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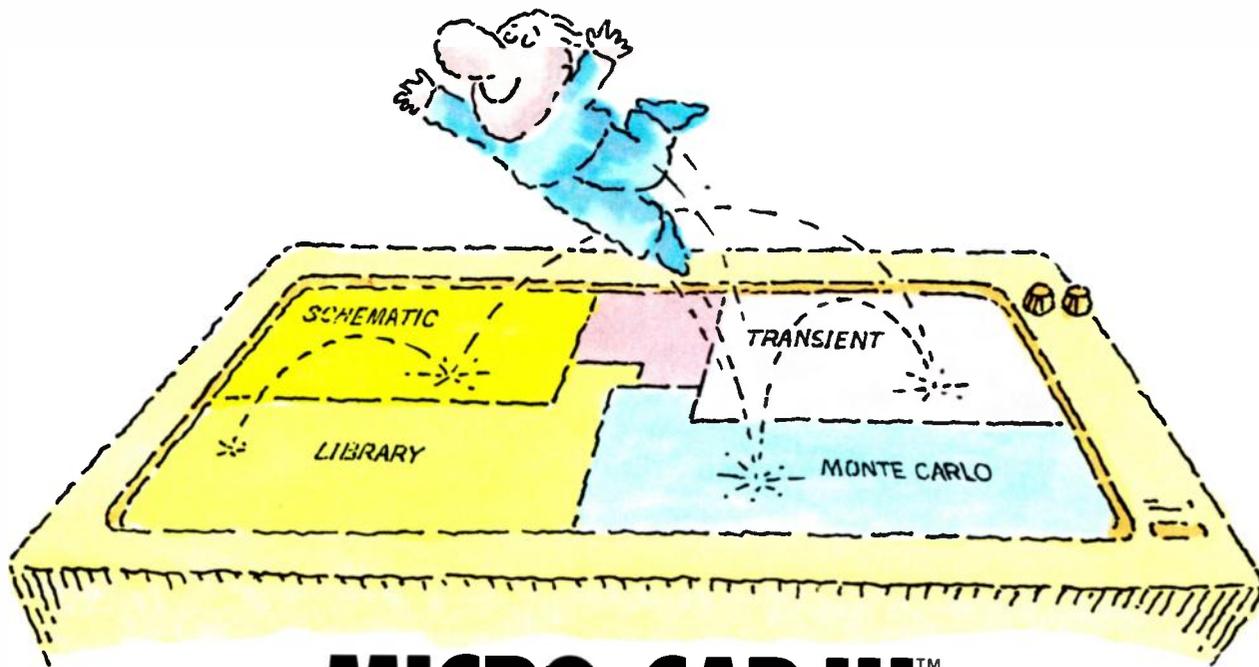
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simulation**

CIRCUIT DESIGN

PRECISION AMPLIFIERS





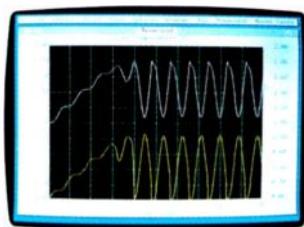
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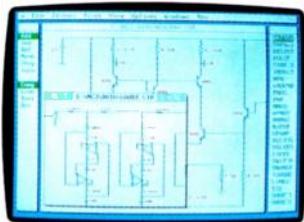
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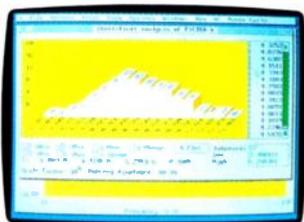
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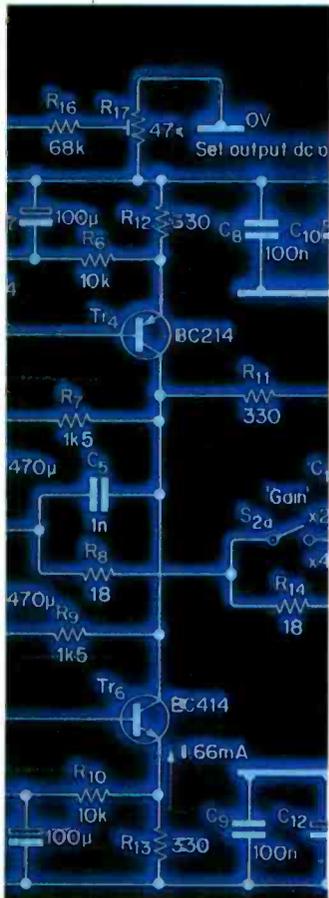
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80286	12	16.5	4	-	2/4/0	109	169	-	-	739	769	869	899	-	-
80286	12	16.5	8	-	3/4/0	139	215	269	399	775	795	905	935	-	-
80286	16	21.5	8	-	3/4/0	189	265	319	449	-	875	969	999	1319	1449
80286	20	27.0	8	-	3/4/0	239	315	369	499	-	925	1019	1049	1369	1499
80386SX	16	21.0	8	-	1/6/0	299	395	489	619	-	999	1095	1125	1445	1569
80386	20	27.5	16	-	2/6/0	509	599	699	829	-	1279	1305	1339	1649	1775
80386	25	34.0	16	-	2/6/0	629	719	809	939	-	1389	1419	1459	1769	1895
80386	25	43.0	16	32	2/6/0	849	939	1029	1199	-	-	1769	1799	2099	2219 *
80386	33	58.0	16	32	2/6/0	1189	1269	1359	1509	-	-	2109	2139	2439	2559 *
80386	33	59.0	32	128	1/4/2	1509	1595	1695	1829	-	-	2499	2529	2829	2949 *
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There is some good news and some bad news. First, the bad news. The worldwide electronics industry is heading for another of its cyclical recessions.

And the good news? The UK electronics industry has atrophied to the point where recessions don't matter any more.

A deeply cynical and pessimistic view, perhaps. But how can you be optimistic when you see small, highly innovative companies fighting for their life and their larger counterparts fighting each other to sell off their electronic interests?

The Liverpool company Rytrak started life in December 1988, the brainchild of two ex-GEC Hirst Research engineers who knew a better way to deposit high quality polysilicon on large glass substrates. The revolutionary low pressure chemical vapour deposition process brought with it the promise of the world's most advanced LCD flat-screen TV and computers, an initial £450 000 venture capital funding from 3i, and the personal blessing of the Duke of Gloucester.

Rytrak built and delivered four of its £350 000 CVD machines and has orders for at least two more. Its products and processes have attracted interest from Sharp, Toshiba and Hitachi. It attributes its present problems to cash flow.

All companies must exist on the basis of sound financial planning but it seems wrong that a spark of high-tech enterprise should find itself extinguished so readily. Rytrak may have got its sums wrong or a likely customer may have suffered from indecision. Either way, a product and process of such commercial potential would not be allowed to founder in Japan.

As regards the destructive practices of larger companies, we can do no better than to quote our friends at *Electronics Weekly* writing in their *Ruminator* column:

"Funny old world, isn't it when last year's tale that the only way to compete in electronics was to forge transnational company link-ups to share R&D and manufacturing costs in semiconductors and telecoms, is succeeded by this year's tale that R&D is unnecessary in the electronics industry and that making telephone exchanges is an unprofitable business to be in.

"We appreciate that this is the softening up process to ease the moment when the UK's share in our only telephone exchange manufacturer, GPT, can be sold off to Siemens, and when the chip process R&D at Caswell can be abandoned which will inevitably lead to the demise of our only significant chip maker, Plessey.

"Curious too when we also see the top brass at STC apparently indecently anxious to shuffle off its interest in our one remaining large computer maker, ICL. And then we remember the apparent relief with which Thorn flogged off our last significant consumer electronics business, Ferguson, to the French. It seems that the bosses want the UK out of electronics in high end communications, components, consumer and computer.

"Yet the 4Cs are the fastest growing areas of an industry which is expected to be the world's largest within a decade. Funny old world, isn't it?"

We could have added a few more abdications of our own, and have done so frequently in this column: the UK semiconductor industry in its virtual entirety with the magnificent exception of Zetex, and a complete infrastructure in the UK's passive component business.

We don't have an easy answer. The hard one goes right back to the beginning of the educational system and students' perceptions of working in industry. However, those who really matter, the British boardrooms, don't appear to have any answer at all.

Or even to have asked themselves the question.

Frank Ogden

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Radio waves at the edge of Eternity

There may or may not be a restaurant at the end of the Universe, but we may soon know what's happening at the outer edge of our own solar system. That at least is the prospect offered by some new calculations from the Polish Academy of Sciences.

In a recently published paper (*Nature*, Vol. 343 No 6267) Andrzej Czechowski and Stanislaw Grzedzielski investigate some mysterious 3kHz radio emissions that were picked up in 1983 by the Voyager 1 and Voyager 2 spacecraft. At that time both craft were between the orbits of Saturn and Uranus.

The precise origin of these emissions remains a mystery; they could have been the result of some interaction between the solar wind and the planets or they could have originated from outside the solar system, possibly in a nearby star. For the purposes of the latest research the origin of the radio waves is much less significant, however, than their behaviour.

What Czechowski and Grzedzielski have done is to examine the characteristics of these 3kHz radio emissions and use them to calculate the distance to the edge of the solar system, or more precisely the position of the heliopause. That is where the wind of charged particles from the Sun meets the flux of particles that form the interstellar medium.

According to the Poles' calculations, the position of the heliopause is between 60 and 100AU (astronomical units) from the Sun — that's 60-100 times the distance between the Sun and the Earth. Pluto's orbit is about 50AU, so unless there is a Planet X at a greater distance, the Sun's electromagnetic influence ends just beyond the planetary system. That in turn is significant, because Voyager 1 is already 40AU out on its journey and could therefore reach the edge of the solar system within the next few years.

Precisely how the Polish scientists reached their conclusion is almost as interesting as the conclusion itself. Whatever the origin of the faint 3kHz emissions, they were observed to drift upwards in frequency at a rate of 1kHz per year until they faded away. This rise of frequency is attributed to Fermi acceleration, an effect that takes place when radio waves bounce off a moving structure. That moving structure, according

to Czechowski and Grzedzielski, is the solar wind moving outwards from the Sun at 400km/s. The other reflecting surface that forms a cavity within which the waves are trapped must inevitably be the fixed heliopause at the edge of the solar system.

The Poles go on to argue that the rate of increase of frequency of the RF emissions in such circumstances must depend entirely on the number of times they bounce off the solar wind. That in turn must depend on the size of the cavity of which the solar wind forms part. Ultimately it provides a measure of the one unknown factor, the distance of the heliopause.

Before Voyager 1 reaches the heliopause, probably in 1996, it will first pass the so-called termination shock, a point at which the particles of solar wind cease to travel at supersonic speeds. This should occur in about 1993 and should be detectable by Voyager as a hiccup in the density and temperature of the wind.

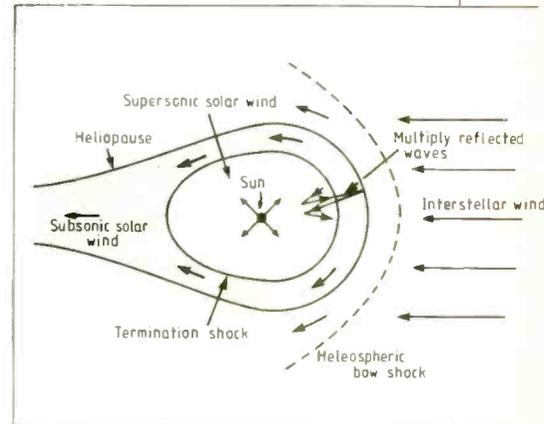
When eventually Voyager 1 reaches interstellar space it will be able to make the first ever observations of the world beyond the Sun's influence. We'll then know what magnetic fields and what chemical elements exist in deep space.

Microwaves in a skid

Albert Wuori of Michigan Technical University in Houghton is researching a novel but highly practical use for microwave energy. It's part of a \$6.5 million US National Research Council programme to find a satisfactory way of de-icing roads.

Over in the States where they've had a particularly rough winter, just about everything has been tried from high-pressure anti-freeze jets to sandblasting devices. The trouble is that anything capable of melting the ice completely uses up prodigious amounts of energy to overcome the latent heat of fusion. Traditional methods such as road salting cause corrosion to vehicles and bridges, while snowploughs damage the roads and usually only scrape off a surface layer of ice.

The problem (as anyone here who can remember far enough back knows) is that ice clings tenaciously to roads, both by chemical bonding and by physical



All being well, these observations will carry on till around the year 2015 when Voyager's nuclear power sources (see *EW+WW*, December 1989) cease to function. By then the craft will be 130AU from the Sun, well outside the solar system.

Astute space watchers may well wonder why Pioneer 10, launched earlier and now 48AU away, won't reach the heliopause first. The answer, apparent in the diagram, is that Pioneer is travelling downwind in what is assumed to be the tailward direction of the heliopause. That way it probably has much further to go.

interlocking with the cracks and crevices of the surface. What Wuori and his colleagues have therefore been trying to do is use beams of microwave energy to zap the ice-road interface.

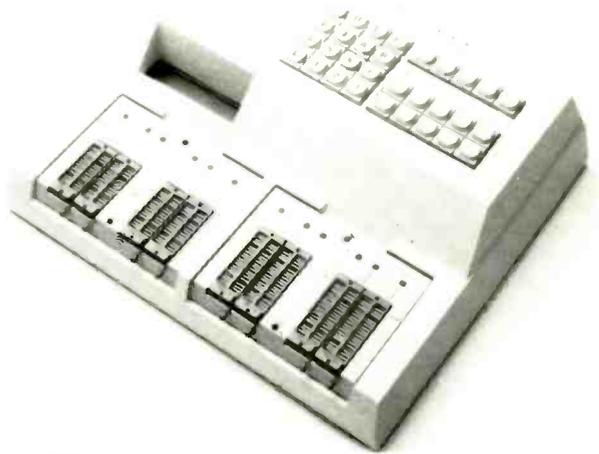
A mobile experimental system designed and built by Research Associates of Syracuse, N.Y. uses about 15kW of beamed energy and can successfully loosen ice as it passes along a laboratory track. Ultimately, the idea is to develop a practical vehicle that will travel at about 20mph, loosening ice and then dispersing it mechanically.

One problem, of course, with such a high-power system would be ensuring that stray microwave energy was kept within safe limits for public exposure, no mean task! The best solution, suggest the researchers, would be to incorporate microwave-absorbing materials in the surface layers of roads likely to be affected by ice. That way much less RF would be needed to do the trick.

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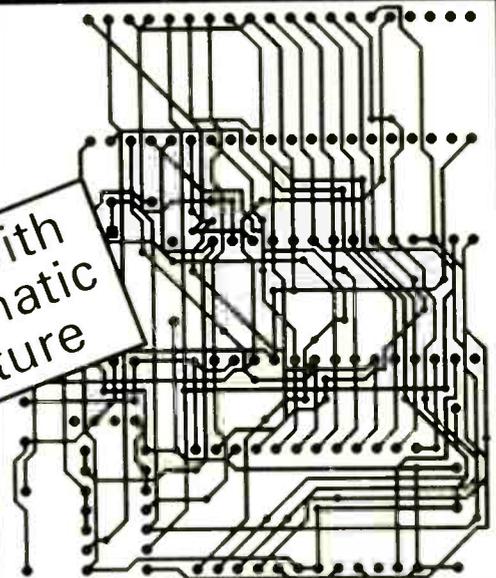


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CIRCLE ENQUIRY NO. 123 ON BACK PAGE

Atomic snooker

Scientists at IBM's Almaden Research Centre have demonstrated their ability to roll individual atoms on a surface and to build sub-microscopic structures one atom at a time.

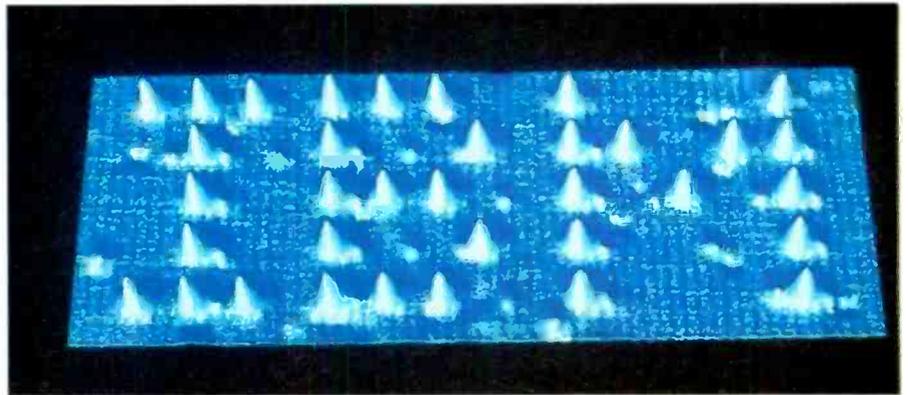
This should make possible a wealth of future applications such as building molecules atom-by-atom, altering individual molecules, making ultra-small electrical circuits far smaller than is possible today or storing data on an atomic scale at densities more than a million times greater than is now possible.

More immediately, the new technique will assist in studying the fundamental behaviour of atoms on surfaces. This knowledge is crucial to everyday industrial processes, such as catalysing chemical reactions and manufacturing semi-conductors.

Physicists Donald Eigler and Erhard Schweizer have described in *Nature* vol. 344 no 6266 how to move individual atoms across the surface of a metal crystal and have demonstrated their new technique by moving atoms of the element xenon into a pattern forming the letters "IBM". Each letter is about 500,000 times smaller than those on this page, while the atoms themselves are separated by only 13Ångstroms.

Later, the scientists created the first-ever atomic cluster built one atom at a time — a chain of seven xenon atoms bound together. These demonstrations required the atoms to be cooled to -269°C to eliminate thermal motions. Eigler says he realised the xenon atoms in the seven-atom chain were bonded together because he could relocate up to three of them at a time by moving an end atom. The ability to position individual atoms in this way has the potential to advance fabrication techniques to a scale of less than 10^{-9}m , more than 500 times smaller than that currently used in semi-conductor chip manufacturing.

The scientists used a scanning tunnelling microscope (STM) both to move the atoms into place and to view their progress. The STM, invented in the early 1980s by Nobel Prize-winning scientists at IBM's Zurich Research Laboratory, can image individual atoms on a metal or semiconductor surface by scanning the tip of a needle over the surface at a height of only a few atomic diameters. By plotting the path of the

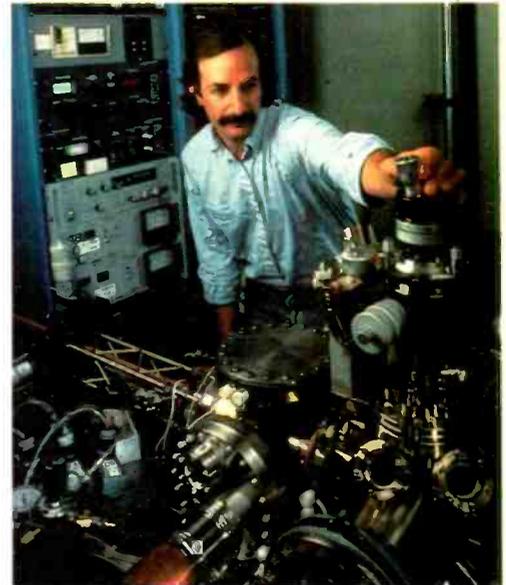


Don Eigler of IBM with the STM, with which he is able to move atoms around on a substrate at will, as witness the company name in xenon atoms.

STM tip, it's possible to make an atomic-resolution image of the surface. Eigler and Schweizer discovered that, by bringing the STM tip slightly closer to a xenon atom on the surface, they could drag the atom across the surface when they moved the STM tip parallel to the surface. At the desired position, the scientists raise the STM tip, leaving the atom fixed in its new location. The process is like moving a magnet by the motion of another magnet nearby, but not touching.

The shape of the surface is important. To another atom, even an atomically smooth surface would look like the bumpy, waffled surface of a large egg carton. To move over the surface, the atom must traverse its depressions and crests. Eigler and Schweizer's method requires that the STM operator maintain a delicate balance of several attractive forces between the atoms. For example, the atom's attraction to the tip must exceed the lateral force that prevents an atom from riding over a crest into the next depression. But to keep an atom from flying off the surface, the force that binds it to the surface must be stronger than the atom's attraction to the STM tip. Although in this technique it is possible to tune the attractive forces by controlling the height of the STM tip above an atom, the process will only work with particular combinations of atoms and surfaces.

In creating these structures, the scientists had to plot the course of the atoms across the surface carefully so the



atom being moved would not collide with, or get close enough to be attracted to, any other xenon atoms along the way.

This achievement grew out of Eigler's desire to understand how atoms and molecules interact with surfaces. To do this, Eigler built an STM that could operate with extraordinary sensitivity and precision. He not only reduced contamination and unwanted atomic motion by placing the STM in a high vacuum and cooling it with liquid helium, he also carefully isolated the STM from vibrations as faint as those from the sound of a person's voice and heat sources as weak as a nearby human body. As a result, Eigler's STM can resolve vertical changes in an atom's apparent shape as small as 0.002Ångstrom — far smaller than any single atom.

Micro-chip fuel cell with a big future

Fuel cells have always been attractive in theory because of their high conversion efficiency (50+ % compared with 35% or so for a heat engine) and also because they can provide endless power if the fuel supply is maintained. The US space shuttles, for example, derive their electricity supplies from fuel cells fed with hydrogen and oxygen gases. The only snag is that conventional fuel cells are technologically complex and hence expensive.

Against that background comes what appears to be a highly significant development from Bell Communications Research in New Jersey. In a letter to *Nature* (vol. 343 no 6258), C. K. Dyer describes a novel fuel cell which can be made using microchip technology, which is potentially cheap to manufacture and which produces a high output using mixtures of hydrogen and oxygen in widely varying ratios. Hitherto, fuel cells have required the input gases to be separated and purified if they are to produce reasonably high output voltages.

Dyer's cell is deceptively simple, consisting of little more than two platinum electrodes separated by a microscopically thin layer of pseudoboehmite, a natural form of hydrated aluminium oxide.

To make his cells, Dyer took a quartz substrate and sputtered on a layer of platinum. About 50nm of aluminium was then RF sputtered on top of the platinum. Subsequent boiling in water

oxidised the entire aluminium layer to form the pseudoboehmite. Finally a thin, gas-permeable layer of platinum was added to form the top electrode.

When fed with mixtures of hydrogen and air in a very wide range of proportions, the cell gave a virtually constant output of around 1V. Useful currents could also be drawn.

As yet, Dyer has no idea of exactly how the cell works, nor why it can maintain its voltage and polarity with such a wide variation of feedstock. Equally puzzling, but gratifying, is the cell's ability to operate at room temperature, given that it has a solid electrolyte. The more practical consequence of this design, however, is that it offers prospects of easy fabrication, cheap production and high power density.

Replacing the quartz substrate with Kapton and using the acid polymer Nafion results in a cell which performs just as well as its laboratory prototype, but has the potential to be rolled up or made with a high surface area, just like electrolytic capacitors. Already power densities of around 100W/kg are achievable, while Dyer believes that 1kW/kg should be relatively easy to manage.

In concluding his paper, Dyer says that, regardless of how the cell works, the fact that it can be easily and uncritically reproduced should lead eventually to a broad range of applications from small lightweight fuel cells to new uses in information processing. (To which one could add electric vehicles,

where recharging with liquid gas or suitable hydrocarbons would give the convenience of petrol with the efficiency and environment-friendliness of electricity.)

Potential new contraceptive?

Novel ways of keeping us out of the family way know no bounds, according to a conference report published in the *General Practitioner*.

Tests on baboons in the USA have apparently shown that sperms can be immobilised and even electrocuted by means of a 'battery device' surgically inserted in the female cervix. The report, I hasten to add, emphasises that satisfactory results, i.e. no little baboons, were achieved by means of a low current that caused no discomfort to the animals (male, I hope, as well as female).



Dr Stephen Kaali, director of the Women's Medical Pavilion in Dobbs Ferry, NY, who developed the idea, reckons, however, that it will be many years before women can switch themselves on or off to order. A major problem, for example, is that of ensuring bio-compatibility, i.e. preventing rejection or some nasty interaction with body fluids.

Also, how could you ever face the prospect of telling your family that they owed their existence to a dud PP3?

Research Notes is written by John Wilson of the BBC World Service science unit.

What shape is an atom?

A fuzzy blob? A miniature solar system? The answer seems to depend on how you look at it; whether like the IBM team you use a scanning tunnelling microscope or whether, like Dr Peter Teubner of Flinders University in Adelaide, you use an active method.

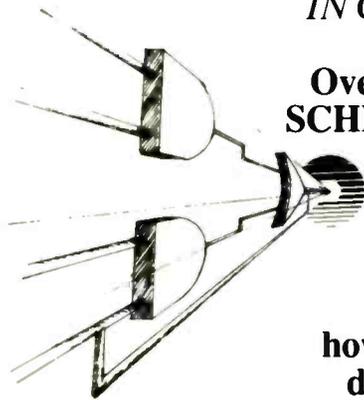
Teubner and his team are reported (*Aust Sc. & Tech Newsletter* vol. 2, no 5) to have measured the precise three-dimensional shape of a sodium atom. The technique involved using a laser to excite the atom and then firing electrons at the excited atom. By observing whether or not the electrons gained energy in the collision as the polarisation of the laser beam was varied, Teubner was able to plot an exact

representation of the atom's p-orbital — effectively the shape of the atom. So detailed are the results that the Flinders teams describe them as having all the detail of real photographs.

The actual shape of the sodium atom, though, is no less intriguing than the technique used to image it. Teubner describes it as resembling a peanut about 10^{-9} m in length! (On that basis one wonders if some of the big accelerators producing condensed matter are just expensive peanut butter machines). Perhaps the team is wise now to set its sights on calcium, which Teubner describes as "a more theoretically interesting but less ambiguous atom".

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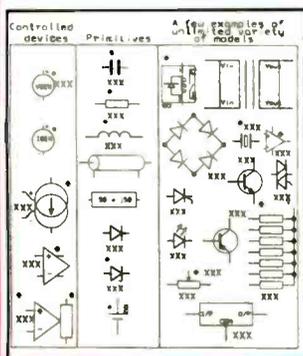
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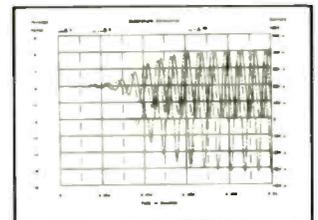
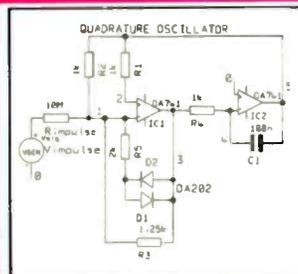
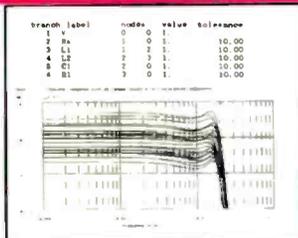


Components

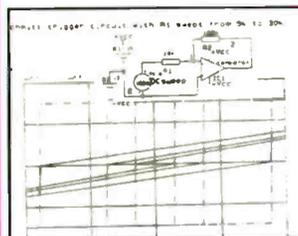
ECA-2 accepts simple two-terminal linear components such as resistors and capacitors; it includes current and voltage sources and transmission lines. Diodes are described by the exponential diode equation wherein (amongst other parameters) the user can define the emission coefficient, energy gap, temperature correction factor, and forward and reverse resistances. This enables real diode characteristics to be matched. Transistors, thyristors and operational amplifiers can also be modelled. These can be saved as macro models and a number of popular devices is supplied on the disk. Furthermore, non-linear functions can be added to any component to enable for example zener diodes and voltage-variable capacitors to be created. It is possible to define components in terms of their real and imaginary parts, for example to define the band-width or phase shift.

Statistical Analysis

A rather pessimistic worst case analysis can be run. It also performs a sensitivity analysis indicating which tolerance contribution is the most important factor, whilst R_{c2} has negligible effect. A more realistic estimate of production yield is obtained by a Monte Carlo analysis which can be tabulated or displayed as a graph. Just 25 runs of a 3rd order Chebyshev filter are shown here.



Transient Analysis This calculates circuit conditions over the prescribed time range at the prescribed intervals. This is a full non-linear analysis which is illustrated here by a quadrature oscillator. The circuit generates two sine waves in quadrature. A small initiating pulse is required and is produced by the pulse generator whose output resistance is made very large so that it has no effect on the subsequent operation. ECA-2 allows up to four points to be plotted and here the quadrature waveforms and the current in R₅ are



DC Signal Analysis

Here the analysis is carried out at a fixed temperature with the signal generators set to dc. An interesting application of this is the Schmitt Trigger where the dc command is used to step the input from -5V to 5V in 0.1V steps. The loop option then causes the voltage to reverse so that the hysteresis loop can be traced. In conjunction with the sweep command, this allows the effect of altering the resistor R₁ from 5 kΩ to 30 kΩ in three logarithmically spaced steps to be observed.

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CIRCLE ENQUIRY NO. 125 ON BACK PAGE

Weak field effect overpowers thermal noise, says report

A recent paper in the prestigious journal *Science* indicates that cellular responses to very weak oscillations in electric fields may not be overwhelmed by the ever-present thermal noise and therefore cannot be dismissed on theoretical grounds. The authors, Dr James Weaver of the Massachusetts Institute of Technology, Cambridge, Mass., and Dr Dean Astumian, of the National Institute of Standards and Technology, Gaithersburg, Maryland, state that thermal noise causes randomisation of cellular processes and is important because it is fundamental, and because other biological response mechanisms have a threshold close to the thermal noise limit.

In their paper in the 26 January issue of *Science* (247; 459-62), titled, "The response of living cells to very weak electric fields: The thermal noise limit," the authors present a physical model in

which cells are considered as possible detectors of very weak periodic electric fields. This yields a general relation between cell size and both thermally induced fluctuations in membrane potential and the maximum change in membrane potential caused by an applied field.

They show that a small but repeated electric field concentrated in a narrow band of frequencies can trigger transformations in the shape of macromolecules, especially enzymes bound to cell membranes. The simplest version of the model provides a broadband estimate of the smallest applied electric field to which membrane macromolecules can directly respond (about 10^{-3} V/cm).

Based on the assumption that the molecules are sensitive to specific frequencies and that the cell has a mechanism to enable the signals to

generate a cumulative effect over time, Weaver and Astumian provide models to show that thresholds for electric field effects can be reduced by a factor of 100,000 below the thermal noise level, reaching levels as low as 4×10^{-6} V/m at 100Hz and 10^{-6} V/m at 1kHz. According to the authors, the optimum coupling would occur in the range of 100Hz to 1MHz, based on the relaxation times of transitions between different structural arrangements of enzymes.

They conclude: "We have shown that the lower limit for a minimum detectable field imposed by competition between an applied field and thermal noise is small. Other sources of noise that might lead to larger minimum detectable fields are not considered but should be investigated. However, the estimates presented here argue that concerns due to very weak environmental electric fields cannot be dismissed on grounds of being swamped by thermal fluctuations."

The model proposed by Weaver and Astumian implies that frequency "windows", like those discovered by American researchers Drs Ross Adey and Carl Blackman, would be expected. Their model is also in line with the experimental and theoretical work of biophysicists like Prof Herbert Frohlich at the University of Liverpool and Dr Cyril Smith at Salford University, showing that, contrary to the view that any electromagnetic field bioeffects are likely to be due only to induced tissue currents, high coherent electromagnetic fields can disturb cellular functioning at very low intensities by acting on coherent internal cellular fields.

Further recent experimental evidence that such coherent electromagnetic processes act as the engine for biological dynamics is reported in another paper, by Emilio del Giudice, a research nuclear physicist at the University of Milan, and others (including Dr Cyril Smith) in the journal *Physica Scripta* (1989; 40: 786-91) entitled, "Magnetic flux quantisation and Josephson behaviour in living systems."

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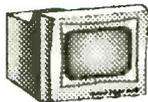


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CIRCLE ENQUIRY NO. 126 ON BACK PAGE

In Brief

It should soon be possible to send short packets of data and facsimile transmissions via the maritime satellite system. The International Maritime Satellite Organisation has successfully shown that its Inmarsat C communications network has these capabilities.

Short data reports, of up to 32 characters, can be sent in a few tens of seconds to almost anywhere in the world, from either a fixed or mobile terminal. This will allow, for example, management and monitoring of unmanned lighthouses; ships' owners or charterers can obtain positional data; and ships can themselves transmit weather data to a base station. And it will be faster and more reliable than either radio telex or morse code.

US Vice-President Dan Quayle was in Europe recently visiting the Headquarters of the European Space Agency. He is chairman of the US National Space Council and thus in charge of developing policy on the final frontier. He met with Prof. Reimar Lüst, ESA Director-General, to discuss joint NASA/ESA operations such as the proposed International Space Station Freedom, and President Bush's plans to send humans to the Moon and Mars.

The extent of the Vice-President's personal involvement with the latter was unrevealed.

A growing part of British Telecom's enormous annual profit comes from the sale of natural fertiliser. This operative is scraping pigeon guano into one of the huge dish-shaped collectors sited at the top of London's Telecom Tower, about 130m above ground level.

BBC denies bowing to political pressure

A controversial interview for a BBC TV programme on the possible health risks from overhead power lines was dropped after electricity industry chiefs complained to the corporation's director-general, Michael Checkland.

The programme, in the *Nature* series, broadcast on March 22, had to be quickly re-edited after Mr Checkland requested the deletion of an interview with Dr Robin Cox, chief medical officer of the Central Electricity Generating Board (CEGB).

In the interview Dr Cox apparently appeared flustered over a question about his reaction to a statement by his predecessor, Dr John Bonnell, in a 1979 letter written to a Mrs Stella Ross, of Innsworth, Gloucestershire, in which he repeated, "my firm assurance that the overhead power lines owned and operated by CEGB at Innsworth and elsewhere will not cause ill effects to your health."

Dr Cox stopped the original interview, complaining of having been ambushed by the question about the letter, which nonetheless had been in the public domain since it was quoted in the book "Electromagnetic Man" by Cyril Smith and Simon Best, published by J. M. Dent in July 1989.

After Dr Cox had read the letter, a second interview was conducted in which Dr Cox described the statements in Dr Bonnell's letter, as "expressing his personal view," despite the fact that they were written on CEGB-headed

paper and signed in his capacity as then chief medical officer.

According to a report in the *Independent on Sunday* (April 1), *Nature* told Dr Cox at the time that it would not promise to use the second interview. However, on March 19 Mr Checkland received a letter from Tim Beaumont, chief public relations officer of the National Grid, asking for reassurance that the first interview would not be used.

Mr Checkland then told Ron Neil, managing director of BBC regional broadcasting, to intervene, according to the *Independent on Sunday* report, though a BBC spokesman denied it was acting under pressure from the electricity industry, saying that the "concern was with fair play", and that "no journalistic points" were changed. However, the production team were "incensed" by what they definitely saw as censorship.

The newspaper reports that Mr Beaumont's approach to Mr Checkland was prompted by a desire to "protect the position and reputation of Dr Cox" and not to protect the industry itself from potential embarrassment. Mr Beaumont claimed Dr Bonnell was "writing in a particular context" and said that there had been no change in the CEGB's position. He is quoted as commenting: "What we say is that we cannot establish any causal link (between power lines and ill health) and we've always said it."



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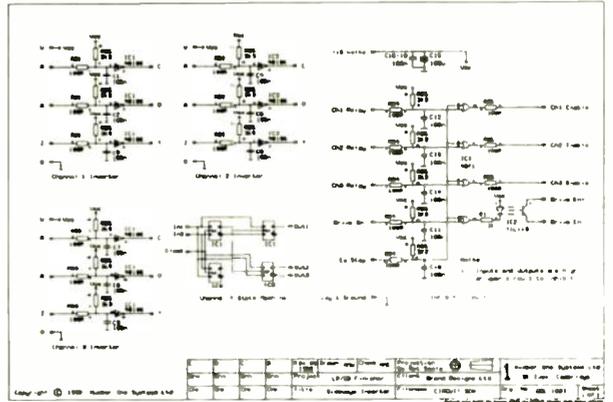
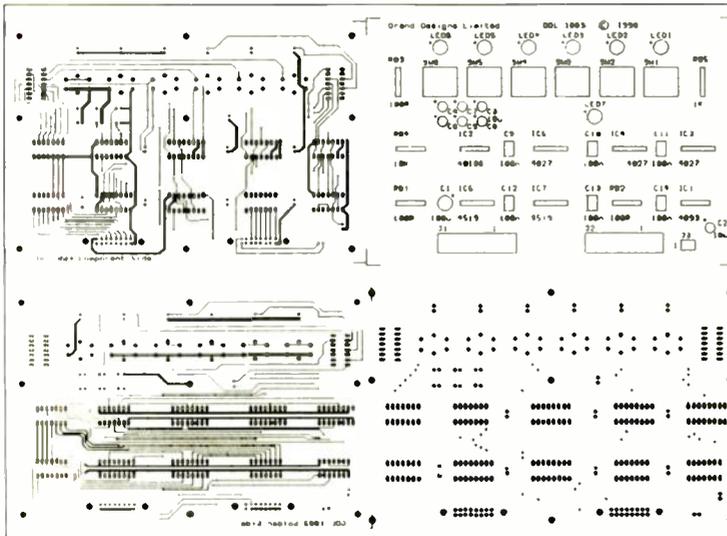
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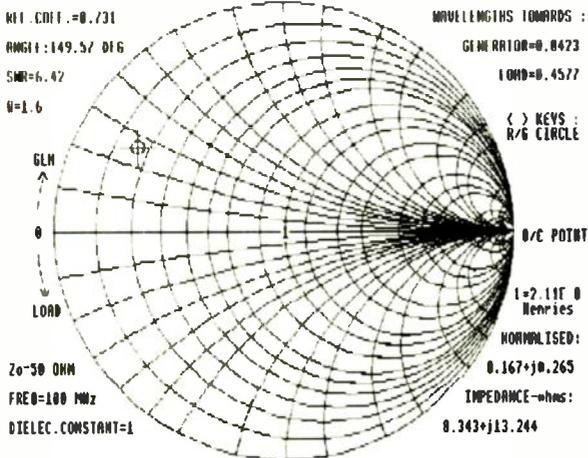
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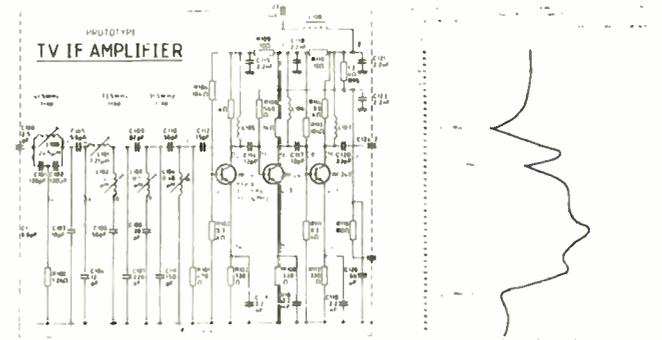
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CIRCLE ENQUIRY NO. 127 ON BACK PAGE

Welsh wary of US radar plot

US Navy plans to site a relocatable over-the-horizon radar (ROTHR) at Brawdy in Wales are causing concern among local residents over possible hazards to health. A local campaign group, Pembrokeshire Against Radar Campaign (contact Lin Miles, 0437 721596), has already been formed after meetings of the local people.

Details of the plans were published in a US Navy budget document for 1991 and uncovered recently by the *Wales on Sunday* newspaper. Although they are public knowledge in the USA, only a handful of key MoD officials here were aware of them. The plans involve a joint US/MoD project, with the UK paying approximately 90 per cent of the total £10 million development cost. The US will pay a further \$90 million for the radar and computer system itself.

The land required is stated as some

37,653 acres (nearly 60 square miles) although a University of Edinburgh military specialist, Michael Spaven, is quoted by *Wales on Sunday* as suggesting that 2,000 acres (around 3 square miles) would be closer to the mark for this type of base. Neither the US Navy nor the MoD would clarify the point.

The ROTHR receiver site and operations centre are to be at Brawdy but the transmitter site (so far undisclosed) will be at least 30 miles away for technical reasons.

The radar's range will be at least 1,800 miles, focused on the Baltic Sea, with a vision span of 63°, and will form part of the US Navy's worldwide network of over-the-horizon radars. Few transmission details are available, except that it will transmit in the frequency range between 5 and 28MHz

and at a peak power of 200kW.

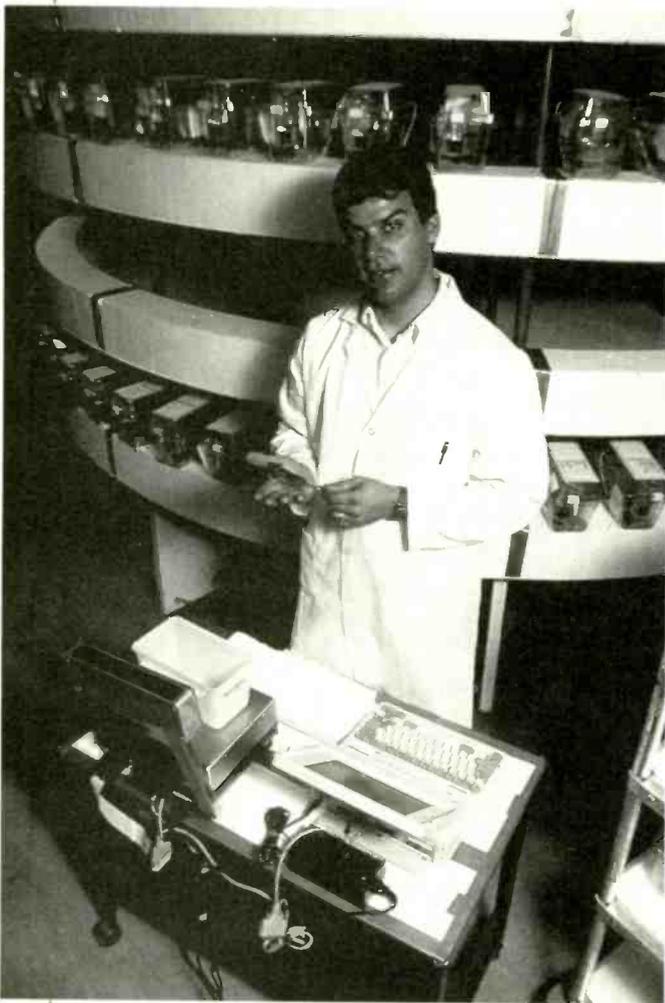
According to Jane's Defence Weekly (7 April), the receiver aerial system needs to be 2590m long, consisting of two parallel rows of 372 supporting poles and 43 equipment shelters. Wherever this is sited it will have a major environmental impact. From a military viewpoint the radar could even pose a danger to aircraft at the nearby RAF airfield, possibly having to be switched off each time an aircraft takes off or lands because of the danger to the electronics on board the aircraft.

In the USA there has been concern and debate for some time about the possible health hazards from the radar, known as PAVE PAWS (Precision Acquisition of Vehicle Entry Phased Array Warning System), sited at such locations as Upper Cape Cod in Massachusetts and Beale Air Force Base in California, and which operates at 450MHz, with a repetition of 18.5Hz.

Much of the evidence and debate is presented in a recent book by *New Yorker* journalist Paul Brodeur ('Currents of Death,' New York, Simon & Schuster, 1989). An overview of the more general research linking health problems with exposure to radar and microwave systems was featured in the March issue of this magazine ('Killing Fields—The Microwave Health Hazard,' pp208-11).

Most recently, a Chinese study reported in the *Journal of Bioelectricity* found significant deficits in visual response, memory and an indicator of immune system function in those living near radar and radio installations for as little as a year. A Polish study of military personnel exposed to radar—by Dr Stanislaw Szmigielski at the centre of Radiobiology and Radioprotection in Warsaw, has found an increased cancer risk of over threefold in certain groups.

It is these and other results that worry local Brawdy residents and which MoD officials will have to try to circumvent. The Ministry of Defence will have to notify the local planning authorities before proceeding with the development but that may be the only chance the public has to comment on the scheme, since the Government has no obligation to put the proposals before Parliament.



Assessing the health effects of low-level radiation: scientists at Georgia Tech, Atlanta, USA, use mice to determine the level of cancer pathogenesis associated with occupational microwave exposure, for instance to air traffic control personnel. The study, one of the largest ever conducted concerning the effects of low level microwaves, will take 18 months to complete. Photo Gary Meek

Current amplifiers from voltage op-amps

Over the years, the semiconductor industry has tried to provide circuit designers with a cheap, high quality, versatile analogue building-block, often with the sole aim of producing a controlled voltage output from a voltage input. The most popular of these networks is the conventional utility, high-gain, voltage-mode op-amp (referred to in this article simply as an op-amp). It has become customary for engineers to think of analogue signal processing in terms of voltage variables and this tendency has resulted in numerous voltage signal-processing circuits.

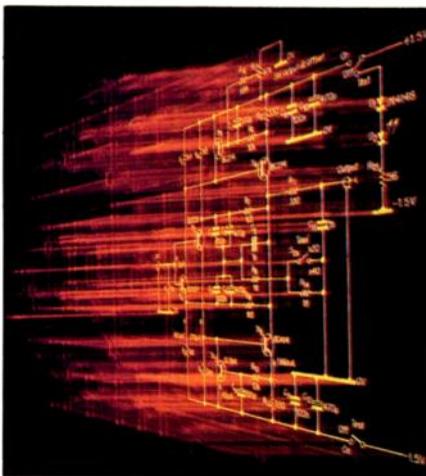
For example, the op-amp is easily configured into controlled voltage-source amplifiers, such as the voltage-controlled voltage source (VCVS) and current controlled voltage source (CCVS or transresistance amplifier), but not so easily into controlled current-source amplifiers such as the VCCS or transconductance amplifier and CCCS. Why should this be so?

Well, negative feedback can be derived by sampling either output voltage or current into the feedback network. Voltage sampling is easy; all that is required is to connect to the output node and ground. The result of voltage sampling is that the output impedance reduces from its 50–100Ω open-loop value, making the closed-loop output performance of the circuit approach an ideal voltage source.

A high output impedance requires a feedback signal proportional to output current; that is, output current sampling. The architecture of the standard '5-leg op-amp' makes this very difficult and current output circuits (CCCS and VCCS) are much less easy to design than voltage-output circuits.

Traditional methods

The most commonly used amplifier



Supply-current sensing extends the capabilities of conventional voltage op-amps to give current-mode operation. John Lidgley and Chris Toumazou survey the methods used and present a universal circuit

having a controlled bipolar current output, and frequently encountered in standard textbook designs^{3,4}, has the load in the feedback loop of a voltage op-amp, as shown in Fig. 1. Although the amplifier's voltage gain varies with R_L , the current in the feedback loop remains fixed, assuming a fixed V_{IN} and R .

A practical limitation of this approach is that the current-driven load

cannot be grounded. Despite this drawback, the circuit does have some interesting features: nearly all the input current is drawn through the load by the action of negative feedback. The input current only differs from the load current by the current flowing into the op-amp, which is extremely small due to the high open-loop voltage gain and input impedance of the op-amp, so the current gain of the circuit is very close to unity.

The circuit shown in Fig. 2 and developed by Howland³ solves this problem of grounding. It acts as a current source, providing $I_L = -V_{IN}/R_4$, for the condition that $R_3/R_4 = R_2/R_1$. If the ratios of these resistances are equal, the circuit will function with a theoretically infinite output resistance, determined by the combined positive and negative feedback action of the op-amp.

However, a major drawback of this combined positive and negative feedback approach is that very small departures from ideal balance conditions either drastically reduce the output resistance or result in instability because the output resistance becomes negative!

A detailed analysis based on an ideal op-amp gives

$$I_L = -V_{IN} \left[\frac{R_2}{(R_1 R_3)} \right] - \frac{V_L}{\left[\frac{R_3}{(R_3 R_4)} - (R_2/R_1) \right]}$$

and

$$r_{out} = -\delta V_L / \delta I_L = R_3 / \left[\frac{R_3}{(R_3 R_4)} - (R_2/R_1) \right]$$

From this equation the output impedance is seen to be negative if the practical values of resistors are such that the denominator becomes negative, which will occur if $[R_3/R_4] < [R_2/R_1]$. If this condition is not met it will not necessarily result in instability, since it depends on the value of load resistance. But if the parallel combination of the output resistance and the load is negative, the circuit will become unstable and will either oscillate or latch-up.

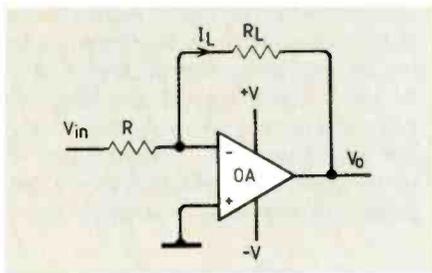


Fig. 1. Most common current-output amplifier, with load in feedback loop. Only possible with floating loads.

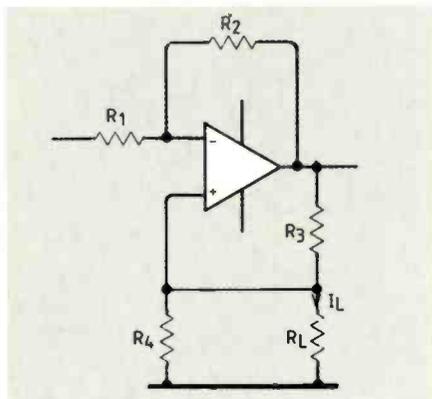


Fig. 2. Circuit by Howland, avoiding restriction of Fig. 1 circuit. Accurate resistor matching required for stability.

The problem arises when the positive feedback exceeds the negative feedback. An alternative way of viewing the circuit from a stability stand-point is to look at the two feedback factors. The negative feedback factor B_N is determined by the potential divider ratio $B_N = R_1/(R_1 + R_2)$ and the positive feedback factor B_P by the ratio $B_P = (R_4/R_L)/(R_3 + R_4/R_1)$. For stability the inequality $B_P < B_N$ must be satisfied. Working through the algebra yields the same result, that $(R_3/R_4) < (R_2/R_1)$ if the circuit is to be stable for the worst case of R_L open-circuit and, perhaps more practically, $(R_3/R_4) < (R_2/R_1)$, where $R_L' = R_L/R_4$.

Other practical problems with the circuit relate to the fact that the positive feedback increases with decreasing load, resulting in a progressive loss of all the well known advantages of negative feedback. For example, the bandwidth decreases with reducing load and the output offset voltage will be more significant with decreasing load.

The Howland design is typical of many similar controlled output-current amplifier topologies which are all based on a combination of positive and negative feedback and which all have the same problem of requiring closely matched resistor networks; as a conse-

quence, they are potentially unstable if the matching is not perfect. To overcome this difficulty, an alternative design approach is required in which positive feedback is not used.

Extending the op-amp

Output current from an op-amp's output terminal must come from the power supplies. By Kirchoff's current law for the standard 5-leg op-amp shown in Fig. 3, if there are no other signal paths to ground, $I_1 + I_2 + I_3 + I_4 + I_5 = 0$ and, since I_1 and I_2 are close to zero, then $I_3 + I_4 = -I_5$, so that by sensing the supply currents, one can derive a current proportional to output current.

The technique of supply-current sensing to provide the conventional op-amp with a well-defined current output facility was first reported by Graeme⁵ in 1974. He showed that a precise VCCS could be achieved by using a pair of complementary field-effect transistors to sense the current flowing in the supply rails of a conventional op-amp, as shown in Fig. 4. Opposing fet current sources Tr_1 and Tr_2 are controlled by the high-gain feedback around the op-amp. The circuit is symmetrical and behaves as a self-balancing bridge; a voltage difference at the input is balanced out by the op-amp's negative-feedback action, causing the input voltages to the op-amp to be virtually equal.

The difference in fet currents produces the output current, the difference current being controlled by comparing the input voltage V_{IN} to the feedback voltage provided by the current sensing source resistors R_s . Input voltage controls output current to within the accuracies of the resistors selected and within the gain-bandwidth and power-supply rejection limitations of the op-amp.

The circuit works well and avoids the output impedance uncertainties of earlier Howland-type designs. However, as a consequence of using two independent negative feedback connections, one for each output polarity, the circuit performance is again sensitive to resistive mismatch.

Supply-current sensing was not an entirely new mode of op-amp operation, since it had been previously used as a means of power boosting a low-power op-amp.

In 1975, Hart and Barker⁶ suggested an alternative method in their realisation of a class-B voltage-to-current converter, shown in Fig. 5. The circuit configuration is similar, with the op-amp as its main gain block, together with a set of complementary current-mirrors used

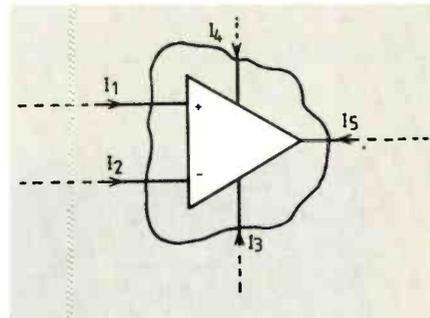


Fig. 3. Current flow into standard five-pin op-amp. Sum of supply currents equals output current.

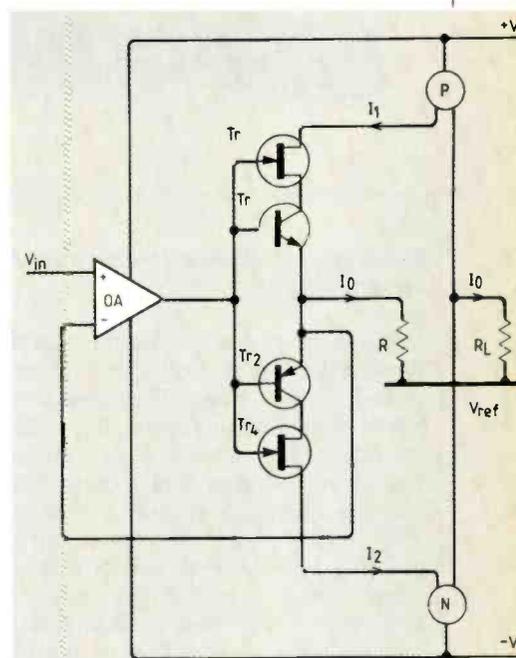


Fig. 5. Circuit due to Hart and Barker, using current mirrors to sense currents in class-B output stage. Accurate resistor matching unnecessary.

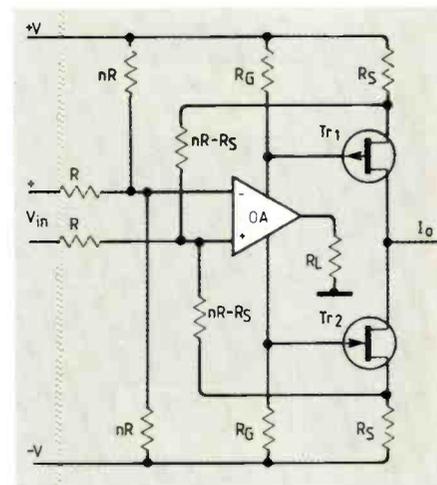
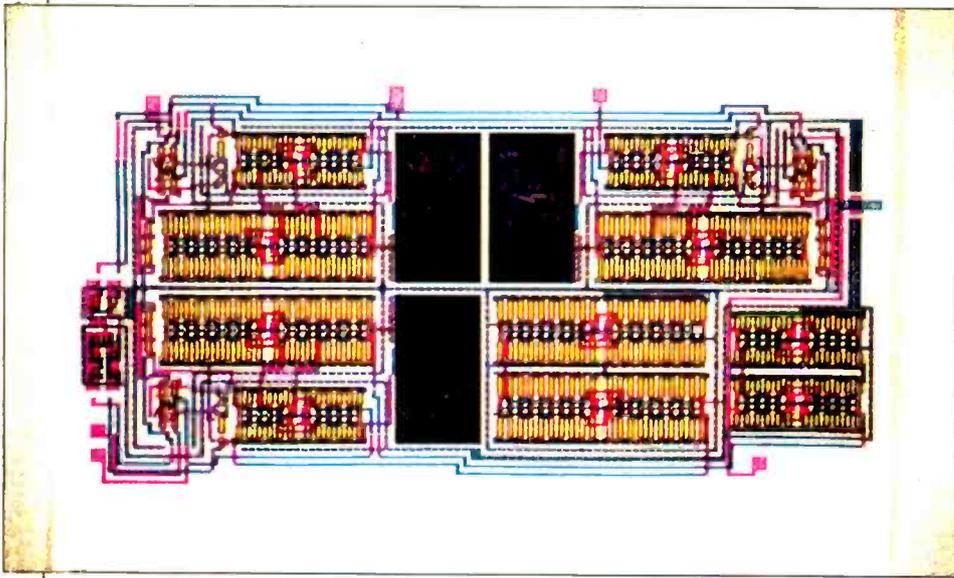


Fig. 4. Graeme's circuit for supply-current sensing with standard op-amp. Difference in fet currents is output current.

DESIGN



C-mos op-amp chip using the circuit of Fig. 16. Voltage gain is 63dB, unity-gain bandwidth 2MHz.

to sense the phase split output current flowing through the collector leads of a class-B output stage. The class-B cascode transistor connection, Tr_1 to Tr_4 , provides the circuit with high common-emitter current gain, a high breakdown voltage and high output resistance. Positive and negative current-mirror circuits, denoted by P and N respectively, could be the improved 4-transistor G.R. Wilson⁷ current-mirror, together with an output fet to improve the overall output impedance and maximum output-voltage swing.

Although Hart and Barker's class-B scheme catered for much higher load-voltage excursion than did Graeme's

proposal and avoided the undesirability of resistor matching, the class-B mode in which the circuit operates resulted in considerable crossover distortion. This problem was identified by Rao and Haslett⁸, who showed that much better high-frequency performance and improved output current drive could be obtained if the output circuit were operated in class-AB. In further work⁹ they related the class-AB voltage-following action to that obtained by the classical class-AB push-pull output stage of an op-amp and showed how the output signal current could now be sensed via a current-mirror arrangement, as shown in Fig. 6.

This led Hart and Barker¹⁰ to use complementary current-mirrors as external current-sensing elements, which resulted in a practical class-AB versatile op-amp structure, shown in Fig. 7, connected as a transconductance amplifier. This circuit is a current-conveyor; a very versatile analogue building-block^{11,12}.

Using the circuit of Fig. 7, they developed a universal operational amplifier converter, shown in Fig. 8. In this design the arrangement of feedback resistors within the op-amp circuit sets the amplifier to any one of the four main amplifier types, namely voltage, current, transresistance and transconductance.

Poor high-frequency performance of these schemes, due to the lateral p-n-p transistors available at that time, was avoided by Huijsing and Veclenturf¹³,

who replaced the positive-supply sensing current-mirror by an elegant negative current-mirror source, shown in Fig. 9, which used a local amplifier with n-p-n transistors. As a discrete design, the high-frequency performance was now limited by amplifier A of the simulated current-mirror source, but a

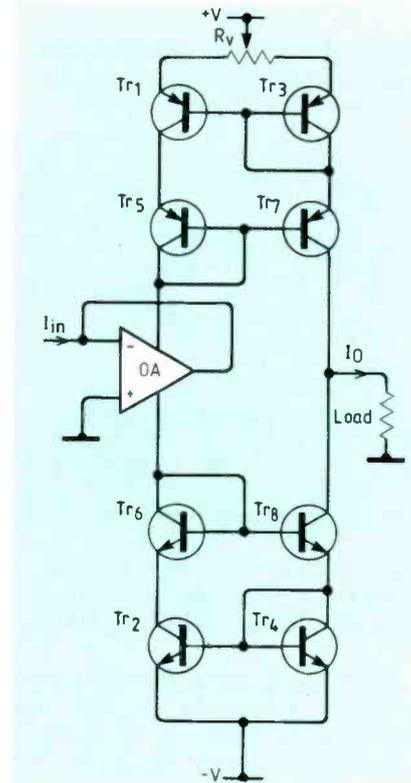


Fig. 7. New circuit by Hart and Barker using complementary current mirrors to give class-AB structure in current conveyor.

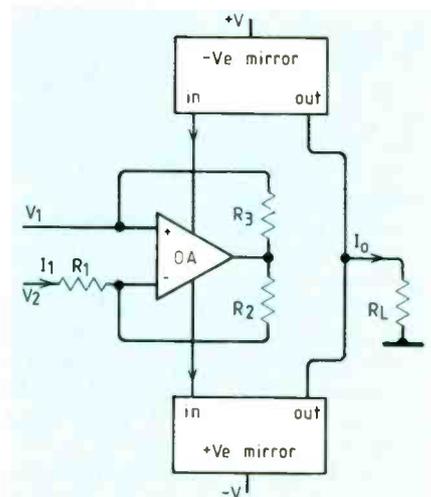


Fig. 8. Universal op-amp converter, developed from circuit of Fig. 7. Basic circuit will operate in any of the four configurations, depending on feedback arrangement.

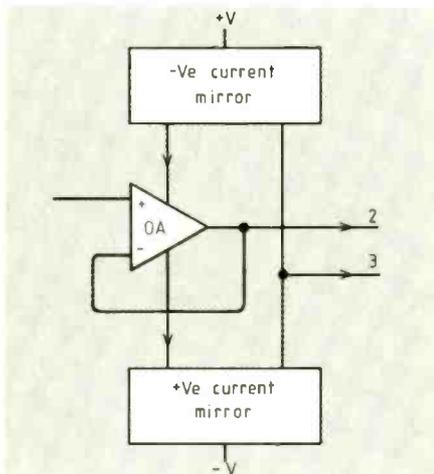


Fig. 6. Rao and Haslett's circuit, which uses class-AB output characteristic of op-amps to allow current sensing in op-amp supply leads by current mirrors.

single-chip realisation was built later, providing a much better high-frequency performance.

A further problem encountered with these class-AB current converter designs is the current output limitation imposed by either the op-amp or transistors used in the current-mirrors. Nedungadi¹⁴ showed that, using a similar converter structure to that in Fig. 8, together with a high current output class-AB amplifier, current outputs greater than 100mA could be obtained while still maintaining high conversion accuracy and high efficiency. This high-current converter technique is shown in Fig. 10.

The essential difference between Fig. 10 and previous schemes is that the outputs of the current-mirrors are fed back to the input of the circuit, while the converter output is taken from the op-amp. This arrangement ensures that only the input current is passed through the current-converter section of the circuit, irrespective of the output current magnitude. For example, with a current gain A, of -100 and an output current I_o , of 100mA, the converter need only supply 1mA, which is easily and accurately achieved using standard op-amps and transistor arrays.

Wilson¹⁵ determined that the output signal distortion in the converters shown in Figs 3-8 was due to the current mirrors being connected in open-loop at the output of the circuit, where collector-voltage modulation effects are significant. By adopting a similar approach to Nedungadi and connecting the current-mirrors in a feedback arrangement, this source of voltage modulation could be reduced considerably.

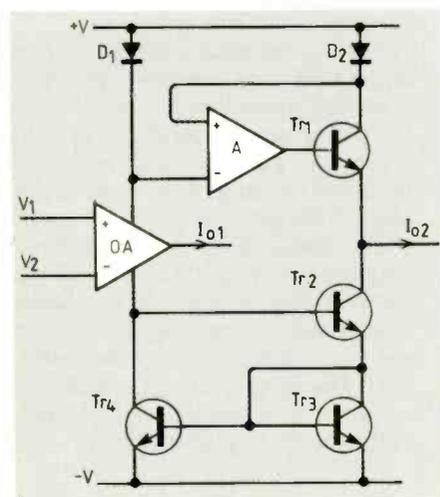


Fig. 9. Improved positive-supply current mirror by Huijsing and Veelenturf gives better high-frequency performance.

During this work he introduced a current-mirror symbol which simplifies diagrams. This symbol and circuits for the low-distortion series and shunt feedback converters are shown in Figs. 11, 12 and 13 respectively. Wilson went on to show how previous current amplifier designs, based upon the open-loop mirror approach, could also be improved to the low-distortion type by simply connecting the mirrors in shunt feedback. In this way, the main distortion component of the converter is referred to the input side of the op-amp. Unfortunately, the current transfer accuracy of the converter is still limited by the current transfer performance of the current-mirror circuits. Furthermore, this feedback arrangement results in much poorer frequency performance than with the 'open-loop' converter structure.

This new trend of using current-mirrors to sense the op-amp's output current and provide well-defined bipolar output properties proved far superior to traditional feedback techniques.

Current followers

The current-follower is the current equivalent in the current domain to the voltage-follower in the voltage domain. It is a circuit with extremely low (ideally zero) input impedance and extremely high (ideally infinite) output impedance. When used with a signal source, it produces a current drive to a load equal in value to the short-circuit current obtainable from the input signal source.

Figure 14 shows a simple but high-performance current-follower which was featured in an earlier article¹⁶, using a 741 op-amp and CA3096 transistor arrays for the current-mirrors. The circuit has a current-gain equal to the current transfer ratio of the current-mirrors, so it is important to use high-quality current-mirrors.

One of the main features of this circuit is its wide bandwidth due to the load isolation from input to output. Also, because the op-amp is connected

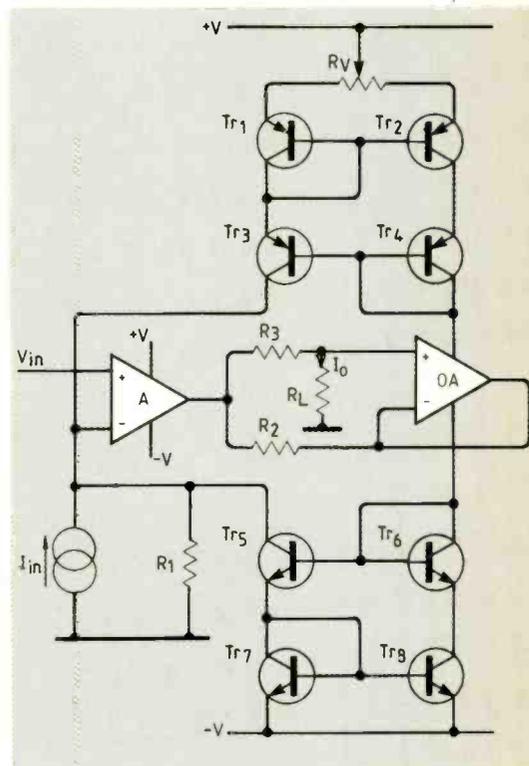
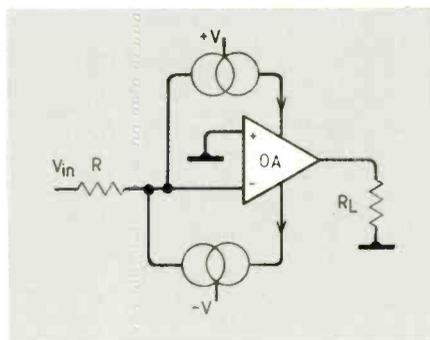


Fig. 10. High-current amplifier by Nedungadi, using circuit similar to that of Fig. 8, but with class-AB amplifier at output.

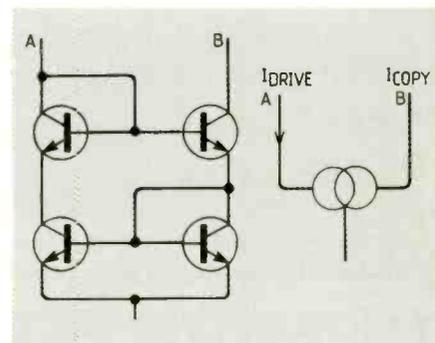


Fig. 11. Current mirror symbol, originated by Wilson.

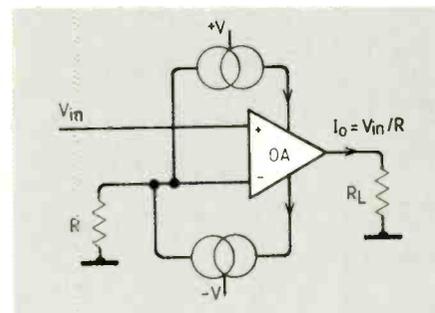


Fig. 12. Low-distortion series-feedback current converter by Wilson, with current mirrors in feedback loop to avoid collector-voltage modulation.

Fig. 13. Shunt-feedback converter, using ideas of Fig. 12.

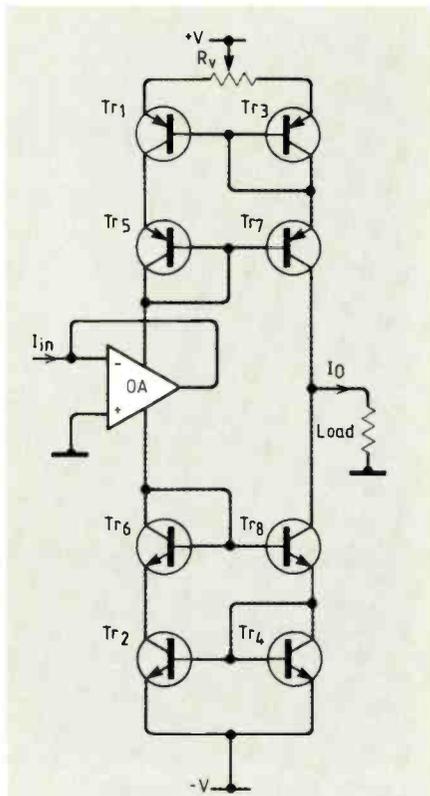


Fig. 14. Wide-band current follower by Lidzey and Toumazou, in which load is isolated from input.

as a voltage-follower with a grounded non-inverting input terminal, the output node of the op-amp is held at virtual ground, providing a very low input impedance and high slew-rate capability, since the op-amp has no appreciable voltage signal swing at its output.

An interesting universal follower-based amplifier was described in ref.16 in which, using only two current-followers and two voltage-followers, it is possible to configure any of the four amplifier topologies (VCVS, CCCS, CCVS and VCCS) without global feedback simply by using interstage resistors, either in series to define input current from a voltage drive or in shunt to ground to provide input voltage from a current drive.

Figure 15 illustrates such a follower-based voltage amplifier using input and output voltage-followers, with an intermediate current-follower to provide voltage gain defined by (R_2/R_1) . One of the most attractive features of this structure is that the normal gain-bandwidth trade-off does not occur; the bandwidth is independent of gain setting and remains virtually constant.

A further feature is the lack of overall output to input feedback. Stabilising negative feedback is used within each

follower block so that no additional feedback is necessary when the gain-defining resistors are added. As a result, any phase lag from input through to output does not affect the stability of the amplifier.

In Fig.15, the bandwidth of each voltage-follower section is close to the gain-bandwidth product of the op-amp. The frequency performance of the current-follower section, shown in ref.16, is determined by the driving source impedance R_s , which in this case is R_1 . The larger the value of R_1 , the higher the frequency performance. Resistor R_1 can be chosen to maximise the bandwidth of the amplifier and the voltage gain can be set independently with R_2 . High gain and wide bandwidth can thus be set simultaneously.

However, there is a limitation on the size of R_2 and hence voltage gain and bandwidth of the amplifier due to the limited output impedance of the current-mirror circuits. Experimental results using conventional op-amps have indicated improvements of more than 50 times the gain-bandwidth capability of the individual op-amps used in the system.

Seven-terminal op-amp

Rather than the five legs of a conventional op-amp, it would be very useful if seven terminals were available, the additional two being the collectors (drains) of the output push-pull pair. This would then allow direct output-current sensing, rather than unnecessarily sensing the whole of the supply current. Such a modification could be easily carried out by semiconductor manufacturers and the device versatility would be extended considerably to allow true output-current sensing, rather than full supply-current sensing.

Sensing the entire supply current is undesirable as it leads to unnecessarily high shot-noise levels. Also, although the power supply rejection of op-amps is generally good, it would be better not to modulate the power-supply connections to the input stages of the op-amp.

The circuit diagram of a c-mos op-amp with uncommitted output drains is shown in Fig. 16, in which the additional two terminals are created by simply taking the output drains to two external pins. The op-amp shown, together with on-chip c-mos current-mirrors, has recently been implemented as an in-

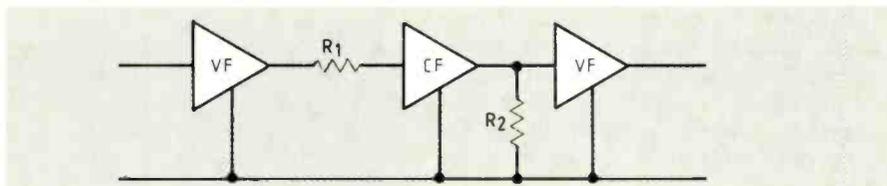


Fig. 15. Follower-based amplifier by Lidzey and Toumazou without overall feedback. Central current follower provides gain.

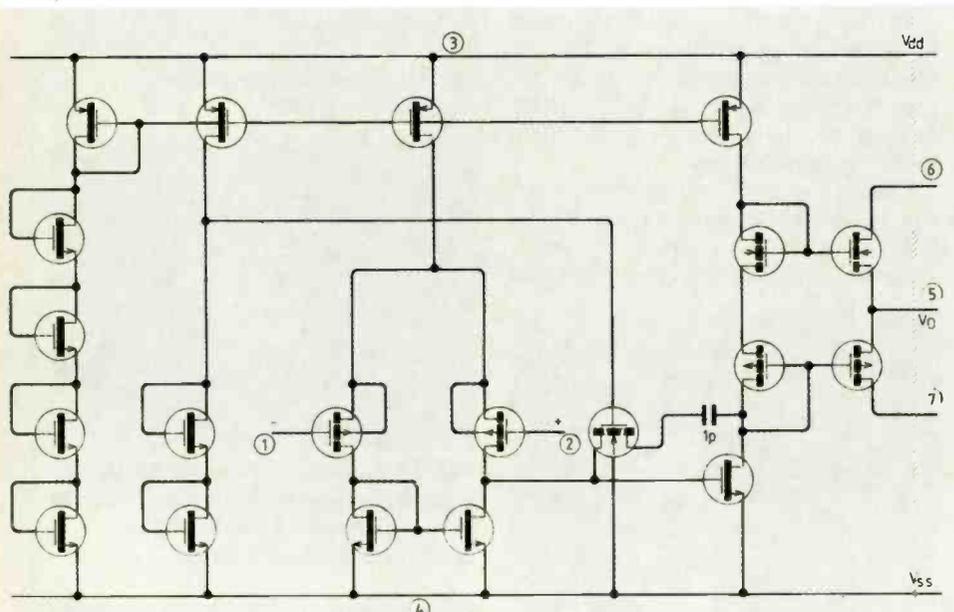


Fig. 16. Authors' c-mos op-amp with free output drains, fabricated in 2micron N-well process.

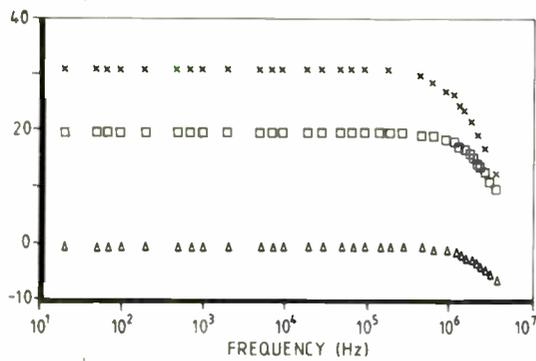


Fig. 17. Measured voltage response of circuit of Fig. 16. No change in bandwidth occurs at gains from 0dB to 30dB.

egrated circuit in a 2µm N-well c-mos process. The circuit comprises three matched 7-terminal op-amps and four high-performance current-mirrors, allowing configuration into any of the supply-current sensing circuits. To ensure good matching, all the op-amps are placed closely together and share the same bias circuit.

Figure 17 shows the measured gain/frequency response curve of the seven-terminal op-amp connected as a current

follower and used in the universal follower-based voltage amplifier of Fig.15. The voltage-followers are the remaining two op-amps on the chip. Results demonstrate that the gain can be varied from 0dB to 30dB with no significant change in bandwidth, as expected. ■

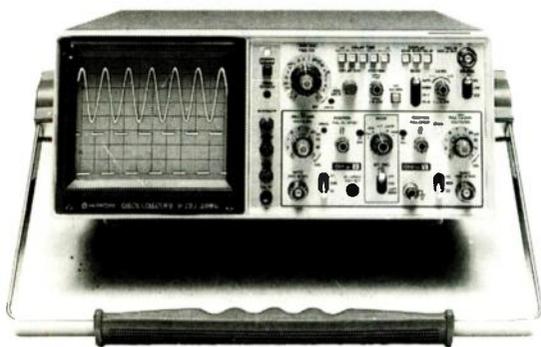
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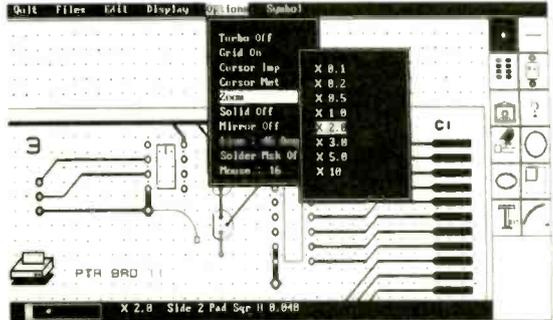
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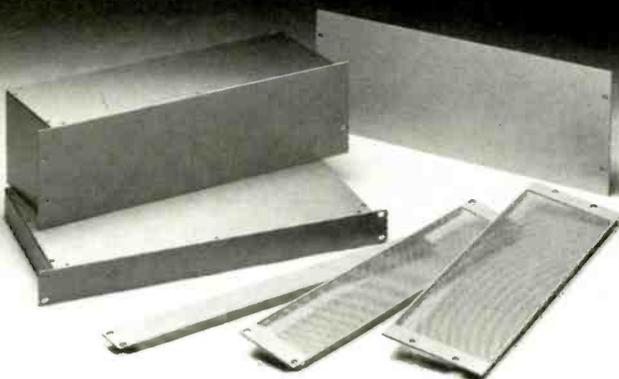
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Once a sequence of electronic keyboards has been built up at a multi-track recording session, one may wish to add conventional instruments to the mix, for instance guitar, vocals or a saxophone solo. Since these instruments can be added live to the synthesiser parts it is both a shame to sacrifice quality and a costly luxury to use a multi-track to piece together the final touches to the mix. Far better to use a mixer to add the saxophone or vocals and guitar to the synthesiser parts and record the result as the master-tape.

The requirement for such a mixer might be: six (mic or line level) input to two output; effect-send sub-mix (in my case to an echo chamber); some form of tone control and clear metering of all inputs (including effect-return); effect-send sub-mix and stereo output.

The block diagram for the mixer filling the specification is shown in Fig. 1. The tone control circuit is a unity gain circuit (all controls flat) so it may be substituted after each input pre-amp to achieve this alternative configuration.

The main justification for using discrete components lies in the decision to rely on a battery supply for the mixer. A further design decision was the choice of the domestic/semi-professional signal amplitude level of 0VU, equivalent to -10dB(V) . The VU is a unit intended to express the level of a complex wave in terms of decibels above or below a reference volume, and it implies a complex wave — a programme waveform with high peaks. 0VU reference level therefore refers to a complex-wave power-reading on a standard VU meter.

The usual convention is to assume that the peaks of the complex wave will be 10dB higher than the peak value of a sine wave adjusted to give the same reference reading on the VU meter. In other words, if we adjust a music or speech signal to give a reading of 0VU on a VU meters the system must have at least 10dB headroom over the level of a sine wave adjusted to give the same reading if the system is not to clip the programme audio signal.

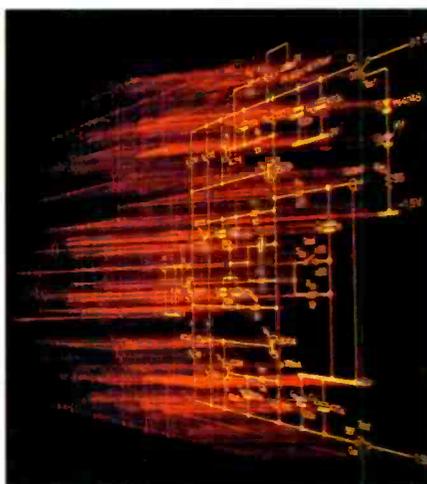
In this mixer, 0VU is set to be equivalent to -10dB(V) . The peak-to-peak value of a -10dB(V) sine wave is:

$$2\sqrt{2(1 \times 10^{-10/20})}V$$

or 894mV. A complex music or speech wave will therefore have a peak-to-peak value of 10dB (or 3.16 times) higher —

Audio mixer design

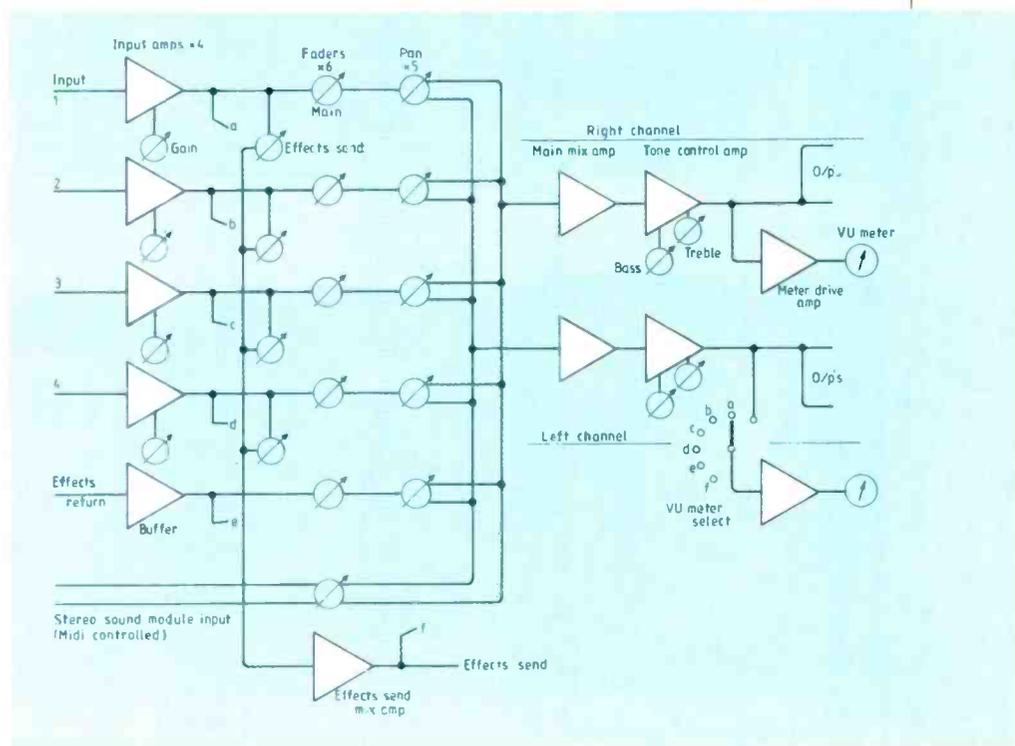
Audio mixers are relatively unsophisticated yet present many conflicting design requirements: headroom, noise contribution, linearity and current consumption. Resolving these stretches analogue design talent to the limit. By Richard Brice.



2.82V pk-pk. The mixer provides more than the stated 10dB headroom since, in practice, more headroom is occasionally necessary.

Well-designed audio circuits should certainly be able to swing 3V pk-pk when running on a single 9V rail. On the grounds of cost I decided that it would be ideal if the mixer could run from a single PP3-type battery for at least eight hours in continuous use. Low-noise op-amps like the NE5534, which would be suitable for the microphone input stages, have a typical supply current of 4mA on a 9V supply. However, the worst-case figure is

Fig. 1. Mixer block diagram.



Continued on page 578

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8mA at 25°C. This would give an I_q for the microphone input stages of 32mA which, running on a PP3 with about 100mA/h capacity, would give just three hours of use simply supplying the mic amplifiers.

Alternative op-amp series commonly used for audio, the LF355 or the TL071, have typical current drains of 1.5mA and may be considered suitable alternatives. However, they are less suitable for the mic input amplifiers. This is because (being fet input) they have a relatively large input noise voltage generator and virtually no input noise current generator, and are therefore relatively noisy when matching a low-impedance source like a microphone.

Input amplifiers

Each of the four input-stage amplifiers is formed from a transistor ring of three. The design requirement is for good headroom and a very low noise figure.

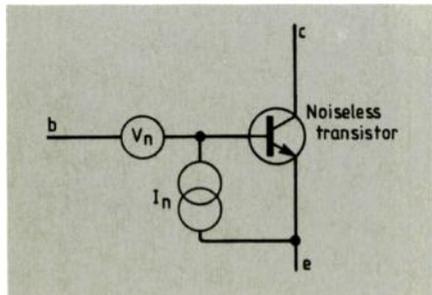


Fig. 2. The noise generator model of a bipolar transistor showing both current and voltage noise sources.

Figure 2 shows noise sources (V_n and I_n) added to a perfect transistor. V_n has the RMS value of:

$$V_n = \sqrt{4kT(r_{b'b} + \frac{1}{2g_m})B} \quad \text{volts}$$

I_n has the RMS value of:

$$I_n = \sqrt{\frac{4kTB}{(2B/g_m)}}$$

Because g_m appears in both equations we can sketch what will happen to V_n and I_n as the value of g_m changes (which is the same thing as saying, as collector current changes)—see Fig. 3. Now it is apparent that the total noise which appears as a signal at the base of the transistor will depend on the source resistance. If this is high then the noise current generated by the current noise generator must be made small because this signal will flow through the large resistance and generate a large voltage signal at the transistors base. If the source resistance is low, then the magnitude of the current generator is not important, but the magnitude of the

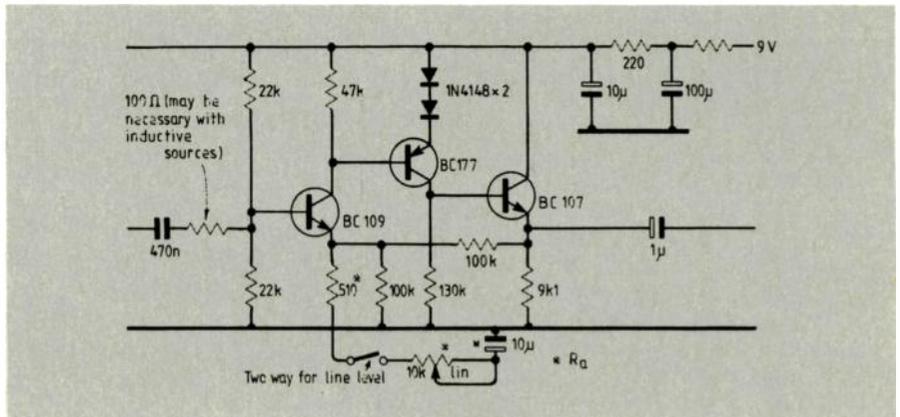


Fig. 4. Ring-of-three microphone channel amplifier.

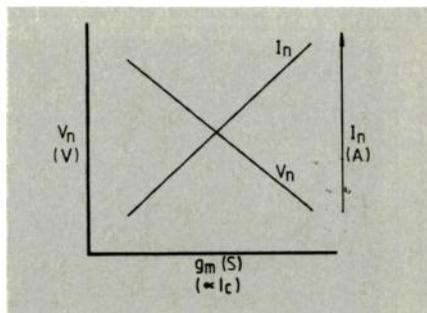


Fig. 3. Sketch of voltage and current noise sources as a function of collector current and transconductance.

voltage generator must be made small.

Referring to Fig. 3 we can see that a low source resistance would seem best matched by a high collector current input stage, and a high source resistance would be best matched by a low collector current input stage. The choice of optimum g_m for low noise is given in a simple equation:

$$g_m = \sqrt{B/R_s} \quad (\text{Ref. 1})$$

Matching the input stage to a microphone would seem to require a collector current in mA to match the low impedance of a typical microphone. Paradoxically, in the final design (Fig. 4) the first-stage collector current is about 40μA. All the above would be very straightforward were it not for flicker noise. Furthermore, while it is necessary to consider only the resistive part of the source impedance when designing noise-voltage generators, the full source impedance must be taken into account when calculating the effect of the current noise generator.

Flicker noise is a type of noise that increases its magnitude with decreasing frequency, hence the alternative name of 1/f noise. It may be modelled as an increase in the current noise generator below a certain frequency (f_f). This frequency, often called the flicker-noise

corner-frequency, is dependent on the transistor type and sample. The magnitude of I_n is better modelled as:

$$I_n = \frac{4kTB}{2B[g_m(\omega_f/\omega + 1)]} \quad \text{where} \quad \omega_f = 2\pi f_f$$

The effect of 1/f noise is further complicated because not only does its magnitude decrease with decreasing collector current (as shown in the above equation) but the flicker-noise corner-frequency itself falls with decreasing collector current. Because an audio signal contains frequencies down to 20Hz, flicker noise must be considered and its reduction ensured by the choice of a low flicker-noise type transistor, like the BC109, operating at a low standing current.

Substituting in the above equations for I_n and V_n with assumptions for the following values: k = Boltzmann's constant = 1.38×10^{-23} ; T = average room temperature = 290K; B = bandwidth = 20kHz. Typical figures for a BC109: Beta = 200; $r_{b'b} = 200\Omega$; $g_m = 2\text{mA/V}$ for an emitter current of 50μA. So V_n , which is the predominant generator for a low source resistance, is:

$$V_n = \sqrt{3.2 \times 10^{-16} (450)}$$

or 380nV RMS or -128dB(V). The important term here is the equivalent resistance term ($r_{b'b} + 1/2g_m$). Its typical value of 450Ω is not much larger than the magnitude of the real part of the impedance across the output terminals of a moving-coil microphone (230Ω). This means that the thermal noise generated in the resistive part of the microphone impedance produces a significant proportion of the total input noise. If we take a typical output voltage from the microphone to be 1mV RMS then this noise will be -68dB.

The magnitude of the current noise generator is:

$$I_n = \sqrt{\frac{(3.2 \times 10^{-16}) \times (2 \times 10^{-3})}{400}}$$

or 40pA RMS.

The total calculated noise-figure for the amplifier fed from a resistive source of 200Ω (and with the gain set to maximum) is about 12dB, this being a measure of the amount by which the total noise from the amplifier exceeds what it would be if the amplifier were totally noiseless. The noise figure of the amplifier gradually deteriorates at lower gains due to the thermal noise generated in the feedback resistors.

It might be thought that the current noise generator term derived above is so vanishingly small that we do not need to consider the increase in this generator due to flicker noise. This would indeed be the case if, as in the above calculation of noise figure, the noise current simply flowed in the microphone resistance.

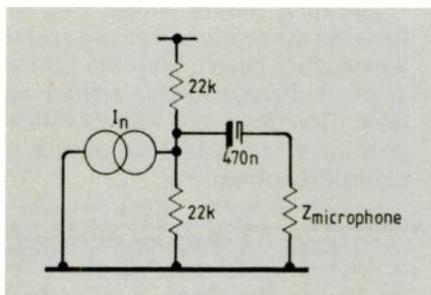


Fig. 5. Reverse view model of the microphone input stage.

However, if we draw the input stage the other way around, as in Fig. 5, it becomes very clear that the noise signal, which may be considered in every way just like the wanted audio signal, is AC-coupled into the microphone load. So, at frequencies below about 1.7kHz, the current generator develops its voltage signal not simply across the microphone impedance but across the potential divider which forms the bias supply for the input transistor. This has an equivalent resistance of 10kΩ, and the noise current flowing across it will generate a noise voltage signal of 124nV in a bandwidth of 1.7kHz — this figure ignores the increased effect of flicker noise at extremely low frequency.

The 10kΩ source resistance at low frequency requires an optimum operating g_m of 1.4mA/V and an operating current of 35μA; thus the final design, like so many other engineering solutions, is a compromise. A low collector current is used to ensure low flicker noise by matching the input stage to the source impedance at low frequency with some sacrifice of noise figure in the mid

Fig. 6. Mixer summing amplifier.

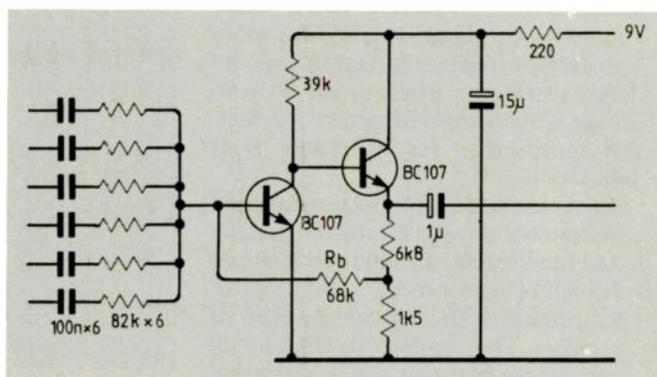


Fig. 7. Fader and panning controls.

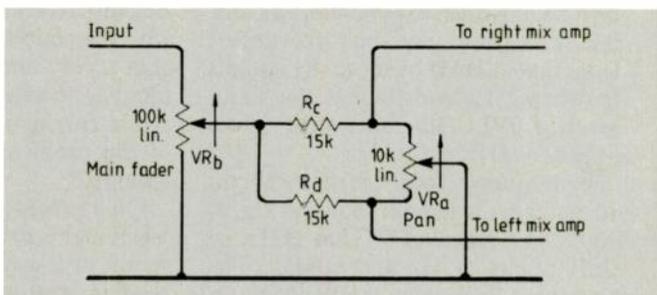
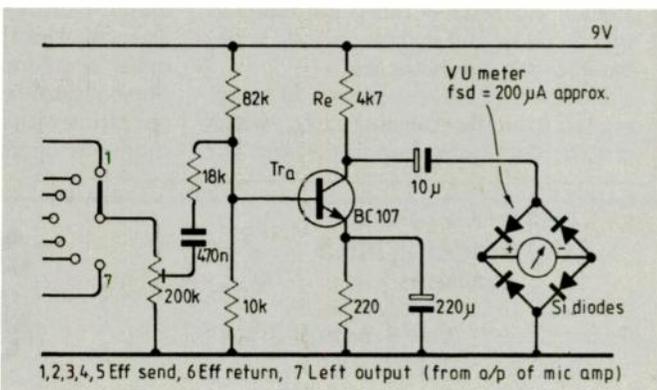


Fig. 8. VU meter circuit fed from current source to reduce effect of diode offset voltage.



and high frequency bands.

The amplifier shown in Fig. 4 has a calculated input noise density of 3nV/√Hz and a calculated input noise current density of 0.3pA/√Hz, ignoring flicker noise.

The frequency and phase response remain much the same regardless of gain setting. This seems to go against the intractable laws of gain-bandwidth product: as we increase the gain we must expect the frequency response to decrease, and vice-versa. In fact, the ring-of-three circuit is an early form of the current-mode-feedback amplifier currently very popular in video applications.

The explanation for this lies in the variable gain-setting resistor R_a . This not only determines the closed-loop gain by controlling the proportion of the output voltage fed back to the inverting port, but also forms the dominating part of the emitter load of the first transistor and consequently the gain of the first stage. As the value of R_a decreases, so the feedback diminishes and the closed-

loop gain rises. At the same time the open-loop gain of the circuit rises because Tr_1 's emitter load falls in value. The current consumption for all four mic pre-amps is 3mA.

Mix amplifiers

The mix amplifiers shown in Fig. 6 are based on a conventional transistor pair circuit. The only difficult decision in this area is the choice of the value for R_b . It is this value, combined with the input resistors, that determines the total contribution each input may make to the final output. In the end, I opted for a value that allowed for unity gain so that an input registering 0VU on the input meter will register 0VU on the output meter with the channel fader fully open. For there to be unity gain through the system there must be some gain because of the lossy fader and pan arrangement shown in Fig. 7.

Other circuitry

Good, clear metering is a real advantage. But led bar-graph displays and the

DESIGN

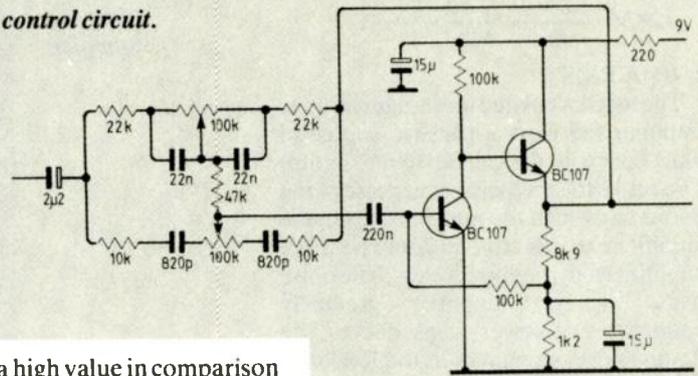
associated ICs to drive them, while otherwise excellent, would take more current from the battery supply than the mixer electronics itself, so the obvious choice is a moving-coil meter. The written specification for VU meter is as follows:

The volume indicator is a standardised instrument developed for the control and monitoring of sound programme. It shall be a root-mean-square type of instrument with a full-wave type of rectifier. The rectifier law shall be intermediate between linear and square law, having of an exponential of 1.2 ± 0.2 . The sensitivity shall not depart from that at 1kHz by more than 0.2dB between 35Hz and 10kHz at an input level of 0VU, nor more than 0.5dB between 25Hz and 16kHz.

The frequency response of the circuit and meter arrangement shown in Fig. 8, relative to a signal of 0VU at 1kHz, is: -3dB points, 15Hz and 60kHz. The response is less than 0.5dB down at 50Hz and 20kHz — this is not really to specification, but it shows a good wide-band response nonetheless.

Notice that the rectifiers and meter are fed from the collector of T_{ra} , which is a current source in parallel with R_c .

Fig. 9. Tone control circuit.



Because R_c is a high value in comparison with the emitter load of T_{ra} the voltage gain is very small during the part of the input cycle when the rectifier diodes are not in conduction. This alleviates most of the problem of the Si diode offset voltage.

On a practical level remember that a VU meter is an indicator of the average power of a waveform; it is not a peak-reading instrument like a BBC PPM meter. Failure to appreciate this (and on a practical level this means allowing the meter needle to swing into the red section on transients) will mean the mixer is operating with inadequate system headroom. In operation the meter needles

should only very occasionally swing above the 0VU reference level on complex programme. Provided input trim levels and main faders are set with this rule in mind, clipping will very rarely be encountered, even under conditions of moderate abuse!

The tone control circuit, Fig. 9, is based on the nearly universal Baxandall tone-control circuit, with the gain element a buffered common-emitter amplifier. The circuit values are modified to give the more gentle response at extremes of boost and cut. ■

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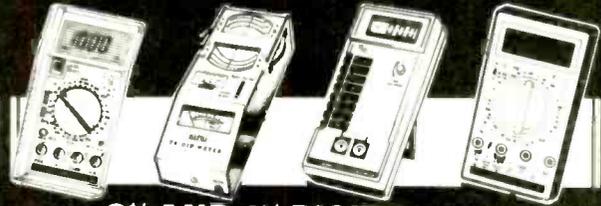


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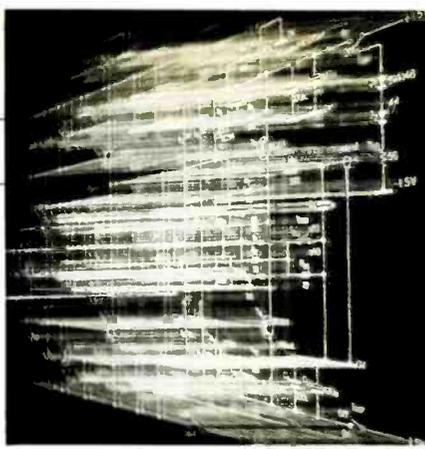
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CIRCLE ENQUIRY NO. 136 ON BACK PAGE



High-accuracy bridge amplifier

Derek Shell describes a differential bridge amplifier which needs only one precision resistor and has no common-mode error

If a Wheatstone bridge is arranged to drive a non-inverting feedback amplifier, with the measurement and reference nodes taken to the op-amp positive and negative op-amp inputs respectively, feedback action will force the bridge into balance. The current injected into the reference node via the feedback resistance must therefore modify the voltage drop across the reference resistance by an amount equal to the bridge current multiplied by the deviation between measurement (transducer) resistance and reference resistance. The voltage drop across the feedback resistance therefore provides an amplified measurement of resistance deviation.

By holding the bridge reference node at virtual ground by the additional op-amp circuit shown in Fig. 1, together with the measurement amplifier circuit, the voltage across the feedback resistance appears at the measurement amplifier output referenced to ground. We now have a two op-amp bridge differential amplifier which requires only one precision resistor, has no common-mode error and suffers from only one op-amp offset error which adds directly to signal input.

Another very useful consequence of the virtual grounding of the bridge reference node and, since the bridge is forced into balance, also of the measurement node, is that the voltages across both upper bridge resistors are held constant at bridge voltage, hence constant

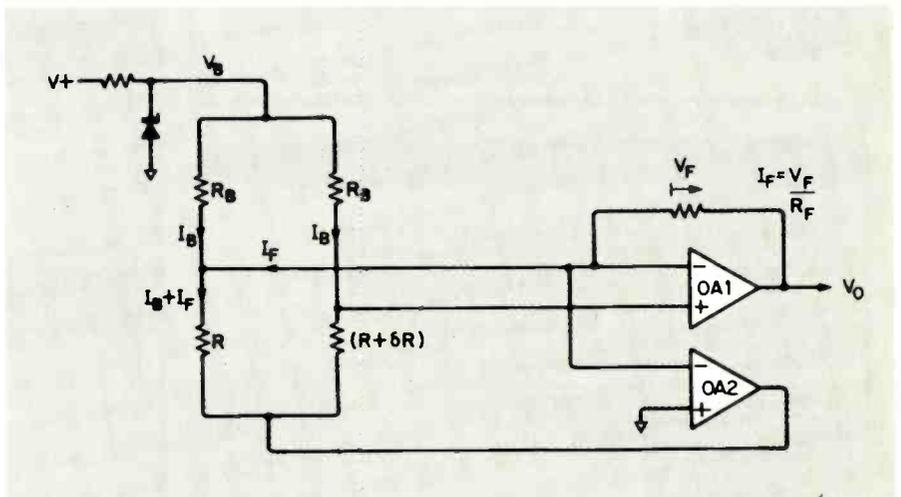


Fig. 1. Non-inverting feedback amplifier holds bridge reference at virtual earth, allowing use of differential bridge amplifier.

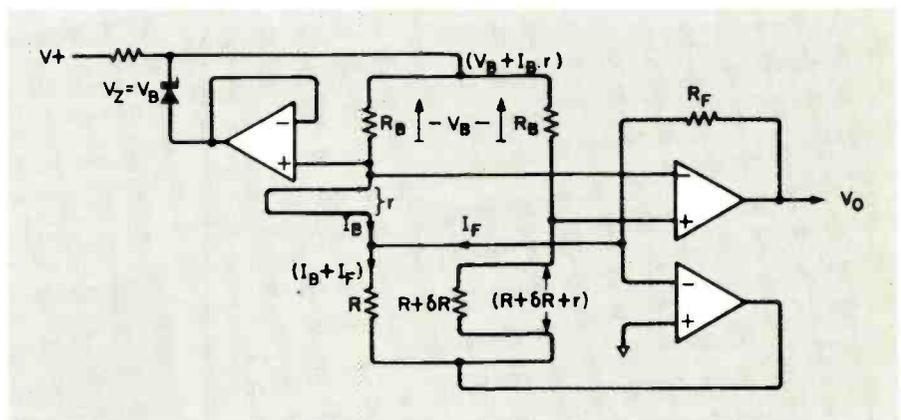


Fig. 2. Dummy lead drop added to bridge voltage holds bridge currents constant.

bridge currents are generated without recourse to constant current generators. The bridge voltage will, of course, require accurate definition, and can be generated by a reference zener and resistance to the positive supply. The derivation of errors from op-amp 1 input bias currents and offset, referring to Fig. 3, is as follows.

$$\left(\frac{V_o - I_{b2} - I_{b1}}{R_F}\right)R + v_{off} = (I_B - \frac{v_{off} - I_{b1}}{R_F})(R + \delta R)$$

$$V_o R/R_F = (I_B - \frac{v_{off} - I_{b1}}{R_F})\delta R +$$

$$(I_{b2} - I_{b1})R - \frac{v_{off}R - v_{off}}{R_B}$$

$$V_o = \delta R \cdot I_B \cdot R_F/R(1 - \frac{v_{off} - I_{b1}}{V_B}) +$$

$$I_{off}R_F - v_{off}R_F/R_B - R_F \cdot v_{off}/R$$

Scale factor error terms are

$$\frac{I_{off}R_F}{v_{off}R_F/R_B} \tag{1}$$

Output offset error terms are

$$\frac{I_{off}R_F}{v_{off}R_F/R_B} \tag{3}$$

$$\frac{v_{off}R_F/R_B}{v_{off}R_F/R} \tag{4}$$

$$\frac{v_{off}R_F/R}{v_{off}R_F/R} \tag{5}$$

Deriving errors from op-amp2 input bias and offset, as in Fig. 4, gives:

$$\left[\frac{V_o - v_{off} + I_B - v_{off} - I_{b2}}{R_F}\right]R =$$

$$(I_B - \delta I_B)(R + \delta R)$$

$$(V_o - v_{off})R/R_F = \delta R(I_B - \frac{v_{off}}{R_B}) + I_{b2}R$$

$$V_o = v_{off} + \delta R \cdot I_B \cdot R_F/R \cdot (1 - \frac{v_{off}}{R_B}) + I_{b2}R_F$$

Scale factor error term is

$$v_{off}/V_B \tag{6}$$

Output offset error terms are

$$v_{off} \tag{7}$$

$$I_{b2}R_F \tag{8}$$

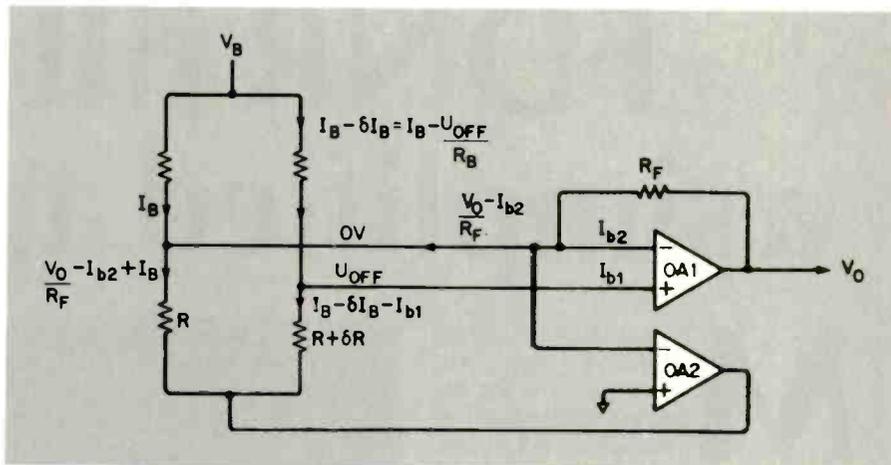


Fig.3. Derivation of errors from op-amp 1 input bias currents and offset

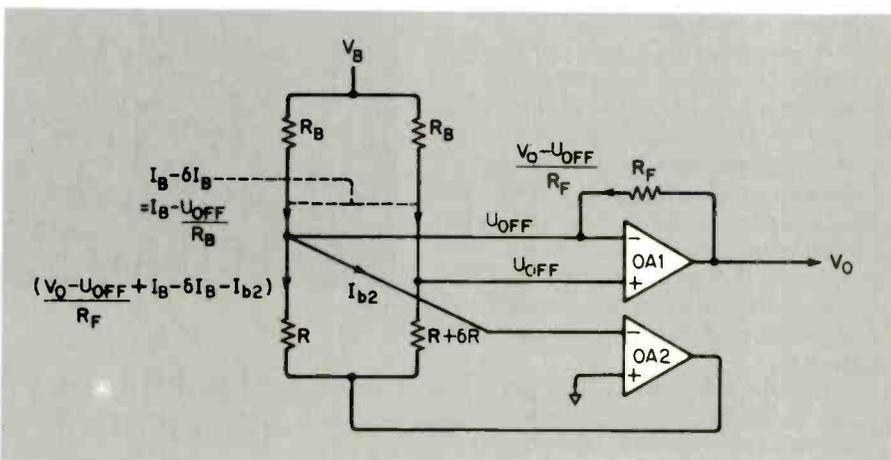


Fig.4. Bias currents and offset errors contributed by op-amp2

In the case where lead resistance compensation for remotely sited transducers must be applied, involving appreciable variation of lead and dummy lead resistance with temperature, the additional op-amp circuit shown in Fig. 2 will maintain constant bridge currents.

Figures 3 and 4 provide detailed analysis of errors introduced by op-amp offset and bias parameters.

Error derivation

Error term (5) expresses an error equivalent to measurement amplifier input offset added to signal and is intrinsic to any amplifier configuration. Excluding this term from consideration, therefore, the accuracy requirement from this amplifier configuration is that the sum of all the other error terms should be substantially less than errors deriving from tolerances in the bridge resistors, feedback resistor and bridge voltage.

With relatively non-stringent values of 0.1mV, 5nA and 10nA for offset voltage (nulled), offset current and bias current for the op-amps, total scale factor and output offset errors of 0.01% and 1mV respectively should be achievable over a wide range of transducer drive and signal amplification requirements, given suitable optimisation of resistance values and bridge voltage.

This bridge amplifier configuration has the merits of simplicity, low component count and high accuracy. Compared with the conventional three op-amp differential amplifier used to amplify a differential bridge signal, it offers a reduced precision resistor count from 3 to 1 with a corresponding reduction in scale-factor error, a reduction in op-amp count, and a halving of op-amp input offset-voltage errors which add directly to bridge signal. ■

FDNR filters — a simplified approach

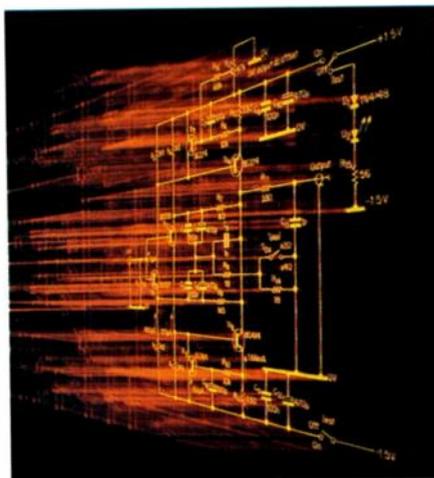
Most analogue circuits are simple in their basic operation, but the authors of textbooks seem to delight in the complicated exposition, sprinkling so much superfluous mathematics around as to obscure the wood with trees.

A typical textbook approach to circuit analysis takes into account all the characteristics of an active device, which results in algebraic equations too long to fit on one line and completely obscures the important principles at hand. How much better, surely, to start with a simplified model to demonstrate the principle and then to indicate any modifications and allowances that must be made to accommodate the vagaries of real devices.

I applied this principle a few years ago, on coming across an article describing an application of FDNR filters in the *Hewlett Packard Journal*. I had heard of frequency-dependent negative resistances before — their use was proposed as long ago as 1969¹ — but apart from noting that the circuits looked fearfully involved, I had not had occasion to take any further interest. This time, however, I resolved to get to grips with them, and knowing that a textbook was only likely to confuse me, set about analysing the circuit from scratch.

FDNR filters are not just a curiosity; they offer a realistic approach to some filtering applications. For instance, the Hewlett Packard 5420A digital signal analyser uses them as input anti-aliasing filters to provide identical, matched, low-pass filters on each of its two input channels.

The technique permits the realisation of filters with the flat pass-band, closely controlled phase response and stable drift-free characteristics required for the two identical input channels of a precision measuring instrument, cha-



Ian Hickman's explanation of frequency-dependent negative-resistance filters is the result of his unwillingness to rely on the maths-spattered textbook method of learning

acteristics which it would be difficult to obtain in the 0-30kHz range with filters using wound inductors.

Of course, it would be possible, if that instrument were being designed today, to consider the convenient and economical switched-capacitor filter IC's available from a number of manufacturers. But, being time-discrete filters, they would still need to be preceded by a

conventional anti-aliasing filter of some kind. And even then, one would be wary of incorporating a switched-capacitor filter, with its high-level clock signal and the possibility of spurious responses, in the front end of a sensitive measurement system.

The FDNR

To see how FDNRs may be used in filters, consider first how inductors and capacitors are used in conjunction with resistive terminations to make a frequency selective network, for example a low-pass filter. **Figure 1(a)** illustrates that the voltage across an inductor is proportional to the rate of change of current through it, that voltage across a resistor is simply proportional to the current, and that voltage across a capacitor is proportional to the integral of the current (or the current is proportional to the rate of change of voltage, which comes to the same thing). It also shows a device called an FDNR or sometimes a D element, with the characteristics shown.

For a constant-amplitude AC input current, the voltage across an inductor rises at 6dB/octave with increasing frequency; across a resistor it is constant in amplitude; and across a capacitor it falls at 6dB/octave. Across an FDNR it falls at 12dB/octave, for which reason it is sometimes called a "supercapacitor" and is denoted by four parallel lines, twice as many as a normal capacitor symbol. Similarly, where the $j\omega$ term associated with an inductor indicates a 90° leading voltage, for a capacitor the $1/j\omega$ or $-j/\omega$ term indicates a 90° lagging voltage and for an FDNR the $1/(j\omega)^2$ or $-1/\omega^2$ term indicates a negative resistance. (Remember that $j^2 = -1$, i.e. two 90° phase shifts give a phase shift of 180° — a sinewave in antiphase or a "negative" sinewave.

For the steady-state condition, which

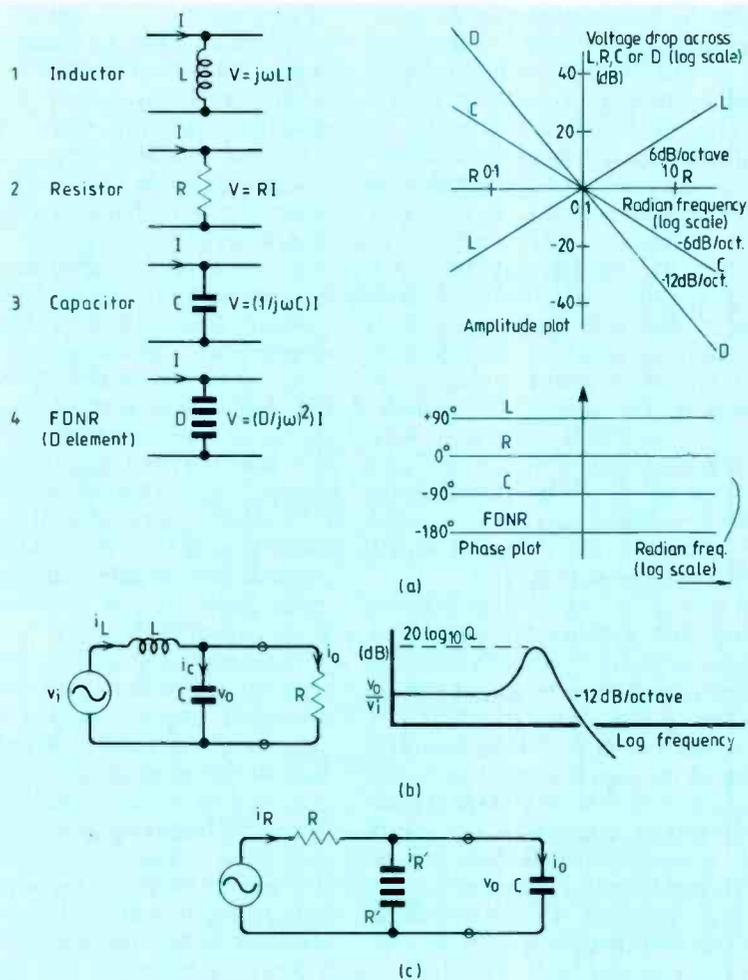


Fig. 1 Frequency-dependent behaviour of various components. Voltage across FDNR falls at twice the rate of that across a capacitor — hence the name “supercapacitor”. At (b) is a passive, second-order, low-pass LC section and its response. The FDNR equivalent is at (c)

For a practical filter, we would choose a value for Q to give the flattest possible pass-band response, rather than the peaky response illustrated.

Figure 1(c) shows the same two-element section with a resistor substituted for the inductor, and an FDNR for the capacitor. To be consistent, a capacitor must be substituted in (c) for the resistive termination in (b). The algebra for (c) is followed through in exactly the same way as in (b) and produces

$$\frac{v_o}{v_i} = \frac{1}{s^2 + sC + 1}$$

This is the same second-order expression for v_o/v_i as in (b), but with the damping term D equal to C, so $Q = 1/C$ for the normalised case where the resistance R and the FDNR R' both equal 1Ω. So in principle, given the appropriate FDNRs, we could realise any LC filter as an RR' filter, where R' indicates an FDNR.

Figure 2 shows an FDNR circuit, in which $C_1 = C_2$ and $R_1 = R_2 = R_3$. It can reasonably be described as a somewhat

is all we are considering here, the complex frequency variable s can be considered as shorthand for $j\omega$, where ω is the frequency of a sine wave in radians per second; so $\omega = 2\pi f$, where f is the frequency in Hertz. My apologies if this recap is a bit breathless, but the main drift of the article is how FDNRs work.

The algebra is fairly simple; the response of the second-order low-pass filter at (b) is as follows.

$$v_o = i_o R = i_c \frac{1}{j\omega C}$$

$$v_i = i_L j\omega L + v_o$$

$$i_L = i_c + i_o = v_o j\omega C + v_o/R = v_o(j\omega C + 1/R)$$

$$\frac{v_o}{v_i} = \frac{v_o}{i_L j\omega L + v_o}$$

$$= \frac{v_o}{j\omega L [(1/R) + j\omega C] v_o + v_o}$$

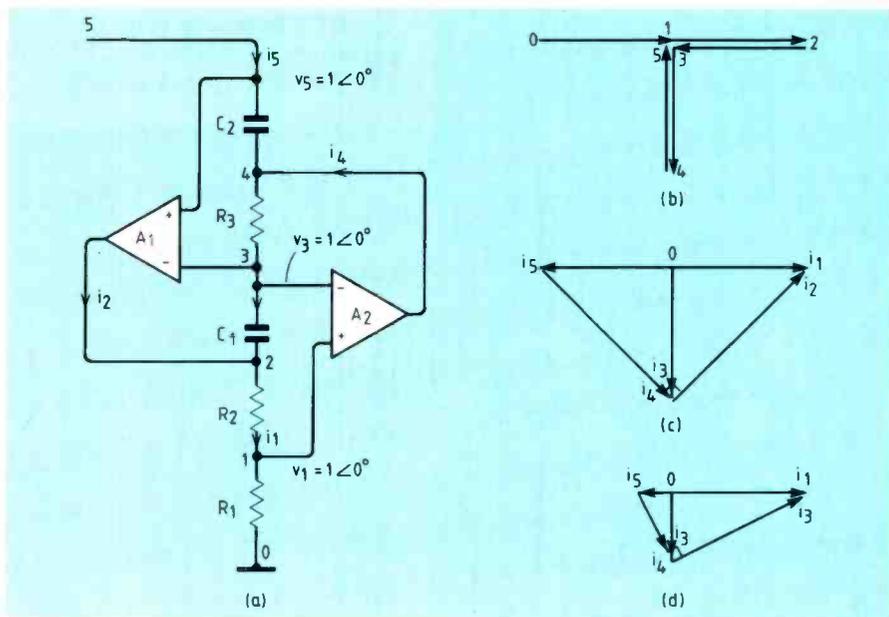
$$= \frac{1}{(j\omega L/R) + (j\omega)^2 LC + 1}$$

and if $L=C=1$,

$$\frac{v_o}{v_i} = \frac{1}{s^2 + s(1/R) + 1} \text{ and } D = \frac{1}{Q} = \frac{1}{R}$$

where D stands for damping, not for anything to do with D elements. The LC section is driven from zero source impedance, feeding a load of resistance R. It turns out that, for the normalised case of $L=1H$, $C=1F$, for which the corner frequency $\omega_c = 1$ radian/s, $D=1/Q=1/R$.

Fig. 2. FDNR circuit (a). $v_1 = v_3 = v_5$ and $i_1 = i_2 + i_3$, $i_3 = i_4 + i_5$. At (b) is voltage vector diagram for (a) when Rs and Cs are equal and $f = 1/2\pi CR$. Current vectors are at (c) and at (d) is current vector diagram for $f = 1/4\pi CR$. Currents i_2 and i_4 are always in quadrature



DESIGN

incestuous circuit — perhaps even reminiscent of the infamous disappearing bird. However, it is not nearly as fiercely complicated as might appear at first sight, being easily unravelled with the aid of vector diagrams as in Figs. 2(b) and (c).

For a start, consider the “extremum” of 0Hz: we can say that, given a DC path from node 5 to ground, the circuit is DC stable. At zero frequency, C_1 and C_2 are effectively open-circuit, so there are simply two op-amps each with its non-inverting input referred to ground and with 100% negative feedback, although in the case of A_1 , the feedback loop

Fig. 3. Evolution of a FDNR filter. Original section is at (a) and (b) shows the FDNR approach. A normalised low-pass filter for the H-P 5420 is seen at (c) and (d) is the active FDNR filter derived from that in (c)

includes A_2 . Now consider the AC conditions and, in particular, the frequency ω at which $1/(\omega C) = R$. For convenience, normalise everything, so that $R = 1\Omega$, $C = 1F$ and $\omega = 1rad/s$.

We need to determine the relationship between the voltage applied to node 5 and the current i_5 flowing as a result, as a function of frequency.

As a start, assume that there is a voltage $v_{1,0}$ of 1V RMS and frequency 1rad/s at node 1, as shown in Fig. 2(a), and define its phase as the reference phase for all the other voltages and currents in the circuit. The double-subscript convention being used here indicates the node at which the voltage is measured and the node with respect to which it is being measured, in that order: thus $v_{1,0}$ is the voltage at node 1 with respect to node 0, ground.

Because of its large voltage gain, an op-amp with negative feedback must have its two inputs at virtually the same voltage — assuming that it is not overdriven, in which case the output voltage would be at one of the supply-rail voltages. We therefore know that the voltage at node 3 with respect to ground is $1/0^\circ$ and at node 5 it is also $1/0^\circ$ for the same reason. Furthermore, node 1 is connected only to an op-amp input, which draws negligible current: therefore, since $R_1 = R_2$ and i_1 flows through both, $v_{2,1} = 1/0^\circ$.

We can now fill in $v_{1,0}$, $v_{2,1}$ and $v_{3,2}$ on the voltage vector diagram of Fig. 2(b) and i_1 on the current vector diagram (c) as shown. Given that $v_{3,2}$, the voltage across C_3 , is lagging by 180° , the current i_3 through C_1 must be lagging by only 90° as shown in (c): since $i_1 = i_3 + i_2$, we can now also mark i_2 , the output current of A_1 . The current i_3 flows through both R_3 and C_1 , so we can mark $v_{3,3}$ in (a) and also $v_{5,4}$ — which must be just the opposite, since $v_{5,0} = v_{3,0}$. Since $v_{5,4}$, the

voltage across C_2 , is lagging by 270° (which is the same as leading by 90°), then i_5 , the current through C_2 , must be lagging by 180° , as shown in (c). We can now finally mark i_4 , since $i_4 + i_5 = i_3$.

Current i_5 is 180° out of phase with voltage $v_{5,0}$ at node 5, i.e. it is flowing the opposite way to what we would expect if we saw a resistance looking in at node 5. What we have is a resistance equal to $v_{5,0}/i_5$, or $(1/0^\circ V)/(1/180^\circ A)$, which is 1Ω .

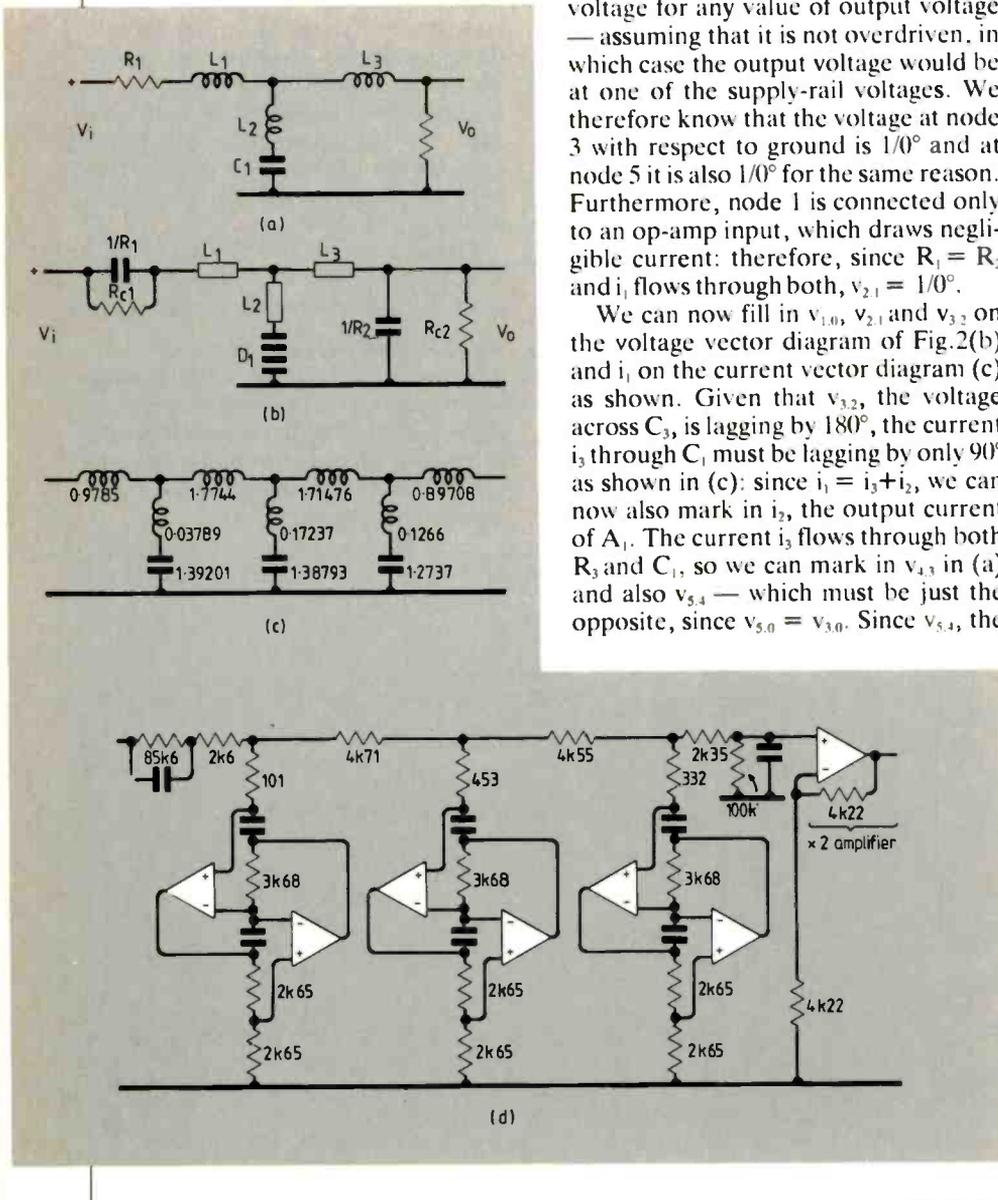
Figure 2(d) shows the current vector diagram for a frequency of $0.5rad/s$. Since, at this frequency, the reactance of C_1 is 2Ω , i_3 is now only $0.5A$ and i_2 is as shown. In consequence, $v_{3,3}$ is only $0.5V$ and, since the reactance of C_2 is also 2Ω , a mere $0.25A$ is enough to provide the $0.5V$ drop $v_{5,4}$ across C_2 . At one octave lower than before, the resistance looking into node 5 is still negative but is now -4Ω . Similarly, at a frequency of $2rad/s$, we see 0.25Ω .

The negative resistance is inversely proportional to the square of the frequency or, to put it another way, if an alternating current of constant amplitude is injected into node 5, the resulting voltage drop falls at $-12dB/octave$ with increasing frequency, as well as being in anti-phase. This is a frequency-dependent negative resistor, subject only to the limitation that one end is anchored firmly to ground.

Since op-amps are not capable of supplying amps of current, except on paper, it is more sensible to choose values for R_1 , R_2 and R_3 in the range $1k\Omega$ to $5k\Omega$. They need not be equal, but the analysis is a little more complicated if they are not. The FDNR will work as described down to a frequency where the op-amp input currents are no longer negligible compared to i_5 ; and up to the frequency at which the op-amp output impedances are no longer low considering the size of i_5 , or where their frequency response begins to fall off, whichever is the lower. At the frequency where $R = 1/(\omega C)$, the input impedance at node 5 is simply $-R$.

FDNR filter design

In many applications, the fact that the FDNR is not floating is no problem. Figure 3(a) shows a three-pole, elliptic, low-pass filter operating between source and load impedances R_1 and R_2 , which would normally be equal. In (b), the transformed filter using resistors instead of inductors and FDNRs in place of the capacitors operates between capacitive source and load impedances; the resistors R_{c1} and R_{c2} are added purely to define the behaviour at 0Hz, where the reactance of the terminating ca-



capacitors has risen to infinity.

In (c) is a normalised seven-pole, elliptic low-pass filter used as an anti-aliasing filter in the H-P 5420A digital signal analyser; while (d) shows the corresponding FDNR filter scaled for a 30kHz cut-off. Resistor R_{c1} is set to 85.6k Ω which, added to the series resistors in lieu of inductors, totals 100k Ω . In conjunction with R_{c2} , a 100k Ω DC return at the non-inverting input of the output op-amp, the DC attenuation is 6dB.

At higher frequencies in the pass-band, the same 6dB attenuation occurs, due now to the reactance of the two equal-value source and terminating capacitors. An attenuation of -6dB in the pass-band sounds a bit odd, but this is because we have been considering the input as V_1 , the source EMF. If, as is usually the case, the source resistance R_1 in Fig.3(a) is considered as the internal or matched resistance of the source and the filter input measured at the input to L_1 , instead, the pass-band attenuation is zero, as expected. The final circuit of Fig.3(d) provides greater than 80dB of stop-band attenuation for frequencies above 80kHz.

Returning for the moment to the prototype normalised seven-pole filter of Fig.3(c), one can see that FDNRs are particularly convenient for low-pass, rather than high-pass filters and also how the choice of T-sections rather than π -sections reduces the number of FDNRs by one.

Design procedure

Practical FDNR filter design proceeds much as normal, starting with a design such as that of Fig.3(c), where the values are normalised; that is, are in Henries and Farads, calculated for a cut-off frequency of 0.159Hz ($1/2\pi$ Hz or 1 rad/s) with a 1 Ω characteristic resistance source and load. The inductor values in Henries become the resistor values in Ohms and the reciprocal of the capacitor values becomes the value of the supercapacitors, D elements or FDNRs, in minus Ohms. The reciprocal of the characteristic resistance becomes the terminating capacitor value in Farads.

The design is then normalised to the desired cut-off frequency by dividing the R and multiplying the D values by the new cut-off frequency — equivalent

to dividing both L and C values by the desired cut-off frequency when denormalising an LC filter. A suitable multiplying factor is then chosen to bring the R values to a manageable level — say 2.5k Ω for R_1 , corresponding to L_1 in Fig. 3(a) — and the same multiplying factor applied to the D elements.

The value of negative resistance required at the cut-off frequency is now determined, and the appropriate value for the two capacitors of each D element can be selected, given the values of R_1 , R_2 and R_3 used, see Fig. 2(a). ■

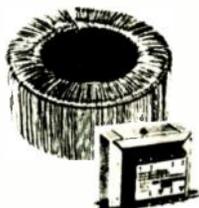
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625	30.06	27.05	19.84	18.79	16.98	15.78	15.03
750	36.42	34.58	25.38	24.01	21.71	20.17	19.21
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MODEMS FOR TELEMETRY

Standards for PSTN-based modem communications are well established. From 75b/s to 14 000b/s, some thirteen standards cover a variety of bit rates, interface specifications and modulation techniques. With modems, increased speed not only increases the cost, but also the likelihood that the received data will contain errors. For this reason, most high-speed communication systems possess some form of software-controlled error correction, which further increases cost.

Many low-cost, low data-rate telemetry systems have used DTMF tones for signalling purposes, allowing a limited amount of data to be sent at a very low speed. Later, single-chip devices for the CCITT V21 (300b/s) and V23 (1200/75 or 1200/1200b/s) standards became available.

V23 evolved from terminal applications where the user was assumed to be able to type at up to 75b/s but could receive from a host computer at 1200b/s. The 1200/1200 specification allowed a higher data rate from the terminal, but the transmission was half duplex, so that only one modem could transmit while the second had to wait until the message was complete.

More recently, V22 allows 1200b/s in either direction at the same time (full duplex), reducing line occupancy and therefore costs. So, how is this improvement achieved?

Telephone-line telemetry for the public utilities is growing fast. Gordon Lindsay presents two circuits for modems using Sierra's modem and controller chips

V23 calls for the use of frequency shift keying (FSK), which is the transmission of two different tones to represent 0 and 1, the receiving modem filtering the signal to extract the data. On the other hand, V22 is a differential phase shift keying (DPSK) system in which the carrier is modulated through four phases, each representing two bits or one dibit.

Most communications textbooks show a proof that a standard telephone line has an optimum data rate of 600 symbols/s due to the multiplexing of PSTN lines, so V22 will operate at up to 600dibits or 1200b/s, since each symbol is two bits.

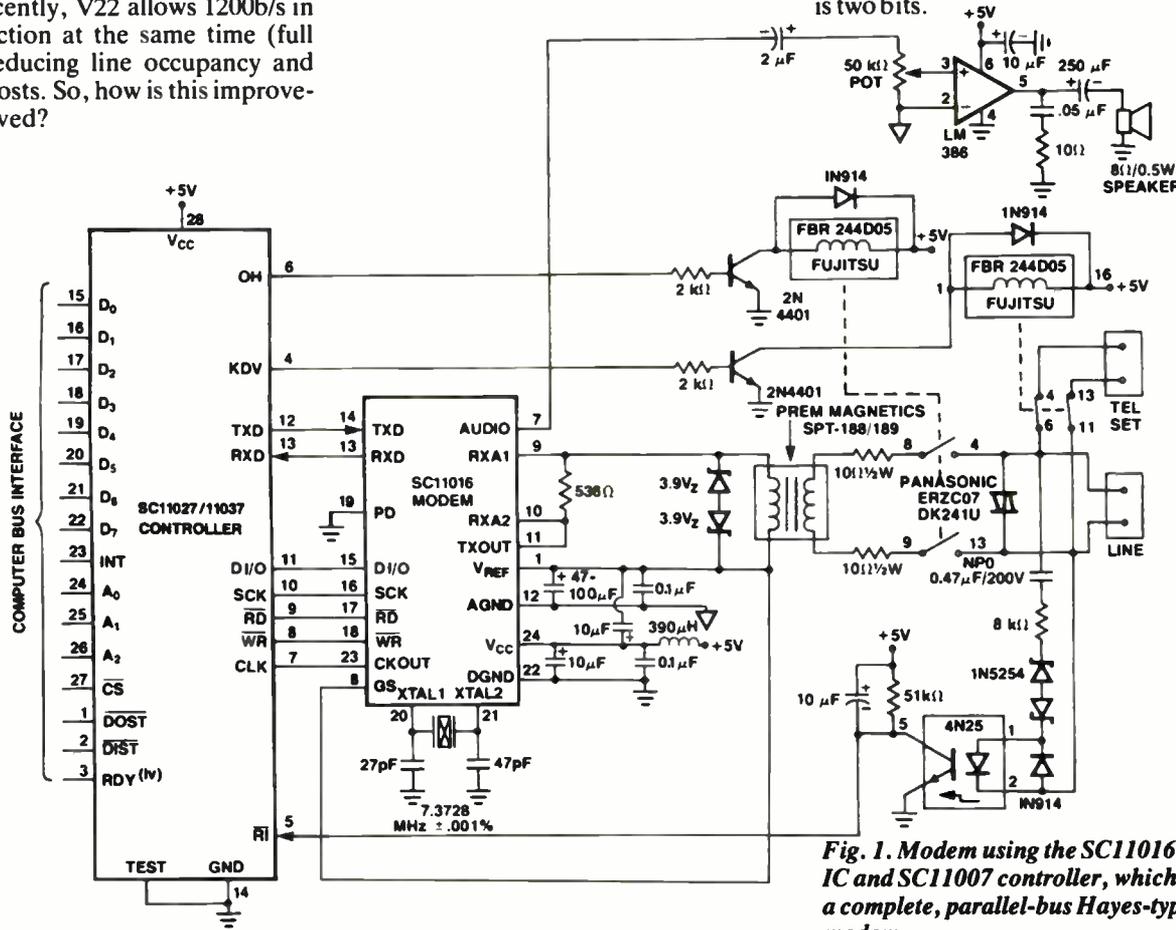


Fig. 1. Modem using the SC11016 modem IC and SC11007 controller, which makes a complete, parallel-bus Hayes-type smart modem

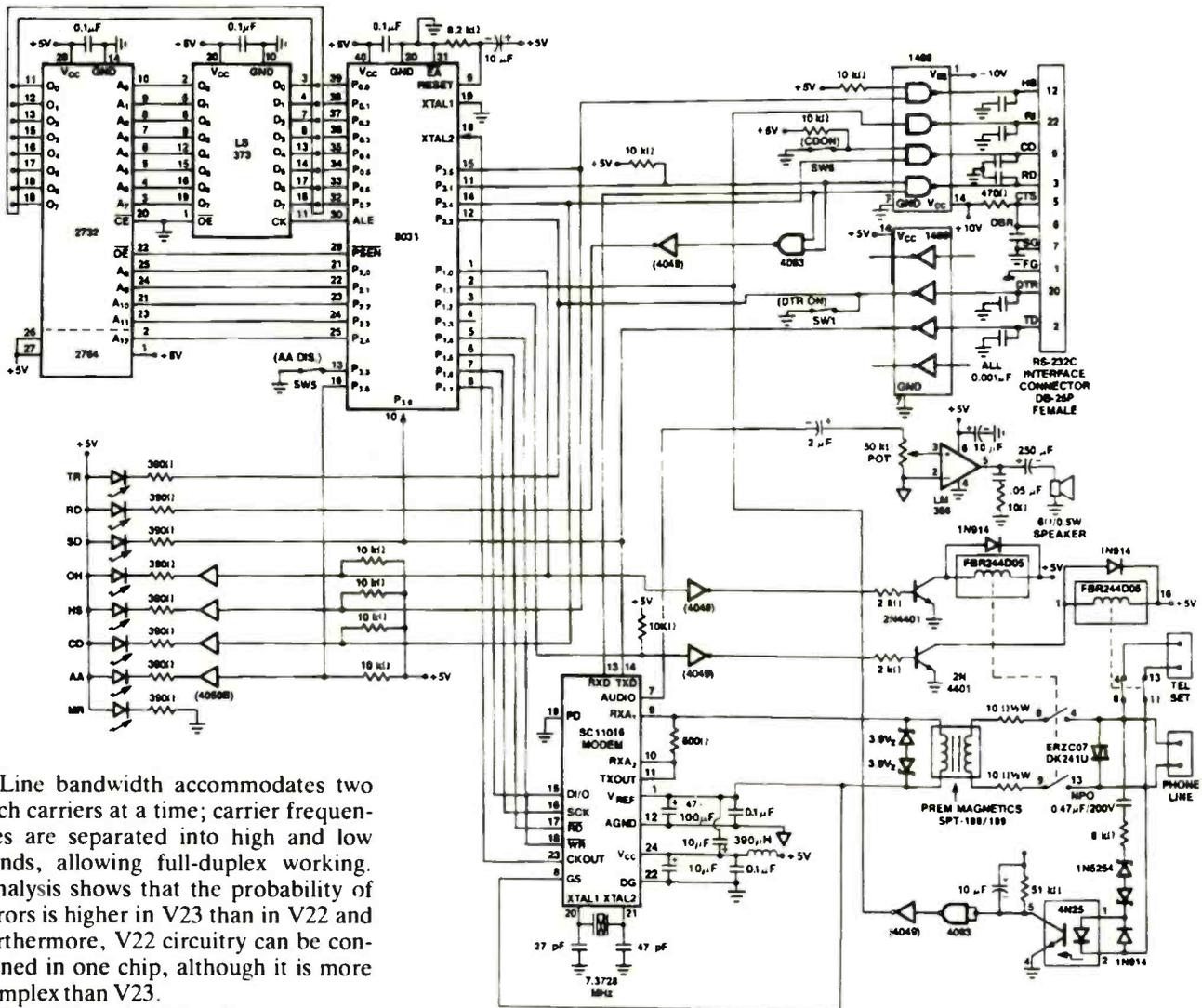


Fig. 2. Stand-alone modem which uses an 8031 (an 8051 without rom) for control to allow connection to any type of host.

Line bandwidth accommodates two such carriers at a time; carrier frequencies are separated into high and low bands, allowing full-duplex working. Analysis shows that the probability of errors is higher in V23 than in V22 and furthermore, V22 circuitry can be contained in one chip, although it is more complex than V23.

The high probability of correct reception using V22 means that such systems do not need error-correction software, except where more than 64Kbyte of data is being sent or in the presence of extremely bad noise; such a large amount of data is unusual in remote telemetry. In many cases, error-correcting modems need a call setup time longer than required for transmission and the use of an error-correcting protocol such as the CCITT V42 recommendation would increase on-line time far too much for it to be an advantage in some systems.

Modem design

Complete modems usually consist of three discrete sections: host interface, modem circuit and line interface. Typical host interfaces are RS232 connections or parallel buses for microprocessors. The line interface is subject to local PTT requirements and must also meet safety specifications. Sierra Semiconductor's modems are configurable in both respects.

In remote telemetry application, Sier-

ra's SC11016 conforms to V21 and V22 and contains all analogue and digital circuitry needed, including an on-chip DTMF dialler. It is controlled by sending commands to the IC through a simple interface to set the mode (transmit, receive or test), data rate, auto answer, transmit and receive levels and autodialling functions. A second chip provides the host interface in either RS232 or microprocessor connection.

Source code, which uses the Hayes AT command set, is available for all the modem control functions to run on an 8031 microcontroller, so that the modem is usable with any type of host equipment.

In initiating a call, the SC11016 generates DTMF tones for dialling and is then set to monitor line status. If the call is answered, it performs the specified handshaking functions and is set to trans-

mit. Line monitoring continues during transmission and is followed by the specified termination sequencer; if the connection fails in mid-transmission, the modem reports the reason for failure. It automatically answers and receives incoming data. The SC11016 operates on a minimum carrier of -43dBm and with s:n ratios of <20dB.

Figures 1 and 2 show designs for two modems: one using a parallel controller and the other an 8031 with the SC11016. The controllers contain all the software in internal rom to allow the use of fewer chips than in the micro-based solution. Modems and controllers are available in surface-mounting packages and a complete system need occupy only 75x75mm.

The SC11016 is a c-mos device, needing a single 5V supply. Software-controlled power-down sets it to stand-by when it is inactive, in which condition it draws 3.5mA. Power-up takes 5ms. ■

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A2599 37.50	EAC91 2.50	EL34 0.95	M8100 5.50	Q5150/45 7.00	YJ1020 42.50	4G57 2.25	6C61 1.50	8CWS 1.50	30P16 1.75	5674 7.50
A2792 27.50	EAF42 1.20	EL34 0.95	M8101 5.50	Q5121/3 5.00	YJ1060 195.00	4G77 2.25	6C62 1.50	8CWS 1.50	30P17 0.60	5675 4.50
A2900 11.50	EB34 1.50	EL34 0.95	M8102 5.50	Q5121/3 5.00	YJ1060 195.00	4J6A 2.95	6C63 1.50	8CWS 1.50	30P18 1.75	5676 7.50
A3283 24.00	EB41 3.95	EL34 0.95	M8103 5.50	Q5121/3 5.00	YJ1071 195.00	4K16 1.50	6C64 1.50	8CWS 1.50	30P19 1.75	5677 2.50
A3343 35.95	EB91 0.85	EL34 0.95	M8104 5.50	Q5121/3 5.00	YJ1290 65.00	4T85P 150.00	6C65 1.50	8CWS 1.50	30P20 1.75	5678 4.50
ACSP3A 4.95	EBC33 7.50	EL34 0.95	M8105 5.50	Q5121/3 5.00	Z77 1.20	4X150A 35.00	6C66 1.50	8CWS 1.50	30P21 1.75	5679 4.50
AC/CS2PEN 8.50	EBC41 3.50	EL34 0.95	M8106 5.50	Q5121/3 5.00	Z300T 6.00	4X150D 55.00	6C67 1.50	8CWS 1.50	30P22 1.75	5680 4.50
ACT22 59.75	EBC81 1.50	EL34 0.95	M8107 5.50	Q5121/3 5.00	Z300C 12.00	4X500A 350.00	6C68 1.50	8CWS 1.50	30P23 1.75	5681 4.50
AH221 39.00	EBC91 1.95	EL34 0.95	M8108 5.50	Q5121/3 5.00	Z359 9.00	5A/102D 9.50	6C69 1.50	8CWS 1.50	30P24 1.75	5682 4.50
AH238 39.00	EBC91 1.95	EL34 0.95	M8109 5.50	Q5121/3 5.00	Z700U 9.50	5A152M 9.00	6C70 1.50	8CWS 1.50	30P25 1.75	5683 4.50
AL60 6.00	EBC91 1.95	EL34 0.95	M8110 5.50	Q5121/3 5.00	Z803X 18.95	5A163K 10.00	6C71 1.50	8CWS 1.50	30P26 1.75	5684 4.50
ANI 14.00	EBC91 1.95	EL34 0.95	M8111 5.50	Q5121/3 5.00	ZM1020 8.50	5A170K 6.25	6C72 1.50	8CWS 1.50	30P27 1.75	5685 4.50
APR12 2.50	EBC91 1.95	EL34 0.95	M8112 5.50	Q5121/3 5.00	ZM1021 8.00	5A180M 9.50	6C73 1.50	8CWS 1.50	30P28 1.75	5686 4.50
APR34 1.25	EBC91 1.95	EL34 0.95	M8113 5.50	Q5121/3 5.00	ZM1022 7.95	5B-110M 10.00	6C74 1.50	8CWS 1.50	30P29 1.75	5687 4.50
APR35 2.00	EBC91 1.95	EL34 0.95	M8114 5.50	Q5121/3 5.00	ZM1023 8.00	5B-254M 11.50	6C75 1.50	8CWS 1.50	30P30 1.75	5688 4.50
AZ11 4.50	EBC91 1.95	EL34 0.95	M8115 5.50	Q5121/3 5.00	ZM1024 10.00	5B-255M 14.50	6C76 1.50	8CWS 1.50	30P31 1.75	5689 4.50
BT16 35.00	EBC91 1.95	EL34 0.95	M8116 5.50	Q5121/3 5.00	ZM1025 9.50	5B-256M 15.00	6C77 1.50	8CWS 1.50	30P32 1.75	5690 4.50
BT58 55.00	EBC91 1.95	EL34 0.95	M8117 5.50	Q5121/3 5.00	ZM1026 9.50	5B-257M 15.00	6C78 1.50	8CWS 1.50	30P33 1.75	5691 4.50
BT17 25.00	EBC91 1.95	EL34 0.95	M8118 5.50	Q5121/3 5.00	ZM1027 8.00	5B258M 14.50	6C79 1.50	8CWS 1.50	30P34 1.75	5692 4.50
BT113 35.00	EBC91 1.95	EL34 0.95	M8119 5.50	Q5121/3 5.00	ZM1028 10.00	5C18A 2.50	6C80 1.50	8CWS 1.50	30P35 1.75	5693 4.50
CIK 27.50	EBC91 1.95	EL34 0.95	M8120 5.50	Q5121/3 5.00	ZM1029 9.50	5C18B 2.50	6C81 1.50	8CWS 1.50	30P36 1.75	5694 4.50
C3M 17.95	EBC91 1.95	EL34 0.95	M8121 5.50	Q5121/3 5.00	ZM1030 9.50	5C18C 2.50	6C82 1.50	8CWS 1.50	30P37 1.75	5695 4.50
CI149/1 120.00	EBC91 1.95	EL34 0.95	M8122 5.50	Q5121/3 5.00	ZM1031 8.00	5C18D 2.50	6C83 1.50	8CWS 1.50	30P38 1.75	5696 4.50
CI150/1 135.00	EBC91 1.95	EL34 0.95	M8123 5.50	Q5121/3 5.00	ZM1032 7.95	5C18E 2.50	6C84 1.50	8CWS 1.50	30P39 1.75	5697 4.50
CI166 125.00	EBC91 1.95	EL34 0.95	M8124 5.50	Q5121/3 5.00	ZM1033 8.00	5C18F 2.50	6C85 1.50	8CWS 1.50	30P40 1.75	5698 4.50
CI534 32.00	EBC91 1.95	EL34 0.95	M8125 5.50	Q5121/3 5.00	ZM1034 10.00	5C18G 2.50	6C86 1.50	8CWS 1.50	30P41 1.75	5699 4.50
CCA 3.50	EBC91 1.95	EL34 0.95	M8126 5.50	Q5121/3 5.00	ZM1035 9.50	5C18H 2.50	6C87 1.50	8CWS 1.50	30P42 1.75	5700 4.50
CD24 6.50	EBC91 1.95	EL34 0.95	M8127 5.50	Q5121/3 5.00	ZM1036 9.50	5C18I 2.50	6C88 1.50	8CWS 1.50	30P43 1.75	5701 4.50
CK1006 3.50	EBC91 1.95	EL34 0.95	M8128 5.50	Q5121/3 5.00	ZM1037 8.00	5C18J 2.50	6C89 1.50	8CWS 1.50	30P44 1.75	5702 4.50
CK5676 6.50	EBC91 1.95	EL34 0.95	M8129 5.50	Q5121/3 5.00	ZM1038 10.00	5C18K 2.50	6C90 1.50	8CWS 1.50	30P45 1.75	5703 4.50
CYNos PRICES ON REQUEST	EBC91 1.95	EL34 0.95	M8130 5.50	Q5121/3 5.00	ZM1039 9.50	5C18L 2.50	6C91 1.50	8CWS 1.50	30P46 1.75	5704 4.50
CCX1140 495.00	EBC91 1.95	EL34 0.95	M8131 5.50	Q5121/3 5.00	ZM1040 10.00	5C18M 2.50	6C92 1.50	8CWS 1.50	30P47 1.75	5705 4.50
CCX1528 3250.00	EBC91 1.95	EL34 0.95	M8132 5.50	Q5121/3 5.00	ZM1041 9.50	5C18N 2.50	6C93 1.50	8CWS 1.50	30P48 1.75	5706 4.50
D3A 27.50	EBC91 1.95	EL34 0.95	M8133 5.50	Q5121/3 5.00	ZM1042 8.00	5C18O 2.50	6C94 1.50	8CWS 1.50	30P49 1.75	5707 4.50
D63 1.20	EBC91 1.95	EL34 0.95	M8134 5.50	Q5121/3 5.00	ZM1043 7.95	5C18P 2.50	6C95 1.50	8CWS 1.50	30P50 1.75	5708 4.50
DA41 22.50	EBC91 1.95	EL34 0.95	M8135 5.50	Q5121/3 5.00	ZM1044 10.00	5C18Q 2.50	6C96 1.50	8CWS 1.50	30P51 1.75	5709 4.50
DA42 17.50	EBC91 1.95	EL34 0.95	M8136 5.50	Q5121/3 5.00	ZM1045 9.50	5C18R 2.50	6C97 1.50	8CWS 1.50	30P52 1.75	5710 4.50
DA90 4.50	EBC91 1.95	EL34 0.95	M8137 5.50	Q5121/3 5.00	ZM1046 10.00	5C18S 2.50	6C98 1.50	8CWS 1.50	30P53 1.75	5711 4.50
DAF91 0.95	EBC91 1.95	EL34 0.95	M8138 5.50	Q5121/3 5.00	ZM1047 9.50	5C18T 2.50	6C99 1.50	8CWS 1.50	30P54 1.75	5712 4.50
DAF96 1.95	EBC91 1.95	EL34 0.95	M8139 5.50	Q5121/3 5.00	ZM1048 10.00	5C18U 2.50	6C100 1.50	8CWS 1.50	30P55 1.75	5713 4.50
DC70 1.75	EBC91 1.95	EL34 0.95	M8140 5.50	Q5121/3 5.00	ZM1049 9.50	5C18V 2.50	6C101 1.50	8CWS 1.50	30P56 1.75	5714 4.50
DCX9 3.50	EBC91 1.95	EL34 0.95	M8141 5.50	Q5121/3 5.00	ZM1050 10.00	5C18W 2.50	6C102 1.50	8CWS 1.50	30P57 1.75	5715 4.50
DCX4-5000 25.00	EBC91 1.95	EL34 0.95	M8142 5.50	Q5121/3 5.00	ZM1051 9.50	5C18X 2.50	6C103 1.50	8CWS 1.50	30P58 1.75	5716 4.50
DET16 28.50	EBC91 1.95	EL34 0.95	M8143 5.50	Q5121/3 5.00	ZM1052 8.00	5C18Y 2.50	6C104 1.50	8CWS 1.50	30P59 1.75	5717 4.50
DET18 28.50	EBC91 1.95	EL34 0.95	M8144 5.50	Q5121/3 5.00	ZM1053 7.95	5C18Z 2.50	6C105 1.50	8CWS 1.50	30P60 1.75	5718 4.50
DE120 2.50	EBC91 1.95	EL34 0.95	M8145 5.50	Q5121/3 5.00	ZM1054 10.00	5C19A 2.50	6C106 1.50	8CWS 1.50	30P61 1.75	5719 4.50
DE122 29.50	EBC91 1.95	EL34 0.95	M8146 5.50	Q5121/3 5.00	ZM1055 9.50	5C19B 2.50	6C107 1.50	8CWS 1.50	30P62 1.75	5720 4.50
DE123 35.00	EBC91 1.95	EL34 0.95	M8147 5.50	Q5121/3 5.00	ZM1056 8.00	5C19C 2.50	6C108 1.50	8CWS 1.50	30P63 1.75	5721 4.50
DE124 27.50	EBC91 1.95	EL34 0.95	M8148 5.50	Q5121/3 5.00	ZM1057 7.95	5C19D 2.50	6C109 1.50	8CWS 1.50	30P64 1.75	5722 4.50
DE125 22.00	EBC91 1.95	EL34 0.95	M8149 5.50	Q5121/3 5.00	ZM1058 10.00	5C19E 2.50	6C110 1.50	8CWS 1.50	30P65 1.75	5723 4.50
DE129 32.00	EBC91 1.95	EL34 0.95	M8150 5.50	Q5121/3 5.00	ZM1059 9.50	5C19F 2.50	6C111 1.50	8CWS 1.50	30P66 1.75	5724 4.50
DF61 3.50	EBC91 1.95	EL34 0.95	M8151 5.50	Q5121/3 5.00	ZM1060 10.00	5C19G 2.50	6C112 1.50	8CWS 1.50	30P67 1.75	5725 4.50
DF91 1.50	EBC91 1.95	EL34 0.95	M8152 5.50	Q5121/3 5.00	ZM1061 9.50	5C19H 2.50	6C113 1.50	8CWS 1.50	30P68 1.75	5726 4.50
DF92 1.50	EBC91 1.95	EL34 0.95	M8153 5.50	Q5121/3 5.00	ZM1062 8.00	5C19I 2.50	6C114 1.50	8CWS 1.50	30P69 1.75	5727 4.50
DF96 1.25	EBC91 1.95	EL34 0.95	M8154 5.50	Q5121/3 5.00	ZM1063 7.95	5C19J 2.50	6C115 1.50	8CWS 1.50	30P70 1.75	5728 4.50
DF97 1.25	EBC91 1.95	EL34 0.95	M8155 5.50	Q5121/3 5.00	ZM1064 10.00	5C19K 2.50	6C116 1.50	8CWS 1.50	30P71 1.75	5729 4.50
DG10A 8.50	EBC91 1.95	EL34 0.95	M8156 5.50	Q5121/3 5.00	ZM1065 9.50	5C19L 2.50	6C117 1.50	8CWS 1.50	30P72 1.75	5730 4.50
DH63 3.50	EBC91 1.95	EL34 0.95	M8157 5.50	Q5121/3 5.00	ZM1066 8.00	5C19M 2.50	6C118 1.50	8CWS 1.50	30P73 1.75	5731 4.50
DH77 1.50	EBC91 1.95	EL34 0.95	M8158 5.50	Q5121/3 5.00	ZM1067 7.95	5C19N 2.50	6C119 1.50	8CWS 1.50	30P74 1.75	5732 4.50
DK91 1.20	EBC91 1.95	EL34 0.95	M8159 5.50	Q5121/3 5.00	ZM1068 10.00	5C19O 2.50	6C120 1.50	8CWS 1.50	30P75 1.75	5733 4.50
DK92 2.50	EBC91 1.95	EL34 0.95	M8160 5.50	Q5121/3 5.00	ZM1069 9.50	5C19P 2.50	6C121 1.50	8CWS 1.50	30P76	

INTERFACING WITH C

PART 4

Signal processing with C unavoidably requires mathematical spadework. Howard Hutchings presents the essentials for serious application development

When computers are used to control engineering systems, the best they can do is to take occasional snapshots or samples of events. Provided events are changing slowly and the computer is sampling rapidly, Shannon's sampling rule will be satisfied. It will be possible to recover the digitally processed signal without corruption attributable to aliasing.

To illustrate the potential hazards of periodic sampling, visualise a rotating radar scanner tracking the progress of an approaching aircraft. The air traffic controller patiently monitoring a flickering cathode-ray tube might observe a steady approach when the flight path could in fact be wavering, vertically and horizontally, about its ideal projected path. Because the radar aerial is rotating relatively slowly compared to the rate of change of the plane's flight path, much of this data is missed completely.

The stroboscopic effects of sampling add to the confusion: an event taking place at one frequency appears to occur at a completely different frequency.

How many snapshots?

It's not my brief to become involved with the sampling theorem, but one ought to be in a position to answer a few questions. The analogue signal presented to the a-to-d converter will be processed into a series of discrete samples: how many snapshots must be taken to characterise the signal completely? To avoid aliasing, it is necessary to sample at a rate which is at least twice the highest frequency present, the Nyquist frequency. The Fourier spectrum of the signal (Fig. 3.1), together with the sampling frequency, illustrates this point rather nicely.

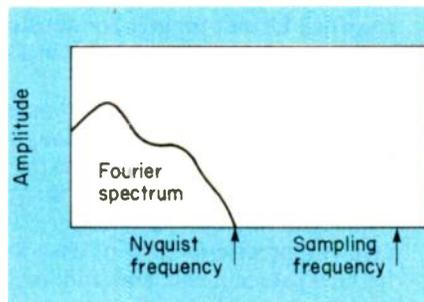


Fig. 3.1. Shannon's sampling theorem in action. To avoid signal corruption due to aliasing, the sampling frequency $f = 1/T$ should be at least twice the Nyquist frequency.

Some a-to-d converters latch the analogue input at the start of conversion and hold it long enough for the a-to-d to complete the conversion cycle. Such a device is called a sample-and-hold. If the input is not latched, then the analogue signal must not change by more than 0.5 LSB during conversion. If it does the digital output will be fraudulent.

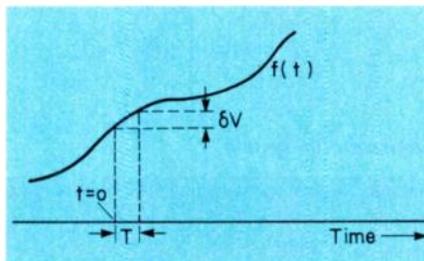
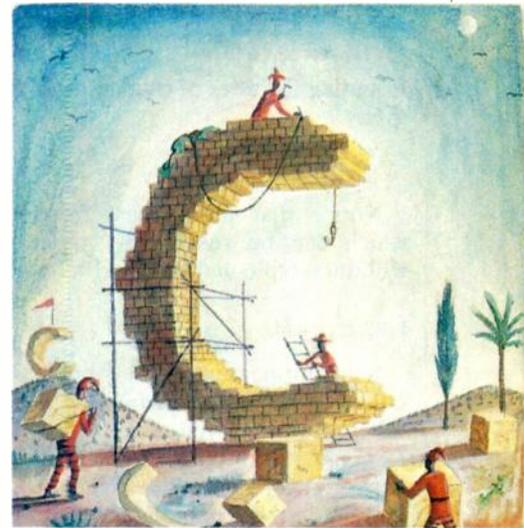


Fig. 3.2. Geometrical interpretation of a-d conversion.

Consider, for instance, the behaviour of an n-bit a-to-d converter without sample-and-hold. Let the full-scale ana-



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logue input be V_{fs} and the conversion time T . For a resolution of 2^{-n} , i.e. $1/2^n$, we may write:

$$\frac{\delta V}{V_{fs}} = \frac{1}{2^n}$$

Hence the smallest detectable change in the analogue input voltage is given by:

$$\delta V = V_{fs}/2^n$$

During the conversion cycle the change in analogue input signal must be less than δV to avoid ambiguity. Using Fig. 3.2 we may assume:

$$T \delta f(t)/\delta t \leq V_{fs}/2^n$$

Expressed simply, at the start of the conversion cycle the rate of change of the analogue input should be less than, or equal to, the full-scale input voltage divided by the product of the conversion time multiplied by 2 raised to the power of the word length. Adopting a sinusoidal input as an easy-to-understand example:

$$f(t) = V_{fs} \sin \omega t$$

Differentiating with respect to time, we may write:

$$\delta f(t)/\delta t = \omega V_{fs} \cos \omega t$$

Murphy's Law:
If anything can go wrong it will

PROGRAMMING

Assume that the conversion starts at $t = 0$:

$$\delta f(0)/\delta t = \omega V_{fs} = 2\pi f V_{fs}$$

Equating the rates of change, it follows that:

$$2\pi f V_{fs} = \frac{V_{fs}}{2^n T}$$

Notice that the highest frequency which can be successfully processed without sample-and-hold will be:

$$f = \frac{1}{2^{n+1}\pi T} \text{ Hz}$$

Substituting numerical values soon convinces one of the serious limitations inherent in this type of converter. For example, the 8-bit AD7820 half-flash converter converts in $2\mu\text{s}$. With sample-and-hold, the sampling theorem predicts the Nyquist frequency to be 250kHz. Without sample-and-hold the frequency is approximately 300Hz.

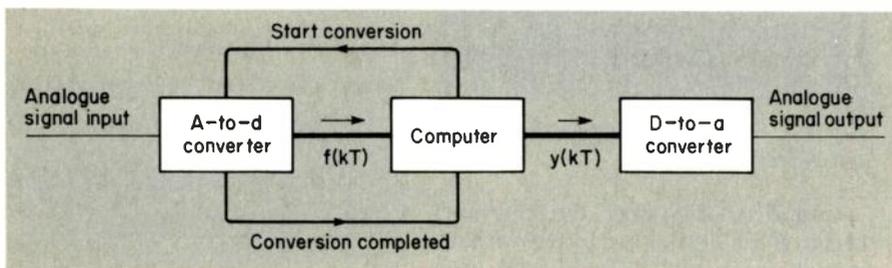
Before leaving this introduction it is worth mentioning that some form of digital signal processing will inevitably follow the sampling. Provided that the constraints imposed by the sampling theorem have been satisfied, it will be possible to recover the characteristics of the signal without corruption due to aliasing. Delivering the data back to the real world in analogue form requires further signal reconstruction, usually through a d-to-a converter and suitable low-pass filter (Fig. 3).

The a-to-d converter is synchronised with the main program by control signals which establish the signal processing time (T). A constant conversion rate and fixed processing time are essential to good design because they allow the behaviour of the system or filter to be modelled mathematically. The processed output is converted back to analogue form via the d-to-a converter.

Mathematical modelling

No-one has yet seen an electron, though some electronic engineers occasionally feel them. But progress is still made. Why should this be so? The answer is due in part to the inherent simplicity of

Fig. 3.3. Processing analogue signals through a digital system.



Ohm's law, which provides a mathematical model that allows us to predict how a circuit will, or ought to, behave. The ability to forecast the behaviour of a circuit or system, before it is built, is a basis for good design.

Electronic signal processing frequently involves the operations of integration, differentiation and time delay. Rather than describe the characteristics of the signal or linear system in the time domain, it is usually more convenient to use the complex frequency domain, or (if the signal is sinusoidal) the frequency domain.

A study of sampled data systems ought to include exposure to differential equations followed by familiarity with Laplace and z-transforms. Any mathematics is auxiliary to the main text, and mathematical methods are adopted only when it is impossible to make any further progress or develop a realistic understanding on the basis of intuition.

Wherever possible, mathematical relationships are shown diagrammatically, in an attempt to dispel the fog of abstraction. But remember that mathematical notation is usually a very terse and precise method of describing the facts. Accept this and understanding will come with use.

Laplace transforms

Laplace transformation decomposes a differential equation into an algebraic equation, allowing much of the manipulation to be carried out in a simpler domain. For example, differentiation in the time domain is equivalent to multiplication by s in the complex frequency domain. This is illustrated in Fig. 3.4 where double-headed arrows symbolise the integral relationships between domains.

Stated formally, the transformation from the time domain to the complex frequency domain is given by:

$$F(s) = \int_0^{\infty} f(t)e^{-st} dt$$

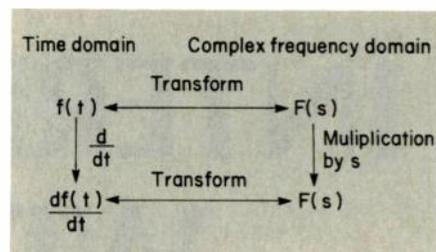


Fig. 3.4. To evaluate the Laplace transform of $df(t)/dt$, trace the top right-hand path giving $F(s)$ followed by the operation of multiplication by s . Alternatively, follow the left-hand path to obtain $df(t)/dt$ and then perform the transformation. The routes are equivalent.

The reverse process of inverse transformation:

$$f(t) = \frac{1}{2\pi j} \int_{\sigma-j\omega}^{\sigma+j\omega} F(s)e^{st} ds$$

is not so straightforward: the complex limits make this integral tricky. Fortunately, it is seldom necessary to employ such heavy artillery in practice.

I make use of the method of partial fractions together with tables of Laplace transform pairs. If you have time, confirm the solidarity of the mathematical scaffolding. This type of recreation increases confidence in the results.

Table 3.1. Elementary Laplace transform pairs

Signal $f(t)$	Transform $F(s)$
$\delta(t)$ [impulse]	1
$u(t)$ [step]	$1/s$
t [ramp]	$1/s^2$
$\exp(-at)$	$1/(s+a)$
$1 - \exp(-at)$	$a/s(s+a)$
$\sin(at)$	$a/(s^2+a^2)$

Electronic calculus

Time domain integration and differentiation are well understood. Provided the problem is selected carefully, classical analysis gives elegant solutions with a minimum of fuss. Electronic calculus compels engineers to adopt alternative methods to describe the characteristics of the linear processors which perform these operations.

The formal integral relationships between domains have already been outlined. One can identify certain key operations and describe the integral domain relationships by means of arrows. Figure 3.5 contains an adequate.

though by no means exhaustive, summary.

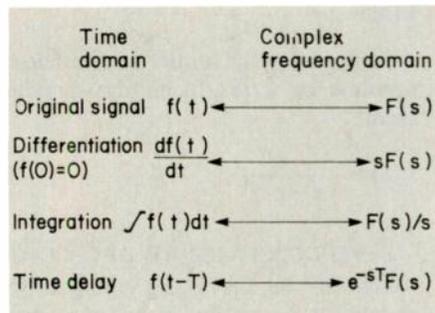


Fig. 3.5. A few important results demonstrating the equivalence of time-domain and frequency-domain signal processing.

Sampled-data signals are composed of a sequence of numerical values which represent the amplitude of the signal at the instant of sampling. Provided the sampling frequency is greater than twice the bandwidth of the signal it will be possible to process successfully these signals using digital methods. However, this is not the complete story. To produce a valid mathematical model of the sampling process, asynchronous conversion must be avoided. Instead the a-to-d conversion subroutine should be synchronised with the main program to ensure the time between samples is constant - this validates the use of z-transform methods to model time delays.

The next order of business is to present some important z-domain properties and to establish the relationships with the Laplace transform. The s-domain model of delayed signals includes the exponential term $\exp(-skT)$ which makes it difficult to use this equation:

$$f(t - kT) \longrightarrow \exp(-skT)F(s)$$

Fortunately there is another transform purpose-built for this application, the z-transform, which models time delays and advances with ease. Simply substitute:

$$z = e^{sT}$$

(time advance of one sampling interval)

or:

$$z^{-1} = e^{-sT}$$

(time delay of one sampling interval)

Figure 3.6 displays the transformations between domains diagrammatically. As an aid to comprehension, recall that the operation of multiplication by s in the complex frequency domain was equivalent to time domain differentiation. Similarly, multiplication by z in the z-domain is equivalent to a unit time delay in the time domain.

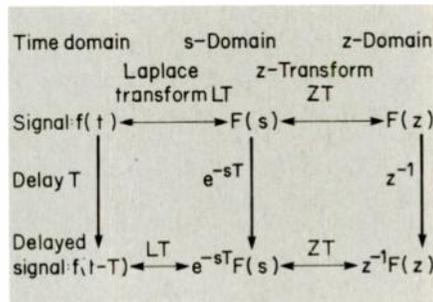


Fig. 3.6. The t , s and z domain representation of delayed signals.

Modelling a-to-d conversion

Referring to Fig. 3.7, notice the output of the a-to-d converter is the set of sample data, $f(0)$, $f(T)$, $f(2T)$, etc. which has defined values only at the instant of sampling.

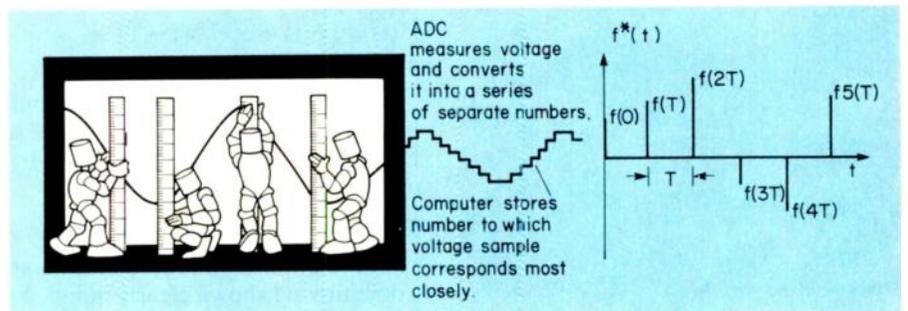


Fig. 3.7. Inside an a-d converter.

As a first attempt at describing the characteristics of the sampled-data signal, consider the a-to-d converter modelled as an impulse modulator, Fig. 3.8. When the switch is in position A, the output $f^*(t)$ is equal to the input $f(t)$. Changing the switch to position B makes the output zero. The interval between samples T depends on the sampling frequency, where $f = 1/T$.

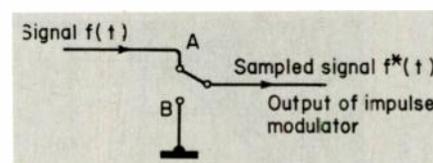


Fig. 3.8. Periodic processing of a continuous time-domain signal can be represented by this simple switching circuit. The asterisk notation denotes a periodic sampled signal occurring every T seconds.

In a practical a-to-d, the switch would be electronically controlled, designed to produce a train of pulses whose amplitudes are modulated by the input signal $f(t)$.

The output pulse train of the impulse modulator assumes the pulse width of each sample to be infinitesimal, while the height is a precise replica of the sampled signal at that instant only. In other words, the output data sequence can be modelled as a set of weighted and delayed impulses. Adopting the notation of sampled signals, the impulse train shown in Fig. 3.9 can be completely represented by the expression:

$$f^*(t) = f(0)\delta(t) + f(T)\delta(t-T) + f(2T)\delta(t-2T) + \dots$$

The Laplace transform of the weighted impulse train is:

$$F(s) = f(0) + e^{-sT}f(T) + e^{-2sT}f(2T) + \dots$$

Substituting $z = e^{-sT}$ we commute from the s-domain to the z-domain:

$$F(z) = z^0 f(0) + z^{-1} f(T) + z^{-2} f(2T) + \dots$$

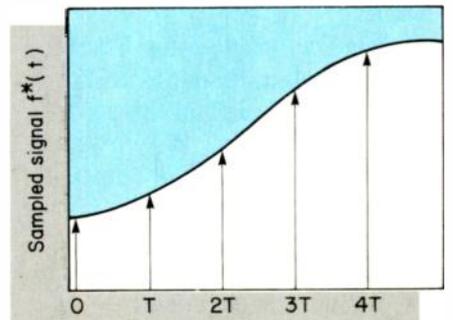
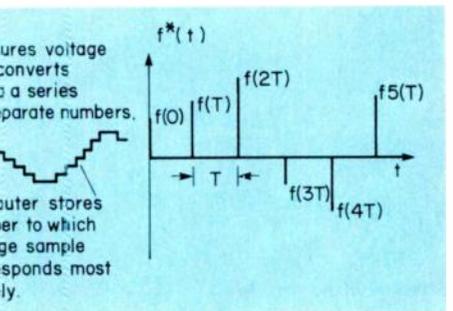
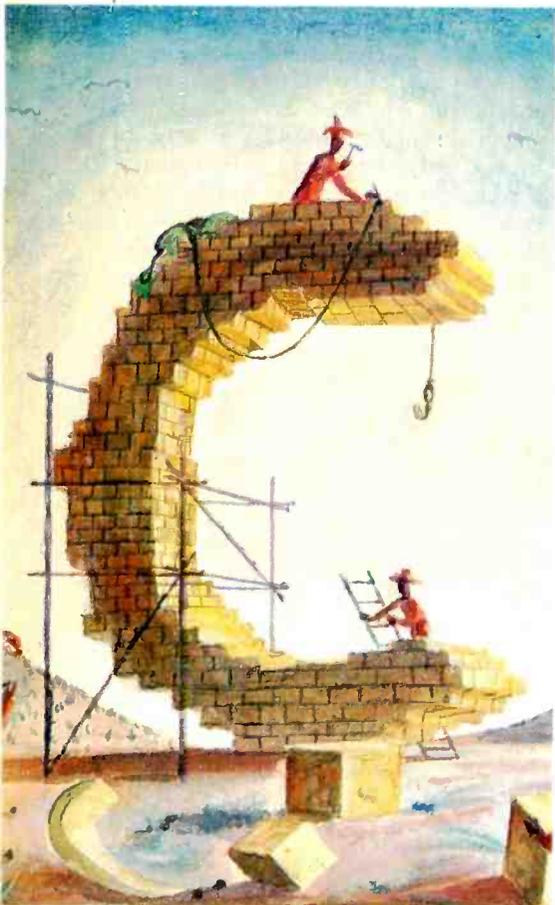


Fig. 3.9. Modelling the sampled signal as a train of weighted impulses. Suppose the signal to be sampled is represented by the line: provided the duration of each sample is infinitesimal, the sampled signal can be modelled as a train of weighted impulse functions depicted by the vertical arrows.

Our mathematical toolkit is now extensive enough to evaluate the z-transform of the decaying exponential signal $f(t) = e^{-at}$, sampled every T seconds as shown in Fig. 3.10.

Inspection of Fig. 3.10 shows that the output of the impulse modulator will be



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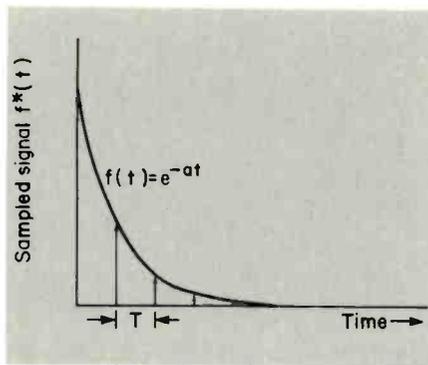


Fig. 3.10. Sampling the signal $f(t) = e^{-at}$ every T seconds.

the data sequence:

$$1, e^{-aT}, e^{-2aT}, \dots$$

Each of these terms weights the corresponding delayed impulse, hence the sampled time domain signal can be expressed as:

$$f^*(t) = 1 \delta(t) + e^{-aT} \delta(t - T) + e^{-2aT} \delta(t - 2T) + \dots$$

The sifting property of the impulse function allows us to write down the Laplace transform directly without having to integrate, so that:

$$F(s) = 1 + e^{-aT} e^{-sT} + e^{-2aT} e^{-2sT} + \dots$$

The relationships between the t , s and z domains are shown clearly in Fig. 3.6, hence the z -transform can be written as:

$$F(z) = 1 + e^{-aT} z^{-1} + e^{-2aT} z^{-2} + \dots$$

The series expansion of $F(z)$ is an infinite geometric series of the form:

$$1 + r + r^2 + r^3 + \dots$$

Provided that $r < 1$ the sum of such a series is given by:

$$\frac{1}{1-r} = \sum 1 + r + r^2 + r^3 + \dots$$

Using this result we can readily ex-

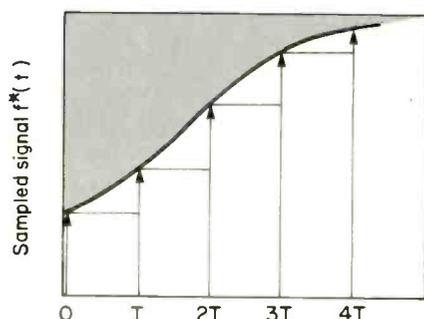


Fig. 3.11. Modelling the sampled signal as a train of weighted pulses. The effect of the a - d converter is to take a sample in zero time and hold this snapshot constant until the next sampling instant occurs one period later.

press $F(z)$ in closed form and write the z -transform as:

$$F(z) = \frac{1}{1 - e^{-aT} z^{-1}}$$

Multiplying the numerator and denominator by z results in the standard form:

$$F(z) = \frac{z}{z - e^{-aT}}$$

Zero-order sample-and-hold

Much of the foregoing analysis was necessary to consolidate our perception of sampled signals. However, practical a - t - d converters do not produce impulses, they process analogue waveforms into rectangular signals or pulses. Perhaps the most common is the zero-order sample-and-hold. Its effect is to take a sample and hold this snapshot constant until the next sampling event occurs one period later. Figure 3.11 contains the details.

Notice that the sampled signal $f^*(t)$ has been decomposed into a train of rectangular pulses each of width T .

$$f^*(t) = f(0)[u(t) - u(t - T)] + f(T)[u(t - T) - u(t - 2T)] + f(2T)[u(t - 2T) - u(t - 3T)] + \dots$$

The Laplace transform of the processed output is:

$$F^*(s) = \frac{f(0)[1 - e^{-Ts}]}{s} + \frac{f(T)[e^{-Ts} - e^{-2Ts}]}{s} + \frac{f(2T)[e^{-2Ts} - e^{-3Ts}]}{s} \dots$$

Taking out the common factor:

$$\frac{1 - e^{-Ts}}{s}$$

shows that the Laplace transform of the sampled data signal is equal to the transfer function of a - t - d (the common factor), multiplied by the output of the impulse modulator examined previously:

$$F^*(s) = \frac{(1 - e^{-Ts})}{s} [f(0) + f(T)e^{-Ts} + f(2T)e^{-2Ts} + \dots]$$

Thus the linear behaviour of the zero-order hold a - t - d converter can be modelled by the transfer function shown below and represented schematically in Fig. 3.12.

$$\frac{F^*(s)}{F(s)} = \frac{(1 - e^{-Ts})}{s}$$

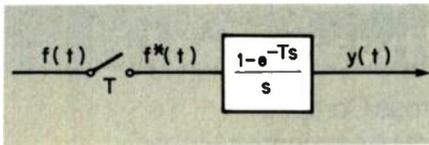


Fig. 3.12. Transfer function model of a-d converter.

The Laplace transform of the sampled signal is more usefully expressed in terms of its z-transform. To simplify the algebra associated with this type of problem simply substitute $z^{-1} = e^{-sT}$ in the transfer function.

$$F(z) = (1 - z^{-1}) Z \left[\frac{F(s)}{s} \right]$$

$$= \frac{(z - 1)}{z} Z \left[\frac{F(s)}{s} \right]$$

The Z symbol means "look up the z-transform of the Laplace transform within the brackets."

Table 3.2 lists some useful Laplace and z-transform pairs.

Table 3.2. Transform pairs

Signal f(t)	Transform F(s)	Transform F(z) of sampled signal
u(t)[step]	$\frac{1}{s}$	$\frac{z}{z-1}$
t[ramp]	$\frac{1}{s^2}$	$\frac{Tz}{(z-1)^2}$
exp(-at)	$\frac{1}{s+a}$	$\frac{z}{z-e^{-aT}}$
1-exp(-at)	$\frac{a}{s(s+a)}$	$\frac{z(1-e^{-aT})}{(z-1)(z-e^{-aT})}$

Example of obtaining the pulse transfer function of the continuous system

$$F(s) = \frac{ka}{s(s+a)}$$

$$\frac{Y(s)}{X(s)} = \left[\frac{1-e^{-sT}}{s} \right] F(s)$$

which may be written in terms of z transforms as,

$$\frac{Y(z)}{X(z)} = (1-z^{-1}) Z \left[\frac{ka}{s(s+a)} \right]$$

$$= \frac{z-1}{z} Z \left[\frac{z(1-e^{-aT})k}{(z-1)(z-e^{-aT})} \right] = \frac{k(1-e^{-aT})}{z-e^{-aT}}$$

The resulting z-transform leads naturally to the formation of a difference equation. Converting from trans-

forms to sequences we use the method shown in Table 3.3 to predict the behaviour of the system for a variety of time constants and/or sampling intervals.

Confident use of the transforms listed in Table 3.2 requires experience and the fastest way to gain experience is by making mistakes, preferably in private! The aim is to establish some confidence in the mathematical shorthand by way of an easy-to-understand example. Consider the effect of applying a unit-step to an ideal integrator. The following investigates this processing operation in each domain and demonstrates equivalence.

Method 1. Signal processing in the time-domain. Standard calculus allows us to write the processed output y(t) as the integral of the input x(t).

Since $x(t) = 1 (t > 0)$, we can write

$$y(t) = \int_0^t 1 dt$$

$$y(t) = t$$

Clearly, the operation of integration has converted the unit-step into a ramp-function.

Method 2. Signal processing in the complex-frequency domain. Inspection of Table 3.1 shows that the transform of the unit-step:

$$x(t) \rightarrow X(s) \text{ equals } 1/s$$

The operation of integration in the complex frequency domain is equivalent to division by s. Figure 3.4 contains the details. In this case the output Y(s) equals the transform of the input signal X(s) multiplied by the transfer function of the integrator H(s):

$$Y(s) = X(s) H(s)$$

$$= 1/s \cdot 1/s$$

$$= 1/s^2$$

To obtain the time-domain response we employ inverse transformation, i.e. use Table 3.2 in reverse to obtain $y(t) = t$. This is clearly in agreement with the time-domain analysis. These are standard results used by control engineers to model the dynamic behaviour of systems and are adopted by electronics engineers to model the effects of electronic calculus.

Method 3. Signal processing in the z-domain. In this case the input signal will be processed through a z.o.h. a-to-d converter and suitable stored program, conveniently modelled by the transfer function H(z). The behaviour of the composite system will be described by:

$$H(z) = \frac{z-1}{z} Z \left[\frac{H(s)}{s} \right]$$

Division by s is equivalent to time-domain integration, so that the z-transform model is given by:

$$H(z) = \frac{z-1}{z} Z \left[\frac{1}{s^2} \right]$$

Using table 3.2 we may write the z-transform of $1/s^2$ directly, so that:

$$H(z) = \frac{(z-1)Tz}{z(z-1)^2}$$

which simplifies to:

$$\frac{Y(z)}{X(z)} = \frac{T}{z-1}$$

Making the interval between samples unity and cross-multiplying gives the result:

$$zY(z) - Y(z) = X(z)$$

To obtain the difference equation we convert from transforms to sequences:

$$y(n+1) - y(n) = x(n)$$

expressed in terms of the current output:

$$y(n) = x(n-1) + y(n-1)$$

Table 3.3 The effect of processing a unit-step through a digital integrator.

Sample number n	Previous input x(n-1)	Current input x(n)	Previous output y(n-1)	Current output y(n)
0	0	1	0	0
1	1	1	0	1
2	1	1	1	2
3	1	1	2	3
4	1	1	3	4

This confirms the anticipated result.

Next month: Using convolution to commute between frequency and time domains.

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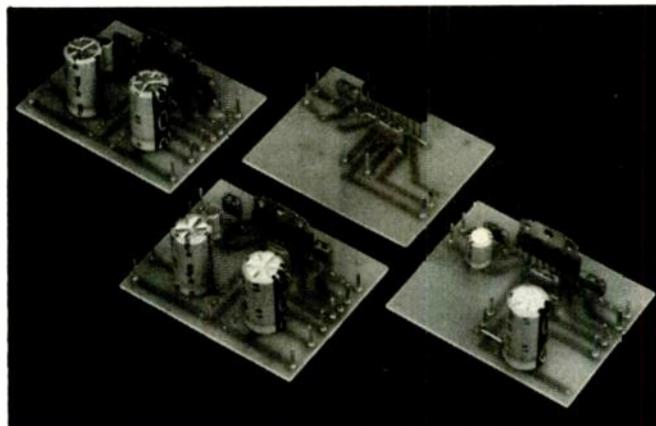
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Low-cost microcomputer. The Flat Controller is a low-cost microcomputer based on a highly integrated Z80 device. Its 64kbyte memory and vectored interrupts are claimed to make it more powerful than other 8-bit microcontrollers, and suitable for embedded applications. The I/O bus permits the addition of further devices, two or more Controllers may be interconnected, and a distributed network may be built using serial links. Ashton Computers, 0386 881256.

4-bit microcomputer. Features of a low-power, low-volume 4-bit microcomputer from Seiko Epson, the SMC6281, include up to 104 segment LCD drive, 1K of rom, 96 words of ram, low-battery detect circuit, timebase counter and stop watch, comparator, and sound generator. It operates from as low as 0.9V, and 2.5μA. Hero Electronics, 0525 405015.

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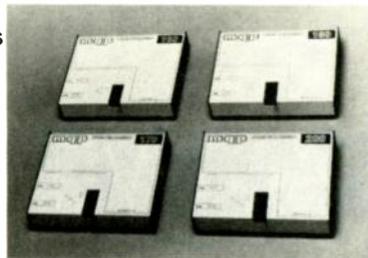
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PASSIVE

Passive components

Miniature potentiometer. The Model 3362 cermet trimming potentiometer, a 1/4in square single-turn sealed device, offers a choice of seven terminal styles and incorporates a multi-wire wiper arrangement which minimises contact-resistance variation. It also has a rotor suitable for automatic machine adjustment and graduated scale for easy orientation. The sealed package makes the device useful in harsh environments such as industrial applications and immersion cleaning processes. Resistance values range from 20Ω to 2MΩ, and contact-resistance variation is within 1% or 3Ω (whichever is greater). Standard resistance tolerance is ±20%, although a ±10% version is available to order, and adjustment angle is 240°. Bourns Electronics Ltd, 0276 692392.

Displays

Flat 5in monitor. A new 5in flat-screen monochrome monitor is designed for instrumentation applications but can also be used in other areas such as security systems. It comes in kit form (prealigned CRT and scan card) or as a standard Eurocassette, with flexible mounting options, and may be driven from either 12V or 15V DC. Screen dimensions are 94 × 70mm allowing a pixel resolution of 512 × 440. Digitran Ltd, 0763 261600.

High-performance VGA graphics.

The INMOS IMSG178 high-performance CMOS colour look-up table device, a single monolithic high-speed IC compatible with the RS170 video standard, is designed to replace TTL/ECL components in computer colour display circuits compatible with IBM PS/2 VGA graphics systems. Its advantages in VGA graphics circuits are said to be reduced component costs, reduced board area usage and relatively low DC power consumption. It is downward software compatible with the IMSG17 family. Hawke Components Distribution, 081-979 7799.

Filters

Filter arrays. The LMF120 and LMF121 mask-programmable switched-capacitor filter arrays are capable of implementing up to a 12-pole filter in one 16-pin, cost-effective package, being designed for applications including communications systems such as mobile phones, anti-alias filters, biomedical instrumentation, control systems, and real-time audio analysers. Both can realise all filter types. The LMF120 has right-half-plane zeros and the LMF121 left-half-plane zeros. National Semiconductor, 0044 793 697572.

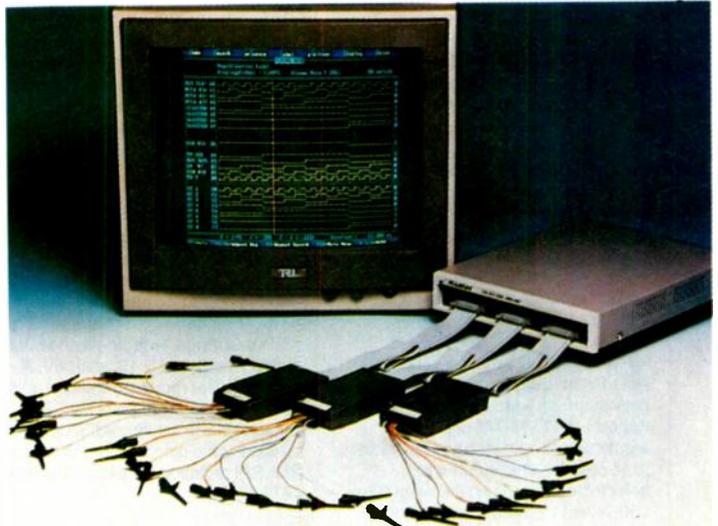
High-performance filter. The PDSP16256, a programmable, variable-length FIR filter, is intended for use in high-performance filters, pulse compressors and converters, providing filter function without software development, with options selected by loading a control register. It contains sixteen 16-bit multiplier accumulators that can be multi-cycled to provide from 16 to 128 stages of digital filtering at sample rates from 20MHz to 2.5MHz. The device may be configured as one long filter, or as two half-length filters for filtering complex data, and can be cascaded to provide filters of any length. Plessey Semiconductors, 0793 518000.

Instrumentation

Precision universal counter. The GT200 is claimed to be the first precision universal counter in a PC-resident format, and to provide all the performance and features found in a bench-top universal counter but with easier programming and lower cost. Designed to run on IBM PC XT AT and compatibles, it can measure frequencies with 10-digit resolution per second of gate time, and make timing measurements to 100ps resolution without averaging. Its two DC-coupled input channels are specified for input frequencies of DC to 75MHz and can accept signals from -5V to +5V. Auxiliary inputs arm the trigger and allow gated counting. The instrument performs frequency, period, time-interval, delay-time-interval, ratio, totalise, gated-totalise, and pulse-width measurements, can sustain 2000 measurements per second, and is supplied with a "virtual" front panel. Amplicon Liveline, 0273 570220.

Portable air-velocity meter. The Data-Trak is a portable air-velocity and temperature multimeter which retains up to 1000 readings, displaying them digitally and computing the averages. Microprocessor-based, it is claimed to offer fast and accurate display of average velocities, individual point velocities and temperatures. Five control buttons provide access to functions, and an optical handheld printer provides hard copies. Dage (GB) Ltd, 0296 393200.

Digital handheld pH meter. The 8000 digital handheld pH meter is simple to use, measuring 0.00 to 14.00pH with an accuracy of 0.02pH. All measurements are displayed on the 12.5mm-high LCD, which incorporates the mode of measurement and low battery indication. A complete kit includes the digital meter, both pH electrode and buffer solutions housed in an ABS case. ETI Ltd, 0903 202151.



IMS's PCL-510 — a complete hardware/software package

LCD detection system. IT'S is a liquid-crystal detection system said to be very precise and versatile, and able to detect hot spots in silicon wafers and packaged devices at intervals less than 1μm. Temperature can be regulated very precisely (±1 C) and tests can be carried out within a temperature range of -10 C to +130 C. It is claimed that the system can analyse a wide range of ICs, including SO, DIP, LCC, PGA, hybrid circuits, or isolated dies, and can be used wherever high-precision temperature monitoring operations are needed. I.R.I.S, 0895 70646.

100MHz logic analyser. The PCL-510 PC logic analyser is a complete hardware software package claimed to turn a PC into a powerful logic analyser, providing 100MHz sampling for eight channels with 8K memory depth or up to 24 channels with 2K memory depth at 25MHz sampling. Features include four sequential trigger conditions, eight data qualification channels, twin cursor (Y and X) for precision timing measurement, 4000 multilevel combinational trigger conditions and variable input threshold between ±10V with 0.1V increment step. Integrated Measurement Systems Ltd, 0703 771143.

Antenna aligner. Propoint is a spectrum analyser unit and TV display combined into one compact, portable piece of equipment, designed as a research tool and developed to meet the needs of data communication dish installation, where high-precision positioning and repeatability are essential. The screen displays the full spectrum of a satellite IF or UHF at a glance, making accurate Yagi aerial and satellite dish alignment a simple operation. Oakbury Components Ltd, 0488 71458.

Oscilloscope power. A pocket-sized power supply module operating from 12V DC, the PM8902, is said to satisfy the need for high-performance portable combination oscilloscopes. It will power

any analogue or digital storage Smart Scope to give true laboratory performance in the field, and may be driven by a car battery, an optional 12V portable power pack, or other 12V DC supply. Output is 115V RMS AC (60Hz) rated at 100W. Philips Test & Measurement, 0923 240511.

Immunity measurement. The TS9980 test system allows automatic measurement of the electromagnetic immunity of radio and TV receivers, tuners, AF and video amplifiers, etc. The menu-driven user interface is claimed to permit easy operation, so that error-free measurements are quick and practical. A significant feature is the interpretation of test results by colour graphs and numerical outputs. All results can be recorded on a printer or a plotter and stored for later evaluation. Rohde & Schwarz UK Ltd, 0252 811377.

60MHz oscilloscope. The Hitachi V680 60MHz, three-channel oscilloscope features delayed timebase, cursor measurement and on-screen readout. The cursor system can be used to measure voltage difference (ΔV), time differences (ΔT), or frequency (1/ΔT). These values, as well as set-up parameters and delay time, can be displayed on screen. On channels 1 and 2, sensitivity can be varied between 5mV/div and 5V/div over the full bandwidth, 1mV/div and 2mV/div are available up to 10MHz, and 0.1V/div and 1V/div are available on channel 3. Sweep time on the main A timebase can be varied between 50ns/div and 0.5s/div, and the delayed B timebase between 50ns/div and 50ms/div. A ×10 magnifier gives maximum sweep time of 5ns/div on both timebases. Thurlby-Thandar Ltd, 0480 412451.

Interfaces

Processor card. A variant on the ECPC single Eurocard PC-compatible processor card is said to drive virtually

NEW PRODUCTS CLASSIFIED

all types of backlit LCD displays including 640 x 200 and 640 x 400 formats. The ECPC-LCD's driver chip and software is claimed to run standard MDA (monochrome) and CGA software. CGA colours being emulated with greyscale and crosshatch patterning. Text fonts stored in ram can be designed with utility programs provided with the card. To save expense, the ECPC-LCD includes a negative voltage switch-mode power supply to generate the additional voltages required. DSP Design Ltd, 071-482 1773.

16/24 digital I/O board. Providing 64 opto-isolated digital outputs and with provision for a watchdog security handler, the TVM-743 digital input/output board also has available up to 16 opto-isolated inputs with selectable debouncing delays of 15 μ s, 250 μ s and 1.5ms. Minimum impulse width is 60 μ s. Input can be selected in groups of four for contact closure toward ground or +24V. Each input may assert an interrupt request and I/O isolation is 1kV RMS. GMT Electronic Systems Ltd, 0372 373603.

Fault-protected multiplexers. Of two new families of fault-protected multiplexers, one claims protection against voltage of up to ± 75 V, the other ± 35 V protection including latches to relieve the microprocessor from maintaining a constant address on the logic inputs. The MAX378 and MAX379 are high-voltage multiplexers with, respectively, 1 of 8 and dual 1 of 4 configurations. These devices are said to permit less than 1nA per channel of input leakage current when the power is off and the inputs are at ± 75 V. With supplies of ± 15 V the device is protected against continuous input voltage of ± 60 V with equally low leakage currents. Maxim Integrated Products, 0734 845255.

Hard-disk interface chip. The ADG9000 incorporates a fully digital data separator simulating analogue PPL with a temperature-compensated internal time reference. Features include write precompensation, clock and R/W clock generation with accuracy claimed better than 7ns, and direct interface with hard-disk controller chips uPD7261Ad (nmos) and uPD72061 (cmos) to support the ST506/ST412 standard with MFM data at 5Mbits/s and enable simple, compact, hard-disk controller units to be built. NEC Electronics (UK) Ltd, 0908 691133.

Power supplies

DC/DC converters. The XC series of 15W, 24V input DC/DC converters are a broad range of isolated and pulse width modulated (PWM) converters, of claimed high performance and efficiency, encapsulated in metal cans with six-sided EMI/RFI shielding. By using cmos and mosfet technology, all converters are said to operate at 100kHz with up to 82% efficiency, virtually independent of load and line. They accept a wide range of input

voltages — 9 to 18V DC, and 36 to 72V DC. Output voltages are single, dual or triple; 5, 12 and 15V. All are designed for mounting onto PCBs and are free-air convection cooled. Remote on/off control is provided and an internal LC input filter is standard. Amplicon Liveline, 0273 570220.

48V input converters. A series of 48V input DC/DC converters for telecommunications applications claim to provide a power range of 5 to 20W, with low ripple and efficiency as high as 83%. The Z series offers a wide 30 to 72V input range, with single and dual output voltages of 5, 12, and 15V. A built-in Pi filter reduces reflected noise from the converter back to the input voltage source, while continuous six-sided shielding is said to minimise EMI/RFI radiation. Amplicon Liveline, 0273 570220.

3000W power supplies. New switching power supplies in the SuperSwitcher range are claimed to deliver from 400 to 3000W but if even more power is needed they can be paralleled since they feature automatic current sharing. SuperSwitchers are said to provide reliable power with up to 155,000-hour demonstrated MTBF. They meet the international safety and emission specifications of VDE, IEC, CSA, UL and FCC and each unit is subjected to 100% dynamic burn-in. They are available in six commonly used single output voltages from 5V to 48V. A 2V rail version is available with a power rating of up to 1400W. Astec Europe Ltd, 0483 756066.

High performance PWM. The SG1825 high-speed current-mode pulse-width modulator (PWM) is claimed to show no output driver "float", to have no pulse-width instability at cold temperatures and to offer lower start-up current and higher ground noise tolerance than the industry standard. For high-frequency switched-mode power supply applications, it is claimed to reduce volume and weight of power supplies and to increase reliability. A new linear Schottky fabrication process is claimed to result in very short propagation delays through the current limit comparator, logic, and output drivers. A four-fold increase in switching frequency is claimed. Britcomp Sales Ltd, 0372 377779.

20W converter. The SB24S05 4000Z is an unisolated 20W DC/DC converter in the industry standard 2 x 2 in package, designed for use in remote mobile and vehicular applications. Operating from input voltages of 8 to 40V, it delivers a precisely regulated 5V at 4A. Typical efficiency is claimed to be 84%, with 80% maintained even at half load. With an operating temperature range of -25 to +75°C ambient air, full output power is delivered without the need for derating. Computer Products Power Conversion Ltd, 0234 273838.

Space-saving converters. The MAX654 6/7 DC/DC converters feature a guaranteed 1.15V start-up, and are

said to continue to function as the input voltage drops below 1V. This new range is claimed to supply up to 450mA (MAX658) of output current with a minimum of external components, achieving typical conversion efficiency of up to 75% in generating a regulated +5V or +3V from a low input source. Kudos Thame Ltd, 0734 351010.

Programmers

Mid-range programming. A new mid-range programming system featuring universal pin drivers for up to 44 pins is targeted at a new generation of programmable logic and memory IC users. The 2900 Programming System is designed to handle the different PLD architectures as well as eeproms, eeproms and programmable microcomputers in a variety of packages, both DIP and surface-mount. Instrumatic UK Ltd, 0628 476741.

Switches and relays

Rugged floatswitches. Two rugged PVC floatswitches claimed to be capable of operating in liquids down to 0.75g and temperatures up to 60°C are interchangeable for normally-open and normally-closed operation and are available for top- or bottom-level indication. Other specifications include contact rating of 10VA, switching current of 0.5A AC or DC, and switching voltage of 200V AC or DC. Wide resistance to chemicals is said to

include many strong acids and alkalis. FR Electronics, 0202 897969.

Transducers and sensors.

Low-cost pressure sensors. Models 90 and 93 are media-compatible, solid-state pressure sensors said to be capable of measuring in both gauge and absolute, pressure ranges from 0-5 to 0-3000psi with infinite resolution and $\pm 0.25\%$ accuracy. Both use 316 stainless steel to protect the sensor from corrosive liquids/gases, with a 1.4-18 NPT pressure fitting welded to the body providing a pressure connection. Output span is 100mV (typ.) with a 1.5mA excitation. EuroSensor, 071-405 6060.

Vision systems

High resolution CCD imager. A frame transfer 1in-format TV CCD imager, said to be capable of offering a full 1032 TV lines per picture width and using standard three-phase clocking, is offered for high-performance commercial, industrial and military systems. The C07 series has four optional forms and offers either conventional fixed barrier anti-blooming, or a gated anti-blooming system providing fast unwanted charge dump with a 4000:1 dynamic range. Full support modules have been developed and provisional camera circuitry giving a 1V video signal output is available now. Intensifier coupled versions will also be available. EEV, 0245 493493.

COMPUTER

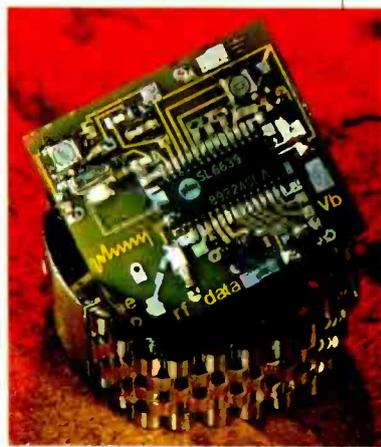
Computer board level products

Flexible MSC. A multi-channel serial card for PCs, the MSC-8, is claimed to be more flexible and to offer better value for money than any other board of its type. It has eight serial channels on one card, each able to be set to RS232, RS422, RS485 or 20mA loop at any individual address location and any interrupt channel. It is intended for use in industrial applications for communicating with large numbers of intelligent sensors and instruments and in lams, shopfloor data collection, time and attendance systems, etc. Blue Chip Technology, 0244 520222.

Floating-point DSP. The DBV32, a fast floating-point DSP board for the VMEbus, is based on AT&T's popular DSP32C processor, and is claimed to achieve speeds in excess of 25Mflops (25 million floating point operations per second). It is suited to fast scientific applications such as signal analysis, instrumentation or rapid numeric processing. Data Beta Ltd, 0734 303631.

Vector/signal processor. The IMS B420 VecTRAM, the latest addition to INMOS' IQ Systems range of TRAnsputer Modules (TRAMS) offers both scalar and vector signal processing in a single compact module.

Demonstration card for Plessey's SL6639 RF data receiver



NEW PRODUCTS CLASSIFIED

It combines the general purpose scalar processing and communication capabilities of the transputer with the functionality of a high-performance vector-signal processing co-processor in a size-4 standard TRAM. The latter is said to make it ideal for many DSP-based systems and numerically intensive application areas. INMOS Ltd. 0454 616616.

Data communications products

Sine wave generators. Two integrated circuits each replace multiple components in the generation of highly precise sine waves. The ML2035 programmable sine wave generator is the industry's first integrated sine wave generator in an eight-pin mini-DIP while the ML2036, in addition to the capabilities of the ML2035, allows control of sine wave magnitude, prevents steps in voltages when inhibiting the sine wave output, and provides clock outputs designed to drive other devices. Ambar Cascom Ltd. 0296 434141.

High-speed modem. Smartmodem 9600, a CCITT V.32-compatible modem for Data Communications Network applications, is BAPT approved for connection to UK phone systems and is intended primarily for leased-line applications. The product contains automatic dial back-up features and is designed for mainframe and lan environments requiring high-speed, full-duplex communications for transferring large volumes of data. Hayes Microcomputer Products. 081-848 1858.

Single-chip coder. A series of cmos PCM coder and decoder ICs are the XR-T3052/3053 3054 and 3057, intended to implement the transmit and receive functions in digital telecommunications circuits. The first three use a companded μ -law PCM format, the 3057 an A-law PCM format. Each contains separate D/A and A/D circuitry, all necessary

TI's MOD2434 single-board modem



sample-and-hold switched capacitor filters, precision voltage reference, internal auto zero circuitry, and a serial I/O interface. Microlog Ltd. 0483 729551.

Miniature receiver. A miniature, low power, direct conversion, RF data receiver for the personal communications market is claimed to be ideal for applications such as wristwatch pagers. The SL6639 is a low-noise, RF cascade amplifier and a data demodulator. Its on-chip channel filters detect frequency shift keyed transmissions and reduce the need for additional external components. This compact solution of a complete radio receiver in one MP28 miniature DIL package permits smaller RF modems, security devices, and pagers that can be built into wristwatches or credit card-sized packages. In contrast to custom chips, the SL6639 opens the market for a wide range of manufacturers. Plessey Semiconductors, 0793 518000.

Multi-standard modem. The MOD2434, a single-board multi-standard PC modem designed to transmit error-free data throughout the world, is said to contain an innovative configuration of digital signal processors and microcontrollers, which, combined with new and more efficient software algorithms, is claimed to result in an impressive array of features. These include auto-adaptation in both originate and answer modes, error-free transmission at 9.6kbaud and compatibility with existing communications software. The unit is also claimed to be smaller and less expensive than any comparable product. Texas Instruments, 0234 223252.

Mass storage devices

Paged memory board. A paged-memory expansion board for STEbus users, SPER, is said to be ideal for implementing rom- or ram-disk system facilities. It allows designers to build systems with rugged solid-state memory for industrial computing applications where dust or vibration might adversely affect a delicate floppy drive mechanism. It can be populated completely with rom to house large application programs, with all ram for fast datalogging tasks, or with a mix of both. Arcom Control Systems Ltd. 0223 411200.

Software

Interactive autorouter. Pads-Push n' Shove uses a costed-Maze algorithm operating in a push-and-shove mode. When a connection being routed is blocked by previously routed tracks, the push and shove feature moves these existing tracks aside in order to make room for the connection being routed. The software is useful in designing analogue PCBs as well as boards that have critical circuits where in the exact track pattern must be controlled. It operates in interactive automatic, steered automatic and manual modes. Lloyd Doyle Ltd. 0932 245000.

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Appropriate technology

I was very interested in your editorial in the May 1990 issue of *Electronics World and Wireless World*. I could not agree more with your statement, "Instructions are complicated because equipment and software offer so much. Users, in the main, haven't actually requested the diversity of features. They appear because equipment and software designers can put them there at 'little cost'."

I would go further and say that I find most systems are designed around what technology can do (system-centred) rather than what people actually need (human-centred). It is becoming increasingly apparent that the systems-centred approach is wrong. For example: factories using large and complex system-centred production control packages are finding that the benefits of these packages are almost as elusive as the staff needed to run them!

I can only hope that we, as engineers, realise where this headlong rush into automation and technologically-led systems in general is taking us before it is too late. Generally automation has been synonymous with *skill reduction*. We should be attempting, through the appropriate use of technology, to encourage the *growth* of the

UK's skill base rather than the current trend of de-skilling through automation.

The UK has suffered enough. We are at a turning point. Let us design systems that prepare us for the year 2000 and beyond. Let us go forward, not with technology that is technically tasty and that scratches some mental itch, but with technology that is appropriate to our *needs*! Let us all adopt the human-centred design philosophy that builds skill rather than kills skill!
A W S Ainger
BICC Systems Development Centre
Hemel Hempstead

A rotten apple

I would like to add my voice to the current discussion on the relative merits of the Apple Macintosh versus the IBM PC family.

Until recently I was a programmer in a York company using both Macs and PCs. Fortunately most of my work was connected with the PC rather than the Mac. I rapidly came to the conclusion that the mouse environment is only of value within art packages, where it is far superior to the cursor keys for drawing purposes. Otherwise it is slower and more awkward to move around in a text editor using the mouse, and involves

the highly annoying need to keep taking one hand off the keyboard. Similarly, the mouse is more cumbersome than typing text responses to a command-line prompt.

Far more annoying is the slowness resulting from using a micro to run a supercomputer-sized operating system containing many inefficiencies and ridiculous design features.

One would have thought that with a 1Mb ram and a basic system file of around 330K the entire system could be loaded at boot time and held in memory thereafter. Yet trivial disk accesses to get data (often only a few bytes) are made all the time, and make operating a Mac without a hard disk a nightmare of switching floppies.

Compiling and linking, say, 100K of C source from hard disk can take about 10 minutes on the Mac, against about one minute on the PC. The code on the PC can be launched for testing in a couple of seconds; on the Mac, 20 seconds is more likely, and the machine will crash more often than not — a rare occurrence on the PC.

Rebooting the Mac and relaunching the development environment can easily take four minutes. A reboot from a System disk, apparently identical to one that works perfectly, can fail with an obscure error such as, "Cannot load needed resource."

Privateer television

The recent provision of large numbers of additional television channels by Sky and BSB might be regarded by some as signalling the ultimate demise of terrestrial television. After all, the cost of the satellite services, however great, must surely be dwarfed by the huge expenses incurred by the BBC and IBA in maintaining a country-wide network of UHF transmitters.

Such economic arguments ignore the technical vulnerability of the satellites. Quite obviously, the uplinks will always be captured by the strongest signals. Could we entertain the prospect of a national television service which may be held hostage by anyone with a big dish and a magnetron?

J. Wilson
London SW20

To identify the missing (and how so?) resource involves breaking to the monitor and tracing back through the incredibly complex chain of code which displays the error message to locate the parameters of the failed GetResource call.

Hardly the ideal machine for the computer illiterate, as it purports to be.
C. G. Bulman
Fishergate
York

General physics

R. V. Harrowell's letter (*EW + WW*, May 1990) proposing a method for suppressing dissident authors, puts me in mind of verses attributed to John Barrett of Harvard University.

I AM THE VERY MODEL OF A MODERN PHYSICS THEORIST
(with apologies to Gilbert and Sullivan)

I've studied all the references
in order alphabetical.
My judgement is, which some of you
may find to be heretical,
A field that really is abstruse,
A field where many screws are loose,
A field quite famous for its spoofs
Is PHYSICS THEORETICAL.

We never do experiments,
We shun the purely practical.
Our best work is in getting grants,
Our budgets are fantastical.
We laugh as others take a dive,
If funding falls we still survive,

We make damn sure our job's secure
In PHYSICS THEORETICAL.

Our scientific breakthroughs are,
to say the least, debatable.
We silence critics haughtily.
Our egos are inflatable.
The rest of science goes along,
Because our last defence is strong,
THEY cannot prove that WE are wrong
In PHYSICS THEORETICAL.

Yours truly,
John Ferguson
Camberley
Surrey

Riddle of Inertia

R. A. Waldron in the May letters (p. 429) glosses over the real problem of motion under gravity. The mass or inertia of a freely moving body measures its reluctance to move faster when pushed. The electrostatic force between two bodies is proportional to the product of the electric charges on them. Similarly, the gravitational force between two bodies is proportional to the product of the 'gravitational charges' on them. However, all experiments show that the gravitational charge on a body is simply proportional to its mass. This *empirical* fact does call for an explanation, and in his General Theory of Relativity Einstein provided one.

Mach's principle attributes the

Continued over page ►

mass of a body to its interactions with the rest of the universe, and Assis in his 1989 paper (Foundations of Physics Letters 2, 301) postulated a 'relational' theory in which the interaction used was Weber's force law for gravitation. On the basis of that paper Graneau claimed in his January 1990 article that the inertia of a body depends on the instantaneous positions of objects in the rest of the universe, i.e. on instantaneous action at a distance. However, towards the end of the paper, Assis wrote, "*The greatest limitation of this model is that it is based on an action-at-a-distance theory. As a result, it is not definitive or final theory, but should be valid in systems with slowly varying motions in which time retardation is not a serious factor.*"

C. F. Coleman
Grove
Oxfordshire

RC oscillators

Recent letters have revived interest in RC oscillators. Around 1938 I was interested in crystal, LC, and RC filters and realised that RC networks could be used as frequency discriminating devices in oscillators. Because of the low effective Q value it was necessary to operate the maintaining amplifier in class-A to obtain a sinusoidal waveform with low distortion and this necessitated the use of an amplitude-limiting device. I used small filament lamps as thermal devices for the AGC system in the maintaining amplifier and oscillators of this type were used in the following few years when I, and many others, were busy on other things. Soon after the war, I used a commercial 5W mains filament lamp and a full theory was given in an article circa 1945 in *Electronic Engineering*. Since then thermistors, isolators, photo cells and fets have been used in place of filament lamps in attempts to reduce distortion and amplitude bounce. A completely bounce-free oscillator with low THD (0.01%) has been produced by the undersigned and an article on it is in preparation.

At the instant of switching on, voltage and current transients

Lumpy Universe

A regular expansion after the Big Bang appears to be entirely consistent with a lumpy universe (Research Notes, May) and in particular with an 'onion skin' model giving separated walls of galaxies.

It is reasonable to assume that, immediately following the Big Bang, matter would be leaving the focal point in all possible directions and over a wide range of speeds.

If all particles originated from exactly the same point in space and time there would be a near infinite series of concentric spherical shells each expanding at a slightly greater rate. Velocity differences, between closely spaced particles, would be negligible. The smallest irregularities of mass would tend to grow under gravitational forces, giving a random distribution of mass concentrations.

are present together with noise, and microphony when valves are used, and these contain frequencies close to those of oscillation which are amplified and fed back, positively, to rapidly build up to the operating amplitude, in a good design. This assumes that no spurious voltage is applied to the AGC system as this can result in a delayed amplitude build-up. It is perhaps of interest to note that because of the relatively poor frequency discrimination of RC networks in oscillators, compared with crystals and LCs, the noise level is usually inferior.

We are living in an age of technical reinvention and although it is important to remind readers of the state of the science it is ethical for the author to ensure that the reader understands what is new and what is old and should be known. Too many authors enjoy the adulations of a semi-technical public by forgetting to give credit to past work by other people and thus, in effect, claim this as their own work. RC networks are no exception to this practice in as much as we have recent examples in letters and articles in *EW+WW*. As examples, a letter

On the other hand if the origin had significant dimensions, in space and/or time, later fast shells would overtake earlier slow shells. Shells would clump automatically in a regular way. Whenever major clumping occurred the resultant gravitational forces would overcome some of the velocity differences: a more massive shell would continue at an intermediate velocity. Irregularities of mass, within each shell, could grow to form larger bodies.

Astronomers were discussing this concept several years ago. It is easy to demonstrate on a computer. Why has it been ignored by mainstream astronomers? Are there weaknesses that appear under deeper analysis?

R.G. Silson
Tring
Hertfordshire

reinvented the circuit of an RC oscillator I patented over 45 years previously and a recent article on RC networks described networks mainly used in amplifiers that were known before the war under the general title of "Wheeler networks." F. G. Clifford
Wetton
South Africa

Gyroscopes

In the March edition of *EW+WW*, there are two letters with comments relating to my letter of December 1989. As an engineer I fully agree that experiment is the final arbiter of any theory, which is why my armchair is very close to the laboratory.

I find it difficult to believe that a difference of 20% from the classical results of mechanics could go unexploited for so long. I for one would be delighted to rationalise such phenomena, should they exist.

One specific case mentioned by Bruce DePalma was that of a pendulum with a rotating bob—this is easily tested using a simple toy gyroscope and two pieces of string. If the axis of rotation is

maintained horizontal by suspending the gyro with equal lengths of string from each end of the axle there is no measurable variation of the periodic time. However, if suspension is from one end only such that the axis of rotation is free to change its orientation then interesting gyrations take place and the system no longer behaves like a simple or compound pendulum. The motion is complex because the gyroscopic effects couple with the pendulous motion and, in general, the detail of the motion depends on the starting conditions. There is no evidence that classical mechanics needs to be modified.

The experiment indicated by Alex Jones was repeated by myself some 15 years ago, except that in place of the bearings I mounted the base on four flexible steel strips such that the stiffness was low in one direction of the horizontal plane. In this case movement of the frame is clearly seen to take place. The higher the spin speed the lower the precessional rate, giving a smaller centripetal acceleration to the centre of mass of the rotor. Thus a smaller force is required to prevent motion of the frame. The horizontal force required to prevent lateral motion could easily be less than 0.1% of the weight, in which case rolling friction would be enough to provide the constraint.

I believe that I have said enough on this matter, save to say that advancement in science does require speculation but it must be coupled with well documented experimental apparatus and procedure. Also it is essential that predictions are made by the proper use of existing theory and compared with experimental results. The application of the theory of three-dimensional dynamics of a rigid body is not easy and therefore false conclusions are often drawn. The effects of friction often make prediction a lengthy procedure, but classical mechanics still gives the most concise explanation of the macroscopic dynamics of bodies at low speeds. H. R. Harrison
Senior Lecturer, Engineering Dynamics
City University
London EC1

Continued over page ►

Alpha-torque Forces

I would like to respond collectively to the alpha-torque letters in the September and October 1989 and February 1990 issues of your journal. Good luck to Simon Young with his MkII water-arc gun. I hope he is careful and will let us know what he found.

A common theme in the letters of Coleman, Bell and Carpenter is that I have ignored electromotive forces. It is my considered opinion that these forces are non-mechanical and therefore cannot be weighed with a balance. For this reason they cannot contribute to the fracture of atomic bonds in wire-fragmentation experiments. Ampère tension, as I have called it, is the strongest evidence in favour of Ampère's force law and against the Lorentz force law. I would go further and argue that we still do not understand how e.m.f.s are produced in wires. At this point textbooks invoke the Poynting vector mechanism which somehow generates an electric field in the wire. The Poynting vector stands for field energy momentum. With an electrodynamic impulse pendulum we have shown that this momentum does not exist. All these experiments are fully described in my book, *Ampère-Neumann Electrodynamics of Metals* (Hadronic Press, Nonantum MA, 1985).

As Carpenter admits, the electrons slip freely through the metal lattice unless they run up against the metal surface. But there are no surface barriers in the way in the direction of current flow. Even in the shortest pulses there appears to occur no pile-up of electrons which could give rise to a Coulomb explosion.

Perhaps I should have outlined why pinch forces cannot explain the liquid mercury fountain. If we consider the more simple example of a cylindrical column of liquid metal, the pinch force will certainly generate axial hydrostatic pressure. Pressure is not unidirectional. It will push liquid up and down. The downward thrust will prevent the

entry of fresh liquid to the column (at the copper stem). This makes the observed circulation of liquid metal impossible.

Bell's comment makes it necessary to say a few words about electron beams in vacuum. Longitudinal Ampère tension would disrupt these beams. In practice, these beams are very stable, which proves that they are not subject to Ampère tension. It does not prevent the electrons from responding to externally applied magnetic fields, as they certainly do in magnetic focussing and deflection. It all goes to show that a current in a wire is something very different from an electron beam in vacuum. It is a fact that charges convecting in particle accelerators do obey the Lorentz force and not Ampère's law.

Carpenter raises a valid point about three types of energy involved in electrodynamic experiments: mechanical work, electric power and stored magnetic energy. Anyone who has looked at my book will know how I have agonised to clarify this issue. I made the mistake of talking about the virtual work principle, which is unambiguous only in the interchange of mechanical work with gravitational potential energy. In fact I may have drawn the alpha-torque forces in the wrong direction, but reserve judgement until more experimental facts come to light. John Carpenter asks whether field theory should be overturned? This is not my objective. I have tried to establish an alternative for the restricted area of metal electrodynamic. In my view two theories are better than one. My remaining comments will illustrate this.

In the last 50 years several EM pumps for liquid metal have been developed for sodium-cooled nuclear reactors. They are all provided with an external electromagnet which is expensive and clumsy. The designers argue that magnetless pumps do not work. Eighty years ago Carl Hering (*J. AIEE*, vol. 42, p. 139, 1923) proved them wrong. He knew Ampère's action-at-a-distance law and pumped liquid metal without magnets in a hundred furnaces.

Science versus subjectivism

I have no quarrel with subjective assessments, when these are scientifically conducted and the limitations born of psychological, logistical and statistical factors are fully appreciated. I am concerned at the virulence of the virus of unscientific, indeed anti-science, subjectivism which is increasingly pursued with religious fervour in audio matters.

I am therefore concerned that the recent article "Better CD" is devoid of any quantifying data. It latches on to the latest "in" idea that everything is subject to RF pollution (riding on the back of public concern about pollution of a more general nature). Even if there was some RF pollution, would this necessarily be a bad thing? I would remind readers that RF "pollution" is essential in order to straighten out the non-linearities in analogue magnetic recording, and it is also quite good at dithering crossover distortion.

Some CD players may not bias the op-amps for zero DC at their outputs; so what! Who says this is a bad thing? As a point of fact, many op-amps perform better when biased away from "centre" — particularly the 741 family which have a nasty crossover region around zero. I might well bias op-amps in the middle myself to avoid use of electrolytic capacitors *only* because they

have a shorter life expectancy. But why use such an elaborate current source when a junction fet with source resistor would do just as well, and with far fewer components? And what is this mystique about strange-valued resistors in non-critical positions?

I am not at all surprised that a panel of self-styled "expert listeners" thought it all sounded better: they always do, and the rating of slam/focus/definition/rhythm/musicality and all that garbage is nearly always in direct proportion to the cost of the product/modification and in inverse proportion to the size of the company promoting it.

Is there any reason why meaningful *objective* measurements should not be presented together with sufficient methodological and statistical data on the subjective experiments, so that the reader can be sufficiently well informed to form his own judgment of their validity?

If you really want to hear the *difference* then why not listen to the difference, by use of a differential amplifier connected between the outputs of modified and unmodified back-ends, fed from a common dac? *Warning*: you might hear nothing!

David Birt
Blethingley
Surrey

Water-arc explosions are far more effective projectile accelerators than railguns. The US Army develops railguns, and recently said they do not understand how water-arc guns work and, therefore, they will not research them.

For 30 years it has been known that lightning strikes which cause forest fires produce hardly any thunder. The loudest thunder claps, associated with the largest lightning currents, are relatively cold. So what causes thunder? Since no suitable electrodynamic forces could be found in our textbooks, meteorologists lost

interest in the subject.

Over the same period of time more than a billion dollars have been spent worldwide on hot nuclear fusion. We are still waiting for a demonstration of ignition! Could it be that the same electrodynamic forces which are responsible for thunder make it so difficult to contain the plasma in Tokamak fusion reactors?

Peter Graneau
Centre for Electromagnetics
Research
Northeastern University
Boston
USA. ■

If the SAS slogan "Who dares wins" applies to scientists then we would expect some fireworks from a budding chemist. If not literally fireworks, then some strong stinks at least. For example, Irving Langmuir, who was to receive a Nobel Prize for chemistry, did not do things by halves. At the tender age of six, he nearly killed himself by sniffing chlorine.

It happened when Arthur, Irving's older brother by nine years, had access to a well-appointed chemistry laboratory. Probably not realising how dangerous chlorine is, he took to sometimes carrying a small bottle of the gas around with him and sniffing it. Taking the bottle home he offered Irving a smell. Irving, "never a cautious type," took a gulp and "nearly strangled on the spot." Several anxious choking days passed before he could breathe properly again.

Father banned further chemistry experiments, except in the classroom. Undeterred, Arthur grew up to become a successful industrial chemist and Irving gained an international reputation for both chemistry and physics.

Irving Langmuir was one of the first immensely successful industrial scientists. He was fortunate to spend most of his working life at the General Electric (G.E.) Research Laboratories at Schenectady in New York, where industrial research was expected not only to yield practical results but to expand knowledge as well. His early successes in improving incandescent lamps must have paid his salary for the rest of his career, and he was left free to let his imagination run riot — to the enormous benefit of electronics and science in general.

Langmuir's career path was set by his choice of postgraduate study. After graduating from Columbia University in 1903 with a degree in metallurgical engineering he left the USA for Germany where he enrolled for a PhD at the University of Göttingen. There he studied physical chemistry under Prof. Walther Nernst, inventor of the Nernst lamp, who suggested that Langmuir should examine the dissociation of water vapour and carbon dioxide around glowing platinum wires. These

PIONEERS

Irving Langmuir
(1881-1957)

"World's Foremost
Scientist"

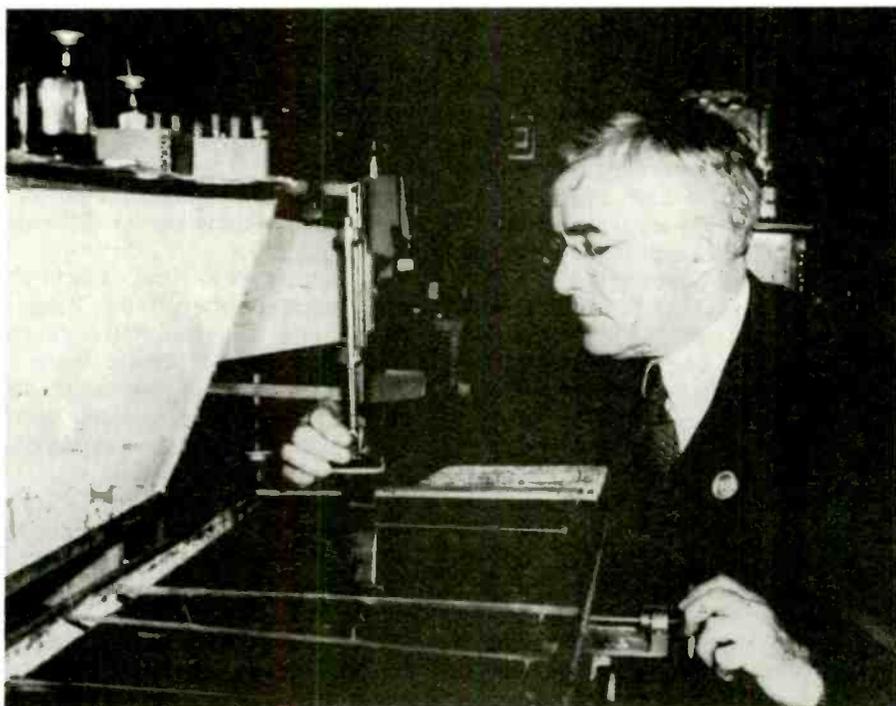
W. A. Atherton

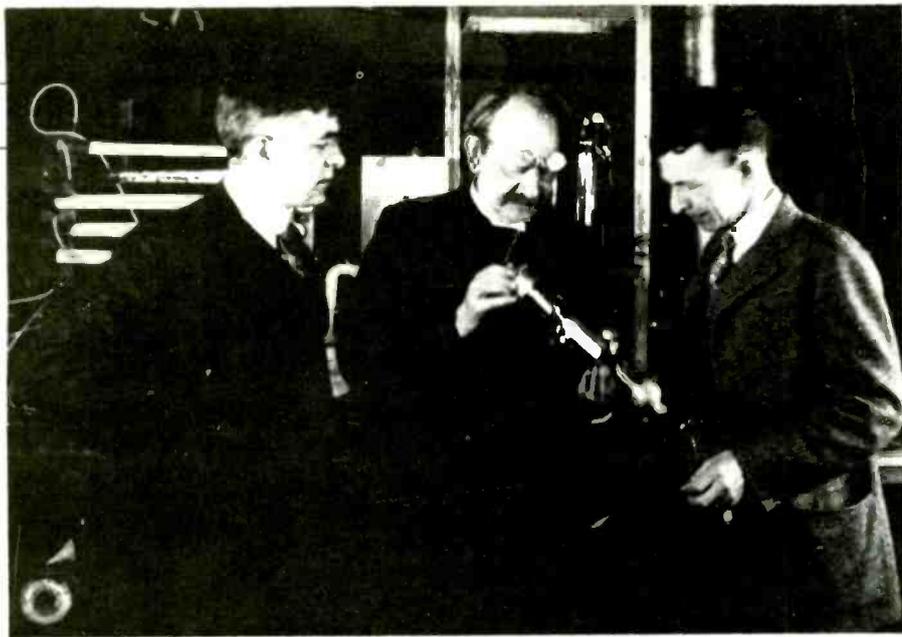
Langmuir with film balance and trough.

studies not only gave Langmuir his PhD in 1906 but the basic techniques which were to flower later at the G.E. laboratories.

On returning to the USA he lectured at the Stevens Institute of Technology in Hoboken, New Jersey. In the summer of 1909 he took vacation employment at the G.E. Research Laboratories. When due to leave he was asked to stay — higher education thus lost a teacher but industrial research gained a genius.

His first task was to study thermal conduction and convection in gases at high temperatures. For this the new tungsten filaments, developed by W. D. Coolidge from an Austrian invention, were used. At first Langmuir felt embarrassed at the amount of time he enjoyed "playing" but discovering nothing, but the director of the laboratories, W. R. Whitney, reassured and encouraged him.





Langmuir (left) and W. D. Coolidge showing the pliotron to Sir J. J. Thomson when he visited the labs in 1923.

Then Langmuir did discover something — namely that hydrogen gas behaves in a very peculiar way at extreme temperatures. In 1912 he explained this as the dissociation of hydrogen molecules into single atoms. It was an important scientific discovery which he used much later in his invention of the hydrogen torch, for which a patent was awarded in 1934.

One of the problems with the early tungsten filament light bulbs was that the inside of the glass blackened with time. It was widely held that a better vacuum would solve this and other problems. Langmuir could not think of any way to get an empty void emptier so he turned the problem inside out by adding extra gas so he could, once again, study the behaviour of gases.

His approach has been likened to treating a case of poisoning by giving more poison. To Langmuir it made sense because he could extrapolate back from his results to predict what would happen if a perfect vacuum was obtained. What he showed was that even the minutest amount of water vapour must be avoided. Beyond that, however, a better vacuum did not lessen the blackening, which was caused by the evaporation of tungsten. Instead of further improvements to the vacuum he suggested the opposite — filling the bulb with gas. "You're dreaming!" Whitney told him, but allowed that he should try it. The result was today's argon-filled light bulb, a money spinner if ever there was.

His work on light bulbs inevitably led him to its closest cousin, the thermionic valve. The tungsten filament of the lamp became the early cathode of the valve and every physicist wanted to improve the emission of electrons from its surface. At the same time as Langmuir was improving the incandescent lamp, others were inventing new applications for radio valves.

When he studied the electron emission from tungsten filaments he found that, at temperatures close to the melting point of tungsten, the emission was orders of magnitude less than predicted by theory. Tracking down the cause of this mystery led him to the discovery of electron space charge, the results of which he described in a law. This recognition and theoretical understanding of space charge was one of the turning points in the development of the scientific design of thermionic devices.

Further studies led to the discovery that the addition of a tiny amount of thoria to the tungsten filament gave an enormous increase in the electron emission. Originally thorium oxide had been added in an attempt to prevent crystal growth of the tungsten metal, but Langmuir's detective work showed that a layer of thoria formed on the tungsten surface, resulting in the vast increase in electron emission. Thoriated tungsten filaments became standard, and the G. E. Labs at Schenectady further confirmed their international importance to what would become known as the electronics industry.

Recognition of his work came in 1920 when he was awarded the Rumford Medal of the American Association for the Advancement of Science. It was not the first, and certainly was not the last, honour he was to receive. On a visit to Japan in the mid 1930s he was described as the world's foremost scientist.

Another way in which Langmuir influenced the design of electronic valves was through his improvements to vacuum pumps. In 1915 he examined Gaede's diffusion pump and, as was his style, gained both theoretical understanding and practical improvements. After the First World War Lang-

muir's mercury diffusion pump became a familiar sight in laboratories.

His work covered an incredible range of interests and he had an international reputation in subjects which we cannot even touch on here. It is no surprise to learn that he pioneered the artificial production of rainfall and snowfall.

Early in the war he had spent time studying the formation of clouds and icing. This required measurements of particle size using Rayleigh's law of light scattering. But how do you turn that into a tool to help win a war? Langmuir's answer was to use it to make far bigger and better smoke screens. The "white" smoke which he and his colleague, V. J. Schaefer, produced became an artificial fog which protected troops during the invasion of Europe. For three days during the crossing of the Rhine a 65-mile front was blanketed with the fog. According to Langmuir's biographer, Rosenfeld¹, its introduction into the Pacific by the US Navy ended the destruction caused by kamikaze attacks.

The story of the invention of this smoke screen began with a new filter for gas masks, which then needed a new smoke generator to produce smoke to test the filters. Later, when a smoke screen was needed, Langmuir and Schaefer were partway there. On 23 June, 1942, Army brass hats were dragged to the top of a suitable mountain before dawn to witness in the valley below a perfect demonstration of how to generate vast amounts of clean white smoke.

Another problem, how to prevent aircraft wings icing up, led to post-war experiments with cloud seeding. It was on 13 November, 1946, that Schaefer first seeded a cloud, using six pounds of dry ice. Langmuir reported that, "The whole supercooled cloud was converted to small ice crystals." In his diary he simply wrote, "It worked!"

Experimental results improved and on a winter's day that year they turned a cloud into snow. But the snow became a storm and the storm a blizzard, and the storm within G. E. over the legal implications led to the company waiving all royalty rights to the patents, and for the first time worrying about what Langmuir was up to.

By the time of his death at Falmouth, Massachusetts, on 16 August, 1957, Irving Langmuir's interests had spanned a range of science from electronic vacuum tubes to changing the weather. He was born in Brooklyn, New York, on 31 January, 1881, the third of four sons of Charles and Sadie Langmuir. His grandfather had emigrated from Glasgow to

Canada and his father from Canada to New York, where Irving was educated at public schools in Brooklyn. He belonged to a big family; besides three brothers he had an abundant supply of aunts, uncles and cousins.

In 1892 a trip by his parents to Europe resulted in his father's appointment as the Paris representative of the New York Life Insurance Co. The four boys were shipped out to Europe and met at Le Havre by their father. Irving continued his schooling in Paris for the next three years. By that stage his love for the outdoor life, especially hiking, was well established, as was all the boys' love for dramatic chemistry experiments — chlorine excluded. Irving recalled that he had a workshop when he was nine and his own laboratory when he was 12.

He completed his schooling back in the USA, claiming that, "Until I was 14 I always hated school and did poorly at it." After that it was Columbia University and on to Göttingen.

Hiking was by no means his only athletic interest. He also climbed, hence his eagerness to drag the Army chiefs to the top of a mountain to witness the smoke screen demonstration. Probably his

greatest climbing feat was to conquer the Matterhorn in 1921, when he was 40 years old.

Skiing was also a favourite activity, and it is said that he introduced the sport to Göttingen. In 1910, according to his biography, he did a lot of canoeing, camping and skating, took lessons in tennis, bowls, horse riding and dancing, went sleighing in the moonlight, swam, hunted, fished, and became a scout-master. His spare time was spent in the darkroom for he was a keen photographer. He also became interested in that new gadget, the motor car.

In his younger days, it is said that he was uncomfortable with the ladies. His diary for 29 June, 1910, records him meeting and dancing twice with a Miss Mersereau. In time he came to use her first name, Marion. They were married in 1912, forming a long and happy union in which they had two children, Kenneth and Barbara.

His wife also gave him a wonderful epitaph: "He was as great a husband and father as he was a scientist." ■

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Langmuir receiving the Nobel Prize in 1932.

Tony Atherton's book, *From Compass to Computer: a history of electrical and electronic engineering*, was published in 1984 by MacMillan



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Not all gyros need a spinning mass. John Nuttall points out that those using optical sensing possess unique advantages which, in some applications, outweigh their high cost

Optical-fibre gyroscopes

Many devices detect rotation: shaft encoders and resolvers, for example, detect rotation relative to a fixed reference. But a gyro is different, because it needs no such reference – it measures absolute rotation. Today, the word is used for any sensor of absolute rotation, even if there is no spinning wheel. In technological applications, gyros are the sensors, acting as an integral part of a measurement or control system.

A typical application for a gyroscope is the stabilisation of an aerial on a moving object. Here, the aerial is mounted on a gimbal system giving two degrees of freedom, the gimbals being controlled to give zero rotation of the aerial. Similarly, an autopilot system requires gyroscopes to detect the aircraft's rotation.

At a more complex level, gyroscopes form the heart of inertial navigation and inertial surveying systems. These aim to measure the motion of a body entirely from within the body (i.e. without the use of radar, beacons or landmarks). Displacement and velocity cannot be measured directly, so the displacement has to be obtained by double integration of acceleration measurements, which are collected in three axes. Of course, one needs to control (or at least know) the direction in which the three accelerometers are pointing, and thus three gyroscopes are necessary.

Optical gyros

The majority of gyroscopes in use today are mechanical instruments, based on spinning masses. This is true for both the least demanding applications (such as some stabilisation systems) and the most demanding applications (such as inertial navigation). However, in recent years, optical methods of rotation sens-

ing have become practical, and there are perceived advantages in cost, reliability and performance. How real these perceived advantages are, will be discussed later in this article.

First on the scene was the ring laser gyro and most gyroscope companies now make an RLG, usually intended for aircraft navigation. More recently, there has been feverish development of fibre-optic gyros. These appear in a number of forms, with confusing sets of initials to describe the variants. In this article, I will describe the original best-known interferometric fibre-optic gyro, which I call the FOG.

The Sagnac interferometer

Optical rotation sensing is based on the Sagnac effect, which was first demonstrated as early as 1912. In the Sagnac interferometer, there are three mirrors and a beam splitter (Fig. 1). (An ideal beam splitter reflects half the light incident on it like a mirror, and transmits

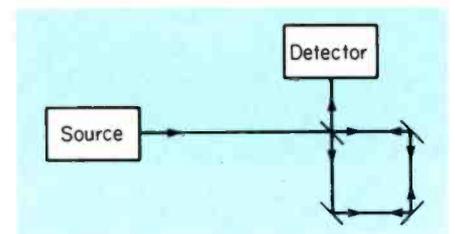


Fig. 1. Sagnac interferometer using mirrors and a beam splitter. Light is sent both ways round the circuit, in which any non-reciprocities show up as a phase shift between the two beams.

the other half like a window.) Light is introduced externally at the beam splitter, and consequently goes both ways round the circuit, and back to the beam splitter. Here some light is reflected back to the light source and some



Ferranti fibre-optic gyroscope, with a selection of components. Small rectangular component is a coupler.



Demonstration model of FOG. Source and detector are in separate die-cast boxes, with coil below. Three polarisation controllers and an integrated-optic phase modulator (top) are shown.

The author with an experimental gyro.

is incident on the detector. It is clear that, if care is taken to ensure that the two beams of light follow exactly the same path, the lengths of the two journeys will be exactly the same. They will therefore interfere constructively at the detector.

There are some circumstances in which it is possible to make the lengths of the counter-rotating journeys different. Such effects are called *non-reciprocities*, and the one we are interested in is the rotation-sensitive or Sagnac effect. The effect of rotation on the passage of light can properly be explained only in terms of General Relativity, but the Sagnac effect can be visualised in a simple way. If the light travels round in the opposite direction to the rotation, the distance the light has to travel seems less than the distance when there is no rotation. On the other hand, if the light is travelling with the rotation, the distance seems greater. The two beams will reach the detector having travelled different distances and will therefore be out of phase with each other.

Using this model, one can calculate the phase shift as

$$\Delta\Phi = 8\pi A\omega / c\lambda$$

where

A = the area enclosed by the light path
 ω = the applied rotation rate (radians/second)

c = the velocity of light

λ = the wavelength of the light (*in vacuo*)

It is usual to enhance the effect by using a long length of optical fibre wound into

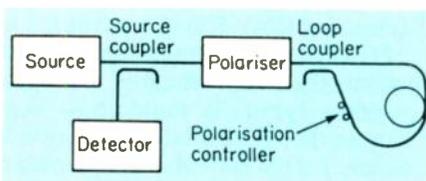


Fig. 2. Optical components of a fibre-optic gyro. The couplers are the fibre equivalent of the beam splitter. Polariser and polarisation control actuator are needed so that only one polarisation of light is used. This configuration is the minimum reciprocal gyro.

a coil of many turns instead of one 'turn' of free space (Fig. 2). Assuming a circular coil of total length L and radius R, the phase shift is

$$\Delta\Phi = 4\pi LR\omega / \lambda c$$

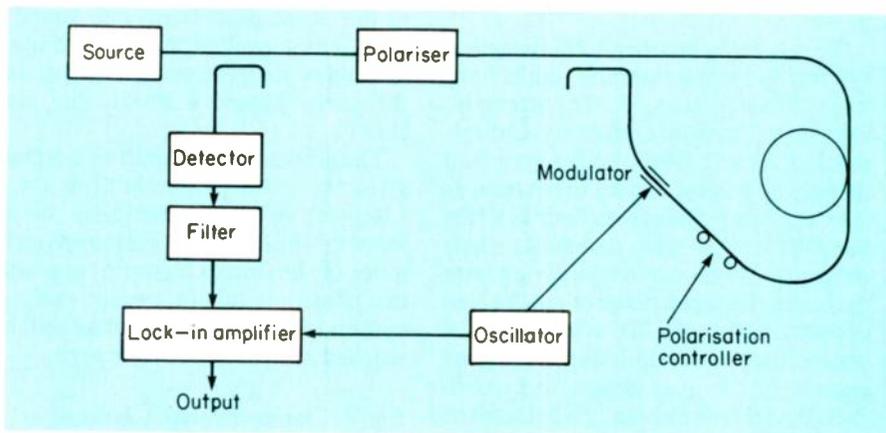


Fig. 3. Open-loop fibre-optic gyro. Detector output is phase-sensitively demodulated at the modulator frequency; output is proportional to rotation rate.

Of course, the phase shift between two beams of light cannot be measured directly. All that can be measured is the intensity of light incident on a photodetector. Hence, for example, a 2π phase shift cannot be distinguished from no phase shift: the amplitude is bound to be some \cos^2 function of rotation rate and therefore multiple-valued. This is not an encouraging start to instrument design.

Also, even after magnification by the number of turns, the Sagnac effect is uncomfortably small. Taking $L = 500\text{m}$, $R = 50\text{mm}$, $c = 3 \times 10^8\text{m/s}$ and $\lambda = 1.3 \times 10^{-6}\text{m}$, gives $\Delta\Phi$ (in radians) $= 0.8\omega$ (in radians/s). For the rotation rate of the earth on its axis ($15^\circ/\text{hour}$), $\Delta\Phi = 6 \times 10^{-4}$ radians, while for $1000^\circ/\text{s}$, $\Delta\Phi \approx 14$ radians, or more than two optical wave lengths.

On the standards of existing instruments, the ability to detect the earth rotating on its axis is not particularly demanding. Thus, the designer of a fibre-optic gyro has three problems to overcome: he has to produce a detection system capable of measuring the small phase shifts; he has to ensure that other sources of non-reciprocity do not drown the effect of slow rotations; and he has

to cope with the potential problem of the gyro output being ambiguous because of the \cos^2 effect.

Measuring the phase shift

The trick required to measure small phase shifts successfully involves the use of a phase modulator inside the sensing loop. It is essential that this component is not in the centre of the loop, but near one end. A modulator can be made by using a piezoelectric device to stretch the fibre in response to an applied voltage, increasing the phase length of that portion of the fibre and hence the time for light to travel through it.

Light which reaches the detector consists of two beams which have passed round the coil in different directions. At any instant in time, the light leaving the coil in the two directions passes through the modulator at different times (separated by the transit time of the loop). Thus if the modulator is fed from an

oscillator, the two beams of light have been through the modulator at times when its effective length was different and the phases of the two beams have been altered by different amounts. The maximum effect is obtained when the modulator period is chosen to be twice the transit time of the loop.

Analysis of this system needs several pages of mathematics but, in short, it reveals that for a modulator angular frequency of ω there will be a large signal from the detector at 2ω , but no signal at ω when there is no non-reciprocal effect in the loop. In the presence of rotation, however, a small signal at ω appears, proportional to the rotation rate. Isolating and measuring this signal is carried out by conventional phase sensitive detection, as illustrated in Fig. 3.

The above is an open-loop detection system, and some stunning results have been obtained using it. For example, workers at Stanford University, California, in the early 1980s, built a gyro with noise level corresponding to rotation as slow as a few millidegrees/hour in a 1Hz bandwidth. On this parameter their device was as good as a typical ring laser gyro, and not much poorer than the best of mechanical gyros. To achieve this, of course, they had to optimise many other aspects of the gyro design too. Also, their device was not really a detector of rotation, but rather a detector of the absence of rotation which had limited commercial possibilities.

Closing the loop

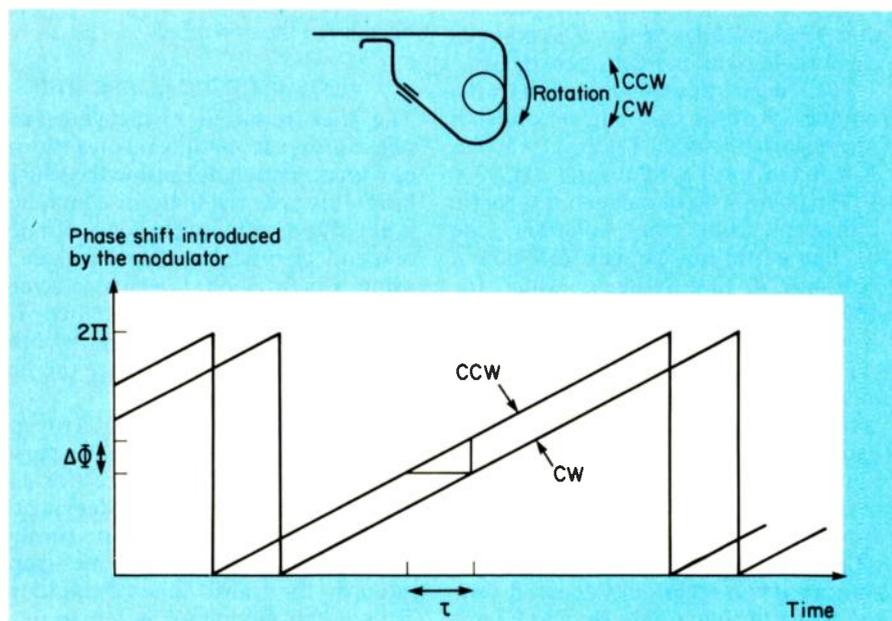
Successful as open-loop detection systems are at giving low noise and hence high sensitivity, a closed-loop

system is needed to get respectable linearity of the sensor. Such an approach is very common in sensor design. The idea is to counteract the rotation-induced non-reciprocity with another non-reciprocity. There are a number of possible 'real', physical non-reciprocities to choose from (e.g. the Faraday effect), but the greatest success has been obtained with an artificial effect, again based on the modulator.

Ideally, one wishes to increase the path length for light passing in one direction, while decreasing it for light passing in the other. That is not possible, but we can again exploit the fact that the two components of the light arriving at the detector at any instant have passed through the modulator at different times. Thus, if a steady ramp is applied to the phase modulator, one beam always sees a smaller phase length in the modulator than the other, and this does the trick. Figure 4 shows this more clearly.

The gradient of the ramp is such that it gives the extra phase shift $\Delta\Phi (= 4\pi LR\omega/\omega c)$ in the transit time of the loop $\tau (= nL/c)$, where n is the refractive index of the fibre. Of course, in reality, the phase modulator cannot carry on indefinitely, but once the phase shift has reached 2π , it can be reset to zero.

Fig. 4. Closing the loop. Clockwise-rotating light passes through the modulator later than counter-clockwise-rotating light by a time equal to the transit time of the loop (τ). It therefore sees a phase shift bigger by $\Delta\Phi$. The gradient of the ramp is controlled so that $\Delta\Phi$ counteracts the rotation-induced phase shifts.



The time between resets is $\lambda n/2R\omega$, and during this time a rotation of $\lambda n/2R$ radians has occurred. This means that each reset of the phase modulator corresponds to a rotation through a certain angle and, by counting the resets, the angle can be found. Thus the closed-loop FOG is an *angle-measuring instrument*, in contrast to the open-loop FOG which is basically an *angular-rate-measuring device*.

Integrated optics

In the case of the gyro in which 500m of fibre is wound into a coil of 50mm radius, the formulae above give a reset every 1.95×10^{-5} radians or four arc-seconds! This illustrates the amazing sensitivities that are considered normal in the gyro business. If the gyro is rotating at 1000°/s, the reset rate is roughly 900kHz. Of course, modern electronics can count at this rate, but it is not realistic to drive a fibre-stretching phase modulator at anything like this speed.

The technology that can provide phase shift with the required bandwidth is integrated optics — the optical analogue of integrated circuits. Then all that is needed is a source, a detector, a coil and the IO chip. All the other optical components — and there are more than I have mentioned in this article — can be built onto the IO chip (Fig. 5).

Integrated optics is a new technology, and it might be that the gyro is its first commercial application.

Other non-reciprocities

Because the rotation-induced non-reciprocity is so small, it is essential to minimise other sources of non-reciprocal phase shift. There are many such effects including the Faraday effect (magnetic fields) and the Kerr effect (electric fields). Effects such as the rate of change of temperature gradient also make their contribution. A special winding system is required — sometimes called 'anti-Shupe' winding in honour(?) of the man who first pointed out the difficulties caused by rate of change of temperature gradient. Careful choice of fibres, sources and other components, and careful engineering of the assembly are all needed to minimise these effects.

Why develop a FOG?

The fibre-optic gyro has caught the imagination of many people throughout the world, and traditional gyro companies have been amazed by the increased interest in rotation sensing by universities and by other companies. The glamour of the research is easy to see; it involves

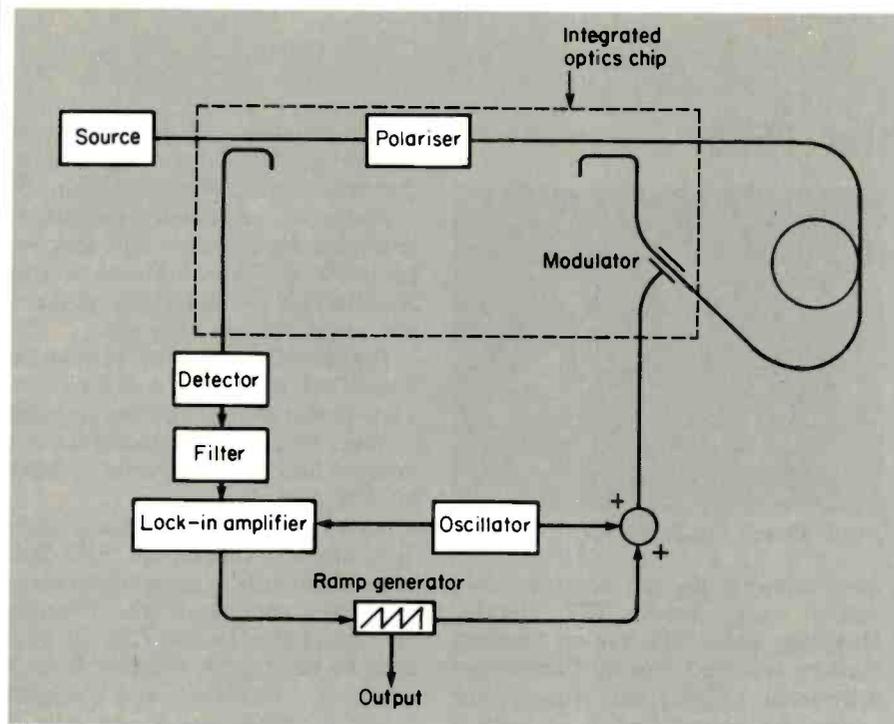


Fig. 5. Closed-loop fibre-optic gyroscope. Optical components are integrated onto a single chip. No polarisation controller is needed if special polarisation-maintaining fibre is used.

you were to try to put together a laboratory model of a FOG today (using commercially available fibre, sources, detectors, and an IO chip of existing design) aimed at a sensitivity of $10^\circ/\text{hr}$, you could spend £10 000 to £20 000 on optical components before assembly. Of course, 'once the demand rises, costs will fall'. However, the market for gyros is not large when compared with, for example, the telecommunications market, and development teams have been waiting for the price cuts for several years. Unfortunately, mechanical gyros are good enough, or almost good enough, to meet many requirements. There is new competition too from simple vibratory gyros (such as the Ferranti PVG) which are suitable when low mass, low cost and ruggedness are more important than sensitivity.

My own belief is that the first significant applications for FOGs will come in an area where cost is not a prime consideration, but where there is a requirement for the FOG's unique combination of abilities.

The author is with Ferranti Defence Systems Ltd, Edinburgh.

BOOKS

SPC Digital Telephone Exchanges, by F. J. Redmill and A. R. Valdar. The SPC of the title refers to "stored-program control" of telephone systems, which has largely replaced relay logic control in modern telephony.

Part 1 deals with fundamentals, introducing telecomms networks and switching systems and the elements of SPC exchanges. In part 2, the principles of digital signal encoding, switching and signal handling in exchanges are described, with a look at the design and architecture of exchanges and the synchronisation of networks.

SPC software and systems form the main part of part 3, which enters into considerable detail in describing exchange control, including integrity and maintainability. Finally, aspects of network operation, planning and the implications of the introduction of integrated-service digital networks are considered in part 4.

pushing a new technology to its limits.

The majority of companies are not aiming to produce a FOG capable of competing with a ring laser gyro. (The Ferranti RLG, for example, can detect less than $0.01^\circ/\text{hr}$, or one thousandth of the earth's rotation speed.) Instead, they are aiming to compete in the larger volume market of around 1 to $10^\circ/\text{hr}$, which is dominated by small, relatively cheap, rotating-mass gyros. Fibre gyros are considered to have a number of advantages over rotating-mass instruments. Personally, I believe some, but not all, of these claims.

The FOG offers a *high maximum angular rate* capability. For example, agile missiles roll at $1000^\circ/\text{s}$ or more, and few mechanical gyros can reach that speed, since a mechanical gyro has to be torqued to follow the vehicle's rotation. The FOG does not mind high rates.

The FOG offers *angle-measuring* rather than rate measuring. If you use the cheaper mechanical gyros you have to derive the angle by measuring rate and integrating, which is an inherently poorer method than measuring angle directly.

A further potential advantage of the FOG is its *dormancy*. This word is used to describe its ability to be stored unused for a long period of time, and to work reliably when required. The FOG is thought to have good dormancy, because there are no obvious deterioration mechanisms.

A FOG has *near zero switch-on time*,

in contrast to mechanical gyros which take some time for the wheel to come to speed. For some applications, this is very important.

It is often said that the FOG can be designed to operate at *extremes of temperature*. Mechanical gyros always have lubrication problems at both high and low temperatures, but the FOG does not have such restrictions. It is claimed that measurements could be taken during the drilling of an oil well, when the gyro would have to survive long periods at 125°C or even 175°C . However, the high-temperature FOG seems some way off, since many of the problems with light sources and other components in such hostile environments are, as yet, unsolved.

Similarly, the FOG shows promise in surviving *severe vibration and shock* which worry the mechanical gyro engineer because of brinelling of bearings and the effect of resonances. It seems likely that a FOG would survive harsh treatment and work well afterwards, but it is not obvious that it will work through it!

It has been claimed that the FOG has the advantage in *size and weight*, but someone was probably joking. FOGs are remarkably large compared with mechanical gyros of similar sensitivity. Some companies are working on very small FOGs, but, since the sensitivity depends on the radius, there are problems here.

Cost is the biggest problem of all. If

5 1/4in disk drive for the PS/2

IBM PS/2 computers do not take standard IBM PC 5 1/4in diskettes. This interface circuit allows a conventional 5 1/4in disk drive to be connected in place of the second 3 1/2in drive on a PS/2. It does not occupy a bus expansion slot.

The problem is level conversion. The disk drives are connected in parallel along a bus that comes out of the disk controller. On the PS/2, this bus uses c-mos tristate buffers for output and c-mos gates for input. The PC uses TTL-level inputs and open-collector outputs with 150Ω pull-up resistors (Fig. 1), the resistor pack being on the last drive in the chain.

Figure 2 shows the interface circuit. Outputs from the drive are pulled up by resistors R₃-R₆ and then shifted to c-mos levels by a 74HC373 tristate buffer. If the 74HC373 is unavailable, a 74C373 will probably do, and the newer 74HCT373 should be ideal.

Signals from the PS/2 to the drive

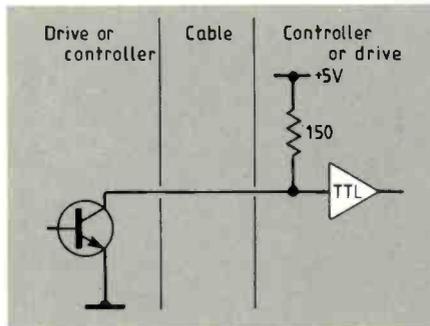


Fig. 1 (above). Fig. 2 (below).

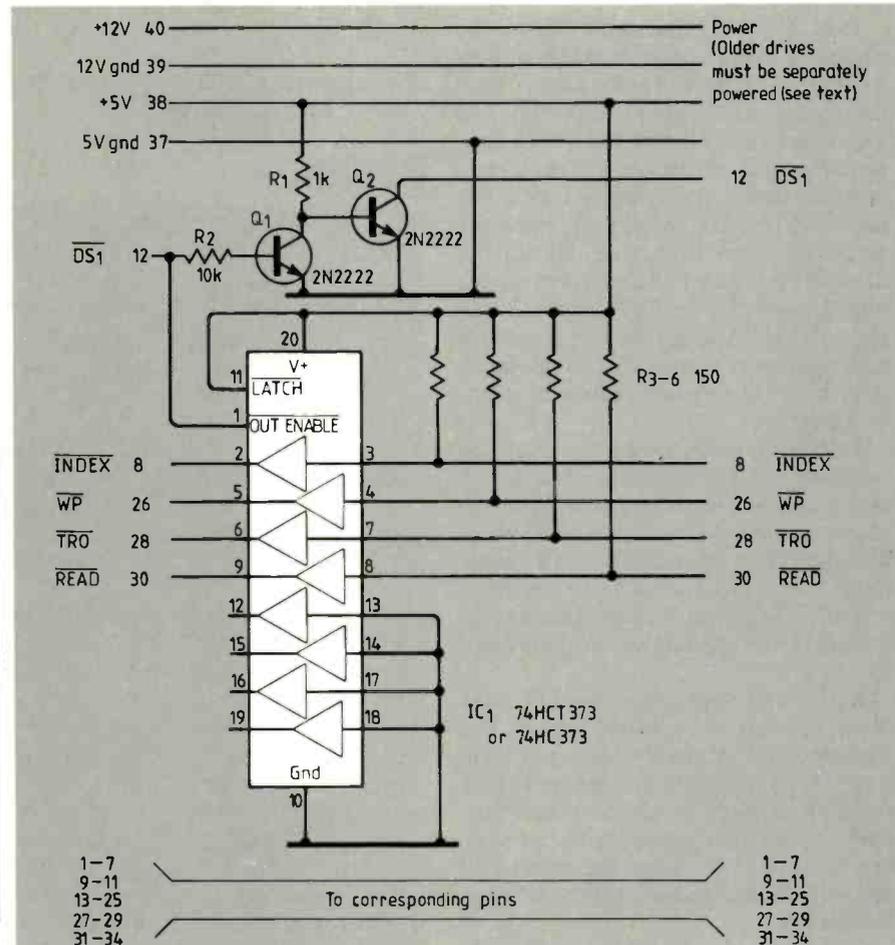
have an easier journey because c-mos outputs can drive TTL inputs. However, the 150Ω pull-up resistors must be removed. For all lines except drive select 1 (DS₁), this is easy - just unplug the terminating resistor pack from the disk drive. But DS₁ has its own terminating resistor soldered in place on each drive. Accordingly, Q₁ and Q₂ convert DS₁ from c-mos to open-collector output.

The two types of disk drives have almost identical edge connectors, the difference being that the 3 1/2in PS/2 drive has six additional positions, two grounded for isolation and four used for power. Of lines 1-34, the interface requires only five to be broken; the rest are wired straight through.

The interface gets its power from lines 37-40; so can the disk drive if it is a low-power model. Caution is in order because PS/2 documentation does not indicate how much power is available for the disk drive.

No special software is needed for the 5 1/4in drive, because the PS/2 BIOS already supports it; it is merely necessary to configure the machine appropriately. On the PS/2 50 this is done by booting the machine from the reference diskette and selecting "Change configuration" to edit the information stored in c-mos ram. The F6 key cycles through possible values for each item, and F10 stores the new information. If, on rebooting, the machine comes to a halt with '162' on the screen, this means the information in c-mos ram doesn't match the hardware found; pressing F1 will let boot-up continue.

The basic outline of this circuit was suggested by Clyde Washburn and is used with his permission. Michael Covington
University of Georgia
USA



Versatile temperature compensation

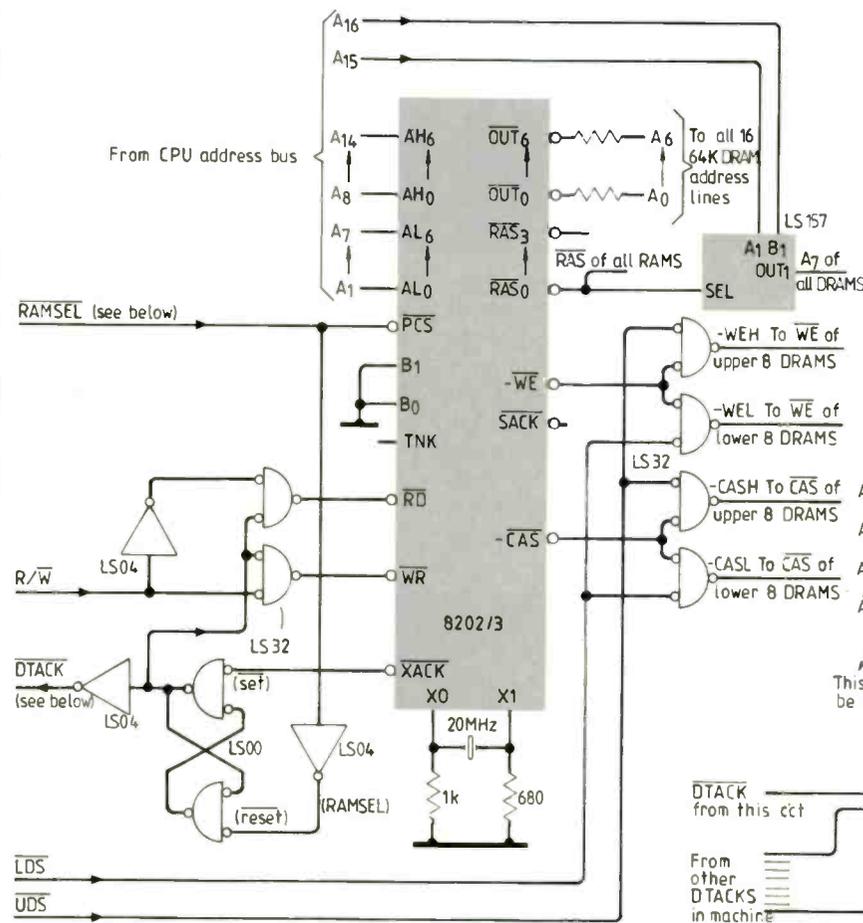
Positive and negative temperature coefficients may be corrected in many critical applications using this circuit.

IC₁ is a temperature sensitive device whose terminal voltage has a positive temperature coefficient of 10mV/K. By inverting this slope in IC₂, and summing varying proportions of the two slopes (R₇, IC₃), any linear correction can be applied to subsequent circuitry.

In the original application, the output was applied to a Varicap diode in the tuned circuit of a drifting RF oscillator.

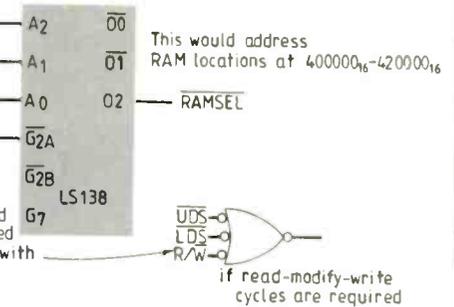
The zener diode D₁ exhibits a very low temperature coefficient at a standing current of 6.5mA, as do most zeners of this particular value. It is chosen to supply a standing current of 7.5mA (including 1mA for IC₁) from the supply rail, which is assumed to be already

Old 16K dram controller handles 64Kbit



A controller for 64Kbit drams can be made with this relatively cheap 16K dram controller, the Intel 8202, by multiplexing two extra address lines external to the chip. On-chip refreshing still works because the 8202 has a 7bit refresh counter, which is all that is needed by a 64K ram. Additional multiplexing is performed by a 74LS157 and all the other gates on the circuit are required to make it work with a 68000.

RAMSEL, generated by the computer's chip-select circuits, activates the 8202. It multiplexes the address lines A1 to A14 and generates all the control signals for a dram. The type of cycle it performs (read or write) is determined by \overline{RD} and \overline{WR} , \overline{WE} and \overline{CAS} and AND'ed



with \overline{UDS} and \overline{LDS} to generate \overline{WE} and \overline{CAS} signals for the upper and lower bank in the memory chip array. The external multiplexer uses \overline{RAS} as a selector and the multiplexed line is used as A7 on the drams.

When a read or write cycle is finished, \overline{XACK} is asserted for 25ns at 20MHz crystal speed, so it is held by an RS flip-flop until the 68000 recognises it (as \overline{DTACK}). After the 68000 has completed its cycle, it negates its control signals and \overline{RAMSEL} should go high, clearing the flip-flop, thus negating \overline{DTACK} .

Buffers should be used on the data bus and memory data lines, but I found it worked well without them. If they are used, they should be enabled by \overline{UDS} and \overline{RAMSEL} , or \overline{LDS} and \overline{RAMSEL} for low bank. Crystal frequency should be between 18.5 and 25MHz; the faster the better.

The circuit should also work using an 8203 with 256K rams, giving 512K, but I have not tried it.

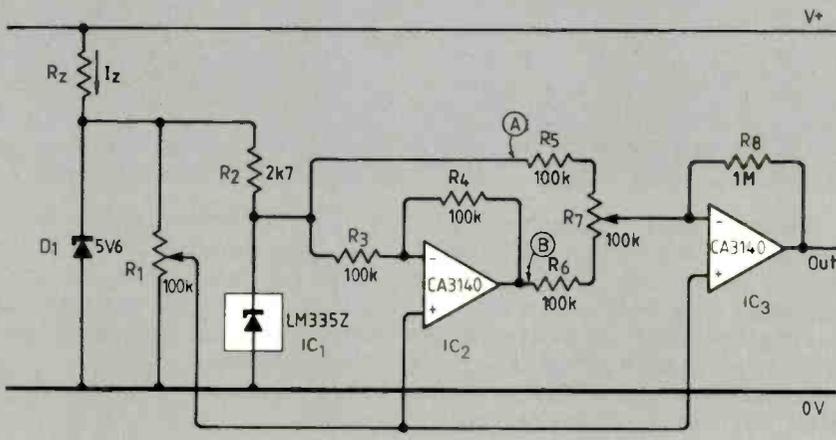
Mark Stephens
Wellington
New Zealand

stabilised and suitable for supplying IC₂ and IC₃, that is 10 to 30V.

In practice, the circuit is allowed to stabilise to normal operating temperature, and R₁ is adjusted to equalise the voltages at points A and B. The circuit is then altered to a different temperature

and R₇ adjusted to provide the desired degree of compensation. With the values shown, slopes from -50 to +50mV/K are possible.

Alan Diverall
Dorking
Surrey



Pulse producer stimulates muscles

This circuit produces pulses which are some multiple of the supply voltage. It was first used in battery-powered equipment to supply pulses for electrically stimulating muscles, but there must be other uses.

When the input is low, the capacitors charge in parallel through the diodes up to nearly the supply voltage. When it goes high, the capacitors discharge in series through the n-p-n transistors until the input goes low again to produce a negative pulse output (Fig. 1). A positive pulse could be produced by inverting the polarity of the whole circuit.

This is simple and more efficient than generating a high voltage and then switching it. It also gives a more precise voltage than a ringing choke.

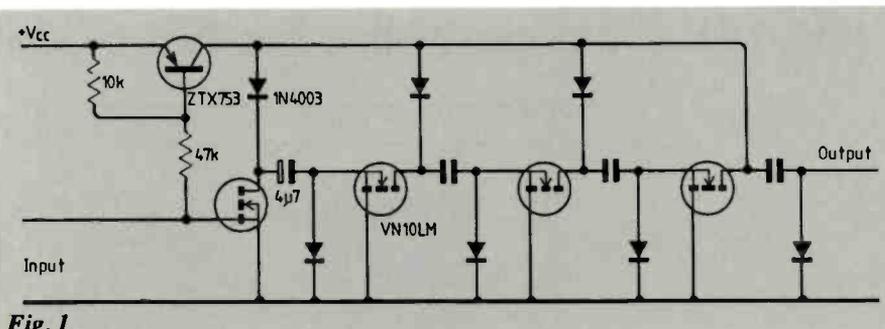


Fig. 1

The n-p-n transistors have no special requirements and transistor arrays such as CA3083 may be used if the substrate is left open-circuit. The p-n-p transistor must be able to stand the full output voltage. Inefficiency in the circuit is caused by the base-current and charge-storage effects in the transistors, as well as theoretically inevitable inefficiency in charging a capacitor from a voltage source.

Charge storage effects can be reduced using speed-up capacitors and Baker clamps, or by mosfets, as in Fig. 2. Mosfets also reduce the drop in output voltage caused by the base current, but care must be taken with regard to gate source voltage rating. S.A.G. Chandler
Engineering Department
University of Warwick
Coventry

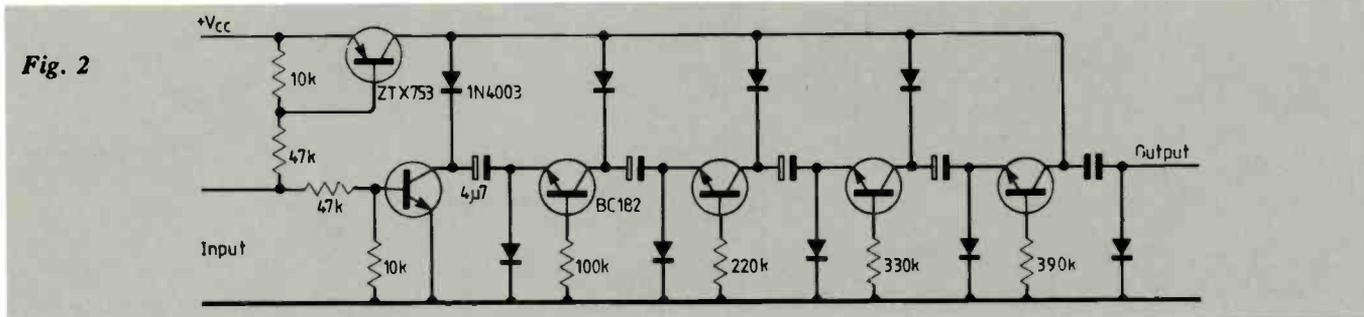


Fig. 2

Battery saving relay control

In the circuit of Fig. 1, the initial voltage 12V is supplied by capacitor C (220µF) for 1s, sufficient to turn the relay on. Then the series resistor R_s (330Ω), chosen to have about the same value as the relay coil resistance, decreases the initial current from 36 to 18mA. The delay between relay operations must be no less than 150ms.

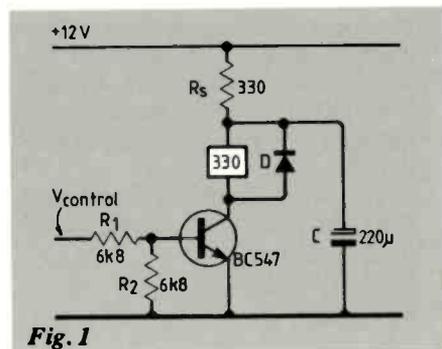


Fig. 1

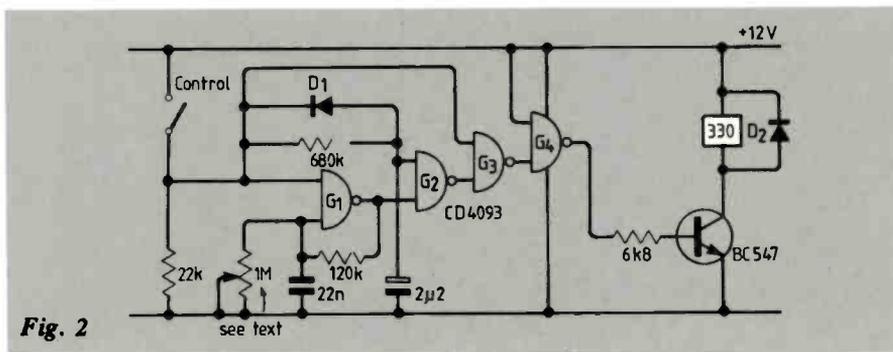


Fig. 2

Figure 2 gives a better solution in terms of power saving; the relay is turned on by unipolar pulses. The minimum duty cycle of the oscillator G1 could be found empirically with the 1MΩ trimmer for a specified relay. In this circuit, with a 390kΩ trimmer, total DC current is about a quarter (9mA) of the 36mA coil current.

Gate G2 introduces a 1s start delay to trigger the relay on at full current. Gates G3 and G4 turn off the output transistor when the control voltage is zero (SW open). Standby current is a few microamps. Kerim Fahme
Aleppo
Syria

The ordinary business type personal computer has become an accepted platform for data acquisition. A truly bewildering variety of signal acquisition and processing cards can now be hooked into the host computer backplane. The software industry, initially slow to respond, has begun to meet the demands of the systems integrators. The advances made in card design, reflected in their sophisticated data acquisition features, are now matched by an equally impressive range of software products which can be used to great effect to control and maximise the card's features. LabWindows, from National Instruments, is an example of a data acquisition software development environment. It greatly eases the writing of software for exercising control over I/O expansion cards or GPIB (IEEE-488) interface cards.

It is simply unacceptable to expect an engineer to write low level software before he can use commercially available data acquisition products. Some GPIB interface cards are quite unfriendly and a great deal of time can be spent specifying the parameters for all the call instructions. If the card is used for an automatic test equipment (ATE) application, a considerable amount of time can be spent in the low level programming: this is an inefficient use of engineering manpower resources. LabWindows' author, National Instruments, has produced a fully comprehensive software development system for instrument control and data acquisition in recognition of the problem.

When LabWindows is up and running on a PC, it presents the engineer with the Microsoft QuickBASIC or C program development shell. This is complemented with a high level set of mouse driven options, through the use of dialogue boxes with menus. An engineer who is familiar with Windows programming will feel instantly at home with the front end user shell of LabWindows.

Tedious low level GPIB commands are replaced with slider icons and control boxes (slider icons and control boxes are shown in Fig. 1, each representing an instrument function or para-

LabWindows developer's kit

The data acquisition and signal processing add-in card market for the IBM PC is currently booming. Allan Brown reviews a software development package which takes the effort out of code writing

meter). As the engineer requires the instrument to perform a set of functions, the respective sliders or control boxes are evoked. Each slider and control box generates a call instruction and these are compiled into a file structure which forms the control software.

LabWindows is a substantial program which is ideally suited to an 80386 machine although it will run sluggishly on XT's and everything else in between including MCA and EISA boxes. The basic product comes on twelve floppy disks. There are two extra disks with the Advanced Analysis utility which only runs on a 386-PC hosting an 80387 coprocessor.

Documentation extends to an awe inspiring total of seven ring-bound manuals which are, in general, very well written. The installation of the program is straightforward and it powers up to

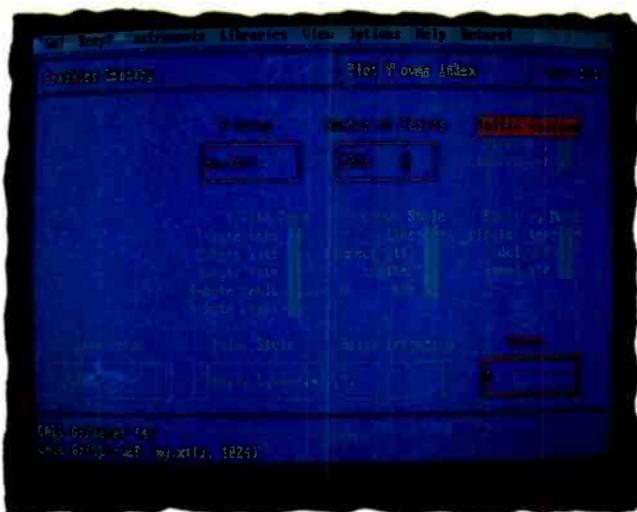


Fig. 1. Tedious low level GPIB commands are replaced by slider icons and control boxes.

```

File Edit Program Instruments Libraries View Options
PROGRAM
dim mx(1824)
for i=1 to 1824
  mx(i)=sin(.24*3.1415*(i))-cos(.85*3.1415*(i))
next i
CALL GfReset (4)
CALL GfYCaru2D (mx,1824)

```

865	866	867	868	869	870	871	872	873	874
881	882	883	884	885	886	887	888	889	890
897	898	899	900	901	902	903	904	905	906
912	913	914	915	916	917	918	919	920	921
928	929	930	931	932	933	934	935	936	937
945	946	947	948	949	950	951	952	953	954
961	962	963	964	965	966	967	968	969	970
977	978	979	980	981	982	983	984	985	986
993	994	995	996	997	998	999	1000	1001	1002
1009	1010	1011	1012	1013	1014	1015	1016	1017	1018

Fig. 2. The lines of code in the top window generate a pair of sinusoids. Numerical output automatically displays in the second window as the program executes.

give the high level language window environment for either QuickBASIC or C. National Instruments has been very thoughtful in presenting a 'Getting Started' manual which provides the engineer with a friendly introduction to LabWindows and an opportunity to get acquainted with many of its features. Extensive support has been given to a wide range of graphics adaptors and the figures presented in this review are of the standard VGA.

The essence of LabWindows is to allow the engineer to write a skeleton program without any input, output, graphics or analysis code included. These features are contained in extensive libraries and can be evoked from the menu options. For example, the first five lines of code shown in Fig. 2 generate a pair of sinusoids. Numeric outputs are shown in the second window which is automatically generated when the program is executed. To display the results graphically, the cursor is moved to the position in the code where the graphics routine is required and the GRAPHICS option is chosen from the LIBRARIES highlight in the command bar at the top of the screen. This leads to a graphics function tree and the engineer is confronted with various options. By choosing 'Plot Y over Index' the control panel with its slider icons and control boxes, as shown in Fig. 1, appears. The mouse is moved to each slider and control box to make the appropriate setting. As each slider or control box is activated, the corresponding low level code is generated at the base of the screen. To test the code, after setting variables, the GO! option in the command bar is activated. Plotting results are shown in Fig. 3. If satisfied, the KEEP! option places the two lines of generated code in the developer's source code. This code may be accessed through the

RETURN! option - as can be seen in Fig. 1. This method of code generation certainly takes the burden out of data acquisition programming and will almost certainly become a standard software design technique for such applications.

Much emphasis is given to the interactive nature of the software. The engineer is able to access instruments and control expansion cards from a high programming level with remarkable ease. This interactive feature is very attractive for program development purposes because it enables the engineer to test sections or modules of code on the real hardware. Users who are proficient in both QuickBASIC and C have the opportunity to write some modules in either syntax. There is an option in the program menu which allows a switch from one to the other. It is therefore possible to mix modules since the compiled modules are language independent.

Much work has gone into the context sensitive help files. This is an appealing feature and is considerably better than the blanket help file so often encoun-

tered. LabWindows also allows the engineer to amend software errors by using the debugging tools found in the Microsoft shell. These include single line stepping, breakpoints and variable watching as the program progresses. It contains all the editing features of the Microsoft shell such as cut and paste, search and copy. These are very useful development tools which require practice to gain a degree of proficiency. In fact LabWindows is such a comprehensive package that the engineer will require quite a lot of practice before he becomes a master of the product. General Instruments is aware of this fact and the initial learning curve, via the 'Getting Started' manual, is very gentle and quickly establishes confidence.

Each entry in the command bar at the top of the screen has an associated function tree. This is a hierarchical structure consisting of a set of dialogue boxes as shown in Fig. 4. The various options are therefore evoked by working through the function tree.

When the LIBRARIES option is activated from the command bar the engineer has the choice of evoking seven options from the dialogue box. These include FORMATTING & I/O GRAPHICS, ADVANCED ANALYSIS, DATA ACQUISITION, RS-232 and GPIB calls. As further options are chosen, more dialogue boxes are generated. The Formatting and I/O allows the engineer to read or write to files in a variety of formats, whether floating point or scientific with power ten exponents, etc. Control can also be exercised over display format.

Graphics

The GRAPHICS option is very well designed and lends itself, with remarkable ease, to a variety of plotting requirements. Not only singular plots as shown in Fig. 3 but also multiple plots as shown



Fig. 3. The plotted results from Fig. 2. If satisfied, the lines of program code may be added to a master file.

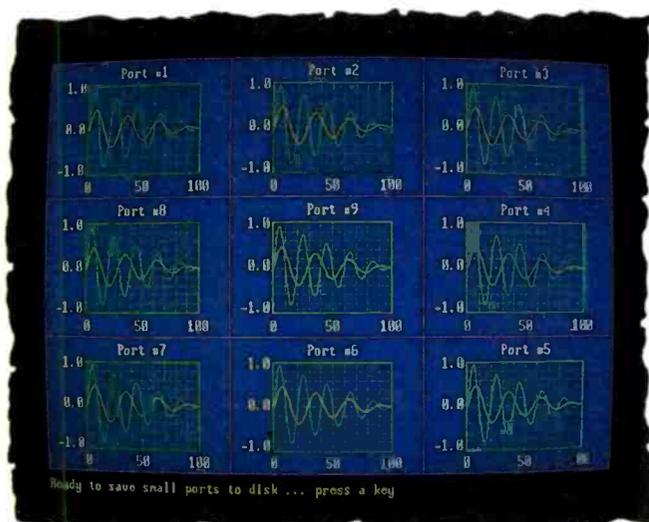


Fig. 4. LabWindows allows the display of multiple traces.

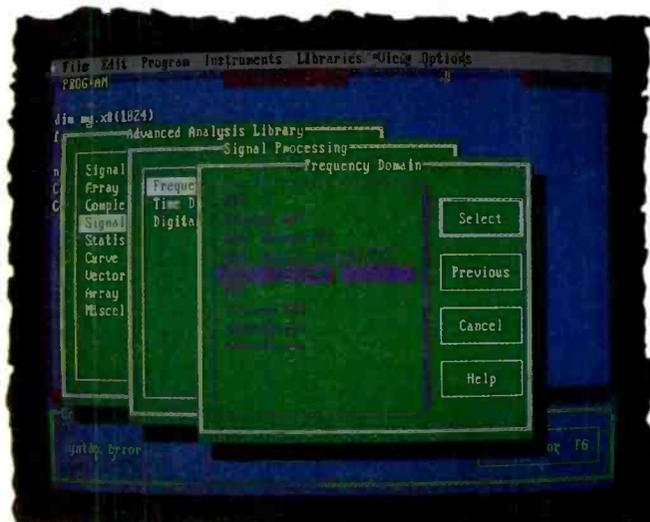


Fig. 5. Part of a function tree from the £895 Advanced Analysis library.

in Fig. 4. Data can also be displayed in scatter plots or even strip charts which is particularly convenient for representing real-time temperature measurements. Many I/O expansion cards have several analogue inputs and, by calling on the options from the graphics functions, code can be compiled to allow the input data from each channel to be displayed directly in the plotting areas – in effect a multiple screen oscilloscope.

Facilities are included to allow the engineer to add text to the graphs and amend the scales as required. Again these tasks are performed by adjusting sliders and control boxes from within the appropriate function panel which generate the corresponding call routines from the library. Hard copies of the graphics displays can be generated and LabWindows has support drivers for 160 different printing/plotting devices. The hard copy device can be connected to the PC either through the GPIB or an RS-232 serial port.

Advanced Analysis

If an application is likely to involve a large amount of data processing, then it would be advisable to add to LabWindows the Advanced Analysis utility. Despite its cost, it is a very impressive feature and greatly enhances the LabWindows product. The code within the Advanced Analysis library has been compiled specifically for the 32-bit Intel 80386 and its 80387 coprocessor. Part of the function tree for this option is shown in Fig. 5 with three dialogue boxes and contains a wide selection of statistical and signal processing operations. The execution time for the 3072 point FFT routine is very fast, even on a 20MHz 386-PC. I should like to see how the

same routine runs on a 33MHz 386-PC with the higher performance Cyrix CX-83D87 coprocessor. The Advanced Analysis library augments the standard library to provide an extensive range of routines. Within the Advanced Analysis library (Fig. 6), the routines can be broadly grouped into analysis, signal processing and statistics. The analysis group includes features for signal generation and a remarkable choice of array operations (1 & 2 dimension and complex). The signal processing section contains frequency and time domain options but, to my way of thinking, the selection is somewhat limited. But the inclusion of Butterworth and Chebyshev digital filter options is welcome.

Statistics is an important tool in the field of data acquisition. The package incorporates an extensive range of statistical processing options within the Advanced Analysis utility. These comprise the standard functions (mean, variance,

median, histogram, etc) and three curve fitting routines. In addition, there are routines for matrix operations — quite indispensable for tackling multi-variable inputs. The use of these library routines is straightforward since they can be generated from sliders and control boxes in their respective function panels.

Although many of these functions are available in other data analysis packages, the Advanced Analysis utility has the benefit of its processing speed. It is one of the few packages which run the 80386/80387 processor pair to its full potential.

Instrument functions

One of the exciting aspects of LabWindows is the ease with which it allows the engineer to control instruments connected to the PC via the GPIB bus or RS-232 bus. INSTRUMENT FUNCTION program files (or alternatively instrument drivers) are modules which reside in a

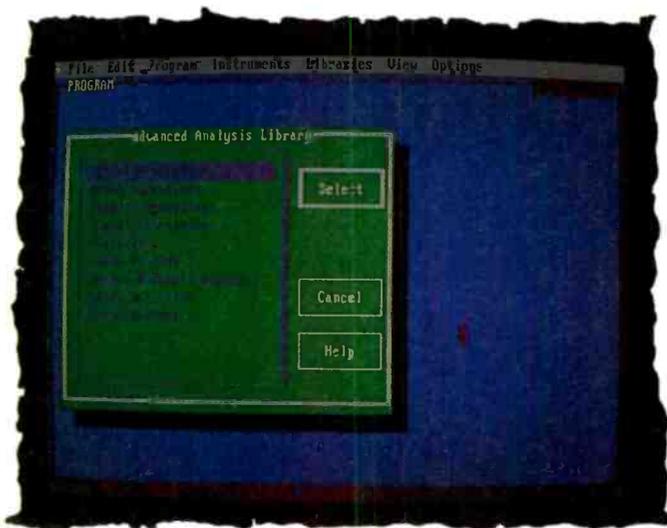


Fig. 6. Another pick list from the Advanced Analysis library showing a grouping into statistical and signal processing functions.

REVIEW

dedicated directory and are accessed from the INSTRUMENTS option in the command bar. When an instrument is loaded, its name is automatically entered into the INSTRUMENTS FUNCTION tree. Moving further down the tree reveals the function files as a set of options which are used to configure the instrument to perform specified tasks (Fig. 7). National Instruments supply an impressive list of instrument drivers for a variety of commercial products — scopes, frequency meters, frequency generators, etc — from a number of different manufacturers.

It may be necessary to create an instrument driver and this is quite an involved process. In fact a complete manual, 'Instrument Library Developer's Guide', is dedicated to the task. To set up an instrument file, a function tree must be created which will consist of a subset of dialogue boxes which generate the functions to be performed by the chosen instrument, similar to Fig. 5. To begin the design process, the FUNCTION TREE EDITOR is evoked from OPTIONS in the command bar, which generates the standard Microsoft shell editor. By choosing the CREATE option from the command bar a dialogue box appears where the instrument name is entered. Help files permeate the whole of LabWindows and the user, when designing an instrument driver, can add custom help files by evoking the HELP EDITOR. This results in the appearance of a help option in the dialogue box. In a similar manner, control panels with dialogue boxes and sliders can be created. Each time a new feature or option is introduced, its corresponding call routine is added to the FUNCTION MODULE file. The completed module will therefore consist of a list of call routines and the module can either be used interactively by the

Fig. 7. LabWindows comes with a large number of instrument drivers to smooth the interface between the software and commercially available instruments.



engineer or added to an executable file during the linking phase of the program development.

Program Compiler

The interactive nature of LabWindows allows the engineer to test each program module in realtime. By the time the program is complete, the engineer will have a high degree of confidence that the program as a whole will satisfy the requirements. The program modules can be compiled in the standard manner under the Microsoft shell using the appropriate compiler for either QuickBASIC or C. On a 383-PC, the Microsoft compilers work at an impressive rate; it is possible to construct a batch file to perform this task for several modules. This creates a number of object files which are language independent. The LWLINK utility is then used to link the LabWindows call routines with the engineer's created object files. When LWLINK is evoked, it generates a screen display as shown in Fig. 8. The created

object files are entered into the region in the top left of the screen and the appropriate libraries, from where the call routines are accessed and highlighted. After the linking operation is complete the engineer is left with a complete program in an EXE executable file format. Extensive use of the LabWindows libraries results in very large files.

DOS provides limited flexibility in program construction. It is customary with OS/2 and UNIX/386 to use a feature known as dynamic link libraries (DLLs). When the object modules are linked, the library routines are not included in the executable file. They are called from hard disk as required by the executable file, resulting in much smaller executable files. LabWindows would be better suited to the alternative operating systems and one may look forward to this in future versions.

In use

LabWindows is well thought out and it matches many of the requirements for developing data acquisition and ATE system software. I have very few reservations in recommending the package apart from program size and data file size. Since LabWindows is a comprehensive tool, it is very likely that many users will have major programming loads for it and I envisage the 640k DOS boundary creeping up very quickly. Before embarking on a large project I would feel more comfortable if there was some provision to include DOS extenders to eliminate the threat.

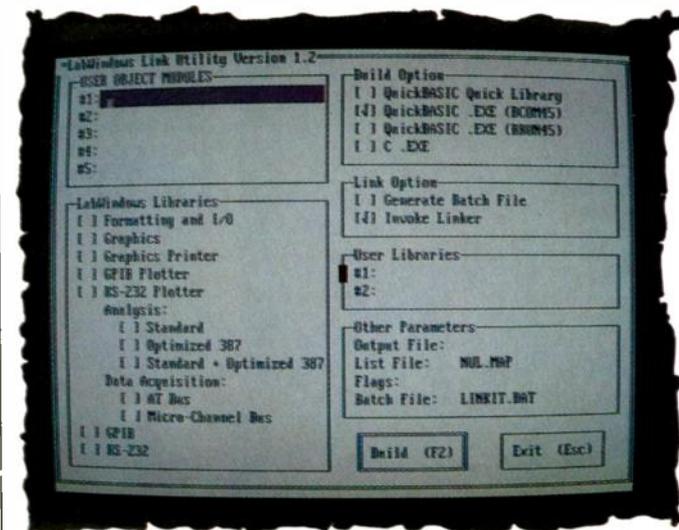


Fig. 8. LabWindows delivers executable code compiled from internal libraries and user generated code although the resulting .EXE file can be unwieldy.

**Supplier: National Instruments UK Ltd, 21 Kingfisher Court, Hambridge Road, Newbury, Berks RG14 5SJ. Phone 0635-523545.
Price: £495 basic package; £895 Advanced analysis module.**

Choosing and using programmable logic

Over the last ten years, the number of different types of programmable logic devices (PLDs) has rocketed. There are now roughly 3000 devices supporting over 300 distinct architectures under a dozen acronyms, including EPLDs, GALs, FPGAs and ASPICs.

The potential for confusion is enormous. Choosing the right part for the job can be a daunting task and getting objective help in decision making may not be easy: the market is intensely competitive, and every manufacturer has its own axe to grind.

The best chance of a disinterested evaluation will come from someone not connected with device manufacture but intimately involved with the devices themselves.

Stephen King is product manager for the Test Products Business Unit of Data I/O, a world class manufacturer of device programming equipment. Together with David Pashley, divisional manager of distributor Instrumatic, he outlined the current state of the PLD product maze and offered a way through it.

King describes the world of programmable logic in terms of four categories: simple or complex PLDs, field programmable gate arrays (FPGAs) and application-specific programmable logic (ASPL). MMI, now part of AMD, set the ball rolling with its programmable array logic (PAL) devices in 1977. PAL is now an MMI trademark, and most of the common acronyms are some company's legal property.

PAL was not the first family on the market; that was Signetics' field-programmable logic array (FPLA), launched in 1975. But where PAL made the running was not only in the introduction of the devices, but also in the first software design tools and its PAL design handbook. It was these three things that "kicked the industry off the ground", King said.

There were architectural differences too. FPLA included programmable

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AND and OR matrices. PAL's OR array was fixed, leaving only the AND array programmable. That made design easier, at the cost of reduced design flexibility.

MMI's design software allowed the engineer to enter designs in Boolean equations, which were automatically converted into a fuse map by the tool. Signetics required all this to be done manually, including logic equation reduction.

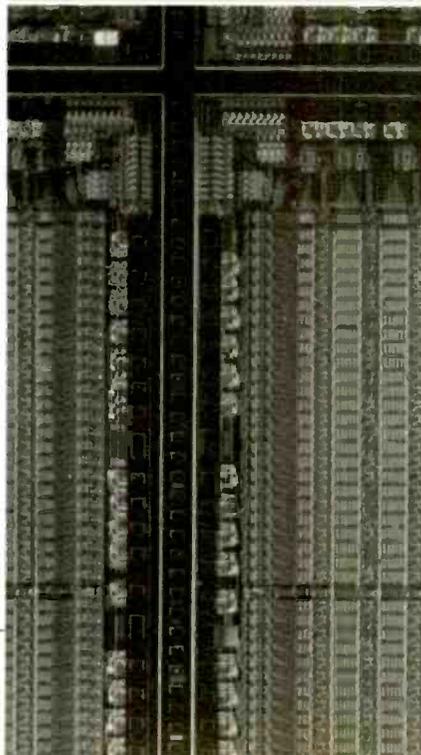
These types of simple PLD remain by far the most commonly used, and the cheapest. The 22V10 is currently the most popular device in the world; it costs 90 cents, the design tools are cheap or even free, and a design can be up and running in a day.

The next big step forward was a result of a change in technology from bipolar to c-mos. Because twice the circuit complexity can be fitted onto the same size c-mos die as bipolar, the functionality of the parts increased dramatically. That happened in 1985, led by companies such as Lattice, Altera and Cypress, and produced a marked growth in the use of programmable logic.

"The complexity of the devices has caused the PLD market to go absolutely crazy," Pashley said. "Whereas a few years ago people were using PLDs just as a substitute for some TTL packages to do glue logic functions, now it's very common to find whole boards populated solely with PLDs performing all the logic functions."

Erasable functions

One of the earliest families was Lattice's generic array logic, GAL, another trademark. These were intended as general purpose arrays, offering



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registered or combinational outputs, such that a single device could do any of the jobs of a number of the simple PLDs. One part fits all, as it were.

These were also the first erasable devices, including Altera's electrically erasable PLDs (EPLDs), which take half an hour under a UV lamp to wipe clean; and erasable PLDs (EPLDs), such as Lattice's GAL, which automatically erase with each reprogramming.

Typically, complex PLDs have a similar architecture to the simple ones: programmable AND array and output macrocells. But they offer greater flexibility, such as asynchronous operation and a choice of feedback paths, allowing an output pin to be tristated and used as an input, for example. This opens up a number of potential applications.

"You can build any function you

want," said King. "Maybe you want a counter: you can go off and buy a commercial counter, but if you want a short count, you have to add extra logic on your board.

"With a complex PLD, you can get the counter to count up to whatever value you want and quit there. You don't need any of the extra logic."

The disadvantage of c-mos is that it is slower than bipolar. Texas has recently launched a 5ns bipolar part, whereas the top speed of c-mos is around 15ns. C-mos is also slightly more expensive than bipolar.

According to Pashley, the price differential extends beyond the extra complexity on a gate-for-gate count. Whereas the simple 22V10 is equivalent to roughly 300-400 gates and costs under one dollar, the newest, biggest PLDs correspond to around 2000 gates

and cost \$50. This happens because nearly everybody makes the simpler types, whereas the more complex ones are proprietary. The differential is expected to reduce, however.

One result of the arrival of complex, c-mos PLDs was that the market blossomed from \$46 million worldwide in 1984 to nearly \$800m last year. Part of this growth is also due to the advent of the next step on the PLD ladder: field programmable gate arrays (FPGAs).

A new level of complexity

FPGAs employ a completely different architecture to PLDs. They have macrocells, but instead of being lined up on the output pins, they are distributed throughout the device and linked by a programmable interconnect matrix. As the name suggests they have

Evolution of the PLD

The first PLDs were made in the early 1970s. Known as programmable read-only memories (proms), they comprise a programmable OR array fed by a fixed AND array (Fig. 1). The AND array is a 'fully decoded' array, meaning that all possible combinations of the inputs $I_0...I_2$ have a unique product term. The size of the array in a prom, due to this full decoding, grows as 2^n , where n is the number of inputs. This can result in a very large and costly device.

Proms tend to be slower than other PLDs due to the switching time of large arrays. Some small proms operate fast enough to find success as logic elements. Moreover, most logic functions don't require that all possible combinations of the inputs should be available, since many of the combinations are invalid or impossible.

The primary uses of proms have been in memory-type applications such as display look-up tables, and software storage known as firmware. Figure 2 shows the written programming conventions.

Figure 3 shows a typical PLD input buffer. Its two outputs are the true and complement of the input, as shown by the truth table. Figure 4 illustrates the written convention used to reduce the complexity of a logic diagram without sacrificing any of the clarity. The traditional representation of an AND gate shows three inputs: A, B, and C. The PLD representation has the same three inputs. This shorthand reflects the three distinct input terms of the prior drawing. The structure of a multiple-input AND gate is known as the Product Term.

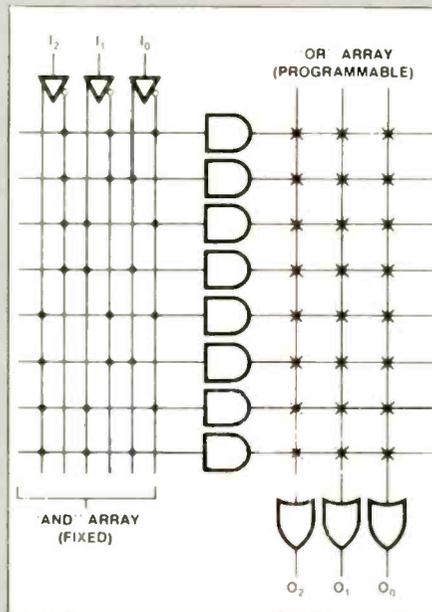


Fig. 1. Basic prom architecture.

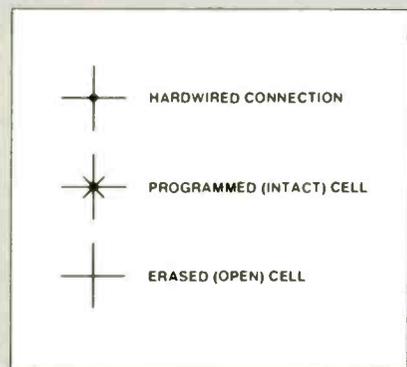


Fig. 2. Programming conventions.

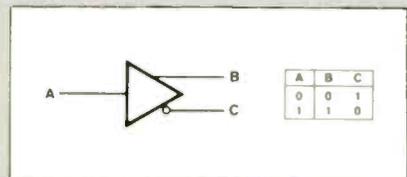


Fig. 3. A PLD input buffer.

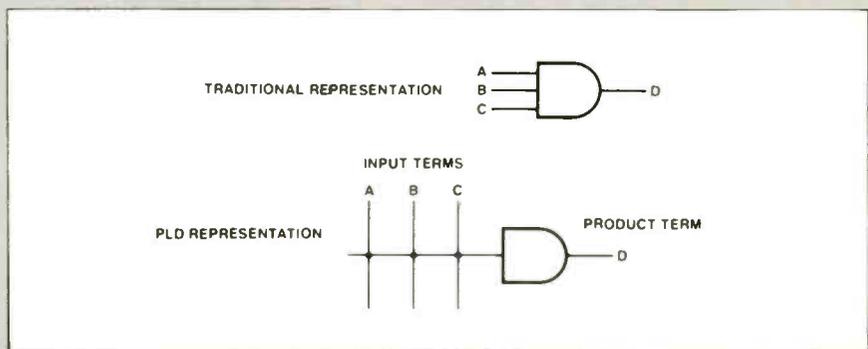
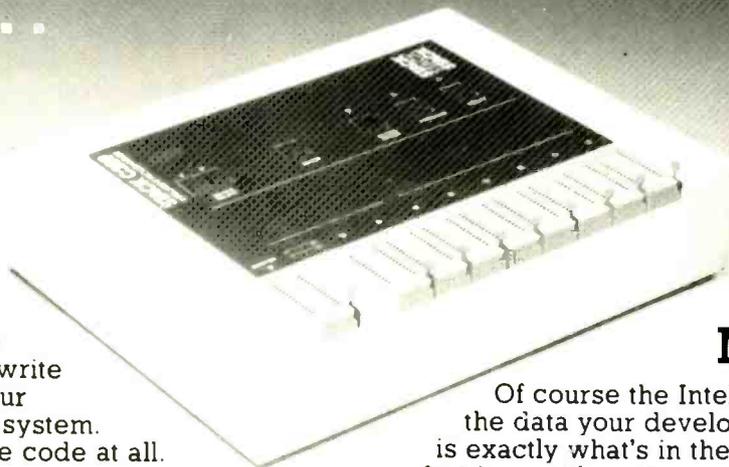


Fig. 4. AND gate representation in a PLD: all terms must be high for a high output.

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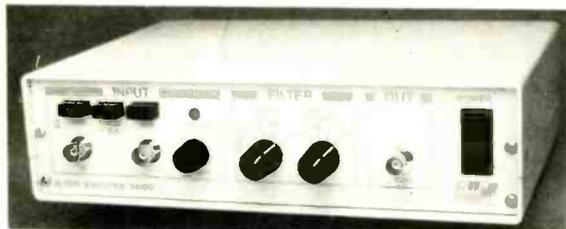
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more in common with gate arrays than the earlier PLDs.

Products vary according to the nature of the interconnect matrix. The Xilinx LCA, for example, is ram-based, and configures itself every time it powers up. Actel, on the other hand, makes rom-based parts that are one-time programmable.

King quoted an application example for the Xilinx part in a video board. The part could be used to emulate EGA or VGA operation simply by downloading the relevant programs into the device. Previously, each emulation would have required a separate IC: now, they can

all be done on the same chip.

The expected appeal of FPGAs as an alternative to custom gate arrays is the avoidance of the NRE costs, which can amount to between \$20,000 and \$50,000 per device and take six months to complete. And may not work. According to King semiconductor manufacturers estimate that roughly half of all custom gate array designs made don't actually do the job when finally plugged into the board.

"That's the big risk with a gate array," he said. "The FPGAs take all that away, in theory. You don't have to spend months simulating. You don't

have to spend that NRE charge with the semi house, and you can do the design and have it working in a week. If it doesn't work, no big deal: you just debug it, fix it and try again."

FPGAs do not currently offer the gate density available from full gate arrays, but they are getting there. However, they are more difficult to incorporate than either simple or complex PLDs and, as Pashley pointed out, their present cost per gate is considerably more.

The fourth and final sector of programmable devices in Data I/O's world-view is application specific programm-

Referring to Fig. 2, the solid-dot connection in the previous figure represents a permanent connection. A programmable interconnection would appear as an X over the intersection. The X implies that the connection is intact, whereas the absence of an X implies no interconnection. The connection of both the true and complement of a given input buffer to a single product term results in that product term always being a logic 0.

The Field Programmable Logic Array (FPLA) was devised in the mid 1970s as a more efficient way to handle logic functions. As depicted in Fig. 5, the FPLA has a basic structure similar to that of the prom; however, both its AND and OR arrays are programmable. To provide a smaller and faster array, the AND array is not fully decoded. Product terms can be shared by any or all of the OR terms.

With its smaller arrays, an FPLA operates faster than a prom. FPLAs find extensive usage in applications where the output functions are very similar, allowing full utilisation of the shared product terms. The dual programmable arrays make the designer's task easier, since he can control all of the functionality of the device.

The PLA (programmable logic array) approach of the late '70s, illustrated in Fig. 6, again varies the array control. This time, the AND array is programmable and the OR array is fixed. This approach offers reasonable performance and efficient architecture for logic functions.

The quantity of product terms per output is fixed by the hardwired OR array. The typical logic function requires some three to four product terms, well under the seven to eight available on current-generation devices. Figure 7 shows a general PLD structure.

The Lattice Semiconductor GAL (Generic Array Logic) architecture uses a programmable AND array driving a fixed OR array. The difference is in the architecture and flexibility of the output functions. The GAL device integrates an Output Logic Macro-Cell (OLMC) on each of its output pins.

The OLMC is configured by the designer on a pin-by-pin basis to implement the desired function.

The programmable polarity feature is implemented by the exclusive-OR (XOR) gate that follows the OR gate from the array. Recalling the truth table of an XOR gate, it can be shown that the data can be

inverted (control = 1) or not inverted (control = 0), depending on the state of the second input, as shown in Table 3.

Programmable output polarity is used extensively in DeMorgan's Law to reduce the number of product terms required to implement a function.

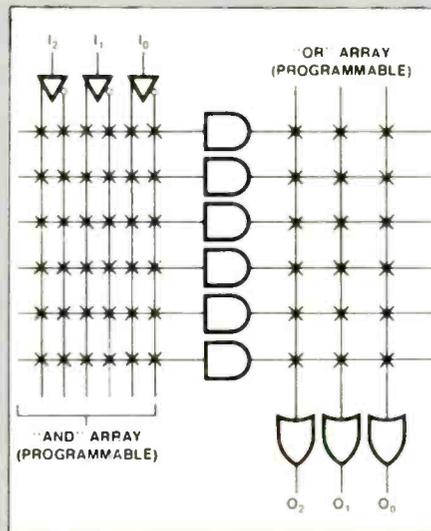


Fig. 5. Basic FPLA architecture.

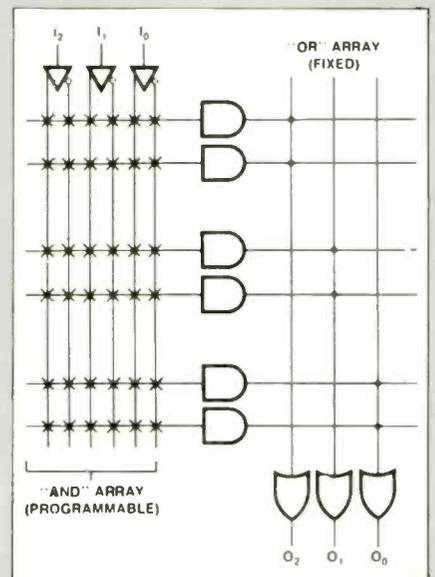


Fig. 6. Simplified PLA device architecture.

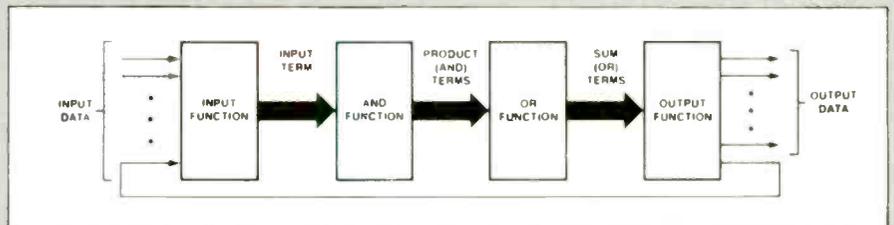


Fig. 7. Signal flow inside a PLD.

	ARRAY	OR	OUTPUT
PROM	AND	Programmable	TS, OC
FPLA	Fixed	Programmable	TS, OC, H, L
PLA	Programmable	Fixed	TS, I/O, Registered
GAL	Programmable	Fixed	User-Defined

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able logic (ASPL), also known as ASPICs. These are geared towards a specific application, but are programmable within that application. They currently constitute about 4% of the market.

Altera's EPB1400 user-configurable microprocessor peripheral is an example. It comprises specialist interface functions, such as a bus port transceiver, input registers and output latches, around a programmable logic core. It includes general purpose logic macrocells for functions such as address decoding, interrupt logic and state machines.

The importance of design tools

King believes semiconductor manufacturers exercise a lot of influence on the user's choice, particularly through offering free or low-cost design tools. He thinks such tactics are sound and useful to users. "It's one engineer talking to another," he said. "That's what it amounts to. They're pretty good about it."

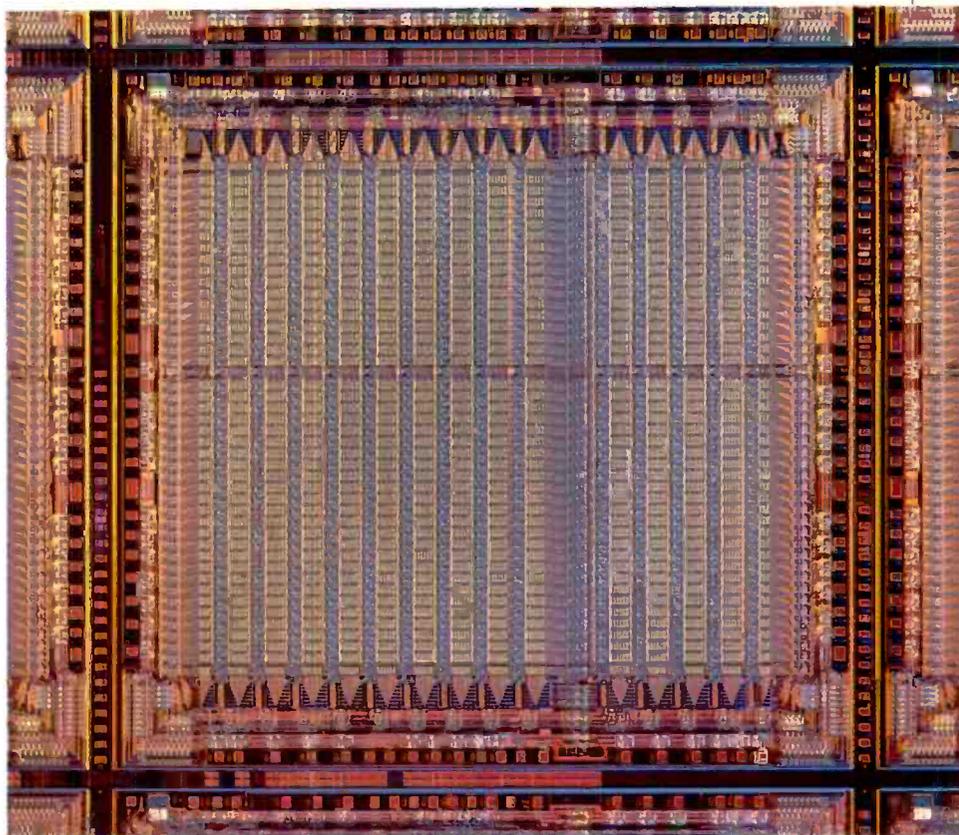
Design tools are key: it was the availability of such tools that got PLDs off the ground in the first place, and they remain critical. Lack of software support is holding the FPGAs back, for example.

With PLD tools, designs can be entered using a variety of techniques, including Boolean equation, state diagrams or schematics. The design software automatically reduces the logic and outputs a Jedec file. Data I/O also makes a test tool called PLDtest Plus which automatically adds test vectors to the Jedec file, which is then downloaded to the programmer to produce and test the device. The whole process is more or less automatic. Not so with FPGAs yet.

FPGAs have taken a similar design route to gate arrays, which typically means using schematics. This may not always be the case: Data I/O's Future Designer system allows FPGAs to be designed in exactly the same way as other PLDs from the same range of methods. But there is still some way to go before FPGAs become as easy to use as PLDs.

"Optimisation is a lot more complex, because you're dealing with several levels of logic," King said. "That's where the device manufacturer has to come into play. He provides the optimisation tools to do that. That's where the real complexity comes into it because it requires tools from a couple of different suppliers."

"Working with programmable interconnects may involve aspects such as delay times, and that's where it gets



Actel's ACT1010 FPGA contains 1200 gates yet can be programmed from an 80386 based PC.

more complex."

Even with the simpler PLDs, the sheer variety of sources of supply and architectures means that standardisation is non-existent. Pin arrangements — the assignment of power and ground, for example — can vary, along with a wide variety of programming algorithms, device speeds and even packaging options.

There are three ways of handling this: limit yourself to a single device type; use a programmer with adaptors; or use a universal programmer. The first option has obvious limitations and the second, according to Pashley, is unpopular with users because of inconvenience or unreliability. King makes a technical argument against the use of programming adaptors.

"Devices during programming are fairly sensitive," he said. "The machines are tuned to deliver maximum performance at the head of the programmer. If you stick another socket in there, it can cause problems, such as noise on the lines."

Data I/O's Unisite programmer uses pin-driver technology to avoid the need for adaptors. But it's not cheap: around £12 000 for the 40-pin version. The recently launched 2900 programmer

provides similar functions for prices in the range of £3500 to £6500, though it is limited to 44-pin devices. Unisite will be upgraded this summer to cater for up to 188 pins.

Testing of devices once they have been programmed is an issue that hasn't been properly addressed yet, King believes. In the case of a full gate array, there is a generally acknowledged path between the customer's design simulation and test at the semi house; with PLDs there is a tendency for in-house buck-passing to develop.

"That's been the big question in the industry over the last year or so," said King, "to identify who's responsible for testing PLDs: is it the production department or the designer? It's important, but nobody wants to own up to the responsibility."

Such issues might be settled some time in the future. By that time, PLDs will have got faster — down to 2ns, King predicts — and FPGAs will have become much more popular. Their prices will have fallen and design tools will have become as easy to use as those for PLDs.

"I don't know if it's the tools that are driving the technology, or the technology that's driving the tools," said King. "But we're definitely chasing each other around right now. It's tough keeping up with the technology because every year there's a couple of hundred new devices added to the pile." ■

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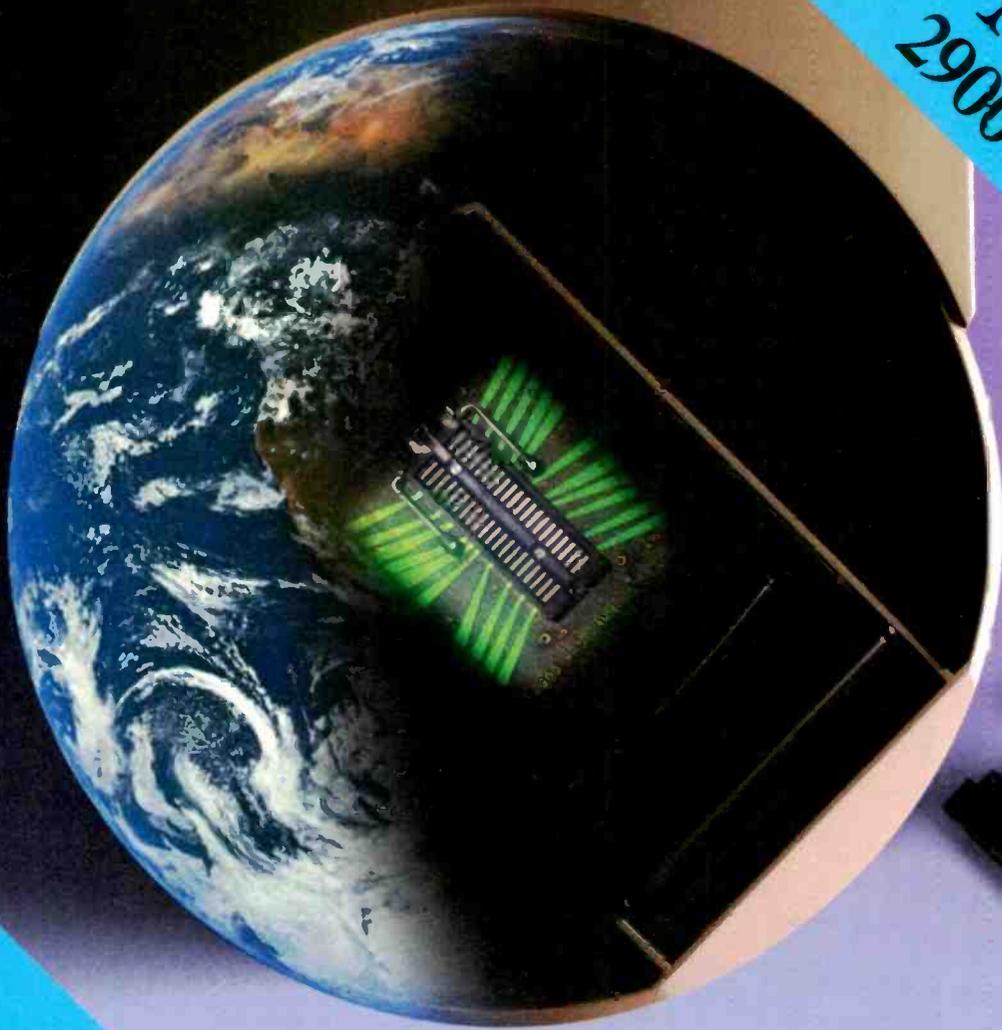
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1	2	6.08	2.09	1	2	6.19	1.98	1	2	10.25	2.20
2	4	7.01	2.20	2	4	10.01	2.20	2	4	13.17	2.53
3	M 6	12.08	2.36	3	M 6	11.60	2.42	3	P 6	19.05	2.64
4	P 8	12.87	2.42	4	P 8	13.84	2.53	4	S 8	21.72	2.75
6	S 12	15.62	2.64	6	S 12	16.62	2.74	6	S 12	27.46	3.19
8	16	18.59	3.08	8	16	19.41	2.91	8	16	31.32	3.41
10	20	25.02	3.52	10	20	29.94	3.24	10	20	44.04	4.12
15	30	31.10	3.63	15	30	33.42	3.45	15	30	51.28	4.40
20	40	44.40	4.12	20	40	37.43	4.01	20	40	59.09	5.22
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CIRCLE ENQUIRY NO. 151 ON BACK PAGE

MICRO-CAP III

under test

Micro-cap III (Computer Analysis Program) is a stand alone, feature-packed analogue simulator, produced in California by Spectrum Software. Although it has no sister programs, it's encouraging to know that the series has been around for a comparatively long time, as software goes. Micro-cap I was introduced in 1982, and the III series reviewed here was first released in 1988.

The authors claim, "Micro-cap III combines the speed and accuracy of spice-like numerical methods with a modern, easy-to-use interface ... which avoids the mistakes that are easily made with spice, even by experienced users." In its latest version 3 format, released March 1990, Micro-cap III (from now on abbreviated to MC3) can handle circuits or macros comprising up to 500 components, with an unlimited maximum node-count. Overall, the authors claim, it is the easiest to use and among the most powerful of all PC-based simulators.

The manual cites, "Speedy analysis, using a proprietary Sparse matrix technique." Spectrum says it has shipped more than 10,000 copies of the Micro-cap family, with most sales made in the USA. The majority are in use in large corporations, in all phases of electronics except RF.

System considerations

With Spectrum's own Windows-like video interface, MC3 requires around 590k of DOS base memory in full use, a borderline condition for many PCs with memory usage that isn't finely tuned. Datech, which distributes MC3 in the UK and Europe, can help with optimising your PC's configuration.

As an immediate measure, I had to disable my machine's EMS and CACHE and reduce files and buffers in

CONFIG.SYS to get the required 590k. A co-processor is optional.

Micro-cap's computational speed can be best appreciated with a 386 machine. MC3 automatically detects the system's highest-resolution graphics mode — CGA, EGA, VGA, MCGA and Hercules are all recognised.

Fast, accurate, easy to use — this computer circuit analysis program gets all the superlatives. Ben Duncan explains why

The hardcopy outputs are specified for HP LaserJet or Epson printers, and HP or HI plotters. I had no trouble interfacing with an Epson/IBM-compatible Star printer, but there were some text offset problems with my HPGL-compatible plotter. The programme disks are copy-protected, with 'hidden files' which prevent them being loaded onto more than two hard disks at one time.

The V3.0 software occupies about 3MB and can be supplied in all the standard floppy formats. Installation, de- and re-installation are straightforward, being taken care of by a special menu.

Operation and schematic entry
The schematic is entered directly. Using the mouse is straightforward and is doubly easy if you've already mastered a schematic drawing program. But it is unfortunate that the manual takes the Windows environment for granted — having graduated via DOS it was unfamiliar to me, and I was left feeling handicapped and frustrated at first. Scrolling

and panning are certainly fiddly, requiring a fine touch on the tiny icons.

All commands can be entered directly from the keyboard. This is good, as judicious use of both mouse and keys is an unbeatably speedy combination. However, if you try to use the keyboard from the start, you'll waste a lot of time hunting through the manual. The quick reference card wasn't functionally ordered and some keyboard commands have been omitted — the software world seems impervious to learning aids or refinements of any kind, even low-budget ones!

Also, there is no keyboard macro facility — the F keys are dedicated. Eventually, I found myself absorbing the key commands one by one, once I had conquered MC3 with the mouse.

Schematic elements can be moved, copied and block deleted. Separate circuit files can be merged. The relevant library page can be directly called up on any part by pointing. You can write descriptive text on the screen. There's also a pull-down notepad screen for each circuit file, but the wordprocessing function depresses me, behaving like PRED, the DOS editor.

Circuits can be macroed up to ten levels. With suitable high level macro-models in place, MC3 should be capable of analysing complex feedback systems in every branch of engineering and the physical sciences.

A circuit containing several op-amp stages, or typically 30 discrete components, can be viewed in one go. Layout density is principally limited by component labels, for which there is a choice of only two positions, and the need to keep these from overlapping for clarity. Circuits with value, tolerance and temperature coefficient appended to every passive part will need to be spread out.

For larger circuits, the screen scrolls, or you can use the three-level zoom

function. On hi-res EGA and VGA monitors, resolution was degraded on levels 2 (2:1) and 3 (4:1) respectively — good enough to display the nodes and component designations sharply. The result is that with a VGA monitor you will be able to operate comfortably at 2:1, as it allows you to see and work on a circuit with about 120 components. Node numbers are visible and updated automatically at all times, though they have an annoying habit of straddling circuit traces.

MC3's "6-way x 3-channel" palette (colour selection) is quite versatile but it doesn't allow nodes to be a distinct colour from the component text. To overcome the way that node numbers can change as a circuit develops, MC3 allows you to label the nodes you want to monitor with fixed text labels, e.g. "Input."

Components are listed on and can be picked from the right-hand COM (component) menu. I was impressed by the ease with which the parts I most used could be dragged into a bespoke order on the top page. To expand the visible drawing area, the COM menu can be put away, assuming you're content to enter from the keyboard. Passive component values can be entered with engineering units (e.g. 1k, 1nF) or scientific notation (1E-9).

The available components aren't systematically depicted in the manual. DIY graphic symbols for custom components and macro circuits can be created with a shape editor. MC3 already includes n-channel, mes-fets (sic), transmission lines, polynomial sources

and logic gate macros alongside all the basic components, sources and drawing aids. The thyristor/triac/ujt family is not represented, but they should be modellable either as bipolar macros, or by including basic mathematical expressions for the transfer function in the component labels, as a model statement. By this route it should be possible to model any conceivable component — even a fictional one — given enough patience.

For example, "25pF*(1+f/1E6)" models a capacitor with a value that varies with frequency. The Micro-cap manual and newsletter give examples of an opto-coupler and a triode tube, among others. For these and other completely new component types, you'd use the shape editor to create the necessary graphic symbol.

Libraries, parameters and printing

MC3 allows an unlimited number of custom component library sets, which could be allocated to different areas of design or to specific projects. Each library set comprises a generic passive label library (i.e. where a commonly used component value or string is labelled "X" or "7" to save entry time) and individual libraries for the five kinds of active device models (diodes, BJTs, J-fets, and mos- and mes-fets), as well as for plain current sources; independent and dependent linear and non-linear (polynomial) sources, such as V controlled I sources; transformers; and transmission lines. Each library can con-

tain at least 50 parts.

Two sample libraries (STD.1.2) are supplied. These include the parts, sources and labels for some 20 sample circuits, as well as semiconductor models for device types that are doubtless common fare in California. With not a "BC" or "2SA" in sight, the listing is not so impressive in the UK, Europe or Japan.

MC3's choice of independent signal sources begins with sine and pulse primitives and setting these up was easy. Sine waves can be programmed with exponentially decaying or ascending amplitude and/or frequency, and non-integral periodicity. You can enter a DC offset and any initial phase and source resistance. Pulses range from one-shots to continuous square or triangular waveforms. More complex waveforms can be generated by summing or adding suitable sources in series.

For example, to design a filter to clean up a noisy automation system in a recording studio, I had no trouble simulating the staircase waveform emerging from the dac, using a chain of square-wave sources with suitable staggered transition periods.

Alternatively you can design and enter custom waveforms using mathematical text expressions or enter them as a user source file. Or where applicable, you can use the waveforms emerging from analysed circuits.

The esoteric process data required to accurately model semiconductors (after Ebers and Moll or Gummel and Poon) is a big stumbling block to the simulator business. Datech suggested it should be available from the makers, but in practice, the makers of discrete transistors ignore requests or claim they don't have an item readily available. In reality, makers are shy about revealing intimate details of their processes to public and competitors alike. In contrast, op-amp makers (beginning with PMI and Texas) are eagerly promoting spice-models for their ICs.

MC3's "Global settings" menu allows you to choose between the EM-2 and GP models for BJTs and diodes, and mosfets and j-fets can also be modelled at two levels of detail. Without any help from the manufacturer, one is faced with modelling familiar transistors using either a curve tracer, or MC3's pep (parameter estimation program). Accessed via DOS, the pep utility computes the EM2 (or level 1) parameters for diodes and transistors using figures that can be garnered from most data sheets, possibly with the help of a few measurements. The manual is quite informative

Micro-cap III - key specifications

System:

PC or PS/2 compatible, >DOS 3.2. 386 processor preferred. Hard disk essential. Microsoft or compatible mouse. At least 575k of available memory, preferably 590k. EMS ≥ 1MB for highest capacity when performing simulations with large numbers of timepoints.

Drawing Elements

- any user-defined symbol, plus:

Passive components: Rs, Cs, Ls, transmission lines, transformers, non-linear magnetic cores.

Active components: Diodes, BJTs, J-fets, mos-fets, n-ch mes-fets, op-amps.

Signal sources: Sine, pulse, constant voltage, constant current, formulae- and file-driven user-defined sources. Algebraic and table-defined complex frequency transfer functions. Polynomial, linear, algebraic and table-defined dependent sources.

Miscellaneous

Switches — voltage, current and time-controlled. Macros, representing other complete circuits.

Computation (Sparse array)

Maximum no. of nodes: unlimited
Maximum no. of parts: 500
Macro nesting: up to ten times

On-Line Utilities

Help.
Notepad for each circuit file.
Scientific calculator.
Shape editor draws new schematic graphics.

Utilities accessed via DOS:

PEP derives Eber-Moll parameters.
TOSPACE creates SPICE circuit files.
CONVERTL updates Micro-CAP II circuits.

at this point, using extracts from Motorola data sheets to illustrate the process.

In practice, there remain some nagging uncertainties, like, "How reasonable were those figures I had to guess?" And, "Up to what frequency can this model be relied upon?" The pep gives a basic clue by plotting the characteristic curves of semi-conductors, enabling a check against published data. Still, I'd like to see programs such as pep echo a number of the main data sheet curves before I felt really confident that my DIY models are reasonable. Also, while some Gummel-Poon data is given for the semi-conductors in the sample libraries, the pep doesn't support GP (or level 2 fet) models, which are regarded as too complicated.

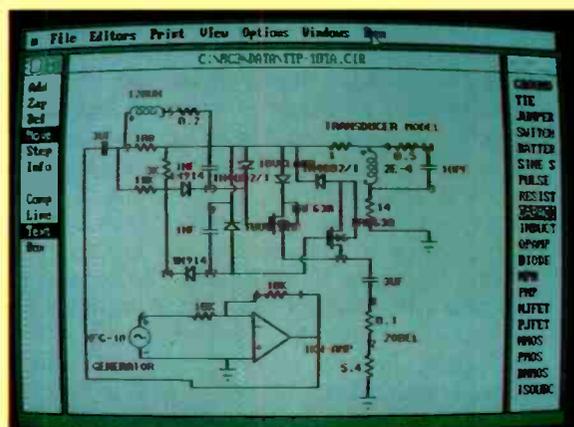
Compared to spice macros, MC3's op-amp models are simplistic, with three cascaded gain stages and two poles defined, and with no access to power rails. Library entry is easy since the data can be taken directly off the data sheet. Overall, I found MC3's op-amps fine for gauging the explicit behaviour of a wide variety of everyday circuits up to 1MHz. Alternatively, MC3 can import, convert and use the more detailed spice and p-spice models. Or, assuming a knowledge of mono-fab process parameters, you can compose your own discrete component macro. The elderly μ A741 is provided as a sample.

Analysing options

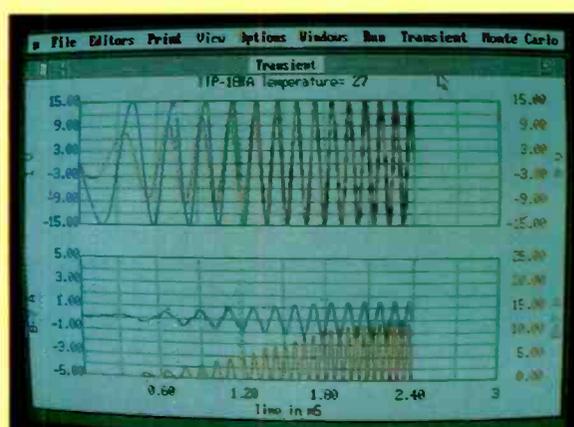
The analysis routines are immediately accessible once the schematic is drawn. If you forget to earth the circuit or connect all the nodes, you get a reminder on the screen and guilty components flash - you won't be able to analyse anything until you've fixed your mistakes. The analysis routines have many features in common. Irrespective of which you choose, both the stimulus (input) and the point being monitored can be single-ended or differential with respect to ground.

Micro-cap III costs £1,495 + VAT. An evaluation version limited to 10 nodes but including every feature, together with the manual, can be obtained for £150 + VAT, with this price credited if you then opt to purchase the full version. UK users receive a newsletter reporting on practical applications, tricks, tips and new models. Datech Ltd., Sidcup, Kent. Telephone: 081-308 1800.

A relatively complex circuit, showing how component labels are apt to overlap. Colours are one of thousands of combinations selected from the EGA palette. Note N- and P-channel mos-fets, descriptive labels ('transducer model') and a custom op-amp ('1kW amp') modelling a high voltage amplifier.



A transient analysis of the circuit (exp 1,2,3) above. Upper screen shows sinusoidal stimulus with exponentially rising frequency, beginning at 1kHz on the left. Overlaid is the output voltage. Lower screen shows current in the zobel branch, and the power dissipated in the zobel resistor, and how it varies with time and frequency. The plot was finished prematurely (at 2.4ms) demonstrating how an analysis can be aborted once the trend is clear.



Plots on the screen are displayed in real time and can be printed as soon as the run has finished. You can also abort an analysis immediately if it is not giving the right result, allowing design by rapid iteration. Results can equally be sent to a plotter in real time, but this slows the plot on the screen. If saved, any subset of the original graph can be reviewed and plotted on the screen, or sent to the plotter (or printer) later, or both. You can also arrange for automatic plotting from DOS batch files. Graphs can be annotated with horizontal or vertical text.

During a run you can choose to dump a numeric listing to the printer, the screen or a file. The same information can be seen in a dynamic numeric readout that can be toggled during a run. To avoid generating too much data you can limit the number of increments, but I could see no way of limiting the output to specific frequencies of interest.

For example, the RIAA replay response for phono amplifiers is defined at arbitrary frequencies, so a call to Datech's support was needed. The answer was to divide the analysis into n runs integral with pairs of the spot fre-

quencies, then use MC3's MERGE function to combine the output files. A SCOPE function allows any portion of the graph currently on the screen to be magnified, but the x- and y-axes are unscaled. Instead, measurement is performed with cursors, their positions revealed in an adjacent numeric display of the x- and y-axis limits and differences.

AC analysis (normally small signal) automatically computes the DC operating point (if required), then sends a 1V sine wave up any required node, and goes on to plot amplitude, phase and group-delay across any desired range of frequency. As with any simulator, results will be increasingly doubtful above (say) 1MHz unless realistic parasitic elements have been inserted. New with version 3 is the ability to perform large-signal analysis by calling up a suitable sine-wave source. Either way, Nyquist diagrams, input and output impedance (or admittances and conductances) and noise (in nV/√Hz) can also be displayed.

Of course, for some circuits, noise figures will be critically dependent on device modelling, and op-amp noise is not included. For gain and noise the

trum admits that the dot matrix (cf. laser) printing facilities are a weak area. This is because a fixed set of binary drawing scales is used to speed up schematic entry and scrolling. Spectrum says higher resolution is planned for future versions.

I found MC3 was good at anticipating (and spotlighting) errors before they had an opportunity to cause hanging or lock-up. Simulators can play a multitude of tricks on the unwary, but MC3's UK users receive a newsletter which goes some way to documenting the (largely unpublished) tricks of the trade.

For example, a transient analysis of a rectifier circuit took 12 hours to run. At least, with MC3, you soon know when a simulation is going to be painfully slow. But by knowing a certain rule-of-thumb, you would change one of the diode's parameters and lo, the simulation would be completed within ten minutes. Repeated high-precision analyses can easily end up filling your hard disk with megabytes of rubbish, and with so many other on-line features, it is unfortunate that there is no facility to erase, sort and back-up files directly from the program instead of via DOS.

Conclusions

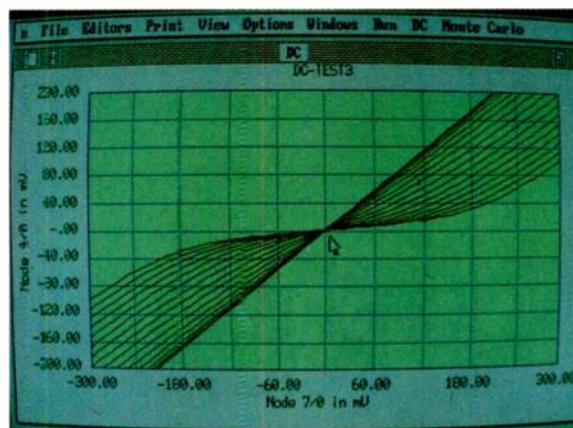
Any mention of analogue simulators gets a variety of reactions from engineering colleagues, ranging from bemusement to outright hostility. And yet all the chips in their computers were designed with simulators – for IC design, this is the only practical way. All simulators have a minimum time for setting-up and getting into gear, and MC3 seems well above average in this area, all the more so once you've invested time in entering regular devices and circuit macros.

Simulation is no substitute for your engineering knowledge, but it adds a mighty weight when you need to shove, and acceleration when you want to shift. It comes into its element when a major circuit needs repeated and minute analysis, or when competing (even competitors') topologies or networks are being compared.

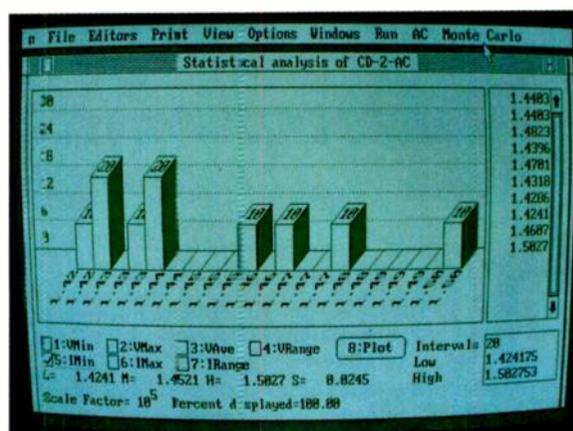
The kind of methodical thinking needed to model and enter a circuit is often enough to trigger the inkling of an answer in your own mind. Sometimes, the discipline of modelling a circuit moves the goalposts, making the model irrelevant. If it ultimately leads to a solution, who cares?

With simulators, the value-for-money judgment is quirky at best. Elementary software costing under £500 has already been reviewed in these

A 'ten run' Monte Carlo plot of a typical CD player output stage with typical component tolerances. The disparate direction of the phase plot just below 1MHz shows the effect of setting the maximum accuracy too high; it was set at 15% to speed up the plotting.



A 3D statistical analysis of Monte Carlo graph (pic 14,15), showing the worst case minimum distribution of the independent parameter, I_{min} .



pages, and Micro-cap III costs barely three times more, and yet (judging by conversations with colleagues) it seems as powerful and far easier to set up and use than the simulators included in some workstation packages costing over £20,000.

Dealing in abstractions, all simulators take a lot of learning, and the cost of the software is plainly irrelevant if it takes 1,000 hours of "playing" to reach the stage of feeling confident of getting accurate results fast. I think the average engineer will get to grips with MC3 faster than with any spice-based program.

Once MC3 (or any good simulator) has been mastered, and subject to the care taken in modelling, the engineer has effectively gained means of investigating the behaviour of global circuits and systems above and beyond the physical limits of engineering, if not the universe. Overall this is such an all-encompassing programme, with uses in so many diverse branches of analogue systems analysis, that it would take several years of full-time exploration to assess and verify each avenue in detail against even the most niche-like realm

of electronic circuit topologies.

For example, version 3 includes (for the first time) the Giles-Atherton state-variable magnetic model. Magnetic material analysis is outside my remit, but it will be of great interest to engineers seeking to design and optimise the magnetic materials used in transformers and other inductive components.

Datech's support was friendly, well informed and reasonably quick, and staff aims to solve any level of technical problem within a variety of response times ranging to a maximum of 36 hours. The software has a few quirks, but (as always) there's scope for refinements. The beautifully presented manual does not do it justice – it's refreshingly readable and cogent as simulator manuals go, but lacks nitty gritty details, and contains quite a few typos and omissions.

Micro-cap III is friendly, competent, workable and well backed-up. You could do a lot worse. ■

AUDIO PREAMPLIFIER DESIGN

Although ceramic and "crystal" pick-up cartridges were of some importance during the early days of transistor-based preamplifiers, if only because magnetic stereo cartridges were relatively expensive, their circuitry was not examined in the first part of this article because the problems and constraints of frequency response equalisation did not apply to them in the same manner as for magnetic cartridges

These cartridges are based on piezoelectric displacement-sensitive transducers, which give an electrical output proportional to the amplitude of the groove modulation. Other things being equal, this would lead to an uncorrected replay frequency-response curve from a disc recorded with RIAA equalisation into a high-impedance load of the form shown in Fig. 1.

In spite of the relatively poor linearity of the transducer element — a typical THD vs frequency graph for a ceramic cartridge at 0.003cm groove modulation is shown in Fig. 2 — these cartridges were popular among amateur constructors; input circuits for these devices, intended to optimise their performance, were described by Burrows^{1,2} and myself^{3,4}.

Crystal pick-ups are rather fragile, both humidity and temperature sensitive and are now seldom found, but ceramic cartridge types are robust, still being widely used in low-priced systems.

In contemporary applications, the mounting of the piezoelectric element and its linkage to the stylus assembly are generally arranged so that, with a

Good preamplifier design has always centred on the available headroom, particularly in equalisation and tone control circuits. John Linsley Hood continues his story of audio development.

nominal 47k Ω resistive load, the cartridge's electrical output is roughly equivalent to that of a velocity-sensitive unit; a conventional, but low-sensitivity RIAA input stage can be used, though with some sacrifice of the low and high frequency response.

Headroom

Very few topics can have generated as much debate in the audio field as the question of the input overload margins practicable with any given circuit, a factor generally referred to as "headroom". Regrettably, much of this debate has been entirely misguided.

The reason for this is that all the signal sources applied to a preamplifier system have specific output limits. For example, in the case of a tape recorder, the recording level is generally chosen so that the peak output signal level is not more than 3dB greater than the normal span of the recording range. At +6dB, the THD would probably have increased, in the case of a cassette recorder, from the typical 0.5% to some 3-5% and at +12dB (4 \times) magnetic saturation of the tape would probably clip the output signal anyway.

Similar considerations exist for both FM tuners and CD players, where a +6dB overshoot beyond the normal maximum output is probably as large as can be anticipated, bearing in mind the rigid constraints which apply, either in maximum frequency deviation or because of the digital encoding of the signal.

In the case of the output from a typical magnetic pickup cartridge, the limits on



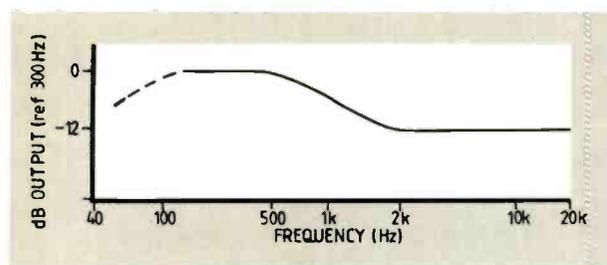


Fig. 1. Uncorrected frequency response of displacement-sensitive piezoelectric pickup cartridge

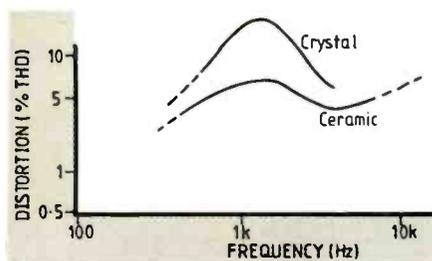


Fig. 2. Typical distortion figures for crystal and ceramic cartridges

output voltage are just as real, though less obvious. In effect, the practicable output voltage from such a cartridge depends on the extent to which it is mechanically possible either to cut a steeply undulating groove on the record surface or for the pick-up stylus to follow it.

This subject was examined in an informative article by Walton⁵ and the maximum recorded levels which were feasible for a vinyl record at various groove diameters are shown in Fig. 3.

Below about 1kHz, maintaining a constant recording velocity would require the amplitude of the groove modulation to increase; even if the groove spacing is adjusted somewhat to allow for anticipated modulation levels, there is a physical limit to this permissible groove excursion. Above about 2kHz, the shape of the rear faces of the triangular-section cutter head means that, as the frequency of the modulation is increased, the cutter tool itself imposes a limit, for any groove diameter, on the closeness between succeeding groove deviations.

Under optimum conditions, Walton considered that, at a 7.5in groove diameter, a groove modulation velocity equivalent to 30dB above 1cm/s at 2kHz was the maximum which could be cut.

The Shure Corporation, which made a speciality of the design of pick-up cartridges with very high tracking ability, claimed of its best models that they were capable of tracking a groove modulation equivalent to +40dB ref. 1cm/s, decreasing with increasing recorded frequency. Shure obviously considered that, on its stereo test records,

the +25dB at 10kHz velocities associated with the "musical bells" track would present formidable tracking problems to most of its competitors.

Therefore, if a recorded velocity of 5cm/s is taken as the normal maximum mid-band signal amplitude, the limitations of the recording process ensure that the maximum velocity the pickup encounters on a heavily modulated part of the recording does not exceed 30cm/s — a 6X overshoot in output signal level. This practical limitation on the pickup cartridge output voltage was also detailed by Wolfenden⁶ using figures from Shure, and by Kelly⁷, who confirmed this magnitude of feasible overload margin.

However, with the same kind of logic which persuades ordinarily sensible motorists that a 150MPH car is a much better proposition — under conditions where the legal speed limit is 70MPH — than one which will only do 100MPH, many circuit designers seem to think it necessary to provide RIAA-stage voltage overload margins in the range 20-30X or greater. By praising such achieve-

ments, reviewers perpetuate the belief that this style of design is both good and necessary.

In reality, the highest output voltage peaks are likely to be associated with blemishes on the record surface and, while it is obviously desirable that such sudden voltage excursions do not produce any prolonged paralysis of the amplifier, brief-duration overload clipping might make the equipment more comfortable to listen to.

Since a typical IC op-amp gain stage, when fed from a $\pm 15V$ supply, is capable of a 9.5V RMS undistorted output-voltage swing, it should be capable of handling the output from any disc whose normal maximum output level is less than 1.5V RMS without clipping.

There are several practical solutions to the problem of ensuring that the output voltage characteristics of the signal source (a low-output moving-coil cartridge might generate $50\mu V/cm/s$, a high-output variable-reluctance design offering 3mV for the same modulation depth) are suitably matched to the amplifier.

Of these, the most elegant is to offer a choice between separate low-sensitivity and high-sensitivity pickup inputs; this approach is almost universally used in the better preamplifiers. Alternatively, a preset gain control can be included at an appropriate point in the circuit chain; a typical arrangement, shown in outline form in Fig. 4, was adopted in the RIAA stage of a design of my own⁸.

An alternative approach is to place

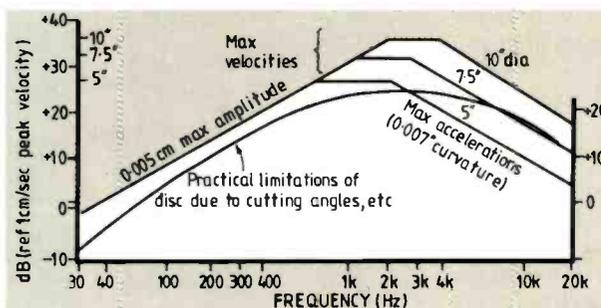


Fig. 3. Maximum recorded levels on vinyl discs at three groove diameters (after Walton)

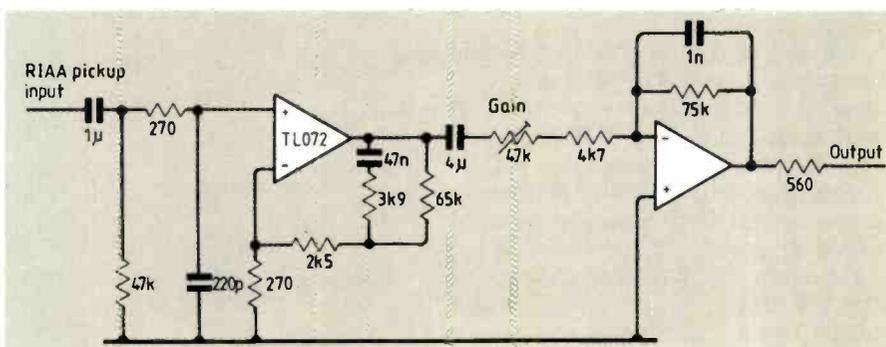


Fig. 4. RIAA equalised gain stage with preset adjustment

Fig. 5. Forward-placed gain control (a), which needs low-noise succeeding stages. That at (b) reduces noise with signal

Fig. 6. Typical passive tone control. Note asymmetrical component values. Non-linear potentiometers were used

Fig. 7. Baxandall feedback tone control, with symmetrical component values and linear potentiometers, is at (a). Switched components allow choice of turnover frequencies, as at (b)

the gain control as far forward in the preamplifier layout as is practicable, usually between the output of the RIAA stage and the input to the filter and tone-control stages, as shown schematically in **Fig. 5(a)**. I adopted this layout for my 75W integrated amplifier design⁹ and I note that it is employed by Quad in all its current designs.

While this choice of gain-control position offers a complete solution to the problems of inadvertent input overload, it carries the penalty that any noise introduced by the various preamplifier stages will be present all the time, even when the input gain control is set to minimum, so great care must be exercised in the design of all the preamplifier stages to keep the noise levels as low as possible.

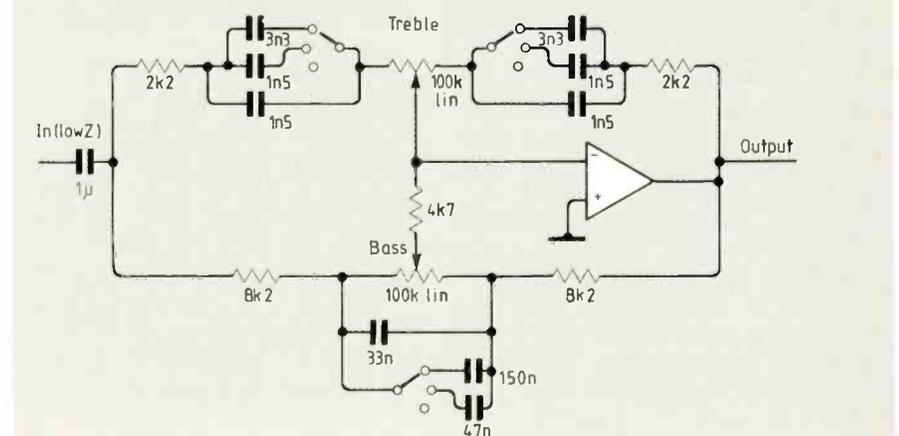
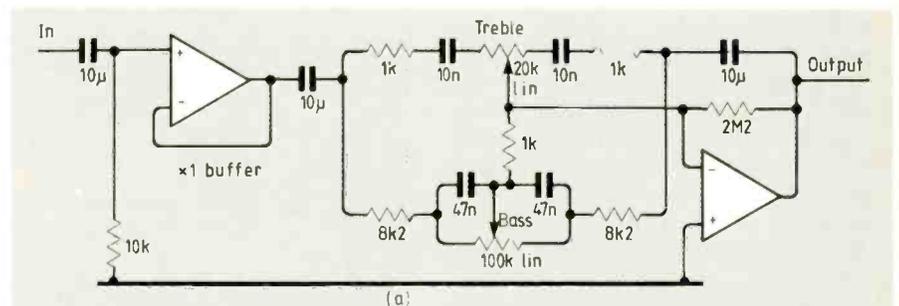
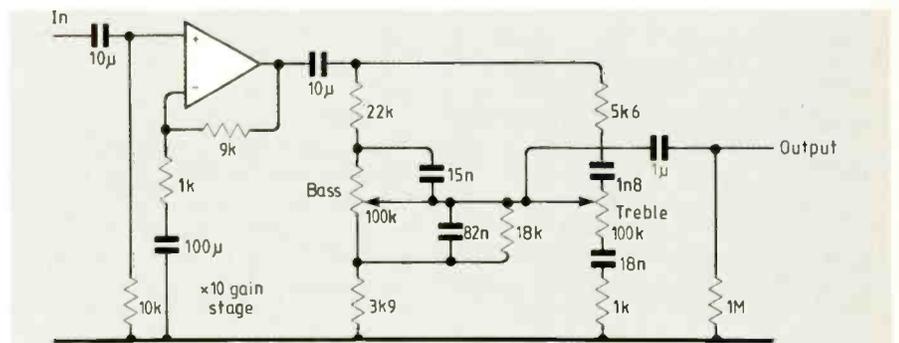
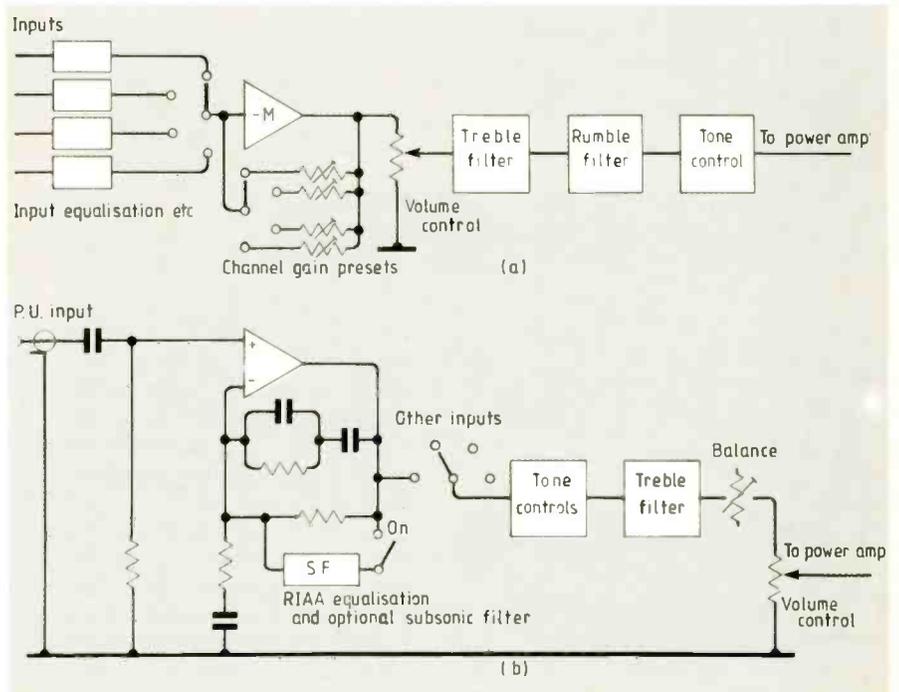
In view of this, the more common commercial approach is to place the gain control between the end of the preamplifier chain and the input to the power amplifier, as shown in **Fig. 5(b)**; for this layout to be satisfactory, the gains of the preceding preamplifier stages must be chosen with care, having regard to the likely signal levels which may be applied to them.

Tone controls

In nearly all early preamplifier designs, some form of "tone control" was employed, and the quality of the design was often judged on the extent of the facilities it provided for adjusting frequency response.

These tone control systems generally consisted of separate "lift" and "cut" applied to the bass (below about 500Hz) and treble (above, say 1kHz) to allow for deficiencies in loudspeaker performance, programme material or listening environment. They were implemented either by passive RC networks of the kind shown in **Fig. 6**, or by a version of the celebrated feedback tone-control circuit due to Baxandall¹⁰, shown in a typical contemporary form in **Fig. 7(a)**.

Among the advantages of the feed-



back layout is that it allows the use of an almost completely symmetrical circuit, based on linear-law controls; to achieve the same frequency response adjustments, component values in the passive circuit must be somewhat lopsided and will not give flat response settings at the mid-point position unless non-linear potentiometers are used.

Also, the feedback system allows the stage to be operated always at the minimum gain level necessary to achieve the required frequency response, whereas the gain stage preceding or following a passive RC tone control network will be operated at full gain all the time, possibly eroding the available headroom margins.

Figure 8 shows the frequency response adjustments which these tone control circuit layouts provide. Clearly, adjustments of such a simple nature cannot remedy all likely shortcomings in the frequency response of the whole audio system and this has prompted designers to offer more elaborate controls.

Graphic and parametric equalisation

More comprehensive facilities can be provided either by means of a simple switched adjustment to the values of the tone-control capacitors, as in the preamplifier layout of my 75W design⁹, shown in Fig. 7b, or by an arrangement in which the gain in each part of the audio band is made separately adjustable.

Most commonly, the pass-band is divided into individually controllable, octave-wide frequency segments, the layout being usually called a graphic equaliser. Figure 9 shows a typical circuit for this type of tone control due to Williamson¹¹, which provides the type of frequency response shown in Fig. 10. As well as the separate, add-on graphic equaliser units now available, Hitachi offers a simplified graphic equalisation system on most of their current preamplifier designs.

A philosophical objection to the graphic equaliser is that, since the frequency response is built up from a series of humps and troughs, the only setting which will give a ripple-free frequency response is that in which all the controls are set to mid-position — a consideration which may affront the purist.

Perhaps because of this, the so-called parametric equalisation circuit was evolved. This is similar in its circuit structure to the graphic equalisation system, except that only one peak or trough is generated and both the ampli-

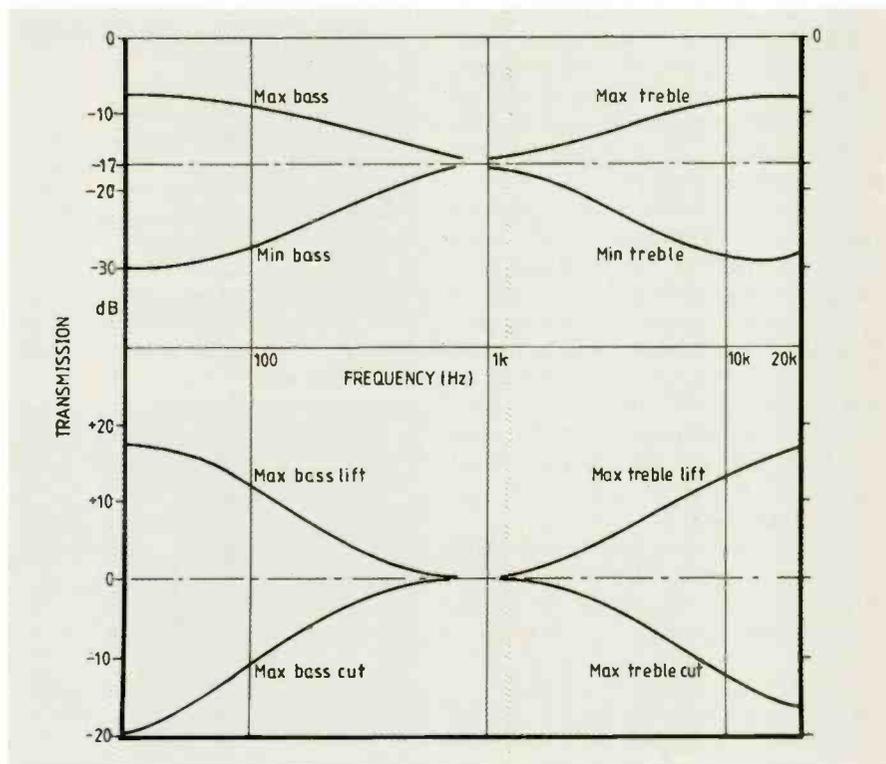


Fig. 8. Performance of passive circuit of Fig. 6. (top) and the adjustment provided by feedback tone control circuit of Fig. 7.

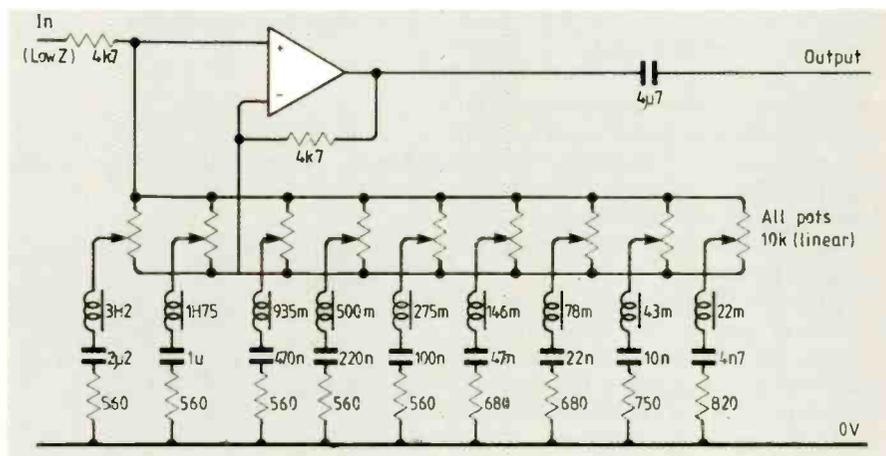


Fig. 9. Reg Williamson's nine-octave graphic equaliser of 1973

tude and frequency of its point of operation can be altered to allow a specific adjustment at a single point in the audio band.

Other tone-control systems

As an approach which would offer local frequency correction, but without a large degree of frequency ripple, I suggested a system based on a switch-selectable group of RC elements¹². The possible frequency response curves are shown in Fig. 11 and give rise to the name "Clapham Junction" tone control.

Better programme sources and greatly improved input and output

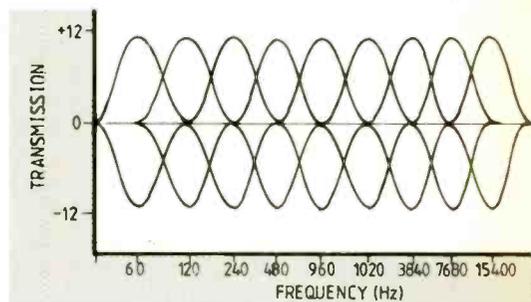


Fig. 10. Response given by graphic equaliser shown in Fig. 9.

transducer systems have reduced the problem of poor flatness frequency response and there is a growing tendency among the purist manufacturers to offer systems without any frequency response modification facilities whatever. This also reduces the cost of the equipment.

A sensible reaction to the improved standards of programme material and transducer performance was the proposal that only a simple adjustment of the slope of the frequency response curve from bass to treble was normally needed to correct for an over-bright or bass-heavy performance. Bingham¹³ published a circuit for a tilt control using the circuit shown in Fig. 12, which gives the kind of frequency response adjustment shown in Fig. 13.

This type of control is used by Quad in current preamplifier designs, using the circuit layout shown in Fig. 14, which gives the effect shown.

Filters

The original reason for the use of bass and treble steep-cut filters was the desire to obtain the best possible reproduction from 78RPM shellac discs pressed from material loaded with emery powder, which causes a continuous background hiss during replay. While this could be ignored, most good-quality preamplifiers offered some form of switched-frequency, steep-slope, HF low-pass filtering to reduce it. There was also usually a steep-cut, 10-50Hz high-pass filter to reduce the rumble caused by poor turntable bearings in the equipment used for the cutting and replay of the discs.

For treble filter systems, there is a choice of active or passive circuit layouts; for attenuation rates of 12dB/octave or greater, passive LC networks of the type shown in Fig. 15 are normally used. However, the inductors needed for rumble filters would be inconveniently large and prone to mains hum pick-up, so active arrangements are preferred.

Sometimes the rumble filter function would be incorporated within the RIAA equalisation network, as in earlier circuits by the author³, Dinsdale¹⁴ and Bailey¹⁵, whose circuit layout is shown in Fig. 16, but such filter stages are now commonly arranged as separate, cancellable, control blocks using one or other of the layouts shown in Fig. 17.

There is a necessary compromise in the choice of filter slope adopted, in that too low a slope is likely to be ineffective, whereas high attenuation rates will cause some degree of audible coloration, which can be particularly notice-

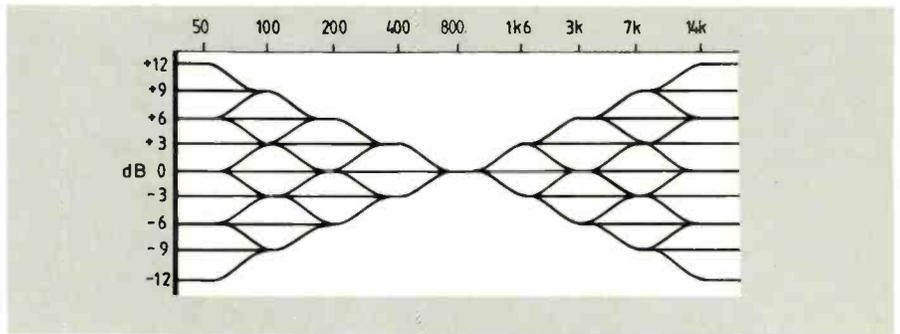


Fig. 11. Author's RC tone control, called the "Clapham Junction" circuit, for obvious reasons

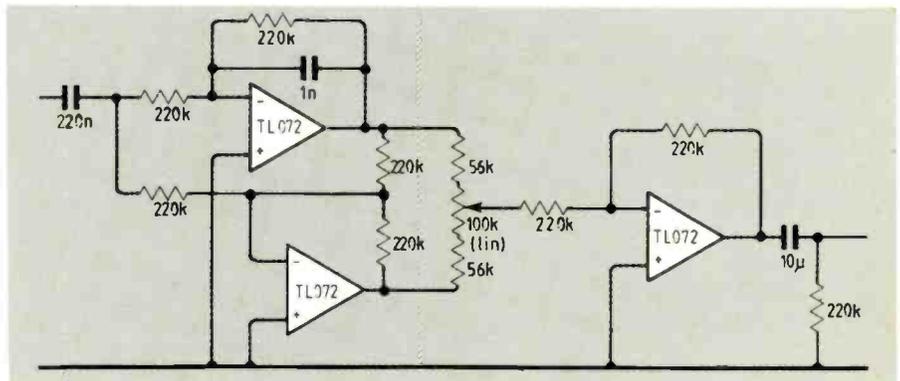


Fig. 12. Tilt or slope-adjustment tone control, which varies slope continuously from low to high frequencies

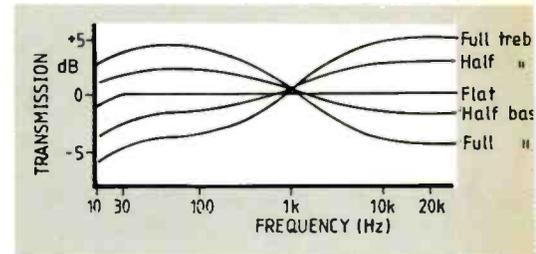


Fig. 13. Adjustments of Bingham's slope control circuit in Fig. 12

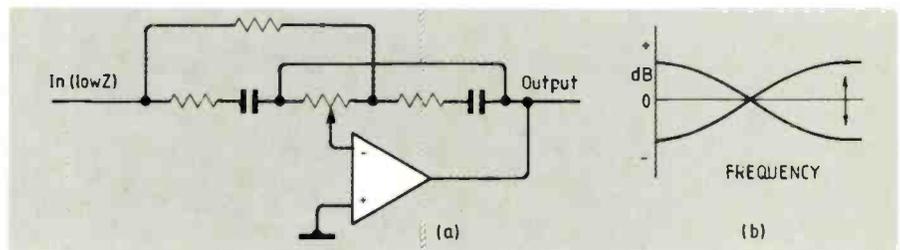


Fig. 14. Quad's slope control at (a) and its characteristic (b)

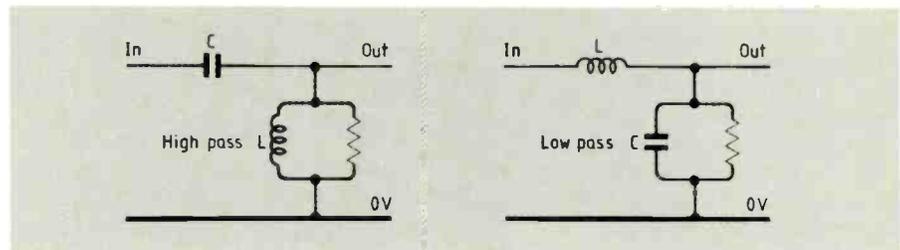


Fig. 15. High-pass and low-pass filters for hum and rumble rejection at 12dB/octave

able in the treble. Current practice is to retain some optional rumble filtering, since any coloration in the bass is likely to be lost among the room and loud-speaker cabinet resonances, which are prominent in this part of the spectrum.

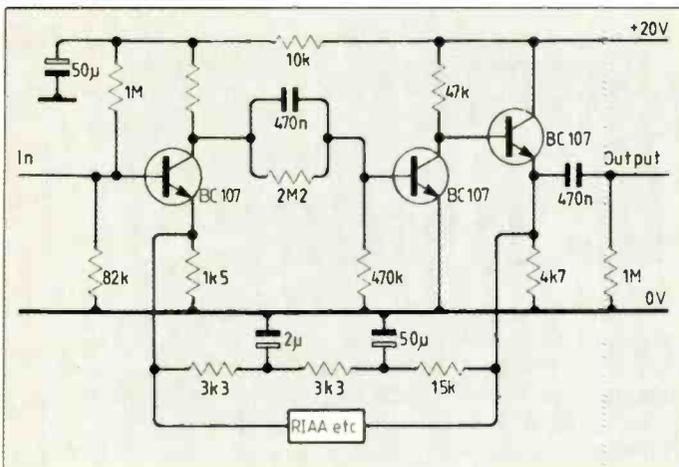
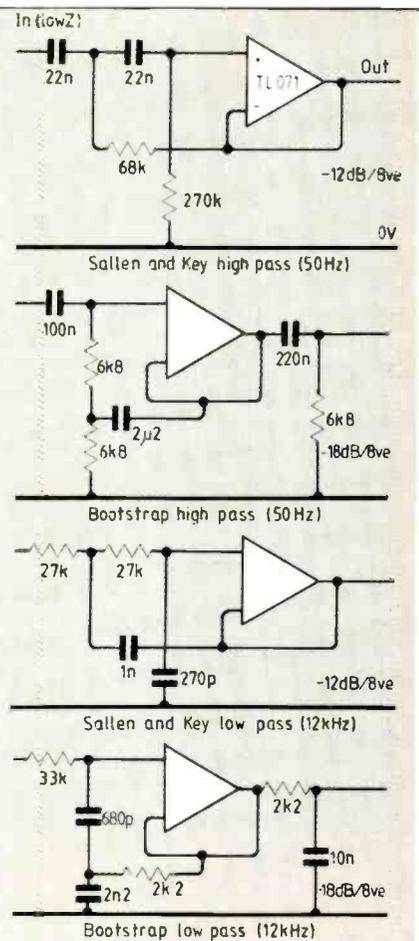


Fig. 16. Rumble filter incorporated in RIAA equalisation stage in circuit by Bailey (1966)

Fig. 17. Unity-gain filters now used as separate blocks which can be cancelled if required



In the next part of this article I will look at ICs and other gain blocks, input switching systems, stereo image-width controls and power supplies. ■

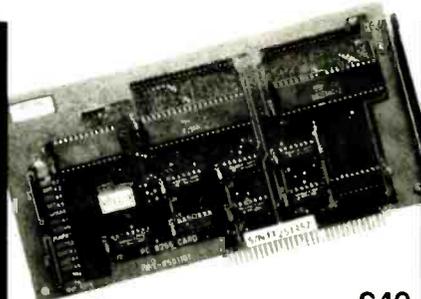
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Gain reduction	2:1
Limiter attack time	13ms
Limiter recovery	1.5ms/dB
Limiter gain reduction	4.6:1
Frequency response (90Hz-20kHz)	±0.5dB
S:n ratio at 0dB	100dB
THD+noise	0.05%

Expander	
Input voltage range	-10dBu
Attack time	26ms
Recovery	3ms/dB
Gain ratio	1:2
Frequency response (20Hz to 20kHz)	±1dB
S:n ratio	100dB
THD+noise	0.2%

Performance of compander using Precision Monolithics SSM-2120 dynamic range processor

Precision Monolithics is at 90 Park Street, Camberley, Surrey GU15 3NY. Telephone: 0276 692392.

The 100kΩ resistor from V_{CC} to pin 10 of U₃ establishes the operating current for the VCA. To keep power supply current to a minimum, all pins of the unused VCA should be returned to ground. Feedthrough trim is optional and can be used to minimise the VCA control voltage feeding through to the output.

Protection limiter

This uses the second rectifier and control amplifier for separate and independent attack and decay times, along with a steeper gain-reduction slope. A threshold control sets the predetermined gain-limiting point for high input-signal levels. Gain reduction ratio is 4.6:1 and typically the onset of gain limiting should be set to +10dBu at the output.

As in the compressor control circuit, the rectifier input current is limited by R₆ and the rectifier also referred to 10μA. Lower precision capacitors and resistors can be used in this case and, as with the compressor, the attack time is much faster than the decay.

The VCA/preamplifier was designed as a system: the VCA was put in the signal feedback loop of the preamplifier principally to prevent preamplifier overload, while keeping the overall noise low, and minimising component count.

Power consumption

The application circuit requires two power supply voltages, ±9VDC; power consumption of the circuit shown in Fig.

1 is less than 15mA from each supply. The design described will operate properly with good dynamic range as the battery voltage begins to fall below the nominal 9VDC. It is assumed that two 9V batteries would be used, but for the smaller hand-held wireless microphones, a single 9V battery would be sufficient. Figure 2 depicts a DC-to-DC converter that will supply the -9VDC.

The converter circuit incorporates an astable oscillator running at 25kHz, which is followed by a capacitor-coupled level shifter and rectifier with a filter. An SE/NE555 timer is used in the "output sink" mode for maximum efficiency and longest battery life.

Receiver expander circuits

Figure 3 shows the control connection of the SSM-2120 (U₃) VCA, rectifier and control amplifier. The control circuit connection to the VCA produces a 1:2 gain expansion curve: if the input rises by 3dB, the output level will rise 6dB. Gain expansion ratio is:

$$= \frac{R_2 \cdot R_4}{R_1 \cdot R_3}$$

The rectifier input current is again limited by a 10kΩ resistor connected to pin 9 of U₃, and the rectifier is biased at 10μA current through a 1.5MΩ resistor connected to V_{EE}. The SSM-2120 rectifier and VCA each have a 100dB dynamic range, resulting in accurate tracking of the compressor.

Small-signal averaging time, attack time and decay rate are identical to those of the compressor/limiter circuit.

Control-circuit gain values provide a control voltage to the VCA section + V_c

Fig. 3. Expander circuit for receiver processor.

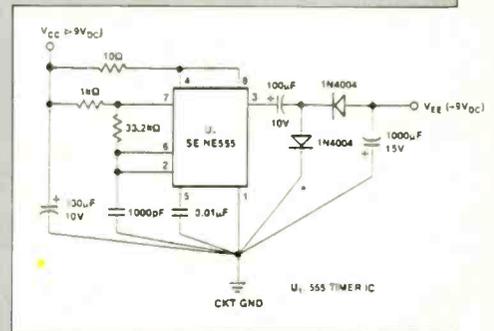


Fig. 2. DC-to-DC converter for ±9V DC at 15mA

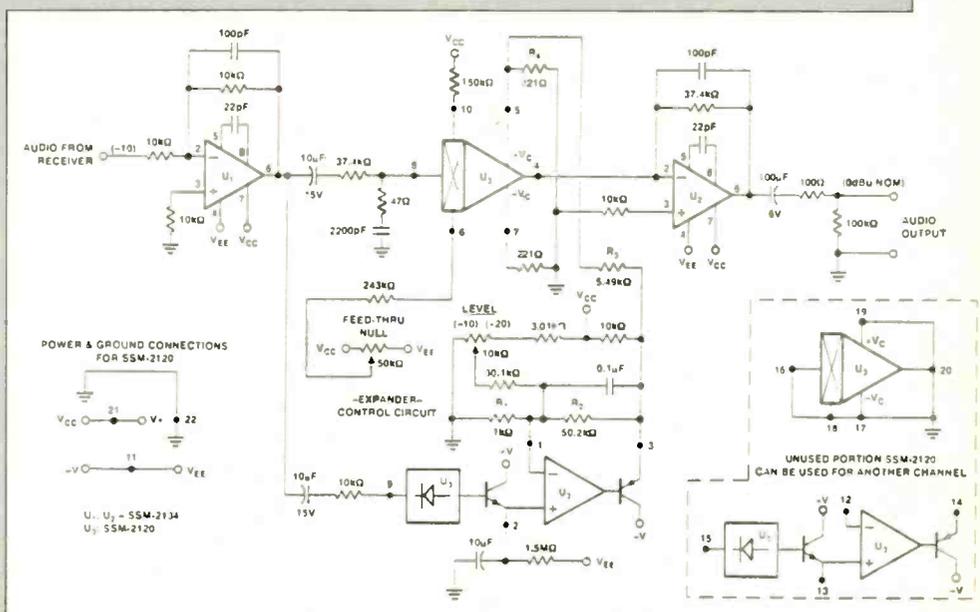
control port (U₃, pin 5), which results in a 1:2 signal expansion characteristic. The level control sets the initial overall gain value and is adjustable from -10dB to +20dB.

Audio signal source

Input amplifier U₁ acts as a buffer between the input signal source (FM receiver), the expander rectifier/control circuit, and the VCA audio signal input. If the signal-source output impedance is below 100Ω, U₁ can be omitted.

The nominal source signal level should be -10dBu; if signal gain or loss is required, U₁ gain structure should be modified to provide -10dBu to the VCA input-current limiting resistor. The 37.4kΩ resistor ahead of the VCA input limits peak signal currents to avoid VCA distortion.

A 150kΩ resistor to V_{CC} sets the VCA input/output current compliance range and a VCA input shunting capacitor minimises signal distortion and keeps the VCA stable by providing a high-frequency path to ground, the exact value being determined empirically. ■



HDTV

is there a consumer market?

Much effort and a significant amount of both private and public (European Community) money continues to be devoted to the development of the compatible HD-MAC 1250/50/16:9 HDTV system capable of being transmitted to the home over a standard 12GHz DBS channel. As made clear in the recent ICC colloquium Image processing for HDTV, good progress is being made with the help of some extremely sophisticated electronics. But there remains the long-term problem for both HD-MAC and the Japanese Hi-Vision 1125/60/16:9 systems that nobody is prepared to guess if or when there will be a consumer-budget display device capable of doing real justice to the resolution possible with HDTV – or indeed what would be the gamma of such displays if based on solid-state technology. Then, again, a year or two back it was suggested that it would be essential to produce HDTV sets that could be sold for about £1,500, a difficult target indeed.

So at present HDTV *broadcasting* can hardly claim to be a market-led technology. There are still those non-engineers would believe that HDTV could turn out to be Concorde of television: a technological triumph, a marketing blind-alley. The pictures produced by 625-line PAL are still capable of considerable improvement and wide-screen 625-line MAC pictures do not leave much to be desired for home viewing. HDTV *production*, however, offers considerable scope for generating very superior master tapes as the basis of broadcast, video or electronic cinematography programmes/cassettes.

Despite the important BBC contribution of DATV (digitally assisted television) that now forms an inherent part of the HD-MAC system, the BBC's £4-million co-production of *The Ginger Tree* was electronically shot on 1125/60/16:9 in Japan, Taiwan and the Isle of Man by drawing on the facilities of NHK and Sony Broadcast, with the tapes standard-converted in Tokyo for 625/60/4:3 transmission in the UK. Producers are enthusiastic about HDTV.

BSB have promised that their film channel will transmit wide-screen (16:9) pictures, compatible with existing 4:3 screens, from the start of service, encouraged by research which shows that viewers prefer wide screens and would be prepared to pay a premium for wide-screen sets.

One of the first attempts to conduct research on the reaction to dynamic wide-screen pictures has been reported by Karen Pitts and Norm Hurst of the David Sarnoff Research Center ("How much do people prefer widescreen (16:9) to standard NTSC (4:3)?" *IEEE Trans on Consumer Electronics*, August 1989, pp160-169). With 180 non-export viewers, the researchers found a strong, consistent preference for 16:9 pictures,

even when this meant some reduction in the height of the display.

The only variable to influence this preference was the comparative size of the images. But even with equal width pictures (i.e. with the height of the wide-screen pictures only 75% that of the 4:3 pictures) there was still a 72:28 preference for wide-screen. (A number of European broadcasters broadcast films in wide-screen format, leaving the top and bottom of the screen blacked out, though this is seldom done in the UK except for the opening titles.)

With a moderate decrease in the height of the wide-screen picture (equal diagonals) there was still a preference for wide-screen images 90% of the time. None of the other variables – size of 4:3 set, viewing distance, male-female, seat location – had any significance, although there were some variations with picture content.

Based on a 4:3 set price of \$400, viewers indicated that they would be prepared to pay an extra \$80 for equal-width 16:9; \$90 for equal diagonals, \$100 for equal area, and \$140 for equal height pictures. For equal height pictures the widescreen premium thus amounts to 35%.

For the purpose of this research, pictures were presented side-by-side on Mitsubishi 35in direct-view monitors with suitable masking to represent either 26in or 20in 4:3 screens.

In Japan, the emphasis seems to be shifting away from the non-compatible 1125/60 component standard, except as a studio production standard, towards NTSC-compatible transmission systems based on a growing family of MUSE image - processing / bandwidth-compression techniques. As well as the full Hi-Vision HDTV, there is now "Clear Vision" based on 1125/60 material, transmitted in an NTSC-compatible format, but which permits the receiver to display each field twice and which includes a reference pulse which permits the receiver to eliminate by correction multipath "ghost" signals. Clear Vision receivers sell for about one-third of the price of Hi Vision, although even so are at a significant premium to current NTSC sets (by a factor of about three).



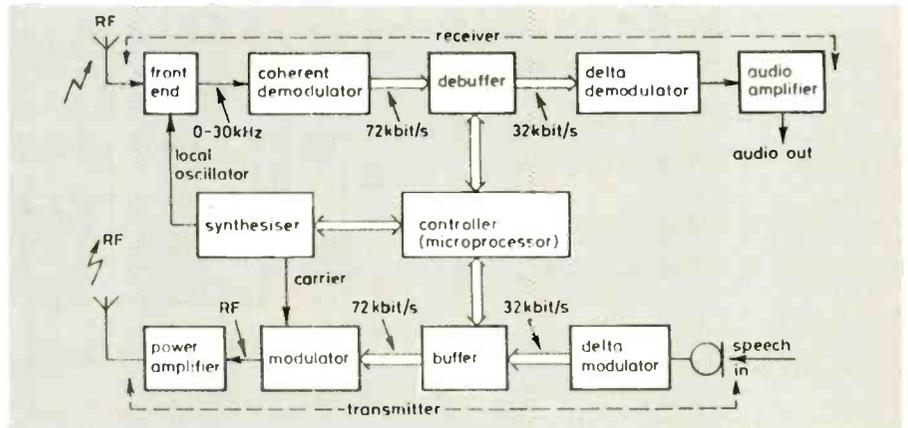
Off-screen shot from a compatible HDTV system: GE, NBC and RCA invented a 6MHz NTSC compatible standard (ACTV). Full bandwidth, 16:9 signals occupy 12.4MHz of luminance bandwidth while standard NTSC sets can decode and display the centre section of the image without modification.

Taming the digital waves

Among the projects that the BBC Design & Equipment Department hopes to sell to industry is a digital audio link suitable for either a single stereo pair or for the six-channel (three stereo pairs) Nicam-3 distribution system in use for several years on BBC Radio. An interesting feature of the new link system, which is now in use as a low-power UHF link between Stockland Hill, Devon and the Channel Islands, is the use of tamed frequency modulation (TFM). This is a form of fast frequency-shift keying in which the abrupt phase changes are smoothed out, resulting in high spectral density and hence an outstanding degree of spectrum utilization.

TFM was originally proposed by Philips engineers in 1978, primarily for digital mobile-radio communications. They showed that digitally-controlled delta modulation (DCDM) speech at 16kbit/s can be transmitted using TFM in 25kHz radio channels, with the power radiated in the adjacent channel some 85dB lower than that in the allotted channel, substantially lower than for other constant envelope modulation techniques. With synchronous detection, there is a penalty of only 1dB in error performance compared with standard QPSK digital modulation.

Details of TFM as presented by Frank de Jager and Cornelis Dekker (*IEEE Trans COM-26*, May 1978) and by D. Mulwijk (*Philips Telecommunications Review*, March 1979) indicated that the power density per bit rate at the edge of the adjacent channel can be -67dB compared with -42dB for filtered QPSK, which has the disadvantage also of requiring linear amplification, -22dB for phase-shaped (Nyquist



Block diagram of a 902MHz cordless telephone handset using TFM. Modulation system details are shown at the bottom of page.

pulse) QPSK and -14dB for minimum-shift-keying. Conventional 16kbit/s QPSK signals would have excessive bandwidth for use in 25kHz channels. In effect, TFM can meet the 70dB out-of-band selectivity values applied to analogue FM transmission.

The basic principle of TFM is to achieve proper control of the frequency oscillator, such that the phase of the modulated signal becomes a smooth function of time with correlative properties. The BBC modem operates at data rates of 676kbit/s (Nicam stereo channel) or 2048kbit/s (Nicam-3 six-channel distribution system for three stereo pairs) on a range of carrier frequencies between 7.5 and 70MHz and can be multiplexed with video links. In one current use, a BBC OB contribution link from Crystal Palace to Broadcasting House London uses a coaxial cable which also carries base-band video and carrier-video. An "over-video" system is being developed for the tele-

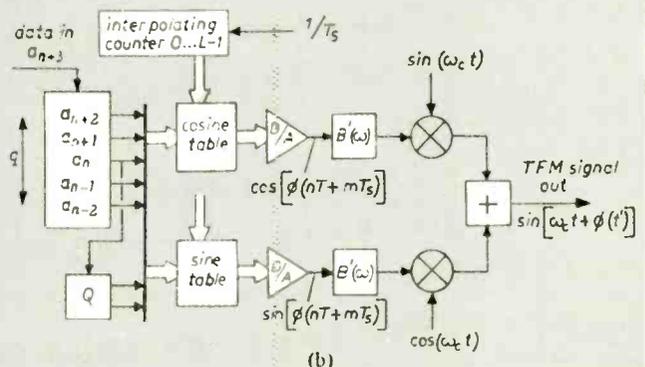
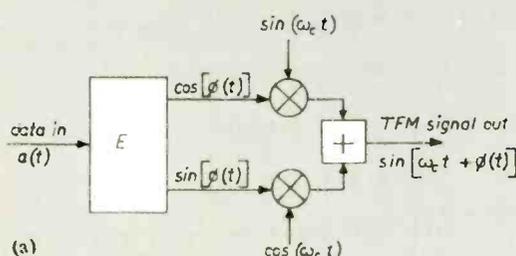
vision service.

The advantages of TFM for digital communications were also explored in a paper by Dr P.D. White and a team of UK Philips Research Laboratories, (*IEE Proc.* Vol 132, Pt F, No 5, August 1985): 900MHz digital cordless telephones. This used TFM for short-range cordless telephones, with the speech coded as 32kbit/s continuously variable slope delta modulation or adaptive differential PCM with a final data rate of 72kbit/s for time-division duplex modulation. This could be conveniently transmitted with a 100kHz channel spacing, with a spectral efficiency rather better than that of comparable analogue systems.

As a digital modulation technique, TFM appears to provide good spectral efficiency for a surprisingly small penalty in performance, though like all digital signalling systems is vulnerable to multipath propagation.

Broadcasting is by Pat Hawker

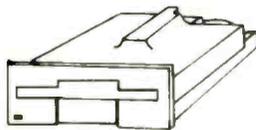
Block diagram of the the BBC's Tamed Frequency Modulation system used to distribute digital data over FM transmission links. It aims to smooth out the phase changes thus distributing the energy evenly over the available passband. Digital voice data at 16kbit/s occupies a standard 25kHz channel with adjacent products 85dB below. Drawing (a) shows the basic layout while (b) is the expanded form. (Philips 1978).



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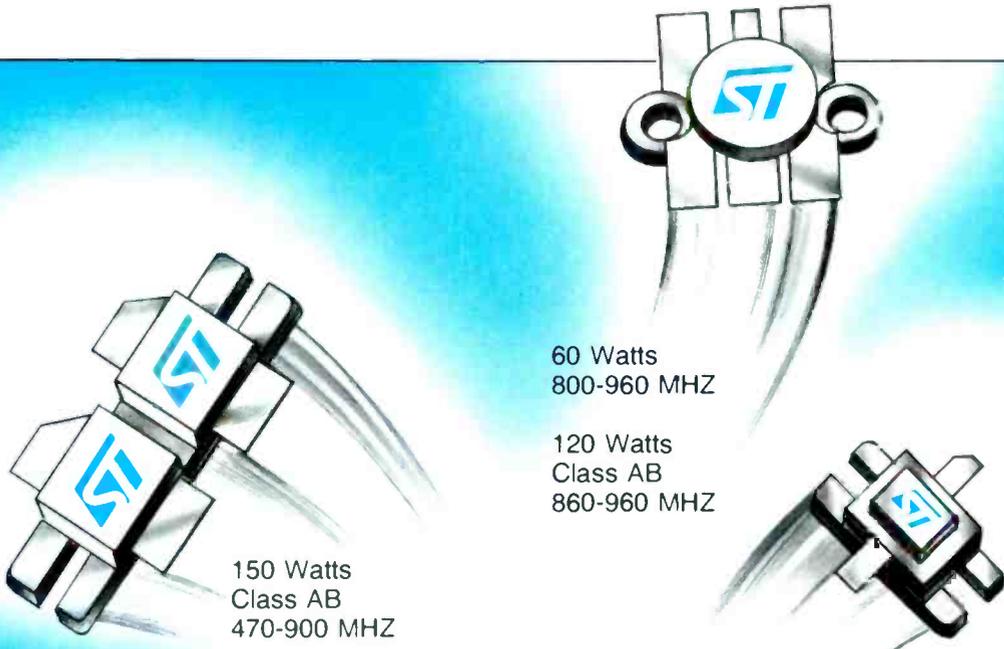
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