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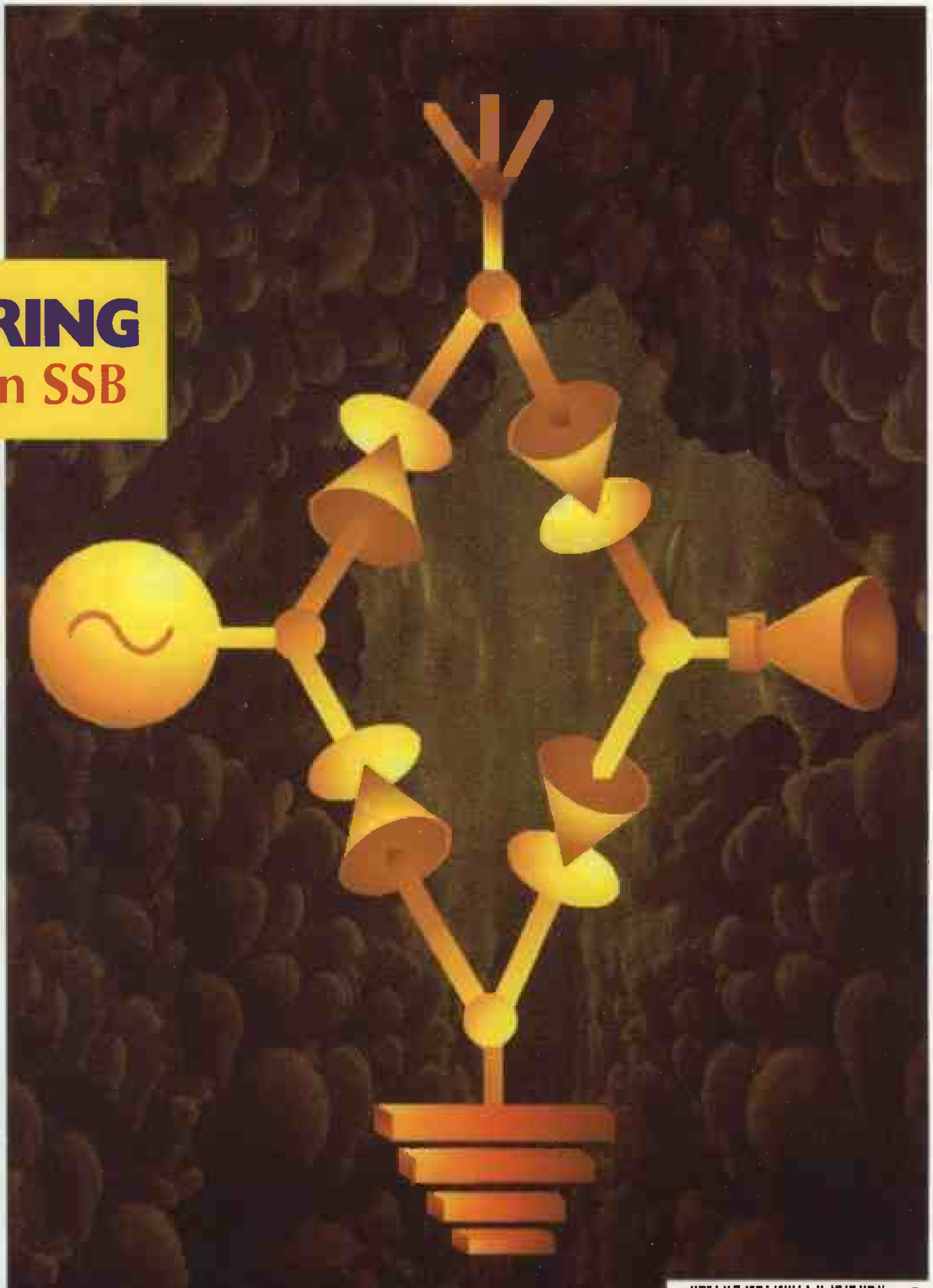

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# Right to eavesdrop?

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Frank Ogden
081-652 3128

**DEPUTY EDITOR**
Jonathan Campbell
081-652 8638

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Derek Rowe

**DESIGN & PRODUCTION**
Alan Kerr

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Lorraine Spindler
081-652 3614

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081-652 3732

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Most people reading last month's lead news story 'Government bans sale of unbuggable phones' will have felt a degree of pleasure at the thought of Big Brother deprived of the ability to snoop on its citizens.

To recap, the DTI has vetoed the incorporation of an advanced encryption algorithm into the new GSM portable phone standard because conversations made using the system will be unable to be decoded by parties other than the intended recipient, at least in real time. This has alarmed both the FBI in the US and our own intelligence services who fear that their targets will be able to communicate freely through relatively secure channels.

A direct result of the veto is that manufacturing and export plans for the new phone system are being delayed or postponed. Although this allows other Far Eastern digital cell phone systems to gain competitive advantage, the real questions seem to concern whether governments should snoop on their citizens.

The instinctive reaction is to applaud any development which reduces a State's ability to interfere yet this seems to be at odds with other, equally instinctive attitudes. For instance, most people would accept that car number plates perform a socially useful function. Similarly, they may be used in evidence when prosecuting dangerous drivers.

Slightly removed but equally acceptable would be the use of number plates an identifier where a car is used in the commission of non-motoring crime.

Continuing the motoring analogy, consider the roads themselves. Anyone who drives in a law-abiding fashion may use them unmolested for their own purposes. The police and security services regularly track target vehicles along a route but most of us wouldn't know and couldn't care.

Likewise the public phone system. The majority of people do not give a second thought about the possible eavesdropping by other parties on a private conversation. In reality, the public phone system is just that: public. Police and security services intercept at will, with or without a court warrant yet most of us lose no sleep over this.

On the contrary, we welcome the proper use of intercept in bringing criminals to book and the placing of terrorists behind bars. It is hard to criticise the use of the phone system's digital technology to facilitate this: for instance the special arrangements on call box lines, the ability to set up a remote intercept with a few keystrokes, the terminal identifiers which are as useful as a car number plate.

The Government has made a mistake in creating a fuss over encryption standards for portable phones. It has simply reminded criminals of the dangers of using the public phone system. They will make a special effort to conduct their business in a way which can't be intercepted by the Government or anyone else.

No parliamentary committee could ever eliminate the occasional abuse of technology by those who are supposed to act in our interests. However, official surveillance systems are, on balance, a good thing in a democracy. Our principal goal should be to increase its strength. A Bill of Rights and a Freedom of Information Act seems to provide the best check against the routine abuse of official surveillance.

**Frank Ogden**

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# Radio chips star at US convention

The world's chip companies look for a quickly growing market in chips which go into pocket telephones and PC-based wireless modems. The problem for the chip-makers is how to provide highly integrated, high frequency chips which operate at low power.

ISSCC '93 covered both sides of radiocomms – the baseband signal processing functions, and the IF/RF modulation functions. Two papers covering the IF and RF modulation and demodulation functions came from National Semiconductor and AT&T.

The National paper described a fully integrated bipolar or cmos front end with built-in image rejection on a monolithic chip. The use of a quadrature two-phase mixer to implement the RF to IF frequency conversion greatly simplifies the receiver hardware by eliminating the costly, bulky filters required with a single mixer.

The 2.5GHz image reject front end, consists of an LNA, complex mixer, bias and power-down stages, and contains on-chip all the required elements including the mixer phase shifting circuitry. The chip consumes 60mW in a 15GHz, 0.8μm bicmos process technology. At 1.89GHz with a 111MHz IF signal, image rejection is 14.1dB, conversion gain 7.6dB and the noise figure is 18dB.

The front end comprises an RF amplifier, an image reject mixer, three phase shifters, output, bias and power-down circuits. The image-reject mixer is composed of two Gilbert-cell mixers. The RC phase shift circuits are connected to both mixers. One phase shifter performs the image rejection while the other loads the other side of the mixer. The dummy section ensures the mixers are loaded equally increasing image rejection by 8dB.

A second paper on the IF/RF function came from AT&T. It described a two chip 900MHz transceiver for the North American IS-54 dual-mode cellular telephone standard.

One chip is a direct up-conversion modulator. The other is a double conversion receiver. Both are made in 12GHz bipolar technology. As the modulator performs

*Continued over*

## BT in touch with virtual reality

British Telecom is developing virtual reality techniques to help in repairing electronic circuits. The picture shows Melanie Collins, a BT technician, at the end of a virtual reality link being instructed by an expert in the repair of electronic equipment.

BT is working on four virtual reality projects – telepresence, data visualisation, emotional icons, and controlling a communications network. Telepresence applications include mobile teleworking between head office and service centres, directing medical operations from remote locations, and electronic news gathering.

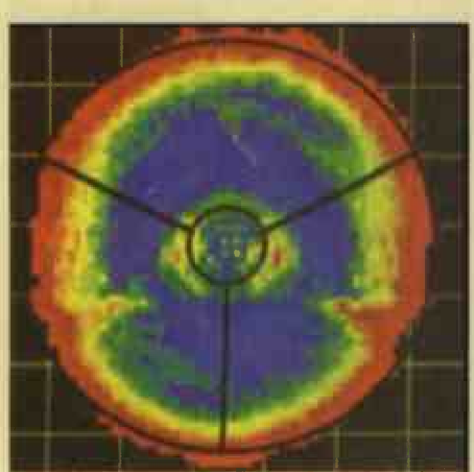
Visualisation techniques can be used, say, to build up a picture of lightning strike patterns as they affect a telecommunications network. Emotional icons can be used to provide a humanised interface with data in network control applications and can also be linked to artificial intelligence to assist decision making.

## Microwave focus

Dish designers can now use a microwave holographic modelling technique to check the phase integrity of their aerial systems.

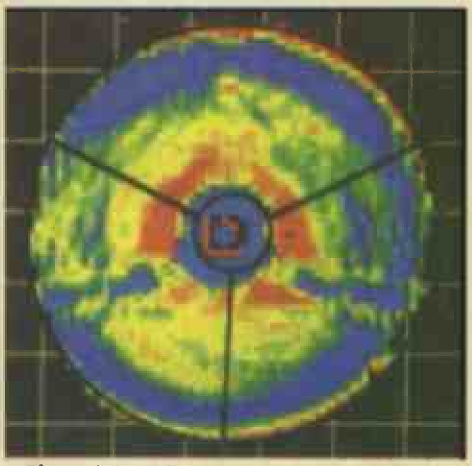
Developed by ERA Technology, these techniques produce a holographic phase map which creates three dimensional models of the reflector's amplitude and phase profile.

The maps help to identify feed and reflector misalignments which can be catastrophic for the large reflector antenna and phased array radars used in military and civil communications. The system, which involves calculation of the aerial aperture field in both complex far-field and near-field patterns, will run on a variety of software platforms including dos. Contact Chris White, ERA, 0372-374151.



Amplitude (0>–20dB)

Phase (180°>–180°)



Phase (0°>70°)

*ISSCC CONTINUED*

direct up-conversion, there is only one local oscillator and no intermediate frequency filtering is necessary. The inputs to the modulator are the local oscillator, and balanced I and Q baseband signals.

The local oscillator input is buffered and sent to a phase splitter to produce two signals in quadrature. These are then mixed with the I and Q baseband signals to produce a single sideband suppressed carrier RF output signal. The modulator is designed to avoid generating spurious signals. The –26dBc adjacent channel and –45dBc alternate channel noise requirements of IS54 are met.

The receiver chip uses two intermediate frequencies with the second made up of a dual path; one for analogue FM and the other for digital DQPSK demodulation. A RF front end amplifies and demodulates the RF band to the first intermediate frequency at which the first channel selective filtering is performed. The RF signal can be attenuated by 26dB, preventing strong wanted channels from overdriving later stages. Operated from 5V the total power dissipation of the set is 310mW.

Combined silicon area is 9mm$^2$.

Two ISSCC papers covered baseband signal processing functions. In the first, Motorola described a DC-coupled compander for reducing background noise levels in cordless telephones. The second, from AT&T, outlined a single chip incorporating transmit, receive, a DPSQPSK modulator, two A/D converters, four D/A converters and associated analogue and digital filters.

A new generation of VLSI technology – the quarter micron generation – was demonstrated at the '93 ISSCC along with the 256Mbit drams that are the fruit of it. But, it seems, even dram designers are having to move up the design hierarchy from designing at the cell level to designing in blocks.

**Memories.** One of the design features of Hitachi's 256Mbit dram was a subarray replacement redundancy technique. Conventional redundancy techniques of replacing defective lines by on-chip spare lines will, apparently, no longer be sufficient for drams of 256Mbit and beyond. Instead a subarray-by-subarray replacement technique has to be used. The downside for using the subarray technique is an area increase of 3.5%. The upside is that using the technique should double yields.

Another key design issue for Hitachi's 256Mbit was how to reduce the data-retention current (comprising both refresh and standby current). Although low voltage operation is effective in reducing refresh current, it causes an increase in standby current due to the subthreshold current of deep-submicron MOS transistors.

Hitachi solved the refresh current problem by reducing the operating voltage from 3.3V to 1.5V. The standby current problem as been addressed by inserting a switching pmos transistor between the wordline voltage and the driver transistors' common-source terminal which limits the subthreshold current to the driver transistors in the standby state. The result is that the total data retention current of a 256Mbit can be less than that of a 64Mbit.

**Analogue systems**. Analogue technology, it seems, continues to push away at the frontiers of active filter technology and three analogue papers at this year's conference demonstrated the design limits in low distortion, high operating frequency and low operating voltage. Massachusetts Institute of Technology demonstrated a 2V cmos op amp where the chip swings very close to either rail with a gain bandwidth of 63MHz. An enhancement technique provides gains of more than 10,000 with operation down to 2V. The swing is symmetrical with the output going to within 100mV of either rail.

In a couple of sessions on emerging technologies, this year's ISSCC showed some interesting hints of where microelectronics technology could be heading. Two papers looked at mixing microelectronics and micromechanics to make smart micro-machines, and an MIT paper discussed extending micro-machining into binary optics.

Tohoku University described the use of multiple wavelengths to extend on-chip parallel computing into a third dimension. A breakthrough potential in nanoelectronics in the use of resonant tunnelling transistors and

**Continued over**

## Manufacturing under the microscope

The UK manufacturing industry has come under the spotlight in a study being carried out by the London Business School and IBM.

Researchers are going round the country carrying out detailed interviews with more than 200 companies in the first part of a major benchmarking exercise for manufacturers.

Its aim, according to Phil Hanson from IBM, is "to answer the questions that have dogged UK industry for years: Is it behind the rest of the world? Are manufacturing practices outdated? How far are we from being really world class?"

The plan is for comparable studies to be made in Germany, Japan, and the USA in the next two years. Manufacturing sites of all sizes are being visited.

The interviewing stage will be completed this month (April) and results of the *Made in Britain* survey, as it is being called, will be published before the end of the year.

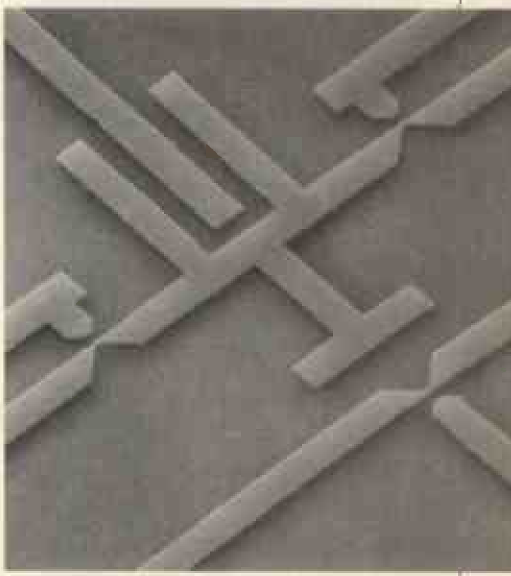
## Single electron may open the door for terabit chips

An experimental circuit could open the way for the first terabit chips. The circuit has been jointly developed by scientists at Cambridge University and Hitachi in Japan.

In the quest to develop a single electron memory cell, they have built a circuit about 0.03mm across with minimum feature sizes of one micron. It stores a charge of about 100 electrons.

The researchers believe the circuit demonstrates the possibility of building smaller true single electron memory cells. Hitachi has already started on work this using a scanning tunnelling microscope. The circuit was described in the 18th February 1993 edition of *Electronics Letters*.

Hitachi expects to see the first working device within two years.



The cell already developed uses two multiple tunnel junctions to control the flow of electrons along a conductive path leading to a capacitor memory cell and a charge sensing detector. The junctions write and erase the memory cell by squeezing a conductive path with an electric field so that electrons are allowed past one at a time. Changing the field strength can turn this current on and off.

As conventional semiconductor technology improves with smaller line widths the problems from quantum mechanical effects may make it impossible for devices to accurately control the small number of electrons that run through the circuit.

The researchers hope that the technology they are developing will be able to control individual electrons paving the way for memory chips with capacities more than 64Gbit. **Leon Clifford,** *Electronics Weekly*

*ISSCC CONTINUED*

diodes was the subject of a paper by Texas Instruments.

**Fuzzy logic**. In the field of neural networks a digital special-purpose hardware accelerator for a model of a biological neuron was the subject of one paper while two other neural nets based on a new device called a neuron mos transistor were described. These devices feature multiple control electrodes capacitively coupled to the gate of the device. Tohoku University described how a logic gate made from a neuron MOS transistor can act as an AND, OR, XOR, XNOR or INHIBIT function by setting voltages on the various control electrodes.

Two papers took fuzzy processing used in cameras into automotive engine control. A Siemens paper described a fuzzy logic coprocessor small enough to be a macrocell in an asic while retaining the ability to process inferences quickly.

Built in 1µm cmos, Siemens' fuzzy controller module includes ram within a chip area of 6.4mm$^2$. Rom is on-chip, 64Kbit in this case but the amount depends on application. The device has a maximum clock speed of 20MHz, and takes 126.5µs to compute a result from four inputs and 1000 rules at 10MHz.

The user can choose one of eight fuzzy algorithms: the aggregation of weights in the rule evaluator can be performed using the MAX operation or the Bounded Sum method; the inference unit computes the possibility distribution of the output (the fuzzy output) with the min-max or min-bounded sum methods; the defuzzification or calculation of a discrete output value is performed using either the centre-of-gravity or mean-of-maxima methods.

The host microcontroller communicates with the fuzzy controller via an 8-bit address/data bus. In the development system, the knowledge based memory (KBM) is a separate device that stores the rules, the membership functions of inputs and outputs, and the knowledge-based descriptors. In final implementation the KBM can be implemented in on-chip rom as well as external rom devices.

**David Manners,** ***Electronics Weekly.***

## ERA sees large impact from neural nets

Artificial neural networking will have considerable impact in the next decade on control, signal, and image processing, according to a report from ERA Technology.

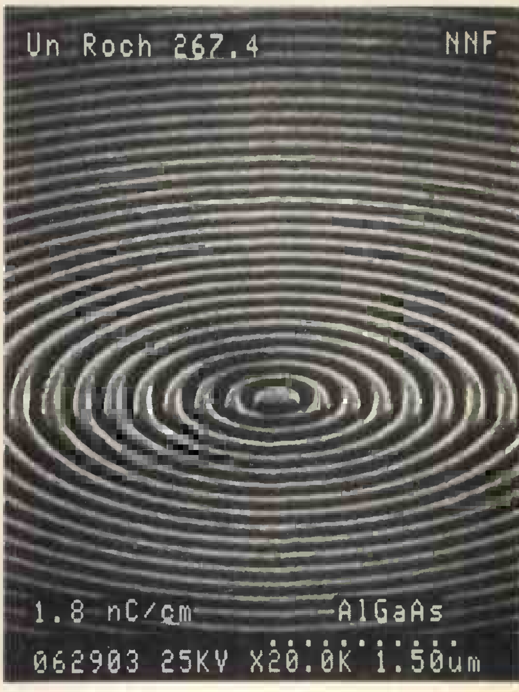
The report looks at recent applications in speech, handwriting, credit rating, and the control of vehicles and robots. It also describes a practical demonstration of neural networking as applied to fabric inspection and ground probing radar.

Neural computing attempts to process data in a similar fashion to the human brain. These networks are trainable and do not need complex computer programs.

ERA says that this can lead to solutions to problems in machine vision, linear control of non-linear plant, pattern recognition, and modelling chaotic systems.

# Scientist find a way round the oval

The first semiconductor laser to produce a low divergence circular beam of light has been developed by scientists from IBM, Rochester University, and Cornell University. The beam can widen or diverge by less than half a degree, compared to 30° for a typical commercial semiconductor laser. From such a laser a point-like beam would be 58ft wide after travelling 50ft, compared with 10.5in with the new laser.




Semiconductor lasers produce an oval beam which causes problems when squeezed into circular optical fibre. Sometimes only 30% of the light gets into the fibre.

The new device is a 150µm surface emitting laser made of gallium arsenide. It has a concentric circle grating (see picture) made up of 600 grooves etched into the semiconductor surface. As the laser light fans out from the centre of the grating towards the grooves, the waves are deflected by the grating's ridges and interfere with each other, producing a coherent laser beam which is emitted from its surface.

A small portion of the light does escape from the sides, but this may end up being useful in developing laser arrays.

## Jedec plans 2.5V chip spec

The Joint Electronic Development Engineering Council (Jedec) plans to complete specifications for chips operating from 2.5V power supplies within the next year. Although 3.3V ICs have only just started breaking into the market, work has started on the development of the new standard that will further reduce the power requirements of portable equipment such as notebook and palmtop computers.

Key members pushing the development include Dec, Hewlett-Packard, Apple, Intel, AMD, and NEC.

According to Michael Pearson, chair of the 2.5V interface task group in Jedec's JC-16 committee, there are two main reasons for lowering the voltage. First, as ICs continue to shrink in size, the amount of power used must also be lowered to let them perform at optimum levels.

The second reason is to save battery life in portable devices and notebook PCs.

While the development of the 2.5V standard has been backed by several industry players, it is also meeting some resistance. Hitachi of America said that it may be a little premature to develop a new voltage standard because it wants to keep only two generations of drams using different voltages, and if there were more than a couple of choices, it would force Hitachi to support additional types of drams.

## EC helps women

The Engineering Council has published a guide listing initiatives to attract girls and women into engineering.

The guide has been published as part of the Wise (women into science and engineering) campaign.

Courses are listed for various age groups starting at 13 years old. Some of these provide girls and women with practical help. And some have hours to suit mothers with children at school or offer childcare facilities.

It is called *Awards, courses, visits* and is being sent to all secondary schools in the UK, careers services, and further and higher education institutions. Awards for girls and women are listed as are details of visits to engineering departments in education and industry.

Copies are available free from The Engineering Council, 10 Maltravers Street, London WC2R 3ER (send A5 sae)

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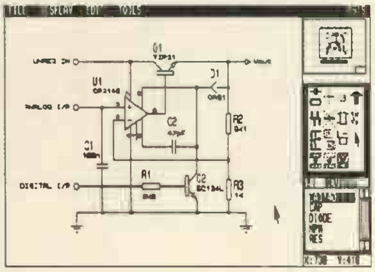

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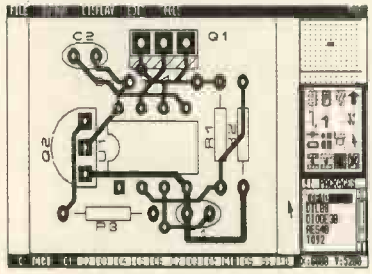

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# RESEARCH NOTES

## Harmonic way to find circuit faults

Greek researchers have taken another step nearer development of the self-analysing circuit by using the harmonic content of the power supply current to reveal critical information about circuit operation.

Some months ago we read of a self-testing cmos circuit that would measure its own power consumption and flag any obvious faults (*Op amp that tests itself*, Research Notes, *EW + WW*, October 92, p. 800). The approach allowed a clearly malfunctioning circuit to be identified and possibly switched out of use. Supply current measurements not only give a go/no-go indication, they can also indicate to some extent the nature of the fault, depending on whether the current is up or down and by how much.

The analysis has now been taken further in a recent paper by D K Papakostas and A A Hatzopoulos of the Aristotle University of Thessaloniki in Greece (*Electronics Letters*, Vol 29, No l). They have shown that even more information about the functioning or non-functioning of a circuit can be extracted by studying the spectrum of the power supply current – in other words its harmonic content – when the circuit is handling a test signal. The fundamental frequency in this case is of course determined by the frequency of the test signal

Their approach is based on the knowledge that $I_{ps}$ in most circuits is critically dependent on a whole variety of parameters, right down to parasitic capacitances. All faults, say Papakostas and Hatzopoulos, can be considered as changes in some branch currents that will cause more or less significant changes in the supply current.

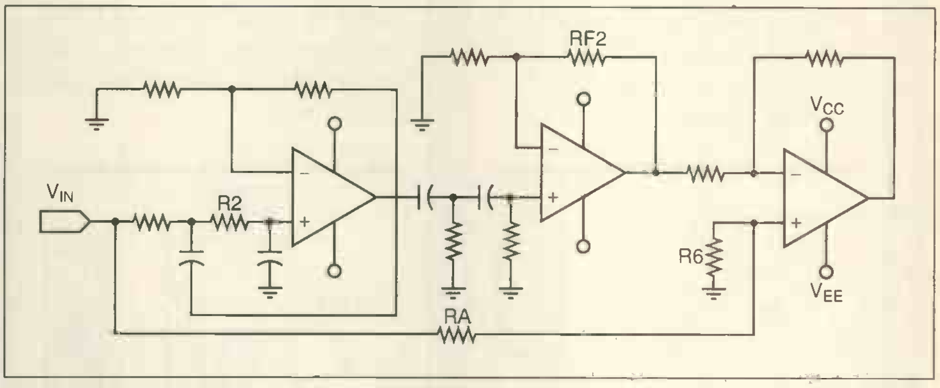
The Greek researchers have developed fault "dictionaries" using not just variations in RMS values of supply current, but also the values of the first five harmonics of its spectrum. Analysis of higher order harmonics was not deemed necessary.

An example of the circuits tested using the harmonic approach is the active filter. Using the *Spice* program, a total of 32 "hard" faults – 15 open circuits and 17 short circuits – were simulated. Shorts were defined as connections of less than 1Ω; open-circuits as more than 10Ω.

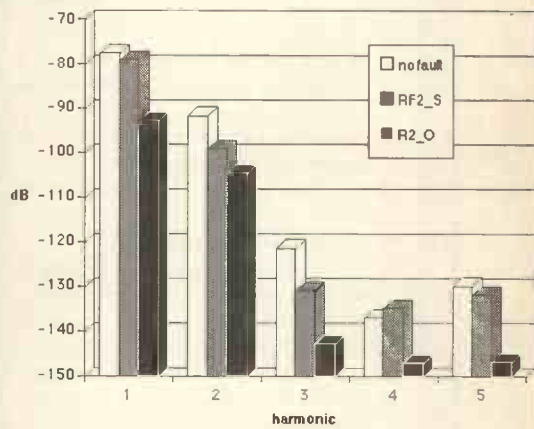
With only the basic RMS current value used, 15 simulated faults were spotted, though only three could be precisely localised. But using the enhanced fault dictionary, with details of spectral harmonics allowed, detection of all 32 faults and localisation of 30 of them was achieved – a vast improvement.

Take the example of resistor *RF2* in the diagram. If this resistor goes short or open circuit, the power consumption remains the same. But if the harmonic structure of $I_{ps}$ is analysed, the two fault conditions can very readily be distinguished from each other and from the normal operating condition.

The only two fault conditions which can be sensed but not precisely located, say the authors, are *R6* open-circuit and *RA* short-circuit.




***Active filter circuit demonstrates the effect of analysing power supply current.***




***Comparing harmonics for RF2 shorted (RF2_S) and R2 open (R2_O).***

## Darker side of the universe

Astronomers analysing data from the orbiting Rosat X-ray observatory have found evidence for huge amounts of mysterious so-called "dark matter" in and around small groups of galaxies. This discovery provides more evidence in support of current theories that the observable sky accounts for only 5% of the total mass of the universe. The rest is thought to consist of dark matter, so named because it radiates no energy and hence can not be observed directly.

Dark matter is important because its existence – or lack of it – places powerful constraints on how the universe will eventually end. If there were no dark matter, the expansion that began with the Big Bang some 15,000 million years ago would, according to modern cosmological theories, go on forever. On the other hand, if 95% of the universe does consist of dark matter, then its gravitational attraction would eventually bring the expansion of the universe to a halt or even reverse it, leading eventually to what has been dubbed the "Big Crunch".

The latest evidence from the Rosat team comes from X-ray images of three galaxies known as the NGC 2300 group, located about 150 million light years away in the direction of the constellation of Cepheus. They show that the galaxies are immersed

in a huge cloud of hot gas about 1-3 million light years in diameter.

The team believe that a cloud like this would have dissipated into space long ago, were it not held together by the gravity of an immense mass. The mass required to restrain the gas cloud is calculated to be some 25 times greater than that of the three visible galaxies themselves.

Dr Richard Mushotsky of Nasa's Goddard Space Flight Center says that one of the galaxies has a very strange shape, looking as if it is running into a wall. But examining an optical photograph shows no "wall" there to see. Yet something – presumably dark matter – must be exerting a very strong gravitational pull to hold the visible galaxy in its rather strange configuration.

Although this latest X-ray data adds to the growing body of evidence that the universe is pervaded by large amounts of this invisible dark matter, astronomers still do not have the slightest idea of what it might consist.

# Bright lights time for the optical computer?

A research team at the University of Colorado at Boulder has built what is believed to be the world's first general-purpose optical computer that stores its own program and processes information using light. It was developed under the direction of electrical and computer engineering professors Harry Jordan and Vincent Heuring.

Heuring emphasises that, while optical processing is not new, this present machine's ability to manipulate instructions defines it as an all-optical computer. Two years ago AT&T Bell Labs developed an optical processor capable of performing calculations with light beams. But that machine relied on electronically-held control programs.

The Colorado machine, described formally as a bit-serial optical computer, consists of a complex array of lasers, optical switches and optical fibres, about the size of a large desk. Laser beams are used to encode the computer's instructions and data into light pulses that are then stored in about 4km of spooled glass fibre. Information fed into one end of the fibre emerges 20μs later. At this rate about a thousand bits of data can be fed into one end of the fibre before one comes out at the other.

Describing this novel form of optical dynamic memory, Harry Jordan says: "For the first time we have a computer in which the program and the data are always on the fly in the form of light, eliminating the need for static storage."

Control beams from other lasers are used to route the light pulses from the memory through individual optical switches for processing. The machine's 66 optical switches – fabricated from lithium niobate by AT&T – can be turned on and off at microsecond rates to perform simple calculations. In addition to the main memory fibre spool, the computer's fibre network includes a number of shorter delay loops that also store data dynamically. All the light pulses are precisely timed with a master laser clock running at 50MHz.

The proof-of-concept machine is comparable in power to a small personal computer, though it is still a along way from being a marketable product. The Colorado team predict that although optical components will increasingly feature in hybrid commercial machines, an all-optical commercial computer is probably several decades away. They believe that the next generation of optical computers will have millions of switches interconnected through free space, using mirrors instead of fibres.

All-optical computers will eventually have a number of advantages over present-day electronic machines, the main ones deriving from the fact that photons – the units of light – can cross each other without mutual interference. Because of this, and because photons do not require wires for travel, an unlimited number of "soft" interconnections can easily be made. Light pulses are also faster and more predictable in their behaviour than electrical pulses and retain their shape better.

First applications for all-optical computers are likely to include super-speed graphics processing and telecomms switching.

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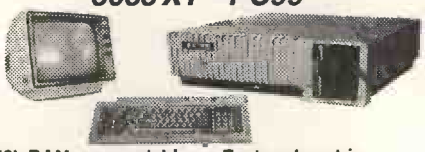

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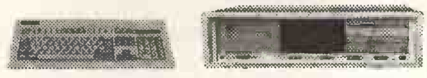

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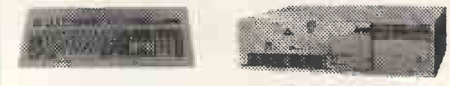

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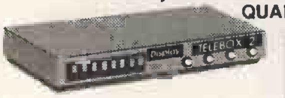

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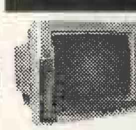

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# Thinnest wire in the world

Scientists in Japan have produced what is probably the thinnest metallic wire in the world – a mere 2nm in diameter. The development could be significant, both for fundamental science and electronics.

The breakthrough that made the latest development possible was the discovery about seven years ago of "Buckyballs", football-shaped molecules consisting of 60 or more atoms of carbon arranged in a lattice structure. Buckyballs, or Buckminster fullerenes to give them their formal title, were named after the inventor of the geodesic dome and have, over the years, proved singularly interesting research subjects. Chemists have used them to trap atoms, while physicists have made them both semiconducting and superconducting.

Just over a year ago, Sumio Iijima from NEC's Fundamental Research Laboratory in Tsukuba, Japan, discovered yet another interesting three-dimensional carbon structure whilst investigating buckyballs. By chance Iijima and his team had synthesised what have now been dubbed nanotubes. These are hollow carbon tubes, just over a nanometre in diameter, built from concentric graphite-like sheets.

Hot on their heels, NEC colleagues Thomas Ebbesen and Pulickel Ajayan developed a way of synthesising nanotubes in bulk. Now, following speculation that these tubes might be able to act as fine capillaries and suck up liquids – rather like super slim drinking straws – the original NEC team led by Sumio Iijima have demonstrated that this is indeed possible.

In *Nature* (Vol 361, No 6410) they explain how the nanotubes were opened up and made to suck up molten lead. The process involves electron beam evaporation of the lead onto the surfaces of freshly prepared carbon nanotubes. The samples were then heated to 400°C – above the melting point lead – and examination by transmission electron microscope showed that many of the nanotubes had acquired tiny nanometre-diameter lead cores.

Precisely how the process opens up the normally-closed tubes and allows the lead to enter is not yet known. The team speculate that it may be the result of a chemical reaction between the lead, the carbon atom tubes and oxygen.

Although the tube fillings appear to be a little disordered structurally – and there is as yet no certainty that these micro-wires consist of pure lead – the prospects are nevertheless exciting. Nanotubes may one day prove to be a practical means of manufacturing extremely fine electrical wiring, far thinner than anything that be created by etching or vapour deposition.

While lead is not the greatest electrical conductor, there is no reason why the technique can not be applied to copper or silver. It may also prove valuable for encapsulating materials that are not chemically stable when exposed to air.

From a theoretical point of view, nanometre-diameter wire can be expected to reveal some of the interesting properties associated with low-dimensional structures in general. Studies of quantum electronics too may be greatly accelerated.

# Study says chip-making puts babies at risk

Pregnant women working in chip fabrication areas are 40% more likely to suffer miscarriages than women working in non-fabrication areas. That is one conclusion to come out of the largest health study of representative samples of the US's quarter-of-a-million workers in the semiconductor industry.

Researchers at the University of California Davis School of Medicine led by Dr Marc B Schenker, a professor of occupational and environmental health, have also shown that the odds of women chip fabrication workers becoming pregnant are about 50% lower than women from non-fabrication control groups.

The precise reasons for these striking differences are not entirely known, but the UC Davis researchers and their collaborators say the findings suggest that exposure to photoresist or developer solvents – including glycol ethers – may be responsible for the higher rates of miscarriage.

In the three-year, $3.8 million study, researchers evaluated the health conditions of 15,000 workers from 14 company sites in seven states. The San Jose-based Semiconductor Industry Association funded the multidisciplinary study which was designed and conducted independently by the University of California Davis. ■

*Research Notes is written by John Wilson of the BBC World Service*

# SSB: third method, fourth explanation

*In spite of the occasional burst of enthusiasm in recent years, the direct conversion SSB receiver has been overlooked. And, in a neglected corner of a neglected subject lies the Weaver receiver. It deserves better. For the professional, it is a natural candidate for use with digital signal processing techniques. For the radio amateur, it is a receiver design that will be novel to most, and is fascinating to play with. By Nic Hamilton G4TXG.*

There are three ways of explaining how the Weaver receiver works in literature. Donald Weaver, in his paper *A Third Method of Generation & Detection of Single-Sideband Signals*[1] uses plenty of sin and cos maths. This is fine for mathematicians, but not so good for engineers. The *Radio Communication Handbook*[2] uses diagrams that resemble articles of evening dress. The method of explanation is reminiscent of a Victorian manual of etiquette on how to tie a bow tie. It starts off with a tie and neck. Then there a few illustrations showing fingers, neck and tie in impossible positions, and finally a perfectly formed knot surmounted by a huge grin.

*Single-Sideband Systems and Circuits*[3] uses the concept of negative frequency with a complex number topping.

Although these concepts are rather tricky to master, this last method is the best; the Weaver receiver becomes simple to understand, and there are many other uses for these concepts[4]. An alternative explanation follows.

The receiver system works by converting the RF input down to audio frequencies in one direct step, without using intermediate frequency stages. For reasons to be discussed later, this is done twice, using two RF mixers, one of which is supplied with an LO signal phase shifted in comparison to the other. The resulting audio signals are then passed through low-pass filters and finally combined in a second 'rotary' AF mixer stage, which is driven by a second LO.

Because the output of the RF mixers is at audio frequency, the Weaver receiver is classed as a direct conversion receiver. However, the receiver is best viewed as a type of superhet, with two parallel audio IF stages between the first (RF) and second (AF) mixers. This article uses superhet terminology. **Fig. 1** shows the block diagram of a Weaver receiver.

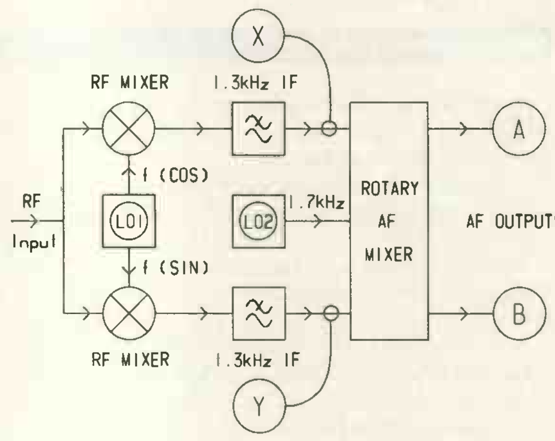
Consider one of the IF low pass filters. It has a cutoff frequency of 1.3kHz, half the width of the final audio output. This may seem rather surprising, but look at **Fig. 2**. This shows the signals to be found at various points on the block diagram. The left hand column shows that there are two possible RF input frequencies that will generate an IF output of 1kHz. One is 1kHz above the 1st LO frequency, and one 1kHz below the 1st LO frequency. So, although the IF audio filter is only 1.3kHz wide, the information passing through the filter is due to an RF bandwidth of 2.6kHz, half of which is below the 1st LO frequency, and half above. This process of getting a quart into a pint pot is achieved by the mixer folding the audio spectrum over.

Unfortunately, this folded signal is unintelligible as ordinary speech, so the job that the Weaver receiver performs is to unfold the audio spectrum into something intelligible.

To do this, a second RF mixer is used, but it is provided with an LO that is phase shifted by 90° with respect to the LO of the first mixer (sin and cos are 90° apart). This extra mixer also provides a folded audio output, however the output is folded differently. The second column of Fig. 2 shows the waveforms on the outputs of the two mixers. Note that, while the waveform at *X* is the same for both RF input frequencies, the waveform at *Y* is phase inverted. This is not to say that the waveform at *X* is useless, on the contrary, it serves as the reference that enables the phase inversion on point *Y* to be seen.

At this point in the discussion, it is simpler to consider these *X* and *Y* IF signals as being connected to the X and Y plates of an oscilloscope to form a Lissajous figure. The result for both RF input frequencies will be a circle (This is why sin and cos are called circular functions). However, the RF input that is 1kHz above the LO frequency will give an anti-clockwise rotating spot, and the RF input that is 1kHz below the LO frequency will give a clockwise rotating spot.

*Fig. 1. The Weaver receiver principle. Although it looks like a direct conversion to baseband system, it is in fact a heterodyne arrangement with an intermediate frequency of 1.7kHz. Individual sidebands are resolved by quadrature product detection at 1.7kHz. Although the IF audio filter is only 1.3kHz wide, the information passing through the filter is due to an RF bandwidth of 2.6kHz, half of which is below the 1st LO frequency, and half above. This process of getting a quart into a pint pot is achieved by the mixer folding the audio spectrum over. The second mixer unfolds it to its full width.*




## Explaining AF mixing

To recover the original audio signal, the *X* and *Y* IF signals are connected to the rotary AF mixer. This imparts an extra clockwise or anti-clockwise twist to the spot's motion. The speed of this extra twist is 1.7kHz, which is the second LO frequency. The resulting outputs are shown in the right-hand two columns of Fig. 2. The outputs are still circles, but with differing rates of rotation, depending on whether the original RF signal was greater or less than the first LO frequency.

Assume that the receiver is to be used to demodulate USB. The upper row of Fig. 2 shows that an RF input of 1.001MHz gives an IF of 1kHz, and the output from the rotary mixer is 2.7kHz. The lower row shows that an RF input frequency 1MHz–1kHz also gives an IF of 1kHz, but this time gives a rotary mixer output of 700Hz.

If the Weaver receiver is used to demodulate LSB, the direction of the extra twist from the rotary AF mixer is reversed. The RF input of 1.001MHz now gives a 700Hz AF output, and the RF input frequency 1MHz–1kHz an output of 2.7kHz.

The circuit of the rotary mixer that gives this extra twist is discussed later. Note that, for the purposes of this illustration, the receiver has two audio outputs with a 90° relative phase shift. This allows the rotary AF mixer outputs to be discussed as circular Lissajous figures. However, for an SSB receiver, only one of the two outputs is needed.

## Mixers

For the direct conversion receiver, the RF mixers' performance is vital: apart from the LO, there is very little other RF circuitry. If a direct conversion receiver does not work satisfactorily, the RF mixer is usually to blame.

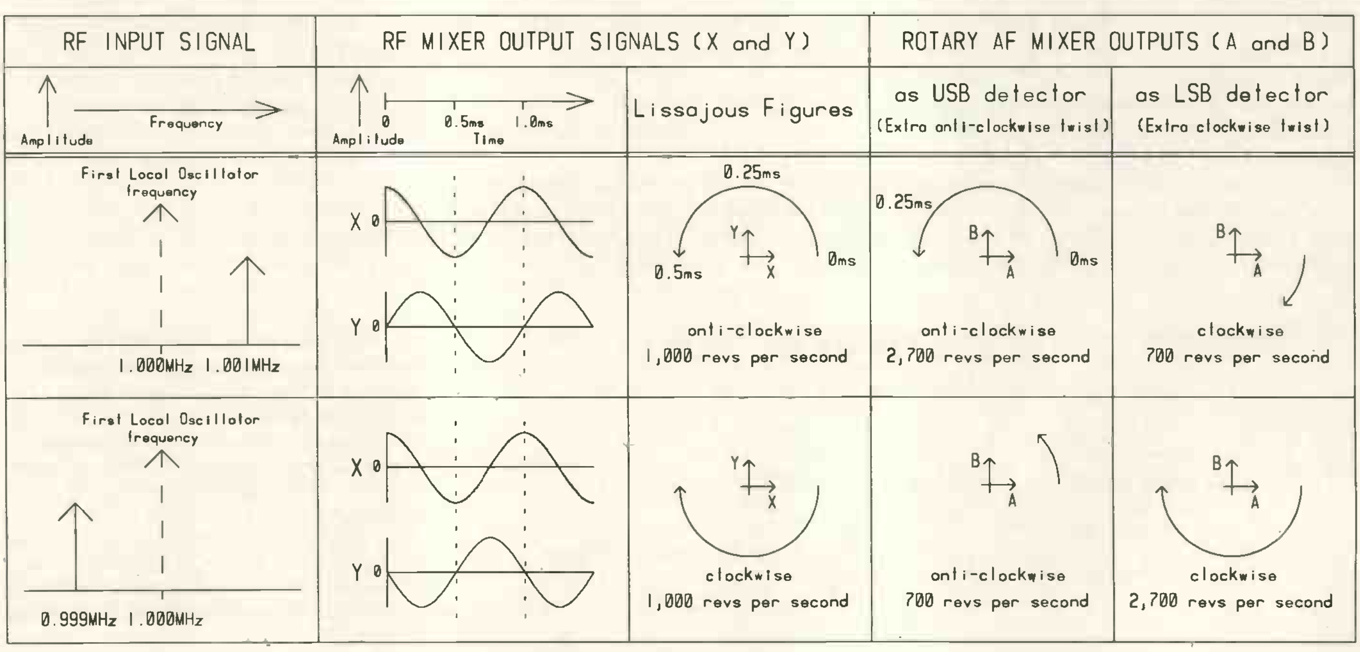
The wanted output frequency from an RF mixer can be either LO+RF or LO–RF. In the direct conversion receiver, the mixer must translate the RF input down to audio frequencies. So it will be assumed that LO–RF, the difference frequency, is the wanted output.

The ideal RF mixer would have two inputs, RF and LO, and the output would consist of just one frequency, the difference between the input frequencies. It would have no harmonic responses. How might this be achieved?

To generate a lower frequency output, the cycles of the input waveform must be lengthened thus the output must be phase-retarded each cycle with respect to the input. To do this, a voltage variable phase shifter is required. It must be able to shift the phase of the input signal by a full 360°, and be continuously variable. A block diagram of a circuit which does this is shown in **Fig. 3**.

The circuit works by splitting the incoming RF signal into a 0° and a 90° component. The two signals are then passed to two balanced mixers. These act both as phase inverters and as voltage controlled attenuators. The phase inverter action means that the phase of the upper mixer's output can be either 0° or 180°, and the phase of the lower mixer's output can be either 90° or 270°. A judicious mixture of these four phases results in a continuously variable phase from the output of the sum-

*Fig. 2 shows the signals to be found at various points on the block diagram. The left hand column indicates that there are two possible RF input frequencies that will generate an IF output of 1kHz. One is 1kHz above the 1st LO frequency, and one 1kHz below the 1st LO frequency.*



## Design considerations

The Weaver receiver has been ignored as a design for analogue receivers. For receivers using digital signal processing it has been considered and rejected[11]. The reasons for this rejection are quoted as these.

"The problem of DC offsets would necessitate AC coupling. However, this would place a notch in the effective receiver passband which could be troublesome for certain modulation modes."

This design proves that the central notch can be made as narrow as 10Hz. The resultant degradation of SSB is negligible. Even when tuning around a strong carrier wave, the notch is quite hard to find by ear. However, the notch does result in a short burst of 1.7kHz from the receiver output each time there is a large change in LO frequency.

For the HF band, the synthesiser would have to cover over two decades of frequency range and provide quadrature outputs. The VCO must have a one octave frequency range. To receive the full HF frequency range, extra divide by two circuits may be used. It is simple to make the frequency dividers provide the necessary 90° outputs.

Gain and phase matching of the (cos and sin) channels have to be accurately maintained over a very wide bandwidth to avoid sideband image problems. This receiver achieved an AF distortion suppression of 30dB without difficulty, and this is almost inaudible. DSP techniques would reduce the distortion.

1/f noise in the mixers and audio amplifiers should be minimised. The receiver's noise figure is about 20dB, which is about 10dB higher than a standard HF receiver. This is mainly due to the choice of amplifier circuit at the input of the LP IF filter. The noise floor could be lowered, but this would reveal 1/f noise and greater hum sensitivity. The advantage would be that, given a clean LO, lowering the receiver's noise floor would further increase its spurious free dynamic range.

RF sub-octave filtering would be essential to prevent unwanted signals at harmonics of the input signal from mixing with harmonics of the local oscillator. This design makes the point that, by running the mixers with an accurate square wave drive, and by using 3rd harmonic response cancellation, the complexity of the preselector is considerably reduced.

Remember that it is standard practice to quote the suppressed carrier frequency of an SSB signal. So, for LSB, the receiver should display LO+1.7kHz, and for USB, the receiver should display LO–1.7kHz.

ming junction. This judicious mixture is arrived at by yet another 0° and 90° splitting of the LO signal.

### Squarewave LO drive

Using a double balanced diode ring mixer as a voltage variable RF attenuator is likely to cause intermodulation products. Imagine that the LO input is a sine wave. As the instantaneous LO voltage nears 0V, the RF voltage has a greater effect on the diode current, so the mixer attenuation varies depending on the RF input waveform. This is another way of saying that it is generating intermodulation products. To avoid this, the mixers must be provided with a square-wave LO drive[5] in order to achieve the required strong signal handling performance. This works because the diode current is at its saturation value most of the time, and passes through the 0V danger zone much more quickly.

In this receiver, the square wave input to the mixer is generated by a frequency divider with a 1:1 duty cycle output waveform. The LO mixer drive of the lower mixer in Fig. 3 can be expressed as a Fourier series:

$$S(t)=4/\pi[\sin\omega t+{}^1/_3\sin3\omega t+{}^1/_5\sin5\omega t+{}^1/_7\sin7\omega t+...]$$

This shows that the second harmonic component of the mixer drive waveform is theoretically zero. In fact it will be present, but will be very small. The result is that the mixer's 2nd harmonic response will be similarly small.

This is not the case for the 3rd harmonic response, which will have a conversion loss only $20\log(^1/_3)$ or 9.5dB larger than the fundamental frequency conversion loss. The upper mixer drive in Fig. 3 may also be expressed as a Fourier series:

$$C(t)=4/\pi[\cos\omega t-{}^1/_3\cos3\omega t+{}^1/_5\cos5\omega t-{}^1/_7\cos7\omega t+...]$$

Compare the two series, and notice that the terms for the 3rd and 7th harmonics change signs. This results in the two mixers giving the same AF amplitude output in response to a 3rd or 7th harmonic RF input frequency, but the outputs are phase inverted with respect to each other. When they are added together in the summing junction, these harmonic responses cancel. The result of this is a mixer in which all the harmonic responses cancel, with the exception of the 5th, 9th 13th etc.

It is difficult in practice to build the RF input phase shifter so that the accuracy is better than 4°. This gives a theoretical 3rd harmonic signal cancellation of 29dB (see box). Add to this the 3rd harmonic conversion loss (see above) of 9.5dB. Thus the minimum 3rd and 7th harmonic rejection should be about 40dB. Inaccuracies of the 0/180° phase shift in the mixer limit the even order harmonic rejection to about 50dB.

### The effects of error

For a pure sine wave input to the receiver, the outputs of the IF amplifier/filters should be equal and have 90° phase shift. However, the phase and amplitude are always slightly in error, so the Lissajous figure, which should be a perfect circle, is always slightly elliptical. This elipticity can be viewed as the result of a small circular component rotating in the opposite direction to the main component. The demodulator interprets this as a small signal of the opposite sideband to the main signal, so, when demodulating USB, the distortion takes the form of an LSB image. This is the worst case in-band AF distortion referred to in **Table 1**.

Note that the harmonic responses of the mixers give a residual output from the two IF filters. Although the resultant frequency seen in the filters will be the same, the relative phase and amplitude between the two signals will be random for any given frequency. Because the two IFs have signals that are not of identical amplitude and 90° phase shift, the rotary mixer will not be certain whether to

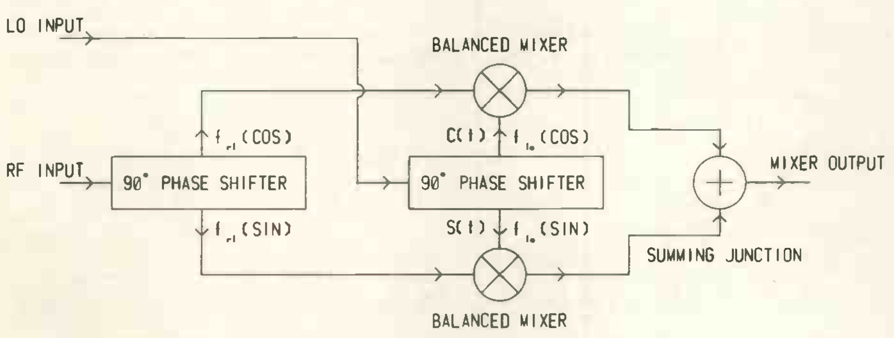



*Fig. 3. The ideal RF mixer for a Weaver principle front end. The circuit works by splitting the incoming RF signal into a 0° and a 90° component. The two signals are then passed to two balanced mixers. These act both as phase inverters and as voltage controlled attenuators. The phase inverter action means that the phase of the upper mixer's output can be either 0° or 180°, and the phase of the lower mixer's output can be either 90° or 270°. A judicious mixture of these four phases results in a continuously variable phase from the output of the summing junction.*

give a USB or an LSB output, and so gives a mixture of both.

Spurious demodulation is possibly the worst fault of direct conversion receivers: strong signals (usually 7MHz AM broadcast) are directly demodulated to audio frequencies, irrespective of the LO frequency. Spurious demodulation occurs because of imbalance in the mixer. The *SBL1* is inherently well balanced: it has all its diodes fabricated on one substrate. The *SRA1H* uses a ring of eight matched diodes in separate packages, and will not be so well balanced. This may explain why the spurious demodulation performance of the two types is similar, even though the *SRA1H* has a higher 3rd order input intercept point.

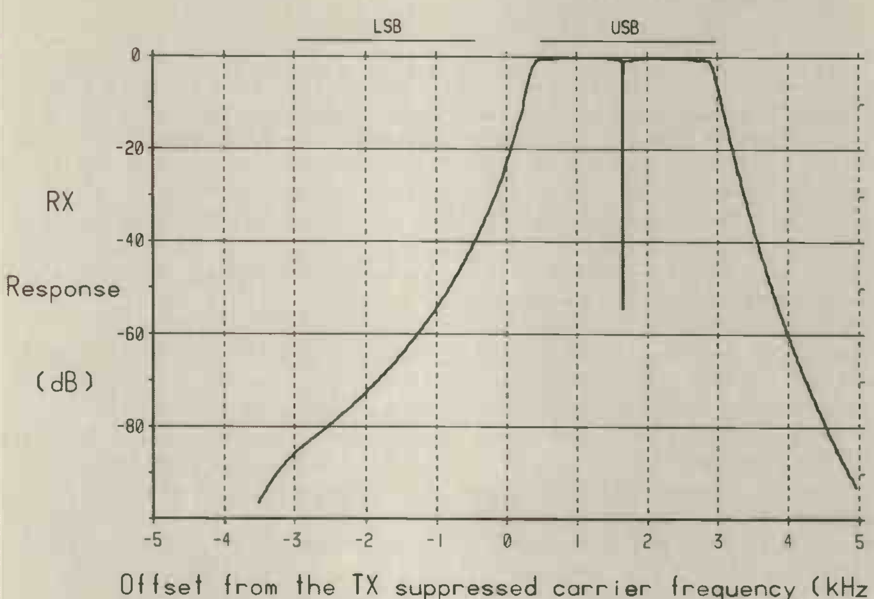
However, the third order intermodulation products start appearing above the receiver noise floor at the same signal level as the onset of spurious demodulation. In fact, for a superhet, an AF detector on the mixer output would be a good indication of mixer RF overload.

In the Weaver receiver, the spuriously demodulated signals are frequency shifted by the second mixer, making them unintelligible. This is an advantage: off tune SSB interference is subjectively far less annoying than, for example, being able to hear the BBC World Service in the background on every frequency.

All direct conversion receivers are prone to pick up power line hum. This is due to a combination of direct pickup and LO radiation[6]. To reduce hum, homodyne and phasing receivers have a 400Hz high pass filter on the mixer outputs. This removes 50Hz and its first seven harmonics without any loss of SSB signal information. The Weaver receiver cannot do this because the IF filters must pass frequencies down to 5Hz. In this receiver, any hum is converted by the AF mixer into two tones close to 1.7kHz, which are impossible to filter out. As a result, the Weaver receiver's mixers must have a hum output much smaller than other direct conversion receiver types.

*Table 1. Performance figures for the prototype receiver. The measurements were made with the RF attenuator and preselector disconnected.*

| | |
|---|---|
| **Frequency coverage** | 1.5 to 30MHz |
| **Detector type** | SSB (USB or LSB) |
| **RX 3rd order input intercept point** | +16dBm |
| **Selectivity** | see graph |
| **Centre frequency notch width** | 10Hz at 3dB points |
| **Sensitivity** | –108dBm for 12dB SINAD |
| **Harmonic responses** (receive frequency 2MHz w.r.t. –108dBm): | |
| | 34 dB at 6MHz (3rd ) |
| | 12 dB at 10MHz (5th ) |
| | 18 dB at 18MHz (9th ) |
| | 21 dB at 26MHz (13th ) |
| | 50±3dB the rest |
| **Sub-harmonic response** (2MHz in) w.r.t. –108dBm: | |
| | 83dB (RX at 4MHz) |
| **AF distortion** | 30dB below wanted tone |
| **Spurious demodulation** (100% AM) | –30dBm for 12dB SINAD |




*Receiver selectivity graph.*

## Implementation

Looking at block diagram **Fig. 4**, the main receiver components are these:

The **attenuator** reduces the signal levels arriving at the mixer when necessary. A 7.2MHz half-wave dipole at night can produce signal levels as high as 0dBm (223mV across 50Ω or 1mW) – causing severe direct demodulation. It is high time automatic RF attenuators were standard in HF receiver design.

The **preselector** performs three functions. The first is simply to minimise the out-of-band RF energy arriving at the mixer. This reduces the susceptibility of the receiver to interference from intermodulation products or direct demodulation. The second function is to attenuate sub-harmonic frequencies. These will suffer harmonic distortion in the isolator and mixer, and give spurious products at the receiver frequency. The third function is to attenuate frequencies at harmonics of the receiver frequency. These may produce harmonic responses from the mixer.

The **isolator** passes the signals from the preselector to the mixer at unity gain, but attenuates LO leakage from the mixer attempting to travel in the opposite direction by up to 80dB. Reference 6 explains the use of an isolator to reduce local oscillator radiation and RF generated hum and microphony.

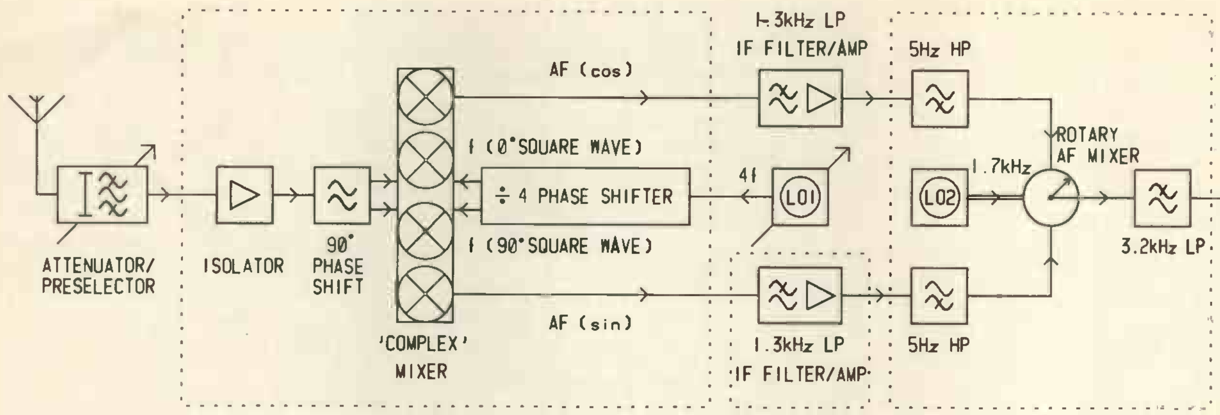
The **90° phase shifter** is a broadband device, but, because its only function is to cancel the 3rd harmonic response of the mixers, the frequency range of the cancellation need only be from the 3rd harmonic frequency of the lowest input frequency to the receiver's top frequency. This phase shifter gives 90°±4° over a 4 to 30MHz frequency range.

The **complex mixer (Fig. 5)** is an implementation of Fig. 3, one to provide the sine output and one the cosine. The mixers are driven by a square wave derived from a divide by four flip-flop stage. The two square waves have a 90° relative phase shift. This phase shift is used to give the correct phase relationship between the two IF filter signals in order that the correct USB or LSB detection can be applied. It must therefore work over the full frequency range of the receiver.

The **1.3kHz LP IF filter and amplifier stages** have a 6th order 0.1dB Chebyshev frequency response. A more complex filter would give better selectivity, but it would make the phase and amplitude matching of the two stages harder. It is instructive (and pretty) to connect the outputs of these filters to the X and Y plates of an oscilloscope. The true phasor nature of the RF input signals can then be displayed. It is, for instance, easy to spot the broadcast stations that have AM to PM conversion, or dynamic carrier control.

The **5Hz high pass filter** removes the DC component of the IF signal that is to be connected to the rotary mixer. Most of this DC component arises from mixer imbalance, and the residue from amplifier offsets. The DC level controls the amount of 1.7kHz in the output, so the spurious content must be removed.

The mixer DC offset depends on the LO frequency, so, for each large frequency change, a step function is applied to the HP filter. The



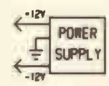




*Fig. 4. Receiver block diagram.*

transient response of the HP filter causes a 1.7kHz ping at the receiver output. So the cut-off frequency of the filter is a compromise between the desire to use the narrowest possible notch width in the centre of the AF passband, and the desire for the ping caused by each frequency change to be shortest.

A 0.5Hz high pass filter would result in less of the AF passband being lost, but would take a long time to arrive at its final value. This would result in a prolonged 1.7kHz whistle every time the receiver's frequency was changed. 5Hz was chosen as a reasonable compromise, and gives a negligible reduction in speech intelligibility[7].

The **rotary AF mixer** is a rotating switch which selects one of eight phases for the AF output. If four phases are used, an audio image appears between 3.8kHz and 6.4kHz, and this imposes a severe filter requirement on the subsequent LP filter. With eight phases, this image is moved up to 10.6kHz to 13.2kHz, well out of harm's way.

The **3.2kHz LP filter** removes the 3rd and higher harmonics of the 2nd LO frequency, and some high frequency mixing products which are generated by the Rotary AF mixer because it operates in discrete phase steps.

Circuits are not given for the attenuator, preselector, LO, or AGC stages since there is nothing unusual or design specific about them.

## Receiver frequency response

The effects of all the audio frequency signal processing can be seen in the graph (see Table 1). The fundamental shape is of two 1.3kHz LP filters glued back-to-back, and centred on the 2nd LO frequency of 1.7kHz. In the centre of the passband is a 10Hz wide notch formed by the 5Hz HP filters. The 3.2kHz LP filter's cut-off can be seen at ±3.2kHz; it causes the receiver response to roll-off faster on the high frequency side of the passband than the low frequency side.

At the AF output shown on Fig. 4, all frequencies less than 400Hz will contain interference and no signal. The subsequent loudspeaker amplifier will therefore be AC coupled. The frequency response of the receiver from the RF input to the loudspeaker output will thus resemble the response graph, but will have an extra notch 800Hz wide centred on 0Hz.

## RF circuit description

The isolator stages use two dual grounded gate JFETs. The circuit[6] has been adapted to a push-pull type in order to reduce the second harmonic distortion. Both live and ground of the RF input are connected to the receiver using 100nF capacitors. The LO input is similarly decoupled. This stops 50Hz hum current from flowing from the ground of the antenna connector or external LO, past the low noise AF input, and into the mains supply earth. A small potential due to this current would be developed at the low noise input of the 1.3kHz LP IF amplifier.

The 90° phase shifter, if made with perfect

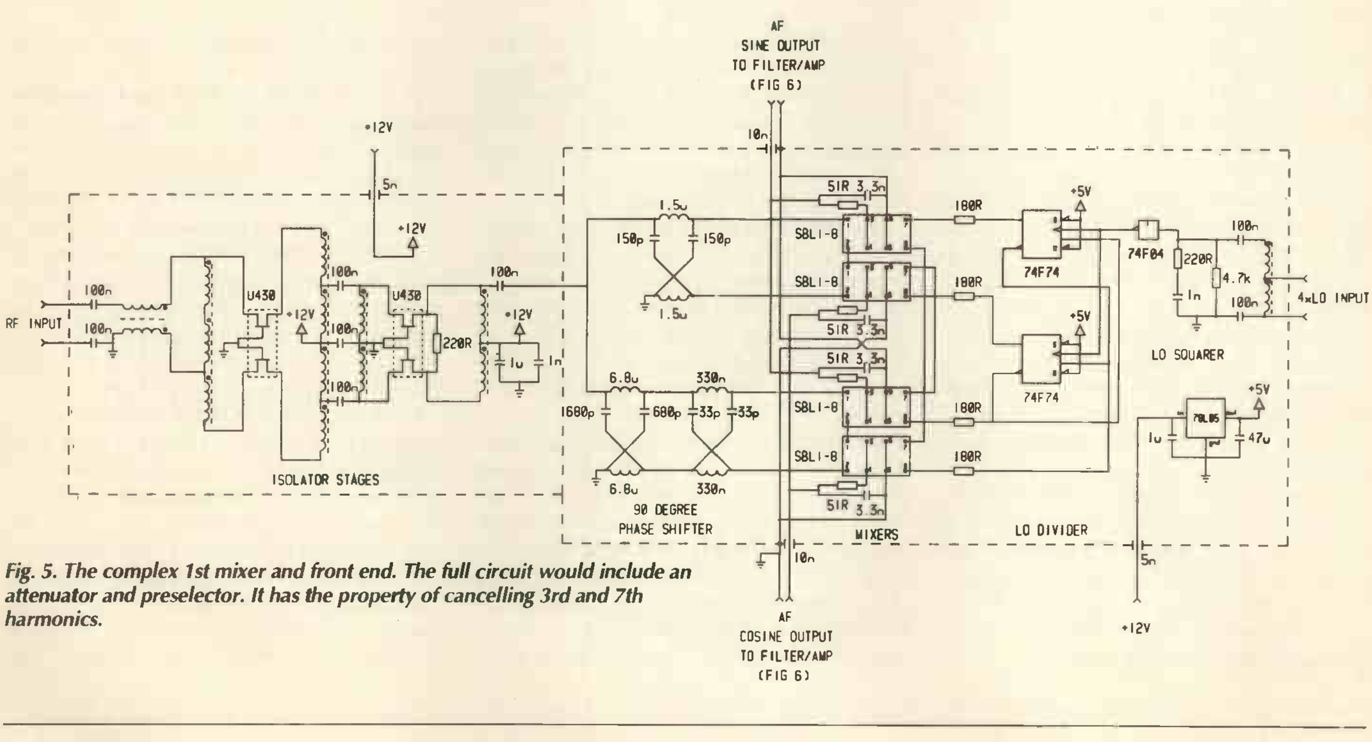



*Fig. 5. The complex 1st mixer and front end. The full circuit would include an attenuator and preselector. It has the property of cancelling 3rd and 7th harmonics.*

components, would give the response shown in the graph, right – a computer simulation. Unfortunately, the most accurate inductors to hand had a tolerance of 10%. As previously explained, a 3rd harmonic rejection of about 40dB could be expected depending on the accuracy of the 90° passive phase shifter. This has not been achieved, see Table 1. Note that this harmonic rejection was measured at 6MHz, which is where the inaccuracy of the 90° phase shift is greatest, so the phase error due to component tolerance was positive, and added to the predicted error.

The phase shifter design is simplified by using a branch impedance of 100Ω, which means that the inductors must be exactly 10,000 times the capacitor values, so the E6 series of component values can be used.

The LO divider is a *74F74*. This provides a square wave drive current of 14mA for the mixers. The *74F* logic series is not guaranteed to operate at a clock rate higher than 100MHz; this would limit the maximum receiver frequency to 25MHz. In practice, all the devices so far tried have operated at greater than 120MHz. For higher guaranteed speeds, use ECL.

The mixer inputs are connected in series, giving an input impedance of 100Ω to match the phase shifter impedance. The IF outputs of the mixers are matched at RF by a 51Ω resistor. At AF, the mixers are mismatched. This reduces the audio band 3rd order input intercept point of the receiver, but improves its noise figure.

## LP IF filter/amp circuit description

The input to the 1.3kHz low pass IF filter/amplifier stages could have used a diplexer[6]. In these, the filtering provided by the diplexer is calculated to be part of the overall filter response. For more complex designs of diplexer using Chebyshev/Inverse Chebyshev or Cauer/Cauer pairs, see reference 8. Be warned that Cauer diplexers have a nasty habit of requiring negative values of output capacitance on the high pass side. The actual circuit is shown in **Fig. 6**. These have to be realised using transformers and some mathematical sleight of hand, and transformers are microphonic.

However, we live in an age where the aim is to avoid using inductors at all costs. So this design uses a second order filter stage[9]. The input resistance is formed by the 50Ω source impedance of the complex mixer circuit. This low input resistance results in a high input capacitance to ground, which provides an excellent opportunity to prevent LO signal escaping from the mixer compartment.

The second audio stage is similar to the first, and has a gain of 20dB. The third stage has the poles with the highest Q, so this was designed using a low sensitivity section. The design procedure for this section produces a fixed R/C ratio, where R and C may be chosen at will.

The 20dB amplifier output stage raises the signal level to a value that is just sufficient to

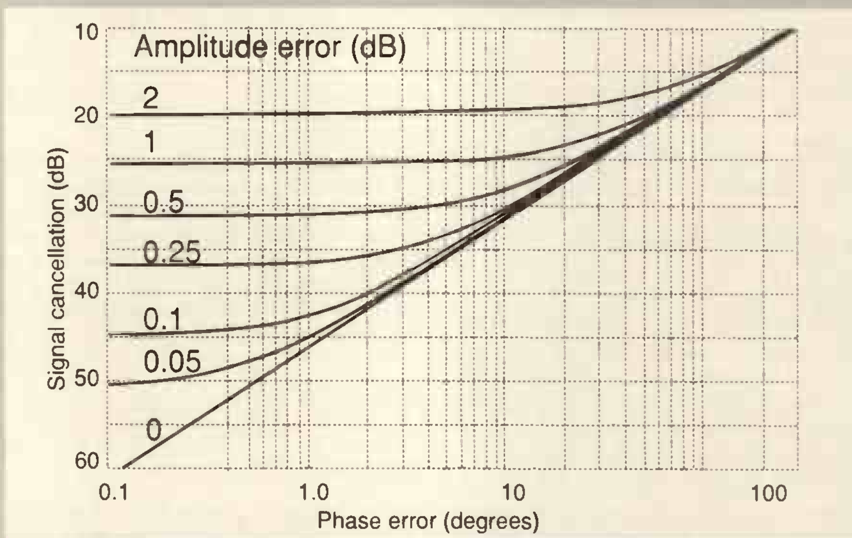
### Signal Cancellation

There are two places in the receiver architecture where the signal path splits into two, which then rejoin. At each of these joining places, the aim is to cancel out unwanted signals. This happens in the RF mixer where 3rd and 7th harmonic responses of the mixers are cancelled out, and again in the rotary mixer where the two low frequency IF signals are combined in order to cancel out the unwanted sideband.

In the first instance cancellation depends on the accuracy of the 90° phase shifting network. This is composed of all pass filters, so the amplitude error will be very small, and the phase error dominant. However, the amplitude error can be dominant; for example, there is a 90° hybrid junction design that has constant phase difference, but has amplitude ripple.

This is the exception. The phase error is dominant in most circuits. Imagine the errors to be due to an RC LP filter. At one tenth of the cutoff frequency, the amplitude error is 0.044dB, and the phase error is 5.7°. Look at the graph; it shows the amount of signal cancellation that may be expected for various phase and amplitude errors. An amplitude error of 0.044dB with no phase error gives a signal cancellation greater than 50dB, whereas a 5.7° phase error with no amplitude error gives a signal cancellation of 27dB. The lower value of signal cancellation will prevail.

There is also a less obvious splitting and combining point in each of the balanced mixers. These work by selecting a phase of either 0° or 180°, depending on the direction of the instantaneous LO current flow through the mixer. The phase shift between these two states must be exactly 180° and the amplitude of the two states must be exactly equal for the even order harmonic responses to cancel completely. In this case, the two paths through the mixer may be thought of as splitting and rejoining in time.



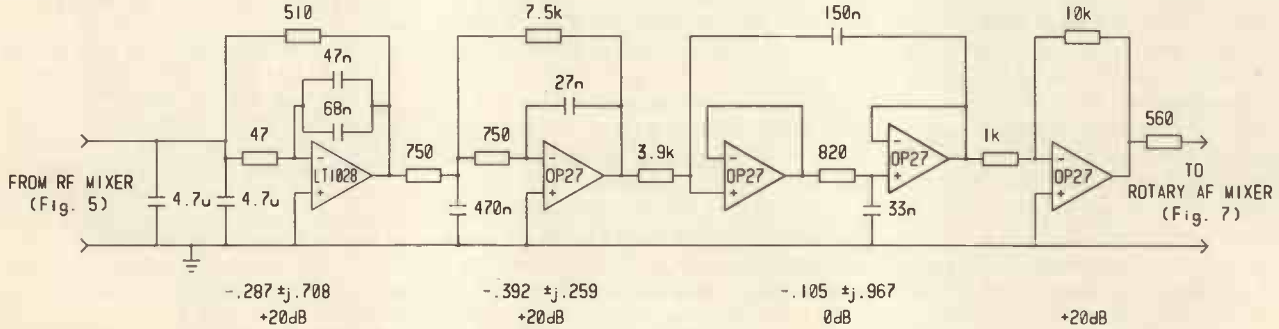




*Fig. 6. 1.3kHz low pass IF filter/amplifier stages*

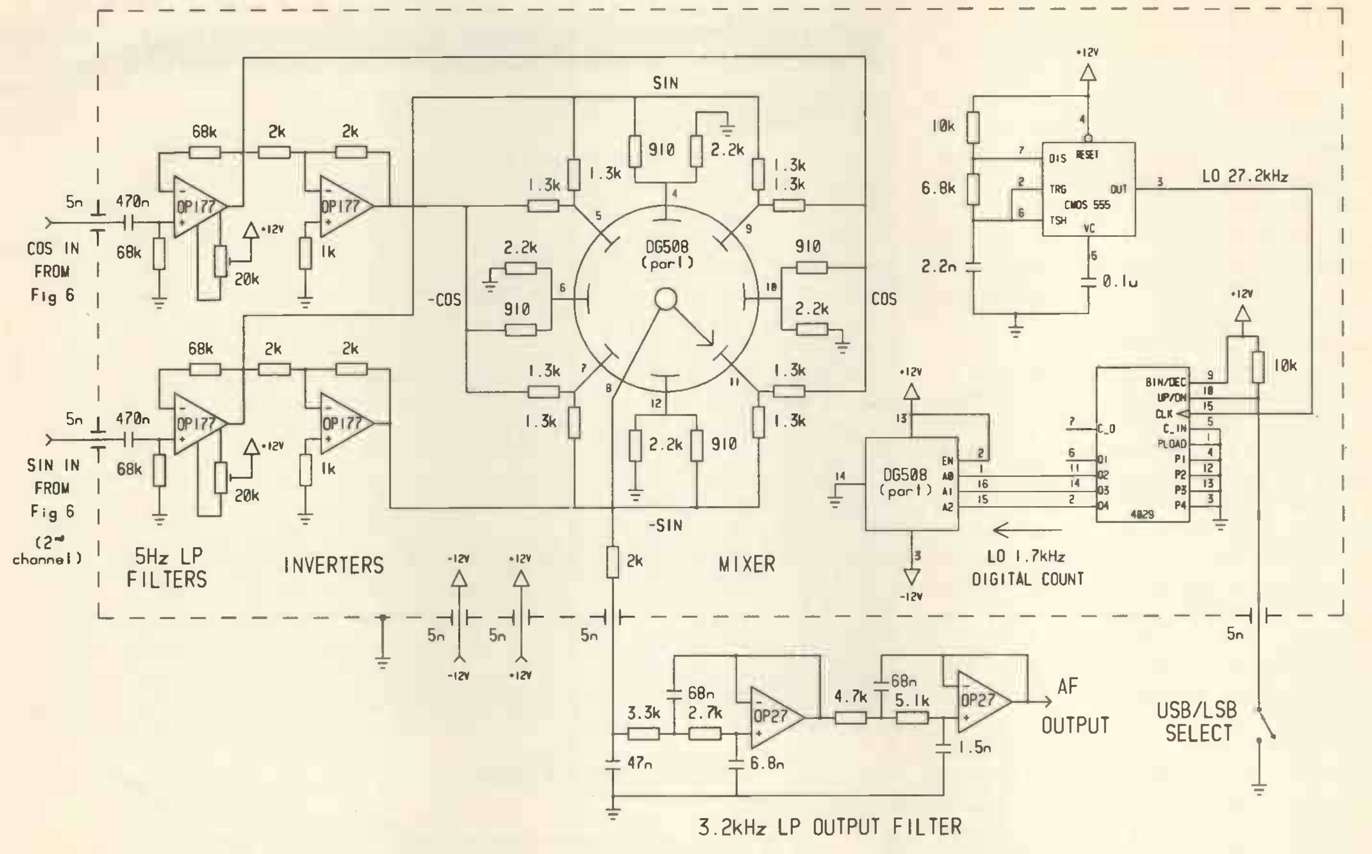



*Fig. 7. Four phase product detector described by the author as a rotary mixer. It contains two inverters: one on the sine channel and one on the cosine channel, their outputs are –sine and –cosine creating four phases to choose from. Four more phases are interpolated by the resistor networks around the analogue inputs of the DG508 analogue switch.*

make the leakage of the 2nd harmonic of the 1.7kHz LO inaudible in the front-end noise.

### Rotary AF mixer description

The 5Hz high pass filters must be first order to keep the transient response free of overshoot[10]. For this a simple resistor/capacitor time-constant is used. An alternative but more complex candidate for this circuit is the critically damped filter, which has a transient response that arrives at the final value rather more promptly than a resistor/capacitor filter.

The *OP117* op-amps have very low DC offset, and would normally require no adjustment. However, a small amount of 1.7kHz leaks from the subsequent mixer circuit into the AF output. The offset potentiometers on the op-amps allow a small amount of signal to pass in anti-phase to the mixer leakage thus cancelling it out.

The **rotary mixer** contains two inverters: one on the sine channel and one on the cosine channel, their outputs are –sine and –cosine creating four phases to choose from, **Fig. 7**.

Four more phases are interpolated by the resistor networks around the analogue inputs of the *DG508* analogue switch. The resistor values are chosen such that, at the *DG508*, the source impedance of all eight phases are equal. Also, the sine, cosine, –sine and –cosine phases are attenuated by a voltage ratio of 0.5, so that all the phases have an equal RMS voltage. There are now eight phases for the *DG508* rotary mixer to choose from.

The second LO is generated by a *555* oscillating at 27.2kHz. This signal is connected to the *4029* up/down counter. As the *DG508* has only three digital inputs, one of the four counter outputs is unused. This gives a choice of using a clock rate of 27.2kHz, and not using the $Q_1$ output of the *4029*, or of using a clock rate of 13.6kHz, and not using the $Q_4$ output. While this last option provides identical control of the *DG508*, some of the 850Hz generated by the *4029* inevitably leaks into the audio output, so the 27.2kHz option is to be preferred. Note that the majority of low frequency circuitry is enclosed by a screened box, and that all the supply and signal connections are made using feed-through filters. This must be done to prevent RF harmonics of the 2nd LO from leaking into the RF stages of the receiver.

The 3.2kHz low pass output filter is a 5th order 0.1dB Chebyshev filter[9]. The filter input resistor has been reduced in value to allow for the mixer source impedance.

In the output of the conventional mixer (balanced modulator), the sum and difference frequencies are present at the same time. This is because the conventional mixer approximates to the rotary mixer by using two phases: 0 and 180 degrees. The mixer cannot distinguish between clockwise and anti-clockwise rotation. The output contains the two frequencies as if the mixer were rotating both clockwise and anti-clockwise simultaneously.

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PC ENGINEERING

# SPICE joins the Windows club

**Spice is now available in a Windows version. John Anderson finds a good package – with one or two shortcomings.**

Users who know *Spice* well will be up and running with *SpiceAge for Windows* with little delay. *Windows* pull-down menus give access to all aspects of the package, including setting test nodes, starting simulations and plotting the graphs. No great increase in functionality is claimed over previous versions, so the question for any potential user is whether support for *Windows* is a worthwhile addition.

Simulation speed is adequate – for small networks. But some circuits require a small time step so that sampling rate errors may be removed. Under these circumstances simulation can take several minutes – even on a fast 486 machine. Even so, processing is still significantly faster than wiring and checking a breadboard.

Overall, the simulation capability is impressive. Frequency or time response can be displayed for any node, with four nodes specified at any one time. In the frequency domain, the response can be displayed as a Bode plot, a complex plot (Nyquist) or as real and imaginary parts. In the time domain, the responses of the systen to a wide variety of predifined generators can be displayed as either XT or XY (Lissajous's figure). Tolerances on a particular response for variations in component values or changes in temperature can be displayed as a family of response curves for each variant.

Fourier analysis of the time response allows determination of harmonics in the waveform. But be careful here as there may not be an integral number of fundamental periods in the transient analysis. The Hanning window option, sometimes referred to as the raised cosine window, can help, enhancing the repetitive nature of the waveform by weighting the data at the centre of the window relative to the edges.

Graphing could be better. Graph scales can be set, but to display the rescaled graph requires a tedious resimulation. On the plus side there is a very useful real time cursor which enables, as the mouse is moved, the XY co-ordinates of the cursor to be displayed at the top of the window. Also the delta mode, implemented when the left hand mouse button is depressed, makes for quick and convenient determination of –3dB points, bandwidth and risetimes accurately from the graph. Scaling is normally automatic, and seems to be sufficient, though a pop up window will control the scale of the graph if required.

One of the sillier aspects of the system is the scrollable tool bar, the "ribbon of buttons" used to input the correct component type letter into the *Spice* model. For example, pointing the mouse at the resistor icon and clicking, results in "R" being input. Of course the "R" can be pressed on the keyboard, and as this is a tool for analysing an existing design, the nodes and components would have already been defined and would only take a few seconds to key in by hand.

## Quiescent analysis and models

A circuit can be analysed for its DC quiescent levels, presented in table form. Unfortunately presentation could be better. The font is small, and when the quiescent conditions are analysed using a Monte Carlo technique on the component tolerances, the table of values is both crowded and illegible.

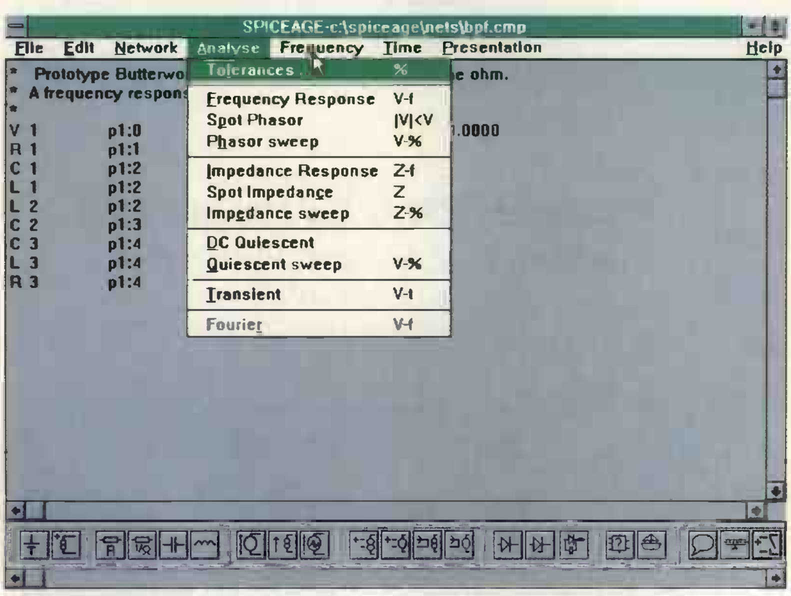
One useful feature is that some models have a temperature coefficient built into their definition, allowing analysis of temperature effects on the circuit response. What a pity this does not extend to making temperature the dependent variable so that attributes can be directly plotted against it.

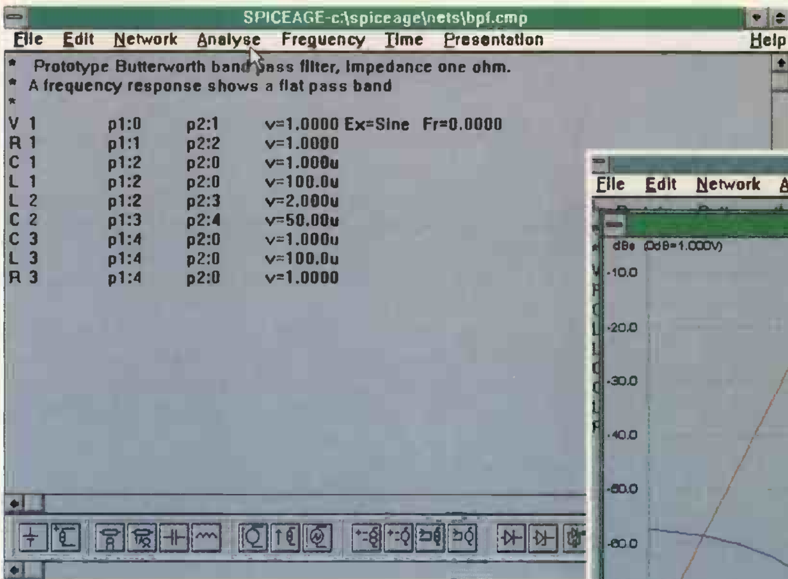
Models comprise a set of primitives, eg resistor and capacitor, and library networks such as transistors and op amps. Library networks created by the user are treated in the same

### System requirements

System requirements are not specified, but the review was carried out using:
Windows 3.1.
Windows 386 Enhanced mode
8Mbytes ram
486 PC
VGA screen
Output to a Windows supported pen plotter or printer.

*SpiceAge for Windows edit window shows the analysis options*






*Left: A simple Butterworth filter definition*

*Below: Displaying a filter frequency and phase response*


way as any other circuit, except their file extension is .LIB rather than .CMP.

But compatibility could be a problem. There are several dialects of the "standard" *Spice* language, so models supplied by op-amp manufacturers may not run immediately on *SpiceAge*. However, the manual provides two methods of defining the same component, including the normal .MODEL syntax of the original SPICE simulators.

## Manual

The package comprises a loose-leaf A5 manual and a disk and the software is not copy protected. The manual appears to have been produced quite cheaply with single sided photocopied pages, but is well written and logically arranged.

Installation is simple and uneventful.

## Spice for beginners?

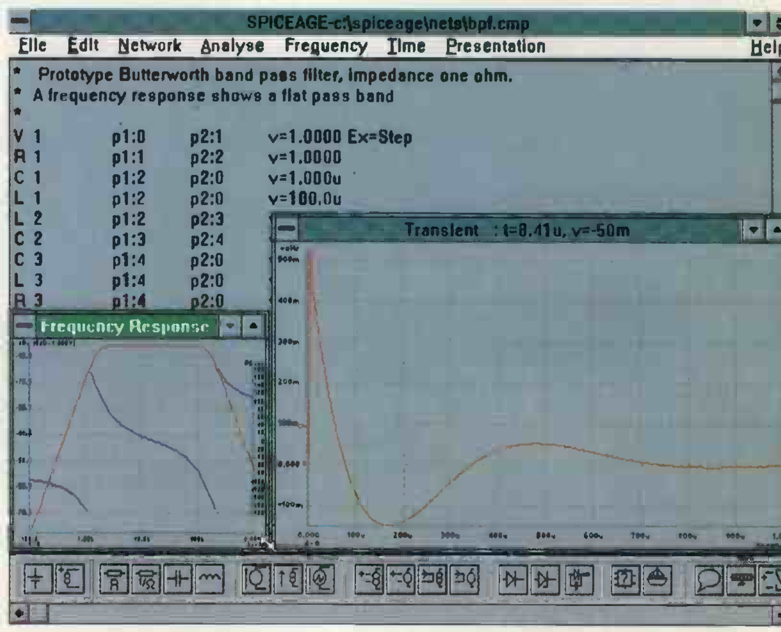
*Spice* is a modelling system for electronic networks, where a series of attributes and parameters of the component parts of the network are individually defined. Modelling takes the form of frequency, time or quiescent simulations.

Emphasis these days is on "right first time", and Spice offers the opportunity to check circuit performance without recourse to any lab work. But beware, *Spice* will not simulate things for which there is no information – eg parasitic inductances and capacitances – and unmodeled modes of circuit operation such as inputs exceeding the supply voltage etc.

Further, aliasing due to the effective sampling rate can cause spurious results. Thus it should be used as a useful test-bed for the circuit operation, and offers an ideal tool to tolerance the circuit.

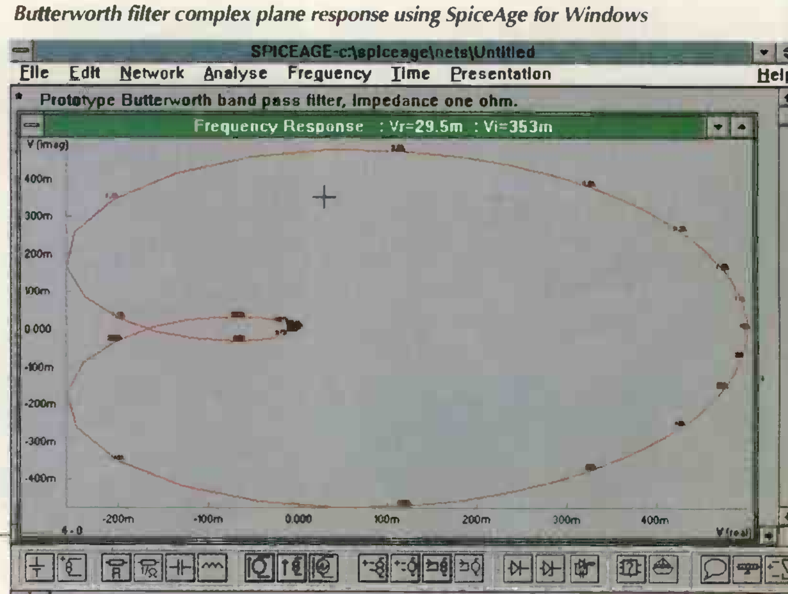
## Half a good idea?

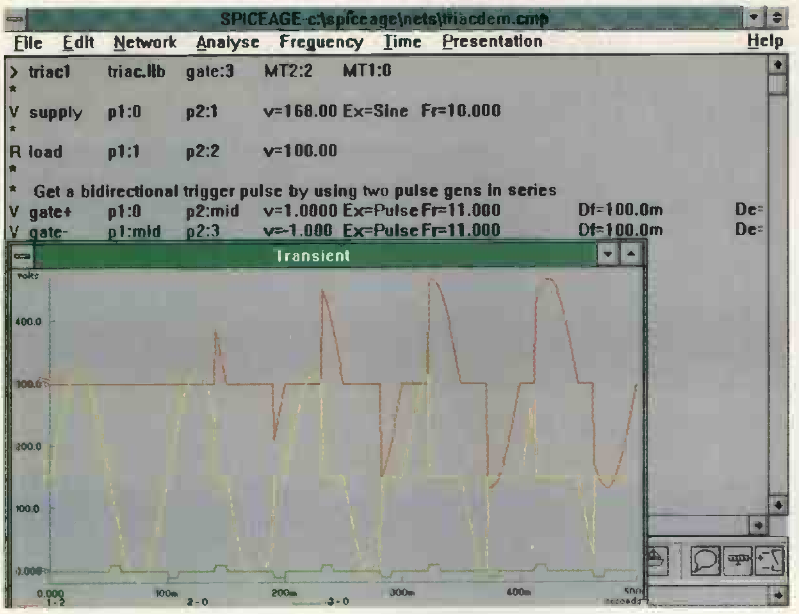
Halfway through entering a circuit I decided I should save the file. But when I tried, the only response was a beep from the computer, no disc access and no message. Indeed as much as I tried I could not save the file either as itself or as an alias. Using the *Windows* clipboard eventually allowed me to save file, and I then looked into the problem. What I found was that there was a syntax error in one line and *SpiceAge* was refusing to save until the file was syntactically correct. Great idea – but a message to that effect would have helped!




*Transient response window for a filter*

*Butterworth filter complex plane response using SpiceAge for Windows*






*Complex time response for a triac circuit*

## Worthwhile windows?

The program's use of *Windows* is a curious mix of good and bad. To its credit are the neat autoscaled graphs of results ready for output as metafiles or direct to printer. On the debit side is the almost illegible text, the inability to bring the edit window to the front and the need to discard analysis windows because changes in the component node list file are not transferred to the current analysis.

The only way round this is to analyse for a particular component setting and icon the graph for later comparison with different component settings.

The different windows that pop up for setting graph colours, selecting the nodes to monitor and other features are utilitarian and intuitive to use.

*SpiceAge for Windows* is a very competent analysis tool, capable of taking on most electronic and dynamic systems models. There are one or two rough edges, but it is, without question, a really good cae tool which every engineer should consider for tackling analogue circuits.

That said, what the package really needs is schematic capture and netlist, so that the node list format can be extracted from the actual circuit diagram. The advantages this would give are clear – minimal effort expended inputting data and a guarantee that there are no discrepancies between circuit and model.

Development of the software has been going on for two years, in a continuous manner (indicated by the test software revision number V1.182).

But its £395 price tag takes it into an area where some potential users might opt for a cheaper dos product instead. That would be a shame because the system under *Windows* is intuitive, easy to use and delivers the results. ■

### Supplier Details

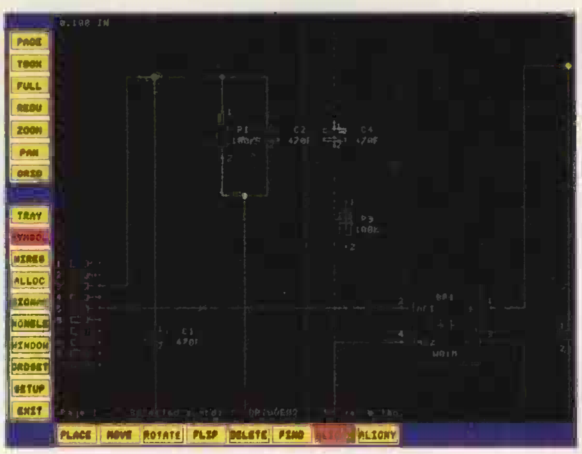
SpiceAge for Windows, £395 + VAT.
Those Engineers Ltd, 31 Birkbeck Road, Mill Hill, London
Tel: 081 906 0155 Fax: 081 906 0969

# RANGER 2: the shareware worth the layout?

*Seetrax has made a version of its Ranger2 software available on shareware. Martin Cummings says hurry while stocks last.*

Last year we reviewed *Ranger1*, an entry level PCB design package with its expensive stable-mate, *Ranger2*. It is a sign of the fast moving cad market that Seetrax is now offering a version of *Ranger2* free of charge, on shareware. The package includes schematic capture, PCB layout and autorouter, output drivers and a manual. Quality, features and performance of the package are a match for most of the competition. So what is the catch?

For many people there isn't one. In an interesting marketing ploy, Seetrax is offering a version of its software with all the features – but with restricted capacity – for nothing. The limitation is 32 parts or 128 component pins. Up to this point the package is a fully functional design tool: beyond the threshold you can design but can neither save nor output.




*ALIGN will ensure that components are exactly placed, ready for connecting up.*

Offering *Ranger2* as try-before-you-buy shareware is no doubt seen by Seetrax as a good way of demonstrating the software. Get a feel for all the functions, climb up the learning curve, then if you are happy, buy the full version without the limitations.

Some non-professionals, and perhaps those working in educational establishments, will probably not find the 32 part limit a problem. To them it is a very attractive offer, and I fully recommend they take advantage of it.

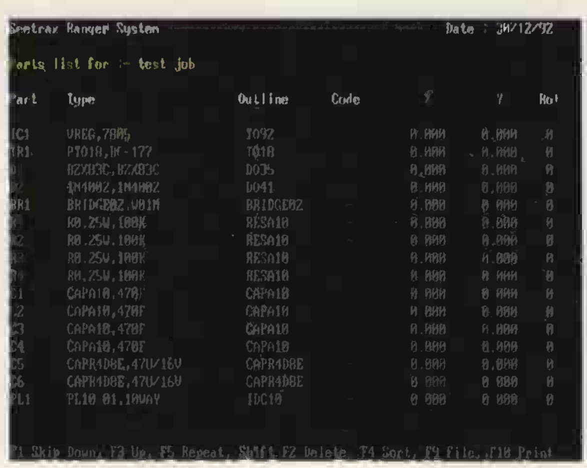

*Components are taken from the tray and placed on the schematic.*

**System requirements**

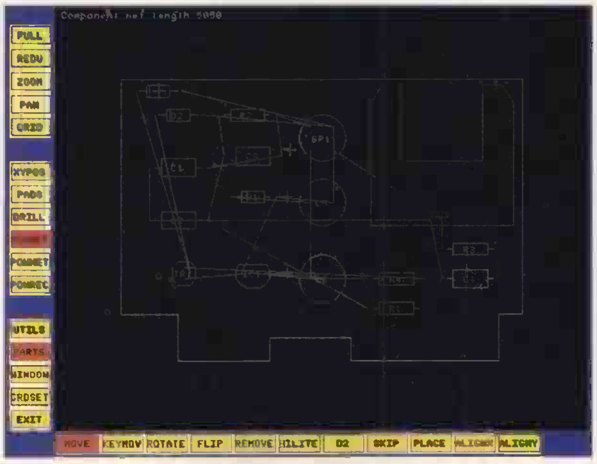
Hard disk
EGA display
Mouse or digitiser pad
Some high resolution screen drivers are supported up to 768 by 1024.

## Using Ranger2

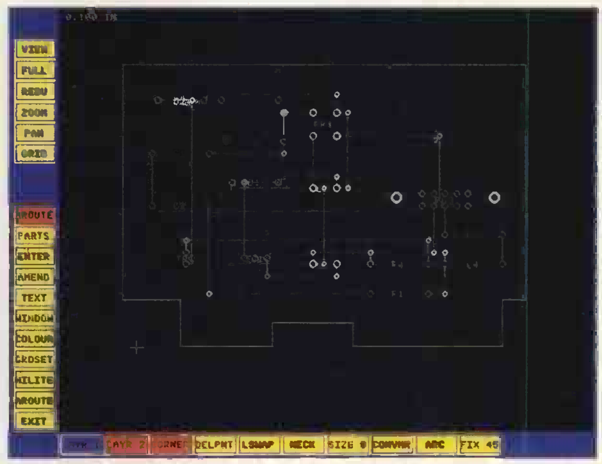
As with similar packages an outer menu or shell co-ordinates all the programs necessary in the design cycle. It organises all the relevant files for a job and allows jobs to be given a meaningful name up to 28 characters long – saving many



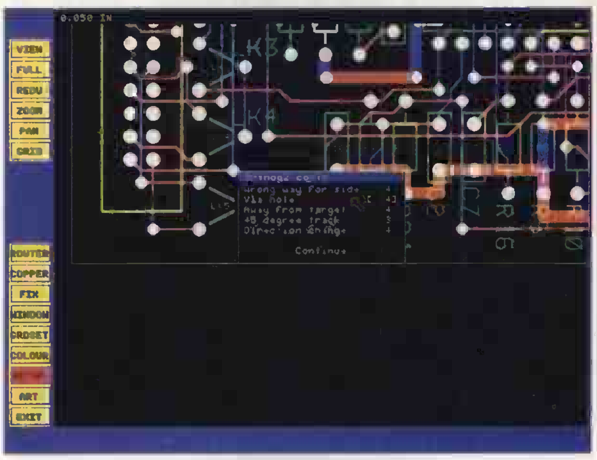

*Parts list can be printed or sent to a file.*



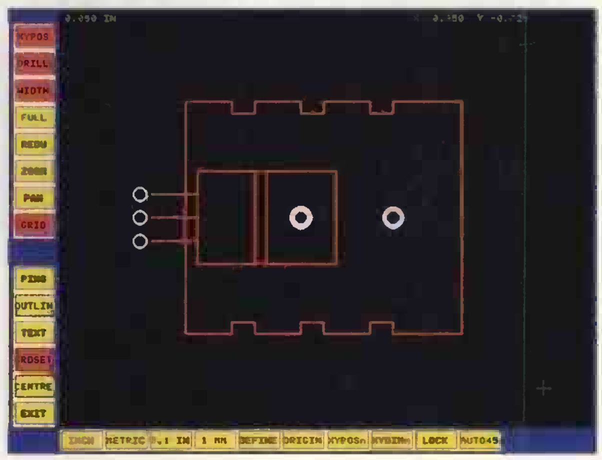
*When moving parts, signals are highlighted to show the effect of the move.*



*Far right This circuitry, routed with the autorouter, is the maximum circuitry possible on the free version.*



*Setting up the costs of the autorouter.*



*A library is built up for each job and can be listed on screen or printed.*

## Manual

The free version of Ranger2 is delivered on two 720k disks together with a professionally printed 90 page A5 demonstration manual. Most of the booklet is a training exercise; fold-out diagrams and plenty of text navigate you through a design from start to finish. The simple two sheet design uses about 30 components so gives an immediate feel for the complexity allowed by the free version.

Those who register and pay the £20 fee are rewarded with a 400 page reference manual.

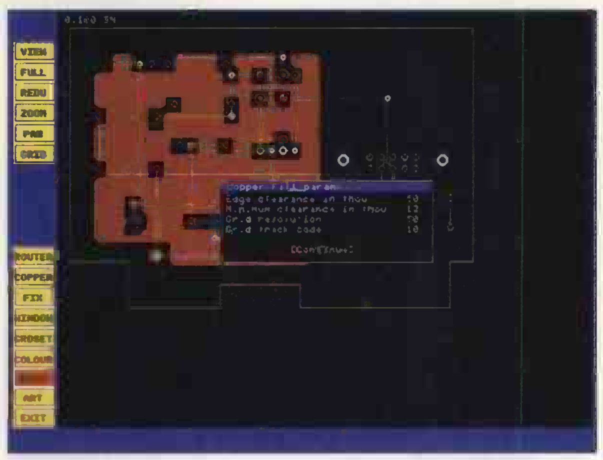
Software and libraries are expanded as they are installed onto the hard disk and eventually occupy 3Mbytes – the installation routine gives full control over where things go by asking questions at every step.

## Ranger 2 brief specifications

Up to eight sheets per design
16 layers
One thou resolution
Component flip for double sided placement
Automatic net length minimisation
Up to 32in square board
Signal highlight capability
Multi strategy autorouter
Gerber photoplotter output
Output for drilling and routing machines

## Supplier Details

Demonstration version (32 parts, 128 pins) free of charge
Shareware registration fee £20
Full version (1400 parts, 16000 pins) £599
Seetrax, Hinton Daubnay House, Broadway Lane, Lovedean, Hants PO8 0SG.
Tel: 0705 591037. Fax: 0705 599036



*Copper fill can be implemented on selected sections of the board, either solid or as a cross hatch pattern.*

hours thinking up awful abbreviations. Like all the main screens this menu is text based, but it is mouse operated and easy to use.

The schematic editor screen layout has menu boxes down the left hand side and along the bottom. They are well organised into screen control functions such as pan and zoom, with a separate set for design work. Operation has a nice feel to it, the menu boxes turn green as you move over them then turn red when selected – there is also no inadvertent operation as was found in the *Ranger1* review.

Five fixed magnification levels can be selected and one or two levels can be jumped depending on which mouse button is pressed. It might be useful to have variable magnification selected by defining a zoom window, but the fixed levels seem adequate in practice. In addition to mouse control, all the screen control functions can also be keyboard selected, for example F1 for zoom, space bar for pan.

The grid can be called up on both schematic capture and layout editors. On schematic capture the grid pitch is set and you have to live with it. Pitch is adjusted according to zoom level and it is important when drawing to know that *Ranger* will snap onto either the full or half grid position.

The artwork editor allows full control over grid pitch – I would prefer full control on both but have to accept that the automatic adjustment is good enough in most cases.

Plenty of scope is available to draw a schematic. It can be spread over up to eight sheets, each of which can be from A5 up to A1 in size, and is an extremely useful function for organising circuits onto sheets relating to function with well defined interface signals between them. (But don't get too excited about this ability if you are limited by the demo software to the 32 parts.)

### 28 volumes of components

To add a symbol, select it from the master library and put it in what is called the tray. Components can be searched for in the libraries. So if looking for a *1N4002*, type in 1 and it can find over 300, 1N and it knows of 33, 1N400 it can find 7 and 1N4002 it finds just one.

The full version of the software comes with 28 volumes of components, each volume being a category, for example volume 12 is Zilog microprocessor devices, volume 25 is A-to-D converters, and so on. The demonstration software only includes 11 volumes although this still leaves a surprisingly good spread of components – particularly for the amount of money involved – and even includes some surface mount devices.

As a design is built up, parts are transferred into a job specific library. The device library editor shows all the components so far used by the job. By clicking on a device a text screen lists seven key details such as symbols per device, terminals per device, which outline to use, and so on. Creating a new device involves entering these details and drawing the graphic symbol to use on the schematic.

Several configuration screens allow pad and drill and track sizes to be defined. Various shapes of pad are possible including square, rectangular and round ended. Track sizes from 1in down to 0.001in can be defined in steps of 1thou and the press of a key will switch from imperial to metric units.

Of the 16 layers, layer 0 is reserved for drilled pads, layers 1 and 2 are the outer copper layers, and all the others can have their function defined as copper, silk screen, ground plane, or several other power planes. The colour for all screen items can be selected.

Moving symbols around the screen is easy and smooth. The align function makes later connection simplicity itself. First of all choose a component with which others are to be aligned. Then, when in either X or Y align mode, click on any other components and they jump into alignment.

The result is a neatly arranged schematic and, more importantly,there is no messy position adjustment to straighten wires when connecting up. The feature is simple and very effective and one that I have not come across elsewhere.

The final step is to allocate part numbers, *R1*, *R2*, etc, and values. Part numbers can be allocated automatically at the press of the ALLOCATE ALL menu box, or it one by one manually if required.

### Completed schematic

Once the schematic design is complete, *Ranger* will generate parts and wiring lists. Both can be printed or edited and with the addition of user part numbers the parts list could form the basis for a bill of materials. Any unused pins can also be listed – although this is probably not as useful as it sounds.

Having defined the outline of the board, component placement is performed manually. *Ranger* attaches each component in turn to the cursor and as it is placed the next is ready and waiting, so placement is as fast as the mouse can be clicked.

During this exercise the signals can be made visible and will rubber band to help optimise placement. The X and Y alignment feature so useful for schematic drawing is equally pleasurable to use for component alignment.

Tracks can be laid manually. But the reasonably competent autorouter will do the job, following four different routing strategies for power, memory, and two types of orthogonal layouts. Any two layers can be routed or the autorouter can be configured to route in single sided mode.

In addition, costs can be selected for certain routing compromises and this will further influence the layout achieved because the algorithm will always try to minimise costs. The autorouter will also fill selected areas with copper leaving clearance around pads and tracks. Copper can either be solid or a cross hatch pattern.

*Ranger*'s design rule checker will check the complete design to ensure a specified clearance has been maintained. Some other packages provide a list of violations in the form of a printout but *Ranger* puts a marker on the artwork so the problem can quickly be located and assessed.

After all the work has been completed and any clearance problems resolved the artwork can be printed or plotted. No guidance on this is included in the demonstration manual so there must be a voyage of discovery through the menu screens (though I have it on good authority that the most recent release includes a chapter on output in the README file).

The design can be output to the usual selection of dot matrix printers, laser printers, pen plotters or photoplotters. Several levels of quality can be achieved with dot matrix printers by selecting single, double or quadruple passes of the print head. For those operating on a shoe string, it is surprising how good a multipass dot matrix print turns out.

### Best user interface

In general, the user interface is one of the best I have come across. Screen update is so fast that it ceases to be noticed. Moving connections and tracks takes a bit of practice, but adjusting the position of components and text is smooth and natural and the commands are all readily available at the press of either a mouse button or key.

There is no doubt that, without the 32 part limitation, *Ranger2* is a competent design tool with an easy to use personality and sufficient features to compete well with the many other packages available.

The version distributed free shares all these benefits and at first sight appears too good to miss.

It is fair to say that many of *Rangers*'s good features are wasted on small circuits; for example the autorouter, multi sheet schematics and automatic component numbering will be of little or no benefit.

There are no catches, but take a look at typical circuits that are likely to be laid out and count the components before excitement gets the better of you.

Those in secondary or further education should also take note. As well as demonstration software the shareware version is also a superb training aid. If you want to learn about integrated schematic capture and PCB layout, or you need to teach others, this is an excellent opportunity and a very good deal. ■

## Did the earth move?

In response to the queries raised by AJ Quinton (*EW + WW*, Letters, August 1992) and Martin W Berner (Letters, January 1993), there is much evidence to show that special relativity's postulate of the absolute constancy of the speed of light is incorrect.

Devices such as the navigational laser-ring interferometric gyroscope, the lasers used for geodetic surveying, the wide-angle Michelson interferometer for Doppler imaging, the global positioning satellite, the monitors of wavelike perturbations in the ionospheric F region, the behaviour of the transponder in Pioneer 10, right down to ordinary police radars, all show that $\lambda v = c \pm v$, not $\lambda v = c$.

If special relativity is incorrect, then the famous zero-velocity result of the 1887 Michelson-Morley experiment, and the less well-known positive result of the 1924 Michelson-Gale experiment favour the geocentric paradigm, that is the earth really is stationary in an absolute sense at the centre of the universe, and that it's the latter that's doing the moving!

This is the opinion of an increasing number of top-notch PhD scientists, such as the Tychonian Society in the USA and the Cercle Scientifique et Historique in Europe (see "The earth is not moving – 400 years of deception exposed", M Hall 1992).

Recent independent research in Australia and Russia has shown that there has been a decrease in the speed of light, which has many far-reaching effects in all areas of science, and implies that the usual estimates for the age of the universe have to be downgraded from billions to only thousands of years (see Richard Milton's "The facts of life – Shattering the myth of Darwinism", 1992).

In addition, a decrease in the speed of light has the effect of shifting light from the stars to the red end of the spectrum, just like the Doppler effect, and voids the necessity to interpret the red shift as being due to an expanding universe.

Contrary to the impression given by the media, close examination of the evidence for the Big Bang makes it turn out to be more of a damp squib (see E Lerner's "The Big Bang never happened", 1992).

***Arenon Goldberg***
*London*

## Light fantastic

There must be many readers who, like Martin W Berner (*EW + WW*, Letters, January 1993) are mystified, if not completely baffled, by the theory of Doppler shift as applied to light.

The mystery, I suspect, stems from Einstein's postulate that the speed of light ($c$) as measured by all observers is independent of the speed of the source or the observer.

An alternative theory (Letters, November 1990 and March 1991), postulates that an observer moving with velocity $V$ towards a light source would measure the velocity of the light to be the vector sum ($c + V$).

The theory is difficult to prove by experiment because, except for high–energy particles, nothing on our planet moves at a speed comparable to that of light. For instance, satellites orbit the earth at about 8000m/s, which is only 0.000027$c$. So, for most practical purposes on earth, $(c + V) = c$.

However, many galaxies beyond the Milky Way are known to be travelling at speeds that are a high fraction of the speed of light, and astronomers measure the Doppler Shift in the light from such galaxies to estimate their speeds of recession.

Einstein's postulate implies that the speed of galactic light reaching Earth from such galaxies is the same as the speed of light from a local source. This, I believe, has lead to some misinterpretation of astronomical data.

For those who are interested in the mathematics: According to Einstein's theory, if a source of frequency $F$ is moving with velocity $V$ at an angle $\theta$ relative to an observer, then the observer would measure the frequency to be $f$ where:

$$f = F\frac{\sqrt{1-\frac{V^2}{c^2}}}{1-\left(\frac{V}{c}\right)\cos\theta}$$

But according to the alternative theory

$$f = F\sqrt{\frac{V^2}{c^2}+2\left(\frac{V}{c}\right)\cos\theta+1}$$

Consider the case when radiation from the source is observed transverse to the direction of motion, that is $\theta = 90°$.

According to the first equation

$$f = F\sqrt{\frac{c^2-V^2}{c^2}} \qquad 1$$

and from the second equation

$$f = F\sqrt{\frac{c^2+V^2}{c^2}} \qquad 2$$

Expressed in plain English, according to Einstein's theory, if the relative velocity $V$ is at right angles to the line connecting source and observer, then the observed frequency $f$ is lower than the transmitted frequency $F$ (as in equation 1). According to the alternative theory, the observed frequency $f$ is higher than the transmitted frequency $F$ as in equation 2. A test using a laser pulse transmitter (prf controlled by USO) and a receiver plus counter/timer would determine which of the two theories is correct. Any suggestions would be welcome.

***John Ferguson***
*Camberley*

## Power response

I would like to respond to two of the comments that have appeared in letters pages about my article "Natural radiation focused by power lines" (*EW + WW*, November 1992).

Harold Kirkham's letter (*EW + WW*, March 1993) neatly pinpoints the areas that need more research, but I can answer some of his points.

First, his comment on the labelling of the graph - it should be "of normal" as he suggests - *mea culpa*.

The Geiger tube I used for the field measurements detects charged particles over 50keV which penetrate the charge space inside the

## Water winner

With regards to Peter Wivell's request (*EW + WW*, Letters, February 1993) for help in his project to improve power distribution to remote villagers in Nepal, my suggestion would be to eliminate or reduce the distance the batteries have to be moved. I do not know much about Nepal except that it is a mountainous country and, I assume, has many fast-flowing streams.

I therefore propose the construction of many small charging stations (one per dwelling, even) using a vehicle alternator or old dynamo driven by a water wheel. A belt drive may be required to run the machine at a suitable speed.

Given a supply of used or military surplus electrical machines, which should not prove too difficult or costly to secure, the technology involved in this is certainly within the capabilities of the villagers.

Another possibility which is rather more innovative is a solid fuel thermoelectric charger. Only copper and iron conductors would be cheap enough, and about 300 hot junctions are required for charging a 12V battery if the temperature difference is in the region of 400K. Nevertheless it may be practical – I seem to remember a report many years ago of a radio powered by a paraffin lamp developed for the eastern USSR.

For a discharge limiter, a simple solution would be a relay whose coil is in series with a zener diode across the battery and whose make contact is in series with load. It might be possible to make the relays locally if magnet wire were supplied. The zener diodes would have to be brought in but are not costly items. To prevent chattering, the relay armature might lock in the released position until manually reset. This is a simple mechanical matter.

***John Woodgate***
*Rayleigh*
*Essex*

tube. Biologists classify charged particles up to 1MeV as most destructive to cells.

The graph showing the zone of interaction is a distillation of many hundreds of field readings with the detector tube parallel, or at right angles, to the line at varying distances.

Here's how the figures break down: tube self count 30/min; average background away from line about 50/min; this gives a net sky plus ground count of 20/min; and this is factored 33% sky, 67% ground or about 7/14.

If the ground count is stable and remains at 14, any increase in recorded count will be due to extra sky particles. To get a doubling of sky rate the total count must rise to more than 58/min.

To make sure that the figures used were conservative, I double checked all field readings against the recording taken on the fixed detector in the lab. Those coinciding with a solar emission peak were not used.

The exact particle dynamics near a multiphase power line are complicated, and will not be as simple as Kirkham makes out. True, the net electrodynamic effect of an individual conductor will be zero over each cycle, but the current and voltage curves may not be accurately in phase and adjacent phases will be 120deg different, so the electrodynamic effect between phases will not average zero unless each particle passes through the electrodynamic balance point between the lines at the correct angle. All interactions will be nonlinear because of the inverse square law effect of distance from the line. The particle interaction between crossing lines, or entering switchyards or distribution networks on the upper floors of large buildings will provide great scope for future analysis!

DE Jeffers of the National Grid Company makes some more general points (Letters, February 1993).

First, I totally agree with Professor Doll's report which found no firm evidence that the electromagnetic radiation from power systems is a direct carcinogenic hazard.

The effect I described is quite different, and fits the observed epidemiological data published since 1976, much better than any other explanation to date. It also has the great virtue that no new disease mechanism has to be invented, because the effects of penetrating radiation are well documented, and current bioresearch shows that all ionising radiation has an effect on cell replication-even at levels well below current safety limits.

The other important point to come from my research is that sky radiation is much more solar emission dependent than suspected and that secondary charged particles are highly organised by the ambient electric and magnetic field as they reach the ground.

This means that people living and working near power lines are subject to wide variations in natural radiation over the solar cycle, especially if they live at altitude or high geomagnetic latitudes where the magnetosphere concentrates incoming solar particles in the auroral zone.

There is some independent evidence for a geomagnetic latitude effect on cancer statistics. Taking the so-called radiation cancers we find that the IARC lists male average incidence at 300 per million for Scandinavia and 150 per million for central Europe. There will also be an altitude effect. It may not be coincidental that the first studies to show a link between power line routing and cancers in the 1970s were based on Denver which is at 5000 feet where the sky background rate will be at least double.

A good test of my hypothesis that solar particles can influence cancer statistics close to power lines would be to see if there is an 11 year period in childhood leukaemia cases.

There's lots more to find out!

***Anthony Hopwood***
*Upton-on-Severn*
*Worcester*

## TV makes you sick

Video games have recently refocused attention on the danger of epileptic fits being triggered by high-intensity visual stimulation at certain pulse repetition frequencies. The effect is of course well known. I remember studies being done on it several decades ago at the National Hospital for Nervous Diseases, London, and the Medical Research Council's Applied Psychology Unit, Cambridge.

Now that television production companies are increasingly using computer-generated graphics for title sequences, commercials, and other visual effects in programmes, the possible danger to the public could be greatly extended.

Some of the optical patterns use regularly pulsating areas of high-brightness, high-saturation colour. The purpose is obviously to capture the viewer's attention by getting him or her emotionally excited through the visual cortex of the brain.

As free-market competition for ratings gets fiercer and Britain's television moves further and further down market as a result, this kind of assault on the sensory nervous system could well intensify.

Do these production companies, and the broadcasting organisations, really know what they are doing? Have the graphics designers and producers paid any attention to the neurophysiological effects on viewers that could result from this kind of visual stimulation?

I don't think I am myself vulnerable to epileptic fits but have certainly experienced short attacks of nausea or vertigo while viewing some of these graphics effects – and I don't just mean aesthetic dislike.

***Tom Ivall***
*Staines*
*Middlesex*

## Twin speakers

In response to PC Meunier's letter "Speakers in series" (*EW + WW*, November 1992) issue, connecting loudspeakers in series is often convenient but I have never seen it advocated for high-fidelity applications. In fact, way back in 1958, GA Briggs in his brilliant book *Loudspeakers* pointed out: "The virtues of parallel working are already well known".

I guess though that non-hi-fi loudspeakers were Meunier's case in point since there is no mention of damping factor.

In cases where different types of loudspeaker are to be connected together then I can imagine that transformers allowing parallel connection might offer a better solution.

Obviously no two speakers are identical, but when the speakers are the same type, surely connecting them in series causes less degradation than introducing all the non-linearities of a transformer into the circuit? As a bonus, the series solution is cheaper, lighter, less labour intensive, and possibly even more efficient.

***Martin Eccles***
*Newcastle-under-Lyme*
*Staffordshire*

## Will to work

I live among what is left of British industry in the north west of England near a wool combing factory. Most mill owner's, like the consumer electronics industry (such as making radios and televisions), took their money and ran when it became cheaper to produce in the Far East.

It is not that industry needs to get off its bottom, people, as we have seen with the miners, are desperate to work. If industry could produce goods that are wanted at a reasonable price, rather than importing goods to an ever increasing extent, there would be hope for us all.

***Peter C Gregory***
*Ashton-under-Lyne*

## The ears have it

I read Ben Duncan's article "Proof of the golden ears hypothesis" (*EW + WW*, June 1992) with much interest concerning what I take to be noise and harmonic distortion improvements by using series connected capacitors.

Rather than explain this away without practical examples, as Phil Denniss did in "Distorted Proof" (Letters, November 1992), I have tried the idea in a Quad 405(2) amp issue 7 on $C_{18}$, $C_{19}$, and $C_{10}$. Improvements I noted were in the near total absence of mains noise (hum). I have noticed no sonic degradation comparing my amp to another of the same specification.

I have also tried an earlier suggestion concerning $R_{12}$ and $C_6$ and my own idea of replacing $IC_1$ with an *AD847* op amp, all with good results.

I consider the practical application of ideas should have a larger voice than those who elect to put pen to paper rather than plug in their soldering iron and at least try.

Furthermore, the market place continues to dictate an interest in valve equipment despite its poor harmonic distortion. Perhaps chasing mathematical perfection without reference to actual listening is an aim which satisfies some, but falls short of any real involvement in why musicians seek to provide enjoyment for us the listeners.

***Chris Daly***
*Tasmania Australia*

## Beavering a way across land

Your reply to Andrew Ainger (*EW + WW*, December 1992) that communication via earth probes "emerged in the early 50s" was roughly right, but you didn't say which century.

Almost from the inception of the telegraph, attempts have been made to signal via probes in earth and water, initially using sensitive galvanometers and later headphones. Ranges up to 1km have been reported.

In World War I, field telephones using an earth return were vulnerable to eavesdropping.

Around 1960, a published article claimed a range of a mile using CW via an audio oscillator/power amp, transmitter, and voltage amp receiver.In the 1960s, I chose earth probe communications as a youth club project. Transmitters used were a 20W valve amp and a 4W transistor portable amp, both having multiple tapped output transformers to match the earth impedance, which varied with moisture content. Receivers were two-transistor headphone amps.

We seldom did better than about 0.5km, although with up to 30 eager young beavers to keep occupied, the situation was not conducive to serious investigation. Contrary to Ainger's experience, there was plenty of interference.

RF filters were required to attenuate broadcast signals (the entire AM band simultaneously), together with 300Hz high-pass filters to remove 50Hz+ harmonics from supply system earth leakage. Even then, higher harmonics were present with up to the 20th being easily heard, and I suspect this was the main factor limiting receiver sensitivity.

A tuned carrier seemed a good idea, and a system consisting of a 4W power oscillator on 80kHz, with an LF converter feeding a receiver with BFO at the other end, was tested. Results were poor, but before this could be investigated the project was abandoned.

There may be a snag. The earth, being a conductor, will radiate, and receive, electromagnetic waves, even if inefficiently (see above, re AM interference). If this happens, the earth station merely becomes another transmitter adding to the existing spectrum load.

Hopefully there may be bands of frequencies too high or too low to be radiated in this way.

I commend Ainger in his efforts, and will applaud if he reports answers to his CQs. I will applaud loudly, with astonishment, if he can generate a signal in Harpenden which is heard in New Zealand.

His linear resistance wire model is not a good analogy for earth impedance. Most of the voltage drops occur in the immediate vicinity of the probes, where earth current density is high.

Between probes, the earth is, for practical purposes, of infinite cross-section. So the measured impedance between two probes has more to do with probe contact with the soil than with conductivity or distance between the probes (though these should be as far apart as possible.

These questions have been well researched by the electricity supply industry and literature is available.

I suggest the circuits shown should include a DC blocking capacitor in series with the probes to prevent electrochemical potentials from the probes upsetting the amplifier DC conditions.

In a permanent installation, some means of isolating the probes during thunderstorms is mandatory since lightning strikes can generate dangerous or even fatal earth potentials.

If he contacts me (by mail, not earth!) I can supply further details of our tests including a method of measuring probe efficiency.

***Ronald Salter***
*Victoria Australia*

## Lipschutz defence

Perhaps the most insulting aspect of WF Blanchard's letter *(EW + WW*, February 1993) is his description of Captain Lipschutz's work as "ideas". As was pointed out in the original article (May 1992), the version of the equipment using the radio method was well worked out and built. It was intended, as was also pointed out, as a simple method of demonstrating automatic plotting, which would have been the preferred method of display for the inertial system to come if the first demonstration had attracted more than a cursory glance.

Lipschutz was not an inexperienced navigator at the time; he already had his pilot's licence and was employed as a radio officer at Lydda. Indeed, his work was held in such high esteem that it was the authorities there who asked him to submit his work. He was, and is, no amateur.

Errors caused by skywaves would be relative to the power of the beam and the main beam would predominate. Even lower frequencies have been used for many years in ADF beacons that surround the globe, with very little trouble from skywaves. Airliners still use beacon-hopping as a valid method of navigation, although the need for it is now, of course, diminished by other means of navigation.

As regards the plotting relative to true north, means not described in the article afforded a method of carrying that out. It was incorporated in a differential drive that would allow the aircraft to bank and turn without affecting the results.

Lipschutz was largely responsible for installing the Adcock antenna at Lydda and was responsible for feeding data to aircrews. He was fully aware of the characteristics of Adcock antennas. But the main thrust of the article was intended to show that the arrogance experienced by Lipschutz probably cost many aircrew lives. Since the equipment was tested and shown to work as expected, it seemed at least possible that someone, somewhere might show some interest, in view of the fact that navigation was in such a parlous state that more crews were lost when trying to find their home bases than were shot down over enemy territory.

It does no one any good at all to denigrate the work of others, particularly in wartime when so much is at stake. Or in peacetime either, for it is still happening. How much highly marketable ingenuity has found stony ground in the UK since the war and yet has been able to flower for another country's profit?

***Philip Darrington***
*Appledore*
*Kent*

## IFF guide

With reference to the correspondence on navigation aids (*EW + WW*, February 1993 and earlier), from 1941 onwards all coastal command aircraft not fitted with centimetre radar had at least ASV Mk II (air to surface vessels, 176MHz), even Beaufighters.

We used to modify IFF (identification friend or foe) sets to act as homing beacons. Since this was then a secondary system, these beacons worked to maximum radar range (100 nautical miles) particularly with Wellingtons which had stacked dipoles on the fuselage as well as yagis under the wings. They were much appreciated by the air crews.

Of course they could have been used by the Germans to attack our bases but I personally saw only one such attack when the hangar which housed the beacon at Wick in Scotland was machine gunned by a reconnaissance aircraft early in 1942.

We used these beacons all over the Mediterranean subsequently.

***Eric Carr***
*Basingstoke*
*Hants*

# GPS

## 5: the software engine

*Having covered the radio and computer hardware, Philip Mattos returns to the software design. Conventionally the reduction of the spread-spectrum signal to a single carrier is done in hardware. To do this in software is unique, and demands a very fast processor.*

The GPS signal arrives at the antenna some 20dB below the noise, at 1.5GHz, spread over a 2MHz bandwidth. We rely on the radio system to amplify it to a level that can be digitised, and to downconvert it to a frequency that can be sampled. It cannot improve the signal to noise ratio however.

Extraction of the signal from the noise requires correlation with a code identical to that used in the satellite to spread the carrier, and the code must be exactly in synchronism with received signal. 'Correlate' means multiply each sample of signal by the appropriate bit of the code, and integrate over the length (epoch) of the code, or several epochs.

This operation gathers in the energy from the 2MHz bandwidth, back to a single carrier, plus the 50 baud data. In the 1ms code epoch it gives 30dB S/N gain; in the 20ms data bit time, 43dB. These are of course theoretical numbers, but practice comes close if it can be made to work.

To do this, one has to simultaneously find and hold the satellite in the time and frequency domain, to an accuracy of about 10ns and 100Hz. This is interesting when the CPU timer has a resolution of a microsecond, and an interrupt response of up to three microseconds. Read on, it can be done without any hardware assistance.

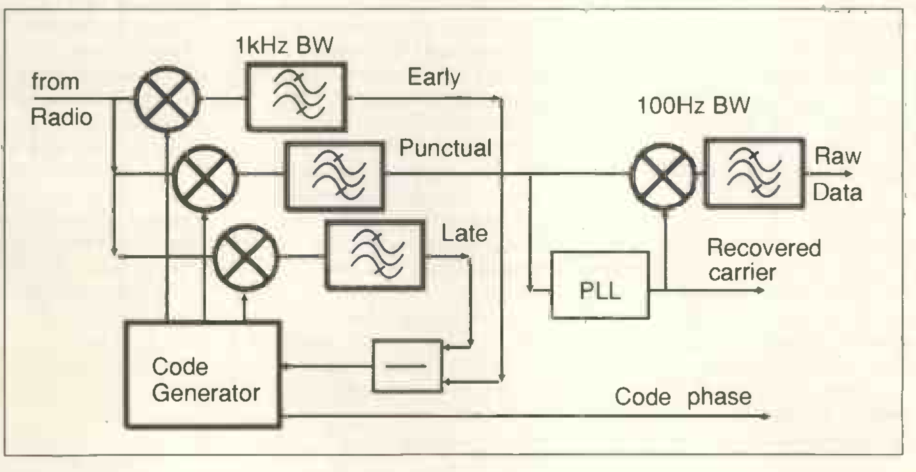
Once the system is locked, it can be easily understood. The traditional all-hardware design, circa 1980 is shown in **Fig. 1**. The signal is filtered only after correlation, so the frequency must be scanned until it hits the filter. However the code generator must also be drifted through all possible phase relationships with the signal, and both must be correct simultaneously.

This process is known as acquisition, and takes considerable time.

Once the system is locked, it must be held locked by tracking the code-phase, the carrier phase, and the extracted data modulation.

The data must be interpreted to give the required information about the satellite orbits

*Fig. 1 Fully hardware system (circa 1980). The original hardware systems adjusted the LO to hit a narrow filter, then ran three correlator channels, to maintain the tracker, the central one to extract the data in a phase locked loop. This took a PC board per satellite.*



and clock errors, then in concert with the timing information derived from the code tracker, the user position is calculated.

Finally, most systems also include a layer of navigation functions, such as distance to next waypoint, estimated time of arrival, off-track error etc

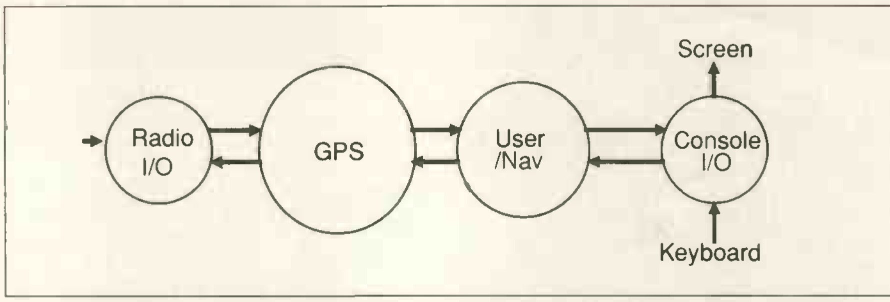



*Fig. 2. All-software system overview. A parallel processor like the transputer allows simple programming, as if there were a CPU per task. Only the console I/O process is hardware dependent, and then only at the lowest level.*

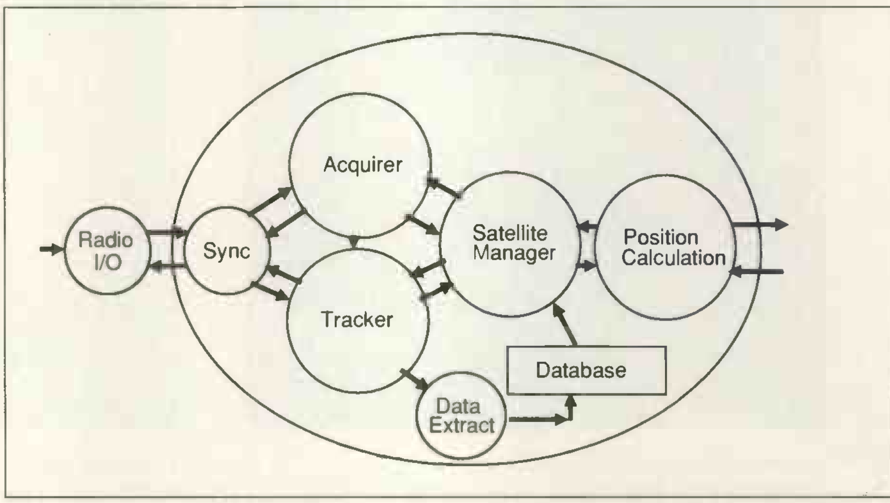



*Fig. 3. Breaking down the GPS task into its sub-processes. GPS is by far the most complex process, both to design and to compute. Most of the CPU time is used in the tracker, after acquirer has finished, but the complex maths is in the satellite manager and position calculator*

## Software structure

The software is organised in a series of parallel processes or tasks, as shown in **Fig. 2**. The transputer is a parallel processor executing any number of tasks apparently simultaneously. It can do this because it has a hardware scheduler, almost a thousand times faster than the equivalent function performed in a software operating system.

It is thus programmed as if there were a CPU available for each task. This makes each task simple and efficient, while the hardware-managed communication between the tasks gives automatic synchronisation.

There are five major processes in the system between the input from the radio, and the output to the screen. The first two run at high priority, the rest at low.

**I/O**. The first is the I/O process that handles the data from the radio. The data is packed bytes of one bit samples. As the system samples at 2MHz, a byte is delivered every 4μs. The I/O task runs once a millisecond, and simply executes an input instruction to load 256 bytes of data from the transputer link. This takes less than a microsecond to execute, but must be done immediately the buffer is full, as it is only 4μs until the next byte. Alternate milliseconds are loaded into two different buffers ("ping-pong" buffers).

**Synchroniser**. There is not sufficient time, even with the transputer's rapid re-schedule time, to manage communications from this task in under the 4μs limit. Thus communication with the next real task, the GPS task itself, is not direct, but via a synchroniser task. There are three shared memory areas between the I/O and synchroniser tasks... the two buffers, and a sync word.

The I/O job does not run on milliseconds timed by the transputer clock, but by counting the bytes of data from the radio. Thus it is the sample clock that controls it. They may or may not be the same... on a portable, they would be, for economy of crystals, but on a PC-based GPS, the transputer clock is supplied by the mother board, so differs. Thus each time a buffer is full, the I/O job reads the transputer timer and puts the value into the sync word.

The synchroniser job reads the sync word, and waits to one millisecond later, plus a 30μs safety margin, before accessing buffer A, then another millisecond, then accesses buffer B. This method guarantees that the sync job is never using the CPU at the instant the I/O job needs to swing the buffers, and also guarantees that processing is never done on an input buffer while it is being written.

The synchroniser job simply idles maintaining sync until it gets a request from the GPS job for a block of data, which it then delivers over an *Occam* channel from the appropriate buffer.

**GPS**. The GPS job is the main part of the task, and is in fact some six tasks internally as shown in **Fig. 3**. The main tasks are

(1) to acquire the satellites, that is to get into lock in the time and frequency domain;
(2) to track them, ie maintain lock, a much easier task;
(3) to manage them, that is select the correct ones according to elevation, geometry, health, etc;
(4) to download data from them, in order to know their orbital positions and other system information;
(5) to manage the data, purging that which is out of date;
(6) most importantly, to calculate the user position.

How each of these is done will be covered in summary later... full detail would fill a book. Acquisition, tracking and downloading will be covered in detail seen from the flow of the signal point of view.

**User/Nav**. The user task is the one that implements the user commands, and generates the information requested for display. It does not actually handle the display, because that is hardware dependent, and thus isolated for portability to different licensee's screens/keyboards.

The navigation task takes the calculated position in WGS84 coordinates, and translates them into the coordinate scheme of the user's choice... my software supports the 47 datums of the GPS spec, in Lat-Long, UK Ordnance Survey Grid Reference, in UTM (Universal Transverse Mercator), and MGRS (Military Grid Reference System). OSGB is supported to one metre resolution for surveyors, or the conventional 6 figure, 100 metre version for leisure use.

Additionally, waypoints and route plans are supported here. Waypoints are simply stored positions, with textual names in my version, and route plans are threads through the waypoints. Normally entered before the journey, or in a yacht even before the season, one route is selected, and the receiver will then advise the distance and course to steer to the next point, the same from the last point, the expected time of arrival and many other parameters.

Part of the user task is the dialogue to allow selection of which parameters to display, because they cannot easily be displayed at the same time due to limitations of screen size.

**Console Handler**. As mentioned earlier, this is essentially a hardware driver to maintain

## PERFORMANCE

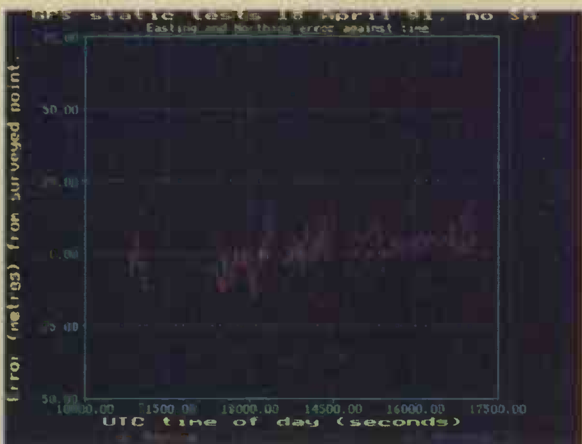
The performance of a GPS receiver is not easily quantifiable as it depends so much on the importance of particular features to a particular user. Accuracy should not be a feature, as the receiver does not contribute significantly to the errors in standalone operation. The errors are essentially the ionosphere and Selective Availability.

The figures show the accuracy of the system at a surveyed point over many hours, with multiple constellations. Left hand figure is just before the end of the Gulf war, when there was no SA, and also benefits from being 0200 to 0500 local time, when there is almost no ionospheric effect. There are only two spikes of error over 10 metres, and the RMS error is less than five metres.

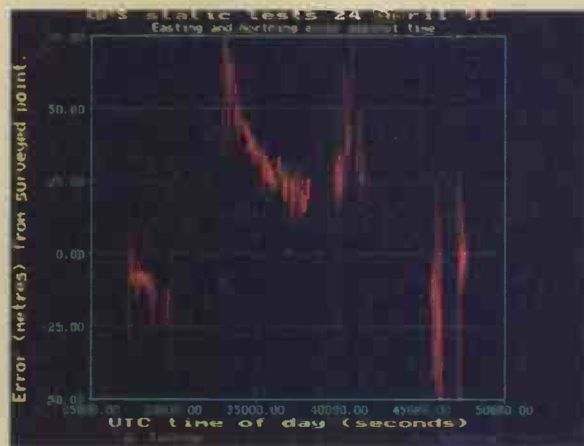
Right hand figure is a week later when SA was reactivated, and the steep curve down from 90 metres of error is typical of SA.

Cold start time, warm start time, and reacquisition time after obstruction are other measures, and for a portable, size, weight and power consumption. One example of a small light portable is the Panasonic *KXG-5500*, which has a low-cost 32-bit transputer as its computing engine.

Tolerance of reflections and partial obscuration by trees are parameters very hard to measure, and are often driven by the antenna in use, but for the car installation are very important. In the car, the user sees only the percentage availability of the system, as shown by the cross on his map-display.




*Performance without SA. The position plot against time for 18/4/91 shows the performance before dawn (no ionosphere problems) with an rms error of about five metres from a surveyed position, over four different constellation*




*Performance with SA. A week later, with SA turned on, and in the day time, the position wanders by up to 90 metres, but two receivers behave identically, so the error can be removed by differential means.*

portability. This is essential, as for example on one clients implementation, the screen driver bus was also the keyboard matrix column driver, so keyboard and screen were inextricably linked.

On another version a dual UART is used, to any appropriate terminal. On yet another, the transputer card is inserted in a PC, and the I/O is over the PC bus to a server program on the PC giving access to the PC's own keyboard and screen.

The level of message passing from user job to console handler is low. For example:

```
user.to.console ! display.string ;
SIZE
string::string
```

is used to put any length of text string on the screen.

### Software walkthrough

The signal processing software is best understood by following the flow of the signal through the software, rather than the flow of control through the code. It is also easier to understand the acquisition process, of finding time and frequency lock with the satellite, after the tracking process is understood, though in real life, they occur in the reverse order.

**Figure 4** shows the down convert and code track part of the software, needed to maintain timing lock. The incoming signal is one bit per sample, with the samples at a 2MHz rate. The transfer frequency from the radio is anywhere from zero to 2MHz, but band limited to about 1MHz. The first task is to mix it down to baseband, using a software local oscillator (LO). The LO is set 4kHz below the nominal carrier, but this has to be found empirically because temperature drifts of the oscillators can move it by a few tens of kilohertz. The output of this mixer is a signal centred at 4kHz, but still 1MHz wide... a problem in hardware, but here we are simply multiplying numbers together.

*A* x (*B* x *C*) and (*A* x *B*) x *C* are identical if there is no numerical overflow, so we can do either first. The benefit of doing the mixdown first is that it can then be shared between all the satellites and not repeated for each.

The second mixer multiplies the signal by a locally generated copy of the satellite spreading code, which reverses the spreading and

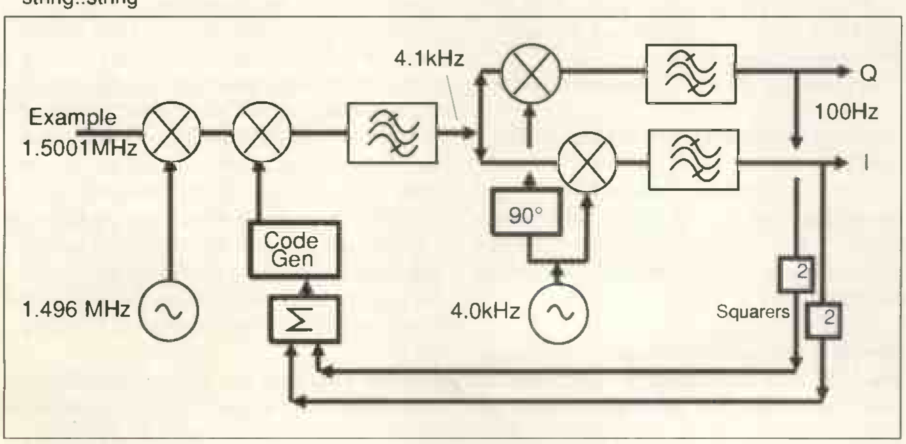



*Fig. 4. Downconvert and code track software. Shown as a signal flow diagram, as if it were hardware, the DSP software is very easily understood. This section runs both when downloading and when positioning, the prime output being signal for downloading, and code-phase when positioning.*

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regenerates the carrier, simultaneously spreading out any interfering carriers.

The benefits of the downconversion and the despreading are not seen until the resultant samples are filtered. As the frequency has been brought down to about 4kHz, not zero, the first filter is reasonably wide at 64μs. This is achieved by adding the samples in groups of 128, with the further benefit that the 2046 samples per millisecond become 16 samples, a much more manageable number.

The next stage is to bring the signal even close to baseband, and reduce the bandwidth further. This is done by mixing it with another software LO, again from a memory buffer, so CPU time is only used at start-up. 32 versions are held, at 250Hz intervals from 0 – 8kHz to allow for the doppler shift between the satellites even after the first LO has compensated for the radio temperature drift. The nearest channel is used, both as in phase and quadrature signals (I+Q): both channels are then low pass filtered to yield just one IQ sample pair per millisecond.

These are passed to the carrier tracker process for downloading data, but are also used internally to control the code tracker.

**Code tracking**. Due to the motion of the satellites, the motion of the user, and differences in reference oscillator frequencies, the relative time offset between the user and the satellite varies. To keep the synthetic code in sync with the received one, it must be adjusted finely – to about 10ns accuracy.

This is achieved by dithering the timing back and forward by one sample period, and maintaining two average amplitudes, one for the early version, and one for the late version. The normalised ratio between these two amplitudes, $(E-L)/(E+L)$, gives an error signal that can be used to manage the tracking in a stable feedback loop.

Note that the error signal does not indicate when to step the code generator phase by one sample. That would result in very coarse tracking. The code generator is run at a pre-calculated rate of drift relative to the reference clock, and the error signal, integrated over several seconds, is used to make fine adjustments to that rate. Only if the error is very severe is there a direct adjustment made to the code phase. This illustrates the flexibility of having all the algorithm in software. Further benefits are achieved by having different time constants for the loop when searching for satellites, and when locked, with a ramp in-between the two.

Independence of code and carrier tracking loops is yet another difference between this and conventional receivers. Many receivers achieve smooth code tracking only by feedback from the carrier tracker. While very effective on a large ship, this makes them fail very easily on land, with trees and reflections preventing the carrier tracker from locking

**Demodulation**. The 2046 samples/ms have now become just one IQ pair per millisecond. More important is that a megahertz wide signal has now become a kilohertz wide, a 30dB signal to noise ratio gain. The residual carrier frequency is now also very low, less than 250Hz.

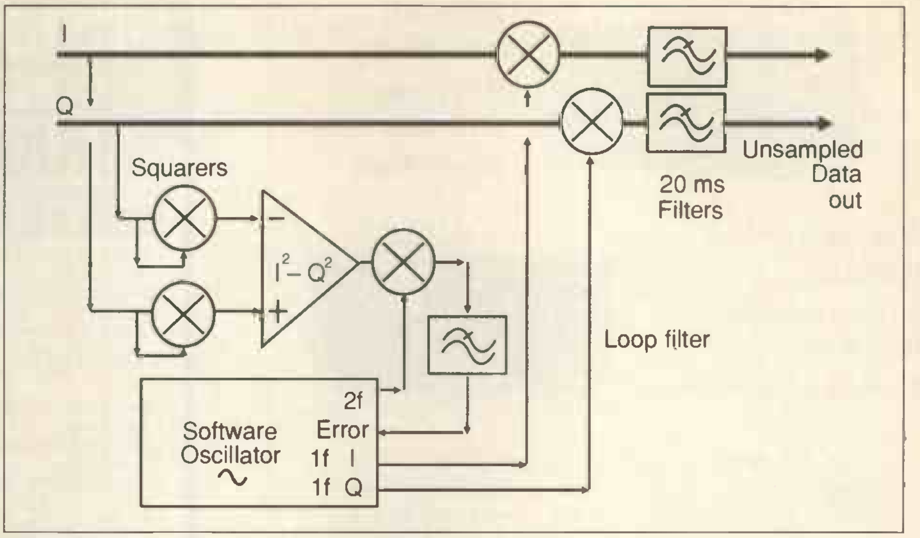



*Fig. 5. Carrier Tracker 2f loop. The carrier tracker only runs during downloading, and again mimics earlier hardware designs... with the advantage that a software oscillator can output several frequencies and phases, and take phase steps on command.*

The final step is to run a software phase-locked loop (PLL) to remove all the carrier, yielding a precise frequency measurement, and the demodulated data, **Fig. 5**.

The IQ pairs are squared to remove the data phase inversions, but this puts in a DC offset dependent on the signal amplitude. To avoid this, they are recombined as $I^2-Q^2$, which removes the DC and gives a continuous signal at twice the carrier frequency ($2f$).

A PLL is run on this, generating both the $2f$ signal to maintain the loop, and regenerated noise free I and Q carriers in phase with the original signals. They are multiplied by the original signals to yield two copies of the modulation. These are filtered in a 20ms filter to match the baud rate and reduce the noise bandwidth, and are then combined to reject the noise present from the image frequency.

The filters are classic integrating filters, and do not reduce the sample population, as we do not wish to smear data peaks. They simply add 20 samples together. However a classic implementation as shown in **Fig. 6a** requires 19 add operations per millisecond. The version of Fig. 6b requires just one add and one subtract, and the storage space for the 20 integers was needed in both versions. More important than the 18 arithmetic operations saved is the indexed loading of the operands.

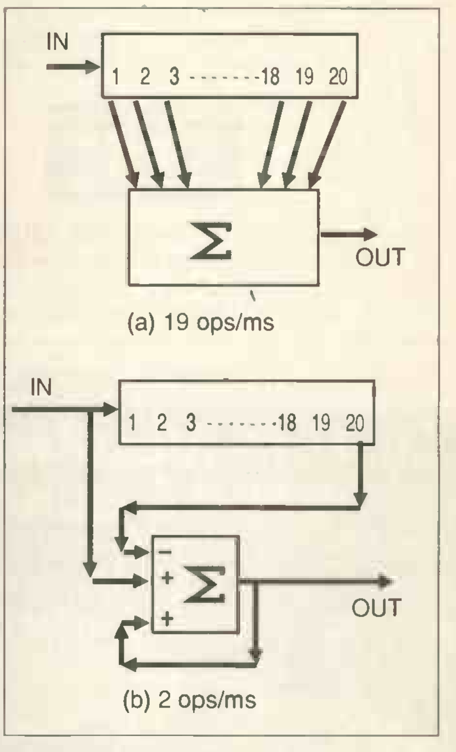
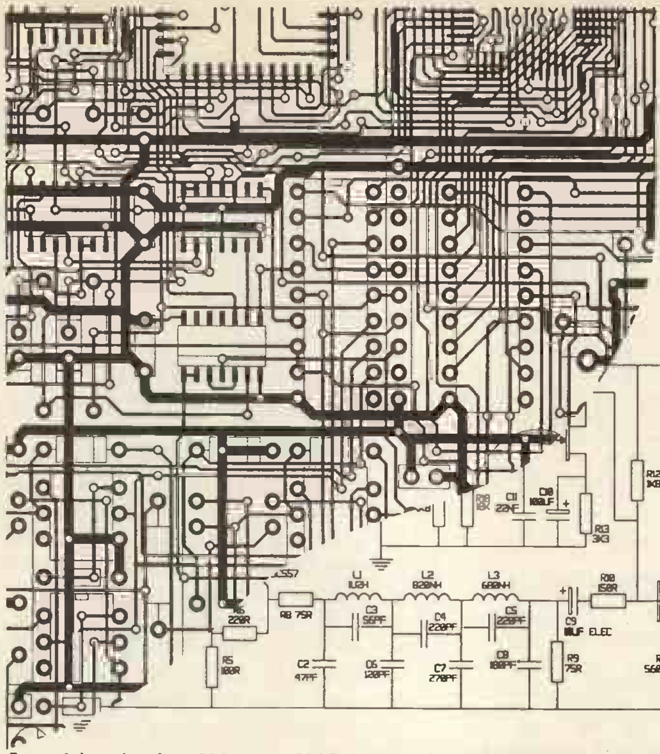



*Fig. 6a and b. Filter implementations. This is a continuous 20ms filter that optimally drags 50baud data from the noise. In fact, at this point the GPS signal is well above the noise, as it has been filtered to a few hundred hertz bandwidth. The two implementations shown produce the same output, but one is ten times faster. Note it must be cleared at startup.*

## Data Extraction

To extract the data and the 20ms timing edges needed for position resolution, the structure of **Fig. 7** is used. The filtered data, if noise free, would be reduced from square waves to ramps, but achieving the same peak amplitude, while any noise at a frequency higher than 25Hz is severely attenuated. In fact 50Hz noise conveniently hits a deep notch. The data amplitude is sampled at three points, nominally early, punctual and late. The sign of the punctual signal is then the data bit, 0 or 1, while the normalised ratio of the rectified early and late signals $(E-L)/(E+L)$, as in the code tracker, is used to advance or retard the sampling point. This is done directly on the phase, however, as the data edges are synchronous with the code epochs, so are not continuous, but must fall on a one millisecond clock edge in satellite time. Thus the tracker has simply to find the correct one of 20 possible phases.

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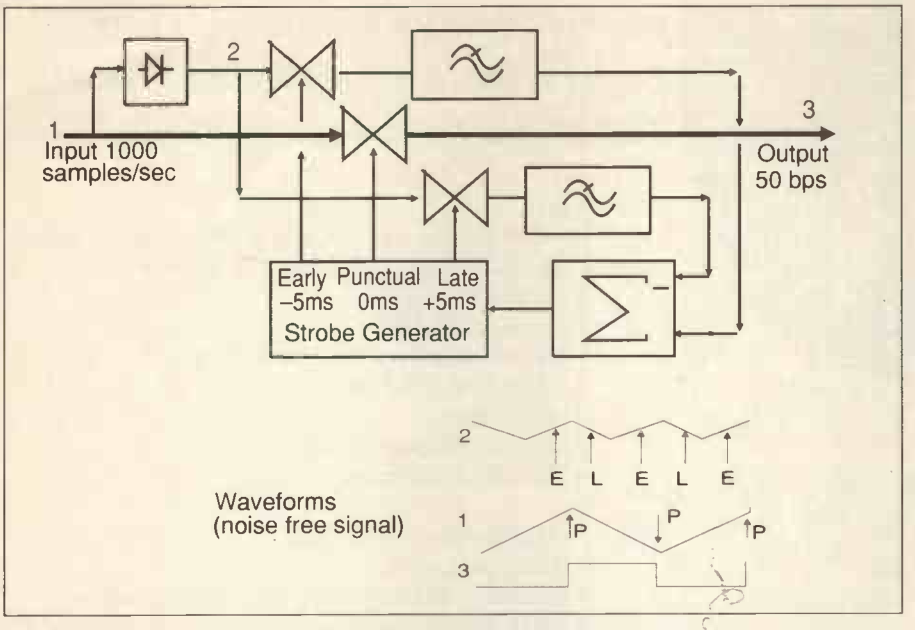



*Fig. 7. Data-bit Extractor. Data is extracted by detecting the peaks and running a narrow bandwidth PLL to maintain correct sampling. The rectifier removes the data from the tracking channel. Note this is dependent on the integration ramp from the 20ms filter. If the filter were omitted, this would need a differentiator to find edges, which gives very poor noise performance.*

The data bits are then scanned for the preamble, a defined pattern that identifies the start of a packet, or subframe in GPS terminology. While acquiring, this must be done on a per bit (20ms) basis, but once found it is automatically maintained by counting, so need only be checked for correctness on a 0.6s basis. The preamble defines both word and subframe sync, so subframe of 10 30-bit words is assembled, and passed to a background task for parity/checksum tests and storage. This is best done on a subframe basis, rather than per word, because the checksum overlaps between words.

## Acquisition mode

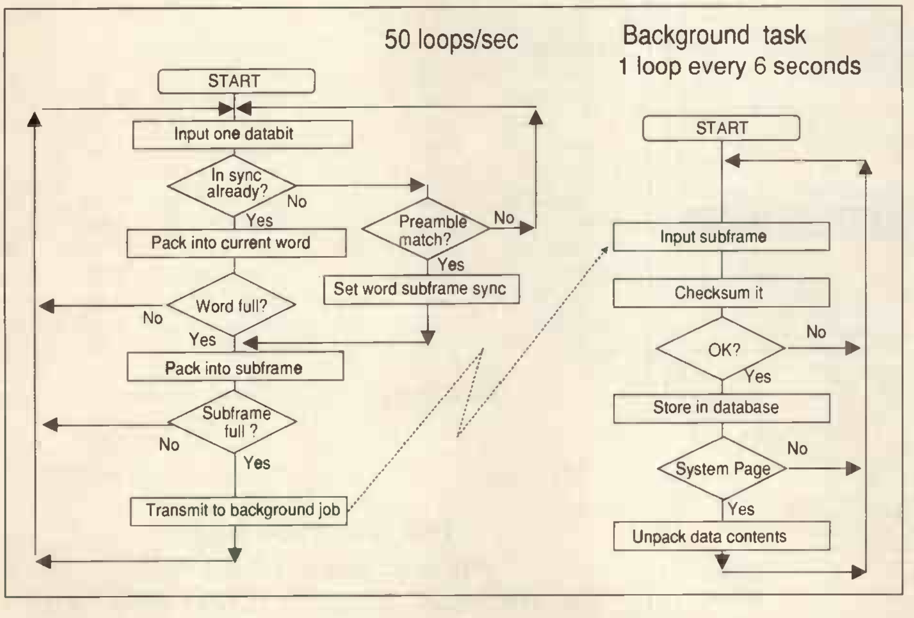
Acquisition is much harder than tracking, because instead of one code phase to compute for, we must do 2046. Instead of one 250Hz channel, we must cover about a hundred. On the other hand, it does not need to be done in real time, because in a software solution, off-air samples can be stored in memory for repeated processing off line.

The code-phase search is unavoidable, though on high sample rate machines (we have done one at 10MHz) the phase step may be more than one sample during acquisition. The same structure is used as the code-tracker described above, but the loop is opened and swept through all the possible offsets.

The problem is also minimised by reducing the bandwidth in steps, not directly to 250Hz. First the output of the correlator is processed with an FFT. This covers 16kHz, and yields output bins of 1kHz bandwidth. This is done for all 2046 correlation phases, on one millisecond of data. If the magnitude of any bin exceeds a threshold, it is mixed down to baseband using 4ms of data. While this is done, a list of the rolling best 10 phases is stored. Finally, the best ten are processed for a further 4ms, to both confirm the amplitudes, and establish a range-rate that can set the tracker off on the right path. The best amplitude of the ten is then run live off-air.

The fact that the transputer can process the spread-spectrum for one satellite at about seven times real time rate means this acquisition process is very rapid. In fact the limiting process is the one second delay inserted to allow accurate range-rate estimation. As a result, the early implementations did not use the subtle trick of calculating the expected frequencies for the 2nd and subsequent satellites, after calibrating from the first, but this has now been added to increase the sensitivity still further.

*Fig. 8. Data operations. Data operations are shown as a flow chart, as they are not continuous. In the foreground, bits are packed into words, and words into subframes. In the background, the subframes are checked for errors, and either archived or decoded.*




## Critical signal processing

Having understood the function of the software at the system level, it may be interesting to see the detail of the most CPU-intensive task, the de-spreading or correlation process. No CPU could be expected to perform an expression evaluation inside a loop, and repeat the loop at a 0.5μs rate. But because of a unique match between the transputer instruction set and this task, it can do it some seven times faster than this!

The critical feature is coding the radio signal at one bit per sample. This obviates the need for AGC in the radio, but more importantly allows the samples to be packed 32 to a word inside the computer.

The most demanding task is the front end of the software, where the two mixing stages and the filter are implemented. Mixing is multiplication. If you cannot accept that, take two sine waves, $\sin(\omega_1 t) = \sin(A)$, $\sin(\omega_2 t) = \sin(B)$

Traditional expansions give

$\sin A \sin B = (\cos(A-B)-\cos(A+B))/2$

The change to cosines is simply a ninety degree phase shift, and the two terms are now the sum and difference frequencies, as expected from a hardware mixer.

In the one bit per sample world, multiplication is an exclusive-OR or differ operation, ie

$0 \times 0 = 0$
$0 \times 1 = 1$
$1 \times 0 = 1$
$1 \times 1 = 0$

This upsets the human brain until it is realised that the binary 0 does not represent a null value, but represents –1, which matches our perception:

$-1 \times -1 = 1$
$-1 \times 1 = -1$
$1 \times -1 = -1$
$1 \times 1 = 1$

The apparent inversion of the result using exclusive OR is not a problem: its only the same as using an amplifier that inverts.

Thus in the transputer we can multiply two

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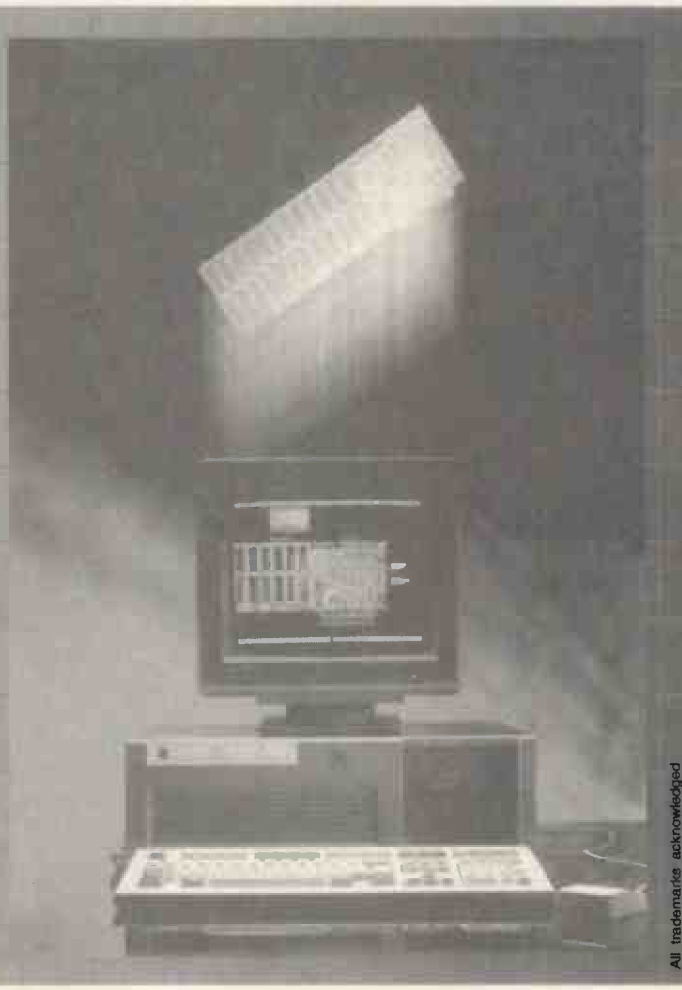

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*CIRCLE NO. 122 ON REPLY CARD*

32-bit variables using the XOR instruction, which will take two machine cycles, 100ns at 20MHz, the slowest available transputer speed, or 66ns at 30MHz. The critical part is that we have multiplied 32 samples in one instruction, or an equivalent time per sample of 3ns or 2ns respectively.

Fine. This is excellent, but many machines have XOR instructions, and at least one other has the serial link I/O facility that allows the data to be loaded this fast in the first place. The final crunch is the filtering operation.

The traditional DSP enthusiast would be looking for a MAC, or multiply and accumulate instruction to do both jobs, but the transputer is a very fast general purpose processor, not a DSP.

It does however have the BITCOUNT instruction. This counts the number of bits set in a word, with a facility to add-in a previous result, making it cascadable over any number of words.

This instruction takes a worst case of 34 cycles on a 32 bit machine, 18 on a 16-bit transputer. Thus the total processing time for 32 samples is some 36 cycles, less than 2μs, for 16μs of signal samples.

However we have left out the overheads. The operands for these instructions have to be loaded, and the results stored, and the whole sequence run in a loop. Loading takes four cycles per operand, and storing is avoided by keeping the rolling result in a register, as the transputer's register stack is deep enough. The loop overheads become negligible by putting 16 copies of the code in line before looping. Thus the resultant time is 44 cycles per 32 samples, 2.2μs on a 20MHz machine, 1.5μs on a 30MHz CPU.

The *Occam* code for the high-speed processing is shown in **Fig. 9**, and the equivalent assembler in **Fig. 10**. Note that with help from the programmer, ie the way in which the expression is nested with no intermediate storage, the *Occam* is just as efficient as the assembler.

*Fig. 9. Occam code to implement downconvert and correlate. The Occam to correlate one millisecond of signal, from 1.5MHz carrier, 2MHz bandwidth to 4kHz carrier, nominally 100Hz bandwidth in 1kHz noise bandwidth, is very simple and requires only six lines (a). However for maximum efficiency, equal to assembler, version (b) is used, with loops opened out, constant indices, and no storing of intermediate results.*

**(a) Simple Occam**

```
SEQ point= 0 FOR 16
  SEQ
    accumulate := 0
    SEQ input = 0 FOR 4
    accumulate :=
    BITCOUNT(LOcode[point*4+input]>< signal[point*4+input],accumulate)
    result[output] := accumulate –64
```

**(b) Efficient Occam**

```
VAL point IS 0:
VAL LO IS [LOcode FROM point*4 FOR 4]:
VAL sig IS [signal FROM point*4 FOR 4]:
result[point]:=
  BITCOUNT(LO[3]><sig[3],BITCOUNT(LO[2]><sig[2],
    BITCOUNT(LO[1]><sig[1],BITCOUNT(LO[0]><sig[0],-64))))

All repeated 16 times in line, through 16 values of point
```

*Fig. 10. Assembler code to implement downconvert and correlate. The instructions generated for the inner loop take less than a tenth of the CPU time per satellite... or acquisition can be done ten times faster than real time, using a single set of radio data*

| Instruction | | Machine cycles (33ns on 30MHz CPU) |
|---|---|---|
| LDL | LOptr | 2 |
| LDNL | Index | 2 |
| LDL | Sigptr | 2 |
| LDNL | Index | 2 |
| XOR | | 2 (inc one prefix) |
| BITCOUNT | | 34 (worst case) |
| | Total | 44 cycles, 1.5 microsecs |

Repeat 64 times in line, indices 0-3, adjusting pointers every 4
Eight extra cycles to adjust pointers (0.25μs) or (16 x 0.25) + (64 x 1.5) μs which equals 100μs of CPU time to process 1ms of real time

When acquiring timing lock with the satellites, the carrier phase is not important, so the code above premixes the carrier and synthetic code in advance. When running live, this is not so easy, as each adjustment to the code-phase distorts the carrier phase. Also, the work of the first down conversion can be shared over several satellites.

Thus in the 2046 samples per millisecond models, the downconversion is done separately.

In the 2000 samples/ms models, where carrier-phase compensation is much easier, as each millisecond is the same number of bytes (250), rather than (256,256,256,255), the combined-then-compensate method is used.

The output of the mix/filter operation is a block of 16 samples every millisecond, each representing 64μs. Each now has a value between –64 and +63, ie, seven significant bits, but they occupy a full machine word, 16 or 32 bits depending on processor. This is important, as sign-extension is very slow. It comes for free here by pre-loading the BITCOUNT operations with –64.

In acquisition, the work done next on these 16 samples is an FFT. In tracking, it is further downconversion in I+Q, some 32 multiplies, around 20μs. However as this now occurs only once per millisecond, it is no longer a problem.

## One or two transputers

The GPS receiver based on the software described here has been running since 1989 in some guise, with first positions in mid 1990, and the software has been essentially unchanged since February 1991. The major effort put into it since then has been re-hosting it onto different CPU boards, different I/O systems for clients, and of course the occasional bug-fix. My serious effort in the meanwhile has gone into the radio design, covered in an earlier article, and into studies as to how it can be improved for the next serious client.

The major deficiency with the current single transputer system is that while it can code track four satellites simultaneously for positioning, it can only carrier track on one satellite at a time (20MHz CPU). A 30MHz CPU could handle two, but the software changes would be so major that it was not worthwhile. This means that data downloading is sequential at cold-start, and on a change of satellite, positioning is interrupted.

These problems are very unfortunate considering that the major benefit of the software approach is the acquisition and re-acquisition time.

The solution is to provide two transputers, with the benefits of full parallel downloading and positioning, and plenty of spare CPU for map handling, as described in the third article in this series. This has now been built and is running in the office. It is transportable, but not portable. It went to the *Wireless World* studios for photography in October 1992, and the maps displayed in the screen-shots were real, not superimposed.

## Applications

The pros and cons of raster and vector maps, colours, pan and zoom will all be covered in the next article, on applications of GPS. The technology is now at the stage where it is a small black box costing a few hundred pounds/dollars that is built in to a larger system, be it a jumbo jet or a combine-harvester

***Philip Mattos is a consultant engineer with Inmos, SGS-Thomson.***

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**Microwave Systems** MOS/3600 Microwave frequency stabilizer – **1GHz** to 40GHz **£1k**.
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**Marconi TF2370 spectrum ANZ** + TK2373 FX extender 1250Mc/s + trk gen – £2.5k–£3k.
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*CIRCLE NO. 123 ON REPLY CARD*

# Analysing performance of current mode op amps

*Current mode amplifiers have much in common with conventional op amps. Potentially they offer much more performance in many applications. Terrence Finnegan lays down the design rules to get the best out of your application.*

Current mode amplifiers are now over ten years old. They originated in their present form from the Comlinear Corporation of Loveland, Colorado, who brought out a hybrid in 1980 and who I think hold the basic patents covering the concept.

But the early devices did not catch on because they were expensive, around £150 each, and Comlinear probably tried to push the technology too far. The *CLC220* for instance had a bandwidth of over 200MHz. But the concept did take hold and Elantec brought out the first monolithic device, the *EL2020*, in the mid 1980s eventually followed by devices from PMI, Analog Devices and others.

Comlinear's approach to design focused on the traditional op-amp's variation of bandwidth with gain.

Conventional op-amps (**Fig. 1**) have a differential, high impedance input stage which feeds several following gain stages. Open loop output of the amplifier to a first order is $V_O=A(s)[V_1-V_2]$, where $A(s)$ is a complex gain function.

With the feedback completed, a sample of the output voltage is applied to the inverting input and the closed loop gain becomes:

$$\frac{V_o}{V_i}=\frac{G}{1+\dfrac{G}{A(s)}}$$

where $$G=\frac{R_1+R_2}{R_1}$$

To see the effect the gain setting $G$ has on the frequency response, $A(s)$ is separated into a numerator $N(s)$ containing the zeros of the frequency response and a denominator $D(s)$ containing the poles of the response. Substituting this into the expression for the closed loop gain and re-arranging gives:

$$\frac{V_o}{V_1}=G\frac{N(s)}{N(s)+G.D(s)}$$

$G$ not only scales the gain magnitude as expected, but it also multiplies the effect of $D(s)$ on the closed loop response. Locations of the closed loop poles are now functions of $G$. This is the chief design failure of traditional voltage mode op amps, mathematically stated, leading to difficulty in compensation and gain-bandwidth product specmanship among manufacturers. The several gain stages also cause propagation delays and slew rate problems (see "*Current alternative to operational amplifiers*" by Frank Ogden, *EW + WW*, August, pp.643-644)).

Comlinear reasoned that if they could develop an amplifier structure which would remove $G$ from the denominator of the closed loop gain expression, then performance would improve dramatically. They succeeded in this aim – and the current mode amplifier was born.

In one of its application notes[1] the company describes operation of the non-inverting current mode op amp.

In **Fig. 2**, the unity gain input buffer forces $V_2$ to equal $V_1$. Current $I_{inv}$ flowing into or out of the input terminal is amplified by a transimpedance amplifier, generating the output voltage. The complex transfer function of the transimpedance amplifier is $A(s)$ ohms and $V_0=I_{inv}A(s)$. Operation is described by:

$$I_{inv}=I_1-I_2$$

$$=\frac{V_2}{R_1}-\frac{V_o-V_2}{R_2}$$

But $V_0=I_{inv}.A(s)$ and $V_2=V_1$, forced by the buffer. So

$$\frac{V_o}{A(s)}=V_1\left(\frac{1}{R_1}+\frac{1}{R_2}\right)-\frac{V_o}{R_2}$$

which after re-arranging and again letting $G=(R_1+R_2)/R_1$ becomes:

$$\frac{V_o}{V_1}=\frac{G}{1+\dfrac{R_2}{A(s)}}$$

Again setting $A(s)=N(s)/D(s)$, we have:

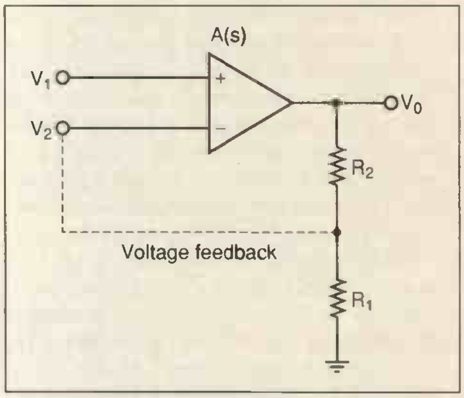



*Fig. 1. Standard non-inverting voltage mode op amp, with the location of the closed loop poles being a function of G.*

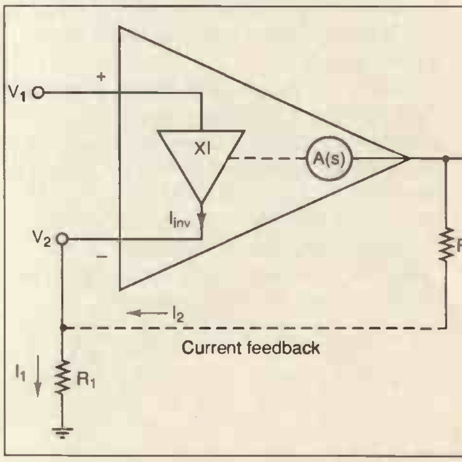



*Fig. 2. Non-inverting current mode op amp. Pole locations – hence performance – can be held constant.*

$$\frac{V_o}{V_1}=G.\frac{N(s)}{N(s)+R_2.D(s)}$$

The closed loop gain equations for the two operating modes are now in the same form and we see that $R_2$ has replaced $G$ in the fre-

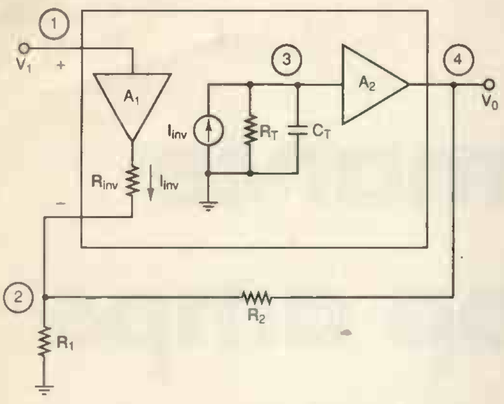



*Fig. 3. Expanded non-inverting current mode circuit.*

quency dependent term of the transfer function. Since $R_2$ can be held constant, whereas $G$ cannot, the pole locations and hence the performance can be held constant.

Gain setting is varied by changing $R_1$, while the bandwidth is adjusted through $R_2$. Most manufacturers design their amplifiers to work with particular values for $R_2$, optimised for the internal capacitances present and for various load conditions. Sometimes $R_2$ is already built into the device to optimise this particular variable and to minimise stray capacitances – particularly across $R_2$, which upset the stability.

Current mode or transimpedance amplifiers, as a rule, are not noted for high gain accuracy, either AC or DC and a more complete analysis will define the AC and DC performance, starting from the more complex block diagram of **Fig. 3**. $R_{inv}$ is the output resistance of the unity gain buffer $A_1$ driving the inverting input terminal. A mirror of the current flowing in this buffer also flows into the transimpedance $Z_T$, developing a voltage across it, which is applied to a second unity gain buffer $A_2$ driving the output.

Assume for the moment that $A_2$ is perfect. $Z_T$ is a complex impedance made up of $R_T$ in parallel with $C_T$ and is a simplified equivalent of one of the internal nodes. To get a feel for the numbers, $R_{inv}$ is 10-100Ω, $R_T$ is 1-15MΩ and $C_T$ is 4-10pf, depending on the device.

Transforming the impedances into admittances, the five nodal equations describing the relevant operation of this circuit are:

$$I_2 = (E_2 - E_1)Y_{inv} + E_2Y_1 + (E_2 - E_4)Y_2$$

$$I_{inv} = (E_1 - E_2)Y_{inv}$$

$$I_3 = E_3Y_T = I_{inv}$$

$$E_3 = E_4$$

$$I_2 = 0$$

Solving these nodal equations gives the expression for the closed loop voltage gain as:

$$G = \frac{E_4}{E_1} = \frac{Y_{inv}(Y_1 + Y_2)}{Y_T(Y_1 + Y_2 + Y_{inv}) + Y_2Y_{inv}}$$

## POLES AND ZEROS: A QUESTION OF TRANSFORMATION

The transfer function of any device is the ratio of the Laplace transform of the output of the device to the Laplace transform of the input signal causing that output, for all initial conditions zero. the Laplace integral which generates the transform is defined as:

$$L_p = \int_0^\infty e^{-st}\,\mathrm{d}t$$

where $s$ is a complex frequency variable of the form $s=\sigma+j\omega$. All possible values of $s$ are therefore defined as points on a plane called the $s$-plane, having axes $\sigma$ and $j\omega$. The input signal is not specified and for convenience, a signal having an easily transformed equation is usually chosen. The simplest signal to use is the unit impulse, which has a transform equal to one. This simplifies the definition of the transfer function to:

"The transfer function of any device is the Laplace transform of the output of the device in response to a unit impulse input".

From this information, the transient response of any circuit in response to any input stimulus can be derived.

When the input signal is a constant amplitude constant frequency sinusoid, as is usually the case for circuit analysis, the output signal at steady state will also be sinusoidal, but of differing amplitude and/or phase. In this case, the expression for the frequency transfer function is simplified and is obtained from the $s$ transfer function by setting $\sigma$=0 and using the substitution $s=j\omega$.

For linear RLC circuits, the frequency transfer function can also be derived using standard AC circuit theory. The familiar equation of a straight line is $y=mx+c$ where $x$ is the independent variable, $y$ the dependent variable, $m$ the slope and $c$ the intercept on the Y axis. The equation may be more conveniently recast for our use as $p=a_1s+a_2$

where $s$ is the complex frequency as before. When a transfer function is the quotient of linear terms, represented by this equation, it is called bilinear. So any bilinear transfer function of the form $T(s)=N(s)/D(s)$ may be represented as:

$$T(s) = \frac{a_1s + a_2}{b_1s + b_2}$$

where the coefficients $a$ and $b$ are real constants and may be positive or negative. If $T(s)$ is written in the form

$$T(s) = \frac{a_1}{b_1}\cdot\frac{s + \frac{a_2}{a_1}}{s + \frac{b_2}{b_1}} = k.\frac{s + z_1}{s + p_1}$$

then we say that $z_1$ is the zero of $T(s)$ and $p_1$ is the pole of $T(s)$. These quantities are located on the real axis of the $s$-plane, at $s=-z_1$ and $s=-p_1$, as shown in the figure. (Poles are conventionally shown by a cross, and zeros by a circle).

For first order transfer functions, $p_1$ is always on the negative real axis, while $z_1$ may be on either the positive or the negative part of the real axis. A pole in the transfer function will cause the output amplitude to fall with rising frequency at –20dB/decade, starting at the pole frequency, and the phase of the output will lag increasingly behind the input with rising frequency, the phase being -45° when $\omega=p$.

The reverse is true for a zero in the transfer function on the negative real axis. Gain will rise at +20dB/decade above the zero, while the phase will lead and will be +45° when $\omega=z$. The effects of a zero therefore tend to cancel out those of a pole and are often introduced in loop compensation to improve stability.

Obviously, a coincident pole and zero will cancel completely and have no net effect. A zero in the transfer function on the *positive* real axis however can be a problem. It will cause the output amplitude to increase with rising frequency, as before, but with a *lagging* phase. Since a corresponding pole cannot exist on the positive axis, there is no way of cancelling out the "right-half-plane zero" and instability can sometimes result, for instance in flyback, boost and Cuk switch-mode power supply topologies.

But the function

$$T(s) = \frac{s - \sigma_1}{s + \sigma_1}$$

is useful in the "all-pass" circuit, because the rising gain characteristic of the zero at $+\sigma_1$ exactly cancels the falling gain characteristic of the pole at $-\sigma_1$, while the two lagging phase characteristics add together giving an overall –180° phase lag when $\omega>>\alpha$. The circuit finds application as a phase compensator, or where several are used together, a differential output between two properly designed circuits can provide an exact 90° phase difference over a wide frequency spread – a useful analogue function.

More complex transfer functions can have multiple poles and/or zeros, which need not necessarily be located just on the real axes. For instance, second-order transfer functions with terms in $s^2$ have pole-pairs which are complex conjugates of each other, which can be located anywhere in the left-half of the $s$-plane.

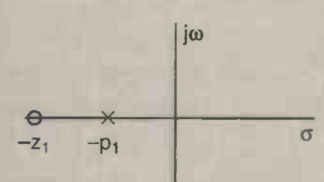
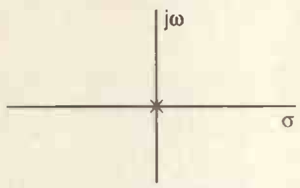
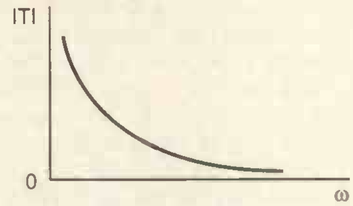
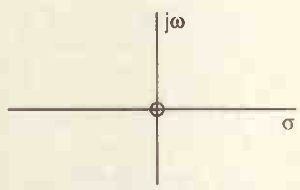
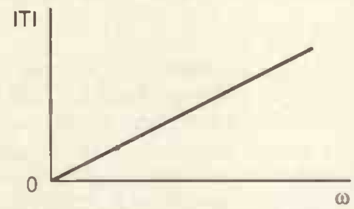
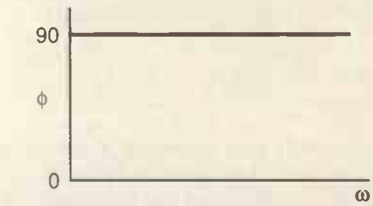
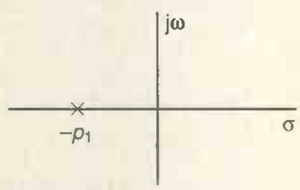
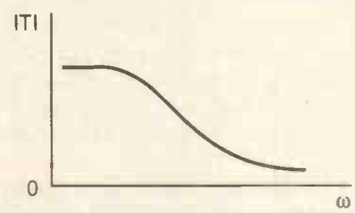
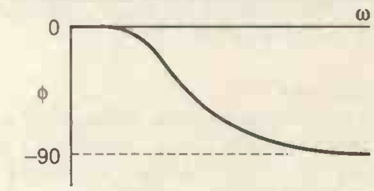
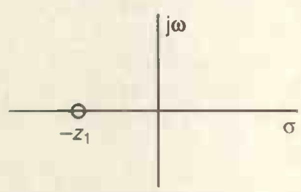
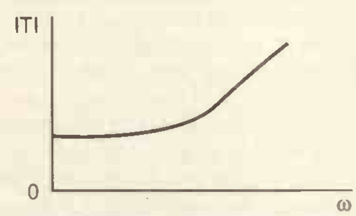
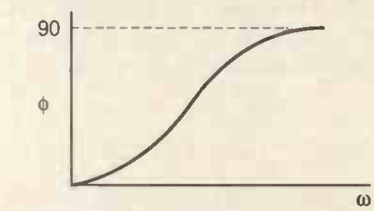
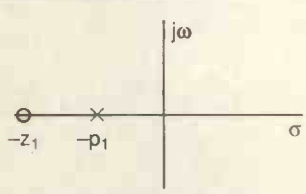
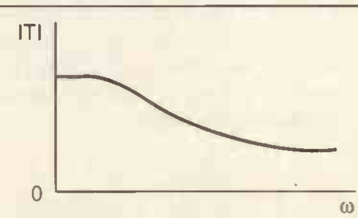
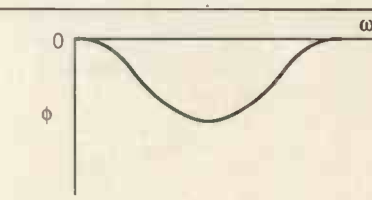
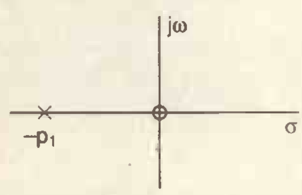
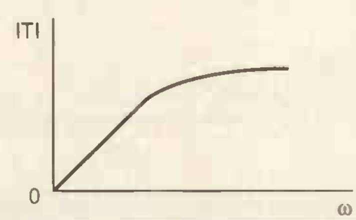
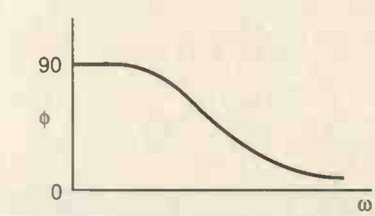
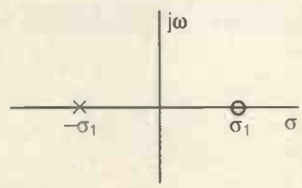
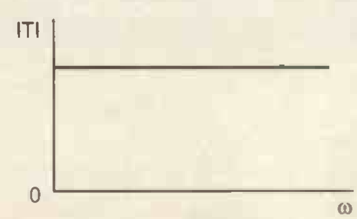
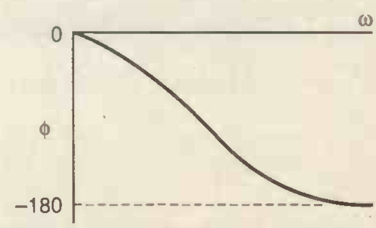



*Fig. 4. Pole and zero location on the s-plane*

*Some relationships for simple first-order transfer functions, between pole/zero locations on the s-plane and magnitude and phase responses.*

| $T_n(s)$ | Pole and zero | Magnitude response | Phase response |
|---|---|---|---|
| $\frac{K}{s}$ |  |  |  |
| $K.s$ |  |  |  |
| $\frac{K}{s+p_1}$ |  |  |  |
| $K(s+z_1)$ |  |  |  |
| $K\frac{s+z_1}{s+p_1}$ $z_1 > p_1$ |  |  |  |
| $K\frac{s}{s+p_1}$ |  |  |  |
| $K\frac{s-\sigma_1}{s+\sigma_1}$ |  |  |  |

### AC behaviour

The important parameters defining AC behaviour are transcapacitance $C_T$ and feedback resistor $R_2$. The time constant formed by these components is analogous to the dominant pole of a conventional op amp and cannot be reduced below a critical value if the closed loop is to remain stable.

Transforming the admittances back into resistances in the closed loop gain equation, substituting $Y_T=(1+sC_TR_T)/R_T$ and simplifying yields:

$$G=\frac{R_T(R_1+R_2)}{(1+sC_TR_T)(R_1R_{inv}+R_2R_{inv}+R_1R_2)+R_1R_T}$$

Now $R_1R_T$ is larger than all the other resistive products and $R_1R_2>R_{inv}(R_1+R_2)$. Hence:

$$G=\frac{R_1+R_2}{R_1(1+sC_TR_2)}$$

showing that the time constant $C_TR_2$ forms the dominant pole and thus defines the bandwidth. There are also subsidiary poles formed by $R_1$ and $R_{inv}$ which modify the shape of the frequency response curve.

### DC behaviour

The gain at DC can easily be computed from the gain equation and simplifies to $G=(R_1+R_2)/R_1$ as expected. But, it is instructive to compare the equivalent open loop gains of voltage mode and current mode op amps, to assess their comparative performance. The gain of a standard voltage mode op amp in the same circuit configuration is given by:

$$G=\frac{V_o}{V_i}=\frac{A_0(Y_1+Y_2)}{Y_1+Y_2(1+A_0)}$$

where $A_0$ is the op amp gain.

Designate $A_{eq}$ as the gain of the equivalent voltage mode op amp which would have the same circuit performance as a current mode op amp, and we can say that:

$$\frac{A_{eq}(Y_1+Y_2)}{Y_1+Y_2(1+A_{eq})}\equiv\frac{Y_{inv}(Y_1+Y_2)}{Y_T(Y_1+Y_2+Y_{inv})+Y_2Y_{inv}}$$

From which

$$A_{eq}=\frac{Y_{inv}(Y_1+Y_2)}{Y_T(Y_1+Y_2+Y_{inv})}$$

$$=\frac{R_T(R_1+R_2)}{R_1R_2+R_{inv}(R_1+R_2)}$$

$$A_{eq}\approx\frac{R_T(R_1+R_2)}{R_1R_2}=\frac{G.R_T}{R_2}$$

$$\approx\frac{R_T}{R_1}\quad \text{when } R_2>R_1$$

Some manufacturers quote this ratio as a "goodness factor" for their op amps. While this may be true at the device level, the in-circuit loop gain is clearly the factor of interest and the above analysis shows that the equivalent open loop voltage of a current mode op amp depends on $R_1$. Also, ratio of the open loop gain to the closed loop gain equals $R_T/R_2$– effectively fixed for all closed loop gains as $R_2$ is either specified by the manufacturer or built into the device.

### Buffer $A_2$

In practice, buffer $A_2$ is a compound emitter-follower, having a finite gain and therefore a finite input impedance directly proportional to the load. Input impedance appears across $R_T$ and so the in-circuit value of $R_T$ and hence the effective loop gain $A_{eq}$ is load dependent. Bandwidth will also be affected by any load capacitance, reflected back across $C_T$. Performance is usually quoted for a given load, often 400Ω, and the actual performance with the user load can be quite different. Do not forget that $R_2$ also forms a part of the load.

### Comparison between operating modes

An interesting comparison can be drawn from this analyses about the way the two parameters of gain and bandwidth transform between the two operating modes. Voltage mode devices have a *fixed* loop gain and a *variable* closed loop bandwidth. In contrast, current mode devices have a *variable* loop gain and a *fixed* closed loop bandwidth. **Table 1** summarises the performance parameters.

We can also indicate the magnitude of the gain-bandwidth product for a current mode device as $GBW=R_T/2\pi C_TR_1R_2$, although whether this has any practical significance is a matter of conjecture.

### Reference

1. Application Note 300-1, "A New Approach to Op Amp Design", Comlinear Corporation, March 1985

## BOOK REVIEW

**Fibre Optic Cabling**, by Mike Gilmore, extends the understanding of optical fibres needed for practical application in telecommunications and data-comms to a more soundly based level, at which theory can confidently be applied to unfamiliar techniques. Nonetheless, this is essentially a practical treatment for engineers and laymen alike, which also covers the installation and commercial aspects of the technology.

Optical fibres in general terms and their use in communications form an initial chapter, as an introduction to more specific discussion. Five chapters cover fibres and connection practice, both theoretical and practical considerations being treated, and are followed by three more on cables, highways in general and highway design in particular. A final chapter on the hardware of optical fibres then presents the choices available in cables and assemblies, connectors, splicing and enclosures. Specifying fibre systems appears to be an undeveloped art, its vagueness contrasting sharply with that commonly found in copper cabled systems; one chapter is therefore an attempt to introduce a little rigour into the process.

Acceptance testing, installation practice and final acceptance testing are all subject to contractual obligations and are therefore supremely important if reputations are to survive a contract; three chapters describe methods of ensuring survival of both reputations and systems.

Documentation and maintenance form the two final chapters in the design and installation part of the book, but the penultimate section is illustrative of the author's experience with real systems — a case study. Future developments such as single-mode fibres and fixed, blown-fibre cables are then discussed in an end-piece. This is no "penny-a-line" text cobbled together by a media man, but is the result of hard-won experience by a practitioner, the author not only being Managing Director of an optical-fibre cabling company, but also chairman of one of the BSI working groups in this area.

*Butterworth Heinemann, 318 pp, hardback, £35.*

**Table 1. Performance parameters**

| Operating mode | Voltage mode | Current mode |
|---|---|---|
| Open loop gain | $A_0$, fixed by device manufacture | $A_{eq}\approx G.R_T/R_2$ varies with the closed loop gain |
| Closed loop bandwidth | $f_{-3dB}\approx GBW/G$, varies with the closed loop gain | $f_{-3dB}\approx 1/2\pi C_TR_2$, fixed by device manufacture |

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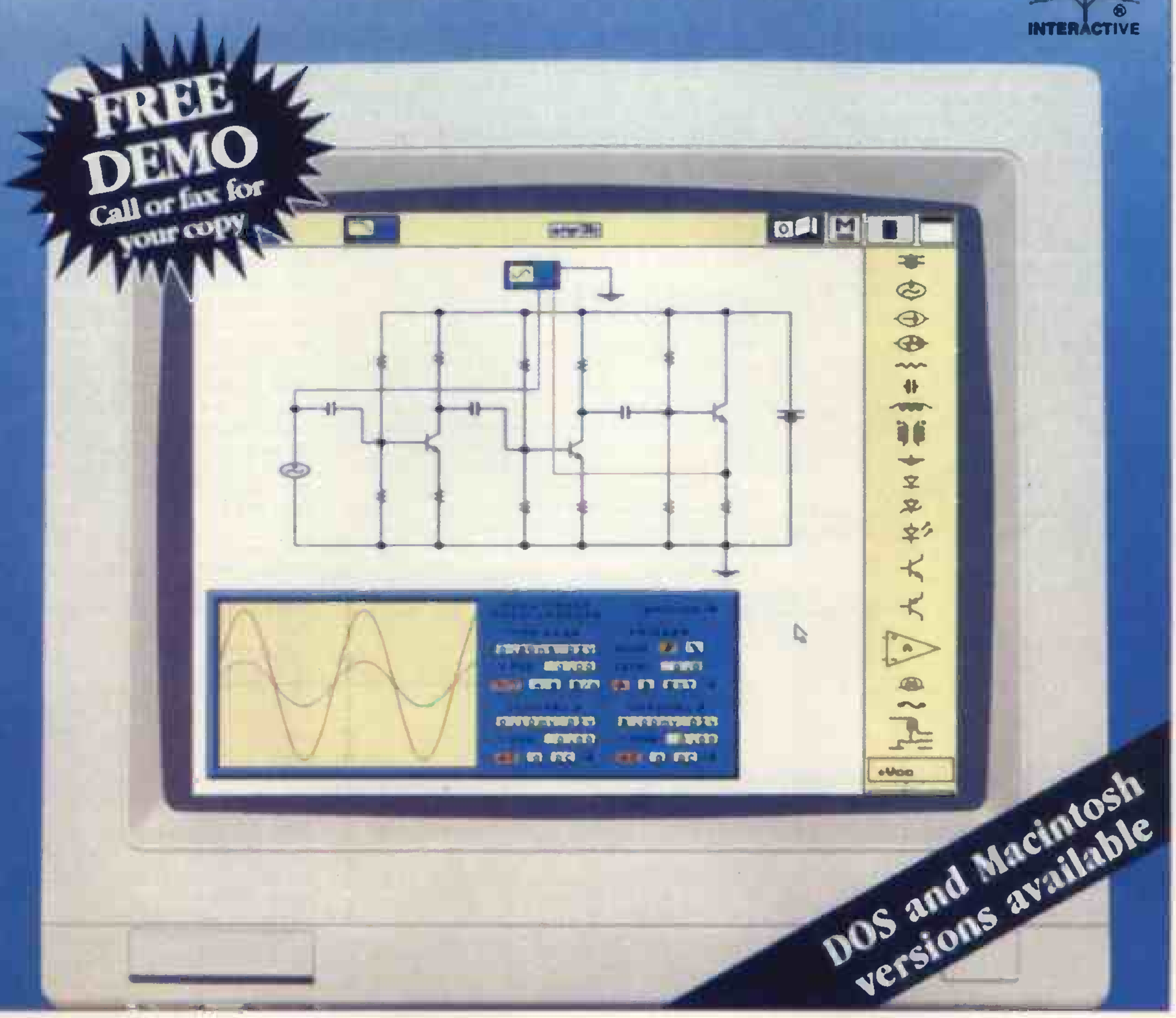


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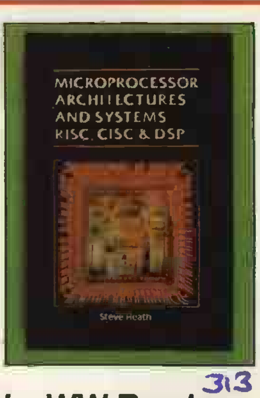


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DESIGN BRIEF

# LOG AMPS FOR RADAR-AND MUCH MORE

***Log amps are useful for the problems associated with identifying radar pulses. But there are many other applications too. Ian Hickman explores their versatility.***

The free-space inverse square law applies to propagation in both the outgoing and return signal paths for RF radiation pulses. Returned signal power from a given sized target is therefore inversely proportional to the fourth power of distance – the well known basic $R^4$ radar range law.

With the consequent huge variations in the size of target returns with range, a fixed gain IF amplifier would be useless. The return from a target at short range would overload it, whilst at long range the signal would be too small to operate the detector. One alternative is a swept gain IF amplifier, where the gain is at minimum immediately after the transmitted pulse and increases progressively with elapsed time thereafter.

Because the scheme has its own difficulties and is not always convenient, a popular arrangement is the logarithmic amplifier. With this, if a target flies towards the radar, instead of the return signal rising by 12dB (amplitude increasing by a factor of x4) every time the range halves, it increases only by a fixed increment, determined by the scaling of the amplifier's log law.

This requires a certain amount of circuit ingenuity: the basic arrangement is an amplifier with a modest, fixed amount of gain, and the ability to accept an input as large as its output when overdriven. **Figure 1** explains the principle of operation of such a "true log amplifier" stage, such as the GEC Plessey Semiconductors *SL531*. An IF strip consisting of a cascade of such stages provides maximum gain when none of the stages is limiting. As the input increases, more and more stages go into limiting, starting with the last stage, until the gain of the whole strip falls to x1 (0dB). If the output of each stage is fitted with a diode detector, the sum of the detected output voltages will increase approximately as the logarithm of the strip's input signal. Thus a dynamic range of many tens of dB can be compressed to a manageable range of as many equal voltage increments.

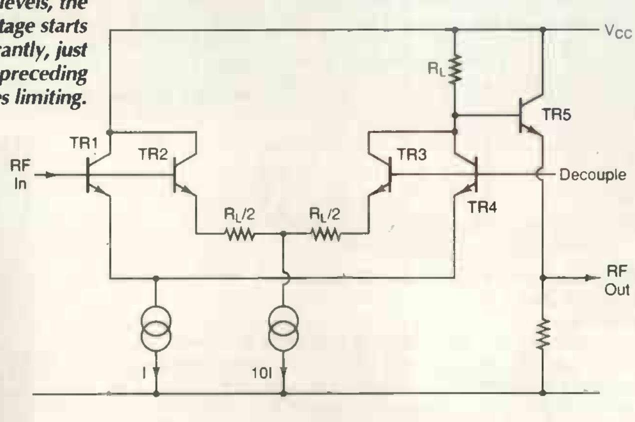



*Fig. 1. True log amplifier. At low signal levels, considerable gain is provided by TR1 and TR4, which have no emitter degeneration (gain setting) resistors. At higher levels, these transistors limit, but the input is now large enough to cause a significant contribution from TR2 and TR3, which operate at unity gain. At even larger signal levels, these also limit, so the gain falls still further. At very low input signal levels, the output from the stage starts to rise significantly, just before a similar preceding stage reaches limiting.*

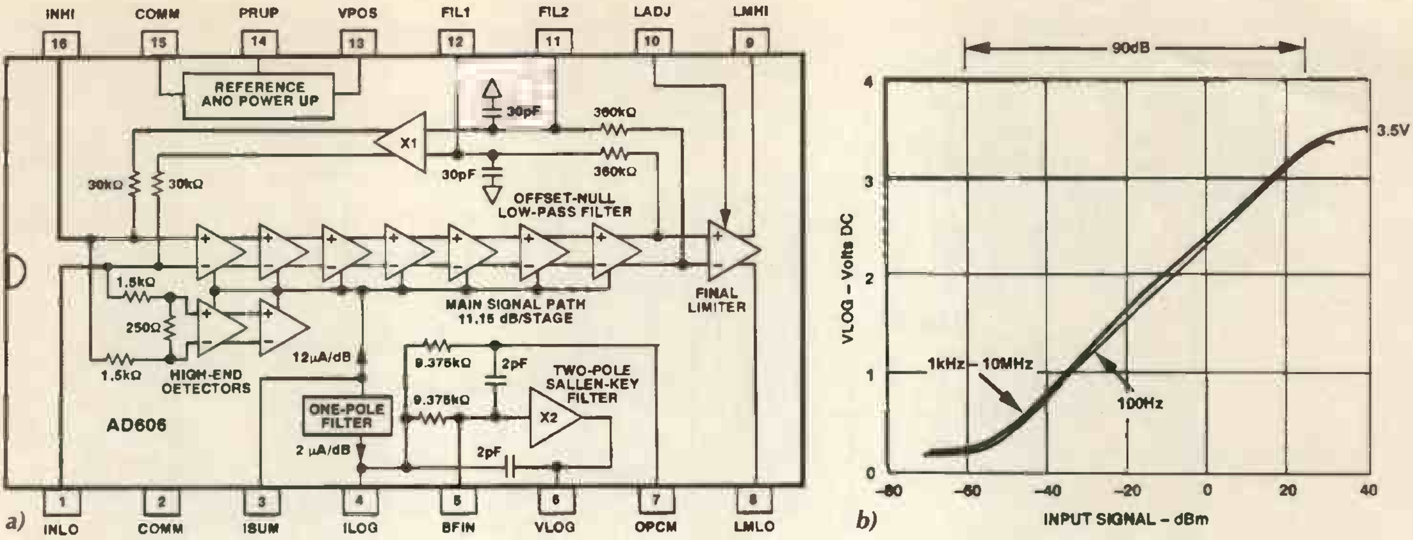
A strip of true log amps provides, at the output of the last stage, an IF signal output which is hard limited for all but the very smallest inputs. It thus acts like the IF strip in an FM receiver, and any phase information carried by the returns can be extracted. The "amplitude" of the return is indicated by the detected log (video) output; if it is well above the surrounding voltage level due to clutter, the target can be detected with high probability of detection and low probability of false alarms.

Many (in fact most) log amps have a built-in detector: if the log amp integrates several stages, the detected outputs are combined into a single video output. If target detection is the only required function, then the limited IF output from the back end of the strip is in fact superfluous, although many log amps make it available anyway for use if required.

The GEC Plessey Semiconductors *SL521* and *SL523* are single and two stage log amps with bandwidths of 140MHz and 100MHz respectively; the two detected outputs in the *SL523* are combined internally into a single video output. These devices may be simply cascaded: RF output of one to the RF input of the next, to provide log ranges of 80dB or more. The later *SL3522*, designed for use in the 100 – 600MHz range, is a successive detection 500MHz 75dB log range device in a 28 pin package, integrating seven stages and providing an on-chip video amplifier with facilities for adjustment of gain and offset (i.e. slope and intercept, discussed below) as well as a limited IF output.

The design of many log amps, such as those just mentioned[1], includes internal on-chip decoupling capacitors which limit the lower frequency of operation to around 5MHz. These are not accessible at package pins and so it is not possible to extend the operating range down to lower frequencies by strapping in additional off-chip capacitors. This limitation does not apply to the recently released Analog Devices *AD606*, which is a nine stage 80dB range successive detection log amp with final stage providing a limited IF output. It is usable to beyond 50MHz and operates over an input range of –75dBm to +5dBm.

The block diagram is shown in **Fig. 2a**, which indicates the seven cascaded amplifier/video detector stages in the main signal path preceding the final limiter stage, and a further two amplifier/video detector "lift" stages




(high-end detectors) in a side-chain fed via a 22dB attenuator. This extends the operational input range above the level at which the main IF cascade is limiting solidly in all stages. Pins 3 and 4 are normally left open circuit, whilst OPCM (output common, pin 7) should be connected to ground.

The 2µA per dB out of the one pole filter, flowing into the 9.375kΩ resistor between pins 4 and 7 (ground) defines a log slope law of 18.75mV/dB at the input to the x2 buffer amplifier input (pin 5) and hence of 37.5mV/dB (typ at 10.7MHz) at the video output VLOG pin 6. The absence of any dependence on internal coupling or decoupling capacitors in the main signal path means that the device operates in principle down to dc, and in practice down to 100Hz or less, Figure 2b.

In radar applications, the log law (slope) and intercept (output voltage with zero IF input signal level) are important. These may be adjusted by injecting currents derived from VLOG and from a fixed reference voltage respectively, as described later, into pin 5. A limited version of the IF signal may be taken from LMLO and/or or LMHI (pins 8 and 9, if they are connected to the +5V supply rail via 200Ω resistors) – useful in applications where information can be obtained from the phase of the

*Fig. 2a (left). Block diagram of the Analog Devices AD606 50MHz, 80dB demodulating logarithmic amplifier with limiter output; 2b (above) shows that the device operates at frequencies down to the audio range.*

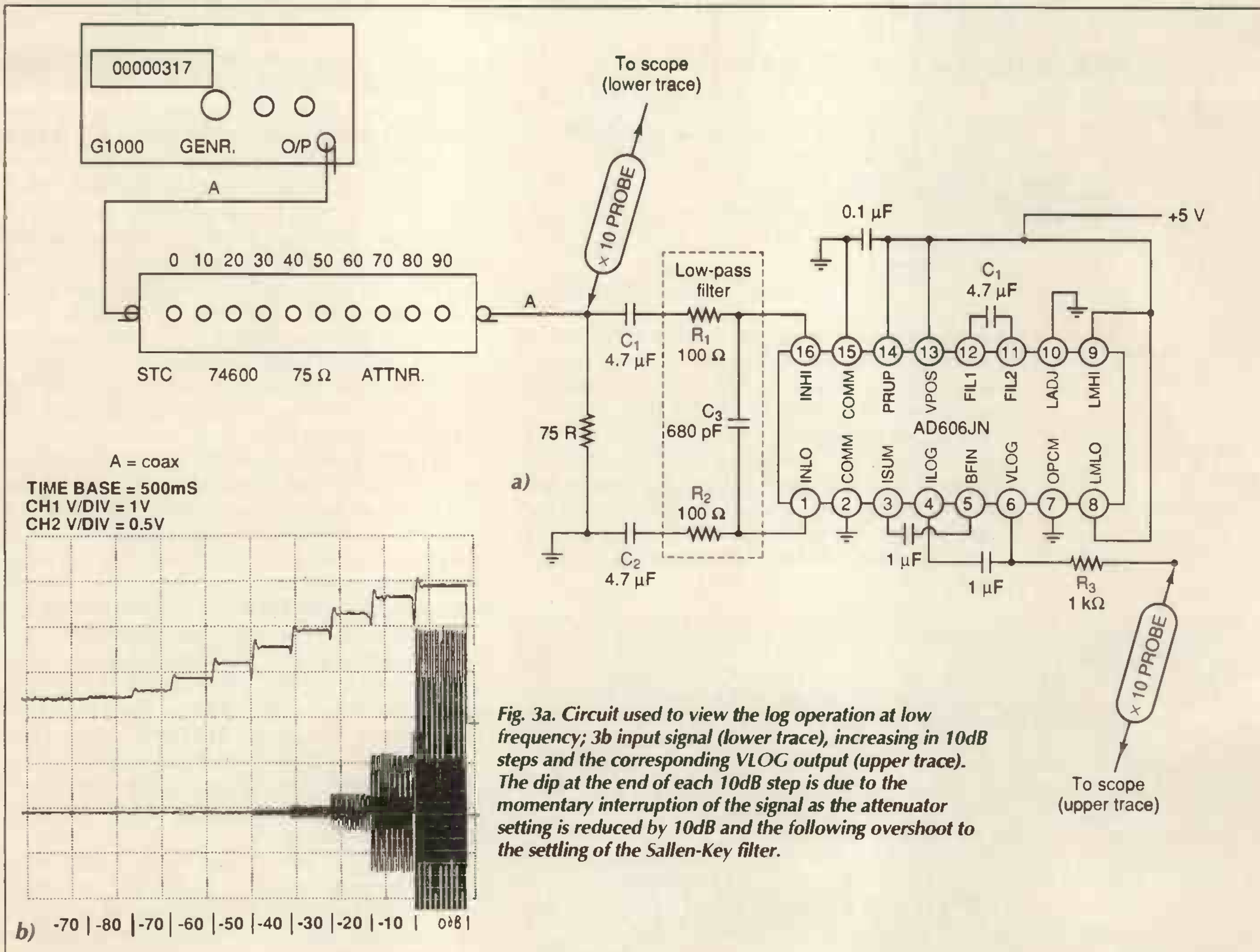



*Fig. 3a. Circuit used to view the log operation at low frequency; 3b input signal (lower trace), increasing in 10dB steps and the corresponding VLOG output (upper trace). The dip at the end of each 10dB step is due to the momentary interruption of the signal as the attenuator setting is reduced by 10dB and the following overshoot to the settling of the Sallen-Key filter.*

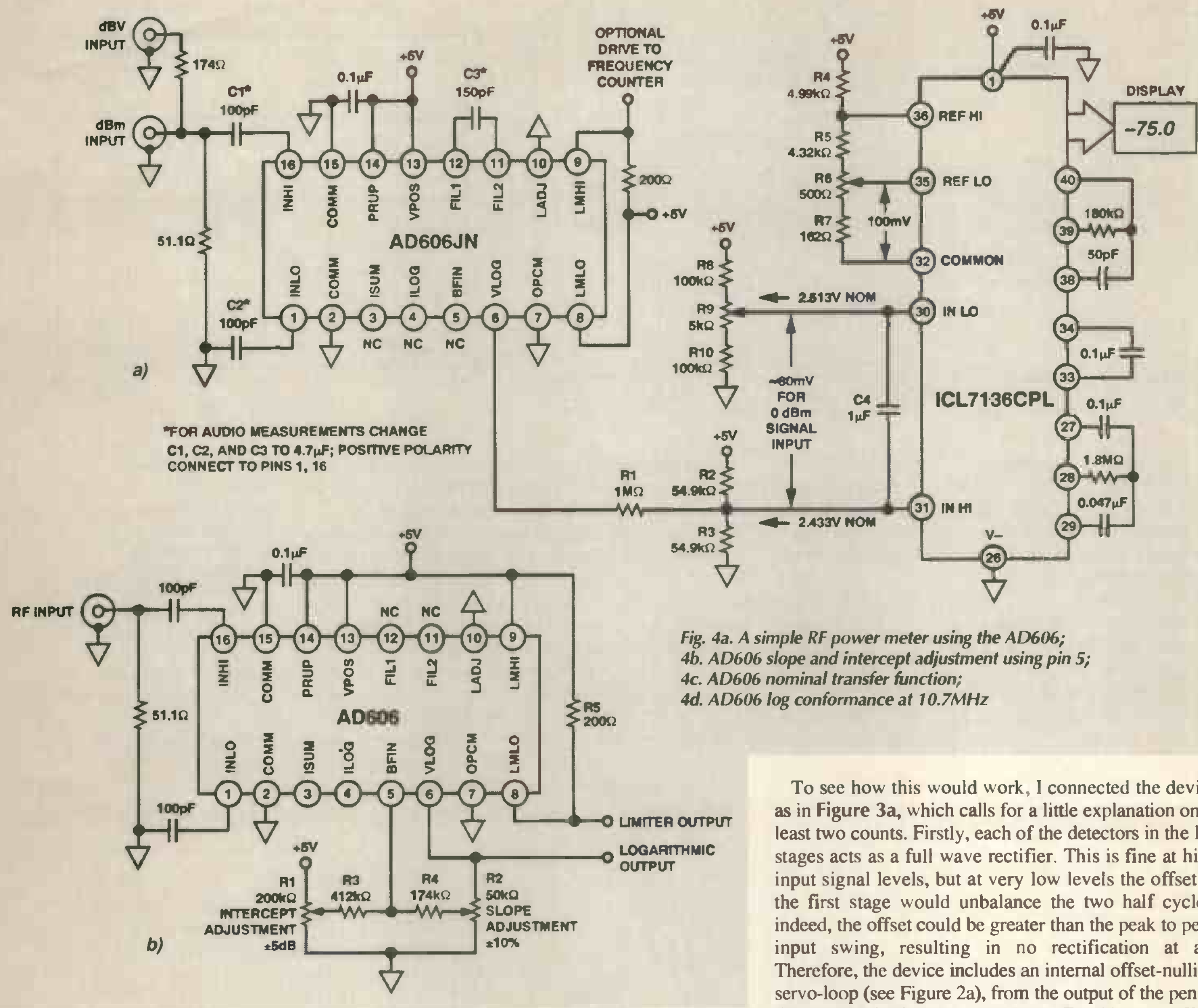



*Fig. 4a. A simple RF power meter using the AD606; 4b. AD606 slope and intercept adjustment using pin 5; 4c. AD606 nominal transfer function; 4d. AD606 log conformance at 10.7MHz*

IF output. For this purpose, the variation of phase with input signal level is specified in the data sheet. If an IF output is not required, these pins should be connected directly to the decoupled +5V.

The wide operating frequency range gives the chip great versatility for applications other than radar; for example, in an FM receiver, the detected video output with its logarithmic characteristic makes an ideal RSSI (received signal strength indicator). It can also be used in a low cost RF power meter and even in an audio level meter.

**Principles of radar**

In radar, a pulse of RF radiation – for example from an aeroplane – is transmitted from an antenna. The antenna (generally the same and probably directional) then receives the echo.

The radar designer faces a number of problems; for example in the usual single antenna radar, some kind of a switch is needed to route the transmit power to the antenna whilst protecting the receiver from overload. At other times all of the miniscule (small) received signal is routed from the antenna to the receiver.

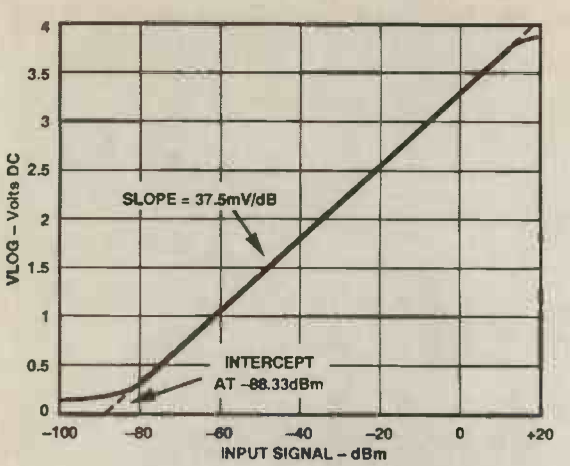
From then on, it is a battle to pull out wanted target returns from clutter (background returns from clouds, the ground or sea, etc) or, at maximum range, receiver noise, in order to maximise the Probability of Detection, PD , whilst minimising the Probability of False Alarm, PFA

To see how this would work, I connected the device as in **Figure 3a**, which calls for a little explanation on at least two counts. Firstly, each of the detectors in the log stages acts as a full wave rectifier. This is fine at high input signal levels, but at very low levels the offset in the first stage would unbalance the two half cycles: indeed, the offset could be greater than the peak to peak input swing, resulting in no rectification at all. Therefore, the device includes an internal offset-nulling servo-loop (see Figure 2a), from the output of the penultimate stage back to the input stage. For this to be effective at dc, the input must always be ac coupled as shown in Figs 3 and 4 and further, the input should present a low impedance at INLO and INHI (pins 1 and 16) so that the input stage "sees" only the ac input signal and not any ac via the nulling loop.

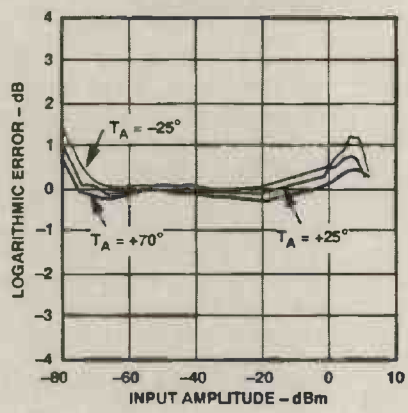
Turning to the second point, the cut-off frequency of the internal Sallen and Key lowpass filter driving the VLOG output at pin 6, is high, due to the 2pF capacitors used, so that at audio, the log output at pin 6 will show a rather squashed looking full wave rectified sinewave. This is fine if the indicating instrument is a moving coil meter, since its inertia will do the necessary smoothing.

Likewise, many DVMs incorporate a filter with a low cut-off frequency on the dc voltage ranges. However, as it was intended to display VLOG on an oscilloscope, the smoothing had to be carried out by the circuit and was in fact done in the device itself. The cut-off frequency of the Sallen and Key filter was lowered by bridging 1μF capacitors across the internal 2pF capacitors. Fortunately, all the necessary circuit nodes were available at the device's pins.

The 317Hz input to the chip and the VLOG output were displayed on the lower and upper traces of the oscilloscope respectively, Figure 3b. With the attenuator set to 90dB, the input was of course too small to see. The




c)




d)

attenuation was reduced to zero in 10dB steps, using the digital storage oscilloscope in roll mode, with all the steps clearly visible on the upper trace. The 317Hz test signal is not very completely delineated, due to having only three or four points per sample at the 0.5s/div sweep speed, but its peak amplitude (about 2V, somewhat in excess of the device's recommended maximum input of 1Vrms) is clearly indicated.

The 80 to 70dB step is somewhat compressed, probably owing to picking up stray RF signals since the device was mounted on an experimental plug board, and not enclosed in a screened box. With its high gain and wide frequency response, this chip will pick up any signals that are around.

The device proved remarkably stable and easy to use, although bear in mind that pins 8 and 9 were connected directly to the decoupled positive supply rail, as the limited IF output was not required in this instance.

**Figure 4a** shows how a very simple RF power meter, reading directly in dBm, can be designed using this IC. Note that here, the slope and and intercept adjustment have been implemented externally in the meter circuit, rather than internally via pin 5. Where this is not possible, the arrangement of **Fig. 4b** should be used.

This is altogether a most useful device: if it is hung on the output of a TV tuner with a sawtooth on its varactor tuning input, it provides a simple spectrum analyser with 80dB range log display. Clearly some extra IF selectivity in front of the *AD606* would be advisable. ■

**References**

1. GEC Plessey Semiconductors Professional Products IC Handbook.

# The healing face of electromagnetic fields

*How can the same portion of the electromagnetic spectrum both hurt and heal? Elizabeth Davies explains that the secret seems to be in our ability to tune them to our needs*

Not all news about electromagnetic fields is bad. Extensive publicity has been given to the role of electromagnetic (EM) fields in the onset of child leukaemias[1]. Worries over our exposure to mains frequencies and the fields produced by transformers have followed. But EM fields are not only thought to cause disease – they are also thought to cure it.

Over the last thirty years, low frequency EM fields have been finding more and more therapeutic applications. In orthopaedics they are used routinely by some surgeons to stimulate delayed bone union[2]. Many limbs have been saved from amputation by this technique.

Long-term venous ulcers, creating holes in flesh right down to the bare bone, have been healed quickly and completely by EM fields, and necrotic hip joints and osteoporotically thin bones have been revived to health and full density by low frequency fields.

So fields that can hurt, seem also to have a role in healing. By modulating cell growth to create a synchronised tumour cell population, a single dose of cytotoxic drug may be used to destroy all the tumour cells in one go, avoiding many of the side effects of prolonged chemotherapy.

## Black box/bone box?

The inherent beneficial effects of EM fields on bone cells seem to stem from the material structure of bone. Bone has piezoelectric properties: it reacts electrically to deformation; it produces streaming potentials in its fluids during stress; and it has semiconductor properties due to its collagen matrix. Natural occurrence of indigenous fields may be essential to normal bone function. Reproduction of these fields by exogenous signal generators, utilising Helmholz coils, (a non-invasive technique) or by implanting electrodes (an invasive, surgical technique) has been shown to heal bones: the electronic signal generators used are known colloquially as "bone boxes".

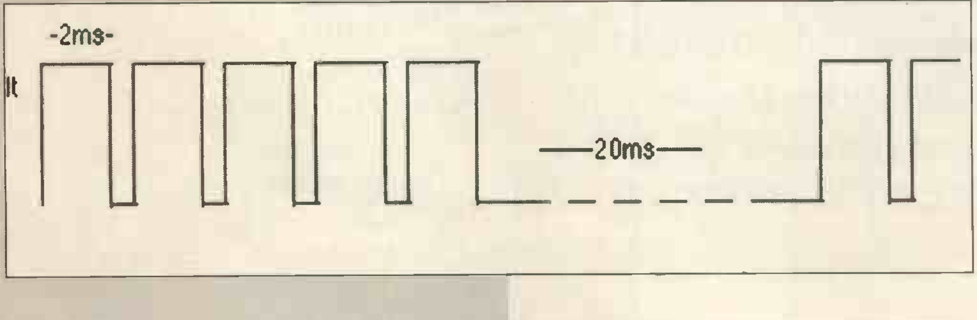
Following years of research (see box – *History)* by 1974 a method had been developed for optimising the waveform of the electrical stimulation, by simulating the endogenous waveform.

Shielded electrodes were attached to the surface of a bone, which was then subjected to mechanical loading – as in normal biological activity – and the electric signals produced were recorded on an oscilloscope.

Mechanical stimulation was stopped and a coil of copper wire placed close to the bone was connected to a Taccusel PIT-20-2A potentiostat. Driving characteristics of the potentiostat were then adjusted until the induced electric field picked up by the electrodes was identical to that produced by the mechanical deformation. Inductive coupling of the signal, via Helmholz coils, avoided signal distortion evidenced by direct coupling, due to different dielectric properties of the many biological tissues involved in transmitting the signal. Variations of the signal were used to treat the damaged bones of live animals, (dogs, rabbits, rats, chickens).

By comparing the results with damaged bones of a similar control group given placebo treatment, the most effective signal was optimised. In this way signals were developed for different pathological conditions of bone and other connective tissue.

*Bart's bone box has been used experimentally, though not clinically.*




## Unclear picture

But the scanty and incomplete picture formed by the research to date, shows that a great deal more time and money must be invested in unravelling the important parameters of both signal and receiver in this apparently important biophysical interaction.

Many orthopaedic therapeutic devices are now on the market, all designed by different companies and generating different signal waveforms, and all claim to have an indispensable characteristic in their waveform

### CONGRESS LANDMARK

The First World Congress for Electricity and Magnetism in Biology and Medicine takes place this June in Florida, bringing together the four main sponsoring journals in this field: *Bioelectromagnetic Society, Biological Repair and Growth Society, Bioelectrochemical Society,* and the *European Bioelectromagnetics Association.* The Congress is also sponsored and supported by the IEEE Engineering in Medicine and Biology Society, by the IEEE Power Engineering Society, by URSI Commission and by other international and industrial bodies. Men and women from a variety of disciplines will be attending and the beneficial as well as the hazardous effects will be discussed. The Congress should lead to heightened awareness and communication about the benign possibilities of electromagnetic fields, and perhaps it will generate a more coordinated international research initiative. It will certainly be a landmark for the future development of healing fields.

*Companies such as EBI are looking forward to the day when non-invasive treatment by EM fields will become an accepted technique.*

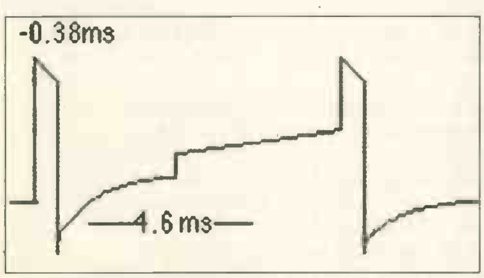
design. Unfortunately, commercial interests and patents have added mystique to the important parameters of each waveform. The waveform is frequently a square shaped pulse, or series of pulses known as a pulse train, but the range of signals is diverse.

The signal from the Bart's bone box, designed by Phil Byrne (Newcastle) induces 4mV in a 98-turn search coil, equivalent to 0.4gauss on a calibrated Hall probe. The pulse is asymmetrical and pulse repetition is 500Hz, gated at 50Hz. So far this signal has been use experimentally, but not clinically.

Bassett's signal used to treat osteonecrosis (dying bone) is a single asymmetric pulse repeated at 72Hz. It induces a voltage of 1-1.5mV/cm in the bone, and peak voltage induced in the 65-turn, 42 gauge wire search coil is 21.5mV positive and 3mV negative. Peak flux density is 5.4mT.

To treat a non-uniting fracture, a signal consisting of pulse trains of 3.5kHz pulses gated at 15Hz, induces a voltage of 2mV/cm in the bone.

Both these last signals have proved effective clinically.




*Osteonecrosis has been treated by Bassett's signal.*

## Frequency dependency

Fourier analysis of the clinically effective signals, reveal maximum power in the low frequency part of the spectrum, about 0.7Hz[5]. Dr Ross Adey, of the Brain Research Institute of the University of California, claims that the biological response to EM fields is frequency and dose dependent[6].

In 1980 Ross performed experiments with cat brain tissue, showing that the binding of calcium ions was affected by weak electromagnetic fields of a strength and frequency similar to endogenously produced fields seen on the electroencephalogram (EEG). He showed how a 147MHz field at tissue intensity of $0.8mW/cm^2$ increased calcium ion efflux from brain tissue only when amplitude modulated at 6-20Hz. Maximum stimulation was at 16Hz, and outside this frequency range window there was no effect. Using a similar waveform, C S Blackman in the US noted a narrow field intensity window, maximal at $0.75mW/cm^2$.

Adey, repeating his experiments, confirmed an intensity window between 0.1 and $1mW/cm^2$, using a 450MHz carrier wave amplitude modulated at 16Hz. Similarly, outside this amplitude window, there was no response. By contrast, ELF fields between 6-20Hz, inducing field intensities lower by several orders of magnitude, *decreased* calcium efflux. The interesting observation was that the frequency and size of the response of the tissue was the same whether it was stimulated or damped.

Consistency of the magnitude of response, 10-15% increase in efflux in either direction,

### HISTORY

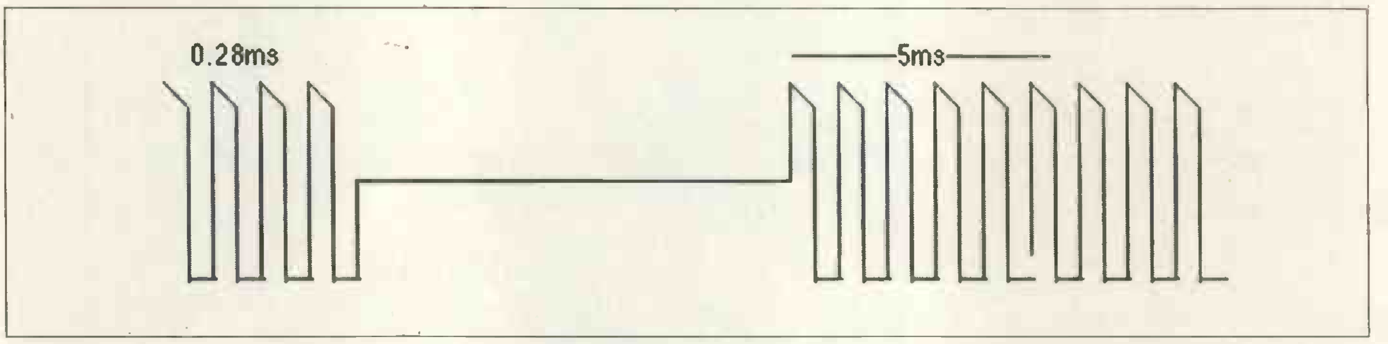
In 1959, the early pioneers of the therapeutic technique, Dr Robert Becker and Dr Andrew Bassett[3], were working on regeneration of amphibian limbs when they realised the potential of the EM field technique for human orthopaedic pathologies. As a result they were persuaded to use electromagnetic fields as a last resort in the case of severe bone trauma – offering an alternative to amputation.

Their success rate, even in the face of complications such as gangrene, was so outstanding that other orthopaedic surgeons soon took notice. As a result Bassett developed his own equipment, marketing it through a company he formed for the purpose, Electrobiology Incorporated, (EBI) of New Jersey, USA.

EBI is still researching and developing new equipment, funding research in laboratories all over the world, and has its own research directors.

In the US, the FDA (Food and Drug Administration) has funded a symposium to review and explore further the possible uses of electromagnetic field therapy. Conclusions of the scientific meeting were that electromagnetic field therapies would soon become a complementary medicine, available to every doctor.

Between 1984 and 1989 some excellent work, funded by EBI, was carried out at Strangeways Laboratory in Cambridge, by Sylvia Fitton-Jackson, Richard Farndale and Clifford Murray[4] using an *in vitro* bone cell model. This, and other in depth studies, helped EBI to develop a clinical signal that proved effective.




*Signal used to treat non-uniting fracture.*

together with the extreme sensitivity to narrow frequency and amplitude windows, suggested to Adey that a particular class of calcium membrane sites was sensitive to low frequency EM fields. Sensitivity was possibly caused by alterations in charge and hydrogen ion binding using quantum biochemical effects.

Adey's discovery has been corroborated independently by many other scientists in different laboratories.

### Electrical/mechanical link

Clinton Rubin's excellent experiments have shown a link between mechanical and electrical stimulation of bone. The effect suggests that longitudinal acoustical waves, as well as transverse EM waves, can be used to modulate cell growth. Rubin found a growth response of bone cells, maximising at 15Hz, and went on to show how this could be reproduced by externally applied EMF, with a maximal response, again, at 15Hz. He achieved the same stimulatory effect on bone growth with a higher amplitude ($10^{-3}T^2/s$) square pulse, containing only a fraction of the harmonic at 15Hz, as with a lower amplitude ($10^{-4}T^2/s$) pure sine wave – when that sine wave was at 15Hz.

The Cyclotron Resonance theory proposed by Liboff[7] and others attempts to explain the effect of EM fields on living matter through interaction with ions at a cell membrane level. Interestingly, it predicts a resonant response for calcium ions around 15Hz.

### Neglected therapy

Researching EM field therapy involves handling large numbers of variables and reproducibility is difficult. As a result some excellent workers have been glad to leave the field.

Nevertheless, many respected researchers who have seen for themselves the subtle metabolic changes effected by EMF, are positive that this effect will be used in the next century, when our understanding has grown. The technique could provide a complementary and even supplementary therapy for many diseases. Fine tuning of the body by EM fields might provide a remedy for presently incurable diseases, such as multiple sclerosis and osteoarthritis.

Yet this innovative non-invasive therapeutic technique remains underfunded and languishing in relative obscurity, applied and researched only by a few visionary and committed individuals. Many doctors are unaware of the existence of a therapeutic application of EM fields, and some biologists still adamantly deny any possible effect of insubstantial "fields" upon solid flesh and bone.

The effects are subtle, there is no doubt, and their safety needs to be further explored. Only collaboration of electronic engineers, mathematicians, biologists, doctors, physicists, biochemists and chemists will unravel the knot of bioelectromagnetic interactions. ■

*Elizabeth Davies is currently completing her PhD research thesis. She is employed in the Pharmacy Department, University of Brighton.*

### CALLING ELECTRONICS ENGINEERS

The signal/tissue response must be researched more thoroughly – different tissues seem to respond differently to different signals. But one practical problem is that only a small range of signal parameters can be investigated thoroughly with one biological model in the time scale of a normal research project. Researchers have used a range of models and signal parameters, but few have attempted a really substantial study using one model for a whole range of signal parameters, or vice versa. The problem lies in the temporal and financial constraints imposed on any piece of research.

What is wanted is a good, inexpensive variable research tool, so that basic research can be standardised. Researchers need a signal generator, inductively coupled to Helmholz coils, that will produce, athermally, square pulses as well as sine waves – of different frequencies, amplitudes, continuous or gated in trains.

The unit would also need to generate, via a second pair of coils, a DC magnetic field of variable strength. With this addition, it could also be used to test the cyclotron resonance theory.

Such a tool could be used with a number of specified biological models, to exhaust all of its parameters, and a completed map compiled. Until this happens research remains enigmatic, exciting but fragmentary and lacking in systematic coordination of investigative effort.

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3. CA Bassett. 1992."Bioelectromagnetics in the Service of Medicine" *Bioelectromagnetics* 13: 7-17, 1992.
4. C Murray, R Farndale. 1985. "Modulation of Collagen Production in Cultured Fibroblast by a Low Frequency PMF" *Biochim et Biophysica* Acta: 838: 98-105.
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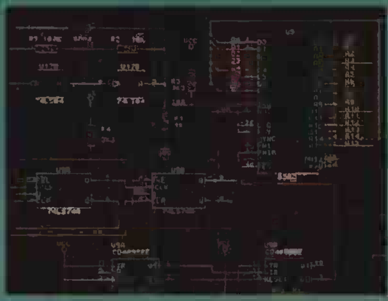
6. R Adey, Bawin. 1980. "Non-equilibrium Processes in Binding and Release of Brain Calcium by low level EMF." *Bioelectrochemistry:* Ions, Surfaces and Membranes" Ed. M. Blank. Publisher: *Advances in Chemistry Series*. 188: American Chemical Society.
7. AR Liboff, BR McLeod. "Kinetics of Channelized Membrane Ions in Magnetic Fields." *Bioelectromagnetics*: 9, 1988.

# FROM CONCEPT TO ARTWORK IN 1 DAY



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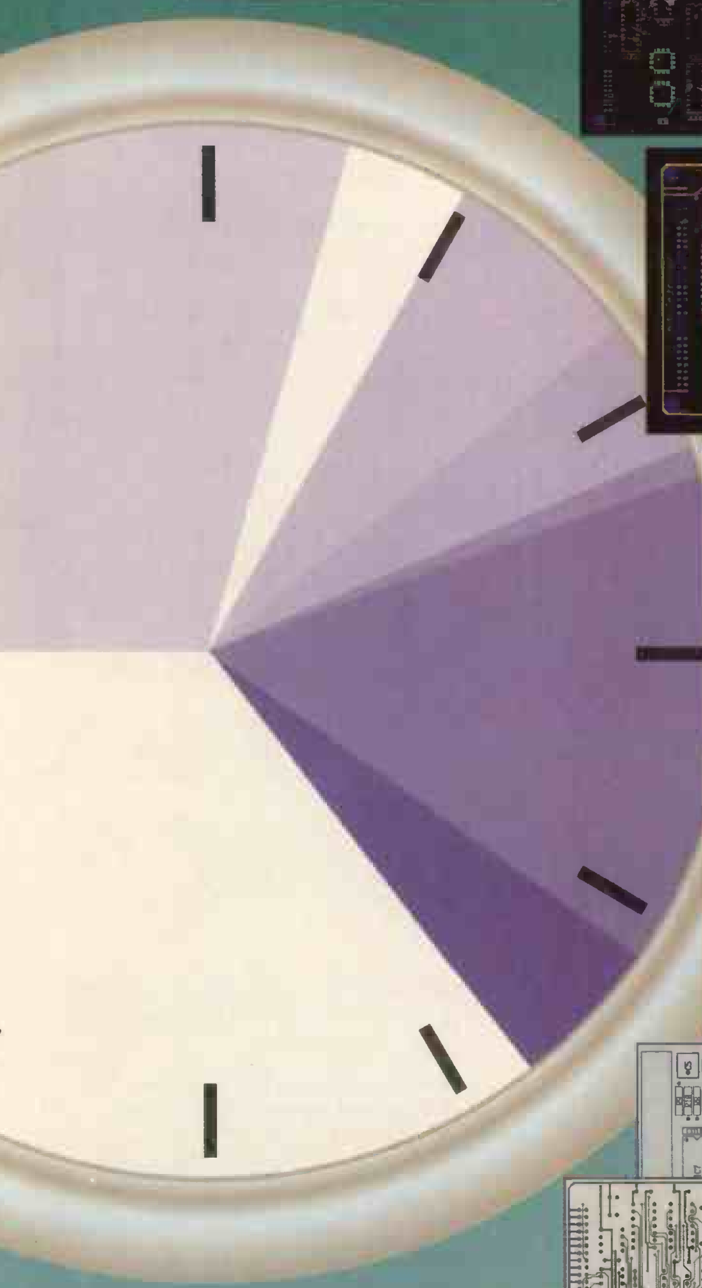

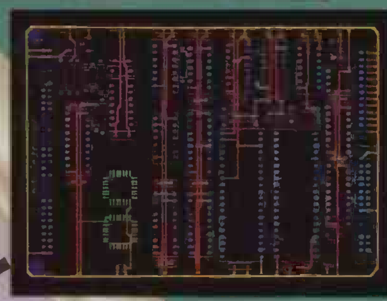

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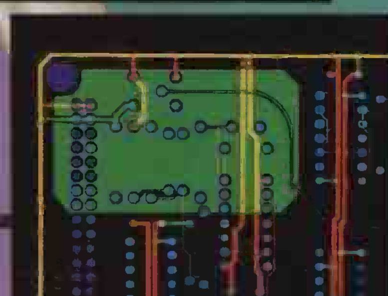

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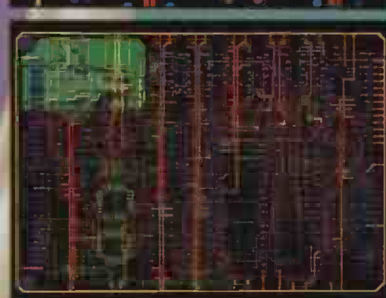

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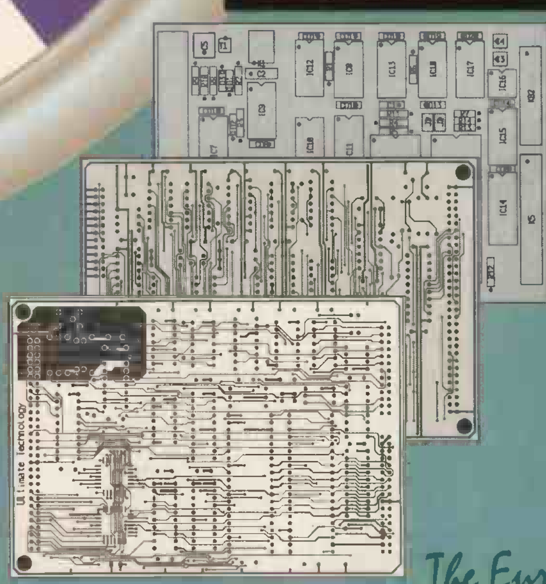

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# CIRCUIT IDEAS

*Send your circuit design ideas to The Editor, Electronics World, Quadrant House, The Quadrant, Sutton SM2 5AS. UK*

## Preset on time for battery equipment

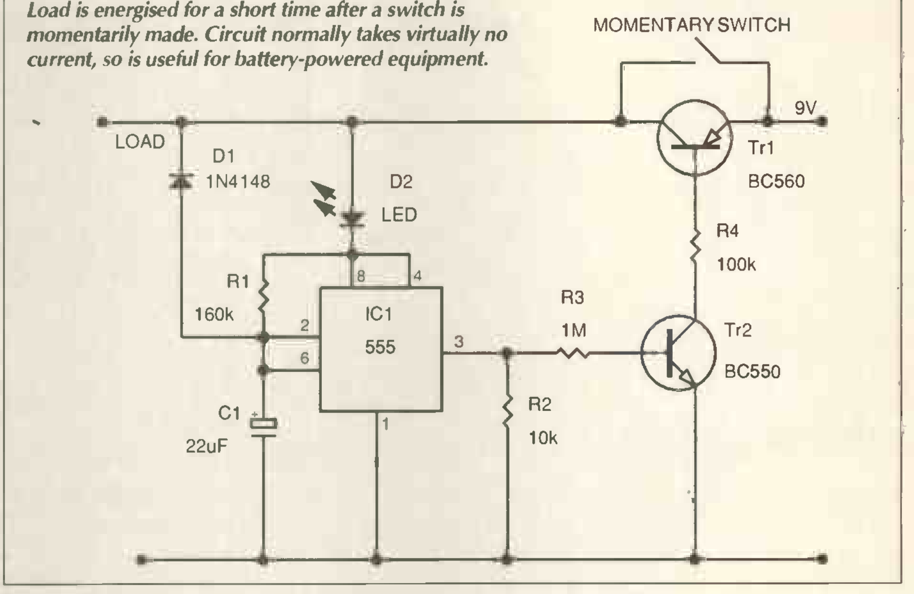
This circuit was designed to switch an alarm on for a short time after a switch is momentarily made, while normally drawing no current.

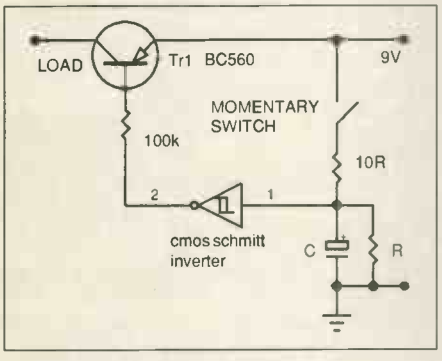
Operating switch $Sw_1$ applies voltage to the load and to the *555* timer $IC_1$. Current drawn by the *555* illuminates the led, $D_2$, to indicate that the circuit is on and also to reduce current drain by lowering the supply voltage to the timer. At switch-on, pin 3 of the timer is high, so that $Tr_{2,1}$ are both conducting. After the momentary switch contact, $Tr_2$ still supplies load current until $C_1$ charges through $R_1$ – about 2s with the values shown. Capacitor $C_1$ eventually triggers the timer, pin 3 goes low, both transistors turn off and the circuit becomes quiescent. Diode $D_1$ discharges the capacitor when the load voltage collapses.

Resistor $R_2$ is an additional load to avoid an intermediate state in which feedback puts the timer into a linear configuration.

***Steve Winder***
*Ipswich*

*Load is energised for a short time after a switch is momentarily made. Circuit normally takes virtually no current, so is useful for battery-powered equipment.*




*Alternative arrangement recommended by our editorial consultant*

## State machine for 2.5s division ratio

In Circuit Ideas for December 1991, p.1051, Yongping Xia proposed a method of pulse frequency division by 2.5. My method uses a programmable logic device and state machine technique, thereby showing that PLDs can be used for asynchronous logic.

**Figure 1** shows input and output, which are to be repeated until the starting input/output relationship is repeated, the divider then cycling in a loop.

Input conditions for every state transition are shown in **Fig. 2**, the state chart compiled by inspection of the waveforms in Fig. 1. Each of the ten states is given an unique state code and, since the operation is asynchronous, it must be a Gray code sequence in which the progression is by a change of only one code bit each time.

To program the PLD, one must prepare a file to define pin functions, state code bits and logic conditions for active inputs and outputs. In the several design software packages such as Cupl, Abel and some shareware software from the manufacturers mentioned in this journal for July 1989 p.667, there is provision for defining transitions using the format:

Again, in an asynchronous system, the PLD clock input must be programmed and connected in the inactive state. A rough idea of a practical circuit is shown in **Fig. 3**, but a manufacturer's data is needed for a working design.

***J Austin***
*Wallasey*
*Merseyside*

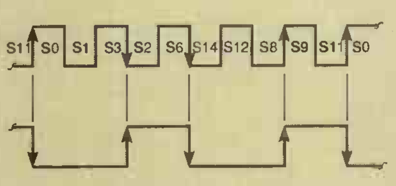



*Fig. 1. Input and output waveforms of 2.5 divider. Cycle repeats when starting conditions are repeated.*

```
present(state) if(input condition)   next(state)
               if(input condition)   next(state)
               out(output).
```

*Fig. 2. State chart of divider system. Ten states are needed, numbered in Gray code, since the system is asynchronous.*

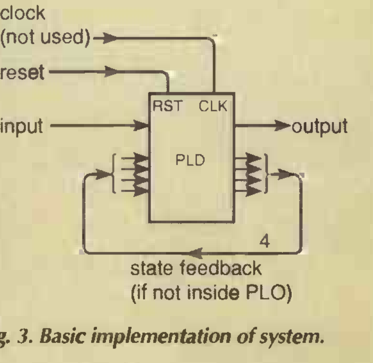



*Fig. 3. Basic implementation of system.*

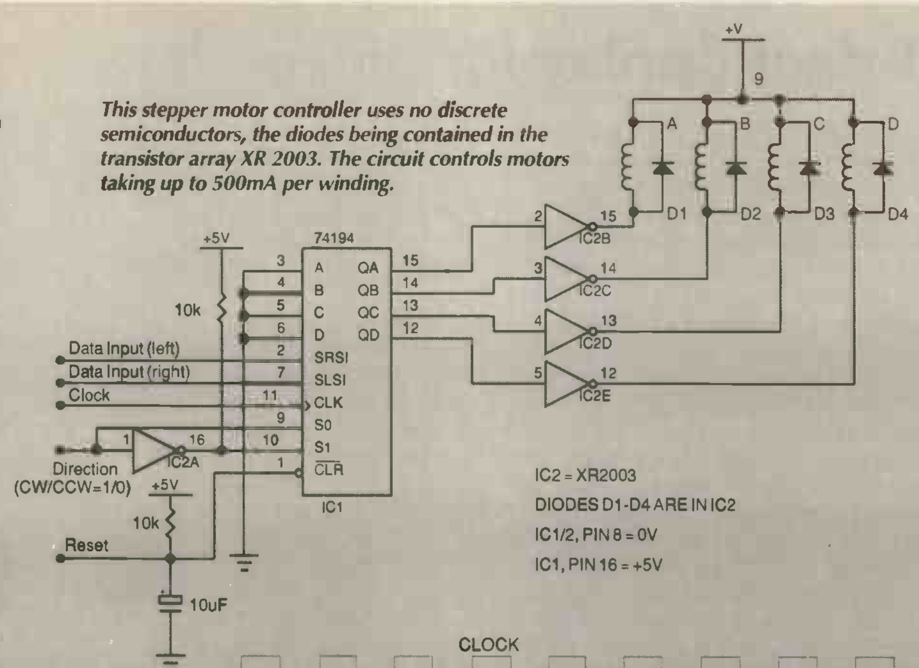
# Stepper motor control

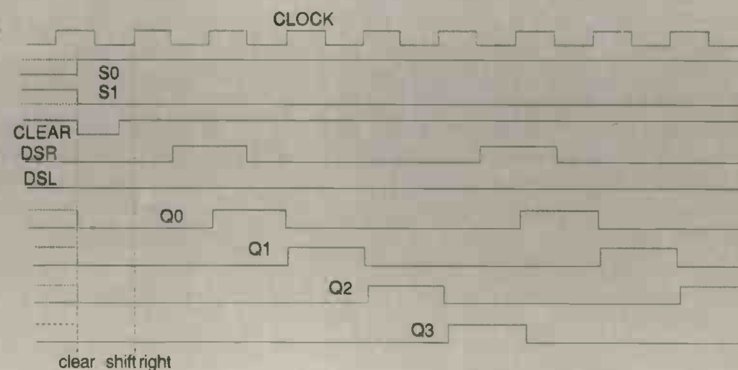
Two chips, a universal shift register and a darlington transistor array, form a stepper motor controller with no visible discrete devices; the free-wheeling winding diodes are in the array. Pulses into the data shift right/left inputs of the *74194* universal shift register produce logic sequences for both directions of motor rotation, depending on the polarity of the direction control signal to the $S_0$ and $S_1$ inputs; the input data is sequentially output from $Q_{0-3}$ or $Q_{3-0}$ at each positive-going clock pulse.

Each device in the *XR2003* seven-transistor array handles a 500mA continuous collector current at up to 45V and devices may be paralleled. The inverter on the direction input to the *74194* is part of the *XR2003*.

***V Lakhshminarayanan***
*Centre for Development of Telematics*
*Bangalore*
*India*

*This stepper motor controller uses no discrete semiconductors, the diodes being contained in the transistor array XR 2003. The circuit controls motors taking up to 500mA per winding.*




*Waveforms seen in the circuit for clockwise motor rotation.*

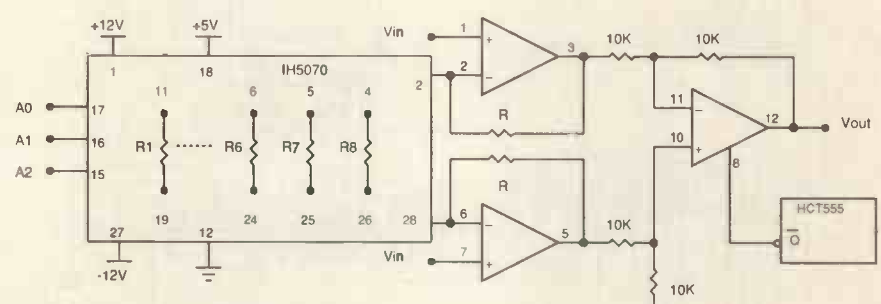


# Programmable instrumentation amplifier

Gain of this three-op-amp amplifier is given by $A=(1+2R/R_x)$, $x$ being the value of one of the resistors in the *IH5070*. Selecting this resistor by the $A_0$-$A_2$ lines produces a programmed-gain amplifier. The *TAB1042*, as well as being a conventional op-amp, is also an analogue switch, shutting down when no bias current goes to pin 8. A timer feeds pin 8, so that the operating time is programmable from about 1s to 24h.

***Kamil Kraus***
*Rokycany*
*Czechoslovakia*

AMP IS TAB1042
SUPPLIES ARE +/-12V




*Three op-amp instrumentation amplifier, programmed in time range and gain, which is set by selecting one of the IH5070 resistors by the A0-A2 digital input.*

# 4-digit display for binary data

To present 14-bit binary signals on a seven-segment display, this circuit uses three ICs, one of them an eprom.

In the diagram, the Q4 and Q5 outputs of the free-running counter $IC_1$ drive the A14 and A15 addresses of the 64Kbyte eprom $IC_3$, input data being taken to A0-13. Output from $IC_3$ 00-06 is a seven-segment drive signal for the display, taken via current-limiting resistors.

Outputs Q4 and Q5 from $IC_1$ drive a 3-to-8 decoder $IC_2$, whose outputs Y0-3 select one of the four dispays, since the input data chooses four different addresses. As an example, if the input is 1A4Chex., equivalent to 6732, the address is 1A4Chex. when $IC_1$ Q4 and Q5 are 0, $IC_2$ Y0=0 and the right-hand display operates. Since Y1,2,3=0, the other three are off. If Q4=1 and Q5=0, the address is 3A4Chex., Y1=0 and the next display comes on. In this way, if 1A4Chex., 3A4Chex., 5A4Chex. and 7A4Chex are programmed with the seven-segment of 2,3,7,6 respectively, the displays show these characters one by one, flicker being reduced by a high scan speed.

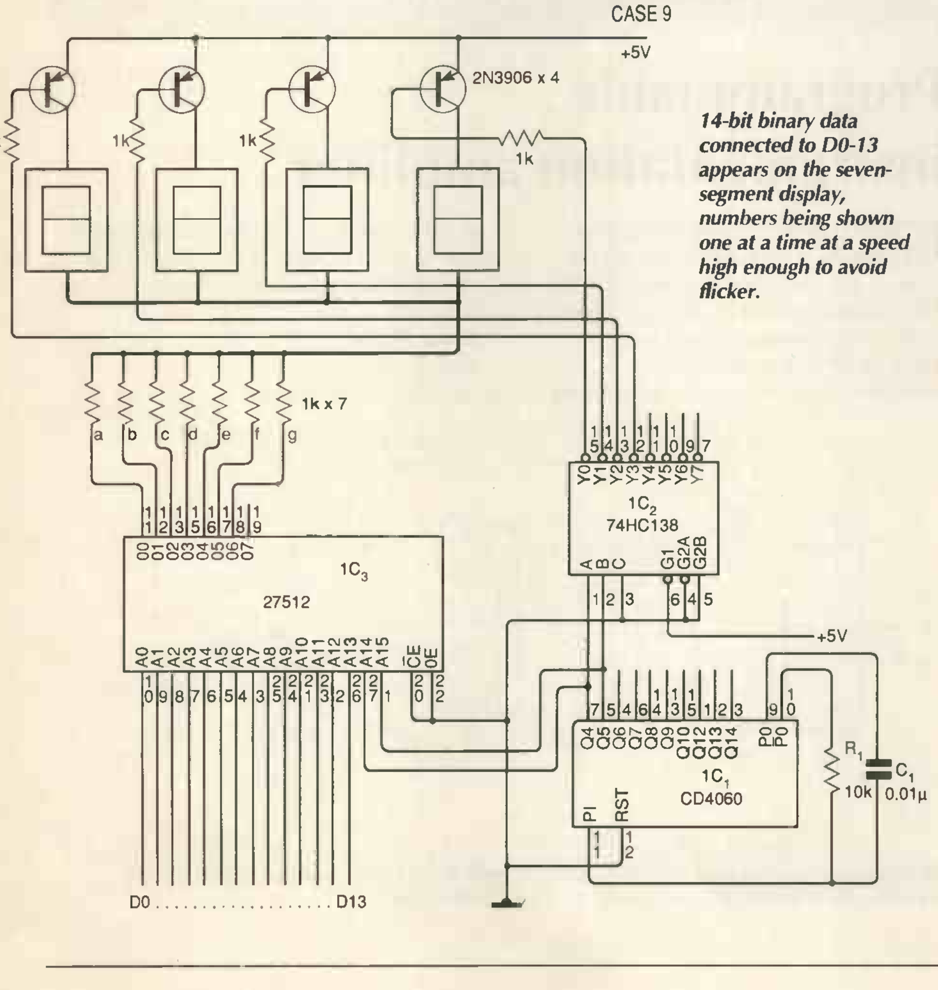
Maximum display is 9999 and the display is off for greater numbers.

The following QuickBasic listing generates the eprom files, files 1 to 4 storing the seven-segmant forms of four digits right to left.

```
DIM N(4)
DIM SEGMENT(4)
OPEN "DATA1" FOR OUTPUT AS #1
OPEN "DATA2" FOR OUTPUT AS #2
OPEN "DATA3" FOR OUTPUT AS #3
OPEN "DATA4" FOR OUTPUT AS #4

FOR NUMBER = 0 TO 9999
    N(1) = NUMBER-INT(NUMBER/10)*10
    N(2) = INT(NUMBER/10)-INT(NUMBER/100)*10
    N(3)=INT(NUMBER/100)-INT(NUMBER/1000)*10
    N(4) = INT(NUMBER/1000)
    FOR I = 1 TO 4
        SELECT CASE N(I)
        CASE 0
        SEGMENT(I) = 40
        CASE 1
        SEGMENT(I) = 79
        CASE 2
        SEGMENT(I) = 24
        CASE 3
        SEGMENT(I) = 30
        CASE 4
        SEGMENT(I) = 19
        CASE 5
        SEGMENT(I) = 12
        CASE 6
        SEGMENT(I) = 02
        CASE 7
        SEGMENT(I) = 78
        CASE 8
        SEGMENT(I) = 00
        CASE 9
        SEGMENT(I) = 10
        END SELECT
        WRITE *I, SEGMENT(I)
    NEXT I
NEXT NUMBER
CLOSE #1
CLOSE #2
CLOSE #3
CLOSE #4
```

***Yongping Xia***
*West Virginia University*
*Morgantown WV, USA*




***14-bit binary data connected to D0-13 appears on the seven-segment display, numbers being shown one at a time at a speed high enough to avoid flicker.***

# Transconductance squarer

An op-amp and a dual fet combine to give an output $kv_i^2$ when $v_i>0$.

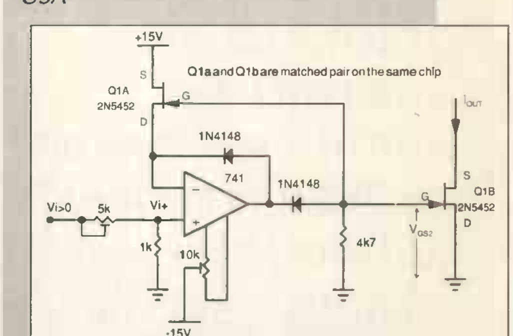
National Semiconductor's *2N5452* n-channel dual fet has good matching between the two devices and low output conductance. Normally, the voltage between the fet gates and the non-inverting op-amp input is constant at $V_p$, the pinch-off voltage of the two transistors. $V_{GS2}$ is $(v_i^+ + V_p)$ and $i_{out}$ is proportional to $v_i^2$.

The coefficient k is adjustable by means of the 5kΩ input variable, the two diodes and the 4.7kΩ resistor ensuring that $i_{out}$ does not exceed $I_{DSS}$ and affording negative feedback should $v_i^+$ become greater than $|V_p|$. Output voltage must lie within the 5-15V range.

Including an absolute-value detector at the input produces a true squarer, in that either polarity of input gives the same output.

With a *741*, the circuit works at several kilohertz.

***Alexandru Ciubotaru***
*University of Texas at Arlington*
*Texas*
*USA*




***Current at the output of this simple circuit is proportional to the square of the input voltage, for inputs greater than zero. Absolute-value circuit makes it a true squarer.***

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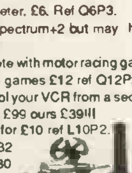
**PC STYLE POWER SUPPLY** Made by AZTEC 110v or 240v input. +5 @ 15A,+12 @ 5A,-12 @ .5A,-5 @ .3A. Fully cased with fan, on/off switch, IEC inlet and standard PC flyleads. £15.00 ref F15P4

**TELEPHONE HANDSETS** 10 brand new handsets with mic and speaker only £3.00 for 10 ref 3P146R

**BENCH POWER SUPPLIES** Superbly made fully cased (metal) giving 12v at 2A plus a 6V supply. Fused and short circuit protected. For sale at less than the cost of the case! Our price is £4.00 ref 4P103R
**SPEAKER WIRE** Brown twin core 100 feet for £2.00 REF 2P79R
**720K 3 1/2" DISC DRIVE FOR £9** Brand new units made by JVC complete with tech info just £9.00!! they have a metal tab instead of a button and you may want to fit an led. Combined power and data cable easily modified to IBM standard. ref L9P2.
**MONO VGA MONITORS £59** Standard IBM compatible monitor made by Amstrad Ex display Our price just £59. Ref 59P4RB.
**CAR BATTERY CHARGER** Brand new units complete with panel meter and leads. 6 or 12v output £7.00 ref J7P2.
**CUSTOMER RETURNED SPECTRUM +2**
Complete but sold as seen so may need attention £25.00 ref J25P1
**CUSTOMER RETURNED SPECTRUM +3**
Complete but sold as seen so may need attention £25.00 ref J25P2

**AMSTRAD 1640DD BASE UNITS**
**BRAND NEW AND CASED**
**TWO BUILT IN 5 1/4" DRIVES**
**MOTHER BOARD WITH 640K MEMORY**
**KEYBOARD, MOUSE & MANUAL**
**OUR PRICE JUST**
**£79!!!!**

**SCART TO D TYPE LEADS**
Standard Scart on one end, Hi density D type (standard VGA connector) on the other. Pack of ten leads only £7.00 ref 7P2R
**OZONE FRIENDLY LATEX**
250ml bottle of liquid rubber sets in 2 hours. Ideal for mounting PCB's fixing wires etc. £2.00 each ref 2P379R
**VIEWDATA SYSTEMS**
Brand new units made by TANDATA complete with 1200/75 built in modem infra red remote controlled qwerty keyboard BT appproved Prestel compatible, Centronics printer port RGB colour and composite output (works with ordinary television) complete with power supply and fully cased. Our price is only £20.00 ref 20P1R
**COMMODORE 64 COMPENDIUM** Pack consisting of a Commodore 64 computer, power supply, data recorder and software. All for £69 ref O69P1.
**PPC MODEM CARDS** Made for the Amstrad PPC1640/1512 range these are plug in modules that operate at 2400 baud. No data. £15 ref Q15P5.
**AMSTRAD LQ3500 PRINTER ASSEMBLIES** Entire mechanical assemblies including print head, platen, cables, stepper motors etc etc. infact everything bar the electronics and case! Our price just £10 ref O10P3.
**AMSTRAD DMP4000 PRINTER ASSEMBLIES** Entire printer assemblies including print head, platen, cables, stepper motors etc. Everything bar the electronics and case. Our price just £20 ref O20P2.
**TOROIDAL TRANSFORMER** 146VA with tappings at 8v, 10v and 32v will give 50v at 3A or 32 at 4A etc. Centre tapped primary. £9 ref O9P2. Fixing kit is £2 ref O2P1.
**AERIAL BRACKETS** Wall plate 7.5" sq complete with rawl bolts, 10" stand off brackets with standard tube clamps. Will take up to 2" mast. Substantial bracket (would take body weight). £7 ref O7P1.
**TV SOUND RECEIVERS** Popular units that with the addition of a speaker act as a tv sound receiver. Ideal as a stand alone unit or for connecting into HI FI!. £12 ref Q12P4
**2,000 COMPONENTS FOR £3** Yes thats right! just send us £3 and you can have 1,000 resistors plus 1,000 capacitors! Our choice of value. Order ref O3P1.
**ETRI FANS** Mains, 11 watt 80mm diameter. £6. Ref O6P3.
**LIGHTGUNS** Originally made for the Spectrum+2 but may have other uses (good stripper). £2 Ref Q2P3
**GX4000 GAMES CONSOLES** Complete with motor racing game, psu and joystick £15 ref Q15P3. Extra 4 games £12 ref Q12P2
**VCR RABBIT SYSTEM** Lets you control your VCR from a second set using the VCR remote control. Retail £99 ours £39!!
**CAMERAS** Customer returned units. 3 for £10 ref L10P2.
**STEAM ENGINE** Standard Mamod 1332 engine complete with boiler piston etc £30 ref 30P200



**HANDHELD TONE DIALLERS**
Small units that are designed to hold over the mouth piece of a telephone to send MF dialling tones. Ideal for the remote control of answer machines. £5.00 ref 5P209R
**AMAZING TALKING COINBOX!**
Fully programmable talking, lockable coinbox BT approved, retail price is £79 ours is just £29! ref J29P2.
**ANSWER PHONES £15** Customer returned units with 2 faults one we tell you how to fix the other you do your self! £18 ref J18P2 or 4 for £60 ref J60P3 BT approved (retail price £79.95!! each)
**COMMODORE 64 MICRODRIVE SYSTEM**
Complete cased brand new drives with cartridge and software 10 times faster than tape machines works with any Commodore 64 setup. The orginal price for these was £49.00 but we can offer them to you at only £25.00! Ref 25P1R

**BULL ELECTRICAL**
**250 PORTLAND ROAD HOVE SUSSEX**
**BN3 5QT TELEPHONE 0273 203500**
**MAIL ORDER TERMS: CASH PO OR CHEQUE WITH ORDER PLUS £3.00 POST PLUS VAT.**
**PLEASE ALLOW 7 - 10 DAYS FOR DELIVERY**
**FAX 0273 323077**

**90 WATT MAINS MOTORS** Ex equipment but ok. Good general pupose unit £9.00 ref F9P1
**HI FI SPEAKER BARGAIN** Originally made for TV sets they consist of a 4" 10 watt4R speaker and a 2" 140R tweeter. If you want two of each plus 2 crossovers for £5.00 ref F5P2.
**EMERGENCY LIGHTING SYSTEM**
Fully cased complete with 2 adjustable flood lights. All you need is a standard 6v lead acid battery. Our price is just £10 ref J10P29
**AMSTRAD 464 COMPUTERS** Customer returned units complete with a monitor for just £35! These units are faulty non returnable.
**WOLSEY DMAC DECODERS**
Made for installation in hotels etc as the main sat receiver no data but fully cased quality unit. £20 ref K20P1. Suitable psu £8 ref K8P3.
**REMOTE CONTROLS**
Brand new infra red CONTROLS originally made for controlling WOLSEY satellite receivers £2 ea ref K2P1 or 20 for £19 ref K19P1.
**DOS PACKS** Complete set of PC discs with DOS 3.2, basic, gemdesktop & gem paint. No manuals. 5 1/4" discs £10 ref K10P2
**CORDLESS TIE CLIP MICROPHONE**
transmits between 88-108MHZ FM 5.2cm x 2cm, uses LR44 watch battery. Complete with wire aerial & battery. £16 ref K16P1.
**CHASSIS MOUNT TRANSFORMERS**
240v primary, 12v secondary 20VA £2 ref K2P2
240v primary, 16v secondary 10A (split winding). £10 ref L10P1
**100 RED LED PACK (5MM) £5 REF K5P2**
**12V STEPPER MOTOR** Ideal for models etc. 3" dia £2 ref J2P14
**CAPACITOR BARGAIN PACK** 100 CERAMICS £2 REF J2P2.
**SPECTRUM JOYSTICKS TWO FOR £5 REF J5P2.**

**AMSTRAD PC CASE, POWER SUPPLY AND 720k FLOPPY DRIVE ALL THIS FOR £30 REFQ30P15**

**USEFUL POWER SUPPLIES.** 18v 900mA dc output (regulated) fully cased with mains cable and DC out cable. £6 ref K6P1.
**UNCASED PC POWER SUPPLIES.** Standard PC psu without case, fan etc Good for spare or low cost PC!. £4 ref L4P6.
**RADAR DETECTORS. Detects X and K bands (ie speed traps). Not legal in the UK so only available if you intend to 'export'it. £59 ref J59P1.**
**100 WATT MOSFET PAIR.** Same spec as 2Sk343 and 2SJ413 (8A,140v, 100w) 1 N channel and 1 P channel £3 a pair ref J3P9.
**LOW COST CAPS.** 1,000 capacitors £3 (33uf,25v) ref J3P10.
**VELCRO.** 1 metre length 20mm wide, blue. £2 ref J2P16.
**JUG KETTLE ELEMENTS.** Good general purpose heating element just £3 ea ref £3P8 or 5 for £10 ref J10P3.
**VERY BIG MOTOR.** 200v induction 1.1kw 1410 rpm 10"x7" GEC 1" keyed shaft. Brand new. £95 ref J95P1.
**BIG MOTOR.** 220-240v 1425rpm 2.8A 5/8th" keyed shaft GEC 6.5" x 8" complete with mounting plate. £38 ref J38P1.
**SMALL MOTOR.** Electrolux 160 watt 3,000 rpm. 220-240v 5/8" shaft precision built £18 ref J18P1.
**EPROMS 27C64 PACK OF 10 £7 REF M7P1. 27C256 PK OF 10 £9 REF M9P1. 27C512 PK OF 10 £10 REF M10P1.**
**MODEMS FOR £1.25?** These modems are suitable for stripping only hence they are only 4 for £5 ref J5P3.
**SOLAR POWERED WOODEN MODELS.** Complete with solar panel, motor and full instructions. £9 ref J9P2. 3 diff £20 ref J20P3.
**SOUND OPERATED LIGHT.** Clap your hands and light comes on. Turns after preset delay. (4 AA's req'd). £2 ref J2P3.
**FERGUSON SRB1 REMOTE CONTROLS.** Brand new units ideal for a spare or have two remotes! £4 each.
**5 1/4" 360K DISC DRIVE** Made for AMSTRAD 1640/1512 machines. White front. Our price just £9 ref Q9P1.

## PC CORNER

**PC CASES** Desktop case +psu £51.60 ref BPCC1, Deluxe slimline case +psu £60.00 ref BPCC2, Minitower case +psu £51.60 ref BPCC3, Deluxe midi case +psu £90.00 ref BPCC4.
**MONITORS** Mitac 14" SVGA .39DP £174 ref BPCM02, Mitac 14" SVGA .28DP £202 ref BPCM01.
**MEMORY** 256K Simm 70ns £8.40 ref BPCMI1, 1MB Simm 70ns £26.40 ref BPCMI2, 4MB Simm 70ns £96 ref BPCMI3
**MICE** 2 button serial mouse with 3.5" s/ware. £8.40 ref BPCMI6, 3button serial mouse with 3.5" s/ware £9.60 ref BPCMI7.
**KEYBOARDS** 102 AT UK standard keyboard £18.60 ref BPCMI4, Deluxe keyboard 102 AT UK £26.40 ref BPCMI5.
**SOFTWARE** MS DOS V5 OEM version. £39.60 ref BPCMI8, MS WINDOWS V3.1 OEM version. £42 ref BPCMI9.
**MOTHERBOARDS** 286-16 Headland chipset £46.80 ref BPCMB1, 386SX-33 Acer chipset £82.80 ref BPCMB2, 386SX-40 UMC with 64K cache £110 ref BPCMB3, 486SX-25 UMC with 64k cache £191 ref BPCMB4, 486DX-33 UMC with 256k cache £378 ref BPCMB5, 486DX-66 UMC with 256k cache £515 ref BPCMB6.
**FLOPPY DRIVES** 1.44mb 3.5" drive £32.34 ref BPCDD05, 1.2MB 5.25" drive £38.40. 3.5" mounting kit £5 ref BPCDD07.
**HARD DRIVES** 42MB IDE 17ms £99 ref BPCDD01, 89MB IDE 16ms ref BPCDD02, 130MB IDE 15ms £215 ref BPCDD03, 213MB IDE 14ms £298 ref BPCDD04.
**VIDEO CARDS** 256k C&T 8 bit SVGA card £19.20 ref BPCVC01, 512k Trident 9000 16 bit SVGA card £31.20 ref BPCVC02, 1MB Trident 8900 16 bit SVGA card £45 ref BPCVC03, 1MB Cirrus AVGA3 16.7M colours £48 ref BPCVC04, 1MB Tseng multimedia £82.80 ref BPCVC05.
**ADD ON CARDS** Multi I/O card 2 serial, 1 parallel, 1 game, 2 floppy, 2 IDE hard drives. £11 ref BPCAOC01, ADLIB sound card with speakers £37 ref BPCAOC02. Orchid sound card with speakers £63 ref BPCAOC03.

**EXAMPLES OF COMPLETE SYSTEMS**

**386SX-33 SYSTEM**
386SX-33 board at £82.80, case £51.60, 2MB ram £52.80, 42MB drive £99, 512SVGA card £31.20, 3.5" FDD £32.34, multi I/O card £11 SVGA colour monitor £174, 102 k/board, £25 build fee if required. Total £579.34
**486DX-33 SYSTEM**
486Dx-33 board £378, case £51.60, 2MB ram £52.80, 89MB drive £166, 512 SVGA card £31.20, 3.5" FDD £32.34, multi I/o card £11, SVGA monitor £174, 102 k/board £18.60, £25 build fee if required. Total £939.84.
**ALL PC PARTS AND SYSTEMS ARE GUARANTEED FOR 1 YEAR PARTS AND LABOUR.**
**1993 CATALOGUE AVAILABLE WITH ALL ORDERS IF REQUESTED OTHERWISE A4 SAE FOR FREE COPY.**

**SOME OF OUR PRODUCTS MAY BE UNLICENSABLE IN THE UK**

**IN SUSSEX? CALL IN AND SEE US!**

*CIRCLE NO. 124 ON REPLY CARD*

## COMPUTER ICS
P8271 BBC DISC CONTROLLER CHIP EX EQPT ... £20
SAA5050 TELETEXT CHIP EX EQPT ... £5
2817A-20 (2K×8) EEPROM ex eqpt. ... £2
80C88A-2 used ... £1.25
27C64-25 used/wiped ... £1.50 100+ £1
27S191 PROM ... £2
IMS1400P-45 ... £2
80C31 MICRO ... £2
P8749H MICRO ... £5
D8751-8 NEW ... £10 USED+WIPED ... £6
D41257C-15 (256K x 1 DRAM) ... £1 ea
NEW 4164-15 ... £1
USED 41256-15 ... £1
USED 4164-15 ... 60p
BBC VIDEO ULA ... £10
6845 CRT ... £5
6522 PIA ... £3
AY3-1015D UART ... £2.50
9 × 41256-15 SIMM ... £10
8 × 4164 SIP MODULE NEW ... £8
2864 EPROM ... £3
27128A 250ns EPROM USED ... £2 NEW £2.30
27C1001-20Z NEW 1M EPROM ... £6
FLOPPY DISC CONTROLLER CHIPS 1771 ... £10
FLOPPY DISC CONTROLLER CHIPS 1772 ... £17.50
68000-8 PROCESSOR NEW ... £6
HD6384-8 ... £5
ALL USED EPROMS ERASED AND BLANK CHECKED
CAN BE PROGRAMMED IF DESIRED
2716-45 USED ... £2 100/£1
2732-45 USED ... £2 100/£1
2764-30 USED ... £2 100/£1.60
27C256-30 USED ... £2
27C512 USED ... £3.50
1702 EPROM EX EQPT ... £5
2114 EX EQPT 50p 4116 EX EQPT ... 70p
6264-15 8k STATIC RAM ... £2
GR281 NON VOLATILE RAM EQUIV 6116 ... £5
Z80A SIO-O ... £1.25
TMS27PC128-25 ONE SHOT 27C128 ... £1 ea 100/£70
80387-16 CO-PROCESSOR (OK WITH 25MHz 286) ... £50
7126 3½ DIGIT LCD DRIVER CHIP ... £2 ea
2816A-30 HOUSE MARKED
... £2 each QUANTITY AVAILABLE

## REGULATORS
SANKEN STR451 USED IN AMSTRAD MONITORS ... £5
78H12ASC 12V 5A ... £5
78M05 5V 0.5A ... 7/£1
LM317H T05 CAN ... £1
LM317T PLASTIC TO220 variable ... £1
LM317 METAL ... £2.20
7812 METAL 12V 1A ... £1
7805/12/15/24V plastic ... 25p 100+ 20p 1000+ 15p
7905/12/15/24 plastic ... 25p 100+ 20p 1000+ 15p
CA3085 TO99 variable reg ... 2/£1
L387 5v ½A WITH RESET OUTPUT ... £1ea £50/100

## CRYSTAL OSCILLATORS
1M000 1M8432 4M000 10M000 16M000 18M432000 19M0500
20M0500 38M10000 56M6092 ... £1.50 each

## CRYSTALS
1M0 1M8432 2M000 2M304 2M4576 2M77 3M00 3M2768
3M579545 3M58564 3M93216 4M000 4M19304 4M433619
4M608 4M9152 5M000 5M0688 6M0000 6M400 8M000 8M488
9M8304 10M240 10M245 10M70000 11M000 12M000 13M000
13M270 14M000 14M381818 15M000 16M000 16M5888
17M000 20M000 21M300 21M855 22M1184 24M000 34M368
36M75625 36M76875 36M78125 36M79375 36M80625
36M81875 36M83125 36M84375 38M900 49M504 54M19166
54M7416 57M75833 60M000 69M545 69M550

## TRANSISTORS
BC107 BCY70 PREFORMED LEADS
full spec ... £1 £4/100 £30/1000
BC557, BC238C, BC308B ... £1/30 £3.50/100
2N3819 FETS short leads ... 4/£1

## POWER TRANSISTORS
P POWER FET IRF9531 8A 60V ... 3/£1
N POWER FET IRF531 8A 60V ... 2/£1
2SC1520 sim BF259 ... 3/£1 100/£22
TIP 141/2 £1 ea TIP 112/125/42B ... 2/£1
TIP35B/TIP35C ... £1.50
SE9301 100V 1DA DARL SIM TIP121 ... 2/£1
PLASTIC 3055 OR 2955 equiv 50p ... 100/£35
2N3773 NPN 25A 160V £1.60 ... 10/£14
2N3055H ... 4 for £2

## TEXTOOL ZIF SOCKETS
28 WAY ZIF EX NEW EQUIPMENT ... £2.50
40 WAY NEW ... £5
SINGLE IN LINE 32 WAY CAN BE GANGED FOR USE WITH
ANY DUAL IN LINE DEVICES . . . COUPLING SUPPLIED
... 2/£1.50

## CAPACITORS COMPUTER GRADE
24,000µF 50V ... £3 (£1.30)
10,000µF 100V SPRAGUE/PHILIPS ... £6 (£2.00)

## QUARTZ HALOGEN LAMPS
12V 50watt LAMP TYPE M312 ... £1 ea HOLDERS 60p ea
24V 150 WATTS LAMP TYPE A1/215 ... £2.50 each

## NEW BITS
ETHERNET 4 PAIR TRANSCEIVER CABLE. BELDEN TYPE
9892 ... £60 for 50 metres
SIM RS 361-018 Each pair foil screened + overall braided
screen ... £300 for 305 metre drum
+ plus PVC outer sheath. ON 305 metre drums
10nF 63V X7R PHILIPS SURFACE MOUNT 160K
available ... £30/4000 box
330nF 10% 250V X2 AC RATED PHILIPS 60K
AVAILABLE ... £20/100
220R 2.5W WIREWOUND RESISTOR 60K AVAILABLE
... £50/1000

## MISCELLANEOUS
CMOS 555 TIMERS 20k available ... 2/£1 £30/100
ICM7126CPL CMOS 3½ DIGIT LCD DRIVER CHIP ... £2ea
'SAFEBLOCKS' MADE BY RENDAR . . . MAINS 'RAT TRAP'
USED ... £4 ea
36 CORE 7/0.2mm OVERALL SCREENED ... £50/100m
LITHIUM CELL 1/2 AA SIZE ... 2 FOR £1
PASSIVE INFRA RED SENSOR CHIP + MIRROR + CIRCUIT
£2 each
EUROCARD 21-SLOT BACK PLANE 96/96-WAY ... £25 ea
"PROTONIC 24 VARIBUS" 16.7"×5" FIBREGLASS
MULTILAYER PRH PCB
EUROCARD 96-WAY EXTENDER BOARD 290×100mm
... £10 ea
"PROTONIC 24" c/w 2 SUPPORT ARMS/EJECTORS.
DIN 41612 96-WAY A/B/C SOCKET PCB RIGHT ANGLE £1.30
DIN 41612 96-WAY A/B/C SOCKET WIRE WRAP PINS £1.30
DIN 41612 64-WAY A/C SOCKET WIRE WRAP PINS ... £1
DIN 41612 64-WAY A/C PLUG PCB RIGHT ANGLE ... £1
DIN 41612 64-WAY A/B SOCKET WIRE WRAP (2-ROW BODY)
... £1
BT PLUG+LEAD ... 3/£1
13A MOULDED PLUG+2m lead ... £1
MIN. TOGGLE SWITCH 1 POLE c/o PCB type ... 5/£1
LCD MODULE sim. LM018 but needs 150 to 250V AC for display
40×2 characters 182×35×13mm ... £10
TL431 2.5 to 36V TO92 ADJ. SHUNT REG ... 2/£1
6-32 UNC 5/16 POZI PAN SCREWS ... £1/100
NUTS ... £1.25/100
PUSH SWITCH CHANGEOVER ... 2/£1
RS232 SERIAL CABLE D25 WAY MALE CONNECTORS
... £5.90 ea (£1.30)
25 FEET LONG, 15 PINS WIRED BRAID + FOIL SCREENS
... INMAC LIST PRICE £30
STICK ON CABINET FEET RS NO 543-327 ... 30/£1
LCD DISPLAY sim Hitachi LM016L ... £6.50
AMERICAN 2/3 PIN CHASSIS SOCKET ... 2/£1
HUMIDITY SWITCH ADJUSTABLE ... £2
WIRE ENDED FUSES 0.25A ... 30/£1
NEW ULTRASONIC TRANSDUCERS 32kHz ... £2/pr
Also available 26 slot vari-bus backplane same size +
Price ... NEW
POWERFUL SMALL CYLINDRICAL MAGNETS ... 3/£1
BNC 50OHM SCREENED CHASSIS SOCKET ... 2/£1
SMALL MICROWAVE DIODES AE1 OC1026A ... 2/£1
D.I.L. SWITCHES 10-WAY £1 8-WAY 80p 4/5/6-WAY ... 80p
180VOLT 1WATT ZENERS also 12V & 75V ... 20/£1
VN 10LM 60V ½A 5 Ohm TO-92 mosfet ... 4/£1 100/£20
MIN GLASS NEONS ... 10/£1
RELAY 5V 2-pole changeover looks like RS 355-741 marked
STC 47WBost ... £1 ea
MINIATURE CO-AX FREE PLUG RS 456-071 ... 2/£1
MINIATURE CO-AX FREE SKT RS 456-273 ... 2/£1.50
DIL REED RELAY 2 POLE n/o CONTACTS ... ££
PCB WITH 2N2646 UNIJUNCTION WITH 12V 4-POLE RELAY
... £1
400m 0.5W thick film resistors (yes four hundred megohms)
... 4/£1
STRAIN GAUGES 40 ohm Foil type polyester backed balco grid
alloy ... £1.50 ea 10+ £1
ELECTRET MICROPHONE INSERT ... £0.90
Linear Hall effect IC Micro Switch no 613 SS4 sim RS 304-267
... £2.50 100+ £1.50
HALL EFFECT IC UGS3040 + magnet ... £1
OSCILLOSCOPE PROBE SWITCHED ×1 ×10 ... £12
1 pole 12-way rotary switch ... 4/£1
AUDIO ICS LM380 LM386 TDA 2003 ... £1 ea
555 TIMERS £1 741 OP AMP ... 6/£1
ZN414 AM RADIO CHIP ... 80p
COAX PLUGS nice ones ... 4/£1
COAX BACK TO BACK JOINERS ... 3/£1
4×4 MEMBRANE KEYBOARD ... £1.50
INDUCTOR 20µH 1.5A ... 5/£1
1.25" PANEL FUSEHOLDERS ... 3/£1
CHROMED STEEL HINGES 14.5×1" OPEN ... £1 each
12V 1.2W small w/e lamps fit most modern cars ... 10/£1
STEREO CASSETTE HEAD ... £2
MONO CASS. HEAD £1 ERASE HEAD ... 50p
THERMAL CUT OUTS 50 77 85 120°C ... £1 ea
THERMAL FUSES 220°C/121°C 240V 15A ... 5/£1
TRANSISTOR MOUNTING PADS TO-5/TO-18 ... £3/1000
TO-3 TRANSISTOR COVERS ... 10/£1
PCB PINS FIT 0.1" VERO ... 200/£1
TO-220 micas + bushes ... 10/50p 100/£2
TO-3 micas + bushes ... 15/£1
PTFE min screened cable ... 10m/£1
Large heat shrink sleeving pack ... £2
IEC chassis plug filter 10A ... £3
POTS SHORT SPINDLES 2K5 10K 25K 1M 2M5 ... 4/£1
40k U/S TRANSDUCERS EX-EQPT NO DATA ... £1/pr
LM335Z 10MV/degree C ... £1
LM234Z CONST. CURRENT I.C. ... £1
BNC TO 4MM BINDING POST SIM RS 455-961 ... £1
BUTTON CELLS SIM. AG10/AG12 ... 4/£1
MIN PCB POWER RELAY 12V COIL 6V CONTACTS 2 P C/O
... £1.25
AVEL-LINDBERG MOULDED TRANSFORMER TYPE OB10
15+15V 10VA QTY. AVAILABLE ... £2 ea
BANDOLIERED COMPONENTS ASSORTED Rs, Cs, ZENERS
... £5/1000
LCD MODULE 16 CHAR. X 1 LINE (SIMILAR TO HITACHI
LM10) ... £5
KYNAR WIRE WRAP WIRE ... £1/REEL
OPI1264A 10kV OPTO ISOLATOR ... £1.35 ea 100+ £1 ea
'LOVE STORY' CLOCKWORK MUSICAL BOX MECHANISM
MADE BY SANKYO ... £1 ea
Telephone cable clips with hardened pins ... 500/£2

## DIODES AND RECTIFIERS
A115M 3A 600V FAST RECOVERY DIODE ... 4/£1
1N5407 3A 1000V ... 8/£1
1N4148 ... 100/£1.50
1N4004 SD4 1A 300V ... 100/£3
1N5401 3A 100V ... 10/£1
BA158 1A 400V fast recovery ... 100/£3
BY127 1200V 1.2A ... 10/£1
BY254 800V 3A ... 8/£1
BY255 1300V 3A ... 6/£1
6A 100V SIMILAR MR751 ... 4/£1
1A 600V BRIDGE RECTIFIER ... 4/£1
4A 100V BRIDGE ... 3/£1
6A 100V BRIDGE ... 2/£1
8A 200V BRIDGE ... 2/£1.35
10A 200V BRIDGE ... £1.50
25A 200 V BRIDGE £2 ... 10/£18
25A 400V BRIDGE £2.50 ... 10/£22
2KBP02 IN LINE 2A 200V BRIDGE REC ... 8/£1

## SCRS
PULSE TRANSFORMERS 1:1+1 ... £1.25
2P4M EQUIV C106D ... 3/£1
TICV106D 800mA 400C SCR 3/£1 ... 100/£15
MEU21 PROG. UNIJUNCTION ... 3/£1

## TRIACS ... DIACS 4/£1
NEC TRIAC ACO8F 8A 600V TO220 ... 5/£2 100/£30
TXAL225 8A 500V 5mA GATE ... 2/£1 100/£35
BTA 08-400 ISO TAB 400V 5mA GATE ... 90p
TRAL2230D 30A 400V ISOLATED STUD ... £5 ea
TRIAC 1A 800V TLC381T 16k AVAILABLE
... 5 FOR £1 £15/100

## CONNECTORS
D25 IDC SOCKET FUJITSU ... £2
34-way card edge IDCCONNECTOR (disk drive type) ... £1.25
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# ACTIVE

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

**Television D-to-As.** Maximum conversion rate of 60MHz, with a linearity error of 0.1% are offered by Fujitsu's *MB40730/60* 10-bit digital-to-analogue conveters. The */30* variant accepts ECLinput and provides 0 to -2V output, while */60* takes TTL input to give 3-5V output. Hawke Components Ltd, 0256 880800.

**S/H amplifier.** *LF6197* is a high-speed sample-and-hold amplifier from National Semiconductor, which will acquire a 10V step to within 0.01% in 160ns. It uses ±5V to ±18V supplies and is input-compatible with cmos, TTL or ECL. National Semiconductor, 0793 614141.

**8-channel data acquisition.** Unitrode's *MN7450/1* self-calibrating, 8-channel, 47kHz, 16-bit data acquisition devices contain an input multiplexer, software-programmable-gain amplifier and a 16-bit sampling A-to-D converter. The input multiplexer has make-before-breal operation and latched address inputs, and the amplifier is gain-programmable for A=1,2,4 or 8 with no extra components. *MN7450* allows inputs of 0-5V or ±5V, while *MN7451* accepts 0-10V or ±10V. Unitrode (UK) Ltd, 081 318 1431.

## Discrete active devices

**Fast rectifiers.** For use with high-frequency switched-mode power supplies, inverters and as free-wheeling diodes, GI has available the *Ultra Gold UG* series of superfast rectifiers, which exhibit a recovery time of 15ns. Maximum voltage in the range is 200V and operating temperature is –65°C to 150°C. General Instrument (UK) Ltd, 0895 272911.

**Another fast rectifier.** Rectifiers by GI in the *BYV26D/E* series are miniature glass passivated junction types with short recovery times and peak reverse ratings up to 1kV. General Instrument (UK) Ltd, 0895 272911.

**410mW Zeners.** Surface-mounted zeners from ITT in the BZT52 range offer a 410mW capacity and meet the E24 standard. Thirty versions in voltages from 2.7V to 51V at 5mA have leakage currents down to 0.05μA. Maximum current is 250mA at 25deg.C. ITT Semiconductors, 0932 336116.

**Video switch.** Siliconix's new *DG641/2* cmos analogue switches exhibit an on resistance of, typically, 8Ω and 5Ω respectively. With a bandwidth of 500MHz, current handling capacity is 75mA/100mA. Crosstalk at 5MHz is 87dB/85dB. Siliconix Ltd, 0635 30905.

## Digital signal processor

**Adaptive comb filter video gen.** Motorola's MC141621 advanced comb filter video signal processor separates luminance and chrominance from composite video signals. This being an "advanced" design, it it examines the horizontal and vertical signal transitions before selecting the type of filtering to minimise dot-crawl, cross-colour and colour smear. The device can also be used in A-to-D conversion or as a normal comb filter. Motorola Inc., (USA) (602)244-3818.

**Fast, 3V DSP.** TI's *TMS320C5x* 16-bit digital signal processor range now includes 3.3V devices having increased speed to 25ns, rather than suffering the more common reduction in speed. There is power management to give 1.5mA/Mips and two power-down modes for increased battery life. Texas Instruments, 0234 223252.

## Linear integrated circuits

**9MHz dual op-amp.** A "Butler" front end – a combined jfet and bipolar circuit – is used in Analog's *OP-285* op-amp to give a 9MHz gain/bandwidth, 15V/μs slewing, 250μV offset and 6nV√Hz. Long-term voltage drift is 300μV. Analog Devices, 0932 253320.

**50MHz log amplifier.** Although chiefly intended as a demodulating logarithimic amplifier for signal strength indication in mobile telephones and receivers, Analog's *AD606* is also useful as a limting amplifier in FM demodulators and in wireless lans; both limting and logarithmic outps are present. Dynamic range is –75 to 5dBm to within ±1dB. Analog Devices, 0932 253320.

**850MHz video buffer.** Packaged as a standard op-amp, Harris's new *HFA1112* closed-loop buffer amplifier combines 850MHz bandwidth with 0.033dB gain flatness to 100MHz, 1,700V/μs slewing and programmable gain of +1, -1 or +2. Third-harmonic distortion is -80dB at 50MHz and differential phase and gain are 0.02degree and 0.02%. Harris's UHF1 technology is employed in the new device. Harris Semiconductor, 0276 686886.

**Mosfet driver.** *LTC1154* is a single, high-side mosfet driver that has extensive overload protection, draws only 8μA standby and 85μA on. To drive n-channel fets for high-side switching, an internal charge pump is incorporated, as is programmable over-current sensing. An enable input allows control of banks of the devices. Linear Technology Ltd, 0276 677676.

**Supervisor ICs.** Maxim's *MAX703/4* offer protection to sensitive circuitry against power failure, providing battery back-up, power failure warning, automatic and manual reset, while drawing only 200μA of quiescent current and 50μA in back-up mode. *MAX703* is meant for 5% supplies and the *704* for 10% rails. Ram contents are protected by switching to an emergency voltage when the supply falls below the trip threshold. Maxim Integrated Products Ltd, 0734 845255.

**180W audio amplifier.** Using mixed bipolar/cmos/dmos, SGS-Thomson's *TDA7294* is a class-AB audio power amplifier, working on up to ±40V to give 50W continuous and 180W with a music signal. At 5W and 1kHz, THD is 0.005% and over the full band of 20Hz-20kHz and at 50W, THD is 0.1%. Muting is effective for a short time after switch-on to avoid snaps, crackles and pops. SGS-Thomson Microelectronics, 0628 890800.

## Logic building blocks

**Delay lines.** A choice of four programmable delays is available in the Dallas *DS1020* 8-bit delay line, a cmos device whose delay is variable in 256 steps to maxima of 73.75ns in the fastest device or to 520ns in the slowest of four. Logic states are reproduced without inversion at the output after the delay. There is an inherent delay of 10ns in all the devices. Joseph Electronics Ltd, 021 643 6999.


**Mixed-signal ICs.** Telephone chip. Mitel's *TC200C* is a digital telephone circuit, which has a high-level data control formatter for the ISDN interface. Fully differential audio paths have gain controls for transmit, receive and side tones and a transmit amplifier. A-law and μ-law are implemented, with CCITT G.714 filtering. Transducers interface directly to the device. The use of digital signal processing enables half-duplex, hands-free speakerphone switching and the *TC200C* generates all 16 DTMF tones. Mitel Semiconductor, 0291 430000.

**DTMF receivers.** Mitel's *MT3X7XB* family of low-power DTMF receivers are 8-pin devices for integrated telephone answering machine, end-to-end signalling and fax application. These devices decode all 16 tone pairs into a 4-bit binary form, clocked out synchronously as serial data. The */70B* has a software-controlled guard time, while the */71B* uses an internal counter for guard-time validation with no other guard-time circuit. Mitel Semiconductor, 0291 430000.

## Memory chips

**Data saver.** If the voltage supply to the Xicor's Autostore Novram falls below a preset level, automatically saves the contents of a computer's sram in a shadow eeprom, with no intervention from the computer's processor. At power-up, the data is recalled into the less-than-35ns sram for use. Data is thereby protected during accidental power failures or

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**Compensated oscillators.** As an alternative to oven-controlled crystal oscillators, SEI offers the *OD9208* series of digitally compensated devices that cost less than half the price of oven types and use a hundredth of the power. Short-term stability is better than $2 \times 10^{-9}$ and in the long term better than 0.5ppm in a year. Frequencies range from 4MHz to 25MHz with a tolerance of better than ±0.3ppm. External adjustment can handle ±5ppm. SEI Ltd, 0706 367501.

system crashes. Micro Call Ltd, 0844 261939.

## Microprocessors and controllers

**3V microcontrollers.** New in the Hitachi *H8/300* family of microcontrollers is the *H8/329* series, which is equivalent to the earlier H8/325, but with 3V operation.Four members of the series, *H8/329/8/7/6*, offer 32K rom and 1Kb ram, 24K/1Kb, 16K/512b and 8K/256b respectively, all having 8-channel, 8-bit A-to-D converters, full uart interface, 51 i/o ports, a 16-bit timer and an 8-bit counter/timer. Operating speed is 5MHz. Hitachi Europe Ltd, 0628 585000.

**Embedded engine.** Sharp's *LH72501* embedded engine includes components to control interfaces, buses, system frequency etc., in addition to the CPU, which features virtually all the functions of a PC XT-compatible computer in one chip. At 3V, the chip takes 50mA at 8MHz. Three separate power-saving modes are available, keyboard entries and mouse movements restoring full power. Sharp Electronics GmbH, (Germany) 040/23 760

**PC stereo.** In two chips, Crystal Semiconductor's *CS4215* audio codec and *CS4231* digital audio controller confer 16-bit stereo audio on the PC. Software support, including Windows drivers, is supplied. Design is such that the PC's CPU is not overloaded. Sequoia Technolgy Ltd, 0734 311822.

## Optical devices

**SM leds.** Dialight claims its microLED series to be the smallest surface-mounted leds in the world. They are made in both top-view and right-angle-view forms, the top-view version having a 180deg. viewing angle and measuring 3.2 by 1.27 by 1.2mm. The devices come in a number of colours including blue and infrared emitters and detectors for 700, 880 and 940nm. BLP Components Ltd, 0638 665161.

## Oscillators

**Direct digital synthesiser.** For outputs up to 12MHz, Novatech's DDS-3 synthesiser kit is programmable for sine and TTL/cmos output in 2Hz steps. Output is 1.4Vpk-pk into an open circuit. Output frequency can be changed in less than 250ns and there is better than –90dBc phase noise at ikHz offset, –45dBc spurious and –40dBc harmonic. The kit contains all necessary components, including a PCB. Novatech Instruments Inc., (US) 206 328 6902.

## Power semiconductors

**Audio power modules.** All ILP audio power amplifier modules are now stocked by Cirkit Distribution. The range covers modules fromthose for low-power hi-fi amplifiers to units for high-power stage amplifiers – 15-180W. Bipolar, mosfet and Class-A modules are available, each with heat sink and protective circuitry. Total harmonic distortion produced by the mosfet designs is less than 0.005%. Cirkit Distribution Ltd, 0992 444111.

**500V, 2A mosfet driver.** An "intelligent" half-bridge mosfet and igfet driver IC from Harris, the *HIP2500*, drives a high-side n-cahnnel power switch from low-side logic with no requirement for isolation. Operation is up to 500kHz, switching in less than 150ns, and the 2A drive will turn a 3000pF mosfet gate on in less than 25ns. Harris Semiconductor (UK) 0276 686886.

**PSU for notebooks.** Linear Technology has a 300mA DC-DC converter for notebooks, palmtops and other battery-powered equipment. From two AA batteres, the device will generate 5V at 150mA, or 300mA at 5V from a 9V battery. *LT1108* needs only an inductor, a diode and a capacitor and has a 1A power switch on-chip. Versions with fixed 5V or 12V or adjustable outputs are available. Linear Technology (UK) Ltd, 0276 677676.

**Power plus logic.** TI's Power+Logic family of devices contain cmos logic and power transistors in the same dip package and are claimed to give a 65% dissipation saving and up to a thousand times reduction in power consumption over bipolar peripheral drivers. Available devices are octal designs: *TPIC6529*, an 8-bit addressable latch; *TPIC6273*, a D-type flip-flop latch; and *TPIC6595*, an 8-bit shift register. All have low-side power mosfet output. Texas Instruments, 0234 223252.

# PASSIVE

## Passive components

**High-power, non-reactive R.** Power Film Systems has a range of high-power, non-reactive resistors in air-cooled (CP series) and water-cooled (A series) versions. Both types have alumina ceramic bodies and glazed film elements. Power rating of C types in still air at 40°C ambient is between 10W and 1kW, with a temperature coefficient of 300ppm. The A series offer a coefficient of 100ppm. AMS Electronic Ltd, 0803 200655.

**SM inductor.** The *ACCU-L* surface-mounted inductor from AVX, meant for the cellular communications sector and for the T&M industry, is in a standard 0805 low-profile package, taped for automatic placing. Q values at 2GHz are up to 55, with self resonance between 2.2GHz and 10GHz, inductance range being 2.7nH-15nH. AVX Ltd, 0252 336868.

**Chip thermistors.** Ten components in Murata's *NTH5G* range of chip thermistors operate at temperatures from –40 to 125°C, the range of 25°C resistances being 3.3kΩ-68kΩ. Normal tolerance is 10%, but there is a 5% option. All have a maximum power dissipation of 210mW. Murata Electronics (UK) Ltd, 0252 811666.

**EMI suppression.** Three-terminal capacitors by Murata, meant for interference suppression, have moulded-in ferrite beads to shorten the centre lead and increase filter performance and reduce off-board height. Components in the DSS family cover the range 47pF-22nF. Surtech Interconnection Ltd, 0256 51221.

**Chip capacitors.** Multilayer chip capacitors by TDK in the 16V *Hi-Cap* series have values up to 4.7μF. These ceramic devices use 10μm layers and are packaged in 0402-1206 cases, using X7R, X7S or Y5V dielectrics, the X7R/S types offering greater temperature stability. TDK UK Ltd, 0737 772323.

**Metal-film resistors.** Three ranges of metal-film resistors from Vishay offer temperature coefficients of 15ppm/°C (VY55), 25ppm/°C (VE55) and 50ppm/°C (VC55). Tolerances available are 0.05%, 0.1% and 0.5%. The epoxy encapsulation withstands immersion in solder to 300°C. Vishay Components (UK) Ltd, 0915 144155.

## Displays

**LCD graphics.** Three LCD graphics modules by Hitachi, meant for the handheld computer market, provide a 320 by 240 resolution, which is equivalent to a quarter VGA, in a slim package weighing 200g. Colours available are blue on grey, blue on white and black on white. Two of the modules are back-lit by fluorescent tubes. Hitachi claims that the displays "seriously challenge CRTs and other technologies in a wide range of applications". Hitachi Europe Ltd, 0628 585000.

## Filters

**Microwave filters.** Frequency-selective limiting filters from GEC-Marconi provide selective limiting of high-level signals without affecting coincident small signals at other frequencies in the pass band. Bandwidth is up to 1GHz in the 1.5-4GHz range, frequency selectivity being better tha 10MHz for up to 30dB compression of high-level signals. GEC-Marconi Research, 0245 73331.

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**Filter modules.** Kemo's 1600 series of filter modules is meant for data-acquisition use, and is digitally programmable to cut off over a 255:1 frequency range, three models now available cutting off from 2Hz to 51kHz. There is low variation of DC offset with programmed cut-off frequency and no high-frequency clock breakthrough or input aliasing, the filters being continuous-time designs. Low, high, band-pass and band-stop types are available and there is an evaluation board. Kemo Ltd, 081 658 3838.

## Instrumentation

**Pattern generator.** *GV-698* is Promax's multi-standard video pattern generator, which produces test patterns for pal, Secam and NTSC broadcast systems, any of the formats within these systems being available in the one instrument. The instrument produces eight pages of teletext and Nicam signals. A carrier is tuned by keyboard, manual tuning or by channel number to a resolution of 50kHz. Alban Electronics Ltd, 0727 832266.

**PC frequency generator.** Smooth transitions between all the frequencies needed to run PCs from 286 to 486 turbo are handled by the Integrated Circuit Systems *ICS2655* desk-top instrument. Output is up to 135MHz and consists of clocks for general comms, keyboard, floppy, system reference, bus and CPU. Amega Technology, 0256 330301.

**Sig. gen for modulation analyser.** Rohde & Schwarz's FMA modulation analyser family is augmented by the AM/FM calibrator/AF generator, which supplies AM or FM signals for calibration and an unmodulated 10MHz level calibrator. Outputsinclude single and two-tone signals, stereo multiplex and VOR/ILS/Tacan baseband signals. Together, the two form a complete transmitter test set for broadcast, communications and avionics. Rohde & Schwarz UK Ltd, 0252 811377.

**Large display.** Lascar's *DMX908* 7.92mm dot-matrix liquid-crystal display is an eight-character, two-line type offering 0.5mA consumption (50mA with back lighting) , a low profile, high contrast and a wide viewing angle. Interface is standard, being IDC header. Lascar Electronics Ltd, 0794 884567.


## Literature

**D-to-As for frequency synthesis.** Analog Devices has a 12-page application note on the choice and use of analogue-to-digital converters in direct digital frequency synthesis. The note goes into the background of the subject before discussing the relevant types of A-to-D converter. Analog Devices, 0932 253320.

**Crystals and oscillators.** The 1993 McKnight Fordahl catalogue is now available, being, in effect, a reference manual on frequency-generating devices. It is free on request. McKnight Fordahl Ltd, 0703 848961.

**Data conversion.** MicronetWorks conversion products catalogue is now available, containing data on over 75 families in D-to-A and A-to-D devices and data acquisition. Unitrode (UK) Ltd, 081 318 1431.

## Power supplies

**DC/DC converter.** From a wide-tolerance 5V input, Power General's *HDU1-35* DC-to-DC converter provides 3.3-5V or 2-5V power for mixed-voltage systems. A single output is selected to provide 2Vor 3.3V, 24W or 33W. Case size is 2in square and .0825in in height. It offers remote shut-down, continuous short protection and a pi filter internally for low EMI. Stabilisation and regulation are both 1%. Gresham Power Electronics Ltd, 0722 413060.

**Miniature 20MHz oscilloscope.** Hitachi's *V209* oscilloscope measures 215 by 110 by 350mm, but has many of the features of a standard bench instrument. The display is 63.5 by 50.8mm and sensitivity 5mV/division (1mV with multiplication) and the timebase speed is from 0.5μs/division to 0.2s/division. The two channels may be chopped, added or alternated or used in X/Y form. Thurlby Thandar Instruments, 0480 412451.

## Radio communications products

**900MHz power module.**Two modules from Motorola in the *MHW804* series operate in the 806-940MHz frequency range, accepting a 1mW input and producing an output of 4W. Power needed is at 7.5V. *MHW804-1* works between 800-870MHz, while MHW804-2 is an 896-940MHz device. Motorola Inc., (USA) (602)244-3818.

**GaAs RF power.** Motorola has a range of gallium arsenide RF power modules for use in cellular telephones. These surface-mounted devices provide a minimum gain of 26.5dB on an RF input of 5mW to give 1.4W output. Over the four bands, efficiency is 60% and noise is -95dB maximum. Motorola Ltd, 0908 614614.

## Switches and relays

**Mos relays.** Optical mos relays from OKI in the *OCMXXX* range use a GaAs led, and fet input/output, offering 400V switching and up to 4kV isolation. Both DC and AC/DC types are on offer. OKI Electric Europe GmbH, Germany 213115960.

## Transducers and sensors

**Displacement transducer.** At a maximum rotation speed of 150rpm, Vishay's ECO rotary displacement transducers have a life of over 10 million cycles. Depending on the model, linearity is from 0.25% to 2% and the servo or bush-mounted devices are available in four models of size 9 with a diameter of 22.2mm. Vishay Components (UK) Ltd, 0915 144155.

# COMPUTER

## Computer board level products

**Audio for virtual reality.** The *Convolvotron* from Crystal River allows the addition of "3-D" audio to virtual-reality applications, using headphones. Several different modes can be used: up to four independent sound sources can be positioned in an anechoic environment; a single source with reflections from six surfaces are simulated; and further PC cards work in parallel to support more sources. Inputs are four synchronised 16-bit A-to-D converters sampling at 50kHz and outputs are

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two independent 16-bit D-to-A converters for conventional stereo. Division Ltd, 0454 615554.

## Software

**Design on 386/486 PCs.** *Schematic Design Tools 386+* from Orcad is a true protected-mode design tool using 32-bit addressing and data structures. It is compatible with all current Orcad products, and existing STD designs and libraries are translatable into STD386+ format. ARS Microsystems Ltd, 0256 381400.

**CAD translators.** Instead of the normally rather messy methods adopted to use files generated by a given cad system in a different environment, the *RSI-Translator*'s range of intelligent database translators accept cad files in the one format, interpret and assign all the information and write the data out in a different format. Software up-grades are accepted. Betronex Ltd, 0920 469131.

**Half-bridge simulation.** *HB-SIM* by Design Automation is intended to ease the design of half-bridge (Class D) power amplifiers or power converters. The prgram simulates steady-state periodic time-domain waveforms up to 1000 times faster than *Spice*, or half a second on a 33MHz 486 PC. Also presented are power output, power input and power loss in each component, and harmonic spectra. Any parameter can be swept, the plot of swept parameter and result being drawn. Design Automation Inc., (USA) 617 862-8998.

**PC multimedia.** *PictureBook 2, MicroEye 2C video-in-VGA* and *TV/teletext* capture are cards and software to enable a PC to control and use a wide variety of video and teletext information, and to compile "electronic books". MicroEye 2C merges live video from cameras, video or Canon Ion with VGA displays in a window up to full size. TV1is a teletext decoder card to provide a *Windows* application with teletext, and *PictureBook 2* is a *Windows 3.1*-compatible system for the creation of books from a variety of inputs, including live video. Digithurst Ltd, 0763 242955.

**Heat-transfer computation.** *HotBox* from CHAM is software to assist designers to plan and check cooling strategy in electronic equipment. Solving fluid-flow and heat-transfer equations, *HotBox* calculates flow patterns and temperature distribution, displaying them in an easily interpreted manner. It runs on most computers, including the PC. Concentration, Heat, Momentum, 081-947 7651.

**Windows accelerator.** A Windows accelerator card running at 800 by 600 pixels and a 16-bit palette is introduced by Surtech. *Sprinter 2* has 1Mbyte of ram, is powered by the Weitek *5186* processor and accelerates Windows activity by up to 25 times. It supports 286, 386 or 486 dos and is supplied with *Windows 3.0* and *3.1* drivers and high-resolution dos drivers for *Lotus 1-2-3, AutoCad, Word, Worperfect* and *Ventura Publisher*. Surtech Interconnection Ltd, 0256 51221.

# Diodes in temperature measurement

While almost any silicon diode is usable as temperature-measurement transducer, giving about –2mV/°C slope in a reasonably linear manner, Motorola's *MTS102* silicon temperature sensor is designed for the job in automotive, industrial and consumer use. It provides higher accuracy at ±2°C from –40°C to 150°C and comes in a TO-92 package. The electronics needed to make a complete system consists of diode excitation, offsetting and amplification and Burr-Brown's application bulletin *AB-036* discusses these requirements in detail.

**Figure 1** is the circuit diagram of the simplest type of system. A current source for exciting the diode gives the best results, since resistor bias suffers from the effects of power-supply variation, particularly in low-voltage, single-rail circuits. Fortunately, Burr-Brown's *REF200* dual 100μA current source/sink perfectly matches the *MTS102*, which is specified for 100μA working. In the circuit of Fig.1, one of the sources excites the diode and the other supplies offset current.

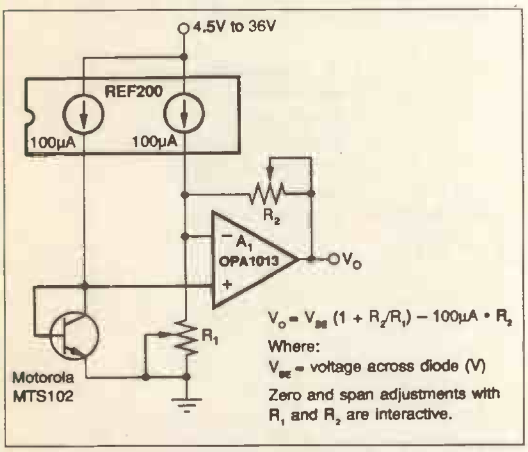


*Fig. 1. Half a Burr-Brown OPA1013 dual op-amp, a B-B REF200 current reference and a Motorola MTS102 silicon temperature sensor make up this basic temperature measurement circuit.*

Choice of op-amp for signal conditioning is not difficult; any precision type is usable, but B-B recommends its *OPA177* low-cost type for ±15V supplies and the *OPA1013* dual single-supply device for 5V working. Its inputs common-mode to ground and the output reaches to within 15mV of the supply voltage. There are two disadvantages to this basic circuit: span and zero adjustments are interactive and the output decreases for increasing temperature. If the output goes to a digital system, neither of these should be important.

Transfer function is

$$V_0 = V_{BE}(1 + R_2/R_1) - 100\mu A \times R_2,$$

$R_1$ is

$$\frac{(\delta V_0/\delta T)\times(V_{BE25} + T_C \times (T_{MIN} - 25^\circ C)) - (T_C \times V_1)}{100\mu A \times ((\delta V_0/\delta T) - T_C)},$$

and $R_2$ is

$$R_2 = R_1 \times \left(\frac{(\delta V_0/\delta T)}{T_C} - 1\right),$$

where $V_{BE25}$ is the diode voltage at 25°C, $V_1$ is the $T_{MIN}$ output voltage, $T_C$ is the diode temperature coefficient and $T_{MIN}$ is the minimum process temperature. $V_{BE25}$ lies between 0.580 and 0.620 for $T_C$ between –2.315 and –2.183 in mV/°C.

The circuit of **Fig. 2** affords independent adjustment of span and zero by virtue of

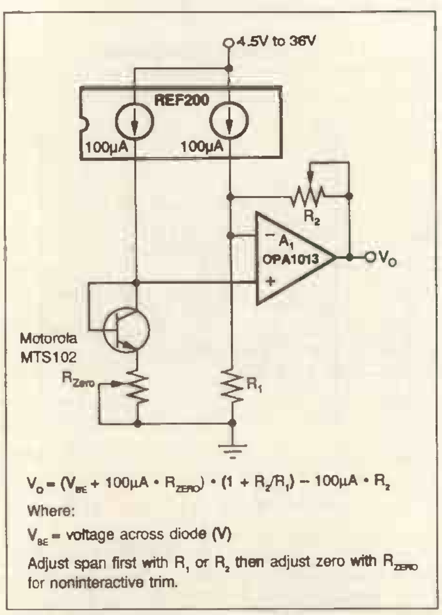


*Fig. 2. Addition of $R_{zero}$ avoids interaction of span and zero controls.*

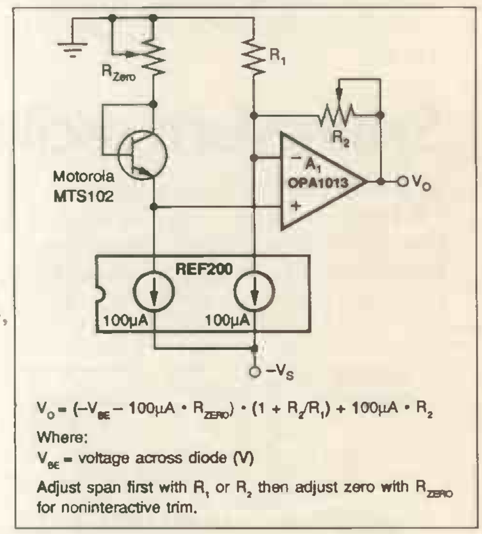


*Fig. 3. If temperature-to-voltage inversion is a problem, this arrangement avoids it, but needs a negative power supply line.*

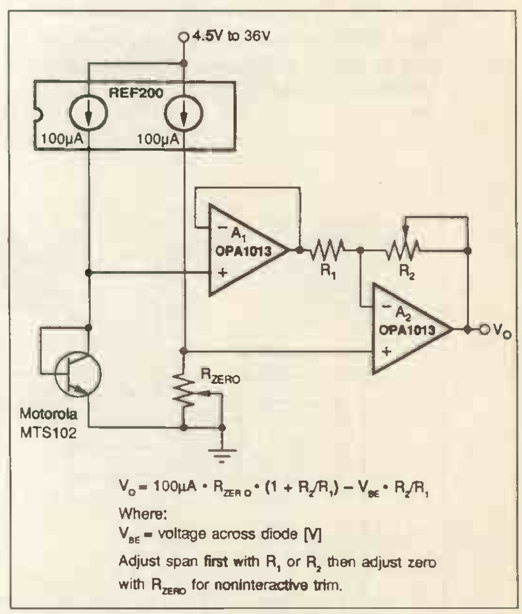


*Fig. 4. Combination of Figs. 2 and 3 gives independent control of span and zero and non-inverting output, but uses a single-ended positive supply.*

$R_{zero}$ in series with the diode. A method of calculating component values is given in the bulletin.

This circuit also has the possible drawback that its output is inverting, which brings us to **Fig. 3**, in which the temperature-to-voltage conversion is positive, although a negative power supply line is needed. Otherwise, the circuit is the same as that of Fig. 2, with the diode and reference reversed.

To obtain the best of both worlds, the circuit in **Fig. 4** is a non-inverting, single-supply design. The sensor output goes to the inverting op-amp input, the buffer $A_1$ preventing sensor loading.

Finally, the arrangement of **Fig. 5** measures the differential temperature between two sensor diodes.

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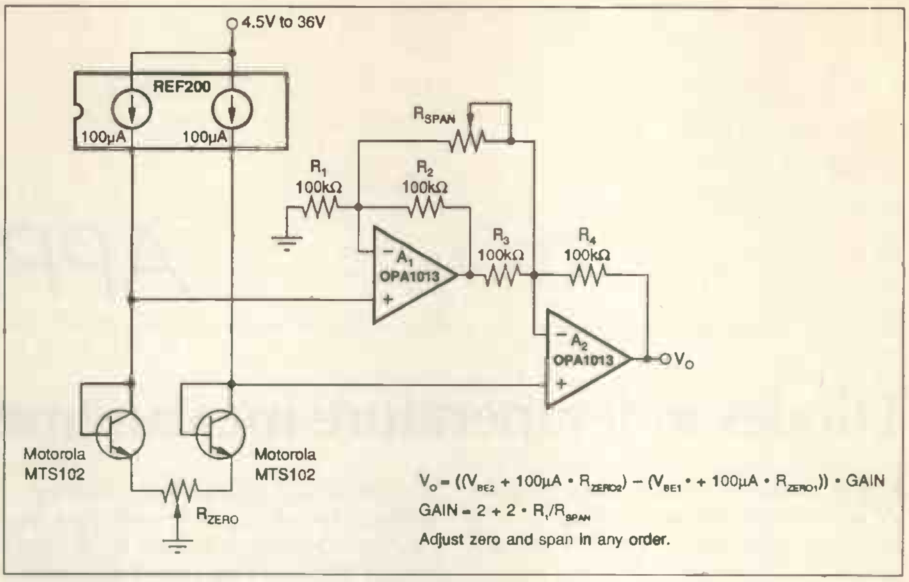



*Fig. 5. Differential circuit using both op-amps in the OPA1013.*

# Synthesised oscillators for radio

Motorola's *MC145170* is a frequency synthesiser with a very wide range — from a few hertz to 160MHz. Application note *AN1207* gives enough information to allow the design of two oscillators: an HF type for use at 9.2MHz and 12.19 MHz, and a VHF oscillator for up to 160MHz (or rather less at temperatures over 85°C).

**Figure 1** is the basic configuration: the *MC145170*, a filter and the voltage-controlled oscillator, plus an output buffer, the contents of the Motorola chip being indicated in **Fig. 2**. Operation is as normal; the reference oscillator and the VCO feed the multiplying and dividing counters, their

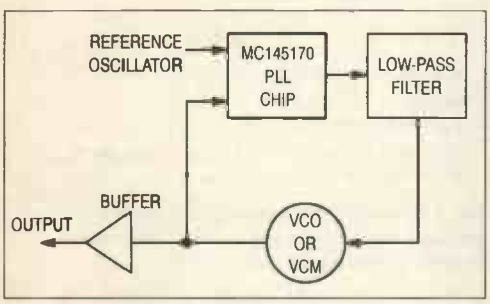



*Fig. 1. Basic block diagram of a phase-locked-loop frequency synthesiser, in this case using Motorola's MC145170 PLL IC and a low-pass filter.*

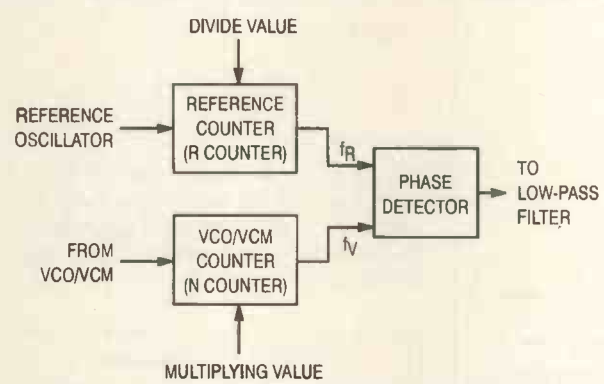



*Fig.2. Essential constituents of the Motorola chip, which also has the reference oscillator and the divide and multiply counters.*

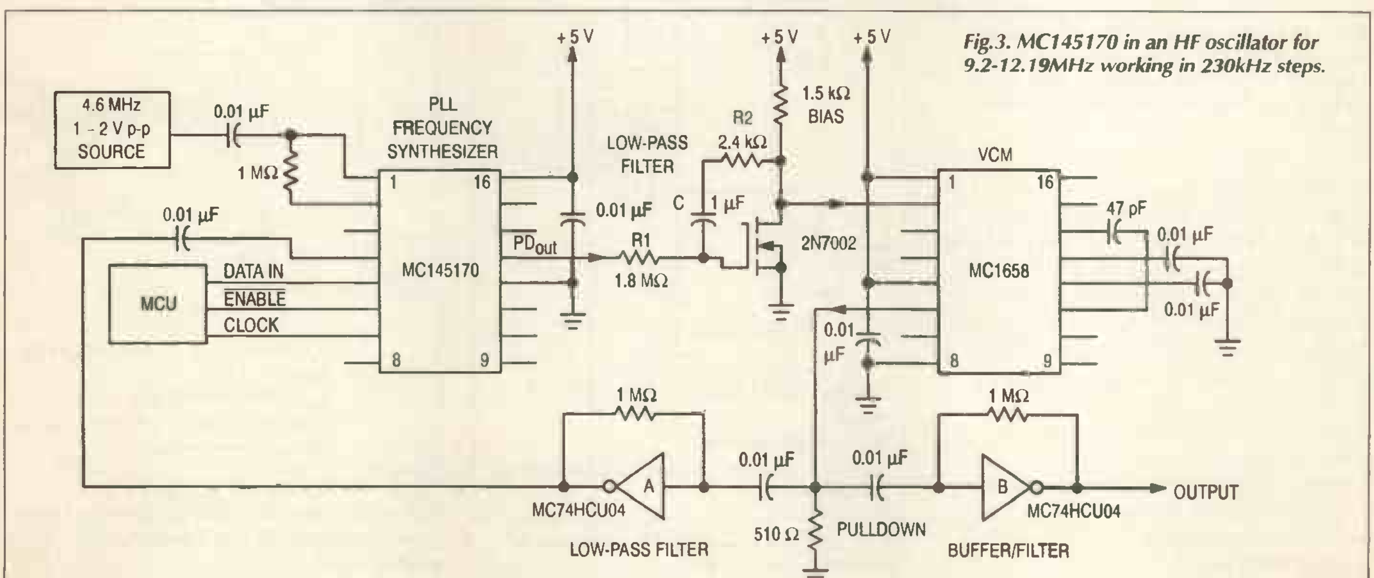



*Fig.3. MC145170 in an HF oscillator for 9.2-12.19MHz working in 230kHz steps.*

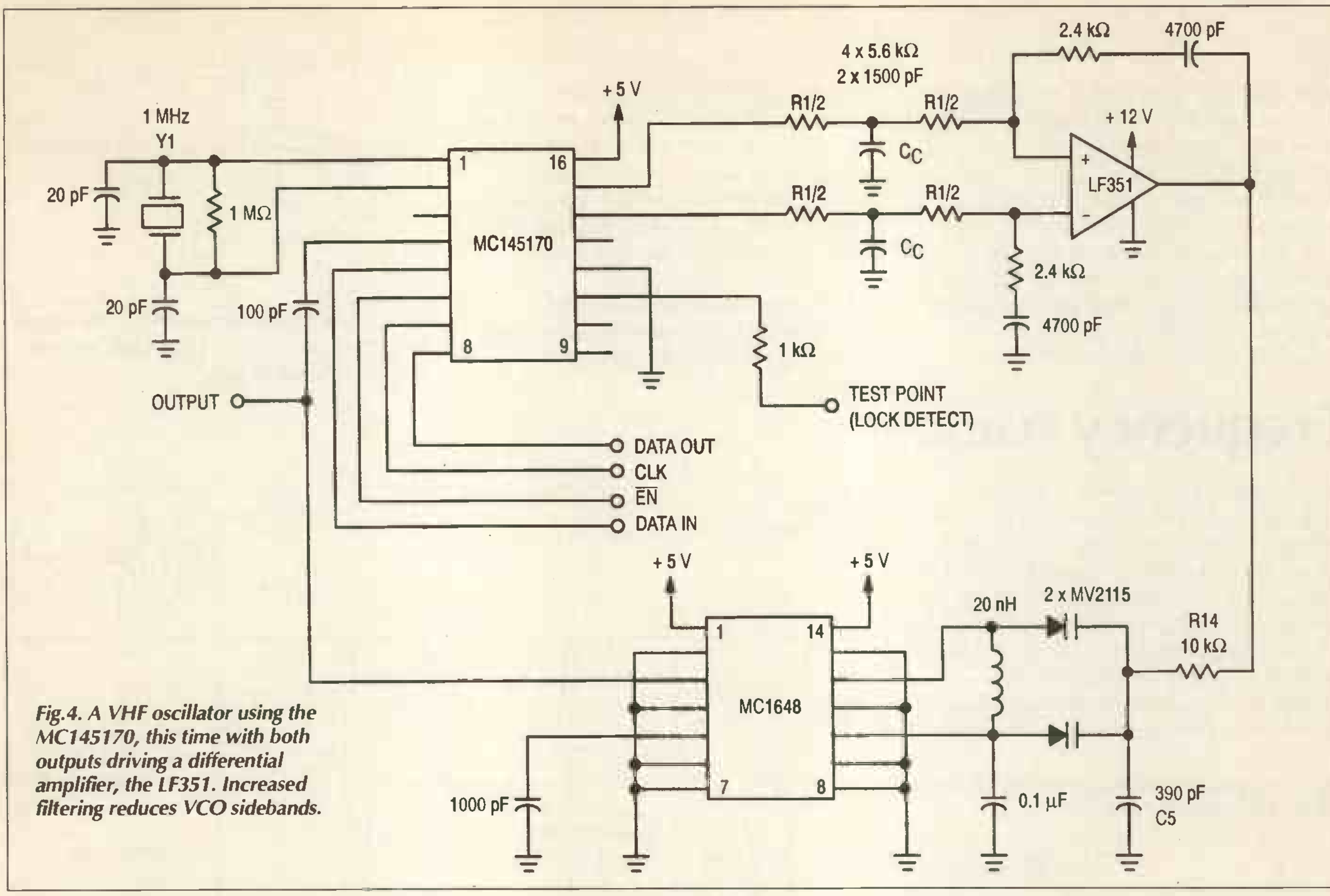



*Fig.4. A VHF oscillator using the MC145170, this time with both outputs driving a differential amplifier, the LF351. Increased filtering reduces VCO sidebands.*

outputs being compared in the phase detector. Either a sine oscillator or a voltage-controlled multivibrator followed by an integrator can be used.

**Figure** 3 is the circuit of the HF synthesiser, which has a resolution of 230kHz, a lock time of 8ms and overshoot a maximum of 15%. In this case, a square-wave output is acceptable and an *MC1658* voltage-controlled multivibrator can be used; since its input loading is fairly large at 350µA maximum, the fet in the active filter avoids filter response degradation. Resistor $R_1$ and $C$ are the filter components and the application note provides a method of component calculation.

To program the circuit, three registers must be programmed: the *C* register, which configures the device, setting the phase detector to the correct polarity, turning off unused outputs and activating the phase-detector output; the *R* register for the divider providing the phase detector reference; and the *N* register, which sets the tuned frequency. Both *C* and *R* registers are programmed once on power-up.

The buffer A cleans up the *MC1658* output before feeding it back to the synthesiser, since any spurious signal would cause miscounting in the *N* counter. A buffer's response is low enough to perform the filtering.

**Figure 4** is the VHF synthesiser, which has a range of 140-160MHz in 100kHz steps and uses both $f_R$ and $f_V$ outputs of the *MC145170*; hence the op-amp. Filter calculations are again given in the note. In this case, the filtering is enhanced to avoid a small amount of reference frequency feeding through the filter and giving rise to larger VCO sidebands. Enhancements consist of the op-amp feedback components together with the split $R_{1/2}$ and $C_C$, and the input circuitry to the *MC1648* VCO – $R_{14}$ and $C_5$.

***Motorola Ltd,** European Literature Centre, 88 Tanners Drive, Blakelands, Milton Keynes MK14 5BP.*

# More gain from the SL6140

GEC Plessey's *SL6140* is broadband amplifier IC with AGC, providing 15dB of linear gain into 50Ω at 400MHz, or over 45dB into 1kΩ. The AGC controls gain over a 70dB range by voltage applied to an external series resistor. Bandwidth depends on the load, being 25-400MHz for loads of 1kΩ-50Ω.

Application note *AN45* describes a method of tuning input and output to match a 50Ω source and load and to give increased gain over a smaller bandwidth. **Figure 1** is the circuit diagram of a single-ended 100MHz amplifier having 35dB power gain.

At the input, a parallel tuned circuit across the differential inputs has the signal applied to one end via the coupling capacitor, the

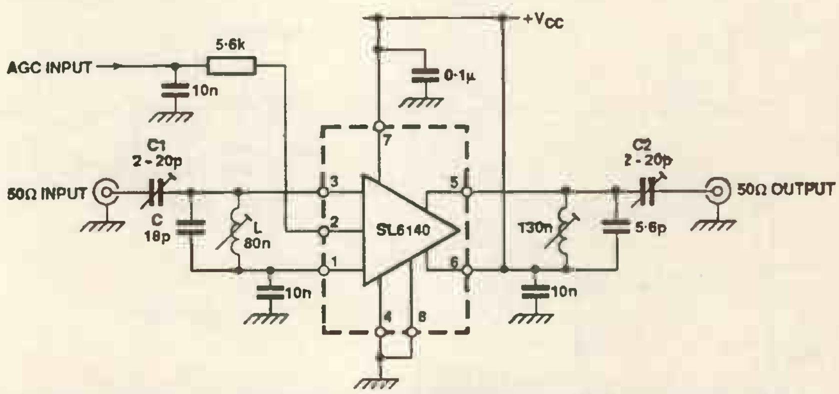



*Fig. 1. Tuning input and output of GEC Plessey's SL6140 wide-band amplifier increases its stage gain from 15dB to 35dB, albeit with reduced bandwidth. This is a 100MHz amplifier matched to 50Ω in and out.*

other end being decoupled. Since capacitor $C_1$ forms part of the matching network, the tuned frequency is given by $f=1/[2\pi(\sqrt{L*C*C_1})/\sqrt{(C+C_1)}]$.

Output matching is done by connecting a parallel tuned circuit from one of the open-collector amplifier outputs to $V_{CC}$, the coupling capacitor again forming a part of the matching network from a high amplifier impedance to the 50Ω load. Coupling capacitors $C_{1,2}$ are adjustable to set maximum gain, short of oscillation caused by too high impedances at input and output.

Since only one amplifier output is used in this circuit, an improvement is brought about by transformer coupling to the output, as shown in **Fig. 2**, in which both are in use to effect a further 6dB increase in gain.

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

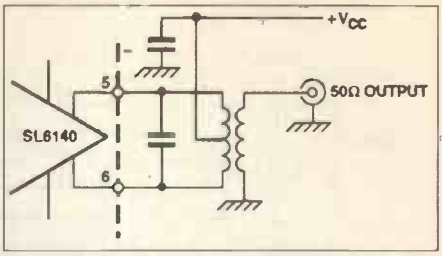
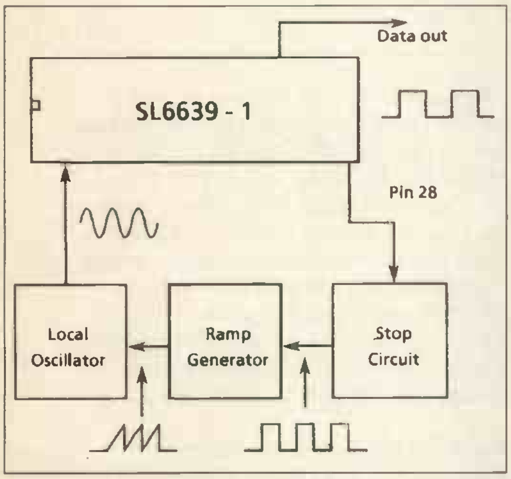



***Fig. 2. Using transformer coupling at the output brings the other amplifier output into play and gives another 6dB of gain.***

# Frequency scanner

GEC Plessey's *SL6639-1* is a direct-conversion FSK data receiver for up to 200MHz working and featuring extremely low power consumption –about 3.7mW. It is intended for use in pagers, direction indicators, security systems and remote control. Application Note *AN96* describes a scanning system to be used with the *SL6639-1*, which scans the local oscillator, detects a transmission and the received data and goes on to scan the rest of the band. The system needs no expensive components, runs at audio frequencies and requires only minor changes to the *SL6639-1* demo board.

***Fig. 1. Block diagram of scanner system for GEC Plessey's SL6639-1 FM data receiver.***



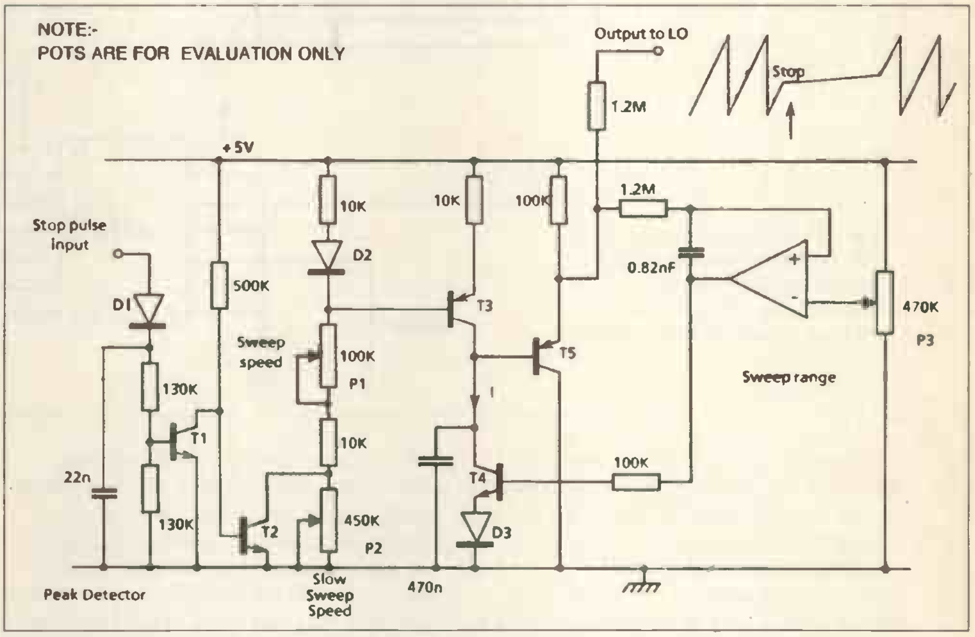




***Fig. 2. Ramp generator. Ramp rate set by P1, rate in slow section by $P_2$. Pulses from stop circuit detected and used to switch capacitor charging current to low level, slowing the ramp.***

**Figure 1** is the system diagram, which shows that a ramp generator scans the local oscillator through the required band. The appearance of a signal from the receiver test output to the stop circuit causes a series of pulses which are used to slow the ramp rate, so that the scanning slows to a rate, settable by $P_2$ in **Fig. 2**, low enough to allow the receiver to detect the correct number of bits.

Stop pulses come from the circuit of Fig. 3, which consists of a high-impedance input amplifier, a gyrator simulating a 1H inductance to resonate at 1kHz, and a comparator. When the output of the LC tuned circuit exceeds the threshold set on the comparator, pulses are produced. All the op-amps are contained in one *TAB1043*.

A constant-current source, $D_2$ and $T_3$ in Fig. 3, feeds a capacitor to produce the ramp waveform. Potentiometer $P_3$ sets a volatge on the comparator input which, when exceeded by the ramp, generates a reset pulse at the comparator output to switch on $T_4$ and take the ramp back to zero. When stop pulses are detected by $D_1$, $T_1$ switches $T_2$ off, reducing capacitor current and

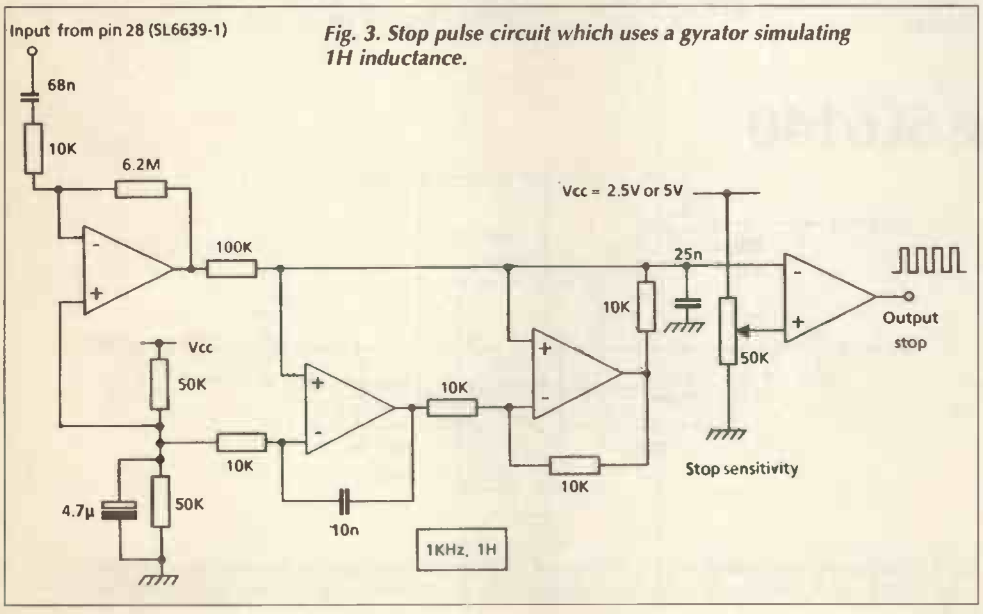



***Fig. 3. Stop pulse circuit which uses a gyrator simulating 1H inductance.***

slowing the ramp to a rate set by $P_2$.

**Figure** 4 is the local oscillator, a circuit designed for the *SL6639*, but with no crystal. It is a Colpitts type with a Varactor diode tuning control accepting the ramp input. ■

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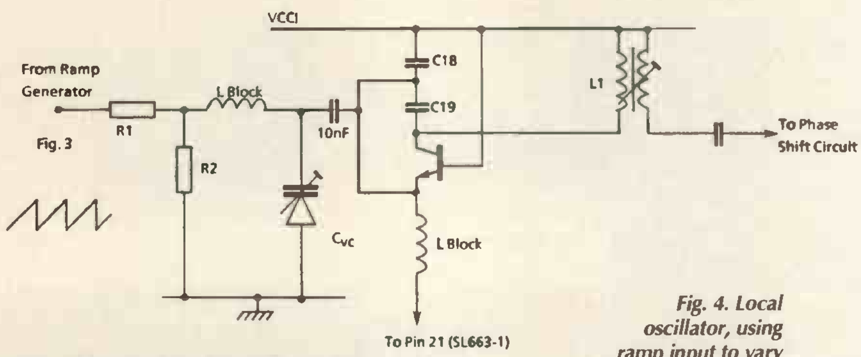


R1 and R2 are chosen for desired frequency and sweep range.

Varactor Diode Capacitance Appears in Parallel with C18.

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

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

*Fig. 4. Local oscillator, using ramp input to vary diode capacitance and therefore tuned frequency.*

## SL6639 OPERATION

The incoming signal is split into two parts and frequency converted to baseband. The two paths are produced in phase quadrature and detected in a phase detector which provides a digital output. The quadrature network may be in either the signal path or the local oscillator path.

The input to the system is an FSK data modulated signal with a modulation index of 18, $f_1$ and $f_0$ represent the "steady state" frequencies (ie modulated with continuous "1" and "0" respectively).

When the LO is at the nominal carrier frequency, then a continuous "0" or "1" will produce an audio frequency, at the output of the mixers corresponding to the difference between $f_0$ and $f_c$ or $f_1$ and $f_c$. If the LO is precisely at fc, then the resultant output signal will be at the same frequency regardless of the data state; nevertheless, the relative phases of the two paths will reverse between "0" and "1" states.

By applying the amplified outputs of the mixers to a phase discriminator, the digital data is reproduced.

# INSTRUMENTS TO BUY

*MX1010F MX1100F*

## FREQUENCY COUNTERS

**MX1010F** and **MX1100F** are 8-digit frequency counters offering a broad range of features.
**MX1010F**: 1Hz to 100MHz, sensitivity of 15mV and resolution to 0.1Hz, data auto set, 10:1 attenuator, high impedance input – £129.00 plus VAT (£151.58).
**MX1100F**: 1Hz to 1GHz, features as MX1010F except ranges 70MHz to 1GHz and 50Ω impedance. £160.00 plus VAT (£188.00).
**SC-130 and SC-40** are full featured, microprocessor-based, hand held frequency counters providing portability and high performance. Both instruments provide measurement of frequency, period, count and RPM plus a view facility enabling min, max, av and difference readings.
**SC-130:** 5Hz to 1.3GHz, 8 digit readout, sensitivity typically 10mV, high impedance input, battery condition indicator. £109.00 plus VAT (£128.08).
**SC-40:** As SC-130 except 5Hz to 400MHz. £89.00 plus VAT (£104.58).

*FG 506 FG513*

## MULTIMETERS

The **180 series** of high performance multimeters provide advanced features and are supplied complete with probes, battery and rubber holster. The case is dust and splash proof making it ideal in most environments. Designed to meet IEC348 Class II safety standard.
**183**: 3$^1/_2$ digit large LCD display, ACV, DCV, ACA, DCA, resistance, continuity buzzer, diode test, hold, basic accuracy 0.5%. £39.50 plus VAT (£46.41).
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*THE 180 SERIES*

## MULTIMETERS (2)

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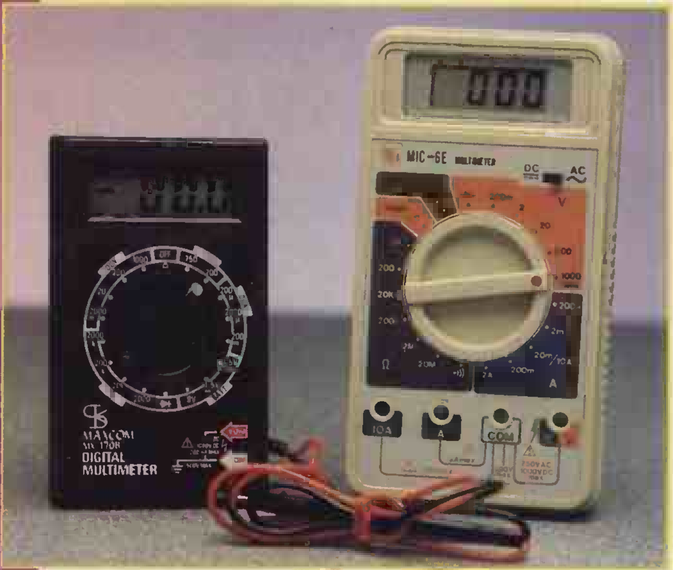
**MIC-6E**: 3$^1/_2$ digit LCD, ACV, DCV, ACA, DCA, resistance, diode test, buzzer.
£33.50 plus VAT (£39.36).




*MX170B MIC-6E*

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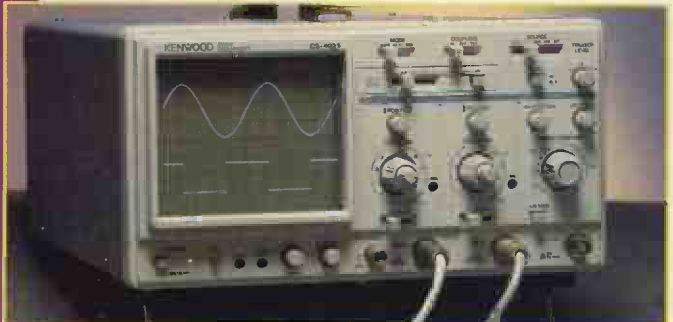


*20MHz 2-CH OSCILLOSCOPE*

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## FUNCTION GENERATOR

The **MX2020** 0.02Hz – 2MHz sweep function generator with LED digital display offers a broad range of features. Output waveforms include sine, square, triangle, skewed sine, pulse and TTL. Lin and log sweeps are standard as is symmetry, DC offset and switchable output impedance from 50Ω to 600Ω. The digital display provides readout of the generators' frequency or can operate as separate 10MHz frequency counter. £199.00 plus VAT (£233.83).

## LCR METER

The **MIC-4070D** LCD digital LCR meter provides capacitance, inductance, resistance and dissipation measurement. Capacitance ranges are from 0.1pF to 20,000µF plus dissipation. Inductance ranges from 0.1µH to 200H plus a digital readout of dissipation. Resistance ranges from 1mΩ to 20MΩ. Housed in a rugged ABS case with integral stand it is supplied complete with battery and probes at £85.00 plus VAT (£99.88)

## FOUR INSTRUMENTS IN ONE

The **MX9000** combines four instruments to suit a broad range of applications in both education and industrial markets including development work stations where space is at a premium.
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3. A 0.02Hz to 2MHz full featured sweep/function generator producing sine, square, triangle, skewed sine, pulse and a TTL output and linear or logarithmic sweep. Outputs of 50Ω and 600Ω impedance are standard features;
4. An auto/manual $3^1/_2$ digit LCD multimeter reading DCV, DCA, ACV, ACA, resistance, and relative measurement with data hold functions.

The MX9000 represents exceptionally good value at only £399.00 plus VAT (£468.83).

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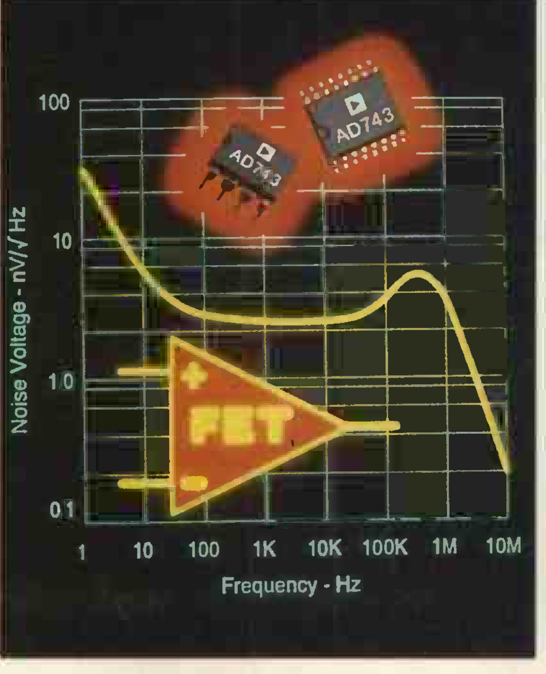
# OPERATIONAL AMPLIFIERS
## the supreme activators

*Although external components often determine an op amp's performance, the chips themselves are not always trouble-free: oscillations and noise represent two possible areas of difficulty.*

*Serialised from the book Troubleshooting Analog Circuits by Robert Pease.*

External components often determine an op amp's performance. But op amps aren't always absolutely trouble-free: Oscillations and noise are two possible areas of difficulty, among others. It is mostly the components around op amps that cause many of their problems. After all, the op amp is popular because external components define its gain and transfer characteristics.

So, if an amplifier's gain is wrong, you quickly learn that you should check the resistor tolerances, not the op amp. If you have an AC amplifier or filter or integrator whose response is wrong, you check the capacitors, not the op amp. If you see an oscillation, you check to see if there's an oscillation on the power-supply bus or an excessive amount of phase shift in the feedback circuit.




If the step response looks lousy, you check your scope or your probes or your signal generator because they're as likely to be the problem as the op amp is. The overall performance of your circuit is often determined by those passive components. And yet, there are exceptions. There are still a few ways an op amp itself can foul up.

Before we discuss serious problems, you should be aware of the kind of op-amp errors that aren't significant. First of all, it generally isn't reasonable to expect an op amp's gain to be linear, nor is its nonlinearity all that significant.

For example, what if an op amp's gain is 600,000 for positive signals but 900,000 for negative signals? That sounds pretty bad. Yet, this mismatch of gain slope causes a non linearities; of about 10μV in a $20V_{p\text{-}p}$ unity gain converter. The voltage coefficients and temperature coefficient errors of the feedback resistors will cause a lot more error than that. Even the best film resistors have a voltage coefficient of 0.1ppm/V, which will cause more non linearity than this gain error.

Recently I heard a foolish fellow argue that an op amp with a high DC gain such as 2,000,000 or 5,000,000 has no advantage over an amplifier with a DC gain of 300,000 because, unless your signal frequency is lower than 0.1 Hz, you cannot take advantage of this high gain. Obviously, I don't agree with that. If you have a step signal, the output settles to the precise correct value in less than a millisecond – not 1s or more. The amplifier with the higher gain settles to a more precise value. It does not take any more time. I guess he just doesn't understand how op amps work.

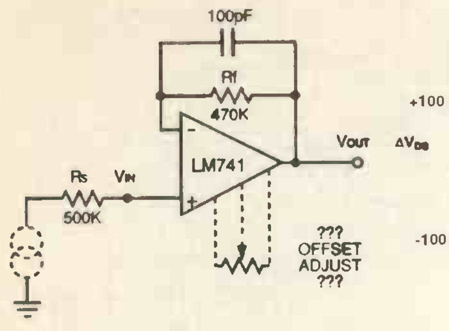



*Fig. 1. If you run an op amp at such high impedances that $I_B$ x R is more than 20mV, you'll be generating big errors, and a $V_{offset}$ trim pot can't help you cancel them out. Please don't even try!*

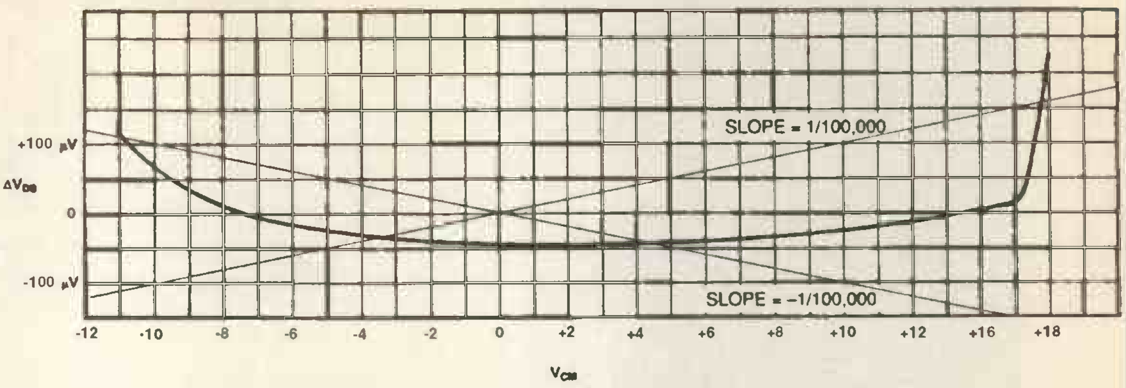



*Fig. 2. The CMRR of an op amp can't be represented by a single number. It makes more sense to look at the CMRR curve, $\Delta V_{OS}$ versus $\Delta V_{CM}$, and note its nonlinearities, compared to a straight line with a constant slope of l part in 100,000.*

Especially since he doesn't even want to talk about gain nonlinearity!

Many old amplifiers had low DC gain and poor gain linearity whereas more modern devices like the NSC *OP-07* and the *LM607* (gain = 6,000,000 min) have much less.

Similarly, an op amp may have an offset-voltage temperature coefficient specification of 1µV/°C, but the op amp's drift may actually be 0.33µV/°C at some temperatures and 1.2µV/°C at others. Twenty or thirty years ago, battles and wars were fought over this kind of specsmanship, but these days, most engineers agree that you don't need to sweat the small stuff. Most applications don't require an offset drift less than 0.98µV for each and every degree: most cases are quite happy when a 1µV/°C op amp drifts less than 49µV over 50°C.

Also, you don't often need to worry about bias current and its temperature coefficient, or the gain error's TC. If the errors are well behaved and fit inside a small box, well, that's a pretty good part.

There is one classical *caveat*. Namely: If you run an op amp in a high impedance circuit such that the bias current causes significant errors when it flows through the input and feedback resistors, do not use the $V_{os}$ pot to get the circuit's output to zero. Example, if you have an LM741 as a unity gain follower with a source impedance of 500kΩ and a feedback resistor of 470kΩ, the 741's offset current of 200nA (worst-case) could cause an output offset of 100mV. If you try to use the $V_{os}$ trim pot to trim out that error, it won't be able to do it. See **Fig. 1**.

If you have only 20 or 40mV of this *I* x *R* error, you may be able to trim it out, but the TC and stability will be lousy. So you should be aware that in any case where the $I_{os}$ x *R* is more than a few millivolts, you have a potential for bad DC error, and there's hardly any way to trim out the errors without causing other errors.

When you get a case like this, unless you are willing to accept a crude error, then you ought to be using a better op amp with lower bias currents.

## An uncommon mode

A good example of misconstrued specs is the common mode error. We often speak of an op amp as having about CMRR (common mode rejection ratio) of 100dB. Does this number mean that the common mode error is exactly one part in 100,000 with a nice linear error of 10µV per volt? Well, this performance is possible, but not likely. It' s more likely that the offset voltage error as a function of common mode voltage is nonlinear. In some regions, the slope of $\Delta V_{os}$ will be much better than 1 part in 100,000. In other regions, it may be worse.

It really bugs me when people say "The op amp has a common mode gain, $A_{VC}$, and differential gain, $A_{VD}$, and that the CMRR is the ratio of the two. This statement is silly: It's not reasonable to say that the op amp has a differential gain or common mode gain that can be represented by a single number. Neither of these gain numbers could ever be observed or measured with any precision or repeatability on any modem op amp. Avoid the absurdity of trying to measure a "common mode gain of zero" to compute that your CMRR slope is infinite. You'll get more meaningful results if you just measure the change in offset voltage, $V_{OS}$, as a function of common mode voltage,$V_{CM}$ and observe the linear and nonlinear parts of the curve of the sort shown in **Fig. 2**.

## How not to test for CMRR

First thing to remember is how *not* to measure CMRR. In **Fig 3**, driving a sine wave or triangle wave into point *A*, will make it seem as if the output error, as seen by a floating scope, will be (N+l) x [VCM/CMRR]. But that's not quite true: you will see (N+l) x [CM *Error* + Gain Error]. So, at moderate frequencies where the gain is rolling off and the CMRR is still high, it is mostly the gain error that will be seen, and the curve of CMRR vs frequency will look just as bad as the Bode plot.

An *LF356* run in the circuit of Fig. 3 gives an error of 4mVpp at 1kHz – a large quadrature error, 90° out of phase with the output (see the upper trace in **Fig. 4**.

If you think that is the CM error, you might say the CMRR is as low as 5000 at 1kHz, and falls rapidly as the frequency increases. The actual CMRR error is about 0.2 mV (see the lower trace of Fig. 6) and thus the CMRR is about 100,000 at 1kHz or any lower frequency.

Note also that, on this unit, the CM error is not really linear: as –9V is approached, the error becomes more nonlinear. (This is a

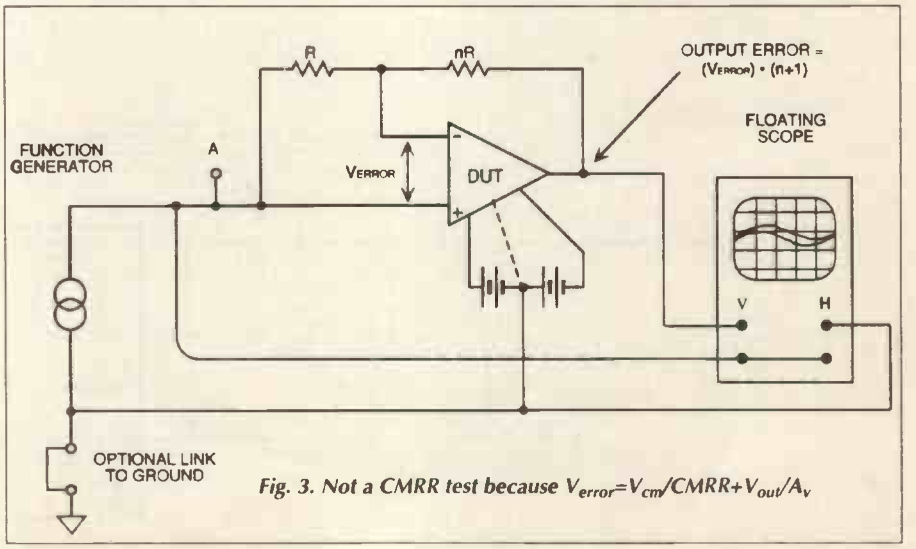



*Fig. 3. Not a CMRR test because $V_{error}$=$V_{cm}$/CMRR+$V_{out}$/$A_v$*

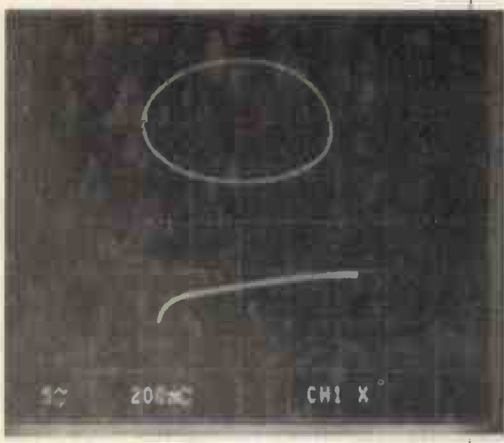



*Fig. 4. Upper trace shows the "CMRR" error taken using the circuit of Fig. 3. But it is not the CMRR error, it is really the gain error, 4mVpk-pk at 1kHz. Lower trace shows the actual common-mode error – about 1/20 the size of the gain error – measured using the circuit of Fig.7.*

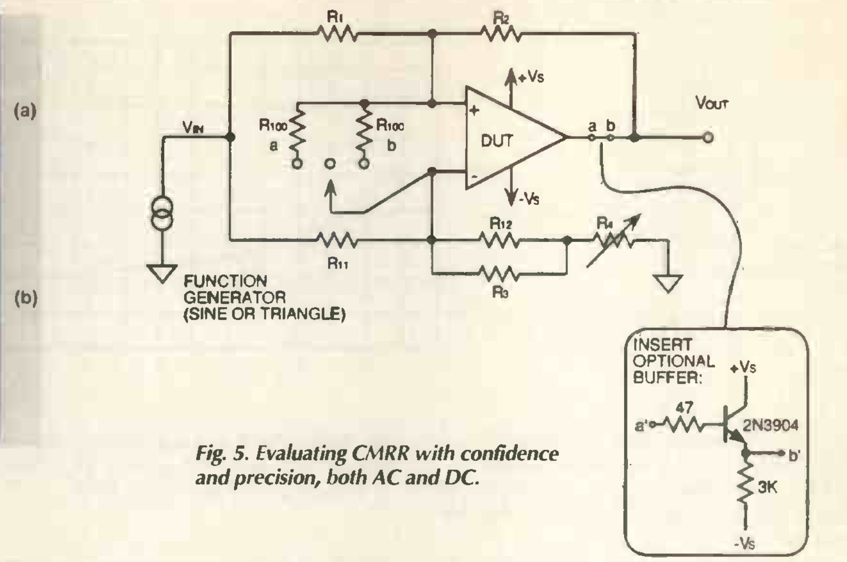



*Fig. 5. Evaluating CMRR with confidence and precision, both AC and DC.*

–9V/+12V CM range on a 12V supply; I chose a –12V supply so my function generator could overdrive the inputs.)

**How to test properly**

So how *can* we test for CMRR and get the right results? Fig. 5 is a fine circuit, even if it has limitations.

If $R_1 = R_{11} = 1k$, $R_2 = R_{12} = 10k$, and $R_3 = 200k$ and $R_4$ = a 500Ω pot, single-turn carbon or similar, the noise gain is defined as $1 + [R_f/R_{in}]$, or about 11. If we put a +11V sine wave into the signal input, the CM voltage is about –10V. The output error signal will be about 11 times the error voltage plus some function of the mismatch of all those resistors.

Connect the output to a scope in cross-plot (X-Y) mode and trim that pot until the output error is very small – until the slope is nominally flat. Whether or not the CMRR error is balanced out by the resistor error is unimportant. Just observe that the output error, as viewed on a cross-plot scope, is quite small.

Now connect in $R_{100}a$, a nice low value such as 200Ω. Computing the noise gain, it rises from 11 to 111. It *was* $(1 + R_2/R_1)$, and then increases to $(1 + R_2/R_1)$ *plus* $(R_2 + R_{12})/R_{100}$. In this example, that is an increase of 100. This means a *change* of $V_{out}$ equal to 100 times the input error voltage (and that is VCM divided by CMRR).

Of course, it is unlikely for this error voltage to be a linear function of VCM. So look at it with a scope in cross-plot (X-Y) mode. Too many people pretend that CMRR is constant at all levels and that CM error is a linear function of VCM. They just look at two points and assume every other voltage has a linear error.

Another good reason to use a scope in the XY mode is to allow visual subtraction of he noise. An AC voltmeter can certainly not be used to detect the CMRR error. For example, in Fig. 4, the CM error is fairly stated as $0.2mV_{p-p}$, not $0.3mV_{p-p}$ (as it might be if a meter that counted the noise was used ). A good amplifier with a CMRR of about 100dB will show the CM error to be about $200\mu V_{p-p}$; as this is magnified by 100, an output error of $20mV_{p-p}$ can easily be seen. With a really good unit having a CMRR of 120 or 140dB, the aim will be to clip in the $R_{100}b$ – eg at 20Ω – and then the Δ (noise gain) will be 1000. The noise will be magnified by 1000 – but so will the error.

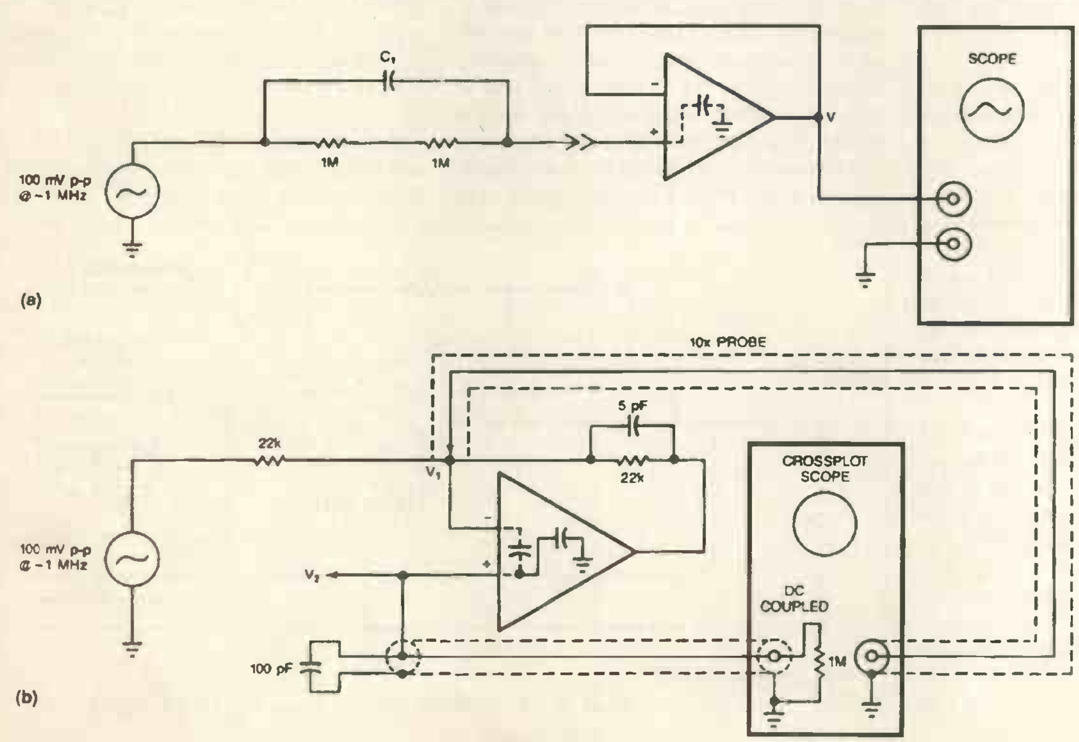



*Fig. 6. Circuit (a) allows testing of an op amp's common-mode input capacitance. When $C_1 = C_A \approx 5pF$, measure $V_1 = V_A$; when $C_1 = C_B \approx 1000pF$, measure $V_1 = V_B$. Then $C_{in} = C_A \times (V_B - V_A)/[V_A - V_B \times C_A/C_B]$. For best results, connect the signal to the plus input of the DUT with a small gator clip. Do not put the plus input pin into the device's socket. Circuit (b) allows testing of an op amp's differential input capacitance. For this circuit $C_{in(differential)} = V_2(p\text{-}p) \times Ct_{otal}/[V_1(p\text{-}p) - V_2(p\text{-}p)]$, where $C_{total} = C_{in(common\ mode)} + C_{cable} + C_{scope\ in} + 100pF$.*

This circuit provides a high resolution view and gives a good feel for what is happening, rather than just showing numbers. For example, with a $22mV_{p-p}$ output signal that is caused by a 22µV error signal, the CMRR really is way up near a million, which is much more useful than a cold "119.2dB" statement. It also teaches that the slope and the curvature of the display are important. Not all amplifiers with the same "119.2dB" of CMRR are actually the same; some have a positive slope, some a negative slope, and some curve, so that if a two-point measurement is taken, the slope changes wildly, depending on which two points are chosen. (Increasing the amplitude of the input signal makes it plain where severe distortion sets in – that is the extent of the common mode range.)

The limitation is that with the noise gain set as high as 100, then this circuit will be 3dB down at $F_{GBW}/100$. So only use this up to about 1kHz on an ordinary 1MHz op amp, and only up to 100Hz at a gain of 1000.

### Bias current

Another op amp spec not to worry about is the differential input impedance: instead, measure the bias current. There is a close correlation between the bias current and the input impedance of most op amps, so if the bias current is low enough, the input impedance (differential and common mode) must be high enough. Generally, an ordinary differential bipolar stage has a differential input impedance of $1/(20 \times I_b)$, where $I_b$ is the bias current. This number varies if the op amp includes emitter degeneration resistors or internal bias compensation circuitry.

Common mode input resistance can easily be tested by measuring $I_b$ as a function of $V_{CM}$. The circuits of **Fig. 6** are quite useful here. Input capacitance data is nominally of interest only for high impedance high speed buffers or for filters, to make sure that the second source device has the same capacitance as the op amps that are already working adequately.

### False error characteristics

Sometimes, an op amp may exhibit an "error" that looks like a bad problem, but isn't. For example, if an op amp's output is ramping at –0.3V/µs, it might be surprising to find that the inverting input, a summing point, is *not* at ground. Instead, it may be 15 or 30 or 100mV away from ground. How can the offset voltage be so bad if the spec is only 2 or 4mV?

Why is the inverting input not at the "virtual ground" that the books teach us?

The virtual ground theory is applicable at DC and low frequencies, but if the output is moving at a moderate or fast speed, then expecting the summing point to be exactly at ground is unreasonable. In this example, $dV_{out}/dt$ equals $2\pi \times$ the unity gain frequency $\times$ the input voltage. So 15mV of $V_{in}$ is quite reasonable for a medium bandwidth op amp, such as an *LF356*, and 50 or 70mV is quite reasonable for an *LM741*.

To make an op amp move its output at any significant speed, there has to be a significant error voltage across the inputs for at least a short time.

Also beware of op amp models and what they might mistakenly say. For instance, the "standard" equation for a single-pole op amp's gain is $A = Ao(\,1/1 + j\omega T)$, implying that when the DC gain $A_0$ changes, the high-frequency gain, $A$, changes likewise. But this is incorrect. There is almost no correlation between the high frequency response and the spread of DC gain, on any op amp. Several ways can be used to get an op amp's DC gain to change: change the temperature, add on or lift off a load resistor, or swap for an amplifier with higher or lower DC gain.

Although the DC gain can vary several octaves in any one of these cases, the gain-bandwidth product stays about the same. If there ever were any op amps whose responses did vary with the DC gain, they were abandoned many years ago as unacceptable.

Op amp spec sheets often give the open loop output impedance as 50Ω. But by inspecting the gain specs at two different load resistor values, the DC gain can be seen to fall by a factor of two when a load of 1kΩ is applied. If an op amp has an output impedance of 1kΩ, its gain will fall by a factor of two when a 1kΩ load is applied. But if its output impedance were 50Ω, as the spec sheet claimed, the gain would only fall 5%. So, whether it is a computer model or a real amplifier, be suspicious of output impedances that are claimed to be unrealistically low.

### Real trouble

What real trouble can an op amp cause? A part may have a bad $V_{OS}$. Or if the temperature is changing, the thermocouples of the op amp's Kovar leads may cause small voltage differences between the op amp leads and the copper of the PC board. Such differences can amount to 1/10 or 1/20 of a Celsius degree times 35µV/°C, equal to 2-3µV.

Remember, too, that not all op amps of any one type have the exact same output voltage swing or current drive or frequency response. Designers can fall into the habit of expecting parts to be better than average. When they receive parts that are still much better than the guaranteed spec but worse than average or "typical," they find themselves in trouble.

### Occasional oscillations

One of the most troublesome problems associated with op amps is oscillation. Just as an oscillator can be built out of any gain block, then it must be accepted that any gain block can also oscillate when not wanted. Op amps are no exception, though most op are well behaved, and only four basic precautions need to be taken to avoid oscillations.

- Always use some power-supply bypass capacitors on each supply and install them near the op amp. For high frequency op amps, the bypass capacitors should be very close to the device for best results. Ceramic *and* tantalum bypass capacitors are often needed.

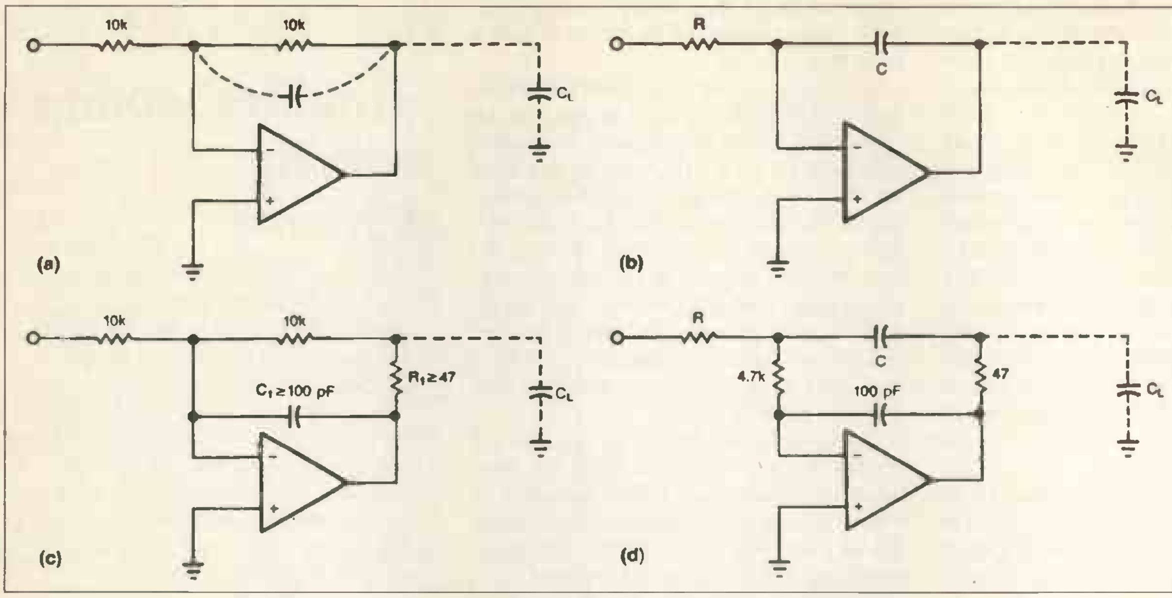



*Fig. 7. The basic inverter (a) and integrator (c) can be easily modified to decouple capacitive loads (b) and (d).*

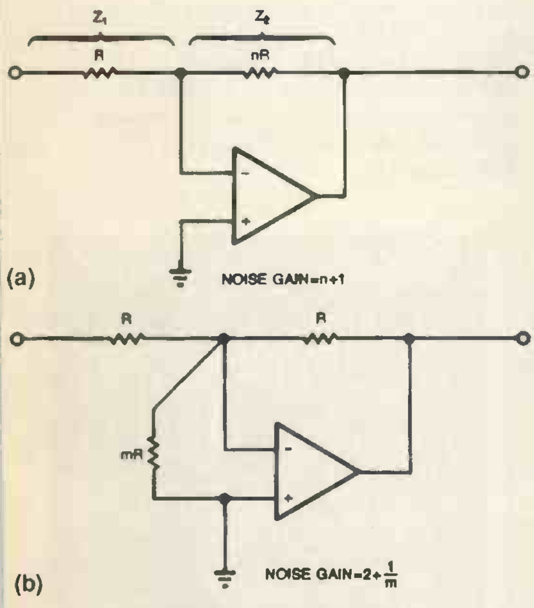



*Fig. 8. By adding just a single resistor, the noise gain of a standard integrator can be tailored.*

Using bypass capacitors is not just a rule of thumb, but a matter of good engineering.

● Avoid unnecessary capacitive loads; they can cause an op amp to develop additional phase shift, making the op-amp circuit ring or oscillate. For example a 1× scope probe, a coaxial cable or other shielded wire added to an op amp used to convey its output to another circuit. Unless the op amp is proven to be stable driving that load, add some stabilising circuits. It does not take a lot of work to apply a square wave or a pulse to the op amp to see if its output rings badly or not. Check the op amp's response with both positive and negative output voltages because many op amps with PNP follower outputs are less stable when $V_{out}$ is negative or the output is sinking a current.

Many analyses claim to predict capacitive loading effects when the op amp's output is resistive. But the output impedance of an op amp is usually *not* purely resistive, and if the impedance is low at audio frequencies, it often starts to rise inductively at high frequencies, just when you need it low.

Conversely, some op amps (such as the NSC *LM6361*) have a high output impedance at low frequency which falls at high frequencies – a capacitive output characteristic. So adding more capacitance on the output just slows down the op amp a little and does not change its phase very much. But if an op amp is driving a remote, low resistive load that has the same impedance as the cable, the terminated cable will look resistive at all frequencies and capacitive loading may not be a problem.

An inverter's and integrator's capacitive load can be decoupled as shown in **Fig. 7**. If the components are well chosen, any op amp can drive any capacitive load from 100pF to 100µF. The DC and low-frequency gain is perfectly controlled, but when the load capacitor grows large, the op amp will slow down and will eventually just have trouble slewing the heavy load.

Good starting point component values are $R_I$ = 47 to 470Ω and $C_F$ = 100pF. These values usually work well for capacitive loads from 100pF to 20,000pF. To make an integrator or a follower required an additional 4.7kΩ resistor as indicated in **Fig. 7d**.

In some cases, as with an *LM110* voltage follower, the feedback path from the output to the inverting input is internally connected and thus unavailable for tailoring. Here we can tailor noise gain. Noise gain is defined as $1/\beta$, where $\beta$ is the attenuation of an op amp's feedback network, as seen at the op amp's inputs, referred to the output signal. For instance, the $\beta$ of the standard inverter configuration (**Fig. 8a**) equals $Z_I/(Z_I+Z_2)$, so the noise gain equals $N+1$. Noise gain can be raised as shown in **Fig. 8b**.

● Add a feedback capacitor across $R_F$ unless it can be shown that this capacitor is not necessary (or is doing more harm than good). This capacitor's function is to prevent phase lag in the feedback path. Of course there are exceptions, such as the *LF357* or *LM349*, which are stable at gains or noise gains greater than 10. Adding a big feedback capacitor across the feedback paths of these op amps would be exactly the wrong thing to do, although in some cases 0.5 or 1pF may be helpful.

Two formulae can be used to obtain considerably improved bandwidth and excellent stability.

For high values of gain and of $R_F$, use:

$$C_F = \sqrt{\frac{C_{in}}{GBW.R_F}}$$

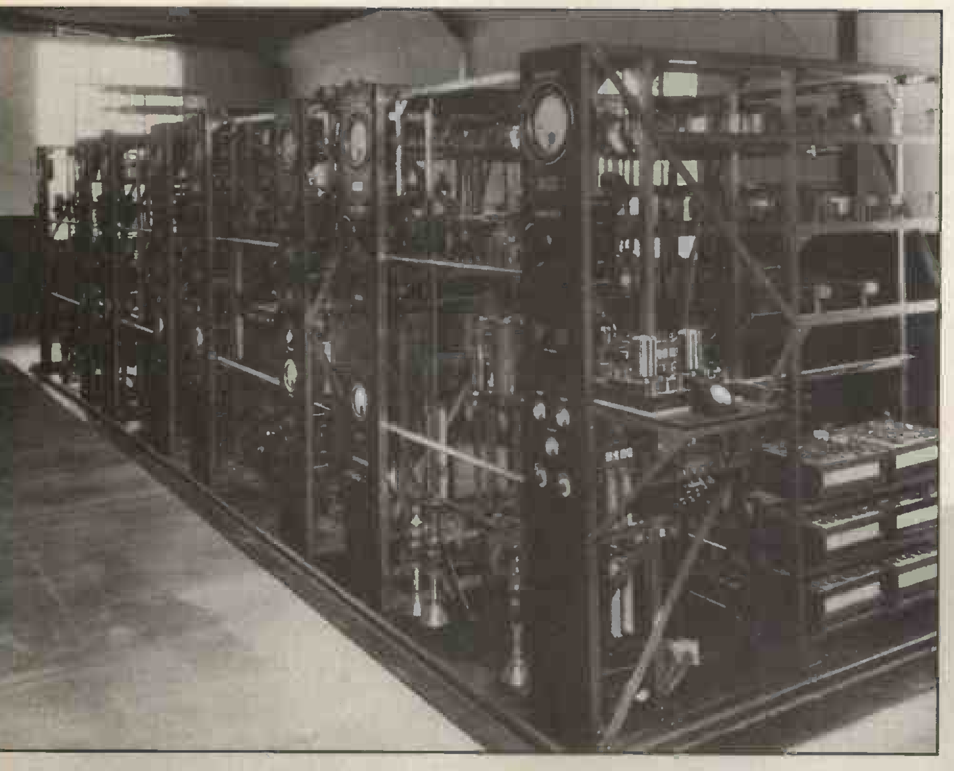
where $G_{BW}$ is the gain-bandwidth product. In those cases in which the gain or impedance is low, such as where $(1+R_F/R_{in})\leq 2\sqrt{(G_{BW} \times R_F\, zC_{in})}$, use the following equation

$$C_F = \frac{C_{in}}{\left(2+\frac{2R_F}{R_{in}}\right)}$$

These equations came from real analytical approaches that have been around for 20 years. The value of $C_F$ computed is not that critical; it is just a starting point. The circuit must be built, trimmed and tested for overshoot, ringing, and freedom from oscillation. If the equation gave 1pF and a clean response can be obtained only with 10pF, you would be suspicious of the formula.

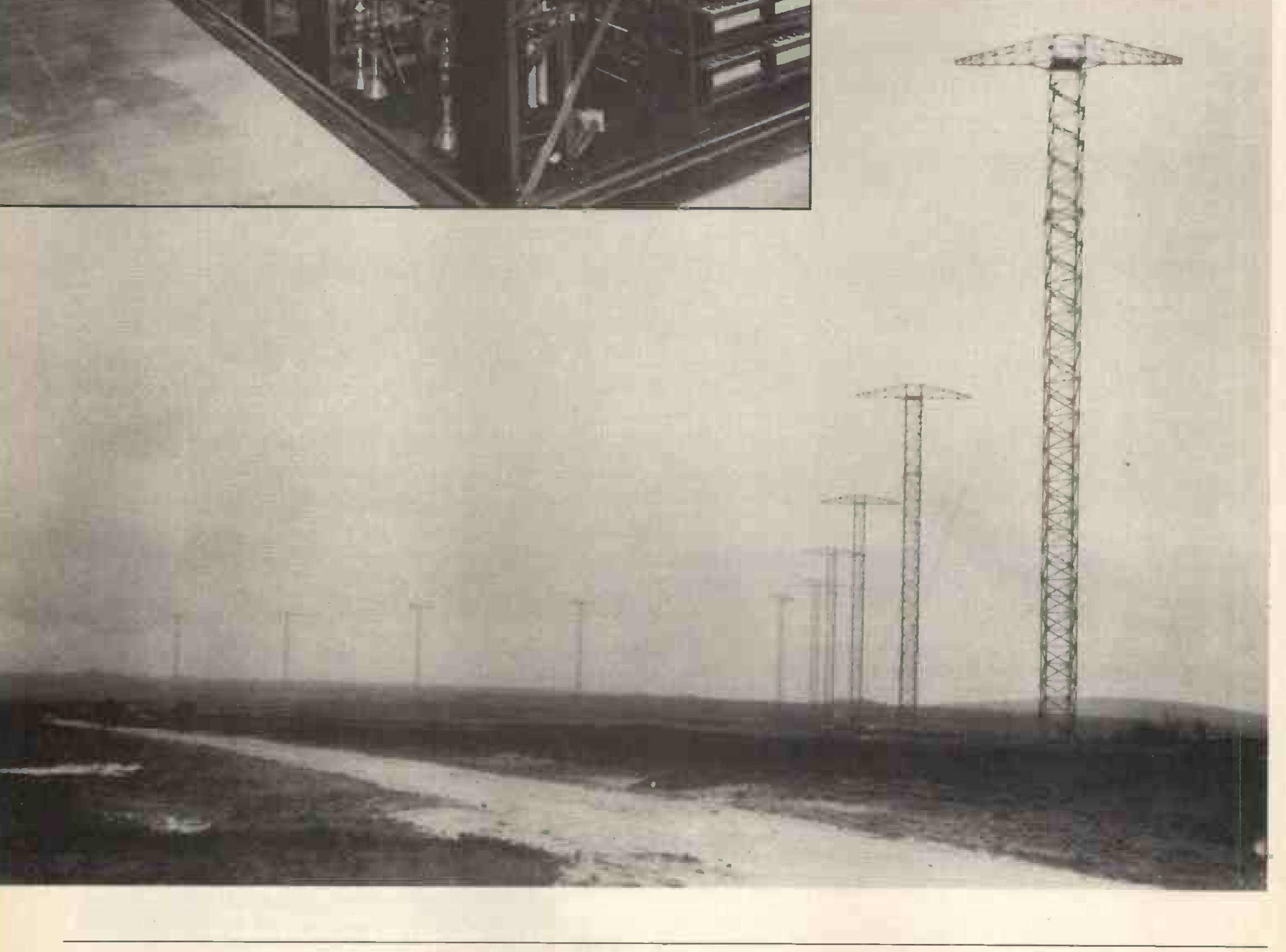
Note that when moving from a breadboard to a PC board, the stray capacitances can change, so the value of $C_F$ must be rechecked. In some cases a separate capacitor may not be needed if 0.5pF is built into the board. ■

# DEVELOPING ANTENNAS FOR BEAM STATIONS



*The four-panel Canadian transmitter at Bodmin. Nearest, to the right of the picture, is the absorber and signalling panel, which kept the load on the generators constant between "marking" and "spacing" in the Morse transmission by diverting the high-tension supply through resistances by type CAM2 oil-cooled valves during spacing periods. In the foreground are two signalling relays, with a change-over switch on the diagonal panel. Two magnifiers, one for each waveband follow, and the end panel contains the No 1 magnifier, or power oscillator and the high-frequency coupling circuits, which convey energy to the aerials.*

*Above The two lines of aerial masts at the Bodmin beam station , opened in 1926 to handle short-wave transmissions to Canada and South Africa. The beams followed great circle tracks, each at right-angles to its line of aerial masts. Alignment required celestial accuracy.*



The shortwave Imperial wireless chain – the beam system as it came to be called – like its long-wave high power predecessor was based on pairs of stations, a transmitter and receiver. Many were in fact conversions of stations begun for high-power links.

Experiments by Marconi and colleagues in 1923/24, using wavelengths between 92 and 32m had established that the shorter the wavelength, the longer the daylight range (when signals were supposedly at their weakest).

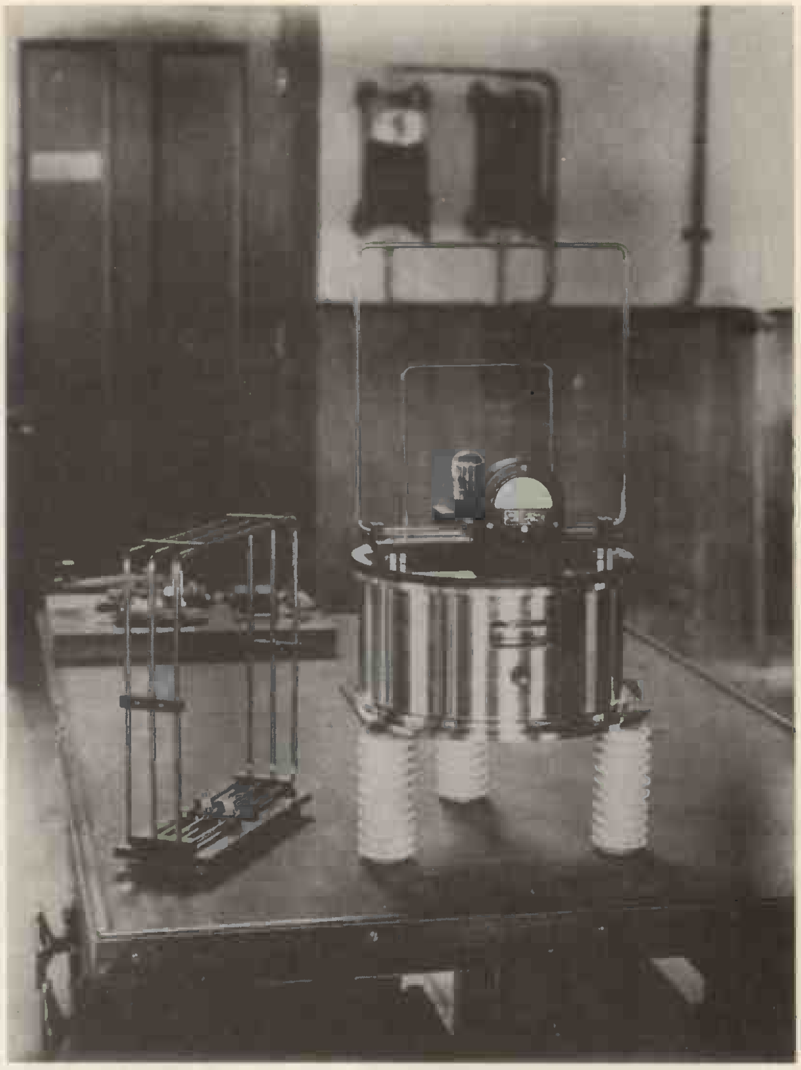
Test transmissions on 32m, using a power of only 12kW, were received in Montreal, New York, Buenos Aires and Sydney, in October 1924. These results came just in time to halt mast construction at some of the planned stations.

Smaller antenna and reflector systems could be substituted. These consisted of a five-mast system, the 287ft high masts spaced 650ft apart, producing a 1300ft antenna path for each of the two discrete wavelengths.

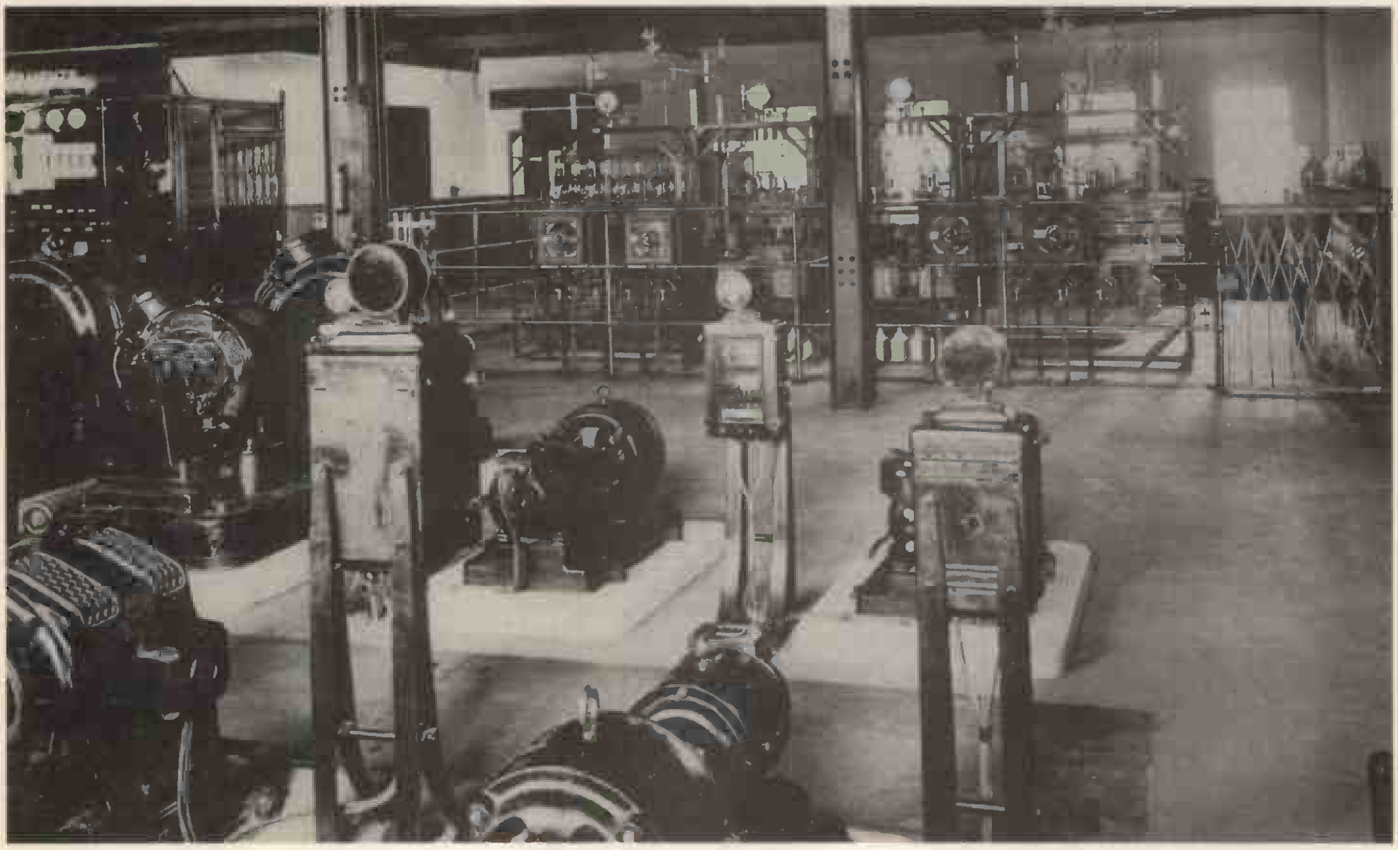
Precise alignment of the masts was of vital importance, and had to be carried out with great care – using fixes on the sun and stars – to ensure that they were square-on to the shortest great circle path to the destination.

The angle of elevation was 10-15° from the horizontal for ranges of 2000 miles upwards. Phasing coils were used between half-wave sections to bring antennas into phase (later replaced by a zigzag array to produce non-radiating phase reversing).

To ensure the efficient transmission of high-frequency power between the transmitter and the antennas, C S Franklin, who designed the array, devised the concentric feeder. This consisted of air-insulated concentric copper



*A standard wave meter, used for checking the transmitted wavelengths.*



*The machinery hall at Bodmin. Power required was less than 20 kW, compared with up to 1000kW for an equivalent longwave transmitter. Rectifiers, in the background, provided high-tension DC anode supplies for the valve transmitters.*

*The Australian Beam Station at Grimsby was designed to work in two directions, west in the morning and east in the evening, both following the great circle path, in order to allow for the effect of the position and altitude of the sun on transmissions. Two aerial systems, either side of a central reflector, were employed, and can be seen here either side of one of the three masts. The reflector can be traced just to the right of the mast by a line of balance weights, used to keep the array taut under wind pressure.*

tubes, held apart by porcelain spacers – in principle the same design as modern coaxial cable. It was carried on iron supports driven into the ground. A symmetrical branched distribution kept the supply to all the antenna wires in phase.

Transmitters used a valve drive taking less than 100W; output was then amplified in three successive stages. Duplicate drives were installed where different wavelengths were used for day and night transmissions, permitting a wavelength change to be effected in about 10 minutes.

New valves had to be designed to overcome problems with the high frequencies in use. These were the oil-cooled CAT (cooled anode transmitting) valves.

The receivers consisted of a single RF stage and demodulator with additional AF amplification stages as required. To cope with frequency drift from the transmitters, the tuned stages were given fairly wide band-pass, while limiting circuits were used to offset fluctuations in signal strength due to the ionosphere.

Bodmin in Cornwall was the transmitting station for Canada and South Africa, with the receiving station at Bridgewater, Somerset. Grimsby and Skegness, on the East Coast, served Australia and India, while Dorchester and Somerton covered North and South America. ■


*The Australian desk at the Central Radio Office, London, where messages were sent to Grimsby, and received from Skegness, by land line, and from where the automatic stations were controlled.*

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# JOHN "RADIO" BROWN

## JIG Brown C.ENG MIEE

who died on January 11 1993, aged 77, was a brilliant electronic engineer and inventor of several devices that proved of vital importance to the War effort in 1940-45.

After the war, he carried out valuable work in the development of medical electronics, notably for cardiac investigations and treatment. But he is perhaps best remembered for his work with the Special Operations Executive (SOE). Communications between the UK, agents in the field and underground groups in occupied Europe and other parts of the world were crucial to success of SOE activities. It was John "Radio" Brown, exploiting his skill, exceptional technical knowledge and attention to the minutest detail, that made this communication possible.

Development work, carried out day and night at SOE station IX by John Brown, laid much of the basis for good communications. His originality in producing light and small transmitter-receivers overcame the difficulties experienced with previous large and heavy equipment that had led to the arrest and often execution of agents operating in the field. Light but powerful suitcase transmitter-receivers designed and produced by Brown, the A Mk I and B Mk II, carried the SOE communications load for the duration of the War.

The equipment usually had a power output of about 20W and operated on Morse code, in the range 3-15MHz – it must be remembered that at the time radio transmission and reception were dependent on thermionic valves consuming relatively high power.

Not content with designing the communications equipment, John Brown also developed the ancillaries, including a pedal generator adaptable for wind drive and a thermocouple charger for the batteries. The charger consisted of a large number of couples housed in a brazier, and in a single night this could fully charge the accumulators.

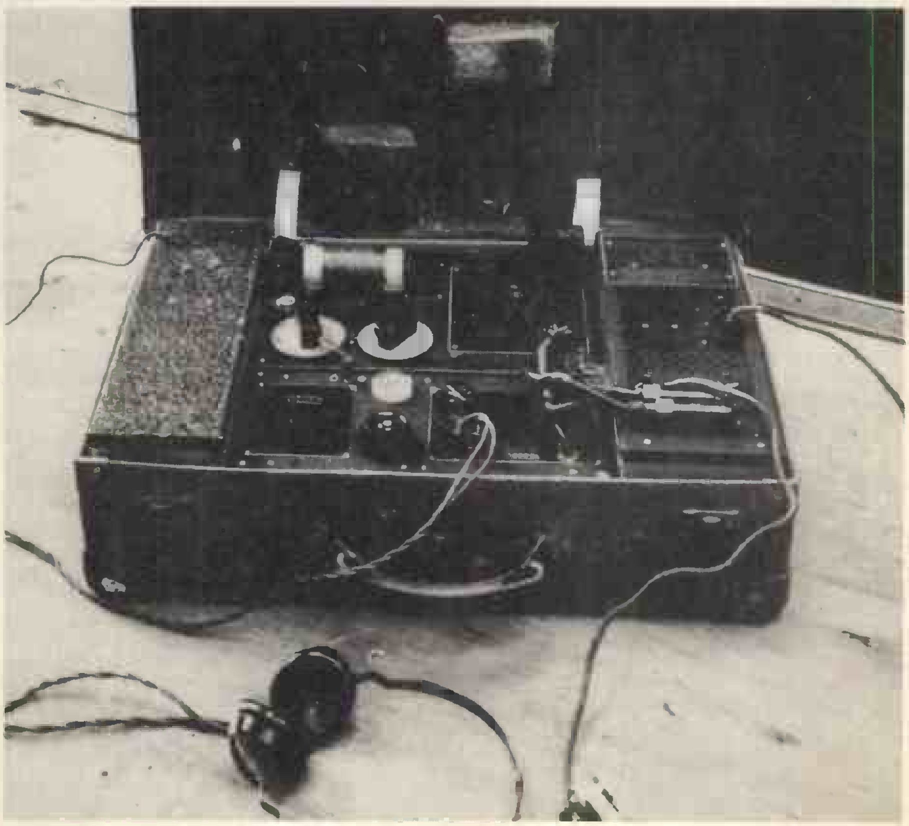
***John "Radio" Brown died on January 11 1993, aged 77.***

In 1944 Britain needed communication receivers covering from 150kHz to 15MHz to enable radio broadcasts to be heard anywhere in the world. The coded broadcasts would indicate the arrival of parachutists, and other SOE activities, and the extensive waveband was needed to penetrate the extensive enemy jamming of so many frequencies being broadcast by the BBC. To meet the requirement John Brown designed, in very little time, the miniature communications receiver (MCR), 20,000 of which were rapidly produced and dropped in occupied countries all over the world.

It is impossible to estimate the value of Brown's work and his contribution to the success of SOE operations. But countless lives were saved by the efficient communications enabled by his designs.

Brown held the rank of Major in the Royal Signals and after demobilisation ran his own company Aveley Electric where he carried out development of important electronics connected with cardiac diagnosis. He travelled to several countries lecturing on clandestine radio techniques.

**Charles B Bovill**



***John Brown's suitcase radio employed a 6V6 as PA and a 7H7 as crystal oscillator. The superhet receiver used three valves. The PA coil output circuits were flexible enough to load up to anything – including a bedstead.***

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## APPOINTMENTS

The EISCAT Scientific Association performs ionospheric research with incoherent scatter radar systems operated in northern Norway, Finland and Sweden. As an important step in the evolution of the EISCAT Scientific Association a new radar station, the EISCAT Svalbard Radar (ESR), will be built on the island of Spitsbergen in the Svalbard Archipelago north of Norway.

For the construction of the EISCAT Svalbard Radar instrumentation a

**System Integration Engineer**

will be installed. This engineer will assist the Radar Project Engineer in technical matters and conduct and supervise digital engineering as well as the interfacing of digital and analogue instrumentation of the ESR system, work on the acquisition of auxiliary subsystems, specification and execution of test procedures, the implementation of monitoring and control systems and the integration, tests and quality control of the ESR system to be installed on Svalbard.

A highly qualified and well motivated individual, possessing a degree in electronic engineering, computer engineering or technical physics as well as practical knowledge of digital and analogue instrumentation is required. Familiarity with modern system software design is important and practical knowledge of RF engineering is desirable. The successful candidate should be prepared to be substantially involved in the practical construction of the radar system. An ability to communicate effectively in spoken and written English is essential. The system integration engineer will initially work at the EISCAT site in Tromsö, Norway, but must also be willing to spend considerable time at the Svalbard site during the system implementation and tests.

This position will be available for a period of four years through the entire construction, implementation, test and initial operation phase of the EISCAT Svalbard Radar. The salary will depend on qualifications and experience and will follow the corresponding Scandinavian scales.

More information on the EISCAT Scientific Association and a detailed description of the EISCAT Svalbard Radar can be obtained from The Director, EISCAT Headquarters, P.O. Box 812, S-981 28 Kiruna, Sweden (Tel.: +46-980-79153; Fax: +46-980-79161). Applications should be sent to the same address by 30 April 1993.

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