vectors of both the geomagnetic field and the powerline electric and magnetic fields in the subject houses. Ratio of fundamental to harmonic components is also needed.

As the powerline fields vary with line loading, the data needs to be recorded over a reasonable period of time, checking where possible how typical the lineloading figures were.

Here we have examined the basis for designing an AC magnetometer suitable for serious epidemiological work if the suggested human cellular resonance mechanisms are to be tested.

If correct, then the odds-ratios for adverse health effects are likely to be much higher than present studies have suggested. ■

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The experimental sensors are being developed by the author who is a member of a workers' co-operative which designs and manufactures environmental monitoring equipment. For more information contact: Alasdair Philips, Delta-T Devices Ltd, 128 Low Road, Burwell, Cambs CB5 0EJ. Tel: 0638 742922.

BALANCING ON THE HIWIRE

John Anderson looks into the latest release of the schematic capture and PCB layout program from Wintek Corporation, Hi-Wire 2

Wintek's original offering was Smartwork, the grandfather of PC-based PCB layout. I bought a copy of this program over eight years ago; while it had some excellent features I was forced to abandon it for a netlist schematic capture system. With this background I was interested to see whether the latest Wintek offering improved on the original, and, indeed, could hold its own against today's plethora of PCB cad systems.

The specification sounded impressive, with schematic capture, PCB layout with optional autorouting, a capability of up to 255 layers and 0.001 inch resolution.

Installation

In some ways installation was well thought through; in other ways extremely tedious. There are seven disks to be transferred to hard disk once a number of inane questions have been answered. They include "Your computer appears to be an IBM PC or PS/2 (y/n)" and asking for the name of your editor and then refusing to accept it because it cannot be found in the current directory.

Once through this, the software was copied quickly and efficiently. One point of confusion is that there is a set-up file that defines where all the key files reside; this must be set into the environment variables. Installation can take care of this automatically, although can be stopped. Strangely, the details of the use of this file do not appear in the installation notes but are buried deeper in the manual, even though the software will not run without it.

The software will also not run without the parallel port hardware dongle in place.

The manual is a thick A5-sized loose leaf boxed document, professionally produced. It contains both a comprehensive tutorial and details of every command. This excellent documentation is let down by the fact that there are only two example files on disk, which are of little instructive merit.

Screen drivers are supplied for most common PC video cards, and a

Fig 1. The HiWire executive main menu

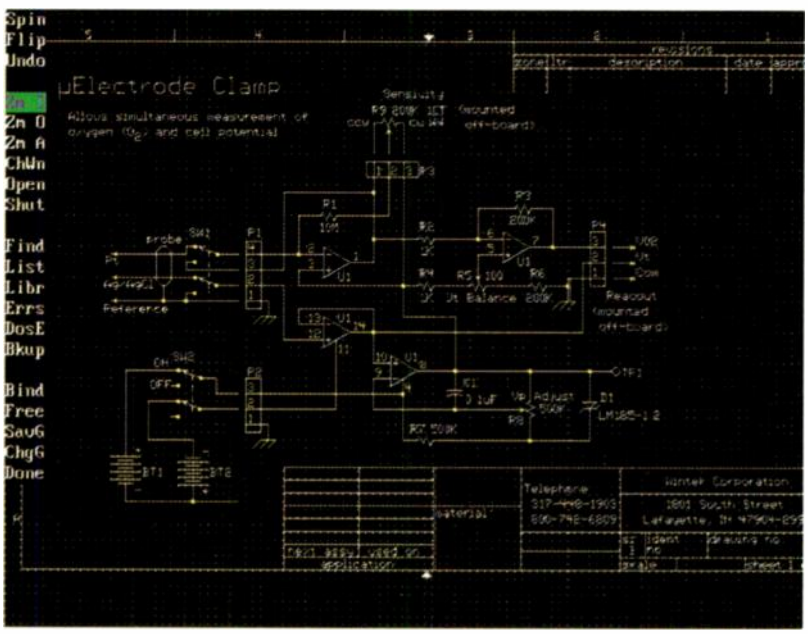


Fig. 2. Schematic capture screen and menu

Requirements
PC compatible or NEC PC9801 running dos 2 or later
640K minimum; with drivers it will use up to 15 Mbytes of extended memory, or 32 MBytes of expanded EMS memory.
Graphics: HGA, CGA, EGA, VGA and Super VGA
Parallel printer port for the dongle
Mouse: not required but recommended
Software can run from either hard or floppy disk.
Plotter output to plotter, dot matrix printer or photoplotter

number of super VGA options are also supported. The program has the ability to use both extended and expanded memory through the appropriate (LIM standard) driver. This means that really large designs are possible.

First impressions

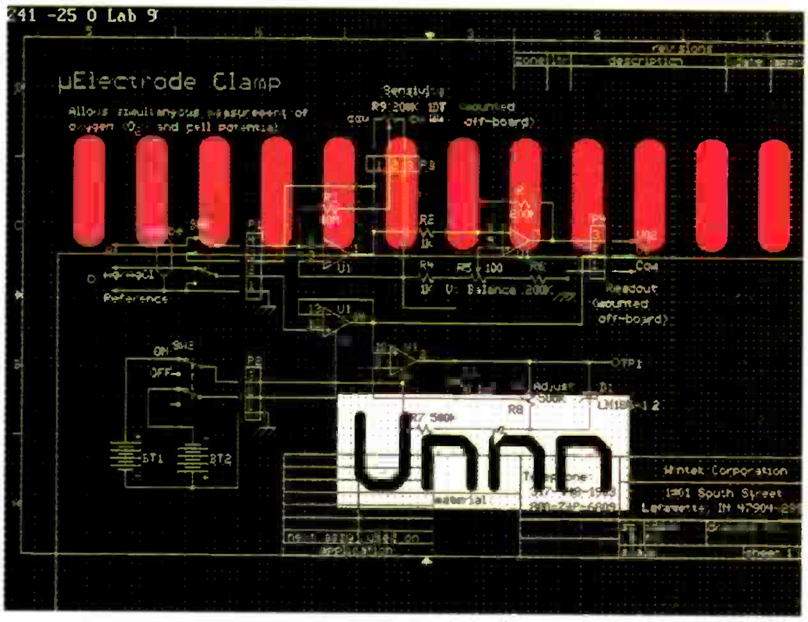
The most memorable aspect of the old Smartwork product was that it could pan very quickly, irrespective of the complexity of the artwork. Although this feature is maintained with HiWire, because computers have moved on since the old PC-XT, if you move the mouse close to the edge of the screen, the whole screen pans across almost instantly. This can be unnerving, and at one point I found the program easier to use with the computer clock reduced from 33MHz to 8MHz!

HiWire is in fact about 25 different programs welded together with an "executive". Each program is responsible for a specific task; eg, editing a schematic or producing a bill of materials, and may be executed from the DOS command line independently of the executive. Although this offers the possibility of producing your own batch file to control the process, the executive is helpful in storing the last activity at the end of a session, so that on returning to HiWire, you can continue with precisely the same file and activity as last time.

Schematics

There are standard libraries supplied with the product which allow direct input of schematic data into the drawing. The method of loading a component seems a bit crude – selection of a library via the second menu uses the 'Get' command from the primary menu – but the second part is quick and efficient. There is an alternative method that uses the list library command and then uses the mouse to select the chosen component from the list.

Fig.3. The danger of a mixed editor: PCB pads loaded onto a schematic



However, you have to enter your own component designations and the software does not track the latest in a series: eg *U1*, *U2*, etc.

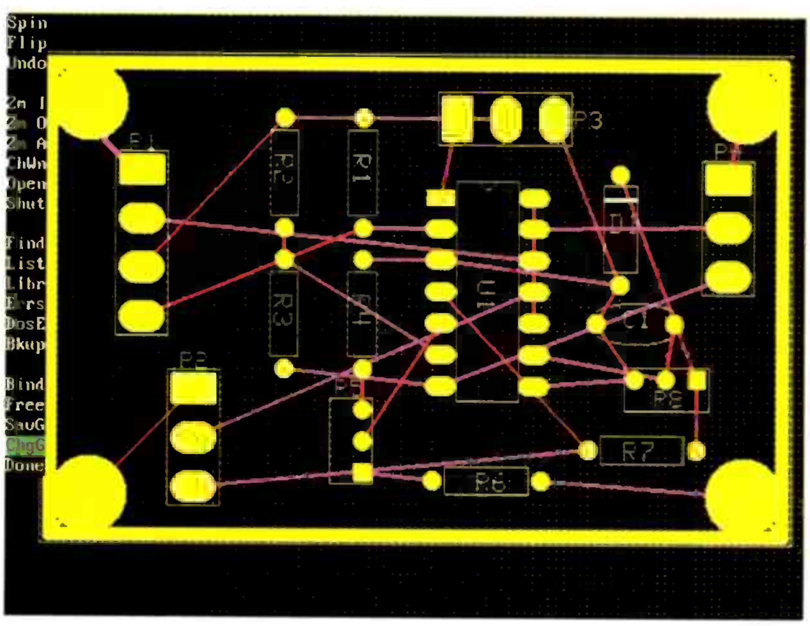
Editing is functional if a bit clumsy. One nice touch is that there is an excellent snap from, say, the line being added to the node to which it is to be connected. There is a good reason for this - the range of display zooms is limited to two in schematic mode, and at maximum magnification it is not possible to resolve either visually or with the mouse the exact attachment point position.

Nets and rats

Hiwire produces yet another incompatible netlist standard, although it does come with a program to translate other net formats to HiWire format (but not back again). There is a feature on the main menu for the autorouter, and although the test version did not have autorouting, the option still attempted to run, sadly without success!

With a successful netlist conversion, and component placement on the PCB complete, the next step is the rats nest. This is generated via another option at the HiWire executive. There are some options here to set the rats nest rules regarding layers.

Fig.4. PCB layout rats nest



PCB edit

The PCB edit facility is very similar to the schematic capture, with the same look, feel and editing commands. This single consistent interface is good from the point of view of learning the editing commands. However, it does allow loading of the wrong type of component: eg a pad footprint into a schematic!

As the editor is essentially a manual process (without the autorouter), with long tracks it can be helpful to identify the entire track which is subject to editing, for example by change of colour. HiWire does not do this, and worse follows. There is no command to break a track into two sections – ie, to convert a single length of track to two lengths with the original endpoints maintained.

In practice, the next step after editing the schematic and

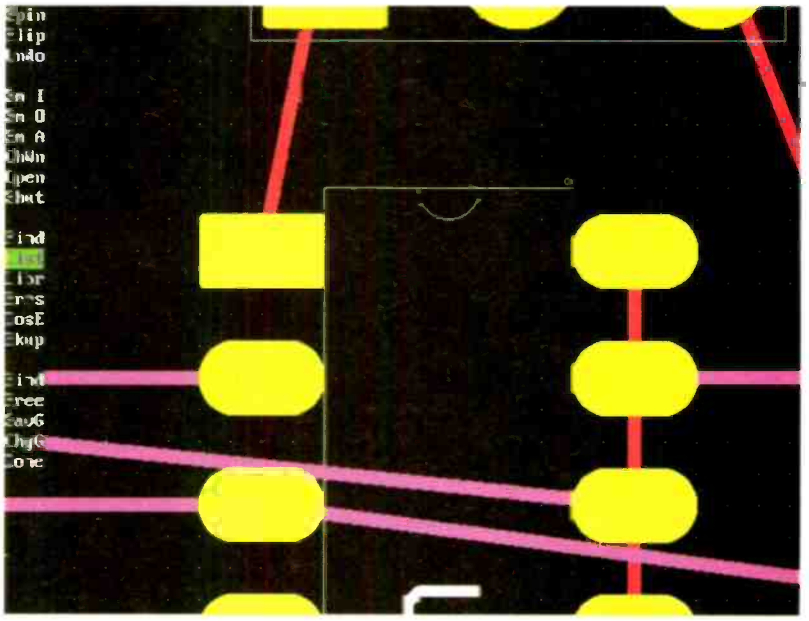
Specifications
HiWire 2, schematic capture and PCB layout
Produced by: Wintek Inc., Distributed in the UK by Riva, 3 Bentley Industrial Centre, Bentley, Farnham GU10 5NJ.Tel: 0420 22666
Price: as tested £695 – autorouting costs an additional £695

generating the netlist is to place the components on the PCB. This is slow tedious work that is carried out manually but should, at least in part, be carried out by the computer.

Design rule file

This is a file that contains all the rules for design rule checking. The design rule file is an ASCII file which could be edited directly. However, there is a method that uses the graphic editor to set the allowed pads and rules – the philosophy appears to be that the graphics editor is used for everything else so why not replace a text editor?

There is a subsequent conversion process which generates the ASCII design rule file.



Utilities

The libraries can be edited in much the same way as a schematic or PCB. The result of requesting library edit is that HiWire unpacks the library and displays all the components available in that library in the editor environment. Once edited, the edit session is closed and Hiwire regenerates the index and packs the file. Although this approach is fully consistent with the rest of the software, it is not necessarily very convenient, with the components packed closely together on a single editing sheet.

Plotting is carried out from the executive. An impressive range of plotting options is available, from HPGL plotters, dot matrix printer plotter emulation and photoplotter support.

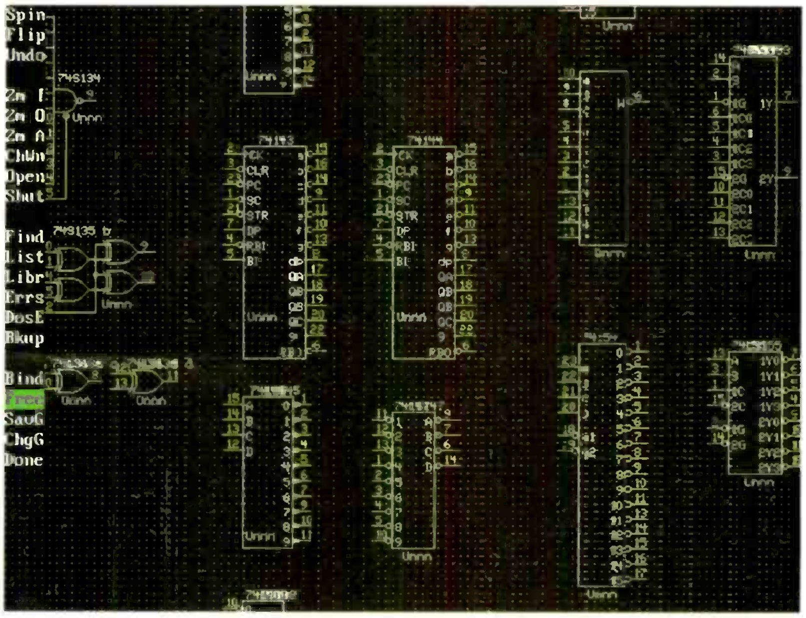
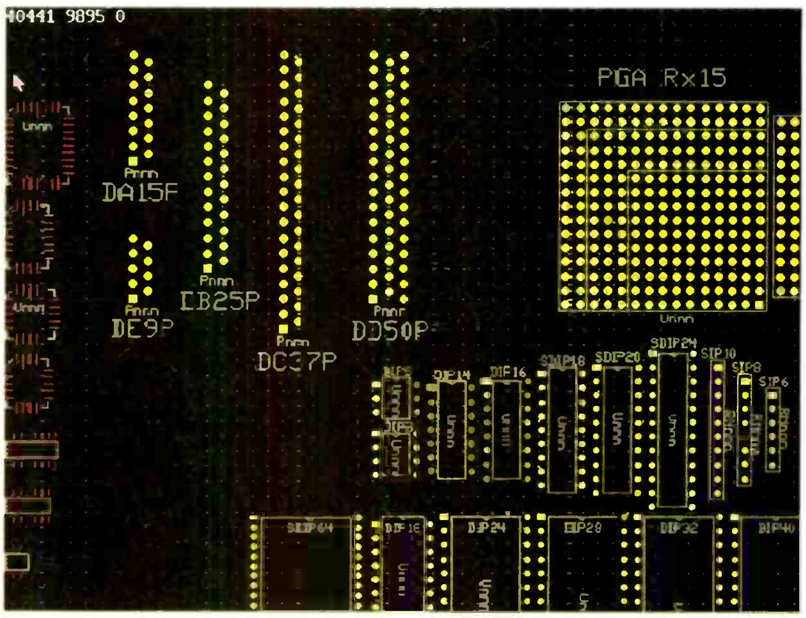


Fig.5. Zoomed-in for more accurate edit - this is not available in schematic capture

Fig.6. Schematic library edit

Fig.7. PCB footprint edit



Conclusions

The fast pan still exists and is as good as the fastest workstation. Although the package is all there, there are some rough edges. Firstly there's the rather quirky user interface, requiring, for example, several user operations just to zoom in and out.

As with its predecessor, Smartwork, PCB data is held in an encoded fashion, rather than in ascii, and this prevents any change to the data except through the use of HiWire itself.

The price of schematic capture and PCB layout software has fallen dramatically over the last two years, and it seems ridiculous to have a relatively low cost utilitarian product like this protected by a dongle. This is a lesson learnt over five years ago by the large commercial software players like Lotus and Borland.

So for PCB designs is HiWire a difficult act to follow? For smaller designs, or indeed for occasional use, on balance I'd be happy to use this software. For larger designs, the manual placement and the user-unfriendly interface would make the task both tedious and slow. It should be remembered that this type of product should be offering improvements in productivity and final quality.

The market has matured significantly since the heady days when Wintek first launched its PCB layout products, and there is now little to recommend this product over the rest of the pack. ■

SIMULATING BLOCK DIAGRAMS

A simple nodal-analysis program can revolutionise the approach to circuit development. Ian White tries out a program that analyses whole systems...

One of the most powerful software tools for electronic engineers is a nodal-analysis program. The usual way this works is that you draw out a circuit, identify the component values and other parameters, and list their interconnection points or "nodes". The software then calculates the properties of the circuit, at DC and/or a range of frequencies. To predict the performance in the time domain, the software does a Fourier transform of the frequency-domain results. Just think what a program could do that analyses whole systems.

If you design systems, you need to optimise performance at the block-diagram level first, so that you can approach the detailed circuit design with a clear specification for each module. Although conventional nodal-analysis software will allow you to simulate a complete system by detailing every single component in every single module, you end up with enormous simulations that run very slowly and present you with too much detail too soon.

Tesla, from the Tesoft company in Georgia, USA, is a different kind of nodal-analysis program that always operates at the block-diagram level. Since it does not need component-level details, Tesla helps you to think about the system as a whole.

On the block

A block diagram considers a system as a collection of 'black boxes', each of which performs a specific function. To analyse the system you only ask what each box does; there is no need to know what goes on inside the box. This is the exact opposite of a component-level analysis, where everything in the circuit has to be described before you can discover what it does.

Since you often cannot model the detailed internal workings of an IC, digital systems lend themselves particularly well to block-diagram analysis. There are many digital logic analyser programs available, and they generally come equipped with libraries of models for well-known digital ICs. These programs work in the time domain: they step through time and determine the state of the system at each time-step by applying logic rules and taking account of transition times and propagation delays.

Tesla operates in the time domain too, but it considers voltage levels rather than logic levels. This makes it equally at home with analogue, digital or hybrid systems.

Example

A simple example will show how Tesla works. **Figure 1** is the block diagram of a phasing-type single-sideband mixer, using modules taken from Tesla's extensive library. Because each of the individual mixers is balanced, neither of the two input frequencies *F1* and F2 appears at the output.

The two possible output frequencies are (*F1+F2*) and (*F1-F2*); but if you remember about the multiplication, addition and subtraction of sines and cosines it is easy to show that the 90° phase-shifting suppresses one of the two mixing products, leading to a single-sideband output. According to the application, the basic system in Figure 1 could be configured as an image-rejection receive or transmit mixer, or as a single-sideband exciter with audio or baseband input.

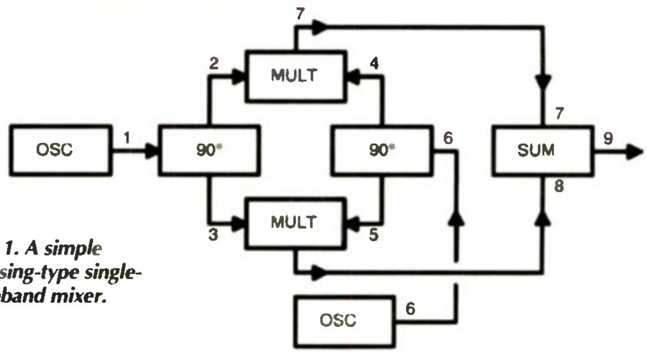


Fig. 1. A simple phasing-type single-sideband mixer.

Tesla input file for the system shown in Figure 1

User label	Node connections	Tesla block	Parameters type
Sig_osc	1	OSC	F=1E4 $V=V_{in}$
Sig_90	1 2 3	NB90	F=1E4
MOD1	2 4 7	MULT	G=1
MOD2	3 5 8	MULT	G=1
RF_osc	6	OSC	F=1E5 V=1
RF_90	6 5 4	NB90	F=1E5
ADD	7 8 9	SUM	G1=1 G2=1

The parameter V_{in} is a user-defined variable which can be entered just before the calculation begins. As configured here, the system will generate the lower-sideband mixing product.

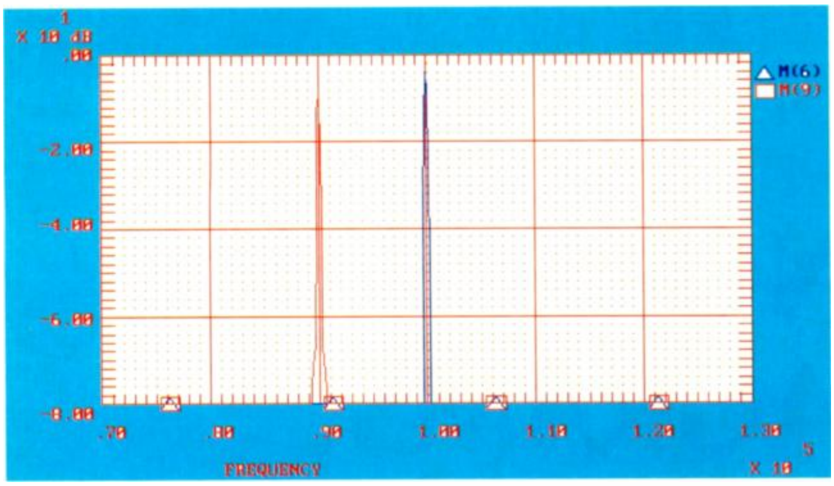


Fig. 2. Output plot from Tesla simulation of Figure 1 at nodes 6 (suppressed carrier frequency) and 9 (SSB output). As expected, only the two frequencies are present and the opposite sideband is suppressed.

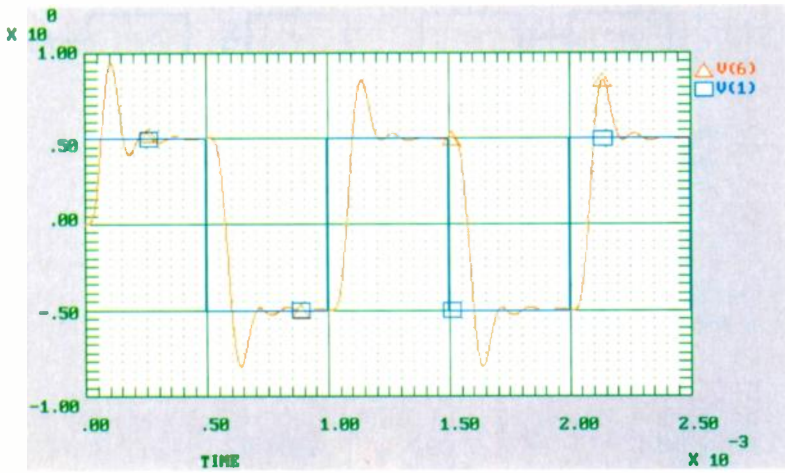
Entering a block diagram

The easy way to set up a new block-diagram system for Tesla is simply to draw it on the screen. Tesoft supplies a symbol library for all the generic Tesla models which is compatible with the industry-standard schematic drafting program OrCad. You can run the relevant part of OrCad from within Tesla and enter all the input data as part of the drawing. Assign numeric labels to the wires and to the input/output ports of all the blocks, and Tesla and OrCad between them will do the rest.

The pedestrian route, which I used for this review, is to sketch the block diagram on paper and type in a data file using an ascii editor - see panel. A nice touch is that you can run your own favourite ascii file editor from within Tesla. The data file then has to be processed by Tesla to generate an internal map of the node linkages and a mathematical model of the system.

This "simulation" step is automatic if you use the OrCad input route, and is rather like compiling the source code for a program. If you entered any of the data in an incorrect format there are helpful error messages indicating whereabouts on the offending line the error occurred. If several messages appear, however, only the first is guaranteed to be meaningful; the rest may well disappear when you correct the one real error.

Fig. 4. Output signal at node 6 (Figure 3) compared with the driving signal at node 1. Note the accurately simulated overshoot, which is typical of phase-lock demodulators.



Having entered and compiled the block diagram, you must decide which nodes to calculate results for. The more nodes you keep, the longer the simulations will take.

If you wish, you can leave some detailed parameters in the data file to be entered manually at this stage, so that their values can be changed without having to go through the "simulation" step again each time. The data file shown in the panel uses V_{in} as such a parameter.

Analysis

Tesla operates primarily in the time domain and you must always perform this type of analysis first. It is something of an art to select the time-frame and step interval, and your choice will depend on the frequencies involved.

For an accurate analysis, the overall time-frame should span several cycles at the lowest frequency involved, while the time-step should be considerably shorter than the reciprocal of the highest frequency involved. In case you intend to do a subsequent Fourier transform into the frequency domain, Tesla indicates the implications for the available frequency span and resolution before you launch into the time-domain calculation.

Output facilities in Tesla are extremely flexible. You can view a colour 'oscilloscope trace' of the waveforms at any of the selected nodes, alter the axes for the most informative display and send the results to a variety of printers or to a pen plotter. Waveforms at digital nodes can be viewed on a logic-analyser type of display which separates the waveforms vertically instead of overlaying them.

The time-domain behaviour of our example circuit is not particularly interesting, so let's press straight on to the Fourier transform which calculates the frequency power spectrum for each node.

Tesla does its best to meet your stated requirements without throwing too much information away. The options for the number of transform points increase by powers of 2 and are chosen automatically - typically 1024, 2048 or the maximum of 4096 points. If you are demanding more than a Fourier transform can possibly provide from the time-domain data currently available, you can increase the frequency interval, decrease the overall frequency span or go back and have Tesla make a more suitable time sweep.

Output

The final output for our example circuit is shown in **Fig. 2**. Theory predicts a perfect single-sideband output signal at node 9 (the suppressed carrier frequency at node 6 is also shown for reference) and Tesla predicts exactly what we expect.

The spectrum broadening in Figure 2 is caused by the trade-off between computing time and the step lengths in the time and frequency domains. Viewed on a greatly extended dB scale there are no significant errors in this simulation above the –100dB level, which should be good enough for anyone!

Even more interesting is to analyse the same block using the MIXGP mixer model, which takes account of imperfections

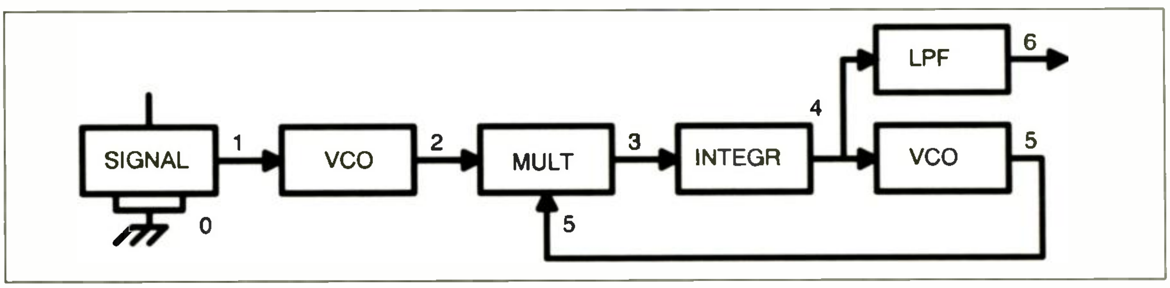


Fig. 3. Simple FM signal generator and phase-locked loop demodulator taken from the Tesla demo disk. See text for details.

Model library

The Tesla model library contains over 60 generic block-diagram models and is divided into five parts:

- Linear analogue functions (including comprehensive filter simulations and real-life amplifiers and mixers with distortion parameters);
- Non-linear analogue functions (including rectifiers, analogue multipliers/dividers, log and square-root functions and a user-defined non-linear function);
- Digital functions (basic gate and flip-flop functions; 4-bit binary counter, adder, comparator and shift register; and a digital phase-frequency detector);
- A-to-D interface functions (including 8-bit converters, multiplexers and demultiplexers and a sample-and-hold), and
- Test and measurement functions.

Each library includes a wide range of generic models (far too many to list here), each of which can be configured by parameters in its specification statement as shown in the panel above. Models can also be grouped into named sub-circuits which can then be called using a single line in the data file, and sub-circuit definitions can call in turn on other sub-circuits.

The Tesla library is so extensive that almost any analogue/digital system can be modelled at the block-diagram level. Equally important, a wide variety of input signals can be simulated by 'building' a specialised signal generator from the test and measurement functions library.

Signal sources include multi-function sine/square/triangle, pulse and noise generators and a DC source. By summing these basic waveforms or using one generator to modulate another, almost any conceivable test signal can be generated. If all else fails, you can generate your own voltage-time data file and use that as an input signal.

The complementary range of measurement functions includes "test instruments" (eg an RMS-reading AC/DC linear/dB voltmeter and a phase meter) which can be attached to any node to make measurements. Bit-error rates in digital communication systems can be simulated and measured using the test signal generator provided in the Tesla library, linked to a complementary error checker.

In all cases the output data can either be plotted or read out to a disk file.

in port-to-port signal rejection and inter-modulation up to 5th-order. You can then predict the sensitivity of a real-life system to mixer imperfections and errors in the quadrature phase shifters.

More complex cases

Tesla really shines in analyses where transient behaviour is important - for example in phase-locked loops and systems that have to deal with pulses and random noise. The free Tesla demo disk deals in fine style with the phase-locked FM demodulator in **Fig. 3**. The first two blocks, *FCNGEN* and *VCO*, form a frequency-modulated signal generator and the remainder of the circuit demodulates the FM signal.

After the mixer, the integrator *INTEGZ* determines the loop bandwidth and damping factor, and the Chebyshev lowpass filter *CHEBL* smoothes the output signal. The comparison between the driving signal at node 1 and the recovered signal at node 6 **(Fig. 4)** clearly shows the effects of the loop damping parameter.

Tesla calculates the node voltages at each time-step, and most analogue models can be configured to simulate output saturation which prevents the instantaneous output voltage from swinging beyond either supply rail. This feature is illustrated on the Tesla demo disk, and shows how extremely useful a simulation program can be.

Design errors can be revealed and corrected at little cost before the real-life system is built. One slight deficiency in voltage modelling is that Tesla cannot automatically model the attenuation effects of fan-out loading or impedance mismatching. Even so, Tesla is so flexible that you could probably devise a 'work-around' solution for almost any specific case.

Supplier details

Tesla version 1.1 $695
Modgen user model generator $495
OrCad/SDT parts library $195
P-Cad PCCAPS parts library $295
Tesla demo disk free
Tesoft,
205 Crossing Creek Court, Roswell, Georgia 30076, USA.
Tel: 0101 404 751 9785 Fax: 0101 404 664 5817

System requirements

IBM PC compatible with hard disk (80286 or better, math coprocessor highly recommended)
512-640K free ram
512K or 1024K disk cache recommended (eg SmartDrive, PC-cache)
VGA (recommended), EGA, CGA or Hercules graphics
Dos 3.0 or later

Model your own

Because some users will always want something different, Tesla offers them facilities for creating their own models. Tesoft encourages an active interchange of user-generated models through a telephone bulletin-board.

Building your own models requires a certain skill, plus an additional Tesla module called MODGEN and the Microsoft PC Fortran compiler. Not having the compiler I could not test MODGEN for this review.

Tesla models are actually compiled Fortran subroutines, called in sequence during a time-stepping simulation, and the purpose of MODGEN is to manage the interface between your source code, the Fortran compiler and Tesla. Like any other Tesla model, a user-generated model simply calculates the relationships between instantaneous input voltages, output voltages and the global variable TIME, for whatever circuit happens to live inside your particular 'black box'.

User interface

Tesla's user interface is certainly a change from the trendy PC world of wimps, GUIs, icons and other items of pop-up jargon. Tesla is completely command-line driven from the input prompt: apart from the excursion into graphics for plotting, output simply scrolls up the screen.

This straightforward approach is a mixed blessing. On the one hand, many engineers prefer typing in commands, because it puts them directly in touch with the workings of the software. To help things along, Tesla has a macro facility for programming repeated runs, or for setting up a batch of long simulations to run overnight.

But even dedicated command-line users like some creature comforts. Unfortunately Tesla's 'glass teletype' user interface knows nothing about the cursor keys or full-screen editing. The latest version has only just introduced the facility for pressing F3 to retrieve the last-entered command line.

Maybe I'm getting soft, but in a command-driven environment such as MS-dos or Tesla, the ability to recall any previous command line, with full cursor/insert/delete editing, should be regarded as a basic essential.

I was disappointed to find that Tesla makes so little use of the PC's keyboard and display facilities, though Tesoft is developing a modern menus-and-mouse interface for the next release.

Pressing F1 at any input prompt will produce a cross-referenced Help screen, though personally I found the crowded multi-colour layout less than completely helpful. In any case, Tesla is not the kind of software you can learn 'on the fly'; you really need to read the manuals, which are very thorough and well-produced. The review copy of the software was undergoing a change to the new name of Tesla which is not yet completely reflected in the program and the manuals, though this caused no problems.

Conclusions

Tesla is a unique block-diagram simulation tool for the PC, and could very quickly prove itself indispensable in a busy system development lab. System designers who are presently struggling with unnecessarily complicated component-level simulations would do well to try the demonstration disk, which is a fully working version of Tesla with a small but useful selection of models.

Since the basic program is advertised as "about the same price as a bench DVM", Tesla could pay for itself in a very short time. ■

FLOATING INTO THE MARKET

Analog Devices' entry into the floating point DSP processor market, the ADSP-21020, should add an extra edge, says Allen Brown.

Fixed word DSP devices have been around for several years and although floating point (FP) DSPs have also been available (the first was the AT&T *DSP32C*) their penetration into the market has been slow. Cost is partly to blame – and the perceived difficulty of designing with FP processors. But now that Analog Devices is vying with Motorola, Texas Instruments and AT&T, we should see much more competition in the cost of FP-DSPs.

The advantage of FP-DSPs is that they lend themselves more favourably to high level languages (such as C) than do the fixed point devices and efficient high level compilers can be readily realised.

FP calculations can be performed directly without having to call FP simulation routines – as is the case with fixed point DSPs. The pressure to code in C rather than assembly language will be even greater now that high speed, low cost target processors are freely available.

One of the advantages of using FP-DSPs is the lower emphasis that the engineer has to place on software overheads (no scaling needed). This arises from the dynamic range of 32-bit FP numbers which is $1.8\ 10^{-38}$ to $3.4\ 10^{+38}$ thus reducing the need to watch for numerical overflows.

ADSP-21020 features

Architecture of the *ADSP-21020* has a lot in common with that of Analog Devices' earlier fixed point DSPs. It is able to perform 40 million floating point operations per second (Mflops) in a sustained mode and 60 Mflops in peak mode using its on-board cache. Its numerical format conforms to the 40-bit IEEE-754 floating point standard and it is able to perform a 1024-point complex FFT (the standard bench mark) in under 1 msec.

With its on-board counter, the *ADSP-21020* can execute looped code without the use of branch instructions; this is often referred to as zero overhead. However for implementing context switches to a subroutine etc, the *ADSP-21020* has a secondary set of registers which can contain the volatile portion of the next task and this can be switched in a single machine cycle (50 nsec). By way of contrast, task switching usually takes several µsec on a 32-bit microprocessor.

The *ADSP-21020* has inherited several features from the *ADSP-2101*, notably the extensive array of addressing modes, including a provision for circular buffers which is quite useful for implementing digital filters.

Architecture

A block diagram of the *ADSP-21020* is shown in **Fig. 1**. It has three functional units which are given over to computational tasks: the ALU, the barrel shifter and the multiplier and accumulator. These three units can act in parallel and can process 40-bit extended precision FP or 32-bit fixed point numbers.

The multiply and accumulate operation takes place in a single clock cycle and the results of each calculation can be deposited in the 80-bit accumulator. The ALU is able to perform operations on both fixed point and FP numbers; likewise, the barrel shifter is able to offer a variety of operations including bit and field manipulation.

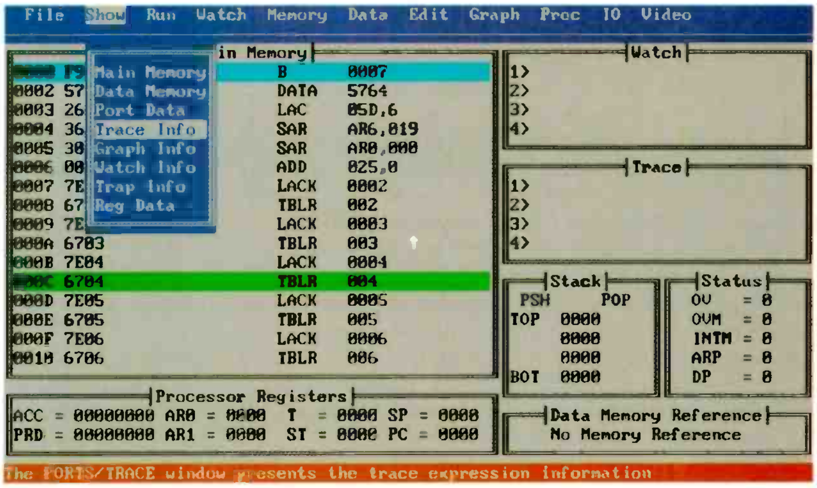
As Figure 1 shows, there are two pairs of address and data buses. These are used with remarkable efficiency to perform a simultaneous access of both program and data memory. Both memory areas can, if required, store operands (data). This means that, when combined with the on-board cache memory, the *ADSP-21020* can simultaneously fetch operands from both external memory areas and an instruction from cache in a single cycle, thus achieving the 60 Mflops performance.

The cache operates in an interesting fashion: normally an instruction fetch from external memory is as fast as a fetch from cache. However when there is a conflict, which can arise when the external program memory is used to store data, then the program instructions are cached. The cache can hold up to 32 instructions. To calculate the appropriate addresses, two address generators are used (*DAG-1* and *DAG-2*), one for each external memory area. The *ADSP-21020* can accommodate memory devices with access times of 35ns: however wait states can be inserted to allow slower (and cheaper) memory chips to be used.

To enable data exchange between the functional units there is a 10-port register file which has two sets of 40-bit registers. Referred to as the primary and alternate registers they play an important role in effecting fast context switching.

The *ADSP-21020* can be interrupt driven and is armed with twelve possible exceptions: four external hardware driven interrupt exceptions and the remainder internally driven through software. When an external interrupt occurs, the *ADSP-21020* automatically stacks the arithmetic status and the mode registers.

Some of the internally generated exceptions allow for fault handling and subsequent recovery – a feature commonly found on general purpose microprocessors. The *ADSP-21020* also has an internal timer (32 bit counter) which generates an interrupt when its count expires – a particularly useful feature for multitasking real-time processing. The address



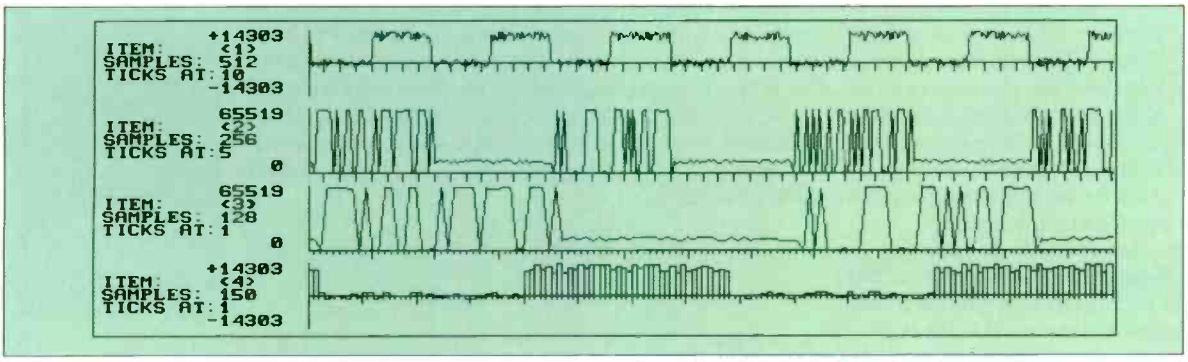


Fig. 1. Block diagram of ADSP-21020 Floating Point DSP.

Fig. 2. Windows are selected in the simulator using the drop-down menus.

bus for accessing the *ADSP-21020* program memory (PMA) is 24-bits wide whereas the address bus for accessing the data memory is 32-bits wide. This represents two memory maps of some 4311M addresses between them. With the two data buses (one for instructions and the other for data) being 48-bits and 40-bits respectively, system designers will not experience too many problems of memory area shortage.

The high degree of parallelism of the *ADSP-21020* is clearly demonstrated by the width of the program memory data bus (PMD) which carries the 48-bit width instructions. A lot of information can be stored in a 48-bit instruction!

Peripherals are memory mapped in the *ADSP-21020* and to enable slow device interfacing, the designer can use a combination of wait states and acknowledge signals. In addition, the *ADSP-21020* has provision for allowing external devices to access the system buses while its own bus buffers relax into a tri-state condition.

Due to the problems involved in amending errors in software when running in real-time, it is now commonplace for chip designers to provide an on-board feature to facilitate debugging. The *ADSP-21020* has a jtag test and emulation feature that conforms to the IEEE 1149.1 standard.

The jtag (Joint Testing Action Group) standard defines an interface protocol for serial scan testing. This can be used with great effect to provide a low cost method of monitoring the chip's register and memory contents as the device progresses through the code. This makes the design of development tools much easier and should result in the appearance of low cost in-circuit emulators from third parties.

Software support tools

To ensure that the full potential of the *ADSP-21020* can be realised, Analog Devices provides an extensive range of well designed software tools. Versions are available to run under MS-DOS and the tools include a C compiler (*CC21K*). The compiler supports the Ansi C standard and is a valuable tool in the development of products for the *ADSP-21020*. However the compiled code it produces usually fails to exploit the full parallelism of the *ADSP-21020*.

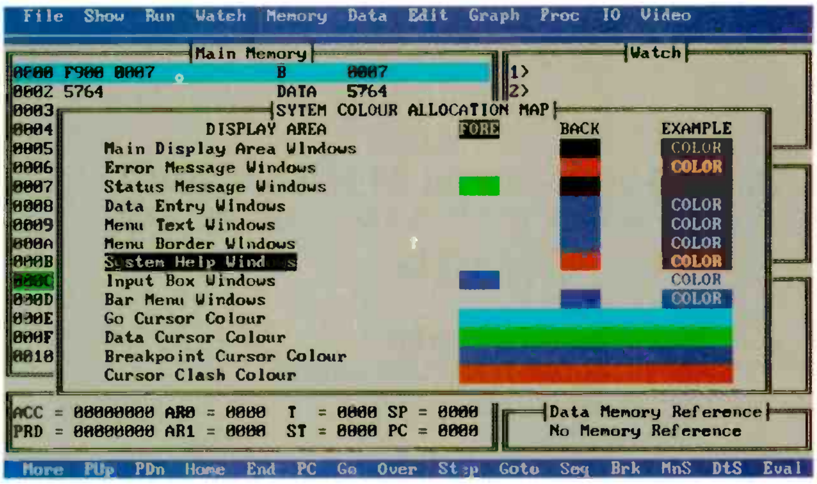


Fig. 3. Selecting break point options from the drop-down menu.

When the processor is required to execute time critical processes, then the user must resort to programming in assembly language. However, the C compiler does allow the insertion of assembly language modules which is very appealing from a software engineering prospective where structured design is often sought.

The assembler (*ASM21K*) initially calls a pre-processor (*A21000*) which sorts out the INCLUDE statements at the beginning of the code. It does have one disconcerting feature: when it is evoked it creates a virtual memory file of 2.2Mbyte on the hard disk.

The assembler offers a range of programming directives, including conditional assembly options. As expected, it will accept numbers in a variety of formats and the user can define the precision of FP numbers by means of the PRECISION directive. There is also an option for creating macros. However with the high cost of fast access memory it is more cost effective to have subroutines rather than macros.

The assembler is supported by a library containing several commonly used maths routines (trig and log functions). As with other Analog Devices software products, before the linker can be evoked, an architecture file must define how the memory is partitioned into ram, rom and i/o space.

Once this file is in place, the assembled modules are passed through the Linker (*LD21K*) which assigns addresses to the machine instructions, resulting in an executable file. One of the switches on the linker allows the user to generate a map file, providing information on where each module is located in the memory map of the *ADSP-21020*. A routine (*SLP21*) will dividie up the executable code into suitable block sizes for downloading to a prom blower to produce roms.

Software simulator

Once an executable file has been created, it can be tested on the software simulator (*SIM21K*). This is a very well-designed windows-based product. Each architectural feature is assigned to a window and the user can choose which feature he wishes to monitor during the simulation of the code.

Figure 2 shows a typical setup with the drop-down menus forming the command line at the top of the screen. From the Register option, the user can select which window is required on the screen by clicking the mouse on the option.

Each window can be moved by attaching the mouse cursor to the window name and dragging the window around. The size of any window can be adjusted by attaching the mouse to either corner of the window and moving it accordingly.

The windows can overlap as seen in Figure 2 and while single stepping, every window is refreshed if its display data is affected by the instruction execution. As seen from **Fig. 3**, the user has the opportunity of setting breakpoints and these are indicated by the instruction being highlighted.

Possible applications

FP-DSPs are particularly good at performing recursive algorithms (multi-stage calculations where the results of the current stage are fed into the next stage). When implementing recursive algorithms on fixed point DSPs, the engineer has to devise clever bits of code to ensure that maximum precision is maintained. This minimises the effects of round-off errors, which can easily lead to instability.

With FP-DSPs this is unnecessary in the majority of applications and naturally leads to faster solutions. As a result of this, one area where the FP-DSP will find application is in the more widespread use of speech compression through linear predictive techniques[1]. The algorithms for performing this task are recursive and the option of using FP-DSPs will ensure a greater algorithm stability with little overhead.

Another application to which the *ADSP-21020* would be well suited is in image compression, using the Fourier Cosine Transform. Given that the demand for image compression for archiving purposes and fast fax machines will continue to grow, the *ADSP-21020* is well placed for these applications if Analog Devices makes it available at a competitive cost and provides the good support which is found with its other DSP products. ■

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Tel: 0932 232222.

References

1. *Digital Signal Processing*. Analog Devices. Prentice Hall (1990) Chap 10.

Developing embedded systems from the top down

New hooks in the silicon are expanding the role of high level languages in microsystems development.

By Julia King

Since first appearance of the 8-bit micro there has been a need for development systems to speed up the programming. Developers of support systems have their work cut out to keep pace with increased silicon complexity. If 8-bit designs needed support systems, imagine where that leaves the current generation of 32-bit devices...

The PC explosion has also affected the route taken by some designers; where microprocessor development systems (MDS) used to be based on logic analysers or specially designed development systems, the PC or workstation has become a workhorse.

Microprocessor diversification has led to stratification in MDS suppliers: there are still great numbers of systems being produced based on the *Z80* despite its age. MDS programming style therefore ranges from code written in assembler to code written in high level object-orientated languages such as C++.

The stuff that dreams are made of? The engineer sees the PC as a VDU and keyboard attached to a full development system running on the target's CPU. Reality is that the PC performs the compiling and interpretative functions but that all time and I/O critical actions occur on the target board. (Source: Comsol)

Dropping a line

The number of lines of code produced by high level languages is far in excess of those produced by machine level languages. The development system therefore has to be capable of compiling lines rapidly enough to be useful. Andy Green of Crash Barrier comments: "With embedded systems, I would be surprised if more than 15% were done with any C in at all. When people are used to getting 4-8k of code, they think there's no point in using C which would blow this up to 16 or 32k, mean using bigger roms, etc". He admits, however, that 98% of applications designed to run on the PC probably use C and that over the long term "virtually everyone will shift up to using C compilers".

Speaking from the opposite camp, Noral's Keith Norton sees a move towards working in higher level languages and allowing debugging at source level. However, he admits that there are problems implicit in the amount of code being used: "The compiler has to be efficient", he says. "A lot more code has to be got down to the target system via the emulator, maybe at speeds above 1Mbaud."

The two opinions highlight the fact that development systems are used for an extremely wide range of applications. These might vary from simple enhancements to a basic weighing machine to developing a processor for a guided missile system.

At the lower end of the market, Chris Stephens from Computer Solutions sees an emphasis on software development tools rather than on the hardware. Monitors sit on the target board and provide the ability to put in features such as breakpoints (interrupting the chip's working to look at the code; trace memory can then be used to see what has been happening with the code). Stephens also says "interaction with the target system through software is important".

CPU value

In Norton's view, "the CPU is only of value if support tools are available for it". Without design support, emulator manufacturers find difficulty keeping pace. There is a feeling that developers of complex processors should begin releasing design details to MDS producers much earlier than the product launch date. Otherwise, there is no hope of producing support tools early enough to be really useful.

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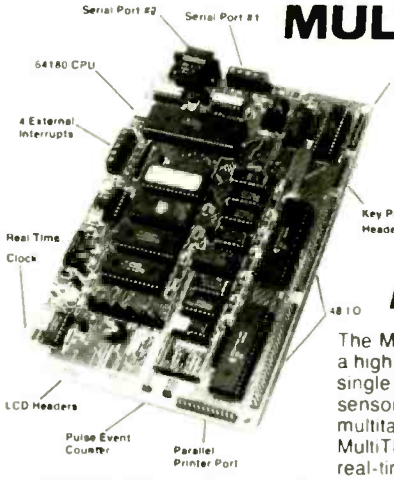

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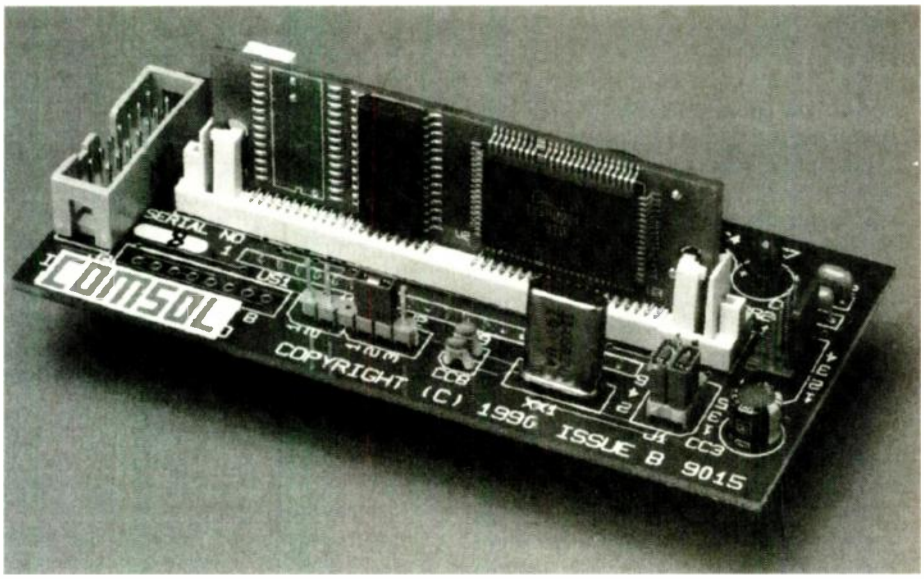


Comsol's ICE 51 for the 8031 processor costs £225. A daughter board replaces the 8031 in the application. The program under development is held in 32kbyte of ram, which also provides sufficient space to enable a monitor program to be held.

Simulation is becoming the preferred development route for simpler devices because it offers a much cheaper development route. However the simulation approach is not really suitable for the development of real time systems, says Norton.

There is a similar divergence where monitors are involved. Norton believes that "monitors are fine as long as the code works. There's no trace information, but they're useful as a complement to emulators".

Monitors may not be able to stand on their own up at the top end of the market but, lower down, there are those who believe in their worth. Cliff Malcolm, Sales Manager of *8051* specialist, Nohau, is a firm believer: "Rom monitors provide full source level debugging down at the C level", he says. They sit as a small bit of program inside the developer's own code on the target board and allow memory to be looked at, changed and generally manipulated.

A monitor is all that is needed "if you have a full working target", says Malcolm. It is much cheaper than going the full ICE (in-circuit emulator) route, with a decent rom monitor costing around £2000 – the price of a debugging option for an ICE itself. "Some CPUs are so hard to emulate that this is the only approach", he claims.

Monitors can also be used to get more out of in-circuit emulators. Given that emulation is only required during a small part of the development cycle, a number of engineers can 'share' one emulator as long as each has his or her own monitor.

Motorola has recognised the type of problems being experienced by MDS producers by improving the emulation interface on its *68332* processor, allowing information to be put in and got out much more easily than before. This is in line with the trend towards a faster interface between the emulator and host.

In Chris Stephens' view, the improved interface might be termed "a free ICE with every chip".

"It's almost got a monitor on the processor, allowing you to talk independently to the processor... It's like putting software into hardware".

An influential Case

Although at the top end, Case tools are starting to be brought into the MDS equation. Ralph Hodgson, European technical director for IDE, says that in the industrial automation field there is a move towards executable models based on finite state machines for embedded systems.

In CIM and shop floor environments, work is being done on producing architectures that can lead to automatic code generation. In projects such as IDE's link with SES Work Bench, an object-orientated approach enables closer integration with development schemes to be achieved.

For instance, software modelling means that safety-critical systems can be looked at from the point of view of what Hodgson terms 'ill time' as well as 'well time'; that is, modelling what the system should not do as well as what it should do.

The safety critical label is equally applicable to banks – eg, wrongful retention of cards at ATMs – as to the more predictable applications such as flight or missile control. Safety criticality is an area increasing in importance: Hodgson says that membership of the DTI's 'Safety Critical Club' has doubled over the last 12 months.

Other developments in the field of processor technology will continue to influence the MDS market. For example, quad flatpack packages don't fit conventional sockets, making it difficult to slot devices in and out. One approach discussed by a group of manufacturers has been to make one pin 'responsible' for halting operation – ie at breakpoints. According to Norton, early reports are that the approach is not working, however.

As developments continue apace, chip manufacturers will have to collaborate more fully with developers of support systems to ensure that the whole market is kept both informed and satisfied. One area that will doubtless see considerable development lies with yet another user interface: the graphical user interface or GUI. Case tools will make extensive use of this.

A very different piece of ICE

Does this mean that the low to mid-range ICE market will die away? Will a new generation

Addresses

Computer Solutions Ltd, Canada Road, Byfleet, Surrey KT14 7HQ. 0932-352744.

Crash Barrier, 18 Oxford Street, Wellingborough, Northants NN8 4HY. 0933-224366.

Flash Designs Ltd, St Andrews House, PO Box 167, Crawley, RH11 9YE. Phone 0293-551229.
IDE, 14-16 Frederick Sanger Road, The Surrey Research Centre, Guildford, Surrey GU2 5YD. 0483-579000.

Lauterbach Datentechnik GmbH, Fichtenstrasse 27, 8011 Hofolding, W.Germany. 010 49 8104-9075

Motorola, The Manager (68332), Literature Distribution Centre, 88 Tanners Drive, Blakelands, Milton Keynes MK14 5BP. (No phone number.)

Nohau UK Ltd, Station Mill, Alresford, Hants S)24 9JG. 0962-733140.

Noral Micrologics Ltd, Logic House, Gate Street, Blackburn, Lancs BB1 3AQ. 0254-682092.

Smart Communications, 2 Field End, Arkley, Barnet, Herts EN5 3EZ 081 441 3890 of low cost universal development systems take their place? David Doo of Flash Designs claims to have come up with a modular, low cost PC-based development system which uses cheap software to personalise it for a particular processor.

Conventional ICE is limited to a single processor or family. The Flash system, based on its existing ram/rom emulators, works through the system memory sockets (program or data) to achieve control without having to worry about things like crystal speed. Because it is modular, Doo says that the developer can move from 8 to 16 bits without having to purchase a complete new system.

The software costs £25 for each different processor. It also works with the company's Turbo-Trace real time debugger capturing address, data and external probe status.

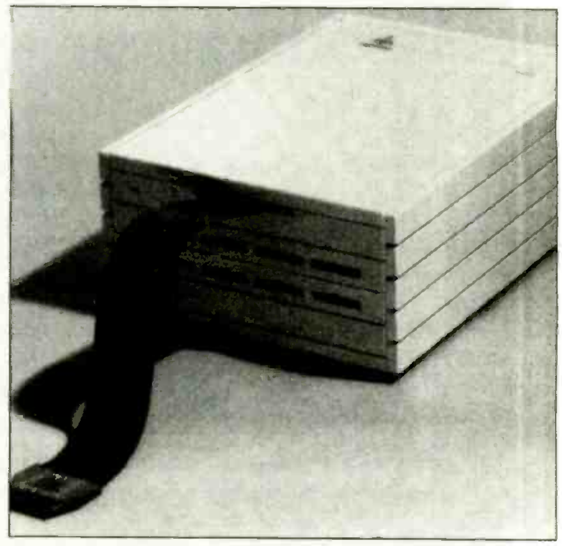
Written in Borland C++ the system software offers mouse driven source view windows (currently four), hex view, micro-watch, data watch and trace. There is also a target memory mapping facility where pods can be mapped as either program or data. ■

System for all seasons

The Trace 32 system marketed by Noral Micrologic and produced by German manufacturer Lauterbach Datentechnik was designed to provide 'a truly integrated, universal development system'. With its emphasis on modularity, it is an example of one of the approaches being followed by developers.

The system can be connected to any workstation and can contain any combination of instruments needed for development (eg, ICE, logic state analyser, simulator, etc). It is based on an active controller with a 32-bit CPU and up to 16Mbytes of ram into which most of the system software is downloaded. The controller performs most of the work.

Although it looks like yet another box, the Trace 32 offers a high degree of flexibility and integration.



SEMICONDUCTOR LASERS SHINE OUT

Microchips packed with lasers operating at different wavelengths will improve optical computing and communications, although no one is quite sure how. By Jeff Hecht

Imagine packing 140 lasers onto a single piece of semiconductor measuring about 2mm by 7mm. Imagine that each laser emits light at a different wavelength, making the surface a bright array of different-coloured spots. Imagine, moreover, that the lasers are made so precisely that each has a wavelength 0.3nm longer than that of its neighbour on one side, and 0.3nm shorter than that of its neighbour on the other. Such an array would have been a mere flight of fancy just a few years ago, but not any more. Communications specialists can now demonstrate one that does nearly all these things – the only exception being that the light it produces is in the infrared range and so is not visible.

Dramatic advances are being made in laser technology, just as electronics was making giant strides a generation ago. Once again, the latest developments exploit improvements in semiconductor technology. To pack a single chip with so many lasers, researchers are now taking advantage of manufacturing processes, such as molecular beam epitaxy and ion beam implantation, that can create layers only a few atoms thick and were developed to produce complex integrated circuits. These techniques also enable single lasers to emit more light at shorter wavelengths than before, giving technologists finer tools to record, transmit and read more information.

Jeff Hecht is Boston correspondent of the *New Scientist* in which this article was first published.

The developers of the laser array see their work helping to improve the performance of high-speed optical communications. Because the lasers will operate at many different wavelengths, the array will be able to transmit data at impressive rates through optical fibres or through free space between computer boards or chips. Similar arrays could be used in optical computers that will process data as beams of light instead of pulses of electric current. Meanwhile, single lasers emitting light at visible (shorter) wavelengths could improve the performance of laser printers by enabling the machines to write faster, and increase the density of data storage on optical disks. But the greatest possibilities may be still to come, sparked off by the new capabilities of the emerging technology.

The earliest electronic devices relied on thermionic valves, and the next generation on individual transistors. Today, thousands or even millions of transistors are crammed onto the integrated circuits that are carried on a single semiconductor chip. Laser technology is still in its early stage. Gas lasers, the counterpart of valves, are still widely used, even when onlymW of light are required. Though engineers prefer solid-state technology, almost for its own sake, the older technology often remains cheaper.

Building up the power

The most familiar gas laser is the helium-neon type, a tube about the size of a rolled-up magazine that emits red light – a wavelength that has only recently become available from semiconductor lasers. The counterpart of the single transistor is the semiconductor laser with just one light-emitting element on a chip of gallium arsenide. It was invented nearly 30 years ago in the US and is now the most common laser, usually buried deep in the bowels of equipment like fibre-optic communication networks, laser printers and CD players. However, it operates only at low powers in the infrared range. Larger lasers exist solely in research laboratories and industrial plants; they consist of gas-filled tubes powered by an electric discharge, or crystalline rods powered by bursts of light from powerful flash lamps.

The latest semiconductor lasers may make much of the older technology obsolete – just as solid-state devices took over from transistors and low-power valves – but they will not replace high-power lasers. For the time being, semiconductor lasers are being seriously considered for applications that require no more than about 10 watts of steady light. Semiconductor electronics developed the same way, starting out at low powers and working up to higher levels.

Like integrated electronics, arrays of many lasers on a single substrate have important advantages. In some arrays each laser can be turned on and off independently. At Bell Communications Research (Bellcore) in Red Bank, New Jersey, scientists can do this up to 5 billion times a second in one two-by-eight array, generating a staggering 80 billion pulses per second. Other arrays combine the outputs of many laser elements to generate a single beam of higher power, which in some cases can be steered electronically, with no moving parts. Researchers have barely started to explore the possibilities.

New ways of making small, thin structures of semiconductor materials are improving the efficiency of single lasers. The smaller structures confine the flow of current and light in the semiconductor more precisely. This makes the devices more efficient at converting electrical energy into light, and they last longer, too. These are critical improvements because, until recently, the lifetime and output powers of some semiconductor lasers have been disappointing, especially for devices generating short-wavelength light.

The choice of materials is crucial. Many semiconducting materials, notably silicon, normally do not emit light. The first generation of commercial lasers was made of gallium arsenide, which emits light only in the invisible infrared region. Only in the mid-1980s did semiconductor lasers producing red light go on sale, and they have acquired important uses in displays and bar-code scanners. They are made from indium gallium phosphide and yield less power than gallium arsenide lasers. In July 1991, the 3M company, an American maker of plastics and electronics, announced that it had developed the first semiconductor laser to emit green light, which has a shorter wavelength than red and infrared light, and so can be focused to a much smaller spot. This is an important development because it allows more data to be stored.

The latest semiconductor lasers rely on the same phenomenon as the pioneering devices developed in the early 1960s. The semiconductor must contain two regions "doped" with different types of impurities. In one region, the impurities add extra electrons to the crystal; in the other, they create electron vacancies, or "holes". Applying a voltage across the material, so that the negative terminal is attached to the material with extra electrons, known as n-type, and the positive terminal to the material with holes, known as p-type, causes electrons and holes to move towards the junction of the two regions. Some electrons fill holes at the junction, causing current to flow. Because these devices have two electrical terminals, they are known as "diodes".

The union of an electron and a hole releases the energy that had kept the electron free to move about the crystal lattice. In many traditional semiconductors, such as silicon and germanium, virtually all this energy is released in the form of heat; with the newer generation of semiconductor compounds, such as gallium arsenide and indium phosphide, some of the energy takes the form of light. This is the basis of both light-emitting diodes (LEDs), which are often used as indicator lights or displays, and semiconductor diode lasers.

LEDs emit light in all directions as the electrons release energy spontaneously at the junction. Lasers are more complex. The edges of the semiconductor wafer are made to reflect some light back along the junction plane, stimulating other electrons to release light energy at the same wavelength. The process amplifies the intensity of the light and also directs it along the line between these two mirror edges (one or both of which let some light escape). In practice, semiconductors operate as lasers only when the current passing through the junction exceeds a threshold value; at lower currents, they operate as LEDs.

The first semiconductor lasers, which were made of two thick layers of gallium arsenide with different dopings, produced so much more heat than light that they burned out unless cooled to well below room temperature. The key to improving them has been to confine more of the current and the reflected light in the junction plane (see **Figure 1a**). Engineers found they could do this by sandwiching the junction plane between two thin layers in which aluminium replaced some of the gallium (see **Figure 1b**). Using the same principle, they could confine light to a narrow strip within the junction plane (see **Figure 1c**).Tighter confinement of current flow and light improves efficiency by concentrating energy. The higher the current, the more electrons are available to emit light. The more light is concentrated, the more efficiently it extracts energy from the electrons. By 1980, this technology had produced inexpensive commercial lasers that could operate continuously for many thousands of hours. The layers were 0.1 to 1μm thick, and the strips in the junction plane 1 to 10μm wide. Then developers began to investigate what might happen if structures were made even smaller.

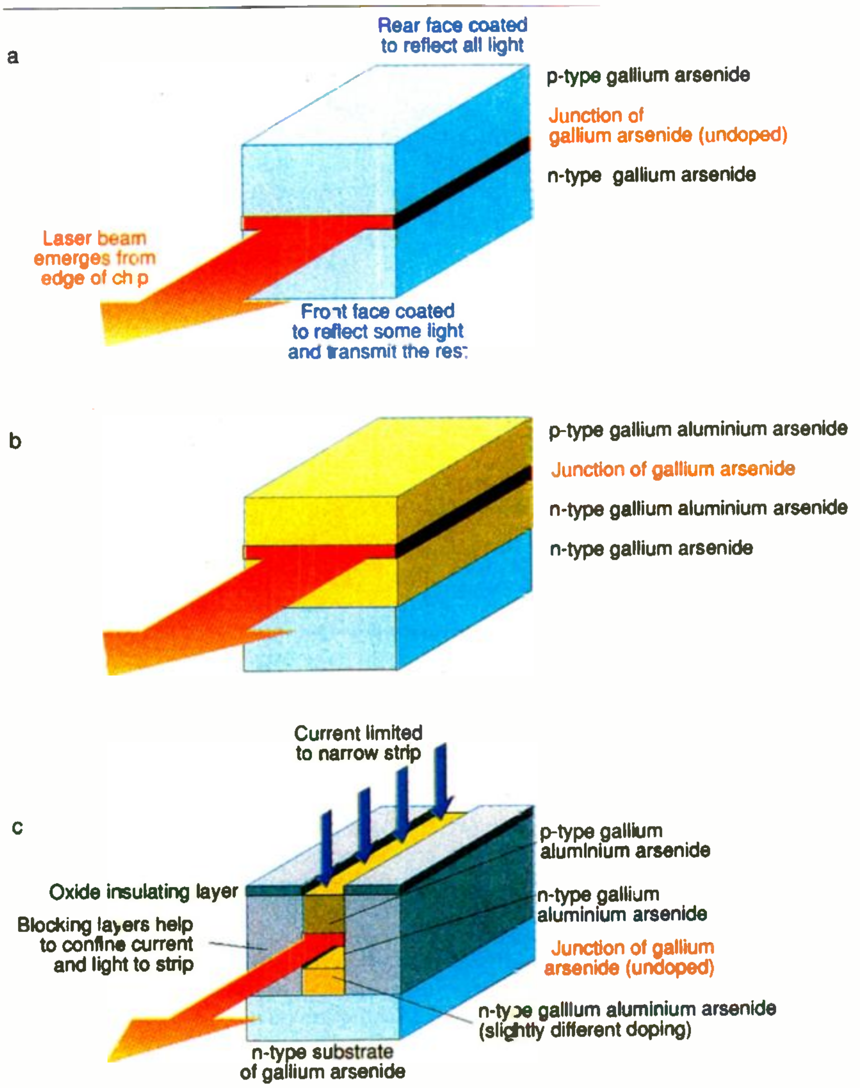


Fig. 1. From the earliest (a) to the latest (c) design of semiconductor lasers, engineers have tried to confine and control the light these devices emit to boost efficiency

A micrometre is tiny on a human scale but large on an atomic one; a 1-micrometre layer is a "bulk" material, in which electrons can have a broad range of energies. That ceases to be the case when dimensions shrink below about 0.02μm (20nm), or about 35 atomic layers in gallium arsenide. In such minute structures, the laws of quantum mechanics limit electrons and holes to certain energy levels. This changes the material's properties, which depend on electron energies. The development of molecular beam epitaxy, which deposits atoms layer by layer, opened the possibility of making such thin layers.

The first step was to make a "quantum well", a layer typically several nm thick, sandwiched between two thicker layers of a slightly different composition. The compositions are

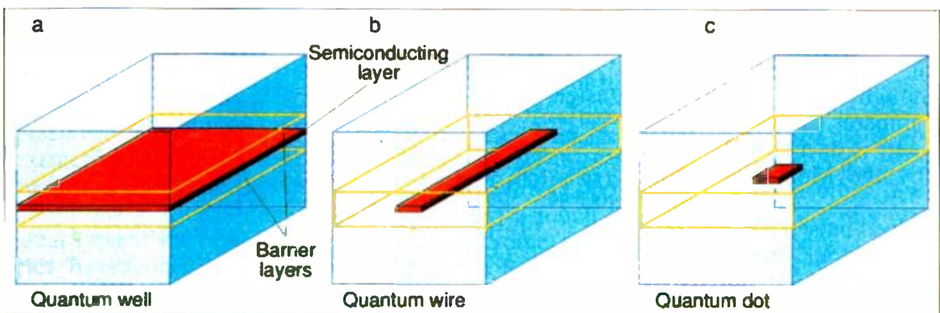


Fig 2 Three steps closer to smaller semiconductor lasers: these structures exploit the laws of quantum mechanics to give more precise control over the properties of a semiconducting material such as indium gallium arsenide

chosen so that electrons that carry current have slightly less energy in the quantum well layer than in the barrier layers that sandwich it. Electrons become trapped in the quantum well if they lack the energy to return to the barrier layers. The barrier layers must be thicker than the quantum wells to keep the electrons from escaping by the quantum-mechanical phenomenon of tunnelling, which lets electrons through thin regions into which they would otherwise have too little energy to pass.

The quantum well is formed at the junction between n-type and p-type semiconductor materials where it can trap electrons and holes, which can be stimulated to release laser light. Stacks of quantum wells can be made by alternating quantum well and barrier layers. This increases the number of trapped electrons and holes, without reducing their density.

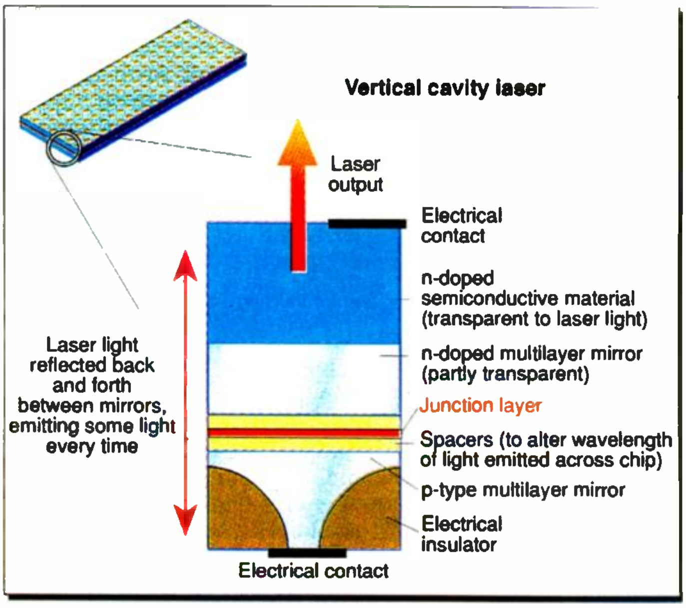


Fig 3 Vertical cavity layers emit light through the surface of the semiconductor device rather than from its edge

Constance Chang-Hasnain's team at Bellcore used three quantum wells in the 140-element laser arrays described at the beginning of this article. The barrier layers were gallium arsenide, while indium was substituted for 20 per cent of the gallium in the quantum wells, giving a compound with composition $In_{0.2}Ga_{0.8}As$. Because quantum wells confine electrons and holes more closely than bulk semiconductors, they make more efficient lasers. They are already used in some commercial semiconductor lasers.

Quantum jump?

Quantum wells confine electrons in only one dimension in a thin layer. The next logical step is quantum wires, which confine electrons in two dimensions – in a narrow strip that is part of a thin layer (see Figure 2). Developers hope that the tighter confinement in quantum wire structures will improve laser performance, further reducing the threshold for laser action and the waste energy that must be dissipated as heat. They even speak of the potential for "quantum dots", which would confine electrons in three dimensions – length, width and height.

The tighter confinement should reduce the current needed to reach the threshold for laser action. Quantum well lasers have threshold currents of about a milliampere, but thresholds could be about a microampere for quantum wire lasers and about a nanoampere for quantum dots, says Eli Kapon of Bellcore, a pioneer in the field. Though the smaller lasers are likely to be more efficient, their individual outputs will be low. The quantum well elements in the Bellcore array emit only about 2mW of light, while the output of quantum wire elements will be of the order of microwatts. Kapon says "that's fine for the kind of applications people are envisioning", such as closely packed arrays of many tiny lasers, each one emitting only a little light, for optical computing or communications.

The first quantum wire lasers were made two years ago, but so far the lowest threshold, obtained by Kapon's group, is only 0.6mA. The major problem is that thin layers are much easier to make than the thin lines needed for the second dimension of confinement. Using present technology, layers a fewnm thick can be formed fairly easily. "There is no equivalent fabrication technique that will give you automatically quantum wires and dots," says Kapon. Quantum confinement requires lines no more than 20nm wide. Though a scanning tunnelling microscope can work at this scale, it cannot do so easily enough for mass production. No one has yet reported operating a quantum dot laser.

Theorists predicted quantum wells, but they were surprised by another property of very thin semiconductor layers—that they tolerate much more internal strain than thicker layers. Researchers discovered this when they were trying to make new types of semiconductor lasers. For a long time, strain was believed to be bad for lasers: "Now it turns out that it seems to be making everything better and better, as long as you don't have too much," says Anders Olsson, head of the solid-state and quantum optics research department at AT&T Bell Laboratories in Murray Hill, New Jersey.

The initial impetus for making strained layers came from problems in growing the semiconductors used in lasers. The thin laser layers are deposited on a thicker substrate of either gallium arsenide or indium phosphide. (More complex compounds of three or four elements are too difficult to grow in large wafers.) Good-quality layers can be formed only if they have atomic spacing close to that of the substrate. If the atomic spacings differ too much, the strain causes the layers to develop flaws – imperfections in the crystal where atoms are missing or extra atoms are inserted – that are likely to make the laser fail. Atomic spacing depends on composition, so this limits the choice of compounds for semiconductor lasers.

For layers a micrometre thick, the restriction is quite stringent – atomic spacing must be within 0.1 per cent. Fortunately, substituting aluminium for more than half of the gallium in gallium arsenide changes atomic spacing by less than that, so lasers of those materials were the first type developed commercially; they emit light at 0.75 to 0.9µm in the near infrared range. However, to match a lattice to indium phosphide – for longer-wavelength lasers – requires the proportions of the four elements indium, gallium, arsenic and phosphorus, to be balanced. Manufacturers do this to produce lasers emitting light at 1.3 and 1.55µm, the infrared wavelengths used in fibre-optic communication systems. These lasers cost much more than gallium arsenide lasers.

In the past few years, developers have grown indium-gallium phosphide on gallium arsenide substrates for lasers at 0.67µm in the red range, and have produced shorter red wavelengths by replacing some of the gallium with aluminium. However, the power output of these devices was low, especially at the shorter wavelengths, and 0.63µm appeared to be a limit for practical lasers. This left shorter wavelengths out of reach, and a gap at 0.9 to 1.1µm in the infrared.

The lattice-matching constraint can be eased to 1 per cent for layers no more than about 10nm thick, which can accommodate much more internal strain than thicker layers. This gives developers a much broader choice of materials, and hence of wavelengths, as they in turn depend on the material. For example, thin strained layers of gallium arsenide with 20 per cent of the gallium replaced by indium – a compound with atomic spacing of 0.570nm – can be grown on gallium arsenide, with 0.566nm spacing. Bellcore used this approach in its 140-element array, which emits light at 940 to 983nm, wavelengths not available from purely lattice-matched lasers.

Developers have also turned to strained layers for visible semiconductor lasers, which suffer from high threshold currents, limited output power and short lifetime. At McDonnell Douglas Electronic Systems, based in Elmsford, New York, lasers with a 7-nanometre strained quantum well of gallium-indium phosphide produced a steady beam up to a record 475mW at 665 to 670nm at room temperature. The current density needed to reach laser threshold was below 375A/sq cm, the lowest ever recorded at that wavelength. In October, Spectra Diode Laboratories, a manufacturer based in San José, California, announced it had made a 1W laser at 680nm. Though the company claims a threshold current density of 350 amperes per square centimetre for the device, the light generated is at a slightly longer wavelength than that produced by the McDonnell Douglas device.

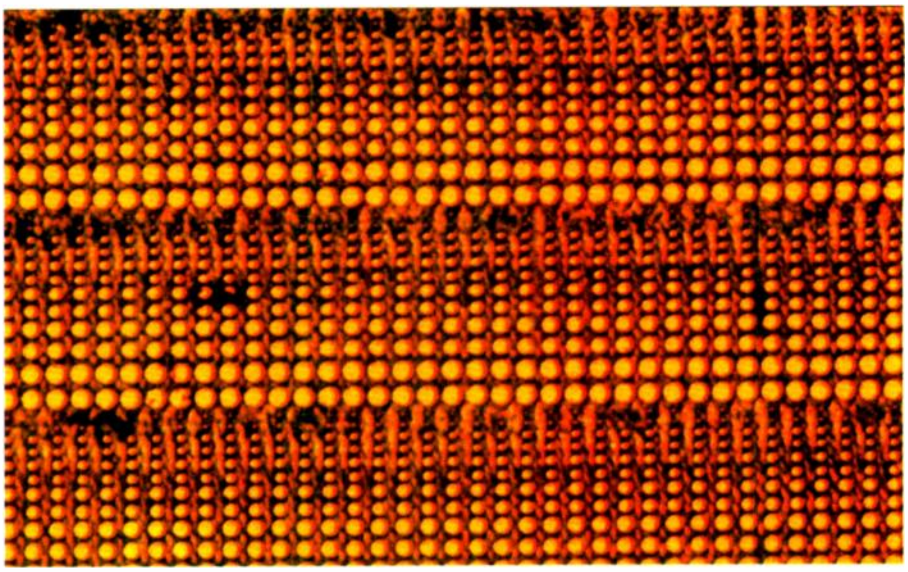
At Bellcore, Chang-Hasnain's group produced pulses of 650mW at 634nm from a laser with four strained quantum wells of indium-gallium phosphide, 2.5nm thick. In between were 4nm barrier layers that also contained aluminium. Because the threshold current density was much higher, 1700A/sq cm, the laser could not produce the high power in a steady beam. However, the shorter wavelength is better for optical data storage and recording, and much brighter to the human eye, making it more attractive for displays.

Strained layers were also used in the most dramatic semiconductor laser breakthrough of 1991 – operation of the first green diode laser at room temperature. Several laboratories had already produced shorter-wavelength blue light by passing the infrared light from gallium arsenide lasers through materials that double its frequency. However, that approach is inefficient, as it converts only a small part of the infrared energy to visible light. A team at the 3M Corporate Research Center, based at St Paul, Minnesota, produced light with a wavelength of 525nm, the shortest ever from a diode laser, by making the semiconductor from doped zinc selenide for the first time. The team's quantum well laser included strained layers to accommodate differences in atomic spacing within the material. Developers hope the work will point the way to new lasers emitting visible light, and a new generation of laser applications.

Integrated mirrors

Another innovation marks an even greater change in thinking about semiconductor lasers than strained layers. For three decades, semiconductor lasers have been designed to emit light from their edges. That is fine for individual lasers, or for linear arrays of many laser elements, but something different is needed if many lasers are to be integrated in a two-dimensional array on a single chip.

The simplest approach is to redirect the laser light. In a standard semiconductor laser, the reflective surfaces at the ends of the chip reflect light back and forth horizontally along the junction plane so that the beam emerges from the side of the wafer. The light can be redirected by etching a mirror tilted at 45 degrees to the wafer surface. A more sophisticated alternative is to etch tiny grooves in the bottom of the junction layer. These grooves scatter or diffract some light at a steep angle, which can then pass through the barrier layers. The latter design can be modified to steer the combined beam from many laser elements over a small angle without moving parts. This isdone by shifting the phases of the separate currents supplying each of the elements.



Turn on the light: tiny chips that emit visible light could herald a new generation of laser applications

Both approaches have been demonstrated, and both have their limitations. Because they emit light from only a small portion of the surface, the number of laser elements that can be squeezed onto a given area of a wafer is limited. Tilted mirrors are difficult to etch, while the scattering grooves occupy much more space on the wafer than the laser elements.

A more direct approach is to place the reflective surfaces above and below the junction layer, so that the laser beam is generated perpendicular to the junction layer and wafer, instead of along it (see **Figure 3**). The beam from such a vertical-cavity laser does not have to be redirected to emerge from the wafer surface. The emitting areas can be circular, so generate a symmetrical beam diverging by 3 to 6°, instead of the 10 to 30° of edge-emitting lasers, which require corrective optics.

Vertical-cavity lasers can also be packed more closely on the wafer than other surface emitters. Typically, each laser element is 10 to 20µm across, and about 300µm from its neighbours. The laser elements can be made smaller and pressed much closer if the material between them is etched away. Two years ago at Bell Laboratories, Jack Jewell etched lasers as small as 1.5µm across, spaced only a few µm apart, thus packing more than a million lasers onto an area of only $0.5cm^2$.

Heat still a problem

There is a trade-off for this high packing density. The power a laser can generate depends partly on the volume of material from which it extracts light energy. If the laser mirrors are placed in the junction plane, as in an edge-emitter, they can extract light from an area of the junction a few µm wide and a few hundred µm long. Mirrors above and below the junction plane extract light from a much smaller area, typically a fewµm to 20µm in diameter. In practice, this means that the mirrors on vertical-cavity lasers must reflect light much more efficiently than those on edge emitters.

Powers available from vertical-cavity lasers smaller than about 20µm will remain low – the highest so far is about 3mW per laser element. Larry Coldren of the University of California at Santa Barbara says that 10mW may be possible eventually, which should be sufficient for the major expected uses of vertical-cavity lasers, in arrays for high-speed communications or computing. However, Coldren warns of problems, including removing waste heat from such a small volume, making current flow easily and uniformly through the structure, and confining light and electrons for efficient laser operation.

The vertical-cavity design, such as the 140-element array, also allows the laser elements on one chip to generate different wavelengths. While semiconductor composition determines the range of possible wavelengths, the exact wavelength is set by the distance between the laser mirrors. The round-trip distance must be an exact multiple of the wavelength, so changing the spacing changes the wavelength or the multiple. If the mirrors are hundreds of wavelengths apart – as they are in edge-emitting lasers – the multiple changes. However, if the mirrors are only a few wavelengths apart, the wavelength changes predictably (the longer the spacing, the longer the wavelength). To produce its multi-wavelength array, Bellcore varied the thickness of one layer diagonally across the wafer, so that it was thickest at one corner and thinnest at the opposite corner. It was this smooth variation in mirror spacing that provided the 0.3nm shift in wavelength between adjacent laser elements.

Advanced semiconductor laser technology will not instantly consign other lasers to the scrap heap, any more than integrated circuits wiped out valves and discrete transistors in a day. Semiconductor lasers are far from matching the highest powers or shortest wavelengths available from other lasers. What they will do is help to create new applications for lasers, in areas such as optical computing and communications. They are likely to bring surprises, and perhaps even new uses for older laser technology, just as integrated circuits created new roles for electronics. ■

Circuits, Systems & Standards

First published in the US magazine EDN and edited here by Ian Hickman.

Enhance the A-to-D process

As a predominantly analogue engineer, I found this a useful primer on an important topic on the borderland between the analogue and digital worlds. The linear nature of predictive encoding is ideal where the next stage is DSP, in contrast to non-linear coding schemes such as the mu-law (US) and A-law (European) companding methods used in telephone networks.

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Predictive coding improves A-to-D performance

By shortening digital-word size without sacrificing digital resolution or accuracy, both predictive coding and companding can improve the efficiencies of analogue-to-digital conversions. While differing in their basic concepts (see "Companding vs predictive coding"), both methods find use in applications requiring greater conversion accuracy than is available with current low-cost A-to-Ds. By using predictive coding or companding, a system's dynamic range can be extended so that shorter data words maintain the signal-to-quantising noise ratio of longer ones.

Highly correlated signals are redundant

Predictive coding achieves its constant accuracy with reduced word size (or increased accuracy with the same word size) because a sampled signal is usually full of redundant information. Any sample can be closely estimated by examining its neighbours; the samples of a redundant signal correlate with one another, and the accuracy of any prediction rises with the degree of correlation.

Specifically, highly correlated sequences, such as voice or music waveforms, contain much redundant information and exhibit relatively small sample-to-sample variations.

By contrast, wild fluctuations in sample values for sequences exhibiting little correlation can be expected. Each sample at the output of a digital noise generator, for example, is statistically independent of its neighbours. For such a signal, predictability is out of the question.

Correlated time sequences have frequency spectra with energies that are concentrated at the low end and fall off at the high end of the frequency range. In the example shown in **Fig. 1**, signal energy is concentrated well below half the sampling frequency (f_s). A standard linear A-to-D converts signals in its full input range at frequencies up to $f_s/2$ (the Nyquist frequency); the A-to-D thus has full-power bandwidth up to $f_s/2$.

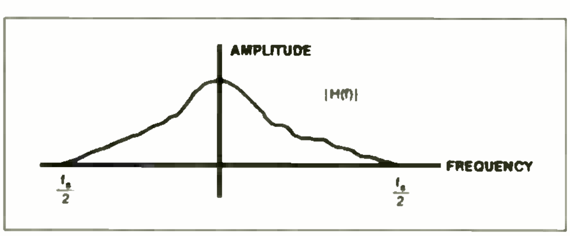


Fig. 1. Most real-world signals concentrate their energy at frequencies far below the Nyquist frequency (half the sampling frequency)

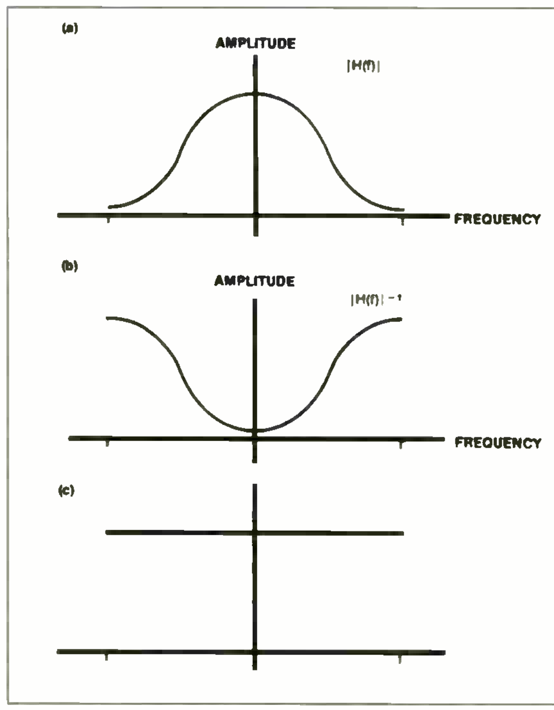


Fig. 2. To get the most out of an A-to-D's dynamic range, the input spectrum (a) should be modified by a filter with the inverse spectrum (b), so the resulting A-to-D input is flat (c).

In the conversion of most naturally generated signals, this capability is not necessary. To obtain maximum signal-to-noise efficiency from the A-to-D, input signals should have a flat frequency spectrum so that the total A-to-D input range is always used up in the conversions (**Fig. 2**).

You can re-examine this principle by remembering that most of a natural signal's energy occurs far below the Nyquist frequency. Thus, for low-frequency components

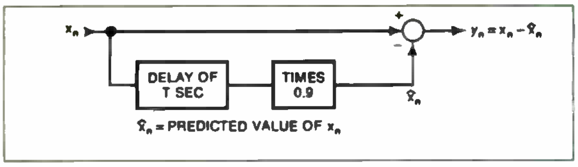


Fig. 3. This predictor uses nine-tenths of the present sample as its estimate of the next one.

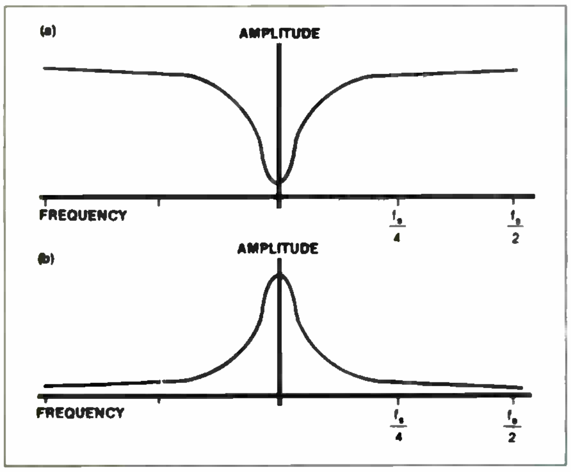


Fig. 4. The predictor shown in Fig. 3 has a frequency response close to that depicted in (a). It closely resembles the inverse of the spectrum (b) and is typical of many real-world phenomena, such as voice and video images.

of the input signal, many more samples than necessary are made.

For example, for a signal band-limited to 16kHz, sampling at 32kHz avoids aliasing and provides two samples per cycle of 1kHz input – the minimum number necessary.

But for a 1kHz input 32 samples are taken per input cycle – 16 times the number needed. If the signal is digitised, the quantisation accuracy per sample for a given word length can be increased by "stealing" information contained in previous samples.

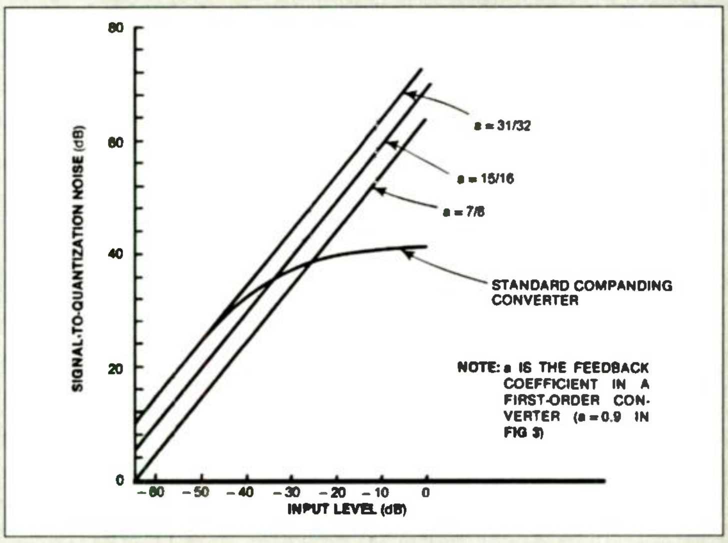
Removing redundancy

The key to predictive coding is removal of signal redundancy – a process termed decorrelation. For example, a simple first-order predictor might use nine-tenths of the most recent sample as its estimate for the next sample. The predictor output would then be the difference between the prediction and the actual signal value (**Fig. 3**). This simple predictor has a frequency response computable by means of Fourier techniques. Because its output has contributions from its input and a delayed version of that input, the network function is

$$Y(j\omega)/X(j\omega)=1-0.9^{-j\omega T}$$

The predictor has a frequency response similar to H^{-1} depicted in **Fig. 4a**. For a 100%-efficient predictor, the input spectrum would have to be the inverse of the predictor's frequency response (**Fig. 4b**).

Obviously, real-world data exhibits time-varying frequency spectra. But for most data sources, the worst-case frequency spectrum can be predicted; a predictor designed with that spectrum in mind offers a substantial fidelity enhancement. The system shown in Fig. 3, for example, closely matches the spectrum of Fig.4b – a spectrum fairly typical of many real-world data sources such as audio and image signals.



Companding vs predictive coding

Essentially a bipolar logarithmic technique, companding can reduce rates in digital data transmission systems. With a compander, an analogue input signal receives very high-resolution quantisation around 0V.

As the input amplitude to a companding A-to-D converter increases, the quantisation becomes more coarse. This process maintains a relatively constant signal-to-noise ratio over a wide input-signal range. Companding proves quite effective in servo systems, where nulling accuracy is important.

Both companding and predictive coding are well suited to voice-signal digitisation. The choice between the techniques generally hinges on the ultimate application. For example, predictive coding proves ideal for digital filtering or FFT applications because predictor output codes are linear; the logarithmic nature of a compander's output code precludes simple interfacing with digital signal-processing software or hardware.

On the other hand, companding is simpler to realise in LSI circuitry as the many recently introduced inexpensive codec chips demonstrate.

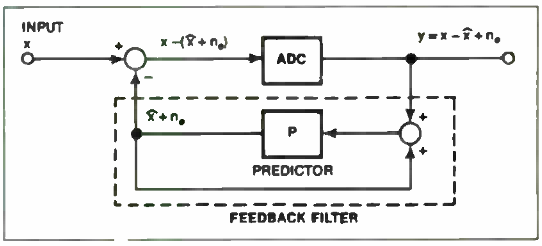


Fig. 5. A predictive encoder embeds a predictor in the A-to-D feedback loop. Here x is the input signal, x^ the predicted value for x, n_o the A-to-D noise and n_e the error-signal noise.

Predictive coding transmits differences

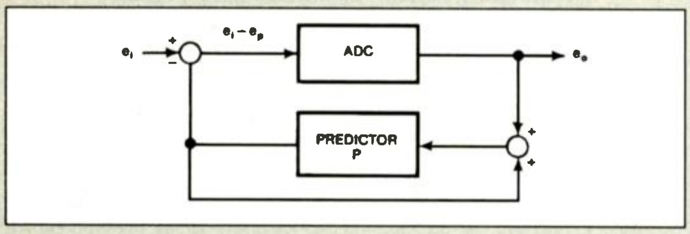
Quantisation error in a predictive encoder won't exceed ±0.5LSB because the loop continually monitors the conversion process and adjusts the present-sample value to compensate for the previous one.

To understand the process examine the figure – a relabelled version of Fig. 5. Here linear analysis is possible because the quantisation is small. The feedback voltage (e_p) is expressed in terms of the input voltage via two loop equations:

$e_0=e_i-e_p$

$[P/(1-P]e_0=e_p$

Combining these equations yields $e_p=Pe_i$. In other words e_p is the predicted value of the input. If the predictor is performing well, e_p will be a good estimate of e_i, and the A-to-D will quantise the difference e_i-e_p. If the A-to-D's total input range is devoted to quantising e_i-e_p, the relative digital error for e_i will decrease when the signal is reconverted to the analogue domain.



The A-to-D converter in this encoder converts the difference between the prediction and the input signal.

Using predictors in encoders and decoders

To construct a predictive encoder, embed a predictor in an A-to-D feedback loop (**Fig. 5**). Incorporating an inverse predictor in the D-to-A circuit will maintain a flat output frequency response when the digital data is finally recovered. Such encoding and decoding circuits offer enhancement of conversion accuracies.

The transfer function of the circuit depicted in Fig. 5 is y/x=I-P; the standard transfer function for a predictive encoder.

In the encoder, the A-to-D output produces an error with each conversion because of its inherent resolution limit. For reasonable input levels, the errors prove independent of the signal output y and are also limited by feedback to values within ±0.5LSB of the true value of y.

Conversion errors are generally unrelated to each other and produce a random, uncorrelated sequence. The frequency power spectrum is very flat and its total power is $\sigma^2= q^2/12$, where σ is the value of an A-to-D LSB in volts.

Because the quantisation errors are uncorrelated, the ADC output can be modelled as the input signal plus noise ($v+n_o$). The noise term (n_e) appearing at the ADC input results from the output noise (n_o) passing through the feedback filter containing the predictor. Because of this feedback, both n_o and n_e are constrained to a value within ±0.5LSB of the theoretical value of the signal (s-s^). Feedback corrects any tendency of signals to cross this boundary. For another, possibly simpler, explanation of the process, see "Predictive coding transmits differences".

A matching decoder (**Fig. 6**) has the transfer function r/y=1/(1-P), where r is the reconstructed output. The complete encoding/decoding process thus yields the original input signal: r/x=(1-P)/(1-P)=I. The noise of the reconstructed output is about 3dB greater than that generated by the A-to-D itself. However, the decoder has signal gain, so the total system signal-to-noise ratio is improved over the A-to-D's inherent accuracy.

The example predictor shown in Fig 3 works well in both encoding and decoding circuits (**Fig. 7**). The decoder gain formula in this case is

$r/y = 1/(1-0.9^{-j\omega T})$.

For low frequencies (below 300Hz), gain is approximately ten, corresponding to more than three additional bits of A-to-D accuracy; an 8-bit converter used in Fig. 7 thus achieves the accuracy of an 11-bit converter.

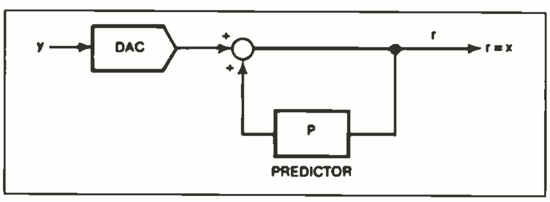


Fig. 6. A predictive decoder unfolds the encoder loop.

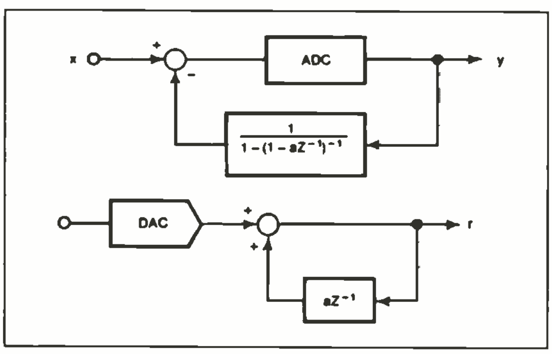


Fig. 7. The predictor depicted in Fig. 3 can serve in both encoder and decoder circuits. Here a equals the feedback coefficient (0.9 in Fig. 3.) and Z^{-1} denotes a delay of one sample time.

Example circuit

Putting all this theory together produces the circuits shown in **Fig. 8**, broken down into components in **Fig. 9**. Illustrating the basic principles of predictive coding, this 8-bit system breaks neatly into several sub-circuits.

A-to-D converter (Fig 9a): the A-to-D used in the system is a typical successive-approximation type, using an 8-bit, fast-settling D-to-A converter, an SAR and a comparator. The sample/hold circuit that usually precedes the A-to-D's comparator can be incorporated in a tailoring filter used to adjust frequency response and full-power bandwidth, as in the case here.

*Predictor (Fig 9a):*the predictive filter and sampler 2 form the heart of the scheme. The sample/hold circuit is timed to sample the D-to-A output after the converter has completed the successive approximation routine and has settled to the comparator voltage (e_c±0.5LSB). The connected filter is essentially a linear hold circuit and low-pass-filter combination. The sampled input turns the combination into a "leaky" discrete accumulator.

A little bit of mathematics reveals that a portion of the present sample of e_i is used as the estimate of the next value of e_i. This single-sample estimation process is termed a first-order prediction. A predictive filter used to perform the estimation is usually a very low-bandwidth discrete filter or integrator with a time constant much greater than the sampling interval.

Decoder (Fig 9b): in the decoder, holding registers store a digital word, which is converted by an 8-bit D-to-A into an analogue voltage. To avoid D-to-A bit-transition glitches, a sample/hold circuit samples the D-to-A output and feeds it into the accumulator sub-circuit.

Accumulator (Fig 9b): essentially the low-pass filter

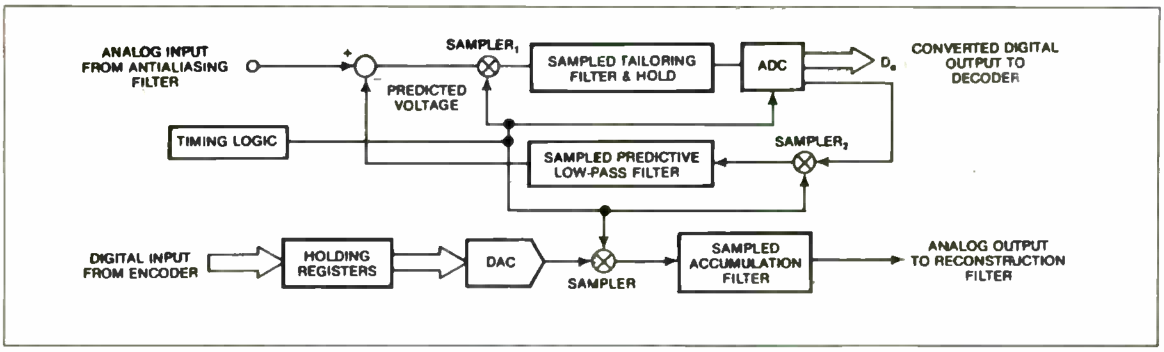
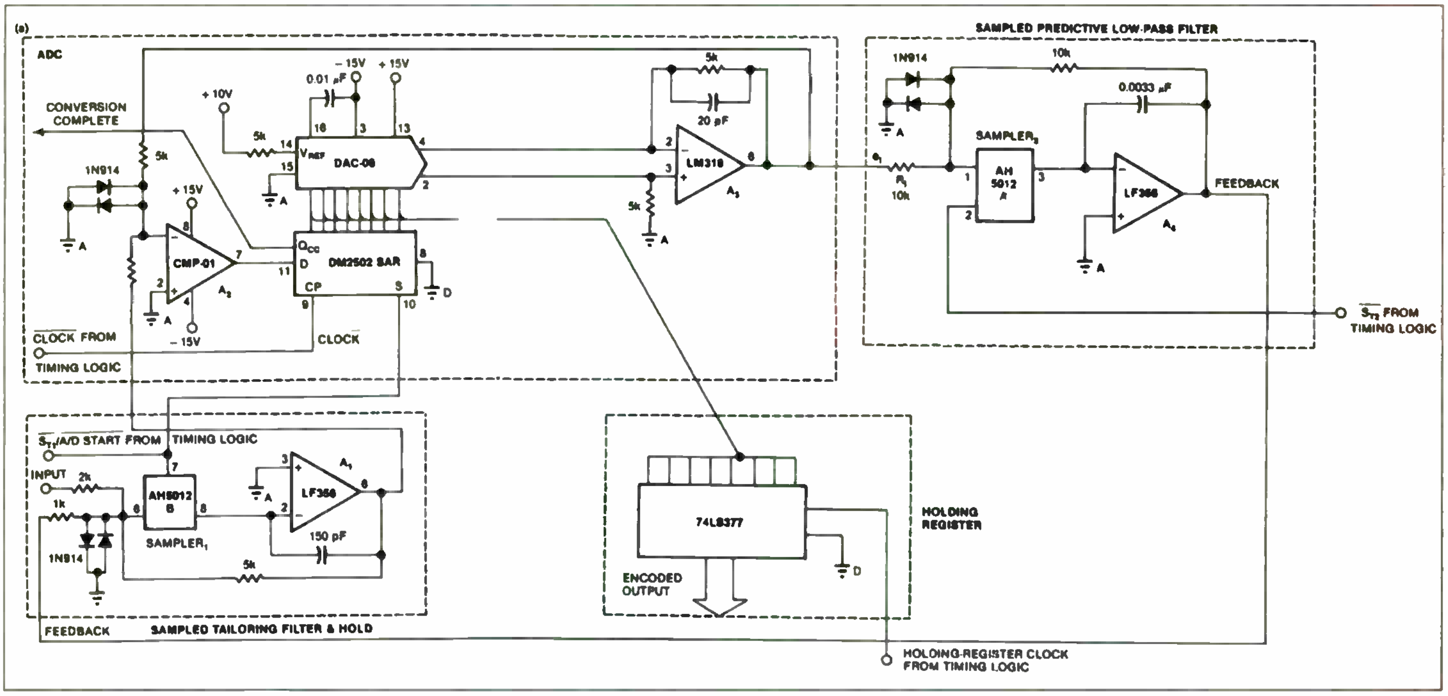


Fig. 8. A predictive/decoder system divides into sub-systems, shown in detail in Fig. 9.



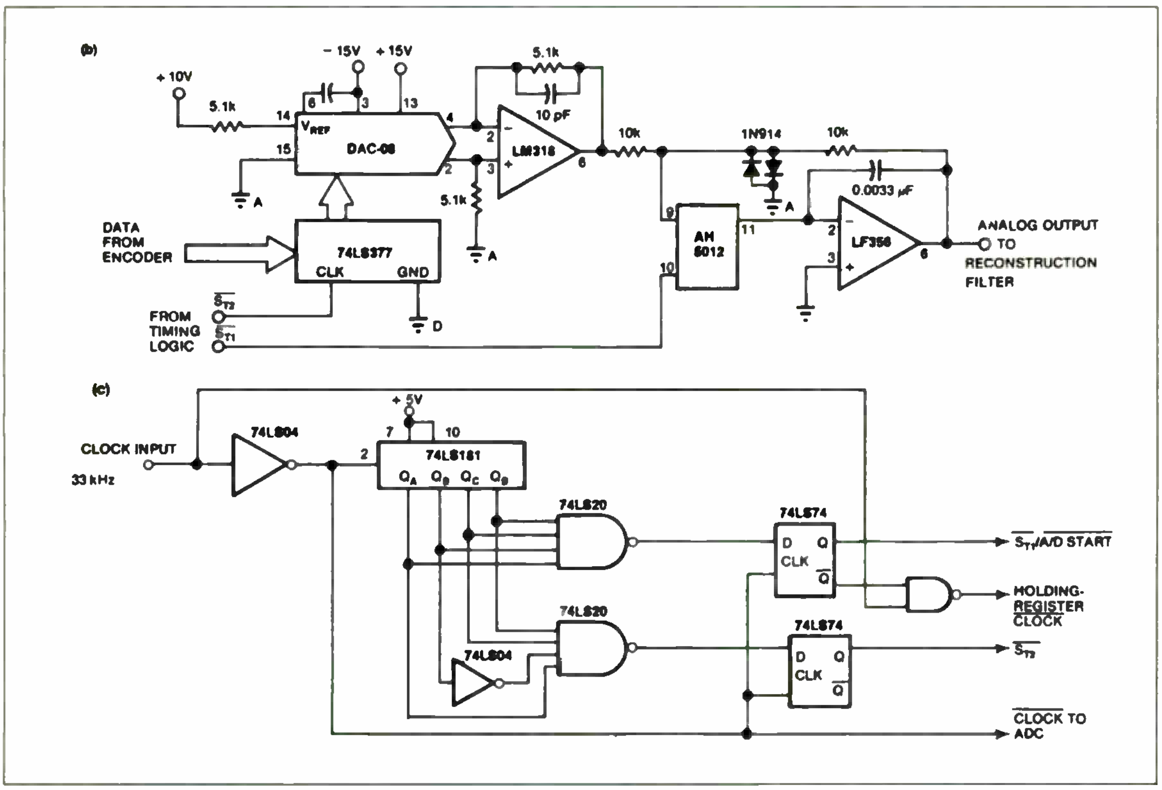
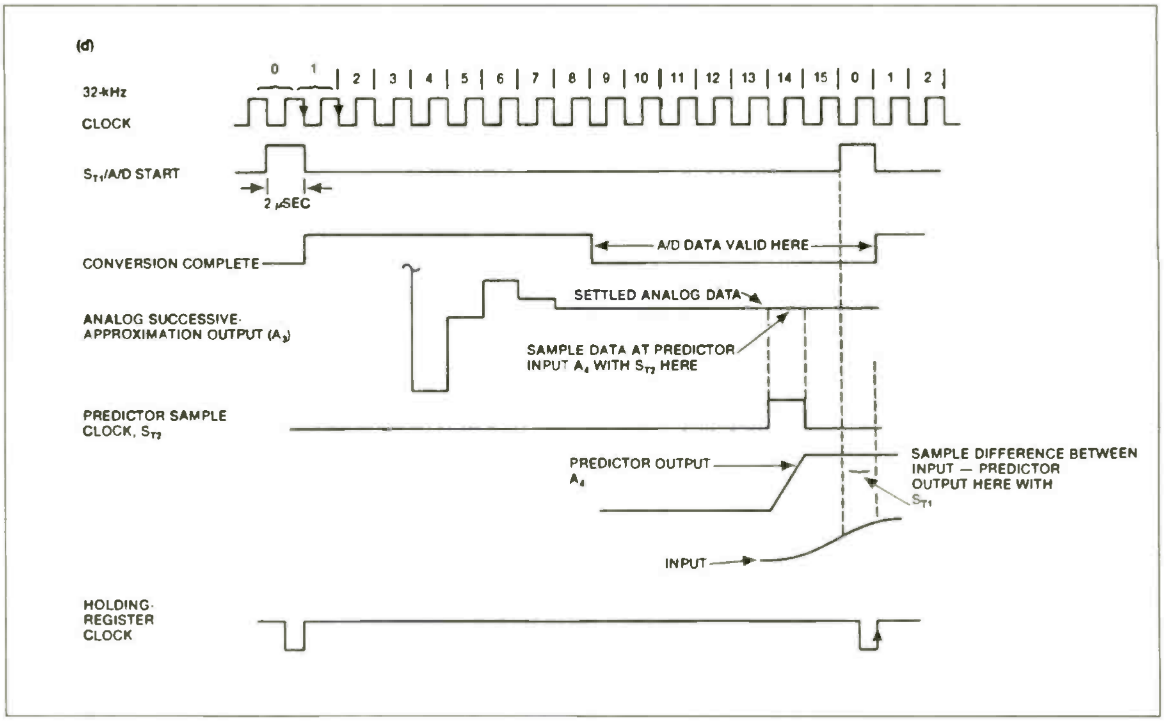
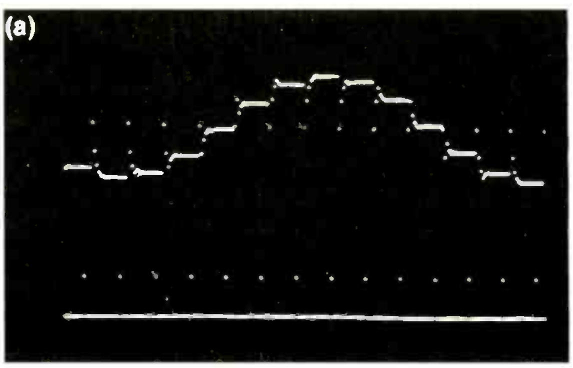


Fig. 9. An encoder (a) uses hybrid sample/hold units, as does a decoder (b). Timing logic (c) controls both sub-systems with signals sequenced as in (d). (Fig.9d is overleaf.)





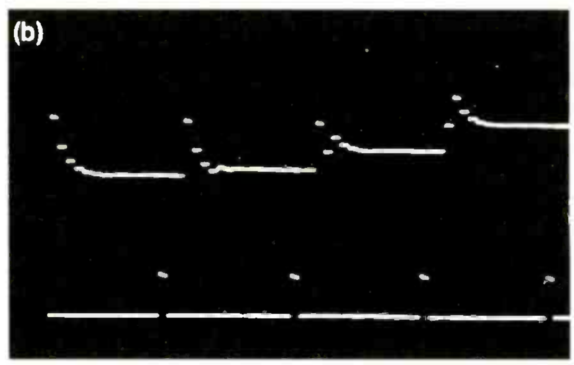
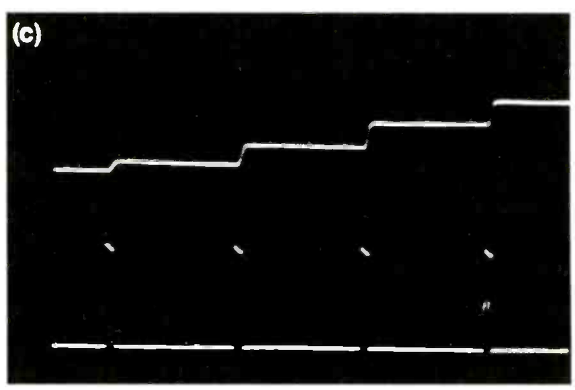


Fig. 10. Observe the successive approximation process;(a) the sample signal appears along the bottom of the trace. Expanding the time axis (b) offers a magnified view of the sine-wave conversion. An expanded view of the sampled-and-held results of each approximation appears in (c). The top trace in (d) is the encoded 2.5kHz sine wave as it appears before (above) and after (below) passing though a seven-pole 12kHz Butterworth filter.



Using predictive coding to save money

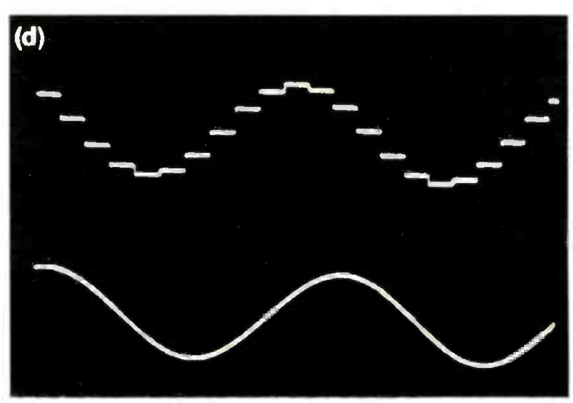
Most applications of predictive coding systems take advantage of the technique's digitisation efficiency, which can produce savings in hardware, memory, channel bandwidth and converter cost.

Video and image-processing systems use predictive coding to minimise data and enhance accuracy. In voice communications, predictive converters can minimise the data rate of multiplexed transmissions.

Predictive coding can also be applied in the field of professional audio equipment. Audio circuitry that handles music signals has always required the best signal-to-noise performance; predictive coding minimises memory requirements for digital reverberation systems while achieving 90dB S/N ratios.

A more subtle advantage of predictive coding relates to current data-conversion technology. Fast 12-bit A-to-Ds are relatively inexpensive and common, but longer-word-length converters (14 and 16 bits) are either slow or expensive.

Predictive coding, in conjunction with today's 12-bit converters, enhances conversion accuracy. A 12-bit predictive encoder, for example, can generate a 14-bit output word when the encoder runs at a rate faster than that necessary to avoid aliasing. Converting signals band-limited to 10kHz at a 160kHz rate and then digitally resampling at twice the Nyquist frequency (20kHz) can yield about three additional bits of accuracy.



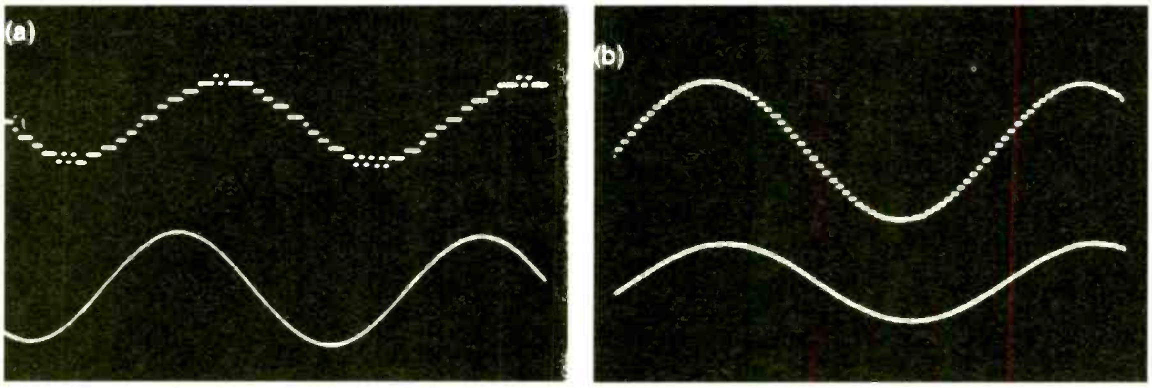


Fig 11 With a 750mV 500Hz sine wave input, a system without prediction suffers from severe quantization errors when encoding a small input (a). With prediction, quantization is more accurate (b). The lower trace is filtered.

used in the A-to-D feedback loop; this circuit has a time constant matching that of the predictor to ensure the encoder's exact inverse frequency response.

System handles 32kHz sampling rate

As with any analogue-to-digital conversion system, D-to-A converter settling times and analogue sampling limitations determine the system's conversion rates. This circuit handles a sampling rate of 32kHz (31µs conversion time). Scope photos, **Fig 10**, illustrate the effectiveness of predictive coding in reducing quantisation errors. Figures 10a, 10b and 10c show various points in the successive-approximation routines, while Fig 10d illustrates a 2.5kHz reconstructed sine wave after it has passed through a 7-pole Butterworth signal-reconstruction filter with 12kHz cut-off frequency.

Standard conversion systems have trouble with low input levels. Emphasising the quantisation error in conversion, **Fig 11a** shows a 750mV pk 500Hz sine wave in a predictive system without an output accumulator. The LSB jumps are plainly visible at the tops and bottoms of the sine wave, indicating severe digital noise contamination. **Figure 11b** depicts the same signal after it has passed through the output accumulator. The photo shows barely noticeable quantisation error.

Richard J Karwoski, Consultant

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SC110A miniature portable oscilloscope

SC110A miniature portable oscilloscope

The SC110A from Thurlby-Thandar is a full feature, single trace analogue oscilloscope packaged into the size of a benchtop multimeter. Fitted with a 32mm x 26mm screen miniature CRT, the bright, sharp image provides resolution and detail associated with much larger instruments. UK designed and built, the internal switch mode power supply draws just 195mA from four C sized batteries (not supplied). The instrument will operate from 4 to 10V DC.

The specification includes a Y bandwidth of DC to 10MHz, 10mV/div sensitivity and an adjustable brightline trigger with AC/DC/TV coupling from both internal and external sources. The X timebase is adjustable from 500ms/div to 100ns/div in 24 steps. The case measures 25 x 5 x 15cm and the instrument weighs about 1kg. SC110A £249+VAT (£292.58).

1021 general purpose oscilloscope from Leader Electronics

1021 general purpose 20MHz oscilloscope

The Model 1021 general purpose oscilloscope from Japanese instrument maker Leader Electronics more than meets its published specification and is of exceptional build quality. Features include 20MHz dual channel operation, 8cm x 10cm display area, 5mV/div Y1/Y2 sensitivity at 20MHz, DC to 500kHz X-amplifier response, variable trigger response, multiple sync conditioning and an overall accuracy better than 3%. 1021 £299+VAT (£351.33)

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PL320K laboratory triple power supply

TS3022S laboratory dual power supply

PL320K laboratory triple power supply

This power supply from Thurlby-Thandar combines three, totally independent power supplies within a single unit: 0–30V at 2A, 0–30V at 1A and 4–6V at 7A for logic supply. The 30V supplies will operate in a bipolar tracking mode for ±30V operation or in a series mode to provide 0 to 60V output. Both supplies incorporate independent remote sensing and independent precision voltage/over-voltage/current-limit preset. Three 3 3/4 digit led panel meters indicate current and voltage to an accuracy of 0.05% fsd. Output stability is typically 0.01% for 90% load change. PL320K £359+VAT (£421.83).

TS3022S laboratory dual power supply

This laboratory quality power supply from Thurlby-Thandar provides two fully floating 0–30V 2A outputs for parallel, series or independent operation. Each supply has its own metering of voltage and current by LCD display; with the output switch to off, the display can be used to preset the voltage and current limits prior to connection of the load. Coarse and fine controls permit output voltage adjustment to within 5mV of a predetermined value. The current limit control employs a log law for precise adjustment down to 1mA. Load regulation is typically within 0.01%. Both supplies incorporate remote sensing. TS3022S £299+VAT (£351.33).

TD201 digital storage adaptor

The TD201 digital storage adaptor from Thurlby-Thandar is a low power, single channel digital storage unit which adds digital storage capability to ordinary analogue oscilloscopes. The maximum sampling rate of 200kHz permits fast transients to be captured while the lowest rate can extend the sampling period to over an hour. The unit stores over a thousand points on the X axis with 256 levels in the Y axis. The internal batteries (not supplied) allow data retention for up to four years. Other features include an AC/DC sensitivity down to 5mV, selectable pre-trigger, roll and refresh modes and a plot mode. The case measures 25 x 5 x 15cm and the unit weighs about 1kg. The TD201 provides the ideal solution for those wanting a well specified and easy-to-use DSO at the lowest possible cost. TD201 £195+VAT (£229.13)

TD201 digital storage adaptor

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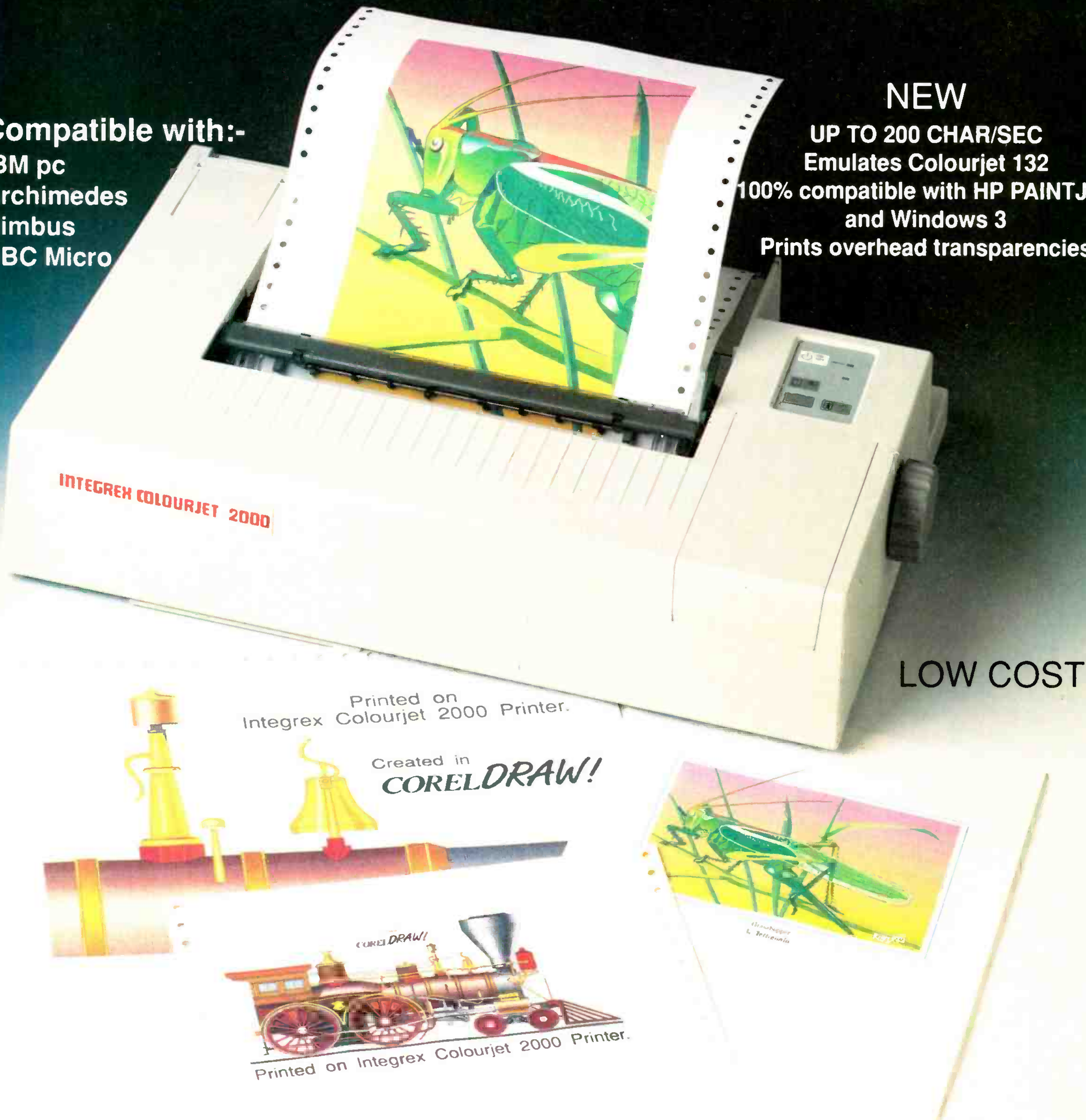

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D-to-A converter on Centronics printer port

Very few external components in addition to an 8-bit A-to-D converter will give an analogue voltage proportional to the data on a parallel Centronics printer port.

Depending on the number between 0 and 255 at the port, the voltage level lies between zero and 2.55V. Port data is stored in the *AD558*'s internal latch as the strobe goes high, although the strobe may be inverted if necessary to take account of those machines in which data is not valid on the rising edge of the strobe.

J Vandana
World Friends AI Group
Tamilnadu
India

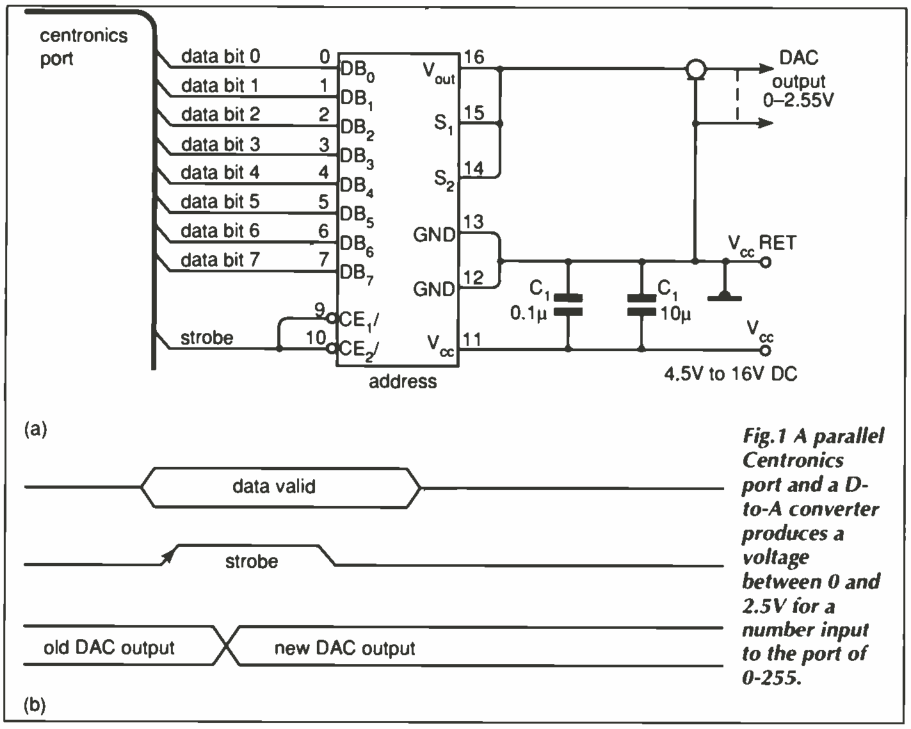
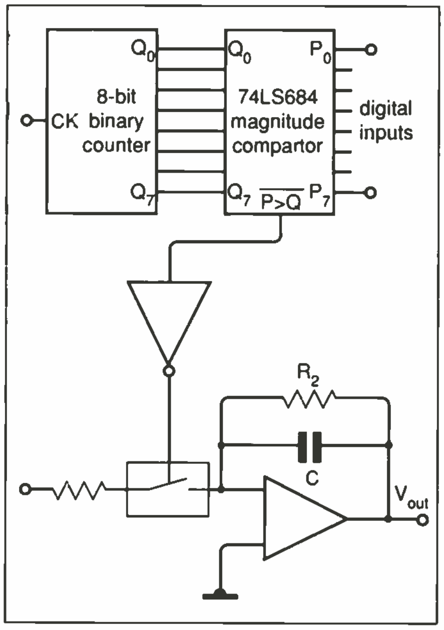


***Fig.1** A parallel Centronics port and a D-to-A converter produces a voltage between 0 and 2.5V for a number input to the port of 0-255.*



Digital-to-analogue converter does not depend on resistor tolerances.

Tolerance-independent D-to-A

If conversion time is not important, this circuit avoids the use of high-precision resistors while still providing extremely high linearity.

In the diagram, the outputs of an 8-bit counter are compared with an 8-bit digital input in the LS684 magnitude comparator. The resulting rectangular wave at the $P{>}Q\backslash$ output has a mark:space ratio directly proportional to the P input and is used, via an inverter, to turn on and off the analogue switch. Direct current through R_1 and R_2 is therefore modulated by the digital input and the amplifier output is $-(NV_{ref}/256)(R_2/R_1)$, where N is the nemerical value of the digital input. Resistor values have no effect on linearity, merely setting conversion factor. Capacitor C smoothes the output.

Linearity depends in the main upon amplifier performance, assuming the frequency is not too high. It should be a simple matter to modify the counter and comparator to provide more bits of resolution.

David J Haigh
Stoke

Simple x^2 converter

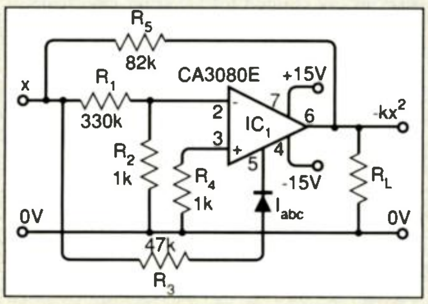
To avoid the complication and expense of using a four-quadrant multiplier to produce an input-squared output, I have used the circuit shown.

An operational transconductance amplifier computes the expression $-kx(x+v)$, where V is the negative supply voltage and k depends on component values. A term $(-kxV)$ is now present at the output, proportional to the input, which must be removed by adding an equal but opposite component through R_5. The result is $-kx^2$, where

$$k=(g_m R_2 R_3 R_L)/[(I_{abc} R_3(R_1+R_2)(R_5+R_L)]$$

In the circuit shown, $k\approx0.1$ if R_L is high. An offset null adjustment and gain compensation provide greater accuracy.

Ian M Wiles
Basingstoke
Hampshire



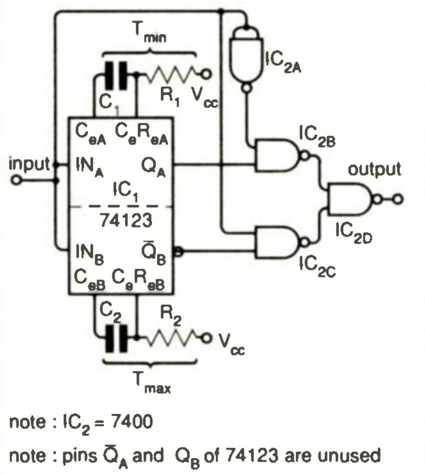
Simple circuit gives kx^2 where k is about 0.1, and avoids need for four-quadrant multiplier.

Pulse-width monitor

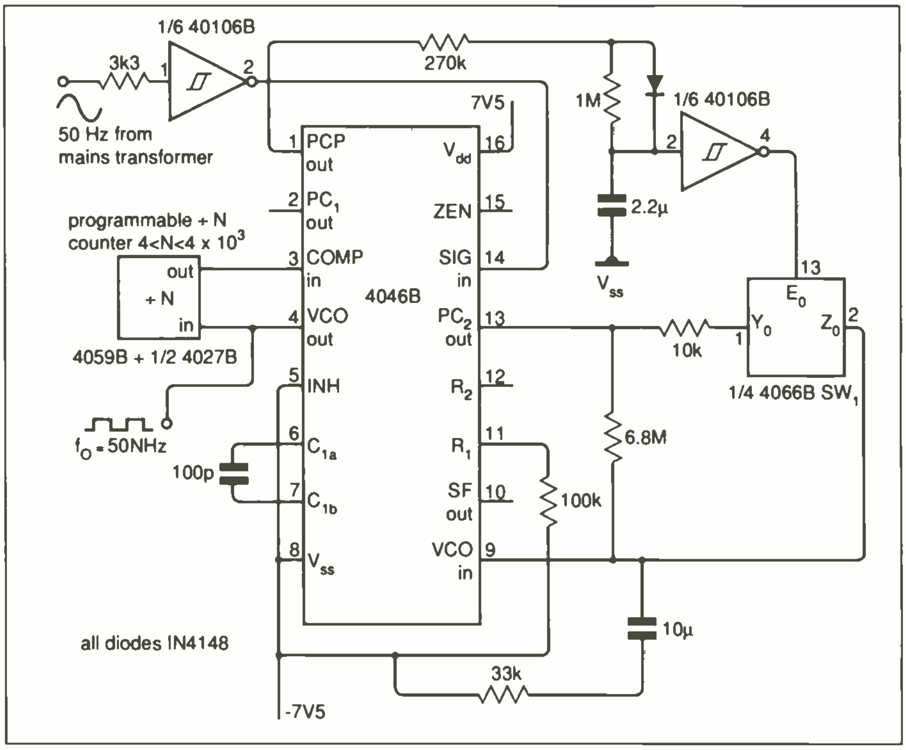
If a positive-going pulse is either shorter or longer than a preset time, this circuit indicates the fact.

Minimum time T_{min} and maximum T_{max} are adjusted by means of C_1R_1 and C_2R_2, the timing circuits of a dual monostable flip-flop. Outputs of the two Nands IC_{2b} and IC_{2c} remain high unless an out-of-range pulse is detected, in which case one goes low and the output of the circuit produces a positive-going pulse.

Frantisek Michele
Brno
Czechoslovakia



Positive-going pulse at the output signals an input pulse outside a preset time range.



Using pulse comparator output of PLL to switch filter in and out of multiplier circuit avoids long wait for lock while avoiding spurious frequency changes due to the effects of input noise.

FRESH IDEAS

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Fast-response PLL frequency multiplier

To avoid the effects of noise in a 50Hz-input PLL multiplier, it is common to use a long-time-constant low-pass filter, which prevents the multiplier responding to the noise. It also, however, entails a delay in the desired response. The circuit described obviates the problem.

The design allows multiplication of the 50Hz input by anything from 4 to 4000 in multiples of 4, ignoring frequency reference disturbances that last less than a second; it takes four seconds to change frequency from 200Hz to 200kHz.

When the loop is locked, the phase-comparator pulses (PCP) output at pin 1 of the *4046B* PLL is at 7.5V, going to -7.5V when it is out of lock. On changing division ratio in the programmable counter, loop response is slowed by the filter from pins 9,14 to V_{ss} and if the loop is not locked after 2s, switch 1 closes and shorts the filter; it then stays closed until the loop has been in lock for more than 0.5s. This delay prevents overshoot and the hysteresis of the closing and opening of the switch enables the loop to track intended frequency variations at the input without imposing a long lock time for large changes.

Richard Pulham

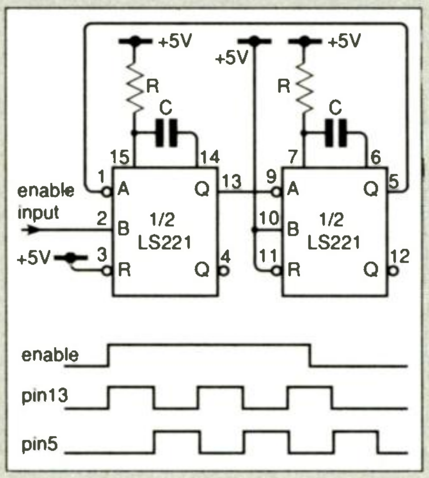
Accurate gated oscillator

Many gated oscillators do not start accurately; they have a tendency to produce a short or long first cycle. This circuit uses a pair of *74LS221*s to make sure the output is clean. the edge fed back from the second circuit terminating the first cycle.

With both timing circuits equal in value, frequency is approximately 1/(1.4CR) up to a realistic maximum of 10MHz.

Differing timing circuits on each flip-flop will give other mark:space ratios.

Jeffrey Borin
Harrow
Middlesex



Two flip-flops ensure accurate starting of a gated oscillator.

Steep-cut, low-pass filter

In Circuit Ideas for July 1987, Tim Mason showed a filter circuit which had the requirements of a steep roll-off above 15kHz with a deep notch at 19kHz. A computer model of the filter gives the result shown in **Fig. 1**, where the peak is caused by the undamped parallel-T section.

In my modified version shown in **Fig. 2**, the parallel-T gives low-pass filtering as well as the notch by attenuating the input to the high-pass section; resistor R_2 dampens the peak by reducing positive feedback. Notch frequency is 19kHz.

To obtain the fairly sharp corner in the response at around 15kHz, 2% components are necessary. A source impedance of <200Ω, such as that from an op-amp, is also needed. Amplifiers in the *TL071/2/3* series, with an open-loop unity gain frequency of 1MHz or better should be used.

Scaling capacitor values up by a factor of 10 would give a corner frequency of 1.5kHz — useful for speech on narrow-band SSB direct-conversion receivers; scaling by 100 would give 150Hz for Morse.

J A H Edwards
Leicester

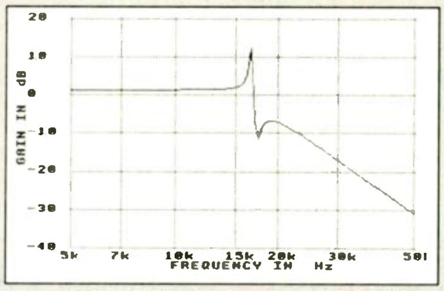


Fig.1. Original steep-cut low-pass filter with notch at 19kHz.

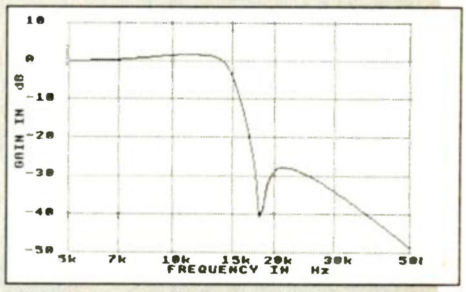
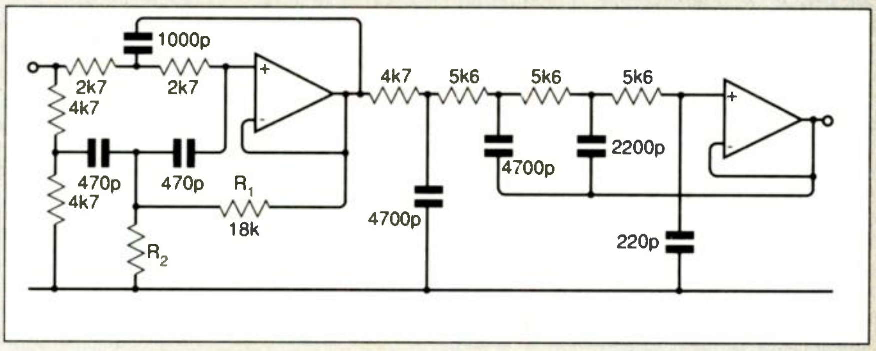


Fig.3. Response of Edwards circuit.

Fig.2. New circuit eliminates peak at 16.5kHz, gives deeper notch and a greater degree of low-pass filtering.



Rumble filter preserves bass

Using a filter to remove very low frequencies caused by, for example, a warped record has an unfortunate tendency to remove some of the music as well. This circuit avoids the problem.

Since rumble is caused by vertical movement of the stylus, the stereo signals are anti-phase and can be removed by passing all out-of-phase components through a second-order high-pass filter to produce a bass mono signal, common-mode signals at all frequencies being left unchanged.

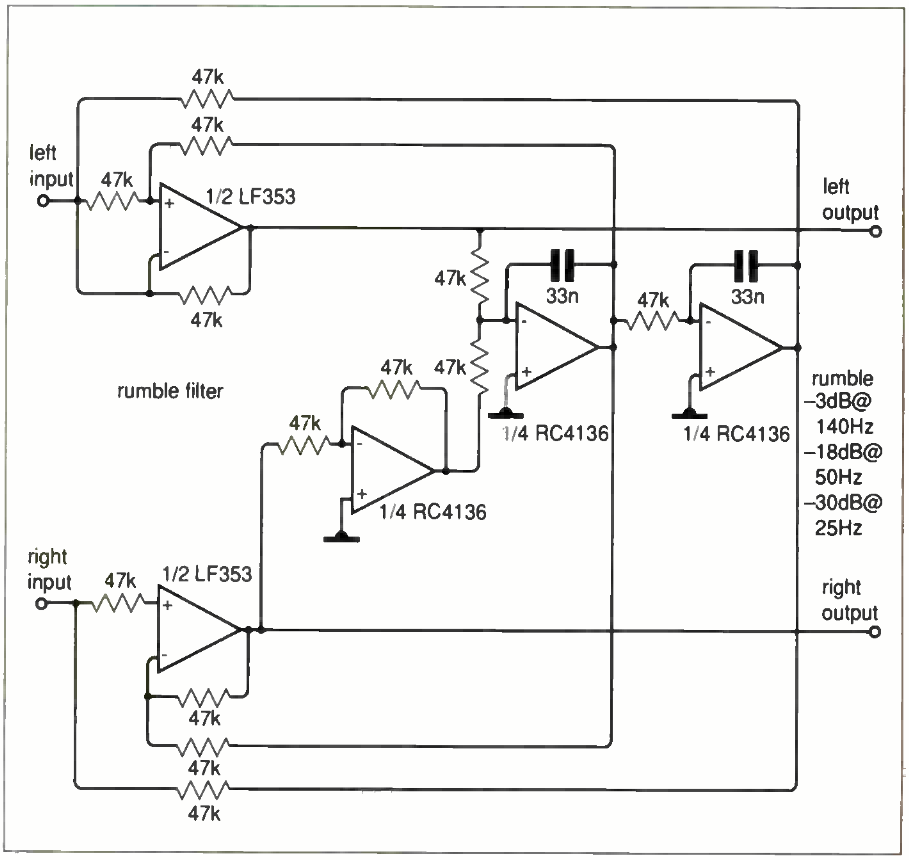
A modification of the biquad filter is used, in which the integrators and inverter need not be of wide bandwidth, an ordinary standard op-amp being satisfactory. Fewer capacitors, a steeper roll-off of the out-of-phase signals and more attenuation are obtained than is the case with earlier designs[1,2].

It is possible that the design may find application in the protection of the cutter head in disc mastering[3] and preventing large speaker-cone excursions at LF.

John Lawson
Cheltenham
Gloucestershire

References

1. Macaulay, J P, Circuit Ideas, *Wireless World*, September 1979.
2. Langvad, J, Letters, *Wireless World*, March 1980.
3. Eargle, J, Sound Recording, 2nd edition, Van Nostrand (1980), ch.10.



Out-of-phase signals caused by turntable rumble are filtered to -18dB at 50Hz, all other signals being unchanged.

LETTERS

Electromagnetic kinetics

Dr Millar's article "Scratching the surface of electromagnetism" (*EW + WW*, December 1991) prompts me to ask some fundamental questions relating to matter, and magnetism in general.

Does anyone know whether a negative charge has been defined in terms of the kinetic energy stored in a moving electron? Or whether a positive charge has been defined in terms of a hole's potential energy in relation to the neighbouring atoms?

The magnetic field of a bar-magnet is often portrayed as static – observe the pattern of iron filings when sprinkled on a surface over the magnet. Is it possible that the lines of flux are continuously moving from one pole of the magnet to the next, at a period related to the frequency of electron spin?

A current is only induced in a conductor when subjected to a hanging (dynamic) magnetic field, suggesting that kinetic energy from the field is transferred to electron movement thereby constituting a current? After all, even with a highly intensive but static field no current is induced.

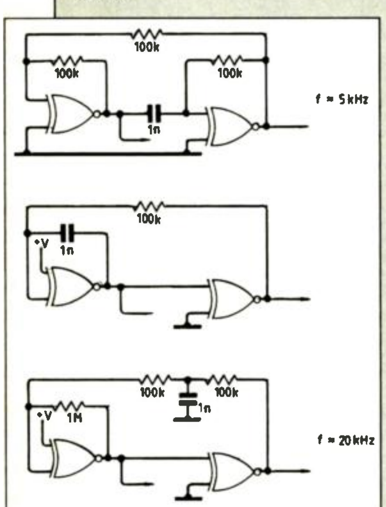
My understanding of kinetic energy is that it is the extra energy possessed by a moving body. But what is kinetic energy really? Perhaps it is simply that each atom has increased energy levels by adjusting the velocity and distance of the electrons around the nucleus?

Is kinetic energy affected by gravity?

George Ho-Yow
Middlesex

Multivibrator musings

The subject has arisen of multivibrators constructed from gates, and their sometime failure to start. This problem can be avoided by regarding each gate as a high gain amplifier and taking the timing capacitor to be open, so that each gate must be biassed into its "linear" operating range. The approach implies negative feedback of some kind, and some conclusions can immediately be drawn:



- No circuit exists using all non-inverting gates.
- If two inverting gates are used, there must be two feedback resistors, one across each; but if three are used they can be connected in a single loop.
- Two gates, one inverting and one non-inverting, can also be connected in a single loop.

The remaining problem is to position the capacitor so that feedback changes from negative to positive at some high frequency.

Three circuits, each using the *MC14077* exclusive Nor gate, have had the unused input wired for invert or non-invert as required.

Circuit 1 might be used when two inverters are available, but circuit 2 uses fewer parts.

Circuit 3 can be useful when the capacitor must be grounded. All three provide square wave in both polarities, and many others are possible.

McKenny W Egerton Jr
Owing Mills USA

Unstable astable

Circuits in your Design Ideas pages regularly use the analogue characteristics of cmos logic gates to provide timers, variable pulse width modulation, oscillators and amplifiers etc.

Such a design approach may be adequate for home experimentation but is entirely inappropriate for any manufacturable product. If you examine the various cmos suppliers' data books you will see that the threshold of these devices is highly variable – typically from 0.3 to 0.7 of the supply rail – and stability is by no means guaranteed with time, temperature or supply rail changes. Further, when biased for use in a linear manner the supply current taken is very poorly defined.

Hence any design based on these characteristics would either have to be such that large variations in performance could be tolerated, or would require that the user regularly adjusts values to retain adequate functionality.

Design techniques based on carefully specified analogue devices for analogue functions and digital ones (including microprocessors where appropriate) for digital functions – though more expensive in initial manufacturing cost – are much more likely to generate products which are easy to manufacture and test, and provide long-term reliable performance. The result of this approach is satisfied customers rather than the satisfaction of getting a quart out of a pint pot. I recommend it.

Allan Hurst
Cambs

Blomley questions

W Groome asks (Letters, *EW + WW*, January) why the Blomley amplifier principle did not sweep class B off the market. I believe the reasons are threefold:

The principle was patented, deterring manufacturers who are averse to paying licence fees, and the associated negotiations can be time-consuming.

The principle of operation is not obvious, but requires some study before it can be understood.

It is not clear that the performance advantages over conventional techniques are overwhelming.

As to the third point, some years ago I built two Blomley amplifiers, closely following the published circuit except for some transistor substitutions. I found the performance no better than that of a conventional complementary bipolar amplifier with reasonably high feedback; in particular there were still crossover effects to be seen on the residual, and attempts to get rid of these failed.

I hasten to add that while this may demonstrate my shortcomings rather than those of Blomley amplifiers, it does to seem to show that the approach is not foolproof.

Interestingly, exactly the same three conditions hold for the Cherry NDFL (nested differentiating feedback loops) scheme – a method for increasing the negative feedback permissible for given stability margins (*JAES*, May 1982). To the best of my knowledge this has also failed to achieve commercial exploitation.

Turning to the class struggle in amplifier nomenclature, I would suggest that class AB be reserved for amplifiers that can develop a significant amount of power in class A, (perhaps 5W?) leaving class B for those with sufficient quiescent current to remove the crossover spikes from the residual, but no more. I suppose this makes a Blomley class B, though such classification says nothing about the principle of operation.

Douglas Self
London

Viva the valve

Far from having disappeared down a deep black hole, the Communications & Electronics Museum is very much alive and kicking, and both of the wireless museums on the Isle of White are open throughout the summer months.

The National Wireless Museum has been at Arreton Manor, near Newport, for the past sixteen years, while the newer one can be found at Puckpool Park, Seaview.

Both will be open again at Easter, I can assure Mr Hawkins (Letters), who probably tried to contact me while I was in the States and my phone recorder went on the blink – it is a typical modern black box filled

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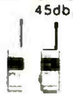

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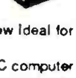

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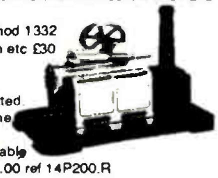

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with chips, and it might well have kept going had it been equipped with old-fashioned valves to keep it nice and warm and dry. But then I love valves...

Douglas Byrne
G3KPO/GB3WM
Communications & Electronics Museum Trust
Isle of White

Radar tracking

Geoffrey Horn infers (Letters, "Delayed Credit" (*EW + WW*, February) that Germany has always been the indisputable leader in radar development. I too have read David Pritchard's comprehensive book, to which Mr Horn refers, but can not agree – even though I am fascinated to learn of the strenuous radar work done in Germany both before and during the war.

It is too easy to simplify and overstate cases. Take, as an example, the statement: "a German demonstrated centimetric radar at sea in 1904". From the book it is clear that the equipment in question was a rudimentary obstacle detector, not a radar. Measurement of range is implicit in the definition of radar.

According to the book, Hülsmeyer neither provided a demonstration nor suggested any real approach to range measurement.

At the conclusion of the report of the trial on the tender Columbus, the book states: "The apparatus used in the trial was not yet arranged for the determination of distance" (p.13).

"The assembly was designed for mounting on the foremast of a ship at a pre-determined height, range-finding being carried out by raising or lowering the assembly on reception of an echo and employing triangulation or simple geometry with the use of a prepared scale of figures" (p. 23).

Could one seriously consider measuring such a tiny angle with any hope of accuracy – and from a floating platform too!

I believe that the lack of a proper range-finding technique was the reason why Hülsmeyer's claim to have been the originator of radar was rightly, but perhaps unnecessarily tersely, turned down by the British Patent Office in 1951.

Certainly Hülsmeyer deserves an honoured place in the history of radio engineering as a far-sighted and courageous inventor: but to credit him with being the "father of radar" – as it appears many in Germany do – is, in my view, over-stretching the facts. It is unfair to many others, of whatever nationality, who later developed FMCW and pulse techniques for the positive determination of range.

I am somewhat surprised to read Mr Horn's comment that the book seems to have received little recognition. I can only say that judging by the number of comments I have heard about it in the last couple of years, it does seem to be read quite widely, well justifying the great effort the late David Pritchard put into it.

Colin Latham
Anglesey

Forty year delay

As a teenager in the late 1950s, I built the "Simple valve voltmeter" detailed in *Wireless World*, December, 1950, pp. 430-432. Unfortunately, it did not work very well, so I put it away as another "failed project".

Recently I came across that issue again and when studying the circuit, realised that R_{13} should be 4.7kΩ not 47kΩ as specified. I rescued my valve voltmeter from the attic, changed R_{13} to 4.7kΩ and having replaced C_1 and C_2 which had died over the years, it now works as specified.

Better late then never!

Jan M Chapman
Bradford

Crossover dissertation

In response to my article "Reducing crossover distortion" (*EW + WW*, October 1990), you published a critical letter from Phil Denniss of Sydney University.

May I take this opportunity to reply.

In the figure, showing a basic class B amplifier, the amplified diode is a nuisance, so it has been removed. To reduce the resultant crossover distortion, the gain components usually found around *B* have also been omitted, to increase to 100% the voltage feedback to Tr_1 emitter. During crossover, point *D* has to rise with infinite speed to prevent distortion, but C_2 prevents that. (Rapid rise at *D* pumps current down through C_2, preventing rapid rise at *D*).

The remedy is to throw S_1. For most of the signal cycle *E* just follows *D*, so C_2 is effectively driven as before.

But during crossover, *D* can now jump fast, without pumping that crippling current down through C_2. It is true that during crossover C_2 loses its compensation function.

But it is also true that during crossover, compensation is not required, because the gain loop is broken (*E* is stationary).

Experiment supports this idea. Input 1V pk at 100kHz to the circuit, and output is the same, but with an enormous bite taken out of the first quarter-cycle. Switch S_1 and you will instantly see an apparently perfect output trace. Switching S_1 reduces crossover distortion enormously. But does it do so sufficiently that the amplified diode can be omitted?

My October 1990 article set out to examine this question, and concluded that with S_1 switched the crossover distortion in the circuit shown is well below the audible level. Making R_4 a current source carries it down much further. This PA circuit outperforms the traditional amplified diode version, and is simpler and safer.

In his April 1991 letter Phil Denniss of Sydney reviews my article and allows my conclusions – but only for the small circuit shown.

A normal power amplifier, he says, would develop the three sub-systems, so that each of my transistors might be replaced by a pair. The longer gain loop that results will slow things down, particularly when the behaviour of power transistors is considered. I may have removed the delay caused by C_2, but there are other serious delays in a large amplifier. These will prevent the fast slew needed at *D* to avoid crossover distortion.

But pundits do not seem to agree: they write of single dominant pole compensation. As frequency rises, the loop gain must fall below unity. Indeed it must do so before frequency rises far enough to create the 180° extra phase lag needed for oscillation.

In practice you pick one point, and hit it hard with a capacitor, so that loop gain falls below unity before any other phase lags become important. In that way any disastrous accumulation of phase lags is fought off.

The clear implication is that at relevant frequencies (up to the 5MHz or so potential oscillation figure) the dominant delays in a class B circuit are at the capacitor point, and not, for example, because of the slowness of the power transistors.

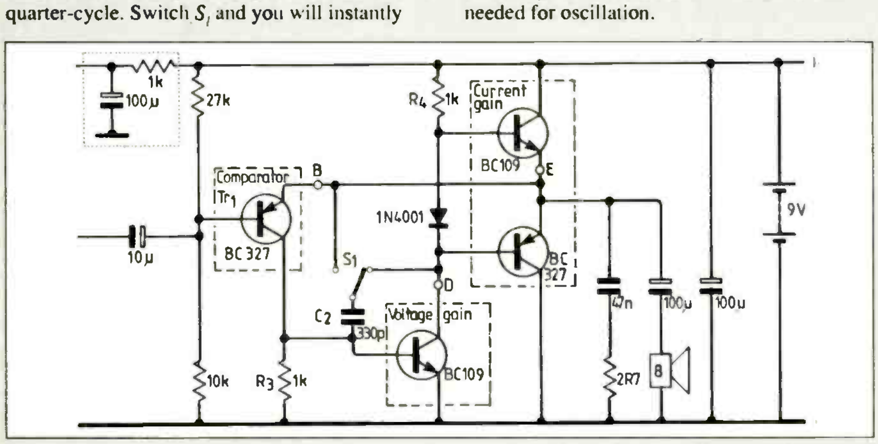
Indeed Mr Denniss himself says the latter can work up to 100MHz. Thus it does seem that reconnecting C_2 may destroy the dominant delay, and allow even powerful amplifiers to dispense with the amplified diode.

I have recently been able to inspect the circuit of the *TBA820M* power amplifier IC. This is a full amplifier circuit, able to produce peak output current of 1.5A, but having just 1mA bias current through the output transistors. Its point *D* must jump very fast, and to allow this C_2 is reconnected as I propose. I hope the idea will spread into the world of discrete component amplifiers.

Mr Denniss observes that reducing the amplifier gain to unity has led to a large value for C_2, and this has brought down the slew rate of the amplifier. Agreed. I have now checked that leaving voltage gain of 11 in the amplifier (as in Figure 1 of my article, but with C_3 reconnected and reduced to 22pF) puts up the 1µs spikes found in the output (trace F) from 0.6mV to 6mV. That is still much less than the 10mV spikes lasting 20µs found by Eric Margam in the output of traditional power amplifiers containing an amplified diode. Thus the proposed system can accept gain, and so recover the normal slew rate, if desired.

The circuit shown will be attractive to teachers, and I include it in my schoolbook *Electronics for You*. Switch S_1, keep leads short (especially BE), and drive from a 10k pot.

Michael McLoughlin
Haberdashers' Aske's School



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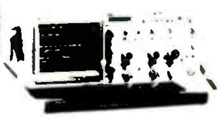

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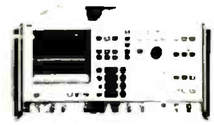

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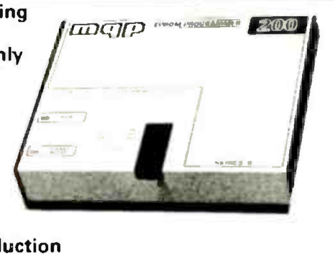

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CIRCLE NO. 120 ON REPLY CARD

Programmable universal filter

Two *MAX682* programmable, switched-capacitor filter ICs contain the sections required to make a psophometric noise-weighting filter. In Europe, this type of filter is used to simulate the response of the ear in audio measurement, while in the USA an equivalent is the C-message filter.

Programming the IC filters will give either type but, this being an American device, the Maxim *Engineering Journal* gives no details of the European filter. Software is, however, available from Maxim.

Figure 1 shows the filtering method, which consists of three second-order band-pass sections and a second-order low-pass type. Loading the filter chips with different sets of coefficients gives other responses, the ones shown in **Fig. 2** controlling filter mode, Q and cut-off frequency for the filter of Fig.1.

IEEE Standard 743-1984 specifies pole locations and f_0 values for the C-message, taken account of by the listed coefficients. Each section contains a pair of Tchebyshev filters with centre frequencies digitally programmable over 1 to 25kHz in 128 steps, pass-band ripple being 0.1dB. Band-limiting is needed at the input to a quarter of the clock frequency and the uncommitted op-amp in IC_1 will give a second or third-order

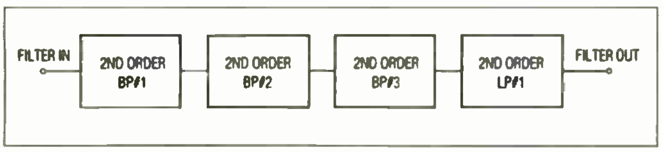
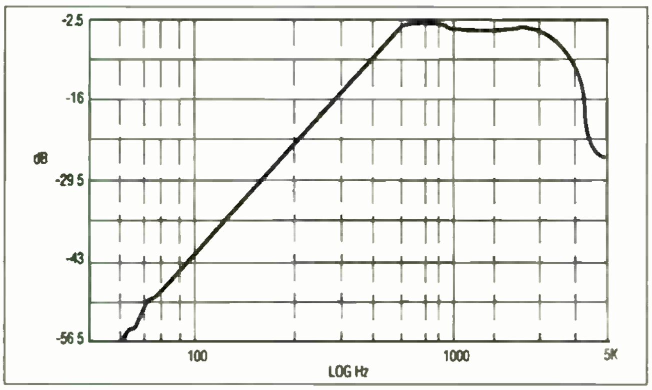


Fig. 1. Form of the C-message or psophometric filter used for audio measurement.

Fig. 2. Two Maxim ICs used to implement the filter of Fig. 1. Loading the filter chips with other coefficient sets produces a different filter type. Data is in software from Maxim.



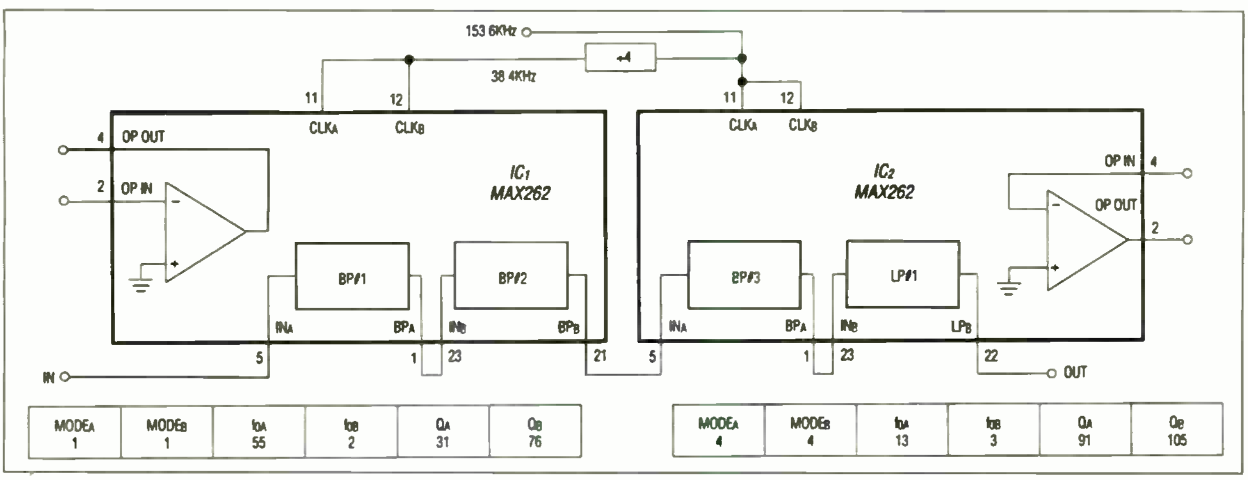
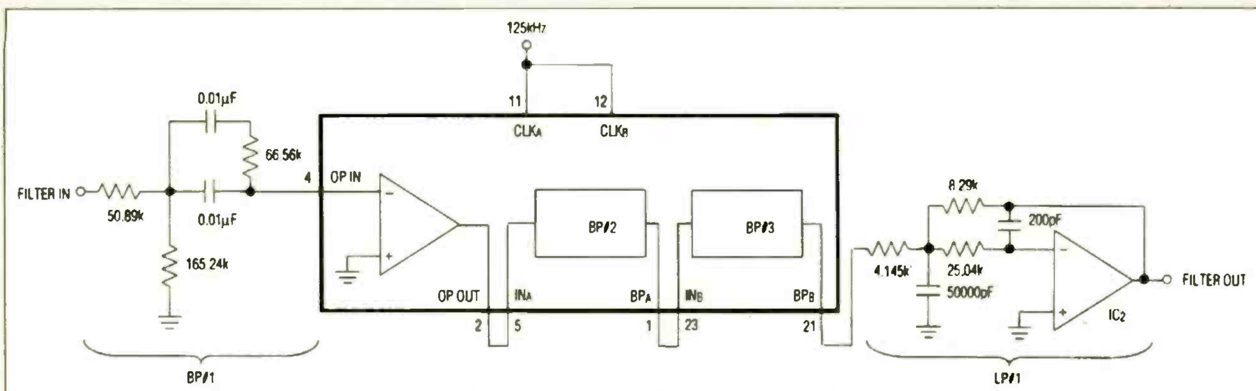


Fig. 3. Frequency/amplitude response of circuit when configured as C-message filter.



CLKA	CLKB	MODEA	MODEB	f_{0A}	f_{0B}	Q_A	Q_B
125kHz	125kHz	1	4	N = 61	N = 6	N = 75	N = 91

filter to do this, the equivalent op-amp in IC_2 also acting as a low-pass filter for that purpose. **Figure 3** is a frequency/amplitude response of the circuit configured in its C-message guise.

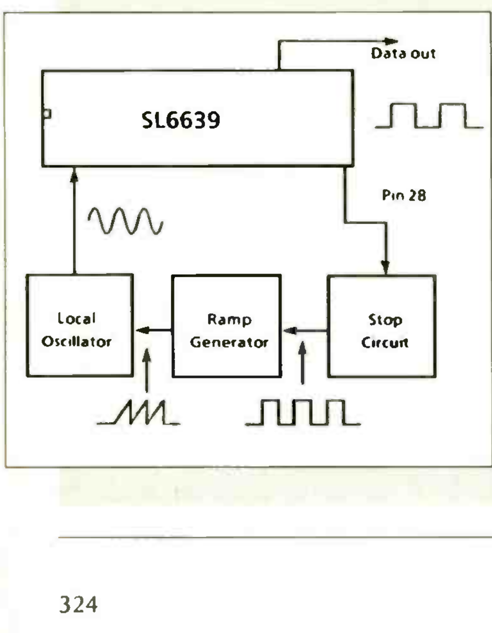
A somewhat more compact filter is illustrated in **Fig. 4**, where only one *MAX262* is followed by an op-amp. The internal, uncommitted op-amp becomes the first band-pass filter and also does the work of the anti-aliasing filter. Second and third band-pass sections are as before and the external op-amp is the low-pass filter and smoother. Using the op-amps as filters does, however, mean that flexibility is lost – this is a one-type filter.

Fig. 4. One of the filter chips and an op-amp will do the job, but the circuit is no longer programmable for different characteristics.

Maxim Integrated Products (UK) Ltd,
21C Horseshoe Park, Pangbourne, Reading RG8 7JW. Telephone 0734 845255.

Scanning the SL6639

Plessey's *SL6639* is a 200MHz direct-conversion FSK data receiver, which has an external local oscillator. It was meant for fixed-frequency working, but Application Note AN96 describes a method of scanning the receiver to detect a transmission, receive the data and move on to look at the rest of the band. Only fairly small modifications need be made to the demonstration board and inexpensive components are used. No RF is carried by the



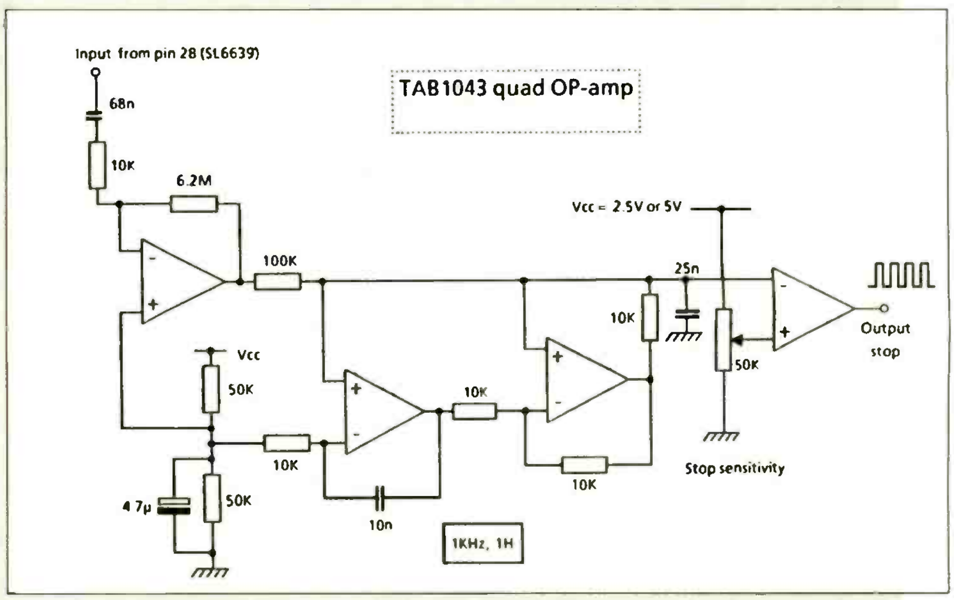
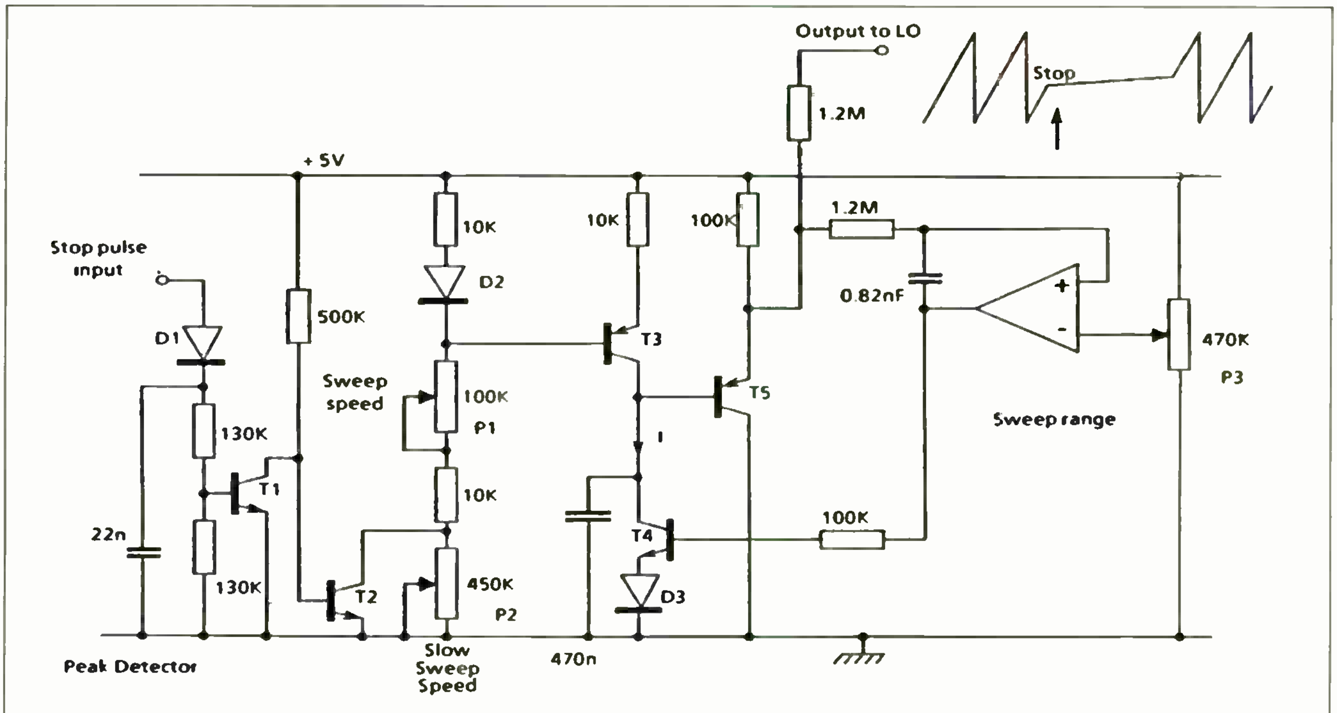


Fig. 1. Outline of Plessey's method of scanning the SL6639 direct-conversion FSK data receiver and detecting data while scanning rate is reduced.

Fig. 2. Stop-pulse circuit to slow down scanning rate to allow data reception. Two middle op-amps form gyrator of one henry inductance.



connections, which allows longish leads to the board.

Figure 1 shows what happens. Saw-tooth from the ramp generator scans the local oscillator through the required band; when a signal is found, the stop circuit emits a train of pulses which damp the ardour of the ramp generator so that the local oscillator shifts the reception frequency through the signal slowly enough to collect all the data bits. With the standard ±4.5kHz deviation, the slow scan takes up 11kHz.

Figure 2 is the stop circuit, which uses a *TAB1043* quad op-amp in a gyrator-plus-comparator circuit. A signal from pin 28 (channel B output test) is first amplified and then taken to an LC tuned circuit in which, the required inductor being 1H, two op-amps form a gyrator. When the output of the tuned circuit exceeds the comparator threshold, pulses appear at the output and are taken to the ramp generator in **Fig. 3**.

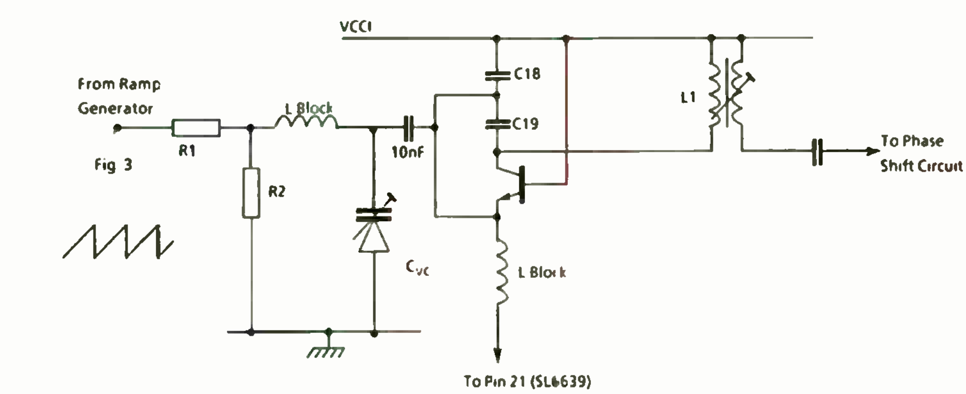
This is a constant-current source (D_2 and Tr_3) feeding the 470nF capacitor. When the ramp voltage across the C reaches the comparator threshold set by P_3, Tr_4 comes on and resets the ramp. Incoming stop pulses are peak-detected by D_1 and, when the DC reaches a given voltage, Tr_1 switches on and Tr_2 off, so that the capacitor current becomes very small and the ramp slows down until stop pulses stop pulsing. The resulting hesitant ramp goes to the local oscillator in **Figure 4.**

With its crystal removed, the *SL6639* oscillator becomes a Colpitts type tuned by a ramp-fed varactor. ■

Plessey Semiconductors Ltd, Cheney Manor, Swindon, Wiltshire SN2 2QW. Telephone 0793 518000.

Fig. 3. Ramp generator to scan local-oscillator frequency and to slow down when stop pulses are recognised.

Fig. 4. Local-oscillator modification, using Varactor fed by ramp generator. Basic oscillator is that shown in data sheet, but with crystal removed.



R1 and R2 are chosen for desired frequency and sweep range.

Varactor Diode Capacitance Appears in Parallel with C18.

$$fc = \frac{1}{2\pi}\sqrt{\frac{1}{L1\left[\frac{(Cvc + C18) \times C19}{(Cvc + C18) + C19}\right]}}$$

The component values depend upon the choice of centre frequency and sweep range required.

REGULARS

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Fast PLD. The first two members of Altera's Multiple Array Matrix(MAX)7000 family of high-density, high-speed erasable programmable logic devices (EPLDs) are now available. EPM7256 is a 256-macrocell, 192-pin device using eprom transistors and is claimed to be the highest pin-count PLD in captivity, while the EPM7032 32-macrocell, 44-pin EPLD is the company's first to be made in eeprom technology. The 7256 supports 70MHz clock rates and contains 10,000 gates. Altera UK, 0844 275285..

A-to-D & D-to-A converters

12-bit data converter. Integral and differential non-linearities of ±0.25 LSB are exhibited by Burr-Brown's DAC65, a fast-settling digital-to-analogue converter meant principally for ATE and direct digital synthesis. Capacitors and a resistor chip are in the package, but off the main chip, for high-frequency spectral purity. Settling time is 35ns to ±0.012% of full scale. Dynamic range without spurious responses is -73dBc with 100kHz sine output and -61dBc at 5MHz; THD is -66dBc at 1MHz. ECL data is accepted.Burr-Brown International Ltd, 0923 33837.

Discrete active devices

High-current pair. A p-n-p device, the ZTX758, and the n-p-n ZTX658 are a complementary pair with 400V C-E breakdown and capable of 0.5A continuous current handling. Maximum pulsed forward current is 1A and collector cut-off 100nA. The pair are completely identical and have a 50MHz transistion frequency with a static gain of 50 at 100mA and 5V supply. Saturation voltage at 50mA and 5mA base current is 0.25V or less. Zetex plc, 061 627 4963.

Linear integrated circuits

Fast video op-amp. AD811 is claimed by Analog Devices to be the fastest IC op-amp working from ±15V and has been tailored for application in the video field; for example, it will drive two back-terminated 75Ω cables. It is a current-feedback type with its gain flat to within 0.1dB over 35MHz at a gain of 23dB bandwidth being 120MHz at that gain. Differential gain and phase are 0.01% and 0.01° and THD at 10MHz is –74dBc. Voltage noise is 1.9nV/ Hz and there is a maximum of 3mV offset. At a slew rate of 2500V/µs and settling time of 25ns to 0.1% for 2V, high-speed pulsed applications are easily handled. Analog Devices, 0932 232222.

Fast, high-current op-amp. OPA671 is an fet-input op-amp which exhibits 107V/µs slewing and an output current of 50mA; it is intended chiefly for line-driving. Bias current is only 5pA but,at 50mA output current, provides ±10V. Settling time is 240ns to within 0.01% and GB product is 35MHz. THD at 100kHz is 0.0006%. Burr-Brown International Ltd, 0923 33837.

One-chip car audio. All the functions of car audio, including stereo decoder, audio processing, ignition noise controller, fast-tuning synthesiser and data conversion are contained in ITT's CAP IC. Processing is carried out by a 16-bit, 15Mips DSP and the main part of the digital circuitry consists of A-to-D-to-A conversion with AM and FM synthesis. One A-to-D converter is intended to cope with ARI/RDS signals.ITT Semiconductors, 081-390-6578.

Stereo D-to-A. For digital audio use, CS4328 from Crystal Semiconductor takes data at standard frequencies of 48kHz, 44.1kHz and 32kHz and outputs line-level analogue signal into 600Ω. The chip contains eight digital interpolation filters, analogue L-P filters, drivers, voltage reference, clock and a calibration microcontroller. Nothing else but a few decouplers are needed. Dynamic range is over 93dB and idle-channel noise figure is 126dB.Sequoia Technology Ltd, 0734 311822.

Fast-slewing current amplifiers. Dual and quad SM current-feedback amplifiers from Linear Technology, the LT1229 and LT1230, are made in a complementary bipolar process which confers almost identical characteristics on high-frequency (600MHz) p-n-p and n-p-n transistors. Over a wide range of closed-loop gain, bandwidth and slew-rate remain "fairly constant" at 100MHz and 1000V/µs; noise is 6nV/ Hz. For cable-driving. differential gain and phase are 0.04% and 0.1°.Micro Call Ltd, 0844 261939.

"Sleep-mode" op-amp. MC33102 is a dual op-amp which employs Motorola's new low-current standby facility. In Sleep-mode, current drain is 45µA per amplifier; when a signal is applied, the amplifier rouses itself in 4µs for an output current of more than 160µA, going back to sleep when the current drops below that level. The amplifiers still function when apparently asleep, in a micropower condition.Motorola Ltd, 0908 614614.

Satellite TV ICs. Four new ICs from Siemens are intended for satellite television receivers. Two of the devices are for the tuner section: the SDA6102X 2.3GHz phase-locked loop giving fastlock on changing channels and the 300-900MHz FM demodulator TDA6140X. The other two compose the main receiver; a MAC/D2MAC/PAL video processor with external de-emphasis for system selection, the TDA6151X, whose sound output goes to the TDA6160X three-channel sound processor for demodulation.Siemens plc, 0932 752623.

Analogue comparator. Harris Semiconductor's HFA-0003L is claimed to be the first comparator to have programmable hysteresis up to 20mV by means of a single resistor or by digital control. This device is pin-compatible with standard types, but exhibits a propagation delay of 2ns, voltage offset of 1mV and tracking bandwidth of 270MHz. Open-loop gain of 70dB eliminates any need for gain stages and the output drives a 50Ω line directly.Thame Components Ltd, 0844 261681.

TV IF amplifier. Both sound and picture IFs are contained in the

Memory chips

4-Mbit eeproms. Arranged as 512K by 8bit, Atmel's AT28MC040 has an access time of 150ns, dissipating only 440mW when active and taking 5mA when deselected. Packaging and pin-out are compatible with previous eeproms. The device offers internal write-inhibit circuitry to protect data from the effects of noise.Atmel (UK) Ltd, 0276 686677.

TA8600N bipolar chip from Toshiba. Picture stages include a three-stage IF with fast AGC, a carrier-synchronised detector and switchable black/white noise inverter. For sound, there are three intercarrier IF stages and detector providing a THD of 0.75% or less and independent sound AGC.Toshiba Electronics (UK) Ltd, 0276 694600.

Mixed-signal ICs.

Video display processor. Yamaha's V9990 enhanced video processor has separate modes for games, audio-visual and office equipment. It provides high-speed drawing and animation and will also support PC crts, LCD panels and ordinary TV receivers. For games there are 512 by 212 and 256 by 212 modes; for A-V, four-bit map display modes for use in NTSC or PAL monitors from 256 by 212 to 768 by 240; the high-res. mode is 640 by 480. Superimposition, digitisation and omnidirectional smooth scrolling are supported. Barlec-Richfield Ltd, 0403 50111.

Optical devices

Optocouplers. Four new devices from Celdis complete a range of 8-pin couplers. All in the HCPL-XXXX range, the 5530 and 5630 are dual-channel versions and the 5500 and 5600 single-channel types. Speed of the 5530 and 5500 is 400kB/s, the other two being 10MB/s devices. Bandwidth is 9MHz and test voltage 1.5kV. All are for commercial use but are certified to MIL-STD-1772. Celdis, 0734 585171.

Oscillators

Oscillators. Hermetically sealed oscillators about 13mm square and working at frequencies from 1.5kHz to 40MHz are available from Salford. Versions for up to 80MHz working can be supplied at a restricted temperature rating. The half-dip, 4-pin units are TTL or HCMOS-compatible up to 40MHz and above that can be ACMOS-compatible. Output is a rectangular wave of 60:40 M/S ratio.Salford Electrical Instruments 0706 367501.

Power semiconductors

Solid-state relays. Three-phase, solid-state relays from Crydom in the 53TP series will switch line voltage to loads. Inputs an doutputs are opto-isolated to 4kV RMS and control voltage may be 90-280V AC or 3-32V DC. Current ratings are 25A or 50A at 25-65Hz.Unitel, 0438 312393.

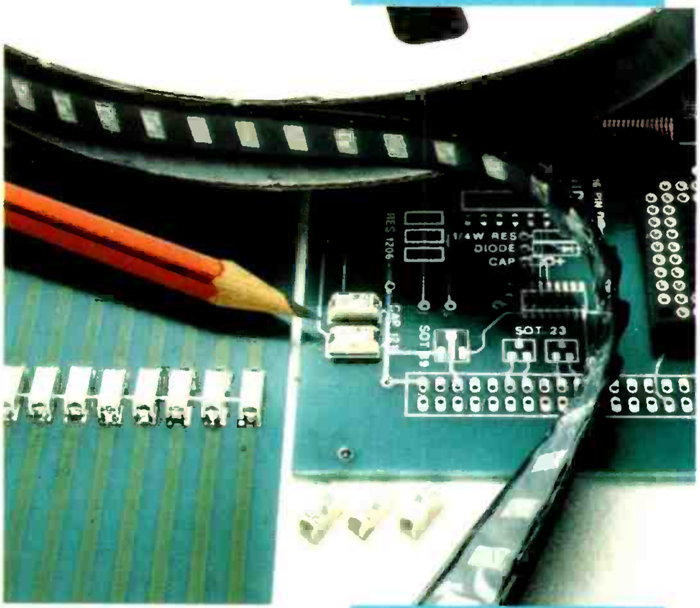
PASSIVE

Passive components

Toroidal transformers. From 30VA to 530VA and with dual or single tapped primaries, the Avel 4000 series of toroids also cover a wide range of output requirements: from 6V to 240V RMS. Secondaries are outside primaries for reduced fields and there is a thermal cut-out. The transformers meet all the relevant international standards.Avel-Lindberg Ltd, 0708 853444.

Displacement transducer. LPW is a family of resistive, linear-motion transducers for measuring mechanical displacement in the range 10-50mm. For both very slow and high speeds of movement, the wiper is a bunched type, which can handle up to 0.2m/s. Mechanical life is 1million strokes and hysteresis is better than 0.05mm.Kynmore Engineering Co. Ltd, 071 405 6060

Digital oscilloscopes. Two and four-channel versions of LeCroy's9300 range of 300MHz oscilloscopes are each available with threememory configurations: 10K, 50K and 1M. Independent 100Ms/s digitisers are on all inputs. The instruments offer fast autostepping, sequence mode, automatic pass/fail testing, a variety of trigger modes, signal processing and fast Fourier transform analysis. Each incorporates a PC-type memory card. LeCroy Ltd, 0235 533114



Surface-mount fuse. From Radiatron, the Schurter fuse is a quick-acting type, is surface-mounted and is said to be the smallest S-M type available, measuring 7.4 by 3mm by 2.55mm high.Ratings from 63mA to 2A are produced, all having a 63V rating and breaking 50A.Radiatron Components Ltd, 081 891 1221.

Wirewound chip resistors. WSC surface-mounted wirewound chip resistors now includes a 0.5W unit, as well as 1W and 2W devices, of values down to 0.005Ω and 0.05% tolerance. Temperature coefficient is ±20PPM/°C.Vishay Components (UK) Ltd, 091 514 4155

Connectors and cabling

Waveguide-to-coax transition. On occasion, it is necessary to take the output of a magnetron into coaxial cable; for example, in medical diathermy, catheters and other applications, compactness and flexibility are needed. NL10316 is a brass transition for CW magnetrons at 2450MHz, which consists of a closed-end waveguide with a hole through which the magnetron antenna protrudes. Output comes from a European metric 7/16 female connector capable of handling 860W into 50Ω.Richardson Electronics Ltd, 0522 542631.

Displays

Vacuum fluorescent displays. On-board microprocessor control, interface logic and DC-to-DC conversion are among the features of the T range of 5 by 7-dot displays from ESD, which now has five new members. The TTL-compatible interface and high-speed parallel bus assists interfacing with a host system, and up to 12 European character fonts are software-defined. ESD Mercator, 0493 844911.

Hardware

Extender board. PS2 extender boards for the PCbus from BICC-Vero fit any spare expansion slot on a PC mother-board, raising an expander card above the rest of the circuitry. Cards can be inserted live and the board design is such that cross-talk between the extended conductors, which are of equal length, is at a minimum. BICC-Vero Electronics Ltd, 0489 780078.

Instrumentation

60MHz oscilloscope. An oscilloscope from Sampo, the SSI-2360 dual-trace instrument, has its own component tester for resistors, capacitors and semiconductors. The salient details of interest are 1mV/div vertical sensitivity, 5ns/div maximum sweep rate, delay sweep, single sweep, beam finder and scale illumination. ECW Instruments, 0376 517413

Clamp-on ammeter. HEME's 2000AC is a simple current meter using the Hall effect to measure current without breaking the circuit.Three autoranged scales of 20A, 200A and 2000A true RMS, from 15HZ to 3kHz, are presented on a 3 1/2-digit LCD, which holds the reading for ten seconds after the clamp has been removed. The instrument meets IEC 348 safety standards. HEME International Ltd, 0695 20535

Optical-fibre power meter. Single-channel power meter, the S370 from Optilas, is programmable for many types of power measurement either by IEEE interface or from the panel. As well as log. and linear absolutes, measurement may be relative to a reference level set by the user, either linearly or in dB. Calibration data for given detectors can be kept in a plug-in rom or the user can program responsivity for detectors. Optilas Ltd, 0908 221123

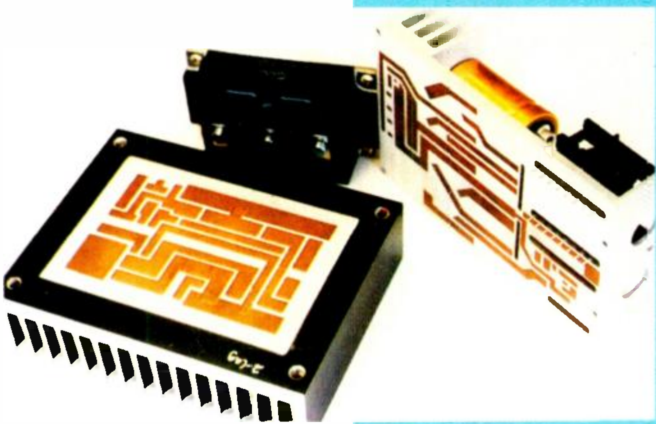
Test receiver. Three types of instrument are gathered into one in the ESVD: GSM field strength and RFI measurement, manually tuned test receiver and automatic test receiver to measure EMI and report results. The equipment covers 20-1000MHz (up to 2000MHz with an extension) and will take an I/Q demodulator to measure transmission quality. Field strength measurement can be automatically made at four frequencies of the GSM network while the test vehicle is travelling at over 60mile/h. Rohde & Schwarz UK Ltd, 0252 811377.

Digital fluxmeter. Handheld fluxmeters from Tandem measure AC or DC magnetic fields from 0.01mT to 2T, using a GaAs probe with good temperature stability. Four models comprise the range and are AC, DC and AC/DC types with 3 1/2 digit displays and an AC/DC model with 4 1/2 digits. Maximum error is less than 3.5% of each range.Tandem Technology Ltd, 0243 532766

Literature

Instrumentation amplifiers. Basic theory, design requirements and application are all explained in Analog's new Instrumentation Amplifier Application Guide. Two appendices examine the specifications met with in this type of circuit and there is a selection chart of the company's amplifiers. Analog Devices, 0932 232222.

UPS guide. In an effort to clear a way through the uninterruptible power supply undergrowth, ERA Technology has published a 250-page guide to the UK companies who make UPSs and power-conditioning equipment. As well as the guide to suppliers,there is a section on power-supply problems and the characteristics of the UPS. Cost is £85 or £75 to ERA members. ERA Technology Ltd, 0372 374151 ext.2234



Power supplies

Wide-input DC-to-DC. PT4900 DC-to-DC converters are designed to accept a varying input voltage and to produce a regulated 5W output in one of three ranges: 9-18V, 18-36V and 36-72V. Line stabilisation is ±0.1% and regulation is ±0.2%. Models are BT-approved for some applications. Dowty Power Electronics, 0722 413060

DC-to-DC converters. Linear Technology have two more micropower converters, giving 5V and 12V or adjustable output. LT1110/1111are in 8-pin dips or SO packs and, since they work at 70kHz, accommodate surface-mounting capacitors and inductors. Quiescent current is 300μA and they operate from as low as 1V (1110) or 2V (1111). Step-up or step-down modes are available and the devices will generate 12V at 120mA for four flash memories. Low-battery detectors are included. Micro Call Ltd, 0844 261939.

Radio communications products

Satcom synthesiser. Sciteq have introduced the VDS-6000 L-band synthesiser which employs the new arithmetically locked loop (ALL) technique. Spurious responses are much less than -60dBc and phase noise at 1kHz offset is -85dBc/Hz. With the addition of a tripler, you get a C-band synthesiser, the VDS-6030, operating inthe 5GHz band. Step size here is 3MHz, against the 1MHz of the VDS-6000. Lyons Instruments Ltd, 0992 467161.

Saws for superhets. Surface acoustic-wave devices from Quantelec for dual-conversion superhets are on offer, to the UK and European approved frequencies of 418MHz and 433.92MHz. They are intended to make it a little easier to use superhets in paging links for remote control systems, where they have previously been bulky and expensive. Each consists of a SAW oscillator and coupled-resonator filter and only a front end and standard 10.7MHz IF are needed to form a complete receiver. Image rejection is provided by the filter and the resonator is within ±0.1% of specification. Quantelec Ltd, 0993 776488

Materials

Power substrates. A new method of manufacturing power-hybrid substrates confers design freedom and miniaturisation. In the method developed by the Norwegian company, the ceramic substrateis sprayed directly onto the metallic base such as a cooling fin at 20,000°C to give a thickness of 0.3mm in 60 passes. Since theceramic is amorphous, it is now flexible and the copper conductors are then sprayed onto the ceramic. Production is to start in early 1993.Compact PowerCircuits a.s, 010 47 33-477 00.

Switches and relays

Miniature relay. One of the smallest relays currently obtainable, the TQ is only 14 by 9 by 5mm and has two changeover contacts, either monostable or bistable. The TQ switches 100W DC and its coils needs 140mW or 280mW, depending on contact arrangement. Plastic sealing allows automatic washing.Matsushita Automation Controls 0908 231555.

Supermini switch. For use where space is restricted, Matsushita's miniature snap-action switches come in pin plunger, hinge lever or simulated roller lever forms and are for PCB mounting. At 7.5mm by 4.5mm by 2.5mm, they are claimed to be the world's smallest. Typical life expectancy is 300k operations at 60/min and the switches handle 5mA at 6V DC, 2mA at 12V or 1mA at 24V, all on resistive loads.Verospeed, 0703 641111.

Transducers and sensors

Accelerometer. The 3145 piezoresistive signal-conditioned accelerometer is claimed to be the smallest of its kind, being 23mm square and 13mm deep. It weighs 13g and will cope with ±2g to ±100g full scale. In the module are a silicon micromachined element, an amplifier, signal conditioning and temp. comp. from -20° to +85°C.EuroSensor, 071 405 6060.

Cable-pressure sensor. A TO-5 silicon sensor responds to pressure drop in pressurised telephone cables. It works over 900-1750mBar with a linearity of ±5mBar when used in the Drallim addressable transducer interface unit (TU). One pair of conductors serves up to 100 TUs, providing both power and interrogation signal. EuroSensor, 071 405 6060.

Contactless linear sensor. For noise-free operation, the Model LP-3U linear-motion sensor uses no contacts and is housed in a cast-iron shell, which also provides protection from life's little knocks. There are two poles to make the output oblivious of lateral motion and minimum sensitivity is 8% of the 14V DC maximum applied voltage per millimetre. Input impedance is 25kΩ. Kynmore Engineering Co. Ltd, 071 405 6060

Temp/humidity measurement. An instrument to determine and record minimum/maximum relative humidity indoors is offered by Maplin at £15.95. It covers 25% to 95%RH in 1% steps at a sample rate of 1/second. No external probe is needed and there is BCD output. A second instrument combines the above with a min/max temperature meter for £18.95. Maplin Electronics plc, 0702 554161

Temperature transmitter. Claimed to be the world's smallest smart industrial temperature transmitter, the TR55 replaces the commonly used analogue components of these instruments with digital technology and also provides digital communication, while still using only the usual two-wire link, which powers the equipment. It can be reprogrammed by a PC fitted with a small interface, all parameters being read on the serial link. Output in 4-20mA or 20-4mA form is programmable. In RFI levels of 30V/m from 20 to 1000MHz, error is less than 0.1%.Thermocouple Instruments Ltd, 0222 734121

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COMPUTER

Computer board level products

64-channel output for PCs. An output board for use with PC XT/ATs has 64 individually programmable, 8-bit D-to-A converters. Each of the converters, which are fast enough not to require wait states even in high-speed machines, provides a 0-2.5V, 0-5V or 0-10V analogue output via one of four 26-pin connectors. The PC64 board scans channels automatically, updating the channel number each time a value is written to a converter, simplifying software drivers and making for quicker updating. Driver software is in C and QuickBASIC.Amplicon Liveline Ltd, 0273 608331.

Industrial PC AT board. A 386SX on an STEbus board provides a PC AT-compatible computer adapted for the industrial environment and with a potential 21 expansion slots. Up to 20Mbyte of dynamic ram is usable, with UNIX in mind. Since the graphics function is an add-on module, the user has the option of display type — crt, LCD, plasma or electroluminescent VGA drivers on the SCIM bus — or none at all, if the application needs no user interface. SCIM386SX is meant for embedded use. Arcom Control Systems Ltd, 0223 411200.

17-bit converter card. A multi-function card from Blue Chip has a selectable resolution of 17bit for both input polarities from 16 analogue inputs, common-mode isolated to 400V. Software selects four input filters, 50Hz, 60Hz, 100Hz and 400Hz. Selectable unipolar/bipolar analogue outputs are provided, as are 16 digital inputs and outputs. Blue Chip Technology, 0244 520222.

8051 micros. Mini and Micro modules made by Phytec are now available here. Minis use the 8051 family microcomputers and are intended specifically for control application. Smallest in the range is only 1.5in by 1in and will host either a BASIC interpreter or a C assembler/monitor with 128Kbytes of memory. MiniModules use the 80C537 or 80C552 and are credit-card-sized, offering the same facilities and up to 256Kbytes of memory. .Hitex (UK) Ltd, 0203 692066.

Development and evaluation

Smallest in-circuit emulator. P-ICE is claimed by ARS to be the smallest ICE in the world at 4in by 3in by 1.25in, emulating 8051, 8031 and 8032 families in real time at clock speeds of up to 16MHz. Command-line structure allows control of emulator, windows and files and will perform loops, conditional control, expressions and line editing. Three complex events can trigger breakpoints and up to 64K software breakpoints can be set. The unit is complete with 128Kbyte of emulation memory, 4K trace buffer and all necessary hardware and software for immediate use.ARS Microsystems, 0276 685005.

Computer peripherals

286 accelerator. FASTCache-SX Plus is a 25MHz board designed to bestow 386 performance on the 286 processor and to give an extra 32Mb of memory. It uses a 386SX processor and allows users of PCATs to run 386-based multi-tasking and 32-bit protected-mode code, as in Windows 3.0, DesQview, Unix and OS/2. Microway say that most applications will be 3-5 times faster and that the optional 387SX increases speed to 5-10 faster than a 287. Microway (Europe) Ltd, 081 541 5466.

Software

DADiSP 3.0. As it presents scientific analysis in a variety of forms simultaneously in a series of windows, this new version of DADiSP is justifiably said to be a graphical spreadsheet. Data can be captured from a number of sources, displayed in any relevant manner in any of the windows, results of calculation or data manipulation in one window updating the displays in all the others. It operates under X-Windows on Sun, IBM, DEC, H-P and Concurrent and on PC ATs and PS/2. There is now mouse support in X-Windows and PC systems and more table-editing facilities, as well as a range of matrix functions. Adept Scientific Micro Systems 0462 480055 ■

A SOUND MODEL FOR AUDIO DESIGN?

Developing rugged power amplifiers for audio and industrial duties involves taxing measurement and analysis. PC-driven analogue simulation can play a valuable role in analysing sub-circuit operation and, with limited physical measurements, is a valuable tool for the designer, claims Ben Duncan.

Anyone who has worked with multiple bipolar power devices and vertical (switching) mosfets will know the sinking feeling of vaporising expensive silicon as early attempts to verify protection or operating stability fail.

Viewing behaviour during realistic output shorts, plotting the V-I demand of real loads against SOA, and finding out what happens at high junction and ambient temperatures are examples of tricky evaluations that benefit from PC simulation, particulary in the initial stages, or when 'reverse engineering' is called for. This article covers the first two, as well as power supply modelling and 'virtual impedance measurement', using MicroCAP-III simulation program[1]. With any analogue simulator, the essence of efficient use and trouble-free passage is to simulate just the bits you need to know about, and to build up models one step at a time.

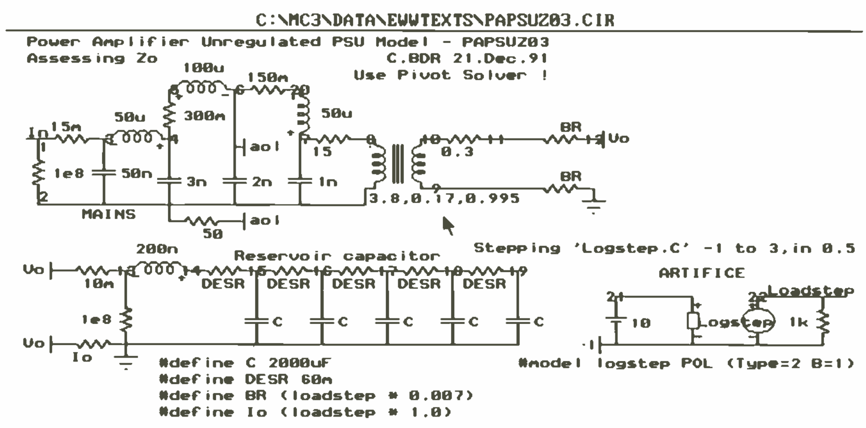


Fig. 1. Simplified power supply model. The importance relates to intermodulation products arising from interaction between audio output stage and supply source impedance

Raw Supply Analysis

The majority of audio power amplifiers employ raw, unregulated power supplies, at least for their output stages. And the majority of power output stages use emitter or source followers – whose performance is critically dependent on being hung from a low supply impedance.

Impedance is readily measured at a spot frequency by observing voltage droop with a given load. However for rails greater than 50V with impedances below 1Ω, and considering dissipation limits in the load test resistor, readings are increasingly obscured by ripple and noise. To complicate matters, raw supply impedance depends on load current and is liable to vary widely across the four decades of frequency that are relevant to audio.

The circuit in **Fig. 1** models a typical low cost power supply. We will use it to make a first order evaluation of the supply impedance vs frequency vs output current, as seen by the output devices. Beginning at the top left is the typical source impedance of the mains where it enters the building (node 3). The first RLC section that follows covers the mains cable, from the fusebox to the wall outlet. At this point, a tie 'aol' signifying 'all other loads' connects 50Ω across the supply in shunt with 3nF. This roughly models the rest of the ring main, with a nominal cable capacitance, and a few electric heaters and/or light bulbs. The second RLC section (nodes 6 to 7) picks up the amplifier's supply cabling up to the transformer.

The transformer model was derived by measuring a sample, in this instance a 60VA toroid. The primary and secondary winding resistances are represented by the resistors in series with the transformer's input and output. The remainder of the transformer's first order specification is built into the definition underneath it, which details primary (LP) and secondary inductance (LS), and efficiency, ie. LP = 3.8 Henries, LS = 170mH, and 99.5%. After measuring the former, the latter can be derived from the frequency response, which is readily measured and translated by repeated AC simulation, until the simulated response and other test conditions match the physical measurements.

Plotting impedance against frequency comes under AC analysis in MC3 and other simulators. This is a small signal simulation, so the

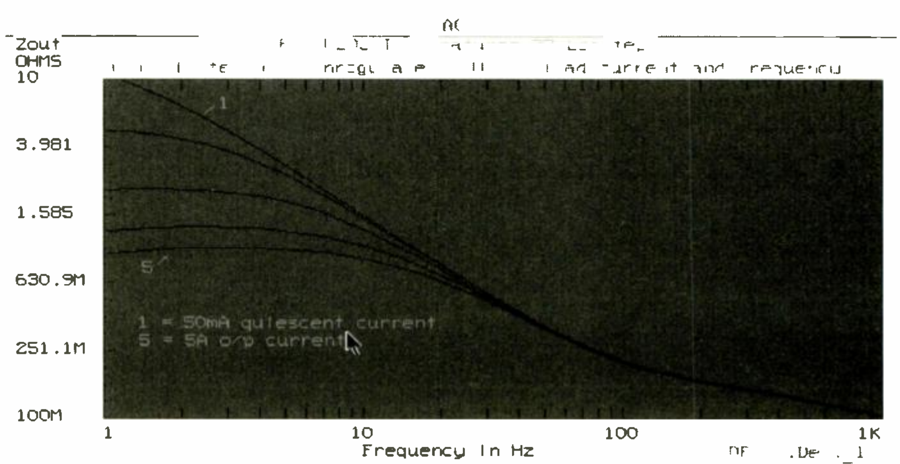


Fig. 2. PSU source impedance vs frequency as modelled at output node V_o. The main problems are likely to be at low frequencies where transformer source impedance is significant

diode bridge has to be reduced to the diode's first order, small signal equivalent – a resistor. As only two diodes are conducting at a given instant, the bridge reduces to two resistors (BR). Although nearly all power amplifier supplies employ bi-phase rectification with a centre-tapped secondary, a single rail, full wave rectifier configuration with single secondary is being dealt with for simplicity. The dependence of the diodes' dynamic resistance on the load current (set by resistor I_0) is handled by stepping both resistors' values in tandem over a typical output current span, ie. 50mA up to 5A. To provide salient information in a few, uncrowded curves, this is done in log steps.

The network underneath the mains and transformer parts mainly represents a single, 10,000µF reservoir capacitor, modelled as a transmission line. On the far right, the 'artifice' comprising a 'PolyVsource' or polynomial voltage source and associated components govern the tandem log stepping of the resistors I_0 and BR. Their different ohmic values are defined in the expressions underneath the reservoir capacitor, by multiplying the 'loadstep' number driven by the simulator's stepping function. The reminder 'use pivot solver' on the screen warns that a reactive network of this size is relatively tough for the matrix solver to get going, though once away, it runs fast. When selected, MicroCAP's pivot solving routine helps initial convergence and avoids erratic plots.

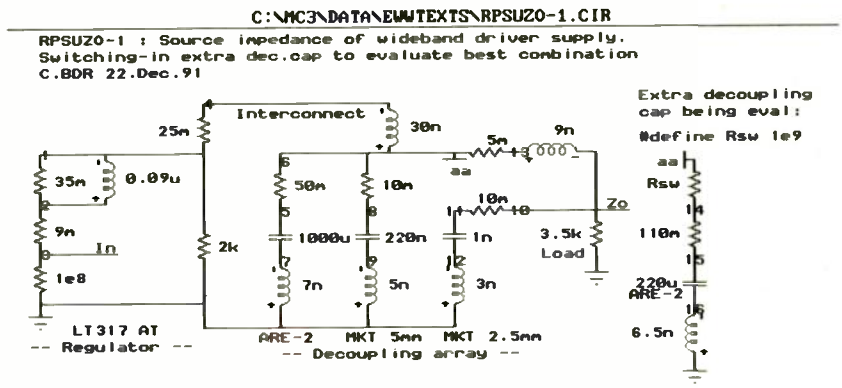


Fig. 3. Regulator model based on the LM317K three-terminal variable component

capacitance with a pulse repetition period that varies with loading and mains waveform purity. While beyond the scope of AC analysis, the moment-by-moment variation can be derived from transient analysis. Secondly, the effect of the transformer's level-dependent losses (notably saturation) are notcatered for. This requires MicroCAP's core modelling process which is beyond the scope of this piece. Resistance modulation in the primary and any secondary fuses isn't covered, but can be included as resistance using the 'loadstep' expression to approximate the variation.

Beyond this, the model has quite general application: by replacing BR with a bridge of representative diodes, you can run a transient analysis and examine start-up behaviour,

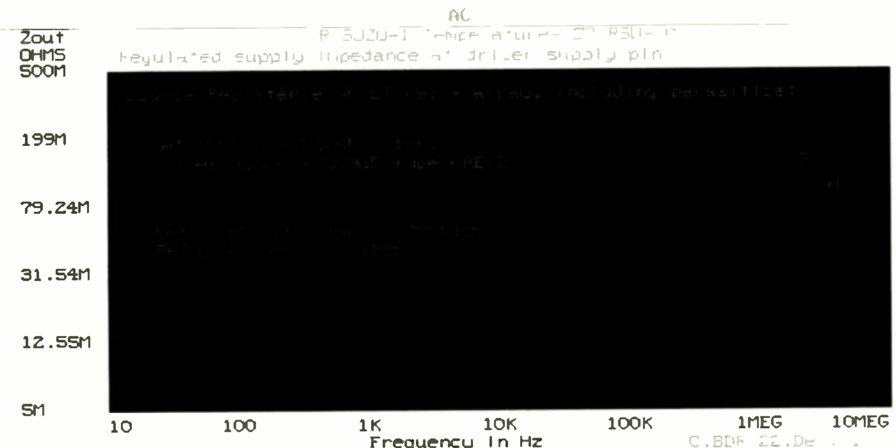


Fig. 4. The output impedance curve derived from running the regulator model shown in Fig. 3.

Figure 2 shows the source impedance vs frequency and loading, reading at the output tie 'V_0'. It shows how impedance under quiescent conditions varies between just over 10Ω at 1Hz down to 100mΩ at 1kHz. The curve is similar to a reservoir capacitor on its own. With increasing current, the low but inductive impedance of the transformer windings is increasingly 'seen' at low frequencies, as diode incremental resistance reduces.

The model gives valuable information but it's important to recognise the shortcomings. First, it neglects the fractional conductance period of the diodes. The real impedance fluctuates between the extremes twice in every cycle. In effect, 100/120Hz ripple modulates the background impedance of the reservoir damping with mains transients and interference rejection, as well as displaying more traditional data, such as the ripple current. This would be measured by plotting the current in the 10mΩ resistor between nodes 12 and 13. Steady state conditions are arrived at by either doing an initial run, and saving the values for the next run, or by simply pre-charging the reservoir. This is done by changing the voltages in the initial values editor, entering +55v on nodes 15 through 19.

Voltage regulator analysis

Driver stages for wideband linear power amplifiers need a supply impedance below 1Ω, up to low RF, as well as being regulated. Performance is particulary important when using high speed (>300V/µs) and wideband (>100MHz) op-amps. But the solo impedance of IC regulators isn't impressive above 1kHz, as gain-bandwidth product is low. Suitable decoupling capacitors help maintain low impedance to VHF, but a bad combination can aggravate existing resonances.

The circuit in **Fig. 3** began as a model of the regulator's output alone, an adjustable three-terminal IC type, the Z_0 of which is commendably low at 50Hz, yet rises markedly above 1kHz. The AC analysis reference input is across the 1e8 resistor on the left. The regulator's Z_0 at node 1 is simulated by the two resistors and inductor beneath. The component values are derived from makers' impedance data. As Z_0 varies over decades of load current, the values will need slight adjustment for other conditions.

The regulator's own Z_0 is followed by the connector and PCB track resistance (25mΩ), and the track and wiring inductance (30nH), up to the local decoupling capacitors. There are three principal ones (1000µF, 220nF and 1nF), and a fourth unit is being evaluated on the far right. For detailed accuracy above 20kHz, the 1000µF should strictly be modelled as a transmission line[1]. This isn't so important for the other units provided they are of multilayer construction.

Physically, the 1nF is located against the IC power pins and is several millimetres from the larger capacitors. Accordingly, 5mΩ + 9nH worth of PCB track is inserted between the 1nF and the latter. Source impedance is read across the load (node 10, alias 'Z_0').

Looking at the plots in **Fig. 4**, the benefit (or not) of adding to the decoupling array is shown by running a second plot with the extra decoupling switched in. Rsw (top right) is the switch, stepping from 0 to 100MΩ. In this instance, the added part produces a slight Z_0 reduction up to the first resonance at 3MHz, where it provides some damping.

Having turned preferred capacitor types into macros, the way is now open to assess which combinations keep impedance suitably low, and comparing different layout schemes. If VHF parasitics are intensively modelled up to 100MHz or more (according to transition speeds), the simulation relevance extends to the design and troubleshooting of digital supply decoupling.

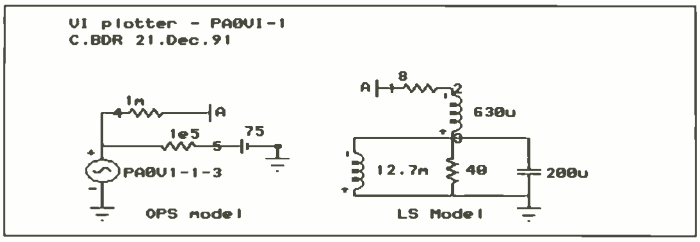


Fig. 5.(above) plots the V-I demand of a given load with a sinewave (or any other) stimulus, over a wide range of frequencies. The sine generator 'PA0VI-1-2' is set up to sweep beginning at 10Hz at T=0, then increasing at 2Hz per millisecond.

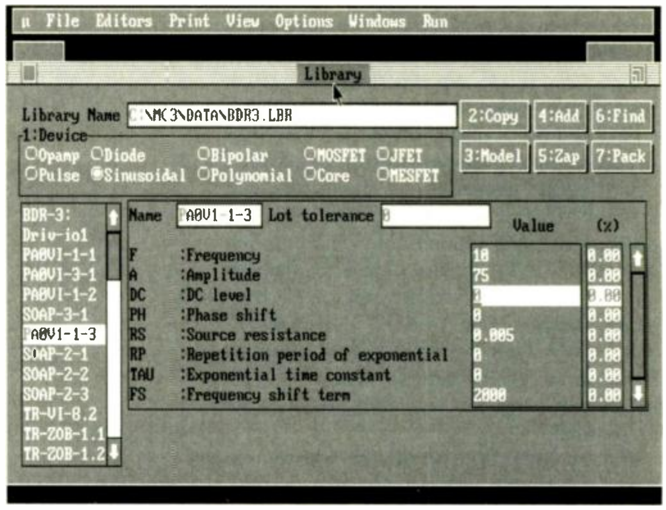


Fig. 6 (right) shows the settings panel in MicroCAP's 'VSI' (voltage sinusoidal) library. Peak voltage swing is 75V, corresponding to a nominal 755W into 4Ω, while the frequency shift function is set at 2000Hz/s.

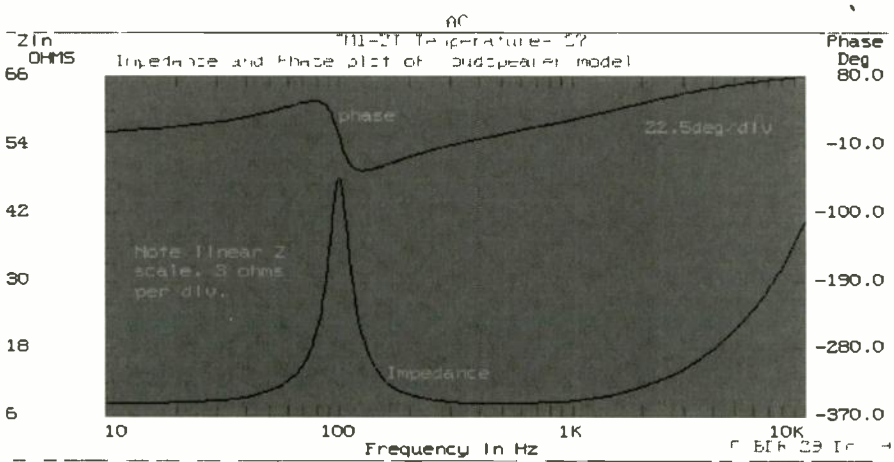


Fig. 7 shows the simulated impedance and phase response. The components (and effects) of a crossover network have been excluded from this loudspeaker model.

The demands of real loads

Most transducers present a complex load and loudspeakers are no exception. Energy storage results in out-of-phase conditions where substantial current flows while the output voltage is low or nil; the output device(s) are sustaining high voltage simultaneous with passing current. If the output devices don't first die from second breakdown, the drastic rise in dissipation may seal their fate as safe operating area shrinks with the increased junction temperature.

A valuable bench test-set for various V-I measurements associated with power amplifiers has been described by Peter Baxandall[2]. However, as presented, it is limited to giving a real time response (on a 'scope) in response to a periodic signal. By contrast simulation can show moment-by-moment transient behaviour.

Figure 5 plots the V-I demand of a given load with a sinewave (or any other) stimulus, over a wide range of frequencies. The sine generator 'PA0VI-1-2' is set up to sweep beginning at 10Hz at T=0, then increasing at 2Hz per millisecond. **Figure 6** shows the settings panel in MicroCAP's 'VSI' (voltage sinusoidal) library. Peak voltage swing is 75V, corresponding to a nominal 755W into 4Ω, while the frequency shift function is set at 2000Hz/s.

A negative entry would sweep downwards, but loudspeaker driver's V-I demand is generally most extreme at the lower end of their range. The generator source impedance is set at 5mΩ to approximate a properly executed feedback amplifier output impedance, at least below 1kHz.

Neglecting Class AB bias, the sum of the collector current in the output devices is the same as the output current, and is monitored across the 1mΩ resistor. The voltage we need is the output device V_{ce}, which is the difference between one supply and the output. The 75V battery simulates the supply, and V_{ce} is implied by reading the voltage across the 100MΩ resistor. The value chosen is arbitrary, so long as it doesn't draw appreciable current. On the right, a typical 30cm, full-range drive unit has been modelled from its impedance data. **Figure 7** shows the simulated impedance and phase response. Note an offset linear scale has been selected for impedance. For simplicity in this example, active speaker drive is assumed, ie. without an intermediate passive crossover network. The transient analysis in **Fig. 8** uses MicroCAP's ability to plot any cardinal physical quantity against any other, in this case I_c (at nodes 4,1 in Fig. 5) vs V_{ce} (at nodes 5,4). To see how the simulation is progressing, drive voltage vs time is plotted on a small scale underneath. The signal frequency can be seen rising at 2Hz/ms, finishing at (10+150) = 160Hz. At T=0, the V-I loop begins at the co-ordinates 0A,75V marked 'start'. Being a transient analysis, some set-

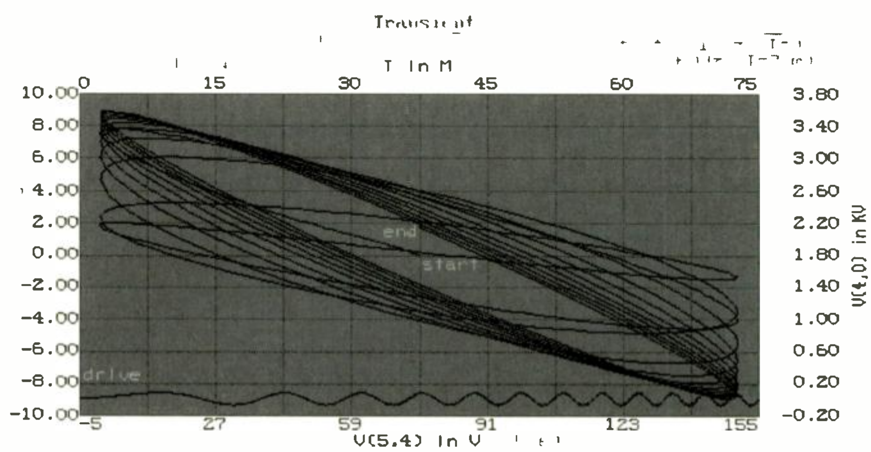


Fig. 8. V-I demand of an 8Ω drive unit viewed in dimensions of voltage, current, time and frequency. Lower sinusoidal plot shows the 75V peak-to-peak drive sweeping during the run from 10Hz at T=0 up to 160Hz at the end (T=75ms). Scales are the top x-axis (time/frequency in ms) and lower right y-axis (amplitude in kV). The small scale is used to fit this plot into the screen. During the run, the V-I ellipse (bottom y- and left x-axis) develops from "start" to "end" using the simulator's ability to plot quantities other than time in transient analysis. The ellipses' position at a particular drive frequency can't be retrieved from this snapshot, but is viewable as a dynamic, numeric readout during the run.

tling is to be expected, so the initial vector is an exception, after which the looping weaves a 'strange attractor' that fills the centre and describes the familiar reactive load parallelogram[2]. At around 25 to 30ms (55 to 70Hz), a reversal takes place, visible as the flat loop lying above 'start'. Answers on a postcard please! As for the V-I limits, the peak demand is +4A at 75V, or 300W instantaneous dissipation.

For device SOA, this is what counts at low audio frequencies although the real (integrated) dissipation is half. Don't be fooled by the apparent 150V (ie. double supply voltage) at 7A in the bottom right corner. Assuming the amplifier is push-pull and operates in Class A-B, this corner represents the current in the –ve device, where its own V_{ce} is 0V while the +ve side device isn't supposed to be passing current (other than bias). For device dissipation, the most critical zone is generally the current around the centre where $V_{ce}=V_{max}=V_s=75V$. We will return to this shortly.

For more detailed work, refinements include modelling the supply frequency-dependent source impedance, output stage feedback, temperature and frequency dependent source impedance and corresponding voltage loss. You may also want to include passive crossover networks, the amplifier output zobel network and the speaker cable characteristics.

Figure 9 models the latter two. A 30m length of cable is simulated with three lumped RLC sections, while the zobel values are typical of commercial designs. With the circuit of Fig. 8, adjusting the generator to sweep from 1kHz up to 20kHz shows that a nominally unloaded amplifier driven at 20kHz has to source several amperes[1] at maximum V_{ce}. The zobel network is the prime culprit.

So a typical cable draw is negligible. But what happens if the open ended cable is momentarily shorted? In Fig. 9, the generator sweeps up from 1kHz, and a time switch has been added to model the short, which lasts for just 900μs. Also, the cable termination, otherwise set arbitrarily high, has been set to 10MΩ for realism. **Figure 10** shows the outcome.

During the short, the current peaks (as you would expect) at tens of amperes. And what happens after the short ceases? The cable sets off a lightly damped oscillation. For clarity, the plot was curtailed at 4ms, but the oscillation hypothetically goes on for several tens of milliseconds assuming the amplifier lives that long...

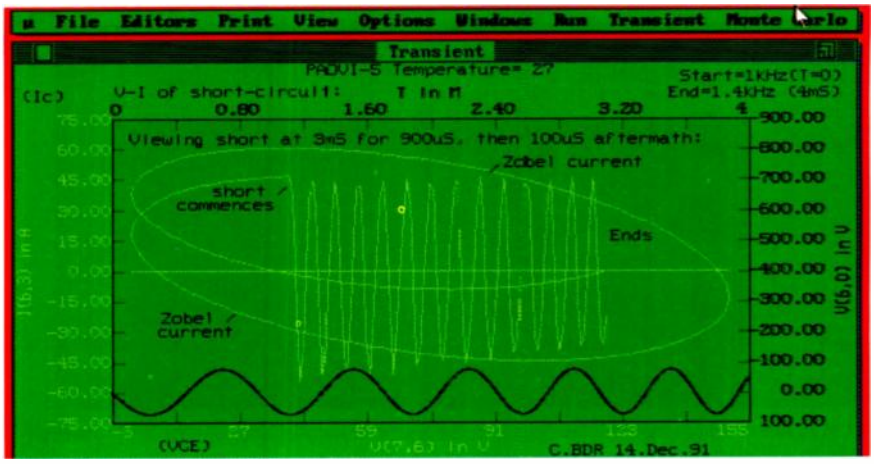


Fig. 10. Will an output short circuit blow the output transistors? This set of curves simulates a momentary speaker cable short. Similar to Fig. 8, this multi-dimensioned transient analysis shows drive voltages vs time (lower plot, top axis and right y-axis) sweeping up in frequency from 1kHz to 4kHz in the 4ms simulation period. It also shows the resulting V-I demand during this period (upper plot, lower x-axis and left y-axis). The V-I plot begins as the central horizontal line – on a larger current scale, it is really a shallow ellipse. At 3ms, the instant or shorting, V-I relations begin describing the elliptical attractor. As the short is removed at 3.9ms, the V-I demand breaks into oscillation up to the point marked "end"

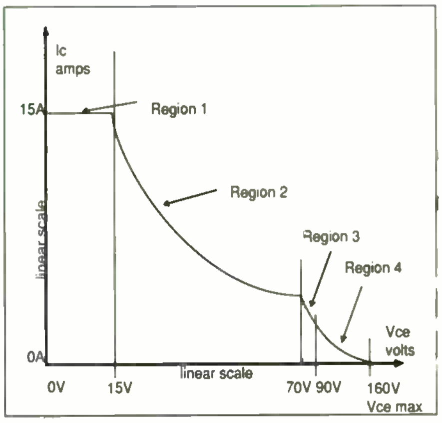
Fig. 9. The model of a loudspeaker cable short circuit ocurring at the end of a 30m length.

Plotting an SOA violation

The typical SOA graph of a linear power device comprises at least two regions. Assuming adequate heatsinking and low V_{ce}, current is limited by internal wires and bonds. The second region is bounded by junction temperature, and at the opposite endpoint, by the relevant breakdown voltage. In bipolar devices (cf mosfets), there is a second region of second breakdown, a state of positive thermal feedback. Crossing the line will destroy or damage the device unless duration is short enough. This region may have one or more breakpoint.

In the lower portion of **Fig. 11**, the SOA model for a number of high power bipolar output devices has been entered as a series of expressions which correspond to the major boundaries of the graph:

Region 1: IC > 15A, V_{ce} <15V.
Region 2: IC > 225/V_{ce}, 15V<V_{ce}<70V
Region 3: IC > [-0.1082(V_{ce}-70)]+3.2 and 70V<V_{ce}<90V.
Region 4: IC> [-0.015 (V_{ce}-160)] and 90V<V_{ce}.



The typical SOA graph breaks down into four areas, each represented as a separate expression in the MicroCap model

Expressions 'Zap1' to 'Zap4' individually model each of the four regions, using the above equations and appending descriptions of the quantities and their mode numbers, ie. V(9,1) signifies 'the voltage between nodes 9 & 1.' Relational operators are also added, so if true, they will evaluate as '1', and '0' if untrue. Beneath them, the 'Total' expression also uses relational and Boolean operators, to switch the 'expire' node (9) to logic state 1 (+1V in transient analysis) if and while the SOA is exceeded in any region. Above the text, 'Zap1-4' and 'Expire' are 'User function sources' (they can be driven by expressions), where the logical voltages are manifest.

The main circuit at the top follows previous simplified, transistorless output stage schemes. Working with paralleled bipolar transistors and the required half-wave drive (for Class AB) is possible, but the complication would slow simulation and wouldn't offer much salient extra data, as breakdown phenomena aren't covered by the classic Gummel-Poon or Ebers-Moll models. On the right of the simplified output stage is the familiar drive unit model (nominally 8Ω). On the far right, a portion of a second drive-unit can be seen. In fact, it is one of four paralleled units. The idea is to see what happens when four of the drive units

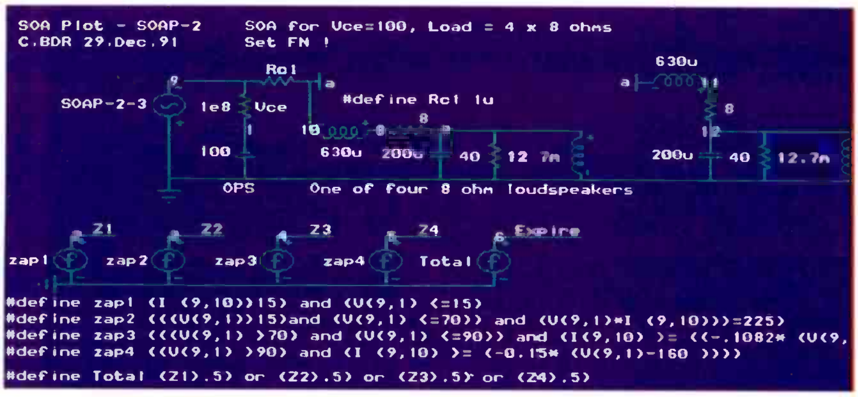


Fig. 11. A model of abuse: four 8Ω loudspeakers connected in parallel although the items marked Zap refer to modelled regions on the SOA curve.

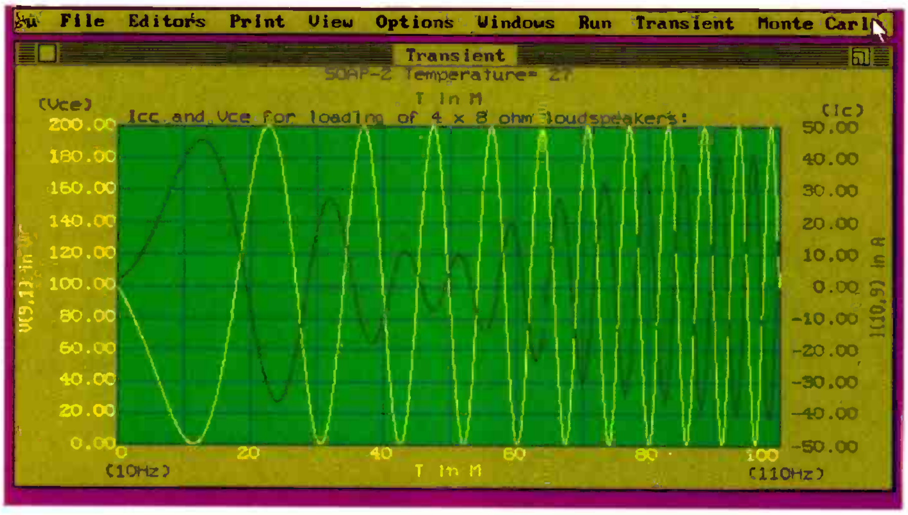


Fig. 12. A plot of collector current and collector-emitter voltage over a frequency sweep

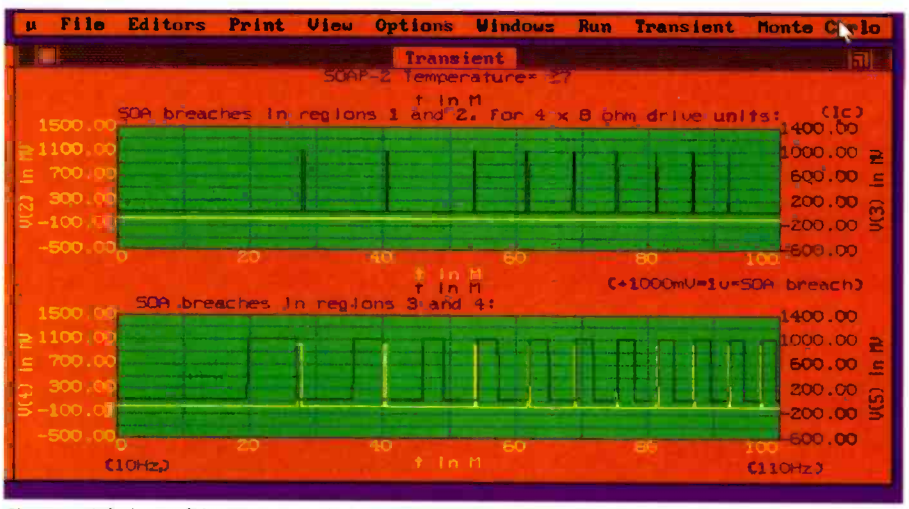


Fig. 13. Violations of the SOA in regions 3 and 4. These may not be important – see text

are paralleled across the output, a common form of abuse in PA systems. On the far left, the generator is set at 100V peak (nominally 1.2kW into 4Ω), sweeping from 10 to 110Hz. Aside from the fact that the heaviest out-of-phase demands are made by LF drivers, SOA is also leanest in this area as there is little derating over DC conditions until PRF significantly exceeds 100Hz.

Figure 12 plots I_c and V_{ce} against time, hence over the frequency sweep. Assuming the output stage is Class AB push-pull, note that V_{ce} above 100V may not necessarily coincide with the indicated current, as most of the latter should be passed by the phantom 'other half'. To see why this is not a problem, turn to **Fig. 13**, where the logical voltages on the four function sources can be seen. With these, we can see in which (if any) regions SOA is being exceeded – and for how long.

MicroCAP vs SPICE types

In general, the circuits shown can be simulated with Spice-based software. However, readers should be aware that MicroCAP has some user friendly facilities not found in most – or any – Spice type simulators. First, it can cope with stepping components to zero values. No such thing exists in the physical universe, but without it, you would have to step to a very small value (say 10^{-12}), and subtract this from an integer, producing a messy step value like '0.99999...e3'.

Few SPICE simulators have the facility of entering expressions (formulae), and it probably won't be possible to plot V against I directly.

Although the individual excursions of plot pairs 3 and 4 (lower screen) are clearly visible, being in different colours, scales are staggered to keep them clear on B+W print dumps. Most of region 4 concerns V_{ce} above 100V, and simple inspection of Fig. 12 will show which alerts are false. The need to do this can be overcome by further developing the expressions and the output stage modelling. Other violations may be eliminated by their short duration. On the upper screen, notice that node 2 – representing bond wire failure – isn't indicating in this instance. If you simply want to know whether SOA has been violated or not, you can choose to plot the sum of the four at the 'Expire' output at (node 6).

When setting up a model of this kind, readers are cautioned that meticulous care is needed in relation to the nodal polarity and character spacing in the expressions. At least you won't be inhaling beryllium dust... ■

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Acknowledgements

The author would like to thank John Szymanski at Spectrum Software, USA, for coming up with the curve modelling and Boolean expressions, needed to plot SOA behaviour elegantly.

Speed and Memory

The models presented here have minimal memory requirements. All run times on an uprated PC-XT with a V30 main processor and 8087 co-processor were under four minutes. The average is about half this. Divide these times by about 10 to 15 for a 33MHz 486 machine. Higher than necessary accuracy settings can have as much, if not more impact in the opposite direction!

Voltage-controlled amplifier systems

Malcolm Hawksford

looks at the design of audio systems using voltage controlled amplifiers.

The voltage controlled amplifier (VCA) is widely accepted in audio systems, particularly in mixing consoles where it enables remote control of both gain functions and more specialised parametric equalizers. Its advantage is that it enables signals to be processed in an essentially analogue environment but has no need for long signal routing to control surfaces.

Computer systems can be overlaid to yield sophisticated control and function automation while allowing signals to remain in the analogue domain. Such considerations have formed the principal thrust for refinement of circuit topology. Lucrative studio applications, each of which may be using 100 or more programmable gain elements, have made the VCA a front line target for competitive manufacture.

In professional audio, the specification required of the VCA is considerable: noise, bandwidth and distortion should be comparable with the best operational amplifiers yet the VCA should offer the added function of gain control.

It should also satisfy the need for accurate calibration, repeatability and temperature insensitivity.

We shall consider only electronic solutions based upon bipolar technology, as these techniques are highly refined. Motorised and precision-switched networks, although well matched on performance grounds, are often too expensive and are space-inefficient for many applications.

A basic, though non-ideal, VCA cell is shown in **Fig. 1**. The circuit consists of two transistors in a differential, long-tail pair configuration where the gain is controlled by the tail current $2I_g$. The input signal V_i is attenuated by the resistor network R_1, R_2 to produce an inter-base voltage V_{BB} that is restricted to a few millivolt. Inevitably this limits the signal to noise ratio (SNR). The output voltage V_0 is derived using a differential trans-resistance amplifier operating on the collector currents of T_1, T_2 such that

$$V_o = R(I_{c1} - I_{c2}) = 2Ri$$

In this example, i is the collector signal current that also circulates through each base emitter junction, where the difference process, in association with two essentially identical but differentially-driven transistors (T_1, T_2) is fundamental to many VCA circuits.

At the heart of the VCA system design equations is the logarithmic relationship between emitter current I_E and base emitter voltage V_{BE}.

$$V_{BE} = \frac{kT}{q}\ln\left[\frac{I_E}{I_S}\right] \qquad \ldots 1$$

where V_{BE} is base emitter voltage; I_E is emitter current; k is Boltzmann's constant, 1.38E-23J/Kelvin; T is temperature Kelvin; q is charge on an electron, 1.602E-19 coulomb, and I_S is saturation current.

Although this equation excludes several parameters found in more general purpose transistor models[1], it nevertheless embraces the target to which a transistor should adhere if it is to be a successful candidate for a voltage controlled amplifier.

In other words, the logarithmic law is the key parametric element.

In practice, although equation 1 is a simplification, it can be approximated with suitable transistor selection, appropriate biasing and by maintaining the collector-base voltage at a low but constant voltage to minimise the contribution from slope impedances[2].

The basic cell of Fig. 1 includes a number of structural details necessary for a VCA.

In its present mode of operation, it is not yet an acceptable candidate, even though it can offer a basic understanding of VCA operation. A positive attribute of the cell is the output difference circuit which implies that the output signal is a function of the difference of V_{BE1} and V_{BE2}.

This feature not only compensates for the base emitter offset voltage of a single transistor but, for perfectly matched transistors, it eliminates the dependence upon the saturation current I_S which is a highly temperature sensitive transistor parameter.

Using equation 1 and assuming the same I_S for both T_1 and T_2,

$$V_{BE1} - V_{BE2} = \frac{KT}{q}\ln\frac{I_{E1}}{I_{E2}} \qquad \ldots 2$$

The two-transistor cell remains temperature-dependent as T (in degrees Kelvin) appears in the equation; however, it is orders of magnitude less problematic than the temperature variation with I_S. The incorporation of transistors that are both parametrically and thermally matched is therefore the second key to VCA design. It necessitates the use of monolithic arrays using transistors that have accurate geometric tolerances and device proximity, fabricated on a common substrate to aid isothermal operation.

Unfortunately, when the circuit of Fig. 1 is driven from the base as a differential amplifier, such that $I_{E1} = I_g + i$ and $I_{E2} = I_g - i$, equation 2 shows a non-linear relationship between the output signal current i and the inter-base voltage $V_{BE1} - V_{BE2}$. Indeed, if re_1, re_2 are the small signal V_{BE}/I_E slope resistances of T_1, T_2, then using transistor amplifier theory[3], we can show

$V_{BE1} - V_{BE2} = i(r_{e1} + r_{e2})$

where

$$r_{e1} = \frac{kT}{q(I_g + i)}, r_{e2} = \frac{kT}{q(I_g - i)} \qquad \ldots 3$$

that is, the cell transconductance g_m only becomes linear for $i << I_g$,

$$g_m = \frac{2q}{kT} I_g$$

which is a severe restriction on dynamic range. The small signal expression for g_m reveals a gain proportional to I_g and since $I_g > 0$ the cell of Fig. 1 can form the basis of a two-quadrant multiplier. Thus by controlling the tail current, appropriate I_E/V_{BE} slope resistances for T_{B1}, T_2 are established which programs the cell transconductance.

However, for a successful VCA application, the problem of non-linearity must be addressed in order to attain both acceptable distortion performance an ' SNR. A number of topological variants will be presented, which yield the following family of gain laws:

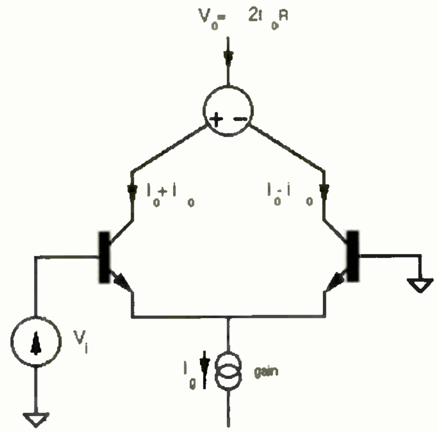
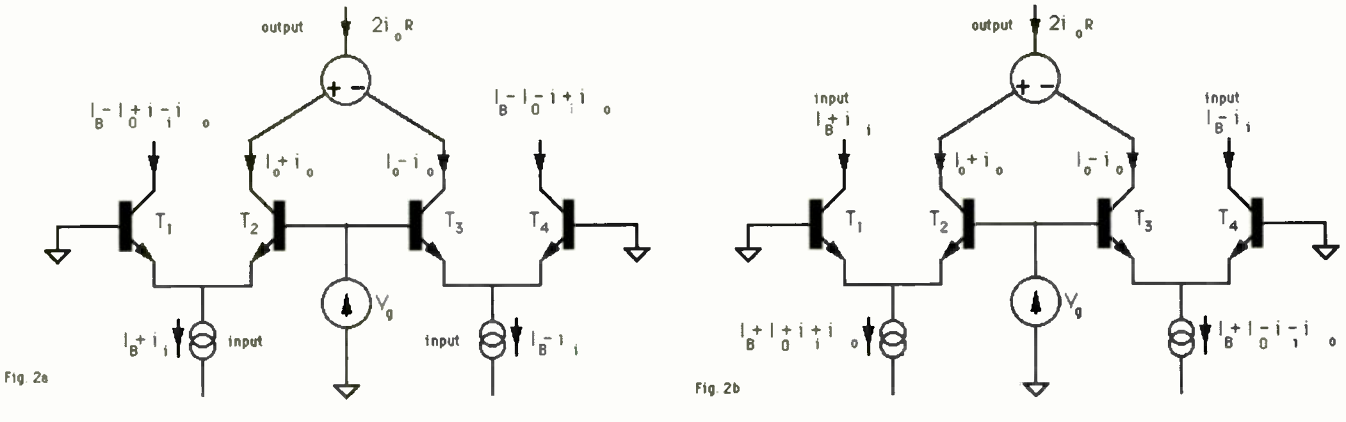


Fig. 1. Basic 'long-tailed pair' two-transistor gain circuit. The circular symbol in the collector circuit represents a differential trans-resistance amplifier in which the output voltage responds to the difference in input current.

(i) Pseudo-logarithmic gain law, but with a maximum gain of unity;
(ii) True logarithmic gain law with capability of gain > 1;
(iii) Linear gain law but with a maximum gain of unity;
(iv) True linear gain law with capability of gain > 1.

It will be shown that groups (i) and (ii) are variants on Fig. 1 but incorporate dual, 2 transistors cells where the base emitter voltages of each pair sum to a constant voltage, while categories (iii) and (iv) use 4 transistors with $V_{BE} = 0$ yielding theoretical temperature independence. We shall also show how a fully complementary topology elegantly solves the

Fig. 2. Two differentially-driven current division cells with input applied to emitters (Fig. 2a) and collectors (Fig. 2b).



bias absorption problem and simplifies the design of both input and output circuitry.

Two-transistor basis cells

Although the topology of Fig. 1 is non-linear, it can be linearised by exchanging the roles of signal and gain control. If the inter-base voltage is equated to a gain control voltage V_g such that

$V_g = V_{BE1} - V_{BE2}$,

then rearranging equation 2,

$$\frac{I_{E1}}{I_{E2}} = e^{\frac{qV_g}{kT}} = \gamma$$

where is a gain parameter incorporating an exponential law. This is a useful result especially where a decibel gain control law is required, such that if

Gain (dB) $= 20log_{10}(\gamma)$

then,

$$\text{Gain (dB)} = \left[\frac{20q}{kT\ln(10)}\right]V_g$$

The expression for γ shows that a constant inter-base voltage (and strictly constant temperature) produces a constant current division ratio between T_1 and T_2, even for large input signals, providing the transistors exhibit good I_E/V_{BE} logarithmic conformity over the working range of emitter currents.

Figure 2 illustrates the two methods of incorporating this constant current division ratio into a VCA that depends upon the entry node for the input current.

However, to enable bipolar operation (as $I_i > 0$), two system extensions are required: firstly, a constant bias current I_B is superimposed upon the input signal current to enable the signal component to go negative and secondly, an additional, identical two-transistor cell is incorporated again with a bias current I_B but now with the signal current of inverted polarity.

The output signal is then derived as a function of the gain cell output currents using a differential, transresistance amplifier, which simultaneously cancels the output bias components and also prevents gain control feedthrough under perfect symmetry.

Thus using the current annotation in Fig. 2 and the result $I_{E1}/I_{E2} = \gamma$), then, for Fig. 2a.

$$\frac{i_0}{i_i} = \frac{2\gamma}{1+\gamma}$$

and for Fig. 2b

$$\frac{i_0}{i_i} = \gamma \qquad \text{(true logarithmic law)}$$

The gain cell of Fig. 2a produces a pseudo logarithmic law that approaches a maximum gain of unity; indeed observation of the circuit shows that as V_g increases, T_2 and T_3 become common-base stages offering excellent bandwidth and distortion characteristics.

The Fig. 2b cell results in a true logarithmic law together with current gain > 1, although there is an additional complication of forcing the collector currents of T_1, T_4 to equal the input currents $I_B + i_i$. $I_B - i_i$. This can be solved using the arrangement of Fig. 2c, where a precision operational amplifier is included in a feedback system to adjust the tail current until the desired collector current is established.

Four-transistor basis cells

The use of four transistor cells where $\Sigma V_{BE} = 0$ enables the kT/q term to cancel and the cell to become theoretically temperature-independent. The gain law is also transposed from logarithmic to linear.

As an example, a new system arrangement is shown in **Fig. 3** where the gain control signal is now a current I_g. Using the result $V_{BE1}-V_{BE2}+V_{BE3}-V_{BE4} = 0$, and noting the annotation of transistor currents, it can be shown that[4]

$$\frac{i_0}{i_i} = 1 - \frac{I_g}{2I_B}$$

Although the gain is proportional to I_g there is a restriction that $0 < I_g < 2I_B$ as the maximum gain is unity, whereon transistors T_1, T_4, as with the Fig. 2a topology, operate as grounded base stages and yield excellent linearity.

An alternative four-transistor cell of low temperature sensitivity[5, 6, 7, 8, 9], again uses the principle that $\Sigma V_{BE} = 0$, but allows a current gain > 1. This cell can be considered an extension of the topology in Fig. 1 in which the additional T_3, T_4 transistors effectively predistort the input signal applied to T_1, T_2 and achieve overall linearity.

Although the process of linearization can be explained by considering the interplay of the transistor I_E/V_{BE} slope resistances, it is simpler to demonstrate by analysis. From the condition

$V_{BE1} - V_{BE2} + V_{BE3} - V_{BE4} = 0$

and using equation 1,

$$\frac{I_{E1}}{I_{E2}} = \frac{I_{E4}}{I_{E3}}$$

when appropriate currents are substituted, it follows that

$$\frac{i_0}{i_i} = \frac{I_g}{I_B}$$

Cell noise performance

Although this has demonstrated how a gain cell can be linearized to operate over a wide range of signal currents, of equal importance is the cell noise performance. Ideally this should approach that of the best operational amplifiers and not exhibit significant noise modulation artefacts with either gain control or signal level.

Noise components in the output signal arise both from the gain cell and associated signal conditioning circuitry. Examination of all the topologies presented reveals at their core a differential amplifier with no local emitter degeneration.

This means that the internal noise sources of each transistor can experience significant amplification.

Because the noise amplification factor as well as the actual noise sources depend upon the gain and state of cell biasing, it is common for the output noise to vary as a function of gain and signal level. This produces modulation noise, where the maximum output noise may not occur at maximum gain, depending upon topology (in practice, cells that offer a maximum current gain of unity also offer low noise at unity gain).

The linearization of the gain cell is crucial in obtaining an acceptable signal to noise ratio (SNR), where it is expedient to select transistors not only with respect to parametric matching but also for noise performance.

However, the noise contribution from associated circuitry should not be neglected as it

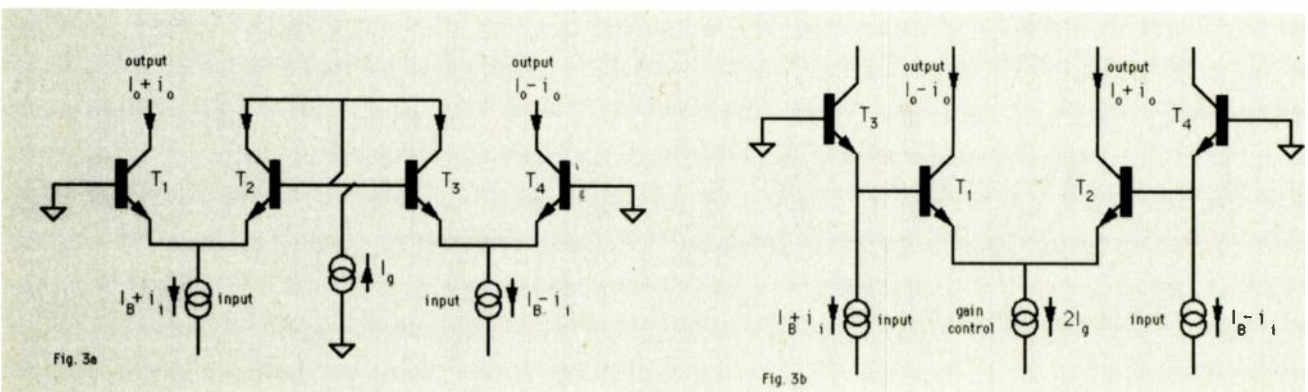


Fig. 3a. Current steering gain cell, after Hawksford[4,13].
Fig.3b. Four-transistor cell after Gilbert[5-9], with linear gain law and current gain capability >1.

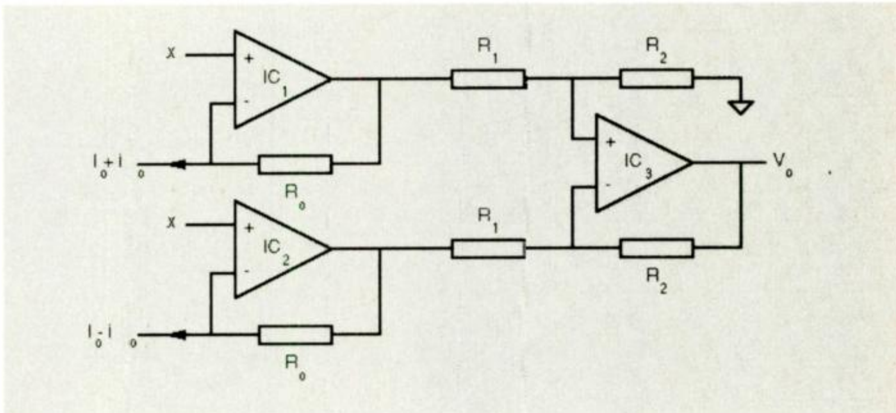


Fig.4. Differential trans-resistance output stage.

can often represent a limit on the absolute noise performance, especially at certain strategic gain settings.

There is not space in this present article to do justice to the subject of VCA noise, but an example may illustrate some of the design problems typically encountered. VCA circuitry can be sub-divided into four principal areas:

(i) input trans-conductance amplifier, often with single ended to differential transformation;

(ii) gain cell topology;

(iii) output differential trans-resistance amplifier;

(iv) gain control signal conditioning.

Consider the differential output stage shown in **Fig. 4** that incorporates three amplifiers where each introduces noise. IC_1, IC_2 are trans-resistance stages that convert the gain cell output currents to voltages while simultaneously maintaining a well-defined collector voltage. Resistor R_0 must convey a current up to $2I_0$, therefore a minimum positive supply > $2\ I_0R_0$ is required. If IC_1, IC_2 have negligible input noise currents, then the mean-square output noise voltage from this cell is $e_n^2 + 4kTR_0F$ and the signal power is $(i_0R)^2$. The best SNR is achieved therefore when $4kTR_0F > e_n^2$.

The second stage IC_3 converts the differential signal from IC_1, IC_2 to single-ended mode, where if the differential gain = 1 (ie $R_2 = R_1$) then the output noise from this amplifier is 6dB above the input noise of IC_3. In the context of a unity gain VCA, this can represent an unacceptable reduction in noise performance.

There are clear problems associated with using too many cascaded amplifier stages including that of absorbing the gain cell output bias currents. These difficulties also extend to the input stage, especially if the more typical, single-ended mode of operation is required.

All of the circuits so far considered are non-complementary and require asymmetric biasing systems to function. The cells must be suspended in a well-defined electrical environ ment both to maintain appropriate biasing and to minimise collector-base voltages and their variations.

This reduces junction heating and prevents the slope impedance transistor parameters from introducing distortion[2].

Because of this, it is common practice to make $VCB \approx 0V$ and to strive to maintain this voltage constant with signal current variations. For example, the use of collector resistors on the gain cell output transistors would be poor practice since it would allow large changes in collector-base voltages to occur.

One solution to these difficulties is to introduce a complementary topology using accurately matched NPN and PNP transistors fabricated on a common substrate. A configur ation pioneered by dbx[10, 11] is shown in **Fig. 5**.

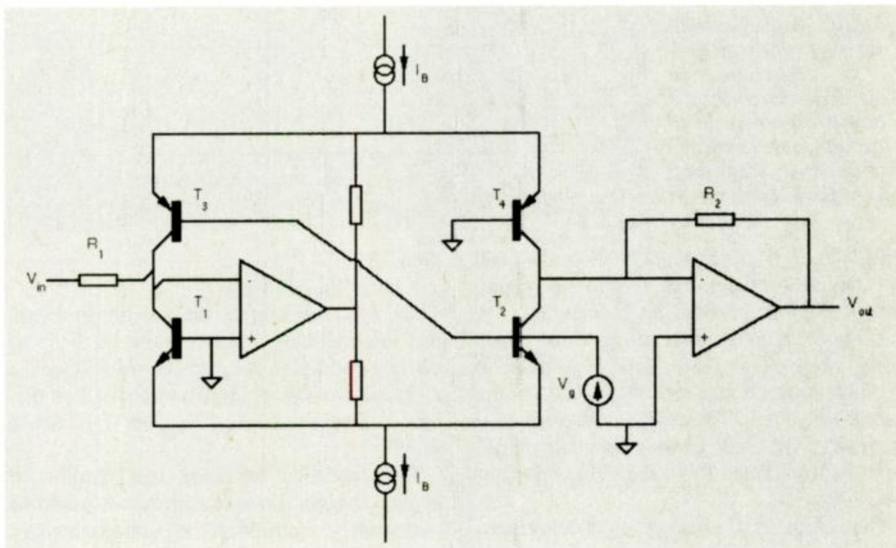


Fig.5. Basic dbx[10, 11] gain cell with support circuitry.

To understand the operation of this circuit, compare the operation of T_1, T_2 with that of T_1, T_2 in Fig. 2a which yielded a logarithmic gain law. Then, turn the T_1, T_2 circuit upside down and substitute PNP devices, thus forming a similar but complementary circuit which can now absorb the bias currents from T_1, T_2 as the biasing of T_3, T_4 and of T_1, T_2 are essentially the same.

The input transconductance amplifier IC_1, by its virtual earth operation, locates the collectors of T_1, T_3 near zero volt as does the transresistance amplifier IC_2 for the collectors of T_2, T_4. Signal current entering R_1 is then steered via feedback control symmetrically into the collectors T_1, T_3 while output current from T_2, T_4 flows via the trans-resistance amplifier IC_3 into R_2.

The control voltage V_g, which is applied to both T_1, T_2 and T_3, T_4 transistor pairs, controls the current division between collectors T_1, T_3 and T_2, T_4, which determines the system gain. The overall VCA gain then becomes,

$$\frac{V_0}{V_i} = \frac{R_2}{R_1}\gamma$$

The topology of Fig. 5 is most elegant and neatly solves many associated VCA problems while reducing the number of cascaded operational amplifiers to two.

If differential operation is required, two separate VCA systems could be used, though gain tracking is paramount. The SNR depends mainly on the gain cell and cell loading factor (that is, the ratio of signal current to bias current[12]) and amplifier noise and reaching a compromise between cell loading and levels of distortion.

Operation can be enhanced by incorporating a number of cells in parallel, although this practice is similar to increasing the junction area.

However, a main performance limitation is the bulk (ohmic) resistance associated with the cell transistors. These have greatest effect at high collector currents where the base emitter slope impedances (r_e) become small.

Bulk resistances are a primary source of non-linearity in gain cells, where the presence of both emitter and base bulk resistance degrades logarithmic conformity by adding an ohmic related voltage in series with the ideal logarithmic device. The main VCA cell equation then becomes:

$$V_{BE1} - V_{BE2} = \frac{kT}{q}\ln\left[\frac{I_{E1}}{I_{E2}}\right] + (I_{E1} - I_{E2})r_{bulk}$$

Consequently, in a current division cell, the

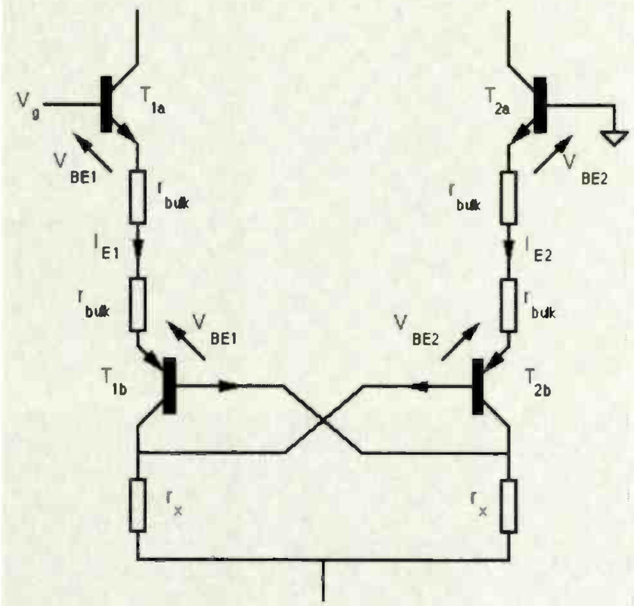


Fig.6. Modified current division cells with bulk resistance error correction circuitry. All transistors are assumed to have parametrically and thermally identical β and to have Ic/Ib current gains.

term $ln[I_{E1}/I_{E2}]$ is no longer held constant and the current division ratio becomes signal dependent, thus introducing distortion.

However, a method of distortion correction exists where a correction signal V is added to the inter-base voltage to compensate for the term $(I_{E1}-I_{E2})/r_{bulk}$. The exact nature of the correction depends upon topology but can in principle be introduced to all the gain cells considered here.

Thus by the dual strategy of selecting transistor arrays to have low bulk resistance (possibly by using parallel transistors) and by using 'ohmic-loss error correction', the inherent distortion can be maintained at low level over a wide signal and gain control range.

To illustrate one example of a distortion correction topology, consider an adaptation of the basic current division cell of Fig. 2a, as shown in **Fig. 6**.

The modified topology includes transistor bulk resistances r_{bulk} and assumes all transistors to be matched and NPN/PNP devices to be complementary. The means of cancelling the effect of bulk resistance can be demonstrated by applying Kirchhoff's voltage law, whereby

$$V_g - V_{BE1} - 2I_{E1}r_{\text{bulk}} - V_{BE1} - \left[I_{E2} - \frac{I_{E2}}{1+\beta} + \frac{I_{E1}}{1+\beta}\right]r_x + \left[I_{E1} - \frac{I_{E1}}{1+\beta} + \frac{I_{E2}}{1+\beta}\right]r_x + V_{BE2} + 2I_{E2}r_{\text{bulk}} + V_{BE2} = 0$$

Hence, if the collector resistance r_x is given by

$$r_x = \frac{2r_{\text{bulk}}}{1-\dfrac{2}{1+\beta}}$$

then

$$V_g = I\ (V_{BE1} - V_{BE2}).$$

Thus, by cancelling the ohmic voltage drops, the internal base emitter voltages become directly equated to the gain control voltage V_g, although we now observe a factor of 2 as the base emitter junctions of $T_{1a,b}$ and $T_{2a,b}$ are in series.

The modified topology can readily be applied to other forms of gain cell. It would be particularly appropriate in complementary guise for inclusions in the dbx cell of Fig. 5. Alternatively, a weighted sum of input and output signals can be superimposed upon the gain control signal[13], allowing the ohmic related voltage drops to be cancelled.

Conclusion

Our discussion has considered a number of factors encountered in the design of a VCA, where the importance of linearity and low noise have been emphasised. We have shown that complementary topologies and error correction for the effects of bulk resistances are also effective tools in furthering the art.

In practice, most bipolar VCAs are variants on the type of topologies presented here where, for example, so called 'class AB biasing' is used with the dbx complementary type of circuit. Such techniques increase distortion but can reduce noise, especially at low signal level.

The maintenance of low distortion at high attenuation is also important where performance can degrade due to transistor current starvation and common mode signal components within a differential input stage[13]. Indeed, a two-stage current steering topology has been proposed where an experimental circuit topology is presented[13]. Because of its non-complementary structure it is also a good illustration of the problems encountered with power supply design and the maintenance of low noise performance.

Most of the problems in VCA design are now well understood and modern fabrication can yield low cost devices of acceptable performance. However, the balance of technology is rapidly changing with the advent of sophisticated and affordable digital processing. Apart from some specialised equipment, the day of the analogue VCA may soon pass into the history books. Even the most sophisticated devices still have, and will probably continue to have, measurable deficiencies, as exposed in Ben Duncan's acclaimed series in Studio Sound[14]. ■

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DESIGN BRIEF

Low distortion audio frequency oscillators

There are ways and means of ensuring that audio reproduction is as accurate as possible. *Ian H[illegible]an* ***looks at some of them.***

Readers of *EW + WW* have long shown a lively interest in high fidelity reproduction, dating back to before the days of the Williamson amplifier. Since an early quasi-complementary design[1] appeared in 1961, many solid state high fidelity power amplifier designs have featured in these pages.

The evaluation of amplifier performance requires a low distortion AF oscillator – or rather two, since intermodulation testing is essential these days. *EW + WW* has published many designs for these, including an early one by Rider. Particularly noteworthy is an APF- (all-pass filter) based oscillator using a distortion cancelling technique[2]. This uses a Philips thermistor type 2322 634 32683, and the circuit achieves a very low THD, namely <0.005% at 20Hz and around 0.0002% at 1kHz.

What sort of performance can be obtained without using a thermistor? The most obvious and cheapest alternative amplitude stabilisation method is to use a diode limiter. This avoids using a specialised and expensive component; moreover it is aperiodic and so completely removes the annoying amplitude bounce often found in instruments using thermistor or AGC-loop stabilisation, when changing frequency.

With diodes, a design based on the SVF (state variable filter) is preferable to the all-pass filter approach, since the former offers an inherent 12dB octave roll-off at the lowpass output. The approach is a great help as the clipping will produce all the odd harmonics one expects to find in a squarewave.

In contrast, the small amount of distortion produced by a thermistor is almost pure third harmonic. This makes distortion cancellation by outphasing in the Rosens[2] design very effective even in an APF-based design. Incidentally, in an SVF-based oscillator, the desirable feature of a linear scale is easily engineered.

Figure 1a shows an SVF-based oscillator which, with the component values given, operates at 1.59kHz. V_l (the voltage at the lowpass output, OUTPUT 1) was 5.3V pk-pk. The Q of this two pole filter is $R5/3R4 = 11$ with the values shown. This modest value of Q was chosen deliberately, to enable the effect of circuit changes on the output distortion to be easily seen. (A Q of 30 would be quite usable and is indeed used in a 20Hz-20kHz 0.04% distortion SVF-based sinewave generator currently in production[3].)

The circuit operates as follows. If an input were applied via a resistor to the inverting input of IC_{1a}, the bandpass output *BP* would be in phase with it at the frequency where the gain of each of the integrators is unity. So if the bandpass output is taken and clipped (an aperiodic operation introducing no phase shift) to a squarewave, the latter can be used as a fixed level excitation input to the filter.

At the filter *LP* (lowpass) output, the fundamental appears amplified by the Q factor while the harmonics are reduced due to the filter's 12dB/octave roll-off. **Figure 2a** shows the waveform across the diodes V_d, which is

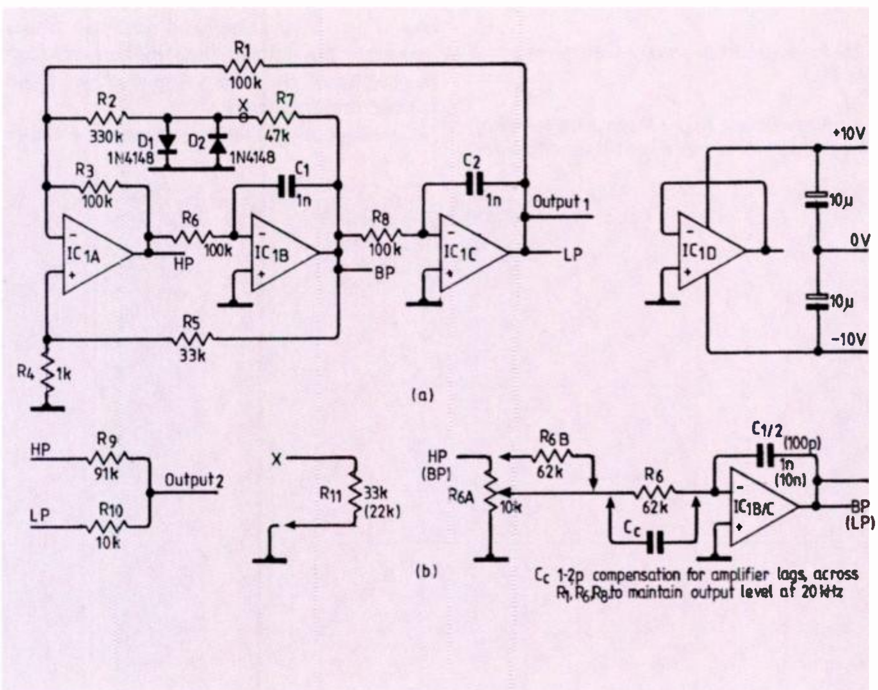

Fig. 1a. 1.59kHz oscillator circuit based on a state variable filter. (A detailed explanation of the SV filter's operation is given in "Analog Electronics", Heinmann, ISBN 0434 907 23 5).

Fig. 1b. Modifications give a nill net third harmonic at output 2.

0.96Vpk-pk and a rough approximation to a squarewave. Assuming that the fundamental compon ent is about 1.4Vpk-pk, the output should be $1.4Q(R_3/R_2) = 4.7$Vpk-pk, not so very different from that observed. If the square-wave input to the filter were ideal, the amplitude of the third harmonic component would be one third that of the fundamental.

At the filter's output, the third harmonic will have been attenuated by a factor of three in each of the integrators, while the fundamental will be amplified by the Q factor of 11. The third harmonic should therefore be one third of one ninth of one eleventh of the fundamental, or 0.34%.

The fifth and higher harmonics will be substantially lower, due to the filter's 12dB/octave roll-off, so the expected distortion is approximately 0.34% and all third harmonic. In fact the measured distortion is 0.2%, due to the smoothish nature of the squarewave of Figure 2a.

With the SVF, the signal at the *LP* output is always in anti-phase to the signal at the *HP* (highpass) output - another advantage over the all-pass design in this application, as it simplifies the outphasing of the distortion. This is achieved by making the circuit operate as a second order elliptic filter at the same time as an oscillator, as follows.

The circuit oscillates at the frequency of the peak of the *BP* (bandpass) response and at this frequency the gain of the two integrators is unity, so the fundamental component of the output has the same amplitude at the *HP*, *BP* and *LP* outputs. On the other hand, the third harmonic is attenuated by a factor of three in each of the integrators.

So by combining one ninth of the *HP* signal with the *LP* signal to give V_2 at OUTPUT 2 as in **Fig. 1b**, the net third harmonic at OUTPUT 2 is nil: we have placed a zero in the filter's response at three times the frequency of the BP peak response. Meanwhile, V_2 has been reduced by about 1dB, by the partial outphasing of the wanted fundamental output. The absence of third harmonic is clear in **Fig. 3a**, showing the residual distortion component in OUTPUT 2. A count of the peaks of the waveform shows that fifth harmonic predominates, but there is a rapid spiky reversal, corresponding to the steep sides of the Figure 2a waveform.

The higher harmonic components of Figure 2a are far from negligible. Nevertheless, the distortion is reduced from 0.2% to 0.095%, a useful if not spectacular improvement. Clearly matters would be improved if the clipping were gentler, the problem being that the *BP* output amplitude into the limiter is ±2.6V peak, whereas the diodes clip at only 0.5V.

Connecting R_{11}, 33K in parallel with the diodes moves the point at which clipping occurs up nearer the peak of the waveform, resulting in more gently sloping sides, **Fig. 2b**. The level of the fundamental component is little affected, since the output falls by only 0.5dB.

The residual is then as in **Fig. 3b**, the distortion is 0.034% and is visibly almost pure 5th harmonic. If R_{11} is further reduced to 22K, with a further 0.5 dB drop in output, the distortion falls to 0.018%.

Thus with R_5 raised to provide an operating Q of 30, a distortion level of 0.006% could be expected, a very creditable performance for such a simple circuit. Indeed, one would need to consider using a lower distortion op-amp such as the *NE5532*[2], the *TL084* used having a typical total harmonic distortion of 0.003% up to 10kHz.

Performance is still substantially inferior to Rosens[2] and the reason is not hard to find. With a second order filter, we can only engineer a zero in the stopband response at one frequency. So although the third harmonic can be out-phased, the filter's response rises again beyond that frequency.

Consequently, the arrangement actually makes the fifth harmonic level in the output worse. This is where the thermistor scores, any small variation in its resistance over a cycle at the operating frequency resulting in almost no harmonic distortion other than third.

It is tempting to speculate that by further waveform shaping in the non-linear network, one could restrict the harmonic distortion in the drive signal applied via R_2 to the filter's input to fifth and higher harmonics. The outphasing could then be modified (R_9 becoming 250K) to suppress the fifth harmonic.

It's true that this would worsen the seventh harmonic in the output, but not by nearly as much, since the ratio of five squared to seven squared is much less than the ratio of three squared to five squared. However, although this is doubtless possible in a fixed frequency oscillator, the necessary settings would almost certainly be too critical to hold in a tunable oscillator covering 20Hz to 20kHz.

Incidentally, the linear scale is organised as shown in Figure 1b. With $R_6 = 62$K, the frequency range is 2kHz down to zero. At mid-track (1kHz) the loading of R_6 on R_{6a} causes the output frequency to be a little too low, since R_6 now sees the pot as a 2K5 source impedance instead of zero at max and min.

This parabolic error can be changed to a much smaller cubic one by connecting R_{6b} $(=R_6)$ from the wiper to the top of track, giving zero error at 1kHz. ■

References

1. *Transistor Audio Power Amplifier*, R Tobey and J Dinsdale, *Wireless World*, November 1961, p565-570
2. *Phase-shifting oscillator*, R Rosens, *Wireless World*, February 1982, p38-41
3. Linstead G3 Sine, Triangle, Square Audio Signal Generator, manufactured by Masterswitch Limited.

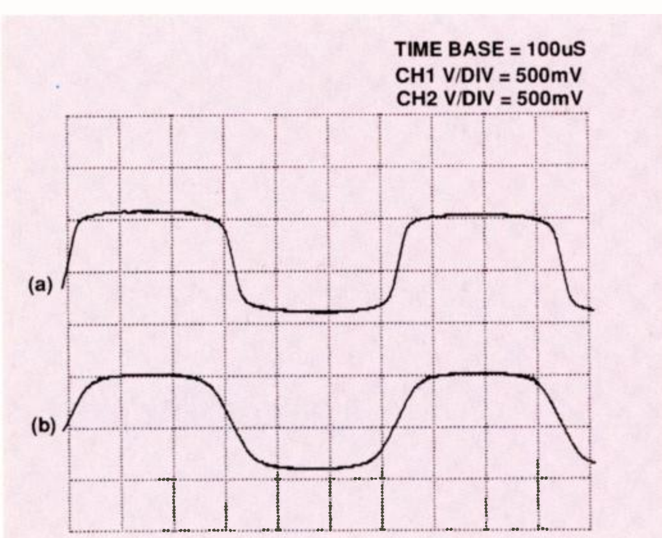

Fig. 2a. Smoothness of square waveform across V_d.

Fig. 2b. Connecting R_{11}, 33K, in parallel with the diodes results in more gently sloping sides

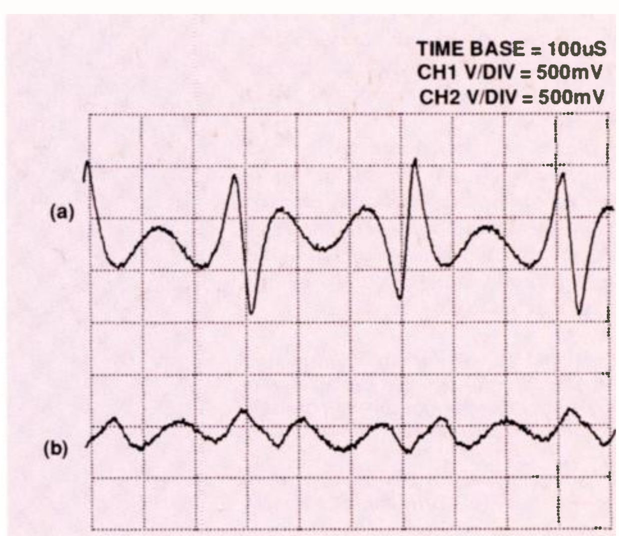

Fig. 3a. Residual distortion component in ouput 2. Fifth harmonic predominates.

Fig. 3b. Residual after connecting R_{11}.

Plans and problems for Channel 5

Against a background of the disastrous fall-out from the 1990 Broadcasting Act, the competing new technologies of cable and satellite, and the contraction in advertising, the Independent Television Commission needs to award the franchise for the new terrestrial Channel 5 network. Transmissions will be squeezed almost entirely into UHF Channels 35 (582-590MHz) and 37 (598-606MHz), two of the four channels comprising the "gap" between Bands IV and V. Channels are being cleared of ancillary services for broadcasters and radio microphones (ch 35), CAA and MOD radars, (ch 36) and radioastronomy (ch 38).

The transmitter engineering plan for Channel 5 has been completed by ITC and NTL engineers in collaboration with the Home Office and DTI and represents a unique exercise in showing just how much coverage can be achieved with such limited spectrum. In its final form, provision is made for 33 transmitters covering an estimated 73% of the UK population. All but six transmitters will use the two channels, exploiting for the first time in the UK, high stability, precision-offset sources, with a tolerance of only ±1Hz in carrier frequency. Horizontal and vertical polarisation will be used for both high- and low power stations.

While the existing main UHF transmitters are stable to ±250Hz, the use of precision offsets derived from either a rubidium source or an atomic clock reference (available from some navigational satellites) will permit higher co-channel signal levels without perceptible interference.

Because of a Home Office ruling that the BBC may not compete to provide transmission services under its present Royal Charter, four of the C5 transmitters (Croydon, Burnhope, Black Mountain and Mounteagle) will be sited at former IBA VHF sites rather than co-sited at existing UHF BBC sites (Crystal Palace, Pontop Pike, Divis and Rosemarkie). A fifth alternative site (Lichfield rather than Sutton Coldfield) had to be rejected for technical reasons since Sutton Coldfield will utilise an image (N+9) channel normally considered a "taboo" channel without cross-polarisation protection unless signals at the receiver are roughly equal.

The 27% or so of people unlikely ever to receive satisfactorily any of the 33 transmitters (the 99.4% coverage of the existing four channels requires some 4000 transmitters at roughly 1000 sites) include the relatively prosperous areas in the south, south-west and south-east coastal areas of England where it has proven impossible to achieve co-ordination with transmitters in France, Belgium and Holland. There will also be many parts of central Wales, the Welsh valleys and the scattered rural communities in north-west and central Scotland. Similarly there are the viewers currently receiving signals from the hundreds of low-power gap-filling UHF relays. Some C5 transmitters, including Croydon and Sandy Heath, will have significantly reduced coverage compared with the existing Crystal Palace and Sandy Heath transmitters and there will have to be careful shaping of the vertical radiation pattern at some of the non-co-sited transmitters to minimise signal strength differences in shadow areas.

Yet another aerial

It has been estimated that some 25% of viewers in C5 service areas should be able to get reasonable reception on existing aerials but a large majority will have to install new aerials to take account of the different polarisation and/or different aerial groups. Use of a simple combiner with a second aerial will destroy polarisation protection and could result in cases of co-channel interference even on the existing channels.

A major uncertainty for would-be C5 programme companies (franchises to be awarded on the basis of a quality-threshold plus highest bid) will be added start-up cost of overcoming interference to video recorders, particularly those which have for some reason been detuned from ch 36 (to which most VCRs are factory-tuned).

Since VCRs use double-sideband modulators this increases the possibility of interference. Work by NTL on behalf of ITC suggest that a total of about 2.5-million VCRs may require retuning or the fitting of filters to overcome interference problems. There is also the reverse problem that VCR, home-computer modulators may interfere with C5 reception. But ch 35 and 37 are already in use in Continental Europe (where the broadcasters are not liable to rectify interference) without major problems arising.

Provided the award of the franchise is made this autumn, he expects C5 transmissions to begin by the end of 1994 or possibly even by the end of 1993.

Pat Hawker

Rebirth of wireless?

Two subjects which arose at the 1991 International Solid State Circuits Conference and were repeated at the 1992 conference (see Update Special) were the rise of wireless connections for telephones and computers and the projected replacement of CRTs by active matrix thin film transistor liquid crystal displays.

The first subject concerned, as George Heilmeier, President and CEO of Bellcore Communications Research put it, "The major paradigm shift under way in telecommunications from wireline communications with phone numbers associated with a specific location to tetherless communications with personal number calling" or, to put it more simply, cordless phone. Heilmeyer concluded that the ideal spec is a power in the region of 1-10mW, less than 0.17kg and coverage areas less than 0.4km for each radio site.

He thought it would not be until 2005 that such a service was fully in place and that the quality of the service and the speed of its introduction would be largely determined by the way in which the spectrum is allocated.

Nonetheless Heilmeyer saw "widespread personal communications services becoming available in the 1990s; electronic communications nearly as effective as face-to-face communications; and personal nomadic computing along with the merging of computers, cellular phones, pagers and electronic mail."

The latter issue re-ran the debate over whether amorphous silicon or polysilicon would be the technology for these displays, but a more immediate point was raised – the manufacturing cost of such panels. Toshihisa Tsukada, senior chief scientist at Hitachi's Central Research Laboratories said the problem with getting these displays out into the commercial market was not, as was commonly supposed, yield – these were over 50 per cent, said Tsukada – but cost. The cost of manufacturing was, he said, around $1500 a panel. That is not yet competitive with CRTs.

David Manners, *Electronics Weekly*

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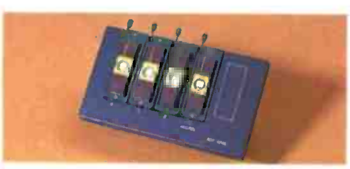

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