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Writing to win

Next month EW+WW, in conjunction with Hewlett-Packard, is launching a major writing award scheme which we hope will unlock the innovative and creative thinking going on in our work rooms, classrooms, institutes and industry.

The person who submits the best electronics design article for publication in this magazine over the period June 1, 1993 to May 30, 1994 will be given a brand new Hewlett-Packard HP54600A 100MHz digital storage instrument worth £2500. This is in addition to our normal authors' fees.

To win this most magnificent prize, the author would be expected to submit an unpublished script of original work concerned with applied electronics at the component level. Designs showing ingenuity in the use of modern devices will be strong contenders, particularly if documented with clear circuit diagrams and concise explanations of circuit operation.

The judging panel, which will include both myself and H-P engineering staff, would hope to see contributions representing all areas of electronics: RF, microwave, audio, video, consumer electronics, data acquisition, signal processing and computer peripherals. The basis for the article may be hobbyist, educational or commercial. Provided that you are the accredited designer of a project and prepared to see details of your work published in full, then we look forward to reading your script.

Although there is only one oscilloscope on offer as a prize, we will naturally wish to publish other suitable submissions. All published material will attract authors' fees which are generous in themselves. For instance, a good design article is worth several hundred pounds to us.

I would be pleased to help potential writers to win the H-P oscilloscope and get their good ideas into print. Simply give me a call at my office to discuss the publishing potential for your latest piece of high technology. I have sets of authors' guidelines for those who require them.

I look forward to hearing from you.

Frank Ogden.

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MPEG-2 digital TV all set to go

The moving force behind the world's first digitally compressed television service – arguably the biggest breakthrough in delivery since the Emitron system consigned Baird to history in 1936 – will be a new video compression system using the MPEG-2 algorithm. It can squash up to four channels in the frequency space required for one analogue terrestrial slot.

When *DirecTv* starts broadcasting to America early next year, with over 150 channels from two co-located 16–transponder satellites, the decoder chips inside the Thomson-made TV sets will be supplied by C-Cube Microsystems, which claims world leadership in the development of broadcast-resolution MPEG decoding.

C-Cube, founded in August 1988 to develop digital imaging technologies, will also be providing the ICs for the encoding system being developed for *DirecTv* by another Silicon Valley company, Compression Labs of San Jose, Cal.

"The big deal is digital TV, not HDTV" according to Alex Balkanski, co-founder and vice-president of C-Cube. "Once one has made the break from analogue delivery, HDTV is just a matter of more chips."

C-Cube has already developed an MPEG broadcast-resolution decoder chip, the *CL950*. This provides a data rate of up to 10 Mbit/s and a resolution of 704 x 576 pixels (in pal) which is superior to existing picture delivery, including laser disc. It follows development of the *CL450*, designed to what is now called the MPEG 1 standard, for computer and CD-ROM applications. This uses a bit-rate of 1.5Mbit/s, to produce a picture of 352 x 288 pixels (pal), about equivalent to VHS standard.

For comparison, a standard analogue TV picture, converted to digital but uncompressed requires 90Mbit/s, taking up virtually all the bandwidth of a 36MHz satellite transponder.

The *CL950*, and a new encoder chip, the *CL4000*, shortly to be launched, are designed to meet the new MPEG 2 standard, developed specifically for broadcast television, which was due to be ratified at the end of March. The *CL4000*, which has cost close to $15m to develop, runs at 40MHz clock and uses 1.2 million transistors and 400,000 gates. Balkanski claims it to be the world's first single chip real time MPEG encoder.

As well as *DirecTv*, applications for the two chips will include full motion video for Philips CD-I. C-Cube is not itself a volume chip manufacturer; to ensure supplies it has partnership agreements with Texas Instruments and AMD.

Because digital compression depends on transmitting only the differences in the TV picture from one frame to the next, it follows that the required bit rate will depend on the nature of the material being transmitted. A relatively static programme, such as an Open University seminar will need less bandwidth than a fast-action sports broadcast. The flexibility to apportion the desired bandwidth, within the overall availability, on a programme-by-programme basis is a characteristic of MPEG.

The MPEG-2 chips will have bit rates of 1.5-10Mbit/s (in practice, 4 to 8Mbit/s) and will be fully scalable between these limits – and beyond them, by putting chips together and dividing the picture between them, to produce HDTV resolution.

Digital compression will enable broadcasters to economise on transponder space; annual rental of an Astra transponder currently costs around £4.8m. Consumers

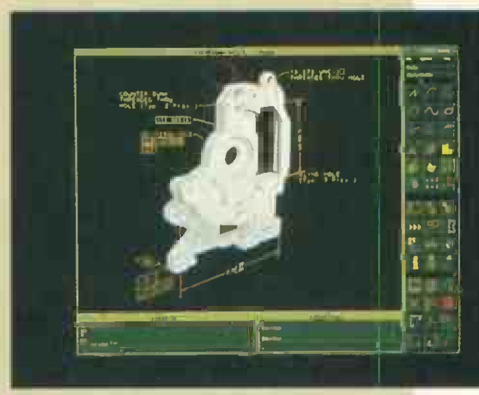
Bright ideas in action

Young Electronics Design Awards finalists Gary Lockton (left) and Richard Coull await judgement on their projects entered for the prestigious awards scheme sponsored jointly by Texas Instruments and Mercury Communications.

Lockton designed a wattmeter capable of calculating and predicting the electricity cost of any appliance connected to it. It uses induced field to obviate the need for a direct connection to the electricity supply.

Coull's project featured an EHT generator to ionise and clean up particulates from vehicle exhausts. A series of charge injecting needles divert particulate matter into a sidestream which is then intercepted by a conventional filter. Richard claims removal of at least 75% of solid matter from exhaust gas by this means.

Other projects reaching the finals included an hydraulic fluid tester which measures boiling point through nucleate boiling detection, a text transmission aid for deaf phone users making use of the standard DTMF key tones, and a snooker score totaliser. The age of entrants ranged from 13 to 22. The scheme attracted more than 300 projects.



The latest version of *I-Deas Master* mechanical design software contains a modeller with a variational geometry constraint management system that combines solid, surface, and wire-frame representations in a single structure. This modeller, shown in the picture, serves as the nucleus for all the other *I-Deas* applications, incorporating as it does nearly all aspects of the design including dimensions, variational constraints, assemblies, drawings, tolerances, and manufacturing data. It was developed by Structural Dynamic Research of Hitchin.

are likely be faced with a high equipment cost. Decoders are heavy on ram which is not cheap in the quantities needed. Expected launch price of the *DirecTv* integrated receiver-decoder is $700 (£500), with, so it is claimed, very little room for reduction as volume sales get underway. The unit includes a modem, for passing pay-per-view information back to the central control unit, as well as *Videocrypt* module. Compression Labs has already developed and marketed compression systems for use in videoconferencing, business television and distance learning channels. It has also supplied the system, brand-named *SpectrumSaver*, which enables Greenland's television service to overcome the problems of satellite reception so far north of the equator: the more robust digital signal eliminates the snow and ghosting caused in analogue by the poor signal to noise ratio at those latitudes. Programmes which were previously distributed to the more remote areas on videotape can now be retransmitted (on VHF) from eight downlinks.

Peter Willis

C-Cube Microsystems (010 1) 408-944 6300.

BBC science coverage is abysmal

The magazine *Science & Public Affairs* representing the views of the Royal Society and the British Association has launched a savage attack on the BBC for its lack of science coverage.

The attack comes from the BBC's own Nick Ross who writes "scientists might well wonder why they bothered to discover and invent radio and television for all the good these media have done for science."

Attacking the broadcast industry for heavy bias towards the arts, Ross asks why so few senior BBC staff are science graduates. He also criticises the gee-whizz style of popular science programming.

He said: "We live in a society that is crippled by scientific illiteracy and pseudo-scientific balderdash."

With renewal of the BBC charter only three years away, Ross believes that scientists should push for a charter forcing the BBC to raise public awareness of science.

Japan works on mind reading computer

Scientists in Japan may have made a breakthrough in creating a system which can read human thoughts before they are turned into words. Fujitsu Laboratories and the Research Institute for Electronic Science of Hokkaido University have isolated the brain wave changes associated with communication even without speech.

The experiment consisted of measuring the brain wave distribution of subjects thinking the sound of the vowel "a." Preliminary results indicated that the speakers' thoughts were observed successfully.

By applying this technology, it could become possible to commercialise a thought input computer by which data can be input simply by thought.

The recognition of silent speech was verified by the measurement of a negative potential distribution generated in the subject's frontal lobe roughly 0.42s after the vowel "a" is thought.

Windows on New Technology

This year will see an unprecedented number of major operating systems being introduced, some capable of running on different microprocessor architectures. It is a battle to determine the necessary components of future computer systems.

Microsoft is the largest developer of operating systems, dominating the market with MS-dos which is tied to the Intel architecture. Microsoft is releasing *Dos 6.0* which improves on the standard with various new features such as data compression. Microsoft also supplies *Windows 3.1*, a graphical user interface that sits on top of MS-dos.

Later this year there will be a 32-bit version of dos integrated with *Windows* and next year there will be *Cairo*, an object-oriented version of *Windows NT*. But it is the eagerly awaited *Windows NT* that is the big news this year since it represents a break with MS-dos and its inherent limitations on memory addressing and multitasking capabilities.

Windows NT is a true 32-bit operating system, offering multitasking in which different applications can be run at the same time and also multi-threading capabilities in which several different tasks can be performed without the overhead of launching different applications to manage those tasks.

It will integrate advanced networking capabilities which are increasingly important in computing. *NT* is also an example of a portable operating system in that it will run on different microprocessors. It is seen to be the key to unlocking the potential high performance of Intel's *Pentium* microprocessor but it will also run on Mips microprocessors, and DEC's *Alpha* chip.

Microsoft also plans other attacks on the operating system market. There is *Modular Windows* which is a stripped down windows version that can be frozen into rom chips, and *WinPad*, an operating system based on windows application programming interfaces to produce a slimmed down operating system designed for PDA devices.

NT's biggest competitors are the various varieties of Unix and IBM's OS/2. Unix continues to make slow and steady progress but it suffers from high system overheads and problems in attracting computer users more familiar with dos operating systems.

Unix is also a multitasking, multi-threaded 32-bit operating system but requires large memories and much hard disk space. Novell, the leading developer of network operating systems, is in the process of acquiring Unix Systems Labs which controls development of Unix System V, the dominant Unix variety. It will be interesting to see if it can combine its *Netware* operating system with Unix to produce a strong competitor to Microsoft's *Windows NT*.

IBM's OS/2 2.0 is also growing steadily and IBM expects to have almost four million users by the end of this year. Like NT, it is a multitasking, multi-threaded 32-bit operating system with a key advantage in that by the time *Windows NT* debuts this summer it will have been around longer; it also takes up fewer system resources than *Windows NT*. Version 2.1 will be out just before *Windows NT* is introduced. IBM plans a portable version of OS/2 that will run on different microprocessors and it will eventually be integrated into a Unix-like operating system through IBM's use of the *Mach* operating system kernel from Carnegie-Mellon University.

Later this year we will also see the first versions of the *Pink* or *PowerOpen* operating system from Apple and IBM. This represents a new type of operating system called object-oriented which eases software development and portability to different hardware platforms. This will come out of the Taligent joint venture between the two companies and is designed to run on the *PowerPC* risc architecture.

Another object oriented operating system due out this year is *NextStep 486* from Next Computer which recently ditched its hardware business to concentrate on its systems software. Next has ported the operating system to run on *486* systems and is preparing to port it to the *Pentium* and other hardware platforms too. Sun Microsystems plans to update its well respected *Solaris* operating system later this year and has plans to port it to non-Sparc based systems.

Matthew Thomas, *Electronics Weekly*

Pentium paves path for faster PCs

The next generation of *dos/Windows* personal computers might be up to five times faster than the present generation of machines following Intel's launch of the *Pentium*, aka i586, microprocessor.

Initial versions have clock speeds of 60 and 66MHz and deliver 63.4 specmarks integer performance, and 54.5 specmarks in floating point. By the fourth quarter of this year Intel is expected to launch a *Pentium* with a speed of more than 100MHz.

The device contains the equivalent of two *486SX* processor cores, a floating point unit, and 64-bit data buses. On performance, Intel claims it is second only to DEC's *Alpha* chip and is twice as fast as the most powerful *486* in integer processing terms. But on the floating point ratings, it lags behind workstation processors from Hewlett-Packard, Texas Instruments, and IBM.

The versions due out later this year though will, Intel claims, have a better performance than all other commercial microprocessors.

Made using a 0.8μm bicmos process and designed with superscalar risc architecture, the *Pentium* has two five stage execution units and can process up to two instructions in a single clock cycle. Later versions will be made using a 0.6μm process.

There are two 8K on-chip caches and a 64-bit burst-mode external bus. It employs 3.1 million transistors, nearly three times as many as the *i486*.

Operating voltage for the new device is 5V, but the 0.6μm units will work at 3.3V providing power reduction for the notebook and palmtop market. In its 5V form it appears to require a fan-cooled heatsink to function.

Intel itself has introduced a housekeeping chipset for the *Pentium* called the *82430* and the company is working with compiler, tools, operating system, and application developers to assist use of the device architecture.

It is possible to have two *Pentiums* in the same computer and it seems likely that dual processor machines will be launched next year. Single processor machines will be upgradeable to dual-processor machines. Though *Dos* and *Windows* do not support multiprocessing, the new *Windows NT* does.

Optical breakthrough for fast PSTN

Researchers at British Telecom have pioneered a technique using only optical components for recovering clock signals from high bit rate data streams. It is part of what BT claims is the first all-optical telephone signal repeater.

It uses a mode-locked laser and an erbium fibre cavity to derive a stream of picosecond optical pulses which are synchronised to the input optical data stream. The incoming 1.54μm pulses are coupled into dispersion shifted fibre which is part of an erbium fibre ring laser.

The non-linear refractive index of the fibre sets up a periodic phase modulation in the cavity which mode-locks the laser. This generates an optical clock signal of 1.54μm pulses which is used to synchronise the amplified data signal.

The ability optically to regenerate a telephone signal without first needing to convert it into an electrical signal, according to BT, will open the way to 100Gbit/s data rates in the public switched telephone network.

Philips TV boss: "never again" for investment in Euro-standard

Rob Oostenbrugge, head of TV manufacturing at Philips, does not try to hide his frustration. After ten years of development, he has watched Europe's mac-based strategy for HDTV fall apart in an orgy of Euro-bickering.

Oostenbrugge has vowed never again to base Philips' TV future on agreements between European governments. He now doubts that any pan-European agreement can be reached on TV standards, and is highly sceptical about talk of a European strategy for digital HDTV.

"When I look at the way we handled mac I am very critical of the way we are now dealing with digital," he said.

Part of the problem is that each country has research groups working on their own brand of digital TV. Oostenbrugge believes these vested interests will make it very difficult to decide on a single digital standard. "Everyone is hanging on to their own ideas. To merge the European proposals would be quite an achievement".

Even if a single standard emerges, Oostenbrugge doubts that the EC can whip into line all the players in the chain, from programme producers through equipment suppliers to broadcasters, needed to turn it into reality. "The question is, is it possible to make everyone in the chain happy ? We are sceptical now. We have been through the whole process before with mac."

As a result, Philips has effectively put European HDTV on the back burner. Oostenbrugge will only move if someone else comes forward with firm proposals. "If there is economics in it and private broadcasters committed to providing new services we will join," he said.

But Oostenbrugge is confident that by taking a pragmatic approach Philips can make money out of advanced TV systems over the next few years. He expects to make bucks out of terrestrial digital TV in the US. In Europe, he sees opportunities in standard definition digital TV delivered by cable and satellite, and in analogue widescreen TV.

Digital technology will take off in Europe, he says, because it will enable satellite and cable TV firms to provide more channels at a lower cost than today, thanks to MPEG digital video compression technology.

"In 1995 digital satellite will happen in Europe," Oostenbrugge insisted. "That's business. By the year 2000 we could have more than 1,000 satellite channels in Europe, of which hundreds are digital".

For terrestrial digital TV he sees France as the best bet, because the French Government has a proven record of seeing projects through to implementation. "I am impressed with the French", he said. If they say they will do something, they do it. With the French you can make a deal".

Oostenbrugge says there is interest in France in a system that carries both an improved definition picture for home TVs and a rugged lower definition signal for mobile receivers, for example in cars.

Irrespective of transmission format, Oostenbrugge sees a strong market developing for widescreen TV. He believes the market will start to take off when PalPlus transmissions start in 1995, enabling both standard and widescreen sets to display pictures from the same signal.

"Once broadcasters start using PalPlus, the high-end TV market will change over to widescreen very quickly," he predicted.

Along with other interested parties, Philips is now urging the EC to give a 500m ECU subsidy to help make widescreen programmes. Oostenbrugge argues that even the budget-conscious UK government has no excuse for not backing widescreen. "If you calculate the extra revenue from VAT through selling 16:9, governments actually get more money out than they put in", he pointed out.

Oostenbrugge admits he is disappointed. "We can't stop it, so let's see how we can make money out of it".

***Karl Schneider**, Electronics Weekly*

Bank sees manufacturing hike

Nearly 13% of all start-up companies in 1992 were involved in manufacturing, the highest since 1980 according to figures released by Barclays Bank.

A statement from the bank said: "Since around 10% of the business stock is in manufacturing, generating an average of 25% of national output, an increase in confidence in this sector is vital for economic recovery to take place."

Despite this, manufacturing output fell 0.4% in 1992. But Barclays predicts a 0.8% growth this year and 2.9% next year ■

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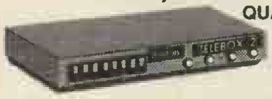

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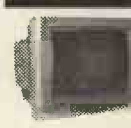

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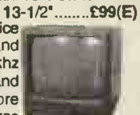

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Lightning link to measure global warming

Climatologists are still uncertain as to whether global warming is actually occurring and if so, how fast. Climate models mostly predict a rise in average temperatures of about 2°C by the middle of the next century. At the moment the average rise (from a baseline before the Industrial Revolution) is not thought to be more than about half a degree at most. But because of the relatively large day-by-day and season-by-season temperature changes, climatologists have great difficulty in detecting this tiny signal buried in a relatively huge amount of "noise".

In his search for an increasingly sensitive global thermometer, Earle Williams, a geophysicist at the Massachusetts Institute of Technology has been studying the unlikely subject of lightning flashes. Williams showed some years ago that there was a strong correlation between the temperature in a particular part of the globe and the incidence of lightning strikes – that is why there are far more electrical storms in the tropics.

Sensitivity of the effect is so great that in one observation in Darwin, Australia, a measured 2°C increase of temperature led to a 100-fold increase in lightning strikes.

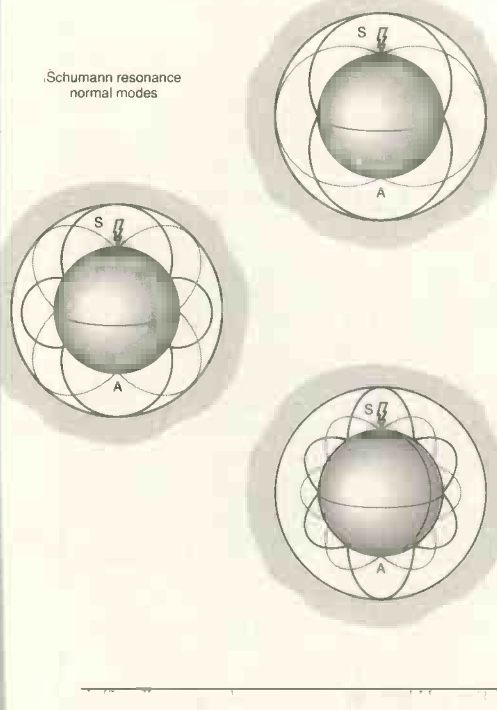
To measure the incidence of lightning on a worldwide basis, Williams has set up an antenna on the roof of MIT to detect an effect known as the Schumann Resonance (SR). The SR consists of low frequency (7-50Hz) standing waves in a global circuit bounded by the Earth and the ionosphere.

Lightning flashes, which occur about a hundred times a second around the world, are constantly exciting these Schumann resonances – a bit like a hammer continually hitting a bell. So amplitude of the SR is highly dependent on the exact incidence of lightning strikes and, given appropriate calibration, is likely to provide a sensitive means of measuring global temperatures.

But there are problems. The extent to which the SR amplitude at any one site can yield a globally representative temperature signal is a matter of debate. Williams says that although SR signals do have a global value, readings vary from site to site. He is therefore comparing his readings at MIT with those made in Alaska and Australia.

Eventually, when the reading are analysed and compared with those collected by other groups during the sixties and seventies, it should be possible to calibrate the SR signal at any given site and use it as a sort of highly sensitive thermometer for global average temperatures.

Another hopeful "electric" thermometer is a quantity called the ionospheric potential, the PD between the Earth and the ionosphere. This potential, which can be measured from balloons and aircraft, may prove to be an even more sensitive indicator of global temperatures. The ionospheric potential is created by the action of lightning, and also by electrified clouds that are not necessarily discharged.



Schumann resonance normal modes

Schumann Resonance models show how low frequency (7-50Hz) standing waves exist in a global circuit bounded by the Earth and the ionosphere.

Quantum leap forward for secret codes

Data encryption is a vital technology these days, not just for obvious military applications, but also for financial transactions and other sensitive situations. The Camillagate affair, where a private phone-call seems to have been bugged, underlined only too well the need for effective coding of voice messages. Now a quantum-based encryption technology under development could help keep secrets secret.

In any encryption/decryption process, the need is to transmit some form of "key" that will unlock the message at the far end. The key can either be built into the equipment (in which case it can be pirated, as in the case of illicit satellite decoders) or can be transmitted along with the data (in which case there's risk of it being intercepted). Under normal circumstances there is no such thing as an entirely secure encryption process. However clever a system is, someone is bound to be one step ahead... Unless, that is you're using quantum encryption.

Tests on a practical system have been conducted by BT Laboratories at Martlesham and by the Defence Research Agency at Malvern. The system hinges on the fact that it is now possible to transmit a key – usually no more than few hundred bits of data – in total security. More precisely, it is possible to transmit the key and to know for sure whether it has arrived securely, or someone has intercepted it. In practice these two situations are virtually equivalent because, if a decoding key is intercepted, no further data would be transmitted on that particular communications link.

The clever aspect about quantum cryptography is that it sends information encoded in individual photons, the fundamental particles of electromagnetic energy. In simple terms, single photons cannot be divided, so either a photon arrives at the legitimate user or it is detected by an eavesdropper. There is no way in which both these options can exist simultaneously.

In practice, information is encoded not just by the presence or absence of a photon, but by its phase. Phase provides additional security against the unlikely possibility of some clever eavesdropper intercepting the message, reading it and then re-transmitting it. Quantum physics rule out the possibility of achieving this correctly more than 75% of the time. So the object of the latest research has been to devise a system that will transmit data accurately for substantially more than 75% of the time.

Quantum cryptography has been tried out in the past, but only over links of a few centimetres. Absorption increases with distance, so some photons never get through to the other end. Of course information is only transmitted when a photon does actually get through. That essentially reduces the rate at which data can be sent. On the other hand because the key is only a few hundred bits, a low data rate is not a serious drawback.

At the moment the system is still experimental, but in recent tests, data of this sort was transmitted with 91% accuracy along a 10km length of optical fibre at BT's laboratories. Ultimately the researchers think that it may be possibly to transmit data with complete security over 100km fibre links.

Clear evidence of molecular rectification

A team of scientists at the University of Exeter and the Cranfield Institute of Technology says it has produced unequivocal evidence of rectification in an organic molecule. Their report (*Phys Rev Lett*, Vol 70, No 2) demonstrates one-way conduction of zwitterions – ions that carry both negative and positive charges, and opens the way to a whole range of new applications in sensors and molecular circuitry.

Ever since the mid 1970s, researchers have been trying to develop the molecular equivalent of a p-n semiconductor by depositing layers of suitable organic chemicals between pairs of metal electrodes. The method has been carried out many times and the rectifying properties have been attributed to resonant electron tunnelling between the electrodes and the charged parts of the organic molecules. The latest research, led by Roy Sambles of the Film and Interface Group in the Physics Department at Exeter, has shown that rectification occurs within the zwitterion and can take place independently of the metal electrodes; in other words entirely within an organic substance.

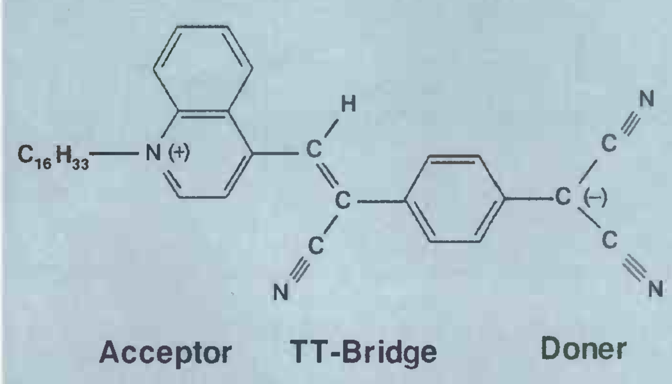
The basic molecular rectifier is made by creating a Langmuir-Blodgett film of the molecule in question by floating it in a single layer on the surface of extremely pure water. One end of the molecule (the +ve end) consists of a paraffin grouping which is hydrophobic; the other negatively charged end consists of hydrophillic cyanide groups. A monomolecular layer of the compound consists, therefore, of an orderly array of molecules, all with their negatively charged ends facing downwards.

Sambles' group lifted this layer onto a silver-coated slide and then repeated the operation seven times to produce a layer seven molecules thick. They then fabricated a number of top electrodes with magnesium contacts.

To show that the rectifying action is completely independent of the electrodes, the researchers did two further separate experiments. In the first case they introduced passive organic spacer layers consisting of omega-tricosenoic acid, a soap-like chemical. The resulting structure – in which the zwitterionic compound has no contact with the metal electrodes and in which no Schottky barrier effects could occur – also shows rectifying properties.

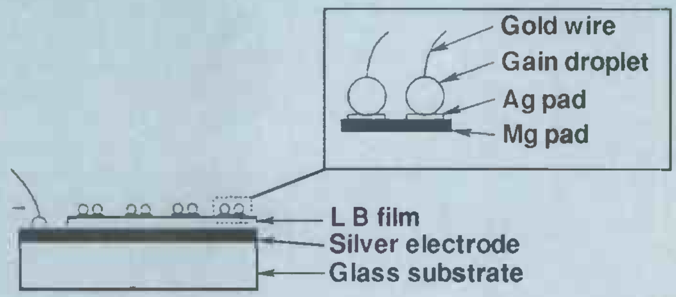
Finally, the team took the original seven-layer molecular rectifier and chemically "bleached" the zwitterion by adding metallic ions. Bleaching of the active molecule instantly destroyed the rectifying action, leaving a device that was almost purely resistive.

The two contrasting pieces of evidence show unequivocally that the rectifier action can take place entirely within an organic molecule. The team now plan to investigate the possibilities of organic photodiode and transistor action.



Structure of the zwitterionic molecule.

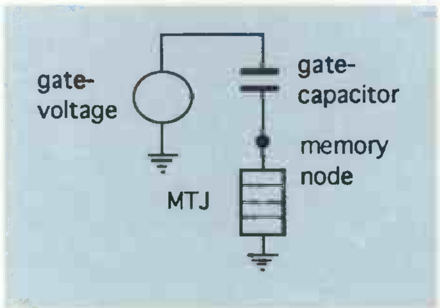
Fabrication of top electrodes.



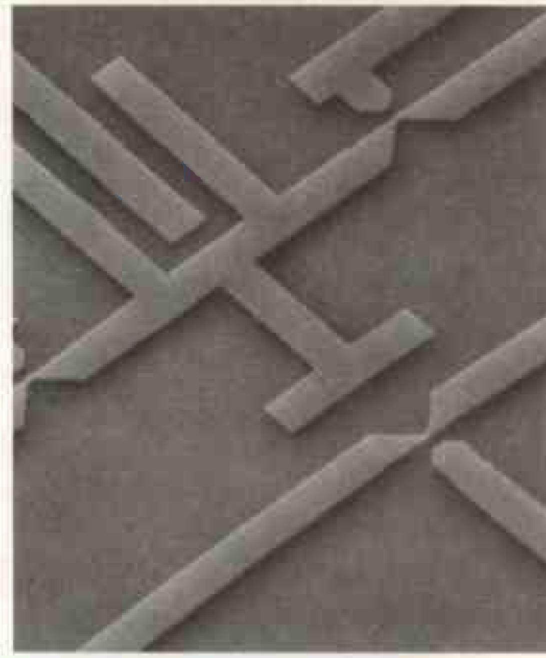
Single electron memory demonstrated

Hitachi Laboratory in Cambridge, in collaboration with the Microelectronics Research Centre at the Cavendish Laboratory, Cambridge University, have demonstrated the possibility of a single electron memory in which one bit of information can be stored by only one electron. The principle has been understood for many years, but this is the first time it has been demonstrated in practice (*Electronics Letters*, Vol 29, No 4)

In the new structure, one bit of information is defined by the precise number of electrons stored at a memory node. The ability to define the number of electrons precisely is made possible by the Coulomb blockade effect, which causes the movement of individual electrons to be controlled. If an isolated region of conductor is made sufficiently small, the change in stored energy due to the gain or loss of an individual electron results in a sufficiently large potential change stopping further electrons from entering and existing electrons from leaving.

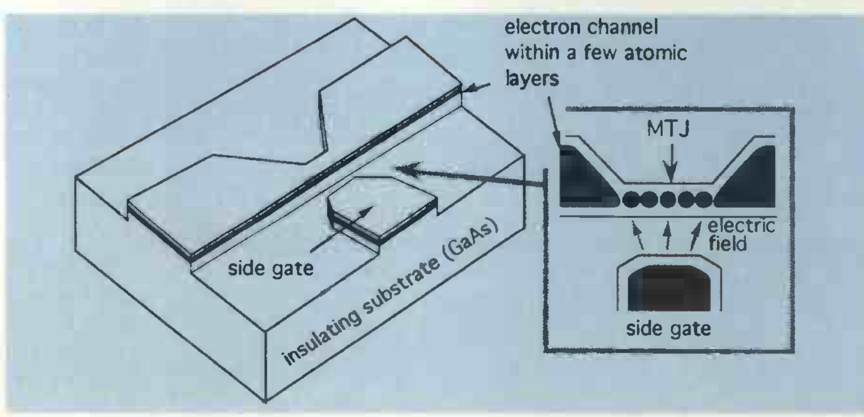


MTJ formed by side-gated structures in delta-doped GaAs material.



Scanning electron micrograph of single-electron memory element with electrometer.

A memory cell consists of one gate-capacitor and one multiple tunnel junction (MTJ).



Principle parts of the single electron memory cell are a gate capacitor and a multiple-tunnel junction (MTJ). The memory node is a small conducting region connected to an external circuit via the MTJ and subject to control by the gate capacitors. Electron transfer to or from the node is only possible through the MTJ, which itself consists of several tunnel junctions in series. The memory node voltage depends both on the number of electrons at the node and on the voltage applied to the gate electrode

To exploit the Coulomb blockade effect the structure must be reduced in size so increasing the electron charging energy. The researchers met this requirement by fabricating a new structure with side-gated channels in delta-doped gallium arsenide. The electron channel is formed within a few atomic layers in an otherwise-insulating GaAs substrate. After adjustment of the side-gate voltage entrance and exit of one electron at a time can be controlled. Although further studies will be necessary to determine the mechanism, the team believe that dopants inside MTJs create tunnel barriers at intervals of several tens of nanometres.

The same fabrication process is used to implement a Coulomb blockade electrometer to detect the voltage on the memory node with minimum interference to the electrons on the node.

Several experimental devices have been constructed using electron beam lithography and they function very reliably. Haroon Ahmed, Professor of Microelectronics at Cambridge, says that at the moment they are still relatively large and may be switching ten to a hundred electrons at a time. He believes that if the line structures are fabricated on a scale of less than 5nm, then they will meet their full theoretical expectations and work at room temperature, rather than 0.1K as at present. To build components on this scale might require manipulating individual atoms using a scanning tunnelling microprobe.

In spite of the enormous practical difficulties, Ahmed is confident that single electron memory chips will be available within another two decades or so. This would enable the creation of one terabit memories consuming a mere 0.1W. Such a memory using conventional semiconductor technology would currently consume about 10kW!

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

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CIRCLE NO. 109 ON REPLY CARD

In the first of a three part series, Dmitry Malinovsky examines phase-locked loops from mathematics to the practical application of PLLs in frequency synthesis and other comms designs

CLOSING THE LOOP

Phase-locked loops, in common with almost any type of electronic system, are easiest to understand when presented as a collection of simpler units.

Most of these building blocks are universal and only form a PLL by virtue of the unique connection used.This article presents the most typical examples of such functionally independent bricks.

All the circuits have been tested by the author while designing frequency synthesizers and other test equipment

Figure 1 shows a VCO circuit often used at about 300MHz with a tuning range of a few hundred kilohertz. A VCO used in frequency synthesis must generate the lowest possible noise at its output and a fet is the easiest method of ensuring this performance.

The tuned circuit is connected to the fet gate via capacitor C_3 from a tap on coil L, which allows the high input impedance of the fet to be used, a fairly small value of C_3 providing a loose connection between the tuned circuit and the fet and the high-gain transistor making it possible to tap from 1/4 to 1/10 of the coil. This gives good frequency stability, since the source capacitance is included in the circuit with a transformation factor less than 1. Elements R_1D_1 stabilise the transistor working point and therefore the output amplitude; the

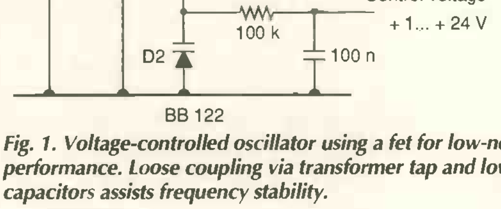


Fig. 1. Voltage-controlled oscillator using a fet for low-noise performance. Loose coupling via transformer tap and low-value capacitors assists frequency stability.

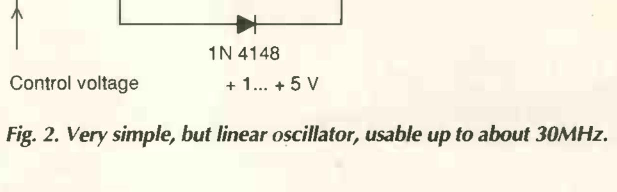


Fig. 2. Very simple, but linear oscillator, usable up to about 30MHz.

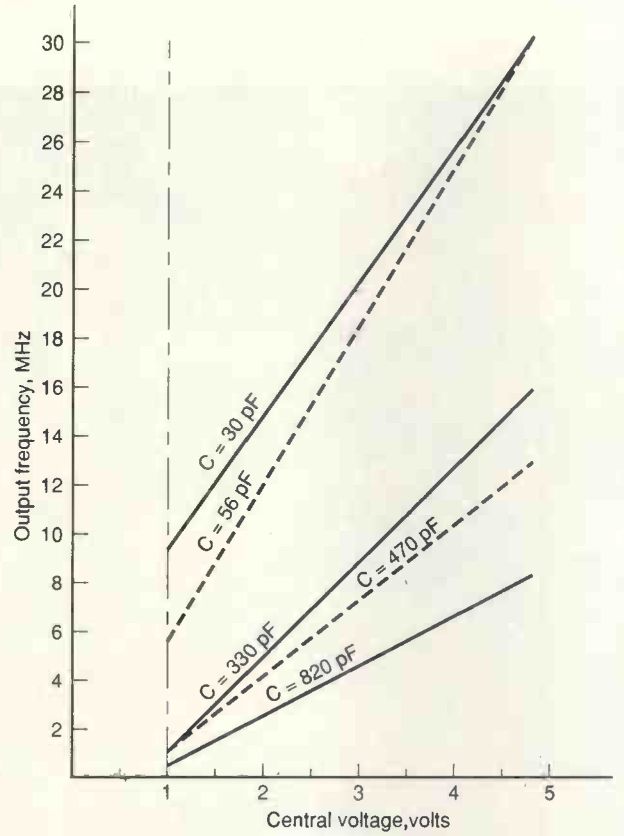


Fig. 3. Transfer characteristic of oscillator shown in Fig. 2 using differing values of C.

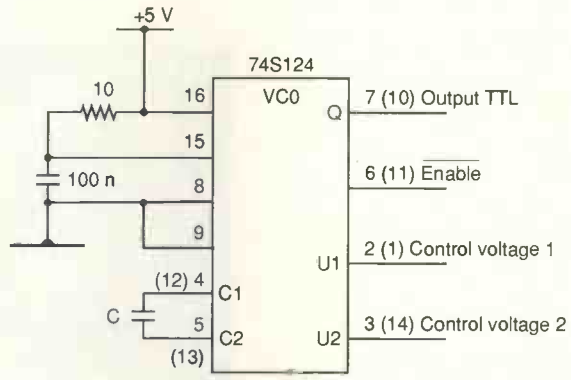


Fig. 4. Dual VCO, giving an output frequency of 5 x 10^{-4}/C(where C is in farads) up to about 60MHz. Tuning range is determined by voltage applied to pin 3.

RC chain in the drain filters the supply voltage. Varicap D_2 determines the tuning range of the VCO, which can be limited by the choice of the additional capacities C_1C_2.

The main merits of the IC multivibrator oscillator shown in **Fig. 2** are very high linearity and almost maximum simplicity; it will work with all types of TTL. **Figure 3** shows the relationship of control voltage to frequency for this oscillator using various values of capacitance. Tuning range depends on the value of the series resistors – increasing resistance reduces the range.

Figure 4 is another IC VCO. A *74S124* contains two oscillators, their outputs being TTL compatible. Tuning range can also be altered here by changing control voltage 2. The VCO is very linear and is used, as well as the design in Fig. 2, in frequency demodulators, which need a high control linearity.

Using no coils, the ECL VCO in **Fig. 5** works at UHF. Here, the mosfet works as a voltage-controlled resistance to set the tuning range, which is changeable by varying the applied gate voltage. Typical tuning range is shown in **Figure 6**. Such a VCO can be used in digital synchronisation systems, frequency demodulators and test oscillators, but has insufficient spectral purity for use in frequency synthesizers.

Control voltage applied to the crystal oscillator in **Fig. 7** changes its frequency by a fraction of one per cent; in the absence of lock, the output frequency of such a VCO is still very stable. Such circuits are used in digital communication systems for the recovery of the carrier, and in frequency standards. In this case, the crystal is working at its fifth harmonic, but it will also work with the resonator on fundamental.

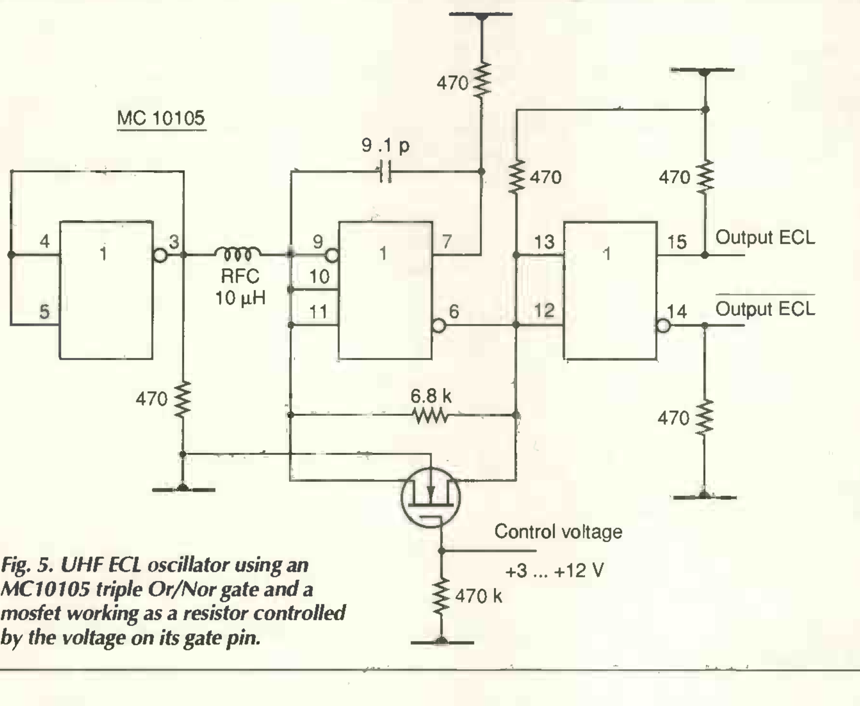


Fig. 5. UHF ECL oscillator using an MC10105 triple Or/Nor gate and a mosfet working as a resistor controlled by the voltage on its gate pin.

Phase detectors

If two inputs to a multiplier or mixer are $V_{in1}(t) = A_1\cos(\omega t + \phi)$ and $V_{in2}(t) = A_2\cos(\omega t)$, multiplying gives two signals $A_{out}\cos(2\omega t + \phi)$ and $A_{out}\cos\phi$, which is at zero frequency and dependent on phase difference. A filter to remove the doubled input frequency completes the phase detector.

The phase detector shown in **Fig. 8** – a double-balanced mixer – was designed more than forty years ago and has been in use ever since. Transformers T_1 and T_2 determine the impedance match.

This type of PD is now giving way to solid-state IC DBMs, of which the Siemens *S042P* shown in **Fig. 9** is a typical representative, working in the frequency range 0-200MHz and having symmetrical inputs and outputs. Its

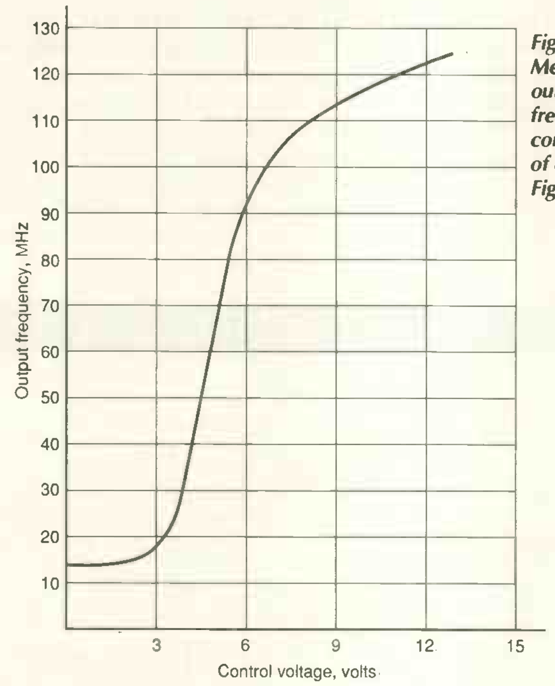


Fig. 6. Measured output frequency vs control voltage of circuit in Fig. 5.

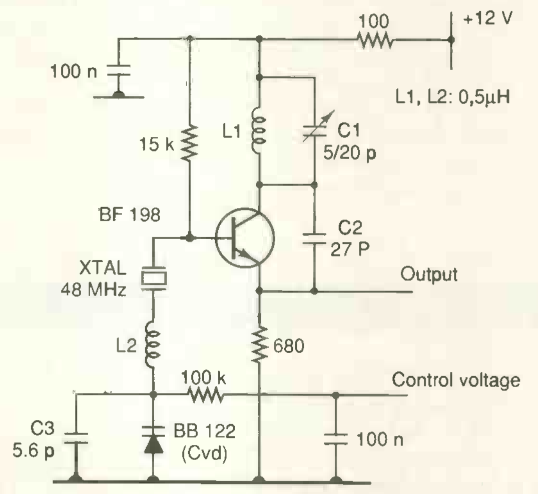


Fig.7. Crystal oscillator capable of frequency variation of less than 1% by varicap diode in series with the crystal.

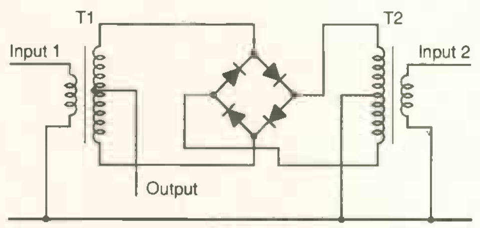
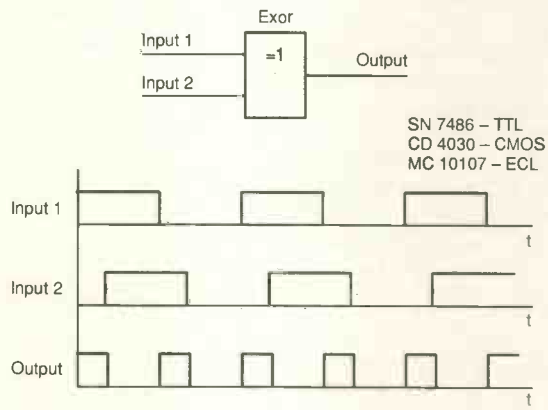


Fig. 8. Early type of double-balanced mixer used as a phase detector, the transformer ratios fixing the tuning range.



Above Fig. 10. Logical exclusive-Or phase detector. Output is a pulse-width-modulated pulse train – zero output when inputs in-phase. A low-pass filter on the output removes the pulses to leave DC.

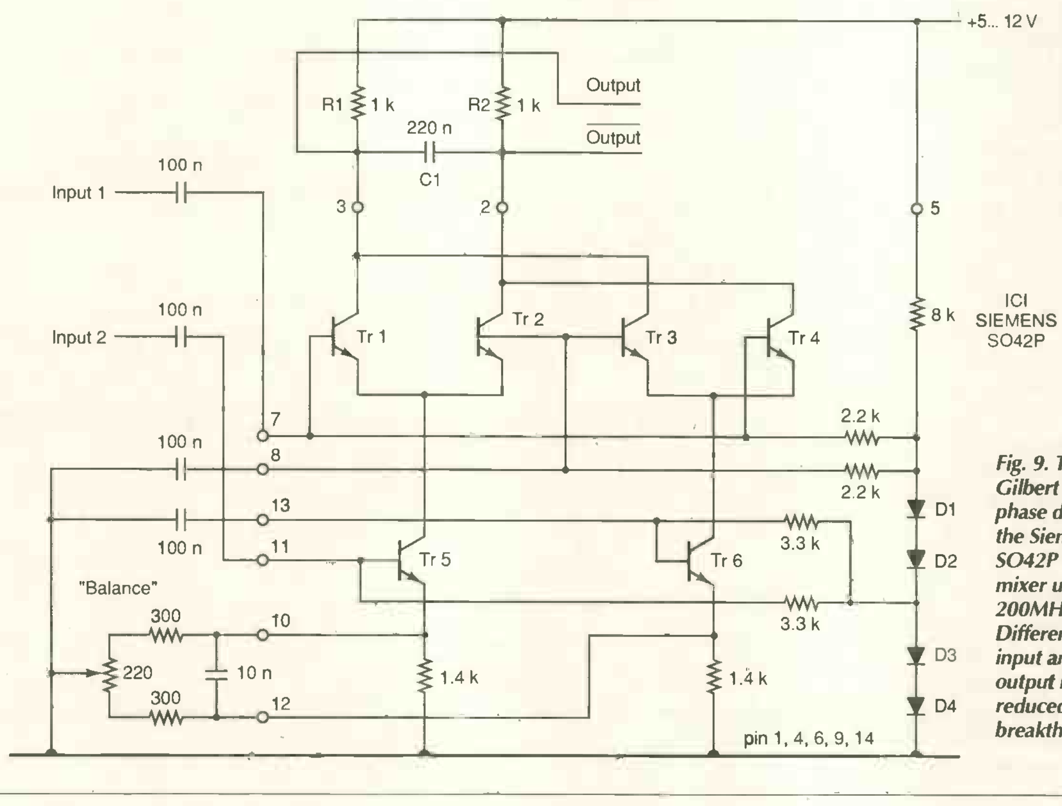


Fig. 9. Typical Gilbert Cell phase detector, the Siemens SO42P balanced mixer up to 200MHz. Differential input and output make for reduced breakthrough

chief merit in comparison with the mixer of Fig. 8 is the balance facility to reduce breakthrough from input to output.

A logical exclusive-Or is a "digital analogue" of a double balanced mixer, a phase detector working in this way being shown in **Fig. 10**. The circuit works as an overloaded DBM with pulsed input signals and does not accept sinusoidal input, unlike the analogue variety.

All three types of PD shown above are "real" phase detectors; they do not work well when input frequencies are off tune by more than 10-20%, which is why search systems were sometimes used in PLLs when there was considerable initial discrepancy between input and reference signals. The invention of frequency-phase detectors (FPD) made such search systems unnecessary. **Figure 11** shows a TTL FPD, but cmos or ECL versions are also made. The device works in the unlocked condition of the PLL as a frequency detector and in the locked state as a PD. Its main drawback is its sensitivity to input phase jitter, whereas a PD using the X-or circuit or the DBM works perfectly with a jitter up to ±45°, which is why edge-sensing FPDs are used almost exclusively in frequency synthesizers with noise-free inputs. The cmos IC *4046* is commonly used; it contains a VCO and two

Phase-locked loop principles

All configurations of the practical phase-locked loop are describable by the typical block diagram of **Fig. I**. Three of the blocks are to be found in all PLLs: the phase detector (PD), the loop filter (LF) and the voltage-controlled oscillator (VCO); elements sometimes absent are the frequency dividers (FD). In digital PLLs in which there is no evident VCO and LF, for example the *SN74LS297*, one can single out the elements having the relevant transfer functions, but working in a digital or pulse regime. Analysis of a classical analogue PLL is therefore a good grounding for the analysis of a fully digital type.

To make a mathematical model of the system shown in Fig. I, first define transfer functions for each block and, since the system is primarily phase-centred, define them as functions of phase, in the s-plane to make life simpler:

input signal:	ω_{in} -- input frequency;
	Φ_{in} -- input phase;
frequency divider by M:	1 / M;
frequency divider by *N*:	1 / N;
phase detector:	$K_d(s)$;
loop filter:	F(s);
voltage - controlled oscillator:	$K_{vco}(s) = K_{vco} / s$;
output signal	ω_{out} -- output frequency;
	Φ_{out} -- output phase

For simplicity, assume the regulating system to be linear and the system to be in a steady state. Using the expressions in Fig.I, derive input/output transforms for the blocks, starting with the phase detector. Since the frequency dividers not only divide in frequency, but decrease the input phase deviation N or M times, put down an expression for the voltage $V_{pd}(s)$ at the PD output.

$$V_{pd}(s) = K_d(\Phi_{in}(s)/M - \Phi_{out}(s)/N) \qquad (1)$$

where $K_d (= K_d(s))$ is the PD's transmission gain in volts/angle (it is usual to define K_d in terms of volts/radian, a radian being degrees/2π). Signal now goes to the LF, whose main purpose is to form the transfer function and to filter HF components from the PD output. Voltage at the LF output is

$$V_{lf}(s) = V_{pd}(s)F(s) = K_d F(s)(\Phi_{in}(s)/M - \Phi_{in}(s)/N) \qquad (2)$$

This goes to the control input of the VCO, which has a transmission gain defined in units of radians/(second volts), i.e. the VCO output signal has the unit of frequency while we are analysing the phase-locked-loop. Phase Φ and frequency ω are related by the classical ratio $\omega = d\Phi/dt$, or in

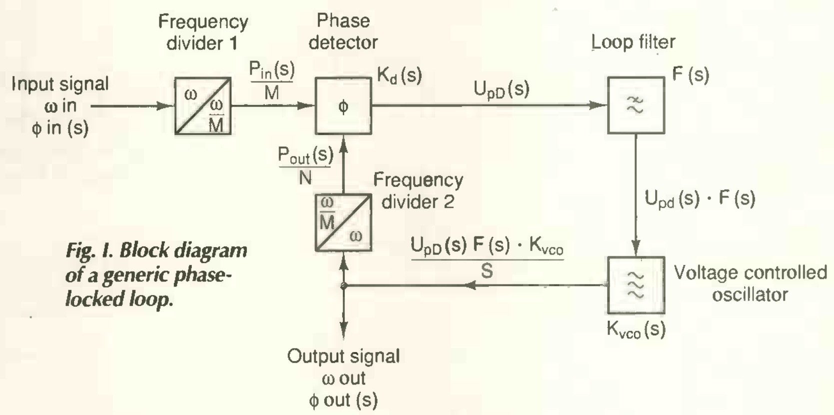


Fig. I. Block diagram of a generic phase-locked loop.

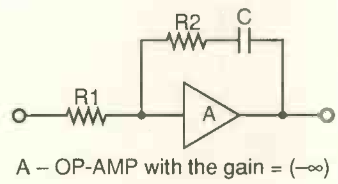


Fig. II. Loop filter for second-order PLL.

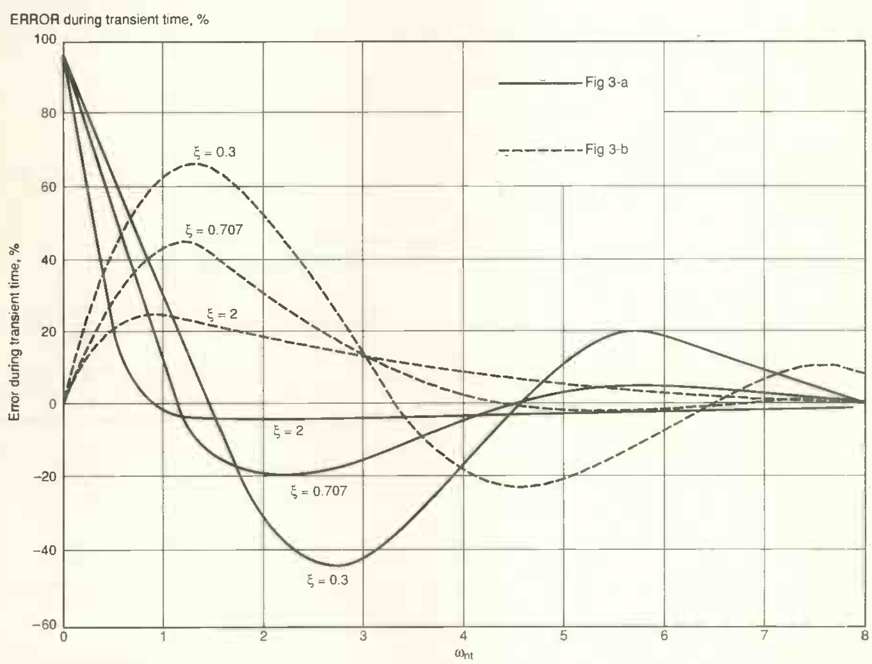


Fig. III. Phase errors in response to stepped change of input phase (a) and frequency (b) for varying values of x.

phase detectors: one exclusive-Or type and an edge-sensing FPD. Frequency range is 0-1MHz. Fast cmos extends frequency range to tens of MHz.

The PD in **Fig. 12** is widely used in frequency synthesizers with very low output phase noise. In principle, this is just a sample-and-hold device in the form of an analogue switch, storage capacitor and high-impedance buffer. This circuit is indispensable when switching noise at the output must be avoided; the S/H device copes well with this task, suppressing pulse noise by 46-60dB even without a filter. Tandem connection of two such devices, with the small penalty of the need to arrange the control signals, allows a noise reduction of 80-90dB.

Loop filters

PLL theory shows that dynamic response is determined by the filter between PD and VCO. Three types of loop filter are commonly used, their circuit diagram and frequency characteristics being shown in **Fig. 13**. It is much easier to optimise such parameters as dynamic error in transient processes in PLL systems with a lag-lead filter than with a lag filter. For maximum pulse suppression, the Cauer filter can be used with its trough corresponding to the comparison frequency.

the complex frequency (s) domain $\omega = s\Phi$. For the phase of the VCO output signal, write down the VCO transfer function $K_{vco}(s)$ in the complex-frequency domain:

$$K_{vco}(s) = K_{vco}/s \quad (3)$$

PLL output signal phase $\Phi_{out}(s)$ is given by

$$\Phi_{out}(s) = V_{lf}(s)K_{vco}/s, \quad (4)$$

or

$$\Phi_{out}(s) = (K_d K_{vco} F(s)(\Phi_{in}(s)/M - \Phi_{out}(s)/N))/s. \quad (5)$$

Substituting $\Phi_{err}(s)$ for $(\Phi_{in}(s)/M - \Phi_{out}(s)/N)$, the phase error between PLL input and output signals, then

$$\Phi_{out}(s) = K_d K_{vco} F(s)\Phi_{err}(s)/s. \quad (6)$$

Since the PLL transfer function T_{PLL} is

$$\Phi_{out}(s)/\Phi_{in}(s). \quad (7)$$

substituting (5) into (7) and omitting intermediate calculations, the expression for T_{PLL} becomes

Conclusions

- According to (3), there is an integrator ($1/s$) in the PLL that, in accordance with the theory of automatic control, results in zero error for the integrated parameter in the steady state. In this case, the parameter is the VCO frequency, which is why there is no frequency error between the signals at the PD inputs in the steady state; if there are no frequency dividers in the PLL, then $\omega_{in} = \omega_{out}$.
- There may be a phase error at the PD, decreasing when K_d is increasing.
- If the PD and VCO have no "inertia", the dynamic PLL response will be determined by the LF parameters and the coefficients M and N (the dividers work as a delay line). The dynamic responses, in this case, are speed of lock and tracking errors during transient processes caused by the input signal changing. Note that the presence of the inertial loop filter or the frequency dividers in the PLL slows down the loop reaction, which increases the duration of the transient processes. Therefore, one should avoid using large divider ratios in the PLL. Natural frequency ωn and damping factor ξ depend on the order of the filter, the second-order type shown in **Fig. II** being preferred. **Figure III** shows phase-error dependence for different values of the damping factor – (a) with stepped change of input phase and (b) with stepped change of input frequency. The error in Fig. III is expressed in percentage of the phase of $\Phi_{out}(s)$ – the VCO output signal in the steady state.

Equations already given determine the static response: steady-state frequency error in the PLL is zero (3); steady-state phase error is determined from (9).

- Figure III shows that, if damping factor ξ is less than 0.707, transient processes are oscillating and if more than 0.707, transient processes are aperiodic, the PLL natural frequency ωn exerting a direct influence on their duration.

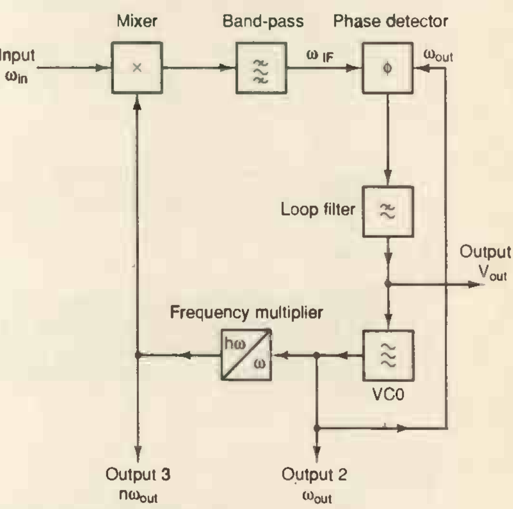
It is not possible to point to a "right" solution for these values, since PLL characteristics depend on the application. For example, in a frequency modulator with carrier frequency stabilisation, the PLL must not respond to the lowest modulation frequency (in high-quality broadcasting about 30Hz), so the LF is a low-pass filter with its cut-off frequency at about 2-5Hz, causing long-duration transients.

On the other hand, the oscillating transients are extremely undesirable when a PLL is controlling motor speed, synchronising two videotape recorders, for example, and in this case one chooses ξ to provide aperiodic transient processes.

- A PLL will reduce noise and disturbances in the VCO output only if their frequency lies within the noise bandwidth, this being $0.625\xi\omega_n$ for a second-order filter.

Applications

Figure IV is the block diagram of a frequency divider based on a PLL but with additional elements: a mixer and a multiplier – the formulae give the functions of the separate blocks. This PLL is useful for measurement of SHF (over 10GHz) signal source frequency (output 2 is used) and for frequency demodulation using output 1. The purpose of the PLL in Fig. V is exactly the opposite: this is a frequency multiplier, having a frequency divider in the feedback loop to make the output frequency higher than the input frequency. In other words, the circuit is that of a frequency synthesizer.

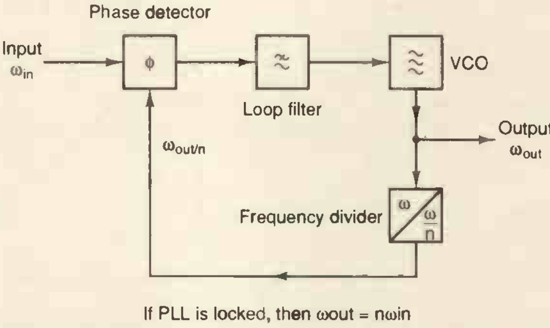


$\omega_{IF} = \omega_{in} - n\,\omega_{out}$
$\omega_{out} = \omega_{IF}$
$\omega_{out} = \omega_{in} - n\,\omega_{out}$
$\omega_{in} = \omega_{out} + n\,\omega_{out} = \omega_{out}(n + 1)$
$\omega_{out} = \omega_{in}/(n + 1)$
If $\omega_{in} = \omega_{in'} \pm \delta\omega_{in}$, then: $\omega_{out} = \frac{\omega_{in'}}{(n+1)} \pm \frac{\delta\omega_{in}}{(n+1)}$

Example: ω_{in} = 10 000 MHz ± 200 MHz
n = 19, ω_{IF} = 500 MHz

Output 2: ω_{out} = 500 MHz ± 10 MHz

Fig. IV. PLL in use as a frequency divider.



If PLL is locked, then ωout = nωin

Fig. V. PLL frequency multiplier. At lock, output frequency is n times input.

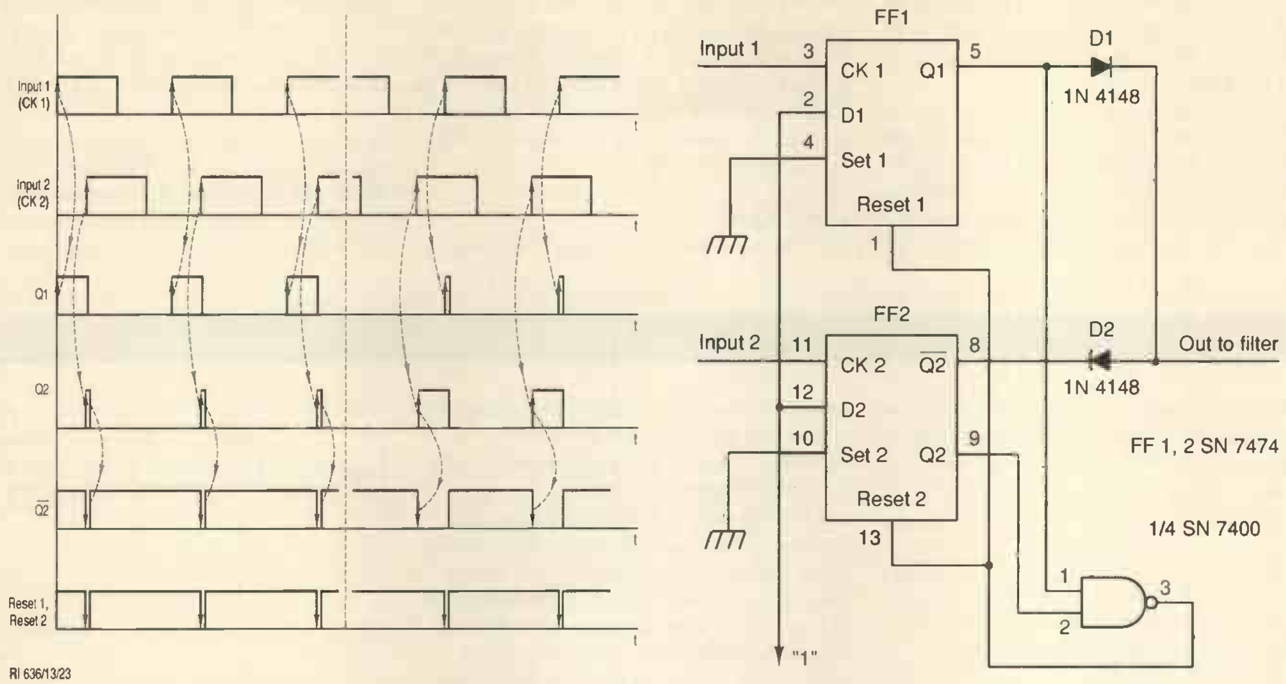
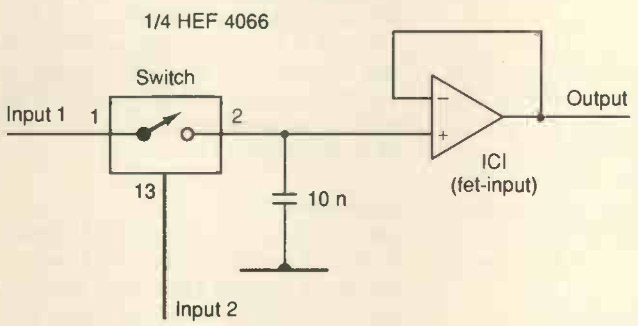
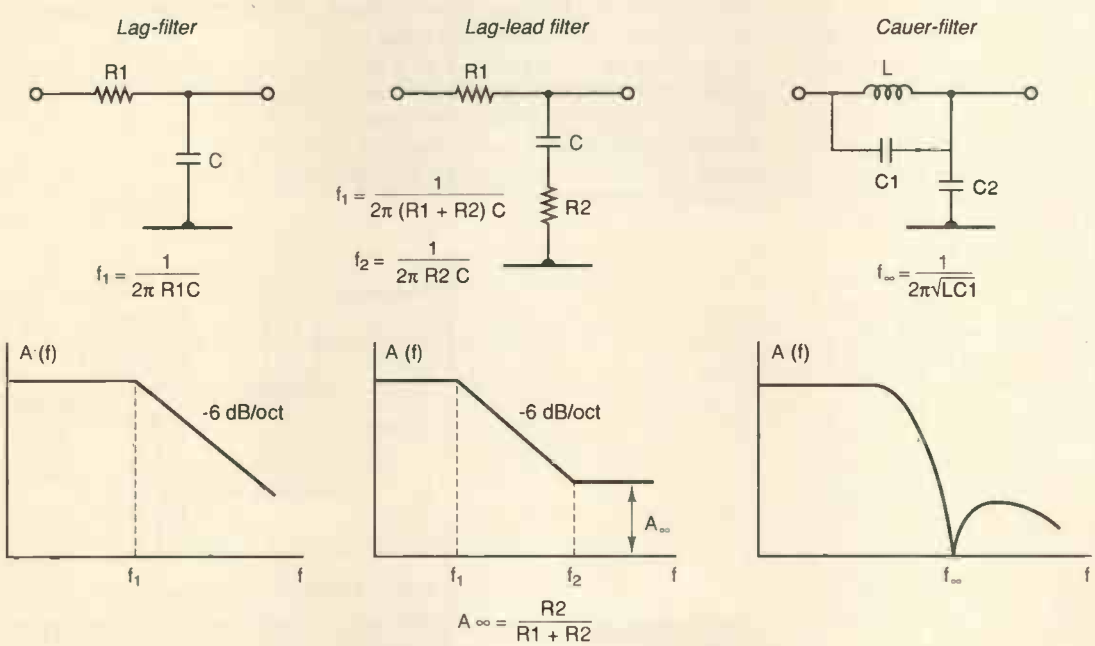


Fig. 11. Phase and frequency detector in TTL.
The time diagram explains operation of a frequency-phase detector. When Q_1=1 and $\bar{Q}_2$ (the inverted output of the trigger) =1 the capacitor is charging; when Q_2=0 the capcitor is discharging. The diagram shows the signals when the signal at input 1 passes ahead of the input 2 signal – and vice versa. The device works on the input pulse front. Duty cycle of the input signal is of no importance.

Fig. 12. Phase detection by sample-and-hold, providing good noise performance on both analogue and digital inputs. Capacitor must be a low-leakage type.





Below Fig. 13. Three types of filter used as loop filters in PLLs. Cauer type works well, but needs a coil.

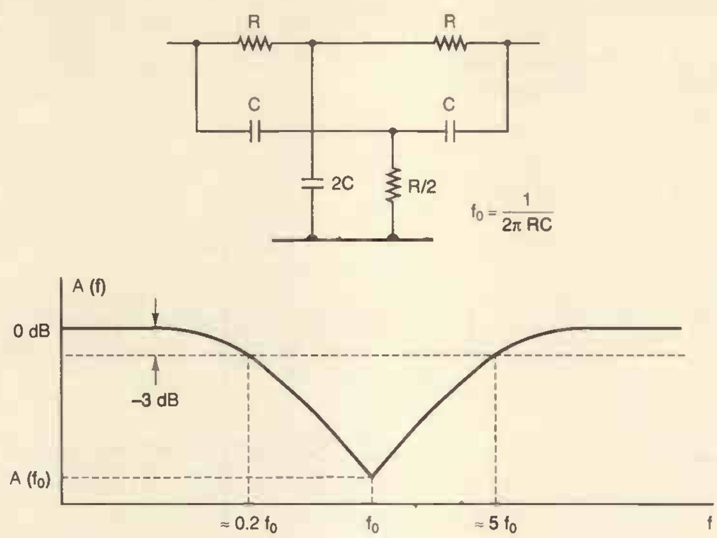


Fig. 14. To avoid coil in Cauer filter, this twin-t preceded by a lag-lead filter produces good results, the trough being better than 60dB with 2% components.

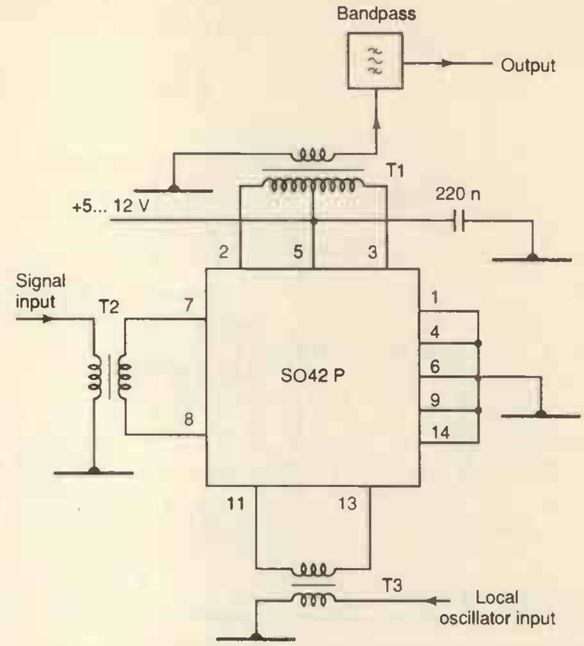


Fig. 15. SO42P used as a mixer up to 200MHz. Input frequency is rejected by 30dB on the mixed output level.

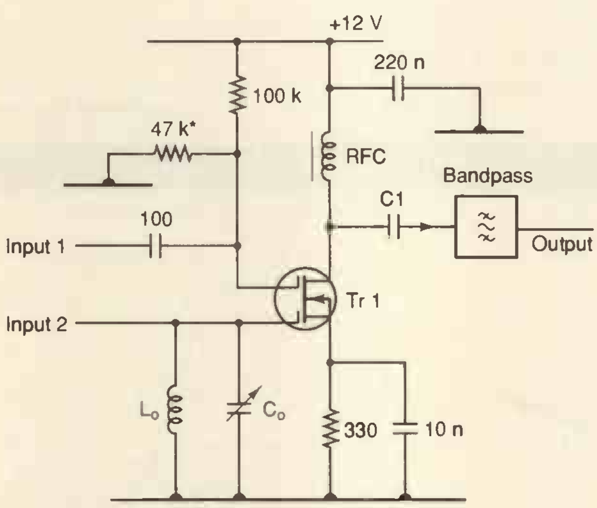


Fig. 16. Mixer based on a dual-gate fet. Both inputs may be wide-band if required.

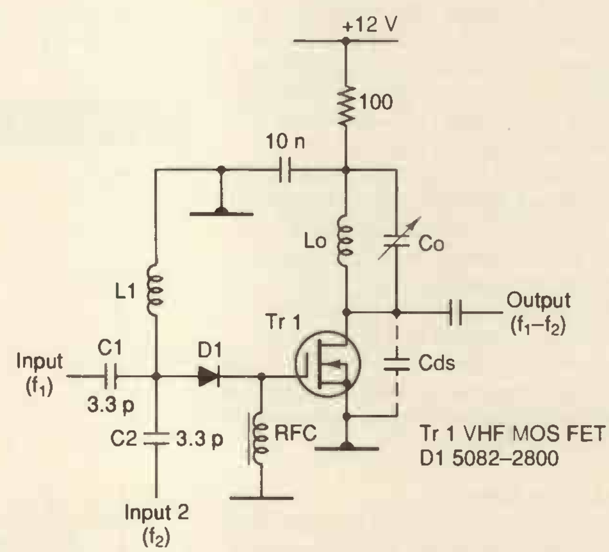


Fig. 17. Diode mixer for up to 2GHz input. Tuned amplifier reduces input-to-output signal feedthrough.

The only problem is the coil — not an attractive idea in microelectronic circuitry – so the compromise of a twin-T filter used after a lag-lead type is often used, as seen in **Fig. 14**. The twin-T is notable for the steepness of its trough, giving up to 46-53dB rejection when 2% components are in use – enough, in most cases.

However, a twin-T can lead to objectionable effects, because of its phase performance: an increase of dynamic error in transient processing and an increase in loop locking time. It is always advisable to analyse the system dynamic response before using the filter in a PLL. In many cases, designers use an op-amp lag-lead filter with R_2C in the feedback loop, which is the same as including an integrator in a PLL to improve dynamic characteristics. But the author always regards with caution the inclusion of active elements between PD and VCO, which is effectively imposing a noise source in the most noise-sensitive place in the PLL. If an op-amp is absolutely necessary, it should always be a low-noise type.

Mixers

The purpose of a mixer is to transfer the spectrum up or down in frequency. Theoretically, any non-linear device, with a filter on the output to select a given frequency, can be a mixer; circuits shown in Figs 8 and 9 function as mixers when provided with filters. A DBM has an advantage over other types in that it can suppress input signals in the output spectrum by up to 60dB at frequencies to 30MHz and by up to 30dB at 400-1000MHz. Symmetrical inputs and outputs improve the suppression of input signals and sometimes simplifies matching of the mixer with signal sources.

Figure 15 shows a basic circuit using a S042P as a broadband DBM, in which transformers T_1 and T_2 determine the working frequency. At low frequencies where transformers are inconvenient, it is possible to use asymmetrical inputs as in Fig. 9.

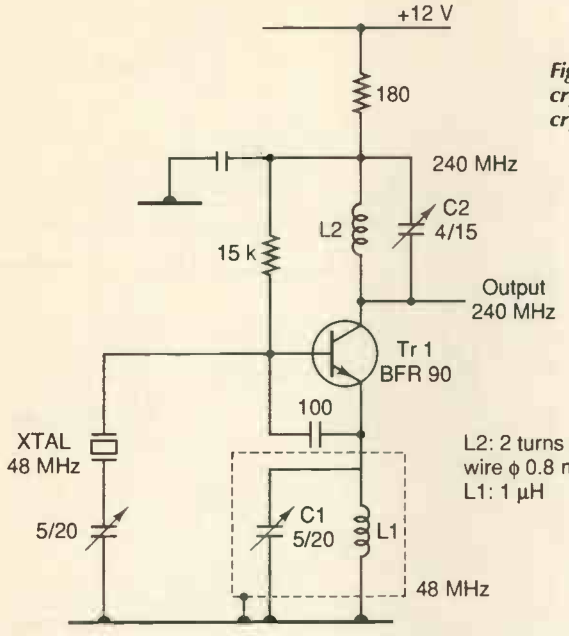


Fig. 18. Fifth-harmonic 240MHz crystal oscillator using a 48MHz crystal.

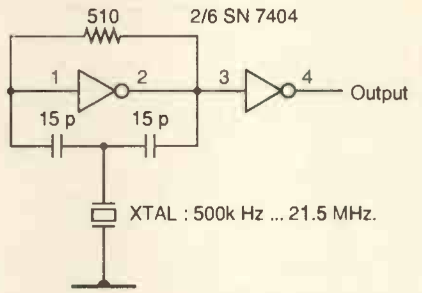


Fig. 19. Terminally simple crystal oscillator with two gates, usable up to about 21MHz.

In many cases it is convenient to use a dual-gate mosfet as a mixer, as shown in **Fig. 16**, often using a lower-level signal on gate 1 because of its steeper characteristic. A resistive voltage divider in gate 2 and the source resistor set a working point, which must be optimized to obtain maximum conversion gain. Input 2 is shown as narrow-band but it can be wide-band as at input 1.

Diode mixers with high conversion gain are often used at UHF/SHF. A fet or mosfet used as a tuned amplifier on the output of such a mixer avoids loading the mixer diode and attenuates input breakthrough to the output, since the transistor amplifies at a lowish IF. In the circuit diagram in **Fig. 17**, a tuned circuit is included in the drain of the fet. This circuit is good for frequencies up to about 2.5GHz; at higher frequencies, microwave versions of the circuit are feasible.

Crystal oscillators

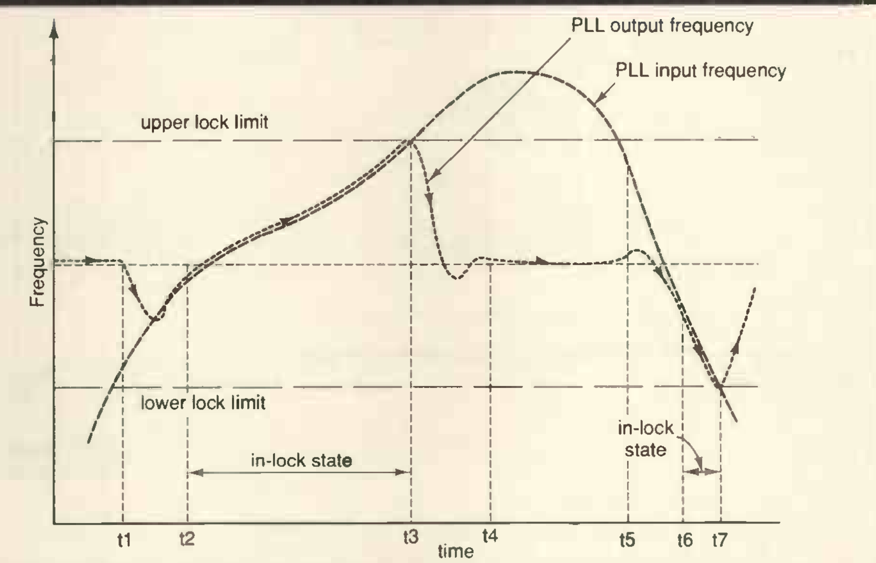
The crystal VCO shown earlier is quite good enough for use as a base oscillator or comparison frequency generator. **Figure 18** shows such an oscillator with a crystal working on the fifth harmonic (compare the circuit diagram in Fig. 7 with this one). The collector circuit is tuned to 48 x 5 = 240MHz; transistor Tr_1 works simultaneously as a crystal oscillator and a frequency multiplier by a factor of 5. A narrow-band output filter with one or two 240MHz sections will suppress unwanted harmonics, but must be shielded from the rest of the circuit.

Figure 19 shows a simple circuit that I have used many times at frequencies from 500kHz to 22MHz; the loose capacitive coupling between the crystal and the buffer ensures high stability.

Crystal oscillators rarely produce a signal exactly the same as that specified by the manufacturer; there are manufacturing tolerances and circuit strays to take into account. For this reason, most circuits include a means of frequency trimming, either inductive or capacitive. Temperature control is essential for extreme accuracy of the order of 0.5 to 0.05 parts per million; without it, 1ppm is about the best obtainable. ■

Continued next month...

LOCKING INTO THE LOOP



A PLL can acquire and track an input signal only if it remains within upper and lower frequency limits of the VCO. The expression "track" is relative. Where the frequency of the input signal varies rapidly ie faster than the time constant of the loop filter but within the capture range of the VCO, the result is frequency modulation of the output signal. This implies instantaneous phase excursions between input and output signals of more than 360°.

Where input signal excursions occur more slowly than the loop filter roll-off point and within the lock range, then phase modulation results (<360°).

The graph shows the progress of signal capture. At point t_1, the VCO output frequency begins to fall to meet the input frequency to the loop. After a period of overshoot between t_1 and t_2 (determined) by the loop damping factor, the system acquires lock and the signals track each other to t_3 where the upper lock limit is reached. The output frequency then falls to its free-running value at t_4 after some overshoot. It holds this value until t_5 where the input signal returns to within the lock range. The system maintains lock from t_6 to t_7 until the input signal goes outside the lower lock range causing the VCO output to return to its free-running value.

PC ENGINEERING

SLICK
system simulation on the PC

System requirements

386 or 486 PC
2Mbyte of extended ram
1Mbyte of hard disc space
Mouse
Windows 3

The power packed into today's 386s and 486s means that system simulation is now possible on the PC. Allen Brown wires up VisSim.

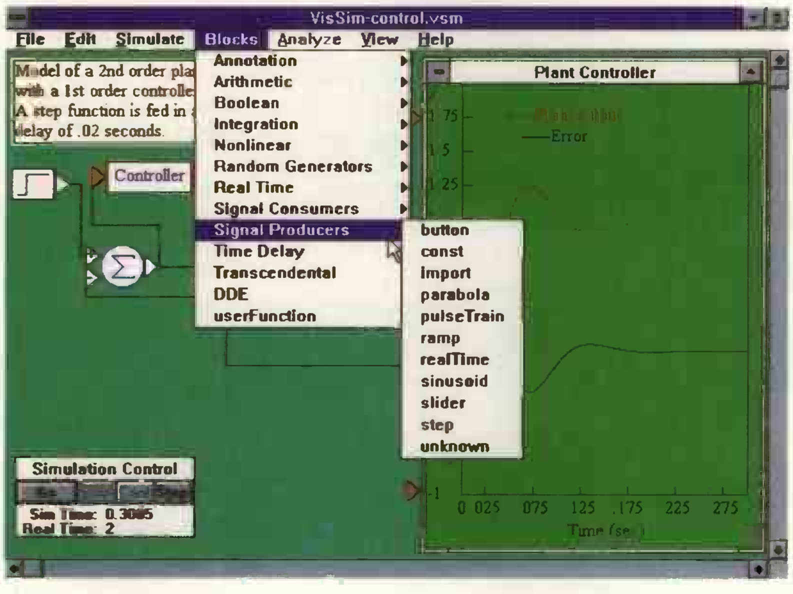
VisSim by Visual Solutions MA, is a system simulation package that can be used not only for electronic engineering but also to great effect in simulating small scale industrial plants, biological processes and mechanical systems. Engineers who feel more comfortable in the analogue world might also feel at home with the it as *VisSim* seems to simulate analogue problems with greater ease than discrete digital systems.

The package itself is graphics intensive – working under *Windows 3* – and includes dynamic data exchange (DDE) to enable the user to pass data between concurrent applications.

In the *VisSim* environment, each system component is represented as an icon (or block), accessed either from the drop-down menu or from one of the standard libraries. Blocks are wired together using the mouse – all in all quite a neat design.

There is a large variety of blocks to choose from, and each heading in the blocks drop-down menu leads to further choices (**Fig. 1**). Design structure is hierarchical, and blocks in a design may represent other sub-systems made up of other blocks, which in turn can represent further sub-systems. Structures are achieved with relative ease – provided the inputs and outputs of each hierarchical level are defined. Custom blocks can be stored to disk, then integrated into the current design when requried.

Fig. 1. There is a large variety of blocks to choose from, and each heading in the blocks drop-down menu leads to further choices.



A model of a Lorenz Attractor, a fashionable feature of Chaos theory, can be used to illustrate the feature (**Fig. 2**). By clicking the right hand button of the mouse on the Lorenz block, **Fig. 3** appears, showing how the Lorenz equations can be implemented from standard *VisSim* blocks. Also shown in Fig. 2 is a characteristic plot of the results of the simulation. Again the plot feature is treated as a block with its respective inputs shown on the left hand side. Most blocks have control parameters, accessed by mouse clicking.

Transfer functions

In the design of any analogue control systems, transfer functions (TFs) are all important. *VisSim* can implement TFs but its method is somewhat unconventional. TFs are designed by stringing together integrators (1/S blocks) with gain blocks. **Figure 4** shows how a first order TF is designed and there is the nagging feeling that the method is a bit cumbersome. To add to the problem the explanation given in the user's manual is inadequate, and is not helped by sloppy use of mathematical notation which mixes Laplace *s* within the same equations. Far more comfortable ways of generating TFs exist, such as entering the coefficients directly.

Much emphasis in the package is on use of integrators, and in fact six algorithms can be chosen to perform the integration process ranging from the humble Euler method to the "Bulirsh-Stoer with Richardson extrapolation". Each algorithm improves with accuracy (and therefore stability) with an increase in computational load – though a speed penalty may not even be noticeable on a 486 PC. But it does sometimes have to be taken into account, such as when *VisSim* is

Manual

VisSim's gradual learning curve means that a working proficiency can be gained in a relatively short time. Familiarity is eased by the information supplied in the Help file, and the Help tutorial also provides useful hints on the operating scheme.

Learning is also aided by the *Getting Started* chapter in the user's manual which enables the new user to gain a confident working knowledge. There is even a chapter given over to mouse technique, and from this point of view the manual is certainly very user friendly. This is clearly demonstrated when wiring icons (blocks) together which is accomplished with total ease by a feature not found on all software packages which use icons for system design. Overall, the user's manual is very readable, reflecting time and thought put into its layout.

Getting Started is supplemented with lots of useful diagrams, and each of the blocks is discussed in some detail.

The Manual is informative without being over-crammed with too much technical detail.

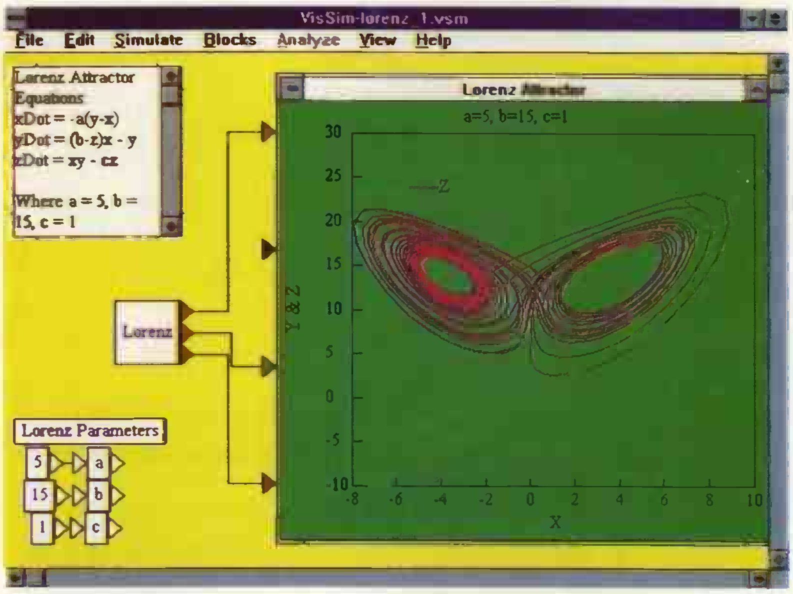


Fig. 2. A model of a Lorenz Attractor used to illustrate VisSim's features.

used in a real-time application. A user must have near total confidence in stability of a control algorithms.

Differential equations

VisSim has an interesting approach to solving differential equations. Based on the assumption that numerical integration is more stable than differentiation, each diferential equation is converted into a type of integral sequence.

Figure 5 demonstrates this, using simulation of a simple harmonic oscillator with damping whose equation of motion is

$$x''(t) = -1/M\,\{-Kx - Bx'(t)\}$$

On the left of Fig. 5 can be seen the second derivative x''(t) followed by an integrator (1/S) to give x'(t), followed by the second integrator to produce x(t). So the equation is implemented schematically, and is quite an impressive feature of *VisSim*, proving to be very useful especially in solving non-linear differential equations. Not only can the solution be plotted in time but phase plots can also be generated with relative ease.

Library features

One of the attractive features of *VisSim* is the number of sample application files supplied. Samples are particularly useful to the first time user who is finding out the package's capabilities. Not only can the user browse through the varied selection of sample applications, but there is a set of additional libraries each containing ready-designed compound blocks. Six compound block libraries span:

- Controls – analogue and digital simulation blocks such as PID controllers;
- A signal generator – simulating a variety of waveforms;
- Tools – a selection of calculation routines, such as average estimator, phase difference and RMS;
- Tutsim – a choice of functions used in Z domain anlysis; analogue filters (Butterworth and Chebychev filter designs), and
- Electro mechanical – A-to-Ds, three phase motor models and armature controlled DC motors.

The compound blocks can prove quite valuable, and once a user becomes familiar with what's available, they can be integrated into system design as and when they are needed.

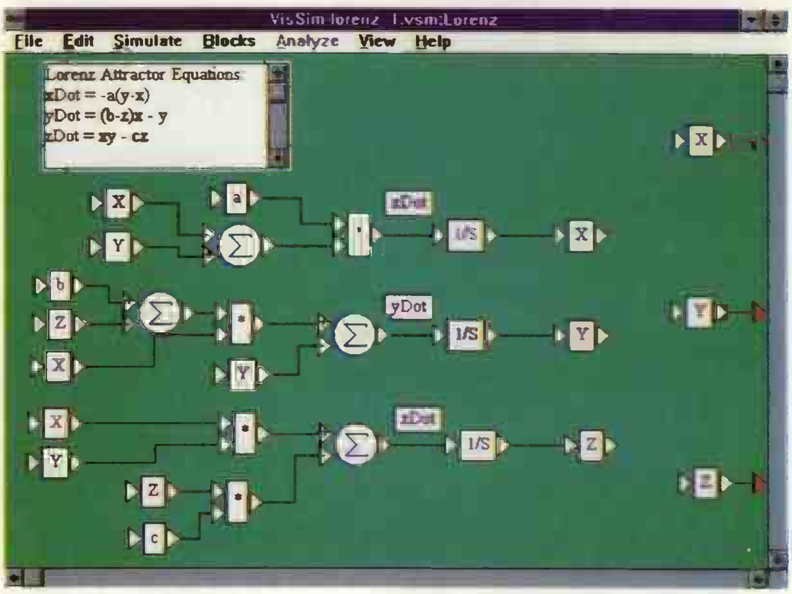


Fig. 3. The screen result of clicking the right hand button of the mouse on the Lorenz block, showing how the Lorenz equations can be implemented from standard VisSim blocks.

Add-on (essentials?)

Irritatingly, analyser options from the menu command bar can only be used by purchasing the extra *VisSim/Analyze* package. The analyser option contains the tools required for frequency analysis and state space modelling (Boyd and Root Locus). Surely no engineer concerned with the stability of dynamic systems would consider these as optional extras.

The functions are essential and should not be marketed as separate add-ons. Some information about the Analyser is given in Help. But not even to mention it in the user's manual is quite an oversight.

VisSim can also be used real-time, when the appropriate expansion cards are inserted in the PC. This is now a standard feature with most software packages which perform data acquisition and processing operation. But with *VisSim*, the real-time feature requires an another optional extra (*VisSim/RT*).

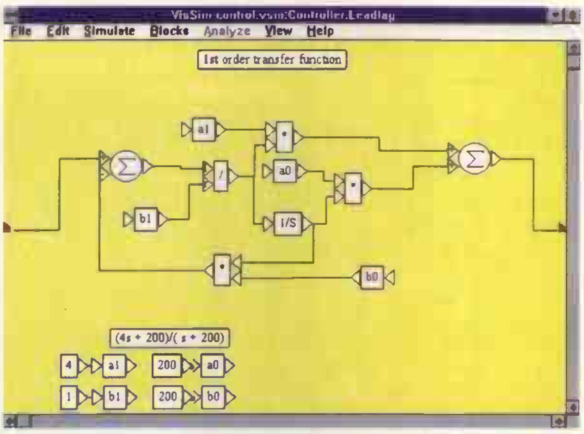


Fig. 4. VisSim has an unusual way of implementing transfer functions and its method is somewhat unconventional.

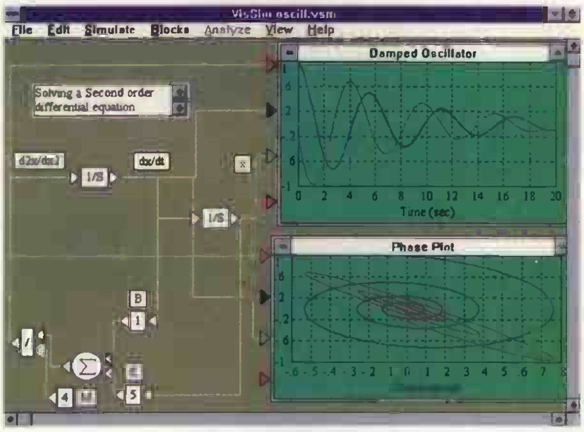


Fig. 5. Based on the assumption that numerical integration is more stable than differentiation, each diferential equation is converted into a type of integral sequence. On the left can be seen the second derivative x''(t) followed by an integrator (1/S) to give x'(t), followed by the second integrator to produce x(t).

It would be useful to see these libraries updated by the manufacturer as more compound blocks become available from other users. *VisSim* will make a useful addition to a working suite of software design tools. It is well designed and remains stable (I have not experienced it falling over). Only two aspects of the package seem a little unattractive: the method of implementing transfer functions, and the lack of frequency analysis features – unless the add-on module is bought.

The transfer function aspect could be a matter of personal taste. But the analysis situation should be remedied with the issue of the *VisSim/Analyze* as standard, not as an extra.

These reservations apart, *VisSim* is undoubtedly a valuable design asset. ■

Supplier Details

Adept Scientific, 6 Business Centre West, Avenue One, Letchworth,
Herts. SG6 2HB. 0462 480055. Cost
Personal version, £495 (limited to 255 blocks), Full version £1595 plus VAT. Analyse option for personal version £295, for full version £695.

An end to spurious oscillations

*Oscillations are rife in analogue circuit design. Robert Pease looks at some of the ways in which problems can be solved. Serialised from his book **Troubleshooting Analog Circuits***

Oscillations are the ubiquitous bugaboos of analogue circuit design. Not only can you encounter oscillating op amps, as described last month, but also oscillating transistors, switching regulators, optoisolators, comparators, and buffers. If you think about it, latched-up circuits are just the opposite of oscillating ones, so I have included them here, too.

Although I obviously cannot tell how to solve every kind of oscillation problem, I will give some general principles and then notes on what can go wrong with various components, including comparators and buffers. This information, along with a few suggested procedures and recommended instruments, will provide a good start. Here are some of the types of oscillations that can pop up unexpectedly:

- Oscillations at very high frequencies – hundreds of megahertz – because of a single oscillating transistor;
- Oscillations at dozens of megahertz arising from stray feedback around a comparator;
- Oscillations at hundreds of kilohertz because of an improperly damped op amp loop, an unhappy linear voltage-regulator IC, or inadequately bypassed power supplies.
- Moderate-frequency oscillations of a switching-regulator loop because of improper loop damping.
- Oscillations at "50Hz" or at "100Hz," or similar line-related frequencies.
- Low-frequency oscillations coming from physical delays in electromechanical or thermal servo loops.

As these general descriptions indicate, the frequency of an oscillation is a good clue as to its source. An electric motor loop can't oscillate at 10MHz, and a single transistor can't (normally) rattle at 100Hz. So when an engineer complains of an oscillation, the first question is: "at what frequency?"

Even though the frequency is often a good clue, engineers often fail even to notice what the frequency was. This omission tends to make troubleshooting by phone a challenge.

At very high frequencies, 20-1000MHz, the layout of a circuit greatly affects the possibility of oscillation. One troubleshooting technique is to slide a finger around the circuit and see if at any point an oscillation improves or worsens. Remember, knowing how to make an oscillation stronger is not worthless knowledge – that information can provide clues about how to make the oscillation disappear.

I remember being very impressed when a colleague showed me that some of the earliest IC amplifiers had a tendency to self-oscillate at 98MHz with certain levels of output voltage. Putting a grid-dip oscillator nearby caused increases or decreases in the problem, when its frequency was near 98MHz. At that time I didn't have a 100MHz scope, but I could see the rectified envelope of these high frequency oscillations on a 25MHz scope.

So, if you see a circuit shift its DC level just because you move your finger near a transistor, you should become suspicious of high-frequency oscillations. Of course, you would never "slide your finger around" in a circuit with high or lethal voltages...

One of the easiest ways to inadvertently cause a very high frequency oscillation is to run an emitter-follower transistor (even a nice, docile type such as a *2N3904*) at an emitter current of 5 or 10mA with the base grounded to RF. In such a case, you could easily get an oscillation at a few hundred megahertz.

Although a good 100MHz scope cannot spot this kind of oscillation, the resulting radiated noise can make other circuits to go berserk and can cause an entire system to fail tests for radiated electromagnetic noise. For example, when the first personal computers were being designed, designers needed a reset function for their processor.

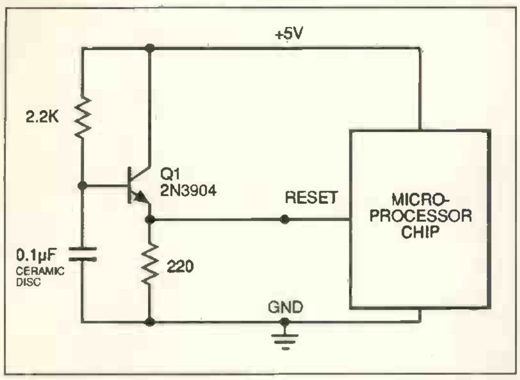


Fig. 1. This was a popular circuit for the reset function until engineers discovered how badly it oscillated.

Several designers decided (quite independently) to use the simplest, cheapest possible reset circuit, as shown in **Fig. 1**. When they had finished their designs they sent the prototype computers to be approved by the FCC; the designs all failed badly. Why? Because the little transistor would run at over 10mA and, with a bypass capacitor at its base, the transistor would oscillate at a very high frequency.

The frequency was so high that none of the designers had noticed it, but as the transistor sprayed around a lot of RF energy up at a couple of hundred megahertz, the FCC examiners noticed it, causing the computers to fail the tests for radiated RFI. They all had to go back and fix it.

For such an emitter follower, a 50 or 100Ω carbon resistor directly in series with the base of the transistor (and not a couple of inches away) can cure this tendency to oscillate. Sometimes a small ferrite bead is more suitable than a resistor because it will degrade the transistor's frequency response less.

Oscillations crop up

Not all problematic oscillations are high-frequency ones. An unstable switching regulator feedback loop can oscillate at low frequencies. For troubleshooting switching-regulator feedback loops, I first recommend a network analyzer to save troubleshooting time.

Optoisolators in switching regulators are another possible cause of oscillation trouble due to their wide range of DC gain and AC response. A switching-regulator IC, on the other hand, is not as likely to cause oscillations, because its response would normally be faster than the loop's frequency.

However, the IC is never absolved until proven blameless. For this reason, you should have an extra module with sockets installed just for evaluating these funny little problems with differing suppliers, variant device types, and marginal ICs. You might think that the sockets' stray capacitances and inductances would do more harm than good, but in practice, you can learn more than you lose.

The design of a slow servo mechanism, such as that in **Fig. 2**, can best be analysed with a strip-chart recorder because the response of the loop is so slow. (A storage scope might be OK, but a strip-chart recorder works better for me.) You might wish to analyze such a servo loop with a computer simulation, such as *Spice*, but the thermal response from the heater to the temperature sensor is strictly a function of the mechanical and thermal mounting of those components. This relationship would hardly be amenable to computer modelling or analysis.

Comparators can misbehave

Saying that a comparator is just an op amp with all the damping capacitors left out is an over simplification. Comparators have a lot of voltage gain and quite a bit of phase shift at high frequencies; hence, oscillation is always a possibility. In fact, most comparator problems involve oscillation.

Slow comparators, such as the familiar *LM339*, are fairly well behaved. If you design a PC-board layout so that the comparator's outputs and all other large, fast, noisy signals are kept away from the comparator's inputs, you can often get a good clean output without oscillation. However, even at slow speeds, an *LM339* can oscillate if you impress a slowly shifting voltage ramp on its differential inputs. Things can get even messier if the input signals' sources have a high impedance (>>10kΩ) or if the PC-board layout doesn't provide guarding.

In general, then, for every comparator application, you should provide a little hysteresis, or positive feedback, from the output back to the positive input. How much? I like to provide about two or three times as much hysteresis as the minimum amount it takes to prevent oscillation near the comparator's zero-crossing threshold. This excess amount of feedback defines a safety margin.

My suggestion for excess hysteresis is only a rule of thumb. Depending on your applica-

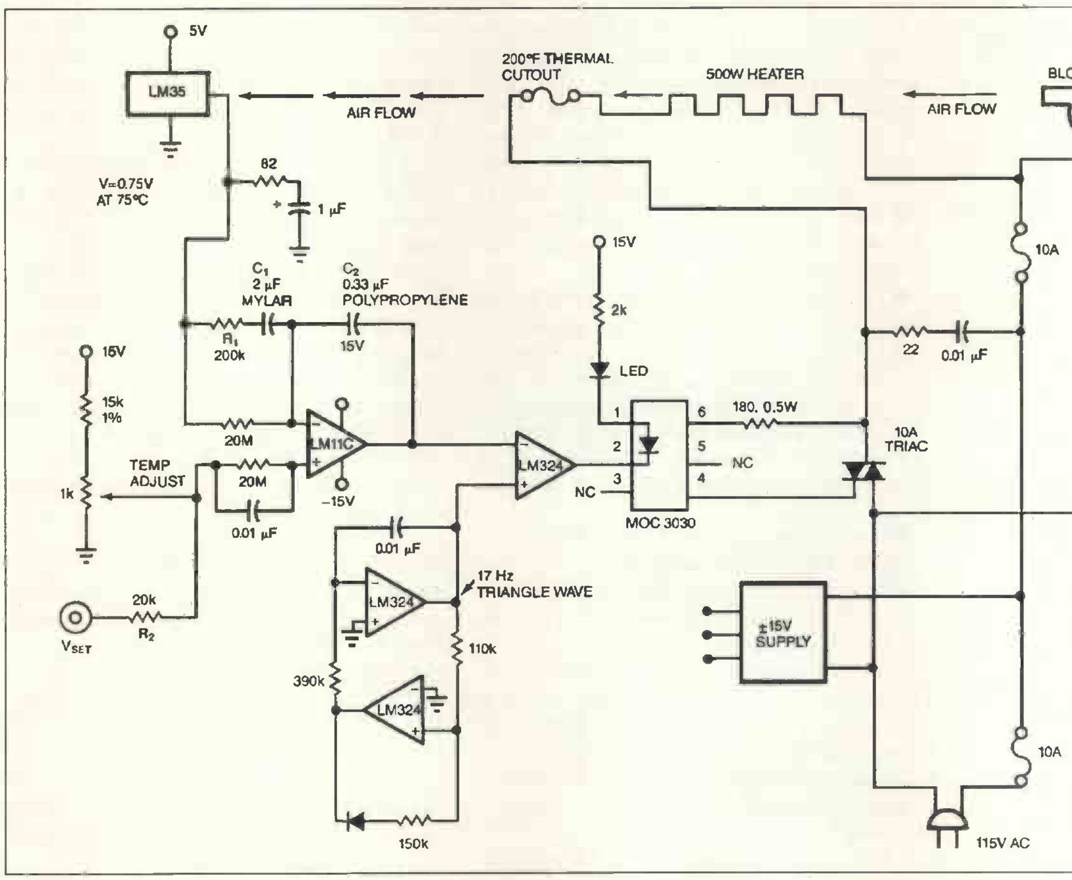


Fig. 2. Stabilising this heater's slow servo loop and choosing the proper values for R1, C1 and C2 involved applying a 1-Vp-p, 0.004Hz square wave, V_{SET} to R_2 and observing the LM11C's output with a strip-chart recorder.

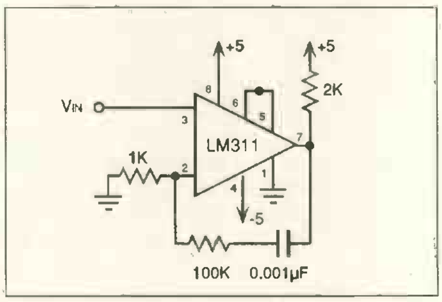


Fig. 3. This zero-crossing detector has no DC hysteresis but 50mV of AC-coupled hysteresis.

tion, you might want to use even more hysteresis. For example, a comparator in an RC oscillator may operate with 1, 2 or 5V of hysteresis, which means you can always use more than my "minimum amount" of excess hysteresis. Also, if you have a signal with a few mV of noise riding on top of it, the comparator that senses the signal will often want to have a hysteresis range that is two or three times greater than the worst-case noise.

Just the right touch

Because you can drastically alter comparators' performance just by touching the circuit with your finger, you should be prepared for the probability that your safety margin will change, for better or worse, when you go from a breadboard to a printed-circuit layout. There's no way you can predict how much hysteresis you'll need when your layout changes, so you just have to re-evaluate the system after you change it.

For faster comparators, such as the *LM311*, everything gets even touchier, and the layout is more critical. Yet, when several people accused the *LM311* of being inherently oscillatory, I showed them that with a good layout, it is capable of amplifying any small signal, including its own input noise, without oscillating and without any requirement for positive feedback.

One special precaution with the *LM311* is tying the trim pins (5 and 6, normally) together to prevent AC feedback from the output (pin 7, normally), because the trim pins can act as auxiliary inputs. The *LM311* data sheet in the National Semiconductor Linear Databook has carried a proper set of advice and cautions since 1980, and I recommend tins advice for all comparators.

With comparators that are faster than an *LM311*, I find that depending on a perfect layout alone to prevent oscillation just isn't practical. For these comparators, you'll almost certainly need some hysteresis, and, if you are designing a sampled-data system, you should investigate the techniques of strobing or latching the comparator.

Using these techniques should ensure that there is no direct path from the output to the inputs that lasts for more than just a few nanoseconds. Therefore, oscillation may be avoided. Granted, heavy supply bypassing and a properly guarded PC-board layout, with walls to shield the output from the input, may help. But you'll probably still need some hysteresis.

For some specialized applications, you can gain advantages by adding AC-coupled hysteresis in addition to or instead of the normal DC coupled hysteresis (see **Fig. 3**). For example, in a zero-crossing detector, if you select the feedback capacitor properly, you can get zero effective hysteresis at the zero-crossover point while retaining some hysteresis at other points on the waveform.

The trick is to let the capacitor's voltage decay to zero during one half-cycle of the waveform. But make sure that your comparator with AC-coupled hysteresis doesn't oscillate in an unacceptable way if the incoming signal stops.

Noisy comparators

Most data sheets don't talk about the noise of comparators (with the exception of the *NSC LM612* and *LM615* data sheets), but comparators do have noise. Depending on which unit you use, you may find that each comparator has an individual "noise band".

When a differential input signal enters this band slowly from either side, the output can get very noisy, sometimes rail-to-rail, because of amplified noise or oscillation. The oscillation can continue even if the input voltage goes back outside the range where the circuit started oscillating. Consequently, you could easily set up your own test in which your "data" for offset voltage, V_{OS}, doesn't agree with the manufacturer's measured or guaranteed values. Indeed, it can be tricky to design a test that *does* agree.

For my tests of comparator V_{OS}, I usually set up a classic op-amp oscillator into which I build a specific amount of hysteresis and a definite amount of capacitance, so that the unit will oscillate at a moderate, controlled frequency.

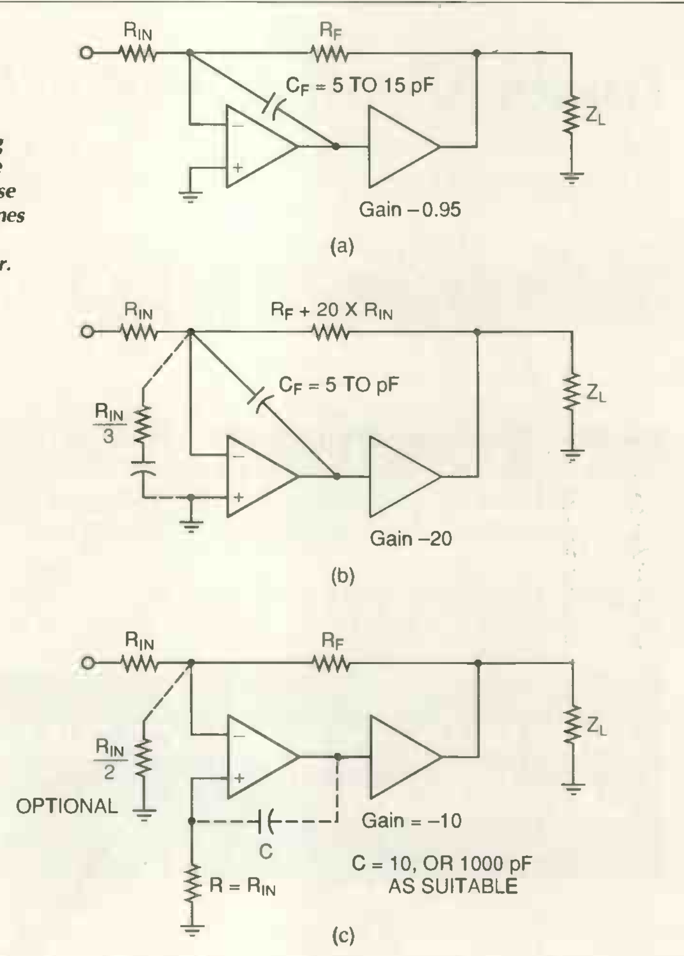
Another way to avoid V_{OS} trouble with comparators is to use a monolithic dual transistor as a differential-amplifier preamplifier stage ahead of the comparator. This preamp can add gain and precision while decreasing the stray feedback from the output to the input signal.

Oscillating buffered circuits

Any circuit that adds current gain can oscillate – even a buffer. Let's agree that a buffer is some kind of linear amplifier that has a lot of current gain. Some have a voltage gain around 0.90 or 0.95. Others have gains as high as 10 or 20 because their outputs must swing 50 or $100V_{p\text{-}p}$ or more. Even emitter followers, which you wouldd expect to be very docile because their voltage gain is less than 1, have a tendency to "scream" or oscillate at high frequencies. So, whether you buy or make a buffer, watch out for tins problem.

Also, a buffer can have a high-frequency roll-off whose slope increases suddenly at 40 or 60MHz, contributing phase shift to your loop, back down at 6 or 10MHz. You can beat

Fig. 4. Depending on the gain of the buffer, you can use these three schemes to stabilise a buffered amplifier.



this problem, but you have to plan. A buffer can also add a little distortion, which the op amp cannot easily cancel out at moderate or high frequencies.

Since buffers don't usually have a spec on this distortion, beware. Also, if you're running the output's quiescent bias current as Class AB, you must be sure that the DC operating current is stable and not excessive. You must set it high enough so that you don't get distortion but not so high that power consumption becomes excessive.

One of my standard procedures for stabilising a unity-gain follower stage is to put feedback capacitance around just part of the loop (**Fig. 4**). This circuit tolerates capacitive loads, because the buffer decouples the load while the feedback capacitor around the op amp provides local stability.

Most unity-gain buffers, whether monolithic, hybrid, or discrete, are unstable with inductive sources, so keep the input leads short. A series resistor may help stability, as it does for the *LM310*, but it will slow down the device's response. When your buffer provides a lot of extra voltage gain, you must make sure that the gain rolls off in a well-engineered way at high frequencies, or the loop will be unstable. If the buffer-amplifier has a positive gain, as in **Fig. 4b**, you can use capacitive feedback around the main amplifier. But when the buffer-amplifier has a gain of –10 (**Fig. 4c**), you may want to apply a feedback capacitor from the input of the buffer-amplifier (the output of the op amp) to the non-inverting input of the op amp.

In some cases, you can achieve stability by putting a series RC damper from the non-inverting input to ground to increase the noise gain, but this trick doesn't always work. Damping this loop is tricky, because there is so much gain stacked up in cascade. The feedback capacitor to the negative input makes this safe and easy, however.

Fail safes

If you have any doubt that your anti-oscillation fixes are working, try heating or cooling the suspected semiconductor device. In rare cases, passive components may be sufficiently temperature-sensitive to be at blame, so think about them, too. Even if a circuit doesn't get better when heated, it can get worse when cooled, so also take a peek at it while applying some freeze mist.

My point is that merely stopping an oscillation is not enough. You must apply a tough stimulus to the circuit and see whether your circuit is close to oscillation, or safely removed from any tendency to oscillate. This stricture applies not only to regulators but also to all other devices that need oscillation-curing procedures.

For example, if a 47Ω resistor in the base of a transistor cures an oscillation, but 24Ω doesn't, and 33Ω doesn't, and 39Ω still doesn't, then 47Ω is a lot more marginal than it seems. Maybe a 75Ω resistor would be a better idea – just so long as 100 or 120 or 150Ω resistors are still safe.

In other words, even though wild guesses and dumb luck can sometimes cure an oscillation, you cannot cure oscillations safely and surely without some thoughtful procedures. And somebody who has an appreciation for the "old art" will probably have the best results.

I certainly do not want to say that technicians can't troubleshoot oscillations simply because they don't know the theory of why circuits oscillate – that's not my point at all. I will only argue that a green or insensitive person, whether a technician or an engineer, can fail to appreciate when a circuit is getting much too close to the edge of its safety margins for comfort.

Conversely, everyone knows the tale of the old-time unschooled technician who saves the project by spotting a clue that leads to a solution when all the brightest engineers can't guess what the problem is.

■

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THE CHIPS ARE DOWN FOR CORDLESS PHONES

The digital cordless telephone, known as CT-2, has promised so much for so long without success that one might be forgiven for thinking it would never happen, at least not at a realistic price. But CT-2 may be about to supersede its aging ancestor, the analogue cordless telephone.

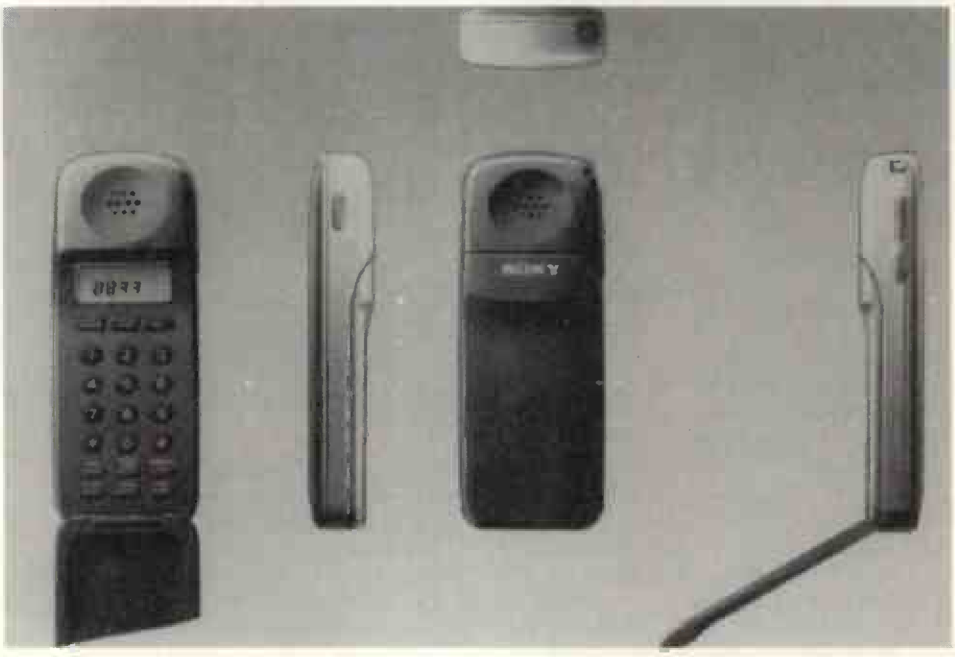
CT-2 designers now have access to low cost, highly integrated semiconductor components. A handset requires just two or three ICs and a handful of discrete components. The new, smaller CT-2's will use fewer batteries, but most importantly they will be far easier to make, and that will cut prices.

Japanese companies such as Sony and Panasonic have quickly seen the significance of the new cordless telephone chips, and stated their intention to produce digital cordless telephones by the million over the next few years.

Advanced Micro Devices is the first semiconductor house to combine all the baseband features for the CT-2 on a single chip which it expects to sell for less than $25. It developed the chip in collaboration with Sony who will use it in its first CT-2 product which will be on sale in the UK high street by 1994.

The *Am79C410* CT-2 controller will be capable of replacing as many as five integrated circuits used in first generation CT-2 handsets. The device includes a speech coder based on the standard ADPCM (adaptive differential pulse code modulation) burst mode logic for the standard CT-2 common air interface (CAI) radio modulation protocol. In addition there is an audio interface with delta-sigma A/D converters and capable of driving a 16-ohm loudspeaker, an 8-bit microcontroller and circuitry to drive a 6 x 6 key pad.

AMD's achievement is that it has combined the ADPCM codec and the burst mode logic with a microcontroller into a single 100-pin package, smaller than the 386 microprocessor.



Small but how cheap? The semiconductor companies hope to build a chip set for $30 by 1995 but the handsets will always be expensive in comparison to the wired variety – particularly if one considers that a similar set of ICs must be included in every base station.

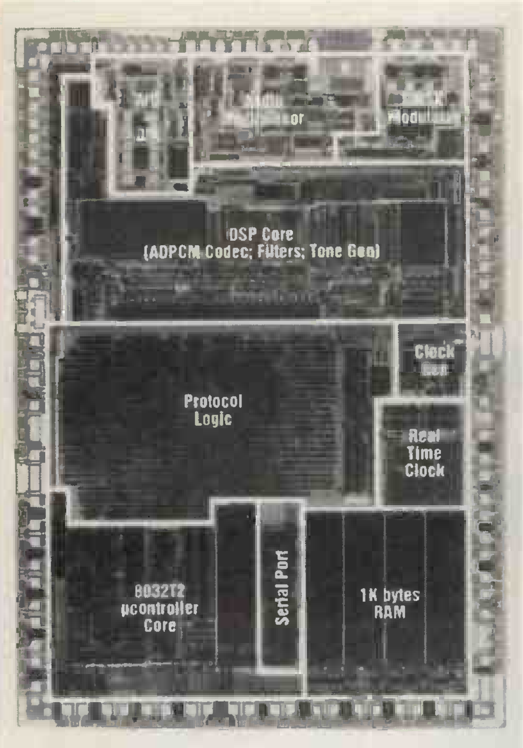
The ADPCM speech codec creates a 32Kbit/s time division multiple access (TDMA) channel to carry the call. The two-way call occupies a single channel frequency with transmit and receive signals interleaved in what is known as a ping-pong technique. There is also a control channel for signalling between handset and base station of up to 16Kbit/s.

AMD has trimmed the performance of the digital signal processing core used for the speech coding in order to squeeze in the 9000 gates of burst mode logic and the bulky delta-sigma A/D converters. Noise suppression has been added to the ADPCM; the company had no reservations about combining the inherently noisy delta-sigma A/D converters with the small signal speech coding circuits.

In place of a general purpose DSP running at 20MHz, the optimised core runs at 10MHz cutting the necessary instruction cycles and programme memory by over 60 per cent.

The benefit is a smaller design and lower power consumption. Working from a single 3V supply its consumes typically 80mW of power during a call and 900µW between calls.

An on-chip 6th order Bessel filter with a 14.4kHz cut-off is used to convert the baseband data into the analogue Gaussian minimum shift key (GMSK) spectral output for the radio interface. And it is the RF circuits which are arguably the hardest part of the handset design to be compressed into one or two chips.



Chip layout of AMD's Am 79C410 telephone circuit

Sony assisted with the *AM79C410* burst mode logic and has also developed a CT-2 RF section in three ICs. But Sony is confident that the design will fit into a single IC if demand justifies the cost of integration.

The RF chips comprise an up/down converter, demodulator and radio transceiver. An external SAW filter is used to generate the CT-2's 866MHz carrier frequency. Sony has set itself a target price of less than $24 for the three.

AMD is not the only chip maker to offer single chip digital cordless telephone designs. VLSI Technology will have, later this year, a two chip CT-2 handset design which uses a separate microcontroller.

VLSI is against embedding the microcontroller because it restricts the designer's freedom in choice of processor. The microcontroller controls the operation of the telephone and may be used to add features which will differentiate products in the market.

While AMD and Sony are attracted by the low cost CT-2 handset market, other semiconductor makers have concentrated on single chip designs for the proposed European DECT cordless telephone standard. The effect of this is that affordable DECT products will be introduced between one and two years earlier than originally anticipated.

Sierra Semiconductor has implemented the more sophisticated DECT handset baseband design in two ICs – *SC14400/1* – which it plans to sell at less than $10. Like AMD, Sierra has trimmed the data processing functions to the bare minimum to achieve the integration.

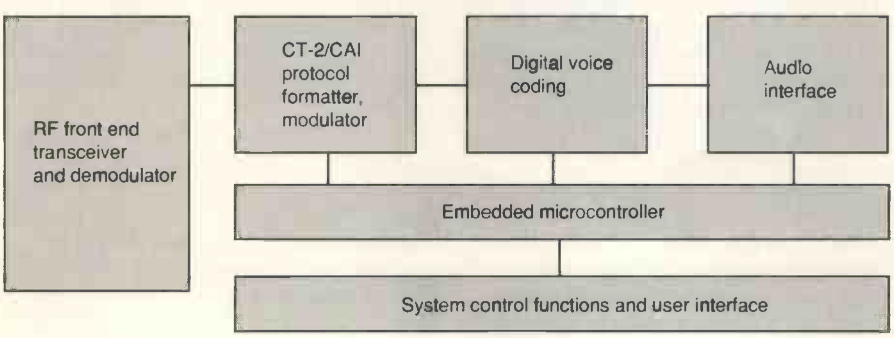
It also uses a 0.6μm cmos process. As a result the company claims the two chips occupy a third of the board space of anything currently available. Sierra also says the chips may be used for CT-2 designs.

Sierra is also collaborating with an RF specialist for the 1.8GHz DECT radio transceiver. National Semiconductor which has as yet only succeeded in compressing the DECT RF section into six components, confidently predicts that total DECT handset component cost will fall to $30 by 1995.

By the end of the year we will perhaps know whether this is achievable. Californian mixed-signal specialist Pacific Communications Sciences, recently bought by Cirrus Logic, claims to have a 1.8GHz DECT design which will require only one baseband IC and three RF chips. ■

Richard Wilson, *Electronics Weekly.*

The operating blocks of a CT-2 telephone handset. Chip designers had to resolve problems of mixing low level analogue signals with the electrically noisy digital signal processing when attempting to integrate all the non-radio functions onto a single chip.



The cordless connection

Cordless phone designers are helped by the technical similarities between what were once thought to be incompatible telephone design specifications adopted by different countries.

The first was the CT-2, pioneered in the UK as early as 1985, originally designed to replace the inferior analogue CT-1 in the home and office. It is also being used in low cost public mobile telephone services in Britain, Hong Kong, Singapore, France and Holland. CT-2 uses a highly spectrum efficient Gaussian modulation scheme and time division of the radio channel. It operates in 4MHz around 866MHz.

Earlier this year the Canadian authorities adopted a more sophisticated variant of the CT-2 design, known as CT-2 Plus. This operates at a slightly higher frequency and has twice the number of radio channels used in the UK. CT-2 Plus is more suited to mobile communications because, unlike basic CT-2, it offers two-way communications in the street and call hand over so the caller does not need to be stationary to make a call.

A pan-European design called DECT – the European digital cordless telephone – was agreed last year. Operating at 1.8GHz with more radio channel capacity and supporting data as well as voice traffic, DECT has been designed for the office. But if handset costs are low enough DECT will also be used in the home.

Finally the Japanese have their own digital cordless telephone design operating at 1.6GHz which is known as the personal handy phone (PHP).

All these designs operate at different radio frequencies, but essentially use the same method of digitally compressing the speech for transmission.

The voice compression coding scheme, which is known by the international standard G.721, is so efficient in encoding voice band signals for digital radio transmission that it has become almost universally adopted. It uses a technique known as adaptive differentiation pulse code modulation (ADPCM) which halves the channel bandwidth required by squeezing the standard telephony μ-law 64Kbit/s bit stream into a 32Kbit/s bit stream.

The G.721 speech codec is only one element of the necessary baseband processing design required in the handset. In addition, there are some 9000 gates of burst mode logic, which formats the compressed data stream for radio transmission; clock recovery circuits and a delta-sigma analogue to digital converter between the digital codec and the audio interface, the mouthpiece and ear-piece.

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GPS

6: Applications

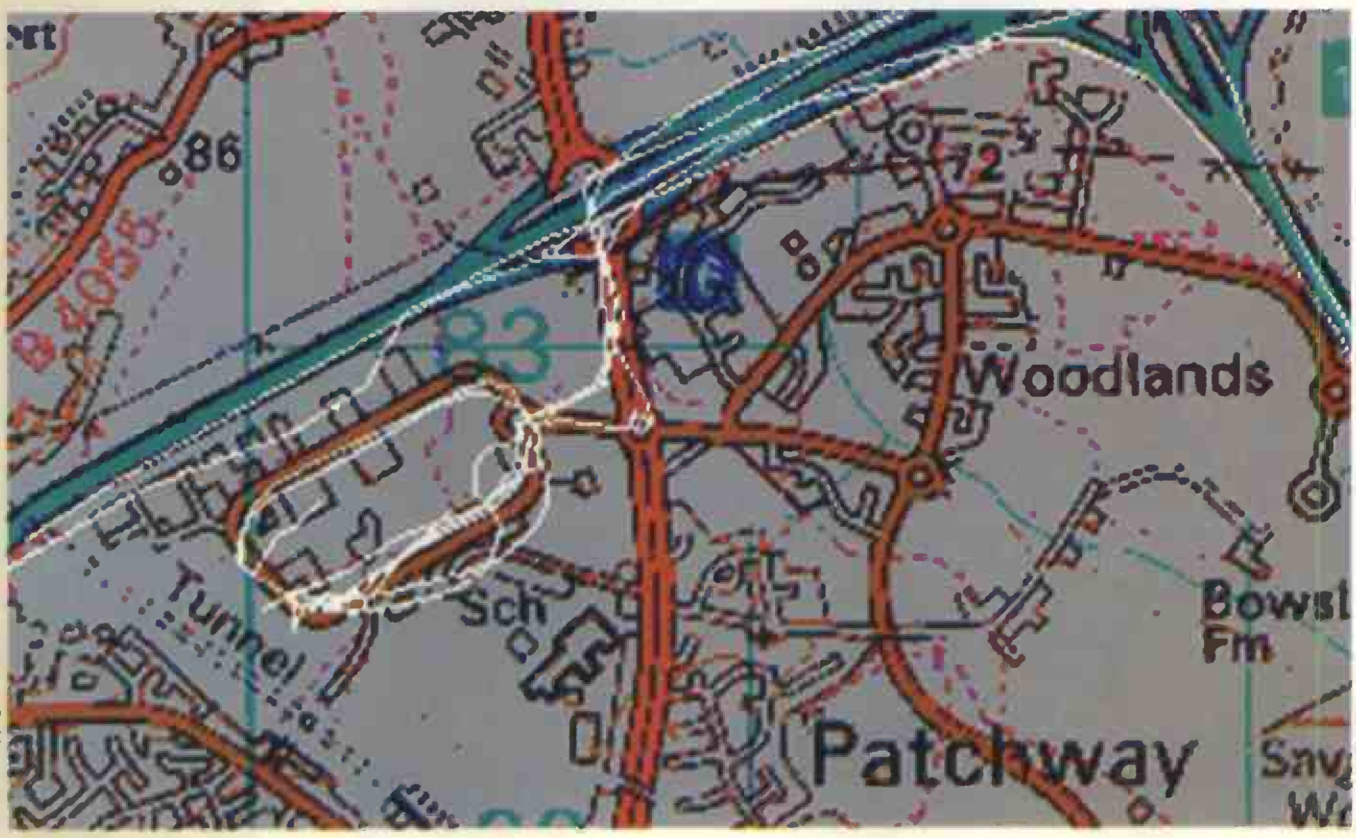
In the final part of his series Philip Mattos describes the applications limits of the system and the fusion of GPS with other sensor technologies for vehicle navigation. We also invite readers to register for further details about a kit of parts using the transputer hardware featured in this series.

Fig. 1. Track of vehicle on raster scanned 50,000 scale map.
The car raster map at 50,000 scale only shows about three kilometres, but is perfect for arrival at a destination. The four off road track sections are caused by blockage/reflections(2), and unmarked roads(2)

The GPS system was developed by the American military, for military purposes in land, sea and air. However over its long development period, receiver technology has advanced so much in terms of accuracy, size, cost and power consumption that civilian use far outstrips military activity.

We recall that civil GPS is capable of 20-30m accuracy in stand-alone mode, but that this is degraded to 100 metres by the American DoD to deny greater accuracy to the enemy. In differential mode, with corrections from a reference receiver at a known location, it is capable of five metre accuracy while, with carrier phase tracking, this can be reduced to a centimetre.

In cost terms, GPS sets with antenna, keyboard and display can be bought for around £500, and bare board-sets for incorporation in larger systems for around £200 in volume. This makes the small black box position sensor a reality and may be fitted to anything that



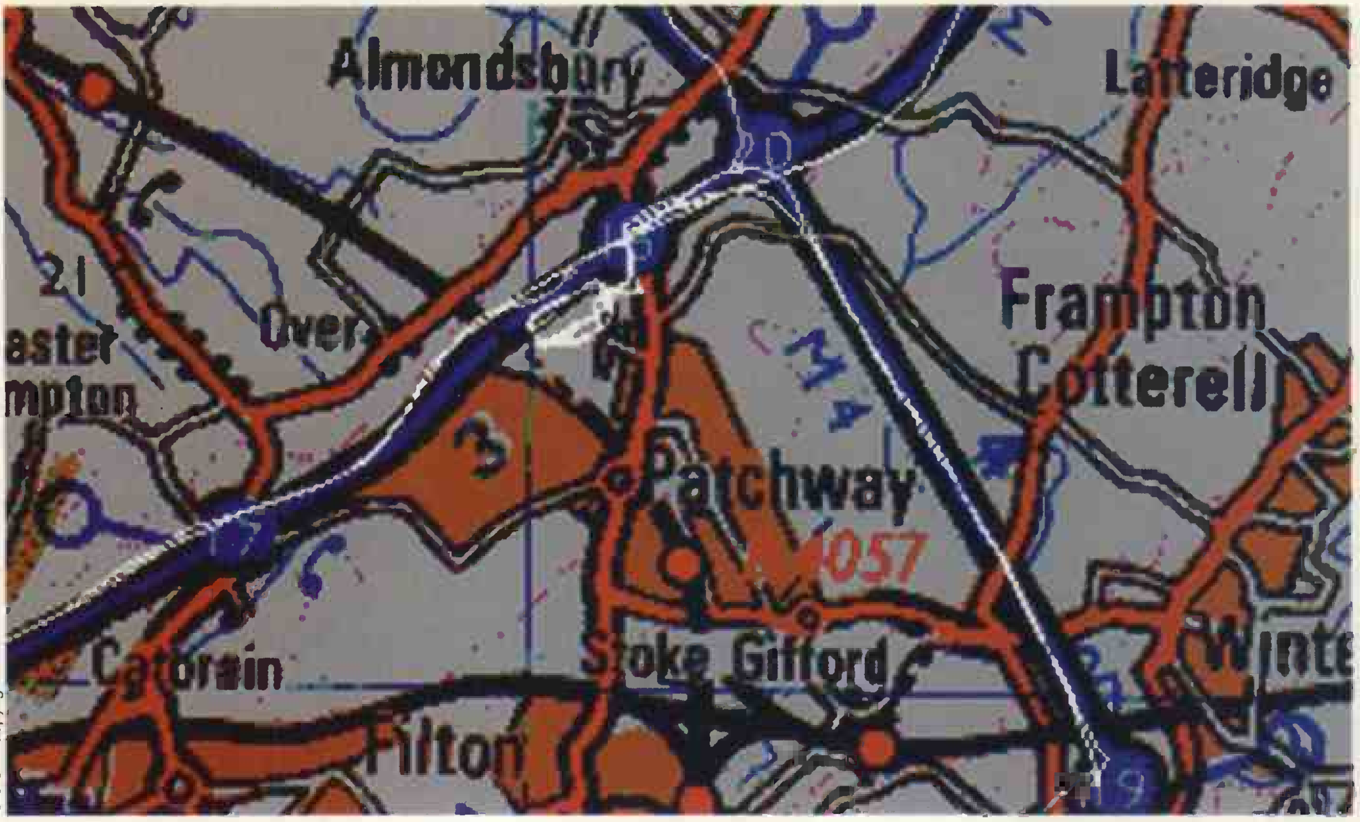


Fig. 2. Track of vehicle on raster scanned 250,000 scale map.
At 250,000 scale, the detail is lost, but about 10 miles can be seen. Note the M5 drawn far straighter, and far wider, than the truth.

moves... Or even as we shall see later, things that should not move, but do, like dams and continents.

At sea

The civilian user first met navigation systems, even navigation itself, at sea, so it was there that GPS established its civilian roots. However it had strong competition. The high volume American Loran systems can be bought for around £200, and in America they work fine. A GPS manufacturer has great problem competing with this, including distributor and retailer margins. Unfortunately too many manufacturers thought they could, and fighting for a finite market at a time of recession saw several companies fail.

Europe did not help. While not well covered for Loran, Europe's Decca stations were aging, and the long debate whether to expand Loran, update Decca or rely on GPS resulted in the devastation of the leisure marine market while consumers waited for the outcome. In the event, Decca won the first round led by the British, while the rest of Europe gave the second round to Loran. In a few years Europe will be covered by all three services.

The professional marine market, for fishing boats, ferries, and similar, though a small market, is perfect for GPS. The sea offers an unobstructed horizon, there are no (high) reflections, and knowing one's altitude exactly allows positioning with fewer satellites. This last was important in the early days, when satellites were few and far between, but is not very relevant now. And there are reflections from the sea itself. These are a problem with a helix antenna and low satellites. A helix is necessary on a sailing vessel that may heel away from the satellites, but a power vessel can use a patch antenna and avoid the problems.

GPS works well with external computer systems so that navigation packages can provide chained routes of waypoints, with course to and from each, distance, expected time of arrival, and many other functions.

In confined channels, such as in the Baltic, ferries use differential GPS to keep them precisely in the channel, even on their own side of it. Production platforms use GPS to locate exactly over the well head left by the drilling platform... even oil prospecting is done under GPS.

Search and rescue operations are managed by GPS, but more dramatic is the automatic distress alert that reports the vessel position. This is an IMO requirement, and effectively means that all commercial vessels carry an INMARSAT-C data-communications satellite transceiver, and a GPS receiver.

The ultimate must be EPIRB... a float-free buoy that alerts low-flying COSPAS-SARSAT satellites, currently on 406 MHz. Historically, the satellite fixes on the EPIRB, to an accuracy of a few kilometres. The next generation reports the vessel position, loaded automatically from the vessel GPS, to geostationary satellites that are always there, rather than intermittent coverage from the low-orbiting ones.

The progress of the future is in ECDIS, or electronic charts with radar information overlaid, and even ADS (automatic dependent surveillance), where vessels scanned by the radar report their name and their GPS position, for accurate display on the ECDIS.

In the air

Aircraft too have always had navigation systems. Decca and Loran do not work well at the speed of commercial jets, but are fine for propeller driven craft. Airliners use two major means of navigation: the autonomous Inertial Navigation System, essentially a set of gyros and accelerometers for positioning across oceans. When near land, they cease being autonomous and rely on radio beacons to give them both range and bearing (VOR/DME) operating in the VHF/UHF band, supplemented by radar control over the air traffic control voice links.

The problem with the INS is that it is an integrating system with no absolute reference, so it gradually drifts. One degree an hour is not unusual. By the end of a trans-Atlantic flight, the error is considerable, but easily corrected as the VHF beacons may be received 50 to 100 miles off shore.

The problem with the beacon system is that it creates lanes, or roads in the sky, causing congestion at each beacon, especially where routes cross.

GPS fixes this problem by allowing the creation of many more lanes, quite independent of the beacons. On land, any decent VOR/DME receiver can do this as it creates a position fix providing a range and bearing to another fictitious beacon. However it is only recently that real computers have been embedded in such equipment to do this.

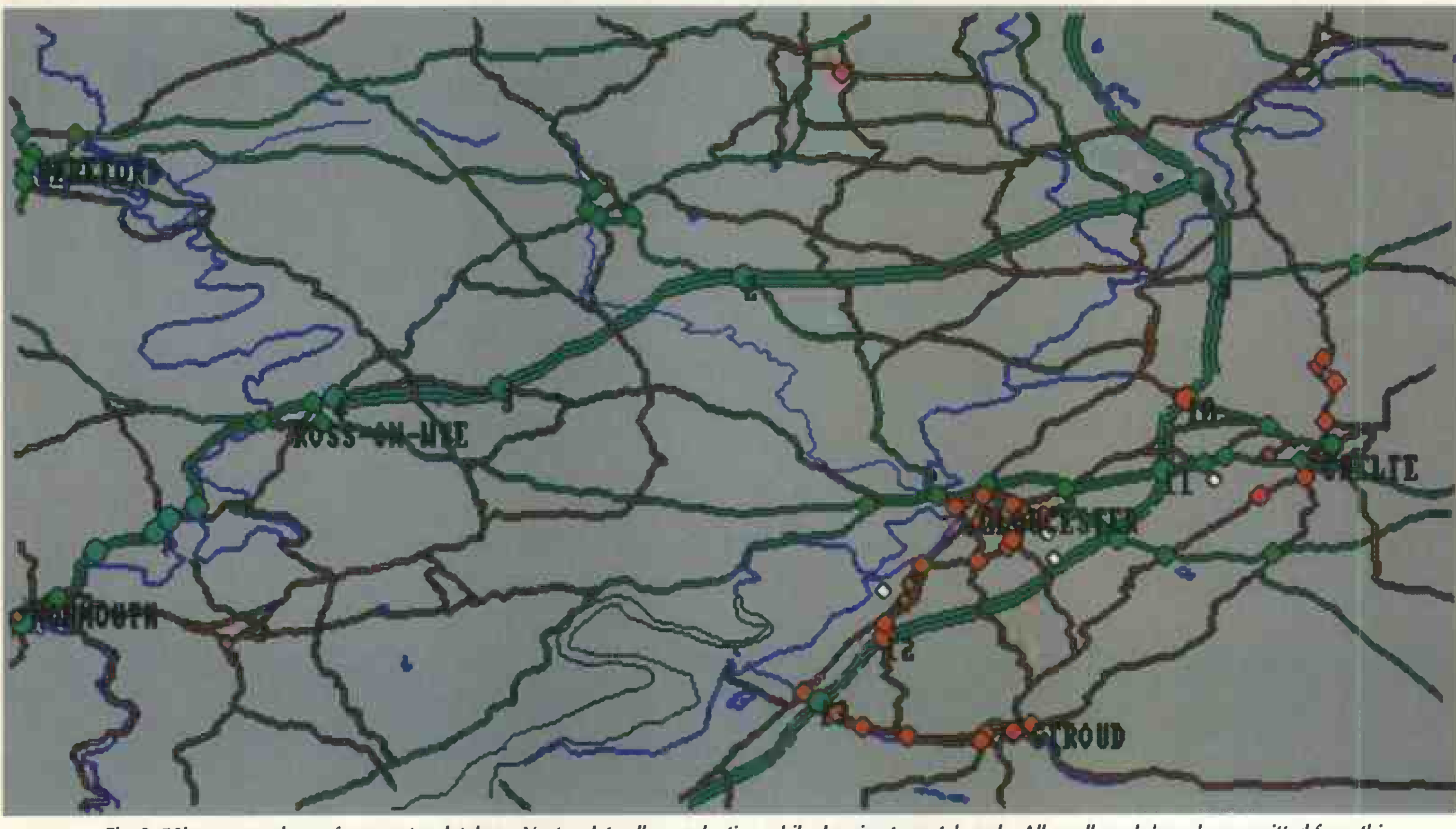
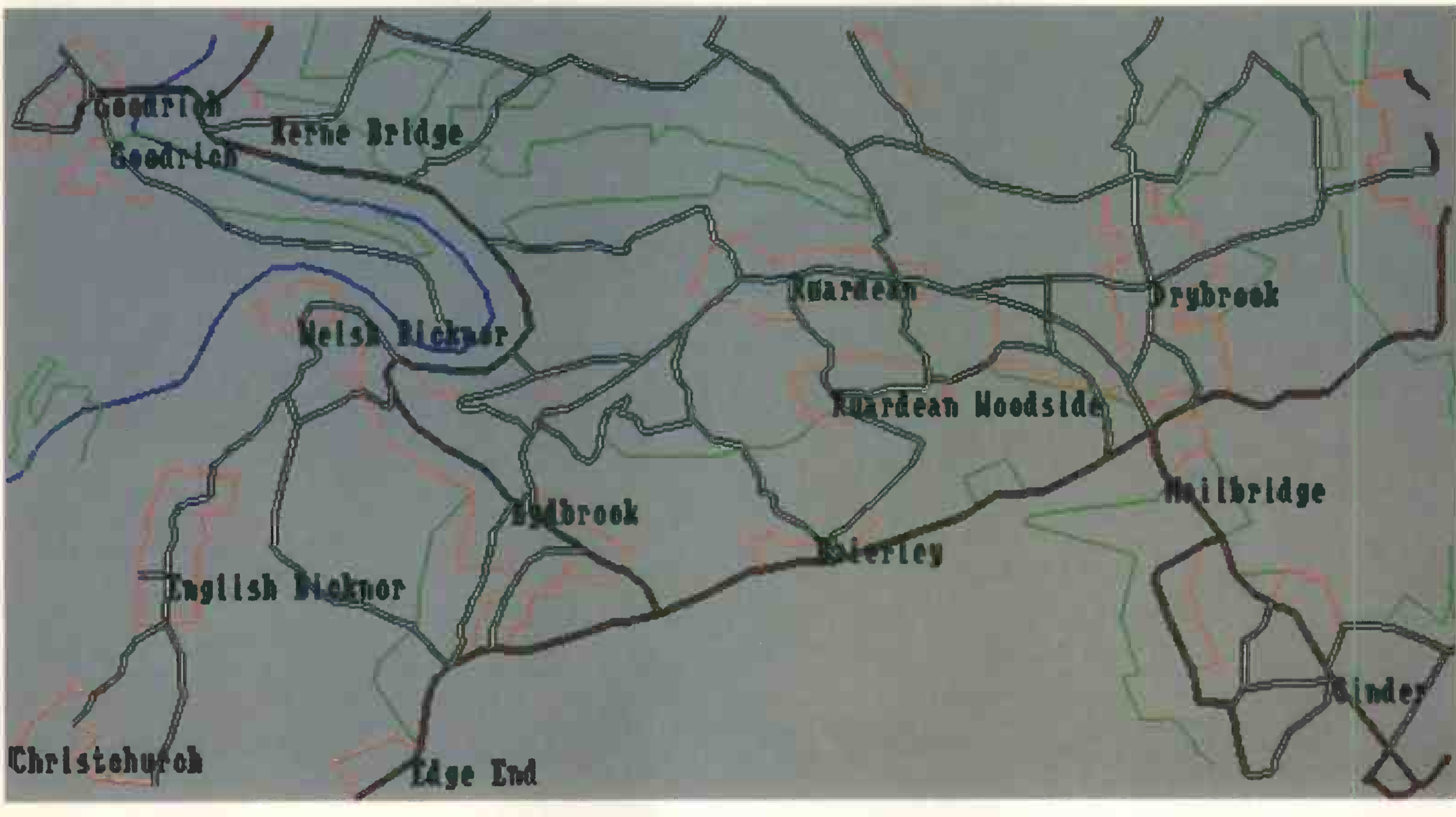


Fig. 3. 50km square drawn from vector database. Vector data allows selection while drawing to match scale. All small roads have been omitted from this map, and all villages. It covers 50 km.

Fig. 4. Detailed map drawn from vector database. Zooming in to display about 7 km, every road, every village, even the area of housing, can be shown.



Only GPS can provide this function over ocean and desert. Many more lanes may be created, and both the cross track and the along-track separations reduced. Satcomms via Inmarsat can report the aircraft positions to air-traffic control, allowing continuous changes of route as weather dictates.

Portable GPS sets will become indispensable for microlights, balloons and light aircraft.

There are snags, however. When GPS stops being an aid, but becomes an essential piece of the navigation equipment, lives begin to depend on it. We can cover for equipment failure on the aircraft by carrying two or even three separate receivers. We can detect a GPS space-segment catastrophic failure and operate with reduced performance. The problem we cannot handle today is the gradual build-up of an error in a spacecraft. Such an error, for example by drifting off orbit, or the clock frequency drifting, would result in a gradual increase in position error that would not be detected until two aircraft, using different constellations of satellites, collided.

The ICAO/FAA standard for such failures is that they be detected and the crew warned within 10s. GPS does not have a mechanism to support this. The control segment monitors satellites all the time, but sometimes cannot upload new data to them, for example to mark them unhealthy, until they are in range of a control station. As these are all near the equator or on mainland USA, satellites over the south pole, or Northern Russia are sometimes out of view for up to two hours.

The Block 2R satellites have the ability to communicate with their neighbours, alleviating this problem. However the GPS message format for the almanac means that while a satellite can report on the errant behaviour of a neighbour if so instructed from the ground, the message is 12.5 minutes long. Thus users, even if listening to it, would have to wait that long for the data.

Even if a satellite detected its own errors, the message frame for the ephemeris/clock data is repeated on a 30s cycle, outside the 10s spec.

There are two solutions at hand. One is to provide enough satellites such that the receiver can itself detect that one satellite is in error from the redundant solutions. If a receiver can receive five satellites at all times, then it can calculate five separate positions, leaving out one satellite each time. Four are wrong, one is right... but which?

If a receiver can always see six satellites, it can calculate some 30 different positions, and all will be correct – except those that use the bad satellite – enabling detection. This is in hand by the work being done on combined GPS/GLONASS receivers. GLONASS is the Russian equivalent of GPS.

This is a simplistic way of looking at it. In practice all six satellites would be included in the Kalman filter solution for a single position, and the residual error of the bad satellite would be seen to soar way over the threshold.

This method is known as RAIM (receiver autonomous integrity monitoring) because the receiver itself makes the decisions on bad satellites. Its easy and effective, but GPS alone will not have enough satellites to guarantee six in view all over the world all the time.

The second method is to monitor the satellite health on the ground, and transmit it up to the aircraft. This is much easier, as the ground station is fixed. If it appears to move, something is wrong. This needs a network of monitoring stations and a ground to ground and ground to air communication system. The commonality of requirements with WADGPS, wide area differential GPS, makes combining the two very effective. This is proposed by Inmarsat on their *F3* satellites to be launched in 1995, being tested now at very low power levels on current space vehicles. They will transmit data in GPS format, on GPS frequencies, so little modification is required to conventional GPS sets. The data will allow WADGPS, but will also give a health indication of every satellite in view every 6s. This is achieved by transmitting the data far faster than the GPS 50 baud rate.

In cooperation with Bristol University and Inmarsat, I have a dish antenna set up to receive the test signals, and have shown that even 1kbaud can be received once the receiver is locked on, but fast baud rates disturb the integrators used to lock on the code trackers during acquisition. This is, however, totally surmountable by using epoch-synchronous integration, so research continues.

Besides remote regions, GPS can be useful for precision approach and landing. Far cheaper than ILS or MLS systems, it travels with the plane, so to the equipped plane, every airport is available immediately. This is important in developing countries, where MLS may take decades to appear. Precision approach can be done using WADGPS for total absence of ground equipment, or using a local monitor for both health and corrections. Even pseudolites have been proposed, but these cause the near-far problem, where they drown out the true satellites.

On land

On land, GPS excels in three entirely separate domains. The first is surveying. The survey industry has developed the use of the GPS signal far beyond the dreams of the system designers, with millimetre accuracies over country-wide baselines. The theory behind this, using the phase of the carrier wave at both reference and mobile stations, was covered in an earlier article.

Such accuracy allows new applications previously undreamt of, such as monitoring the deformation of a dam under the pressure of water as the winter snows melt, monitoring fault lines for movement that might indicate an imminent earthquake, and even monitoring the plate tectonics of the earth. This last is the movement of the twenty or so plates that make up the crust of the earth. Coincidentally and unfortunately, Kwajalein in Indonesia, where one of the GPS control stations is located, is on the fastest moving plate on earth, at about six inches per year. Besides all these exciting applications, surveyors also use GPS rather prosaically to survey building sites, roads with one man operation and no need for line of sight between survey points.

For the GPS manufacturer, however, surveying is not very interesting as the market is so small. The two major players are Ashtech and Trimble, both of the USA, with Wild of Switzerland offering a system based on a Magnavox receiver, and Novatel of Canada offering an OEM module with survey accuracy which uses a transputer!

The second market chronologically relates to position reporting for trucks, buses, trains, and in the future, even freight containers and individual railway wagons. This market is much larger than surveying, but slower to take off as it needs a communications infrastructure to get the position information back to the control centre. For worldwide use, this was pioneered by the combined Inmarsat-C and GPS set demonstrated for Inmarsat by an Inmos/Bristol University team in 1990. Such equipment is now available from at least five

Constructional kit/Newsletter

It is intended to offer a set of pre-assembled boards that the reader may use to make a GPS receiver. However the cost is unlikely to make this proposition appeal to everyone.

The boards combine to make a system that delivers position on, and is controlled from, an RS232 interface. This may be connected to a PC for testing, or taken in a car over a route, then taken back to the PC and the memory dumped to display the track covered. There would be no display in the car unless the user had a portable PC.

The software and radio described in this series of articles are available for licensing. However it is not a viable proposition to construct from component level at volumes below 10,000 units a year. At lower volumes, it is recommended that potential constructors purchase ready made modules to incorporate in their system.

At the time of writing, application engineering data and price information about the hardware mentioned in this series are still being finalised. We are inviting readers with an interest in building either private or commercial GPS systems to write in to the address below to register their interest. We will then put together a mailing list and send out further details as they become available.

We will also be publishing a book based on this series together with additional new material written by Philip Mattos. This will contain a more detailed study of the constructional aspects relating to the GPS system hardware. Further information about the forthcoming book will also be included in the newsletter.

We regret that we are unable to deal with queries relating to this series over the telephone.

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manufacturers. However the cost of satcomms cannot be justified for local communications, and terrestrial infrastructures are taking time to appear.

For service vehicles, such as buses, police, electricity board etc, existing voice VHF networks may be used, and this is progressing well. There are also plans for emergency panic button systems for private cars emerging, but these are a special case because they normally generate no communications traffic until the emergency arises.

Progressing to containers and railway wagons will not happen until we can power the entire system from solar panels. This can be done now for self timed reporting. By self timed, we mean that the system turns on periodically on a local clock, rather than listening continuously for a polling message. In the polling case, the continuous power of the communications receiver cannot be supported.

The third area of application for land-based GPS is the consumer market, be it in a car or portable. There is no real justification for the portable because most land maps do not have Latitude and Longitude on them. At best they have the gridlines shown on the edges only, not overprinted as in the UK Ordnance Survey series.

However the lure of a new electronic gadget seems irresistible, and the portables sell like hot cakes. The first available (1989) were Magellan (USA) and Columbus (UK, transputer based). Now there are offerings from Trimble, and Garmin in the US, Streamline in the UK (transputer based), and JRC, Sony and Panasonic in Japan, the last being transputer based as well.

The biggest volume of all will be the car market. Already established in Japan, it can only run when detailed map data is available in electronic form, and legal impediments to screens in cars are dealt with.

In Japan, the former was handled by the "navigation alliance", where all the manufacturers got together to prepare one common map database, and the latter by sensible implementation... such as only trunk roads displayed if the vehicle is moving faster than walking pace.

In Europe, various half way approaches such as spoken directions or symbolic routes are being tried. Personally I believe these will alienate potential customers. Most want a map that looks like the familiar paper variety; this is what I have implemented. Other approaches are much harder to provide because the computer must understand the map.

Trial maps

Over the summer of 1992 I developed a complete map-display GPS system using either raster scanned maps as demonstrated in the panic system base station in the previous year, or it can use a vector map database. The reason for providing both was the availability of map data. Some 250K data was available in vector form – 50,000:1 scale scanned for the purpose – though such raster images are now available commercially.

The hardware to perform GPS and display the maps was covered earlier in the series. The map work can share the CPU with the GPS because it is performed so rarely. The output from the transputer card is RGB or PAL video. The former is preferable, as it allows higher resolution monitors to be used. However the car market is very cost sensitive so probably only TV quality monitors will be possible.

If PC quality 640 x 480 monitors are used, then raster-scanned 50,000 scale maps are quite acceptable. A 150dpi scan puts about four inches of paper map across the screen, about three miles or 5km. However if lesser monitors are used, such as 320 x 200, then with less than three kilometres across the screen, the detailed display is only suitable for finding ones final destination on arrival. The solution, albeit costly on storage, is to offer two separate map scales.

Fig. 1 shows such a map for the area around the Inmos offices, north of Bristol, where the M4 and M5 motorways intersect. **Fig. 2** shows the same area on a 250,000 map where about 15km can be shown across the screen.

The white tracks show where the car was driven for a demonstration. They show both the best and worst of GPS: perfect tracking through the four-level intersection, and running on reflections when the direct path to the satellite is obstructed. Note that the first is a feature of my software tracker, because with separate code and carrier loops, the loss of carrier lock does not affect positioning.

The reflections can be seen on the M5 just south of junction 16, where we passed three trucks that blocked a satellite low to the west. The same satellite was blocked by a building on the oval industrial park, with about a hundred metre push out to the south-east. Note that the incursion inside the loop road, and the cut-off corner to the A38 north are not errors... we really drove the car inside the loop, on a service road, and there is a slip road to miss the roundabout that is not shown on the map.

Note also what is known as "cartographic generalisation". The map makers distort local features from their true positions to clarify information for the user. For example, the roundabouts are much larger than true size, the motorway far wider. Even the loop road in the industrial park is spread out to allow each building to be shown. Hence the GPS tracks consistently narrower than the loop drawn on the map. On the 250,000 scale map, one also notices that the motorway is drawn far straighter than the truth but then it is almost a mile wide. (A quarter of an inch at four miles to the inch.)

The limiting feature on adapting raster images to screens at different scales is the text size. Other features shrink easily. The simple solution is to use vector maps. **Figs. 3** and **4** show first a 50km square around Gloucester, then a 10km square in the Forest of Dean to the west of Gloucester. These are drawn from the same database. The user can select any scale at the press of the zoom button. Depending on the scale in use, different features are included or omitted. For example in the Gloucester map, all unclassified roads, and all villages, are omitted. In the detail map, the outline of the buildings of each village is shown, and also the edges of forested areas, and every village name. Text is still the limiting feature.

All this information for a 50km square is stored in a 256Kbyte flash eprom card. While production systems would use CD-rom, flash is more convenient for the prototype as it may be rewritten. The 3in x 2in flash card plugs into the transputer board running the GPS system. A portable system would probably continue with plug-in cards, such as PCMIA, in production.

For anyone with a portable LCD television, I can demonstrate a portable map-display GPS today but it will only become a marketable product when LCD screens reach 4in or 6in diagonals. My demonstration version uses a 6in portable CRT screen. The monitor itself takes four amps so although I can carry the screen, I cannot carry the batteries.

While Japan has implemented the true map display, European car manufacturers are far less enthusiastic, preferring symbolic displays or even synthetic voice to direct the driver. How much this is biased by the non-availability of the map data will never be determined. Implementations such as the Bosch *travel-pilot* and the Philips *Carin* are navigation systems giving such driver support, with route planning and directions, but not necessarily themselves having any direct positioning ability. Position can be supplied externally from a GPS receiver, and/or dead-reckoning sensors as described below.

The Panic button or emergency system is a special case where there is no form of position display, map or otherwise, for the driver. The data is sent over a radio link to a control centre, and all the driver sees are the text messages that are returned, such as acknowledgement and ETA.

GPS in towns

Most vehicle navigation systems require some support for the GPS system, particularly to cover periods when the GPS signal is masked by tunnels or tall buildings. There are three major inputs available... distance, direction, and map-matching. In fact there are Japanese systems available that operate solely by these means, without any absolute position sensor.

Distance is the most easily derived signal. It simply counts the wheel revolutions, either at the gearbox, using signals intended for the speedometer, or at the wheels themselves, using signals intended for the anti-lock braking system.

Direction is more difficult to derive. Firstly there is no sensor already present in the car, so one must be added. The only economical absolute direction sensor is the fluxgate magnetic compass, but there are severe problems with distortions of the earth's magnetic field by trains, transformers etc that it is not ideal.

The piezoelectric rate gyro is very accurate, and intelligently integrated can give good results but it is not absolute. One must integrate the output over all time to derive the change in heading so errors can accumulate.

Thirdly one can derive changes in heading from an ABS equipped vehicle by detecting the difference in distance travelled by the left and right wheels. This suffers from problems of wheel slip on acceleration and braking, and again, integration is required.

Thus direction can only be satisfactorily managed by integrating a poor absolute device, the compass, with a good relative device, such as the rate-gyro.

Both distance and direction sensors can be calibrated from the GPS position to allow for different wheel sizes, tyre wear etc, especially at speed on motorway, where the errors of Selective Availability on the GPS can be proportionally reduced by using a large calibration distance.

Given corrected heading and distance measurements, a dead reckoning position is very easy to calculate: the easting is D x sin(*heading*), the northing is D x cos(*heading*) for each elemental line segment making up the total track of the vehicle, and these can be remembered as a total delta-E, delta-N number pair until a new GPS position is available.

Map matching is the adjustment of the sensor-derived position to match the computer record of the roads. That was extremely carefully worded because the map may be displayed or just internal to the computer, and the adjustment may be a correction, as in removing SA errors, or it may, for display purposes, be adjusting a correct position to match an erroneous map, so that the vehicle is displayed on the road. Note that paper maps frequently have deliberate positional alterations as discussed earlier (cartographic generalisation). These are carried through to raster maps, but should not be on vector maps. Unfortunately, on early vector maps, they are still present, because the maps were vectorised for the generation of paper maps, not for navigation.

Map matching is the perfect solution for position determination, but depends heavily on the availability of accurate map data, and on the algorithms used for sensor fusion. The latter, the merging of data from map, dead-reckoning and GPS, is an extremely complex subject, as all three inputs are inherently wrong, due to map distortion, drift and SA/reflections.

As the map is the master database the correct output is the one that visually matches the map. This is moderately simple to implement when the map data is available. Far more difficult is to integrate dead-reckoning and GPS without detailed maps in the computer, as switching between GPS and dead-reckoning tends to produce a combined system with the worst features of both. GPS gives its worst output just as it loses a satellite or mistracks to reflection, so the relative dead-reckoning system starts from an erroneous reference point and has an evens chance of drifting better or worse.

Equally difficult is to do map-matching from GPS without any distance or direction sensors on the car. As the GPS position accuracy is 100 metres without reflections, the correct junction can only be selected if they are always at least 200 metres apart. This means that with an intelligent algorithm that can backtrack, this is feasible in the country, but not in town where the average block is less than 200 metres.

With the map data available to me (250,000 vector data) I have no town roads anyway, so the GPS/map matching combination was possible within its own limitations. The algorithm was extremely simple: find the nearest road line-segment to the current position, excluding those whose angle was more than 45° from the current track. Note the use of the word track since we had no heading sensor. Thus, if parked near a crossroads, the system would behave erratically until the vehicle moved to establish a track direction.

In town, working with much finer scale maps, the Japanese have perfected systems that run on map and dead-reckoning without absolute reference at all. However this fails on long straight roads, and also on motorways, as there are no corners to correct and calibrate the distance sensors. As a result, they are adding GPS to provide absolute position fixes.

The final solution is an intelligent sensor fusion of gyro and magnetic fluxgate compass, wheel sensor and GPS. The gyro compass is accurate, but a relative instrument that drifts. The magnetic compass suffers offsets due to location, and violent swings due to passing metallic objects like buses. Together, they work well, with GPS track calibrating the offsets, and also calibrating the distance sensor. There is no switching from sensor to sensor as satellites and buildings come and go.

This scheme allows the GPS position to be averaged over all time, just as it can be in the static surveying mode. This takes out almost all SA and reflections, providing the vehicle is moving through random obstructions, leaving ionospheric errors as the major ones, as their time-constant is too long to average. Thus we have a 20-30 metres accurate solution to feed to the map matcher when the data becomes available.

Non-positioning applications

It may seem unusual that a global *positioning* system could have uses other than positioning, but the closely associated functions of time and frequency standards are performed with excellence by an almost standard GPS receiver.

Positioning is done by measuring very precisely the propagation time of the signal from satellite to user. An error of 1μs is worth 300m in range, over a mile in position. Any standard GPS set has resolved time internally far better than that. Thus GPS time is used to synchronise systems across oceans, where variable cable delays would make it impossible terrestrially. The BBC time beeps, no longer generated from Greenwich, are timed by GPS receivers.

The modifications required are those to get the internal time out to the user. This is far more difficult than it seems. In fact it is almost impossible due to the delays in implementing an interrupt, let alone executing an instruction.

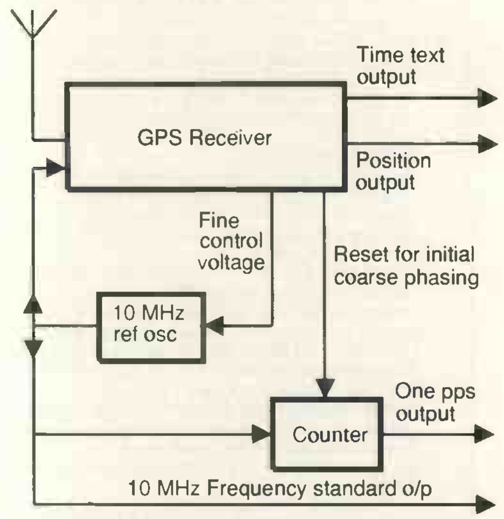

Fig. 5. Timing Receiver. A standard GPS receiver with an external clock generator and divider chain can equal any atomic clock in long term timing accuracy. Short term is degraded to tens of nanoseconds by Selective Availability.

The solution is to generate a hypothetical timing signal in hardware, and feed it into the processor with the GPS signals. The processor then monitors the difference between the signals, and exerts control to pull both the phase and frequency of the external reference until it is perfect. This is normally done internally at 1kHz, to match the GPS C/A code epoch, but the external interface is two electrical signals. On one, an edge every one second at the precise UTC second, and on the other an RS232 text string identifying the exact time.

Note that the GPS satellite message includes parameters which even warn of UTC leap seconds in advance. This is the ultimate equivalent of the RUGBY code clock, accurate to a few nanoseconds with the delay to the user's location accounted for. However, as the GPS signal, unlike 60kHz, cannot pass through buildings, a roof top antenna is required. One must also account for the length of the coax and the delay through the radio since GPS time is that at the antenna.

Given that a perfect reference has been created, normally at 10MHz, standard frequencies are also available from the divider chain used to create the 1pps signal.

In this series we have covered the GPS receiver from antenna to map output and timing beeps. Although I have largely described the working of GPS in terms of my own design approach, I hope the discussion has been sufficiently wide that others can understand the wonders of a system that has been over twenty years in the creation, consuming untold billions of dollars, and finally reaching full coverage this year (1993). It has taken over my professional life for the last five years and there are still plenty of areas within it for me to explore. ■

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

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CIRCLE NO. 116 ON REPLY CARD

LETTERS

CD or NBG?

Ben Duncan's article "How clean is your audio op-amp?" (*EW + WW*, January), prompts me to ask if anyone has analysed just what digitising audio produces in the way of distortion products.

I have been carrying out some quality checks on CDs and am appalled by some of my results.

I naturally expected to find performance far above cassettes, but found that, given a similar playback spectrum curve, on a blind test most people could not tell which was which.

I have no desire to stick my neck out, but it is plain that CDs have an inherent roughness which can only be due to the digital encoding.

It would be interesting to know the true figures. After reading about "dirty" op-amps, I wonder just what is going on and what we can expect from the dozen or so new marvels coming on the market.

I suppose the truth is that the supplanting of vinyls by CDs must be due solely to their convenience. But you would never know it from the media.

Ronald G Young
Peacehaven
East Sussex

Optimum settings

I was interested to read John Cronk's article on the design of a 1.3GHz tuner using a low noise GaAs fet front end (*EW + WW*, March).

The input circuit appears to make no attempt to set the input source impedance to the optimum. Noise figures for devices such as the *ATF10736* (ex *ATF20135*) change very rapidly with this parameter and it is important to include a suitable filter section to set this condition. For example, at 1.3GHz a noise figure of 0.8dB is obtained for a source impedance of 67 + j179Ω, but with a source matched to 50Ω the noise figure rises to 2.05dB.

Construction of circuits at this low microwave frequency is easy using double sided epoxy-glass board. Tracks can be made by cutting and stripping. Capacitors must be surface mounted and coaxial connectors mounted directly onto the board. Ordinary wire-loaded capacitors are at best lossy inductors and it is cheaper to use a couple of turns of wire to give an open circuit.

Stripline design equations are readily available in the published literature and are sufficiently accurate for the design of low noise amplifier circuits.

WBW Alison
Great Yarmouth
Norfolk

Cable con trick cut by Occam's razor

A dramatic difference to our listening, the adverts tell us, would be the result if we chucked away the old fashioned multi-strand flat pair speaker feeder cables and went over to the new silver or linear crystal oxygen free copper cables with added benefit of preferential conductivity.

Quizzing the appropriate manufacturers brought the comment "copper conductors aren't unidirectional". Indeed, the technical spokesman for one distributor, questioned about the need to connect the cables strictly as marked, talked about electrons jumping about, and of the conductor becoming fatigued if polarity was not observed, with the cable having a higher conductivity in one direction than the other.

Pointing out that speakers were driven by alternating signals, so weren't we rectifying the signal, the confident reply was that *they still sound better*.

Duly convinced (of something) we obtained a set of cables and connected them to the switched parallel amplifier outputs.

Using the same speakers we did a blind test from various programme sources including test CDs. Each of us compared the virtually instantaneous switch over from one unknown cable to the other. The first results were crushingly disappointing. No immediate difference could be detected, nor any when a carefully conducted set of double-blind tests followed.

Our suspicions were aroused and it looked as though some theoretical work was necessary to put the apparently outrageous claims for these cables into perspective.

The only properties of a cable which could influence its suitability as a loudspeaker-amplifier link are its series impedance *Z*, its shunt admittance *Y*, and the change (if any) of these parameters with frequency. Furthermore, the cable should, of course, be bilateral (having no tendency to conduct preferentially in one direction, rectifying). But all metallic conductors likely to be considered for this purpose are, by definition, bilateral.

Consider, as a basis for comparison, the well known and widely available 79 strand, twin core, PVC-insulated cable having cores of diameter 1.78mm spaced 4mm centre to centre. The resistance at low audio frequencies along the combined path length is about 0.0137Ω/m and its loop inductance (calculated using standard formulas) about 0.7µH/m. The resistance at high audio frequencies would he higher but for such a stranded conductor is difficult to calculate.

But the increase at, say 15kHz, will be less than the 13% increase for a solid conductor of the same cross-section. The inductance will fall slightly as the frequency increases. The shunt admittance consists of capacitance approaching 77pF/m (standard formulas) in parallel with conductance (measured) of about 2.2 x 10^{-11} S/m, ie entirely negligible.

The effect of these parameters on the frequency response of the

Not trivial

Malcolm Hawksford's acknowledgment (Letters, *EW + WW*, November 1992) of the value of R_g being a capacitor in aiding supply rejection is welcome. But his trivialisation: "After all, if $R_g = 0$, there would be virtually no injection, and no signal either!", is unfortunate. Many power amplifiers of the form shown in his diagram (taken from his reference 3) are likely to use mosfet output stages with considerable input capacitance, highly dependent on output stage transconductance and loading.

Indeed, his own mooted concept of a pontoon buffer power amplifier would see the large voltage swing driver output being loaded by the capacitance of long interconnects to the remote buffer.

His equation for the ratio of output to input transfer functions for inputs V_s and V_{in},

$$\delta = [1/Z_{n1} + (1 + r_2/Z_{n1})/mZ_{n2}]1/g_m$$

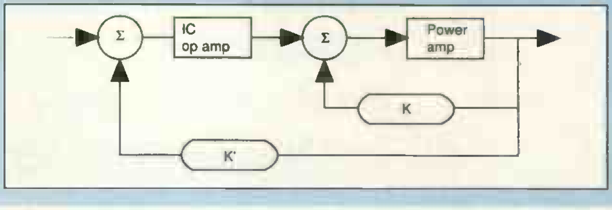
while independent of Rg, clearly shows the benefit of cascading to raise Zn1 and Zn2 and aid supply rejection especially in low gm fet input stages, preferred for their audible transparency.

But, a serious omission is any mention of the substantial advantage to PSRR afforded by enclosing the amplifier in a nested loop with a high gain input stage as shown in the diagram.

Such an approach using a typical IC op amp powered from highly filtered supplies can improve PSRR by 100dB at low to 60dB at high audio frequencies and output stage adverse loading THD by similar amounts. At the same time it shifts the onus for performance in the areas of input device matching, DC drift, common mode, and differential linearity to a monolithically matched and performance defined device according to the error budget.

The considerable improvement in PSRR at low frequencies is invaluable when the large variations in supply voltage of an appropriately sized power transformer at power envelope frequencies are to be accommodated without intrusion.

Greg M Ball
Coolangatta
Australia



loudspeaker/cable combination depends on the loudspeaker impedance and this can vary widely over the frequency range. The worst case occurs at the minimum of the speaker impedance, where the impedance is predominantly resistive.

Assuming this lies (for many speakers) in the region 200 to 1000Hz we have, at a frequency of say 500Hz, a cable impedance, Z, of $0.0137 + j2.2 \times 10^{-3}\Omega$/m and the admittance, Y, is j0.241μS/m. These values give $Z_0 = \sqrt{(Z/Y)} = (182.5 - j155.6)\Omega$ and ν, the propagation constant, $= \sqrt{(ZY)} = (0.0375+j0.044) \times 10^{-3}$/m. So $\alpha = 3.75 \times 10^{-5}$Neper/m and $\beta = 4.4 \times 10^{-5}$radian/m.

Taking a speaker impedance of, say, $(8 + j4)\Omega$ and a cable length of 5m, these figures give an attenuation (input to output of cable) of less than 0.04dB and a phase shift of only 0.13°! At higher and lower frequencies, the effect of the cable is likely to be much less. Speaker impedance increases as the parallel resonance is approached and it increases with increasing frequency (as the voice coil reactance increases) or as radiation and loading effects change the system compliance.

The ideal, of course, is zero attenuation and zero phase shift at all frequencies, but it is highly questionable whether such a slight advantage over the 79 strand cable justifies the high cost (several thousand pounds in some cases that have been advertised).

Some adjustment of the cable parameters could be made on a swings and roundabouts basis – for example, the increase in resistance at the highest frequency, while amounting to less than 13% for the 79 strand, could be reduced by insulating the separate strands, as reminiscent of some current speaker cables. But this would increase the loop inductance. If the go and return leads were intermingled in some way to try to avoid this, the shunt admittance would increase, but in view of the figures quoted above for the 79 strand, it does not seem worth it.

Little wonder then that we were unable to detect any improvement in our listening tests if the worst case example is shown to give an attenuation some 75 times less than the figure (3dB) generally accepted as the smallest change detectable by the human ear.

There is nothing wrong in using the best cable possible but that doesn't mean the most expensive. LCOFC (4 or 6mm^2) cable should cost no more than £2.50/m, but household ring-main cable (2.5mm^2, 30A capacity) or, indeed, a 79 strand (hi-flex) would be just as good. As mentioned, stranded conductors do have a marginally lower high-frequency resistance than solid conductors of the same CSA, but the effect is inconsequential.

In conclusion, for the domestic audio system with loudspeaker cables not exceeding, say, 25m in length, no discernible improvement can be expected when perfectly good cheap, cables are replaced by lengths of supercable. Anyone gullible enough to purchase these cables on the basis of the unwarranted pseudo-scientific claims currently being made in the specialist audio magazines should consider Occam's Razor: "It is vain to do with more that which can be done with less".

Dr BC Blake-Coleman
Dr R Yorke
Bassett
Southampton

Second childhood with whiskers

As a child I enjoyed making crystal sets to listen to Daventry and the early BBC broadcasts. Now, in my second childhood at 84, I am considering taking up the hobby again. From crystal sets I progressed to valves and some truly vast sets, though only of about three valves which were introduced and described in your pages.

Thanks for the memories, but I feel could hardly tackle the modern ones. So, I an writing to inquire first if FM broadcasts could be picked up by a crystal set and, if so, whether any readers know of some traditionalists/antiquarian who has such things as a crystal and catwhisker with which I can get started.

I would also like guidance as to the capacity of the tuning condenser (I can't remember the modern term for it) and tuning coil.

Gerald Carr
London

Old tube

I am trying to find out as much historical and technical information as possible about the old *VLS.492AG* electron tube made in England by Standard Telephones and Cables. Can any *EW + WW* reader help?

It appears to be a tiny cathode-ray tube with the whole flat top being the phosphor-coated display. The tube is a straight cylinder 39.5mm diameter and 167mm long including socket, and is fitted with an ordinary octal base socket (like a PL36 but with all eight pins).

Deflection seems to be electro-static with two pairs of plates at right angles (as usual). The heater seems to work at about 1 to 2V, 1A.

There is also an additional partially erased marking on the glass: the peculiar three-finger duck foot, "*RM*", "*/F*" and "*CV????*" (it could have been *CV1327* or *CV1527*). I read that *CV1327* is equivalent to *Pen1340*, but what is *Pen1340*? Would I be wrong to think that this tube could have served the army?

Christian Steffans
Waterloo
Belgium

Variable Planck

With regard to D Di Mario's "Gravity and electric force link up in black hole?" (*EW + WW*, February), Planck's time, as given in the article, does not correspond with the value of $t_p = 5.39 \times 10^{-44}$s given in "The fundamental physical constants" by E Richard Cohen and Barry N Taylor published in *Physics Today* in 1990. The value used by Di Mario differs from this value by $\sqrt{2\pi}$. He does not explain the difference.

Starting from a dimensional analysis approach, I derive the potential of the electron as

$$V = \left(\frac{\lambda_c c}{4\pi}\right)^{\frac{3}{2}}$$

where λ_0 is the Compton wavelength of the electron and c is the velocity of light. This equation predicts the ratio of the gravitational to the electric force of the electron as $F_g/F_0 = 2.40053(19) \times 10^{-49}$ using the value of parameters as given by Cohen and Taylor.

This is the relationship that Di Mario refers to. As Di Mario indicates, the largest source of discrepancy in the equation is in the poor statistics of the best value for the Newtonian constant of gravitation. I am in communication with Dr Cohen to clarify the matter.

Immo Bock
Randburg
South Africa

War crimes

After reading "The nature of power" (Comment, *EW + WW*, March), I have to ask if you are serious ?

Either I cannot detect your tongue in your cheek and am fooled, or I am justified in being shocked. Your recommendation: "only commit armed forces where there are clear economic goals" forgets that human beings have (non-economic) value in themselves. Economics by itself should not dictate the actions of people.

Do you only justify the National Health Service because it helps the economy? Please tell me you're not serious!

Andrew Gammie
Bath

I am serious in provoking discussion on the reasons for fighting wars. ***Frank Ogden.*** ■

Analogue by any other name

The four phase product detector for SSB described by Nic Hamilton (*EW + WW*, April) has its counterpart in computer science based on pulsed signals and is protected in the UK by patent 2,199,976, an invention for automatic pattern recognition. The principal feature of the patent specification is the use of "average frequency of occurrence" in place of the probability terms of conventional information theory. This results in an entirely automatic method of computation which may justifiably be called direct-in-binary.

In simple terms, this new methodology relies on simultaneous integration and differentiation in a bilinear diphasic arithmetic (ie, modulo four, base two) as geometrical place values rather than as magnitudes of numbers. Under this condition, the constant of integrations can only be zero or one, and a chain of integrations therefore generates a continuous bit string of 0s and 1s as a Turing memory which may be employed in subsequent chained differentiation. The methodology also relies on continuous signals which are reflected onto themselves in parallel loops. Under such a condition the conventional logic gates, such as and, nand, exor, and so on, acquire new functions and may then be employed in neural networks to detect new signal patterns and to ignore old patterns already present in the memory function.

It seems extraordinary that, although it has been known that all information has been encoded in a binary arithmetic since 1948, conventional computer science still requires the acquisition of data in decimal form, conversion to binary form for processing, and reconversion back to decimal for interpretation. Direct-in-binary systems (previously called analogue) offer the advantages of not requiring central processors, operating systems, or programming languages.

Brian Clement
Crikhowell
Powys

Better design with SC filters

Switched capacitor filters are flexible and easy to apply. ***Bashir Al-Hashimi*** *lays down the ground rules for effective design.*

Antialiasing prior to A-to-D conversion is one of the most common uses of filters. Suitable filters may be realised in many ways, with those operating in the frequency range 0.1Hz-100kHz usually built using discrete or hybrid active-*RC* networks. But high performance switched-capacitor (SC) filters are now commercially available, offering the designer a combination of flexibility and ease of use, and giving the advantage of anability to vary the filter band width simply by changing the clock frequency.

The aim here is to show how commercially available SC devices are used to design sharp, frequency variable low pass filters.

SC filters operate on the principle that a capacitor and a switch can be made to simulate the function of a resistor, (see box "Switched-capacitor resistor"). They are often designed using the same methods and configurations as continuous-time active filters, including the state-variable circuits and simulation of LC filters[1]. Resistors in these designs are readily built using SC networks (capacitors and a number of switches), and since SC filters consist of op-amps, capacitors and switches, the approach allows a full filter implementation on a chip.

Recently, greatly improved SC filter IC devices have become available combining ease of use and variation of filter characteristics through changing the clock frequency. But one of the drawbacks is that since the signal is periodically switched, the SC filter represents a sampling system. Aliasing and imaging must be considered – features that continuous-time filters do not possess.

Properties of sampled systems

Frequency spectrum of a continuous-time signal containing frequencies between DC and some frequency F_c, when sampled at rate F_s, will be modified to that shown in **Fig. 1a.** The spectrum now contains components, around the sampling frequency, F_s, called image frequencies, occurring at (nF_s-F_c) and (nF_s+F_c), where n=1, 2... Amplitude of the image component is given by

$$\frac{\sin\left[\frac{\pi(F_s - F_{in})}{F_s}\right]}{\frac{\pi(F_s - F_{in})}{F_s}} \qquad (1)$$

where F_s is the sampling frequency and F_{in} is the frequency of interest. For example, the 1st pair of image components of a 5kHz signal when sampled by an SC low pass filter with F_s of 1MHz, occurs at 995kHz and 1005kHz. The image components appear at the SC filter output as spurious signals, and must be removed or reduced to an acceptable level – achieved using a low pass smoothing post filter placed at the SC filter output. Complexity of the smoothing filter depends on the sampling frequency of the system. In general, the higher the sampling frequency, the less complex is the smoothing filter.

Clock frequency to cut-off frequency (commonly referred to as the corner frequency) of commercially available SC filters is typically 50:1 or 100:1. So the post filter could be either a simple *RC* network or a 2nd-order active-*RC* filter such as Sallen and Key. Clearly, the cut-off frequency of the smoothing filter should be greater than that of the SC filter.

As a rule of thumb, the cut-off frequency of the post filter should be a factor of five higher than that of the SC filter with clock-to-corner frequency ratio of 100:1.

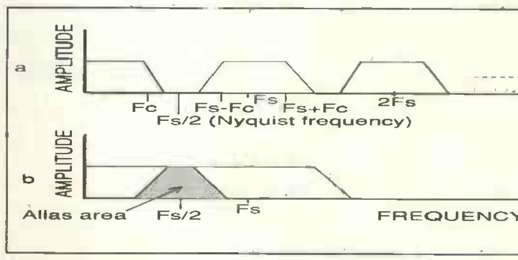


Fig. 1a. Frequency spectrum of continuous time signal sampled at Fs.

Fig. 1b. When frequencies above the Nyquist frequency are sampled, the sampled components "fold back" below the Nyquist frequency, generating unwanted alias signals in the pass band.

SC filters use transistors as switches and the parasitic capacitances between the transistor terminals allow part of the clock signal to appear at the filter output as noise – commonly known as clock feed-through. The amount of clock feed-through varies according to filter manufacturer, but it is typically of the order of 10mV (p-p), and can be eliminated by the post filter.

All sampled data systems are prone to aliasing when input signals exceed the Nyquist frequency, (half the sampling frequency, $F_s/2$). When frequencies above the Nyquist frequency are sampled, the sampled components "fold-back" below the Nyquist frequency (**Fig. 1b**). So signals beyond the Nyquist frequency generate unwanted signals within the pass band, called alias components. To prevent aliasing, a continuous-time low-pass filter is required before the sampled data SC device. Specification of this continuous-time antialiasing filter is similar to that of the post filter.

Several manufactures supply SC filters – eg National Semiconductor and Maxim[2], and general purpose SC filters come in two types: universal and preconfigured.

Preconfigured and universal filters

Preconfigured filters implement a specific filter function – low pass, band pass, high pass

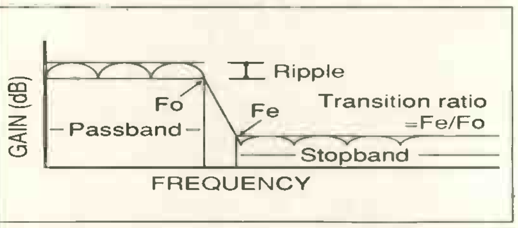


Fig. 2. Corner frequency of the elliptic filters is defined as the point where the filter output attenuation falls just below the pass band ripple.

Response	Elliptic
Filter order	8
DC gain (dB)	–0.1 (typ)
Frequency range	0.lHz-25kHz
Passband ripple (dB)	0.15 (typ)
Stop band attenuation (dB)	>75
Transition ratio	1.5
Clock-to-corner frequency ratio	100:1

Table 1. Typical frequency specification of the Max293 device

or notch – in one of the classic filter responses; Butterworth, Chebyshev, elliptic, etc. They cover the frequency range of 0.1Hz-150kHz and have a fixed clock-to-corner frequency ratio of either 50:1 or 100:1.

Preconfigured filters require no additional components and can come in packages as small as eight-pin dips. All they need to operate is a clock signal to set their frequency response. Most SC filter ICs have an on-chip oscillator which may be used to generate the clock signal. Good examples of preconfigured SC filters include a Butterworth filter with 90dB attenuation at four times the 3dB frequency point, and a notch filter with 30dB depth.

If what is required in terms of filtering functions can not be found in a preconfigured SC filter, then universal IC filters often provide the solution. Universal filters usually contain one to four 2nd-order section per packaged IC. The 2nd-order section is usually based on the state-variable configuration[3], and the circuit allows implementation of low pass, high pass, band pass, notch or all-pass filter functions. The realisation of filtering functions using universal SC filters requires external circuitry – ranging from resistors alone to relatively complex microprocessor control systems – and the design of the external circuitry often proves to be a fairly involved operation. For this reason, SC filter manufacturers usually provide hardware and software design tools to simplify the design process.

Practical application

Antialiasing filters have many different forms and characteristics. But filters with elliptic responses are ubiquitous, chosen because they provide the sharpest attenuation in the transition band (needed to maximise bandwidth and minimise aliasing) with the minimum component count, when compared with other responses.

Consider a requirement for a 5kHz sharp low-pass filter. A number of commercially available SC filters can be used, and one such is the preconfigured *MAX293* device from Maxim **(Table 1)**. One of the attractive features of this filter is the high attenuation (>75dB) provided at 1.5 times the corner frequency. The corner frequency of elliptic filters is defined as the point where the filter output attenuation falls just below the pass band ripple (**Fig. 2**).

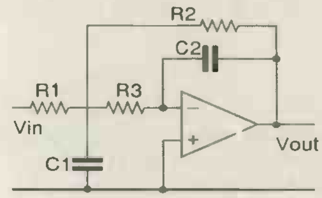
Amplitude response

The *MAX293* filter covers the frequency range of 0.1Hz-25kHz, and has a clock-to-corner frequency ratio of 100:1. So to set the passband edge at 5kHz, a 500kHz clock signal is needed. The clock signal can be derived from the on-chip oscillator with an external capacitor (*C*) given by

$$F_{osc}(\text{kHz}) = 10^5/(3*C(\text{pF}))$$

ACTIVE FILTERS

Analysis of the circuit shown below indicates this filter has a low pass response. It is commonly known as a multiple-loop



feedback filter[5], because there are two feedback paths from the output of the amplifier to the *RC* network. Design equations of the filter are:

$$C_1 = \frac{(0.474Q)}{(w_0 F_0 R)}$$

$$C_2 = \frac{C_1}{(9Q^2)}$$

$$DCgain = \left(-\frac{R_2}{R_1}\right)$$

where w_0 is the normalised frequency and Q is the quality factor of the filter. Both these parameters may be obtained from Table 2 for various filter responses. Parameter F_0 is the cut off frequency of the filter, while R is an arbitrary value of the filter resistor values ($R_1=R_2=R_3=R$) chosen to give practical values of capacitors. Note that this filter has an inverted output.

An example is the design of a 50kHz Butterworth low pass filter (w_0=l, Q=0.707). Assuming that R=20kΩ, the values of C_1 and C_2 are: C_1= 335pF and C_2= 74pF

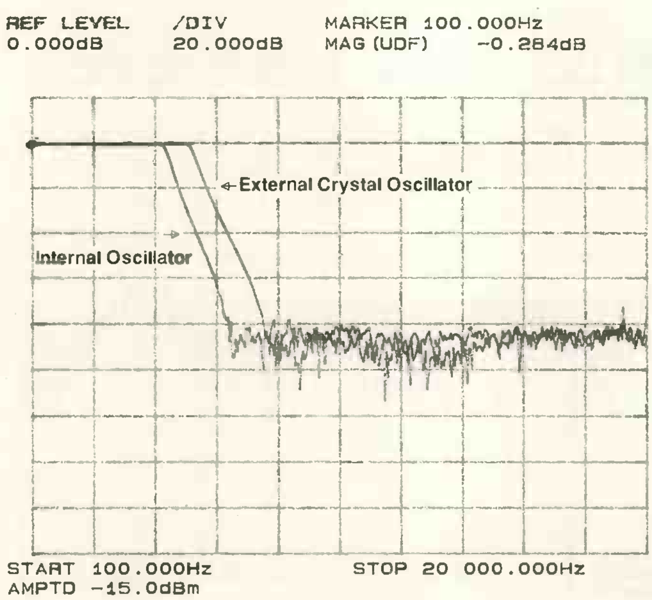


Fig. 3. 5kHz SC filter amplitude response with the internal and an external oscillator.

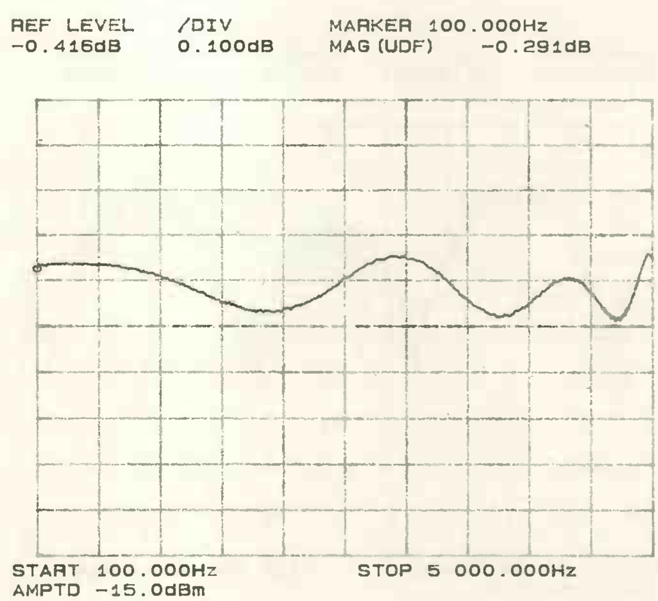


Fig. 4. Pass band ripple of the 5kHz SC filter with external crystal clock.

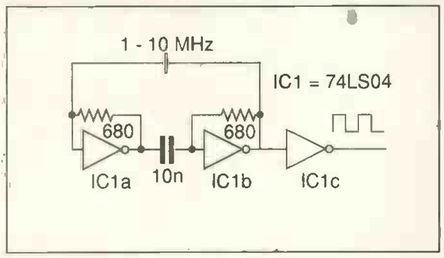


Fig. 5. Simple crystal oscillator consisting of two inverters connected in a ring via a capacitor and a crystal.

To generate a 500kHz clock signal, C=67pF.

Performance of the 5kHz low pass filter using the internal oscillator to provide the clock signal can be compared with the response of the same filter using an external crystal clock (**Fig. 3**). The filter driven by the internal clock has a premature roll-off, due to the tolerance (±15%) of the internal oscillator frequency. Obtaining a well defined pass band would require a variable capacitor (C) and the filter would need adjustment on test.

The filter driven by the external crystal clock performs exactly as predicted, pass band ripple of the filter agrees (**Fig. 4**) with the specification given in Table 1.

To achieve optimum results from the SC filter, an external stable clock must be used to ensure minimum drift. A simple crystal oscillator (**Fig. 5**) consists of two inverters connected via a capacitor and a crystal. Output from the oscillator is buffered through a another inverter.

The filter frequency spectrum around the clock frequency (500kHz) is shown in **Fig. 6** when a 3kHz input signal is applied to the filter. Note that the clock feed-through signal level is about 5mV, and the amplitude of the image frequencies agrees well with that predicted by Eq (1). Placing a simple RC smoothing filter with –3dB frequency point of 50kHz at the output of the filter reduces both clock and image components by about –20dB (**Fig. 7**).

The *Max293* device has an uncommitted op-amp which can be used to build an antialiasing or a smoothing filter. Generally, using this op-amp to build an antialiasing filter rather than a smoothing filter is the more useful choice, since the op-amp experiences some clock feed through.

For the 5kHz filter, a 2nd-order antialiasing low pass filter that can be built with this uncommitted op-amp is shown in **Fig. 8**. The filter has a Butterworth response and -3dB point at 50kHz (Design of the continuous time filter is discussed in box "Active Filters").

To vary the pass-band edge of the low pass filter, the clock frequency needs to be changed. **Figure 9** shows the amplitude response of the *MAX293* device at 5kHz, 10kHz and 20kHz bandwidth.

Phase response

Phase response of elliptic filters is a non-linear function of frequency. Since the derivative of the phase function is a measure of the delay (or group delay) through the filter, a non-linear phase response means that the delay will vary with frequency in a non-linear fashion. The group delay curve of the 5kHz filter is exactly as would be expected from any realisation of an 8th-order elliptic lowpass filter. **Figure 10** shows the effect of the non-linear delay versus frequency characteristics upon a square waveform (500Hz). The output waveform has a considerable overshoot and ringing, due also to the filter truncation of the Fourier series of the square waveform.

Practical considerations

Some practical problems are encountered with IC switched-capacitor filters. They exist with all SC filters but, as before, they can be illustrated by the *MAX293* device.

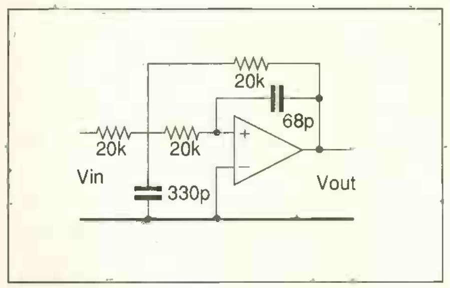


Fig. 8. 2nd order low pass antialiasing filter (F3dB=5kHz) that can be built with the Max293 uncommitted op amp

w_0	Q	
Butterworth	1	0.707
Chebyshev (0.5dB)	1.231	0.864
1dB	1.050	0.957
2dB	0.907	1.129
3dB	0.841	1.307
Bessel	1.73	0.577

Table 2. ω_0 and Q for various filter types.

Generally, SC filters have high output voltage offset – the *MAX293* is typical at about 300mV – and this is significantly higher than that encountered with op-amps. Offset adjustment is usually necessary.

Total harmonic distortion (THD) is important too, as a measure of unwanted harmonics produced at the filter output when a pure sine wave is applied to the filter input. THD arises from non-linearities within the filter and varies with filter type. Elliptic filters have the worst THD specification of all types because of their high Q-sections. Measured THD of the 5kHz elliptic low pass SC filter is better than –70dB. The test input signal has a frequency of 1kHz and amplitude of 5Vp-p sine wave, a 500kHz clock frequency and 20kΩ load. The measured wide band noise level of the 5kHz filter

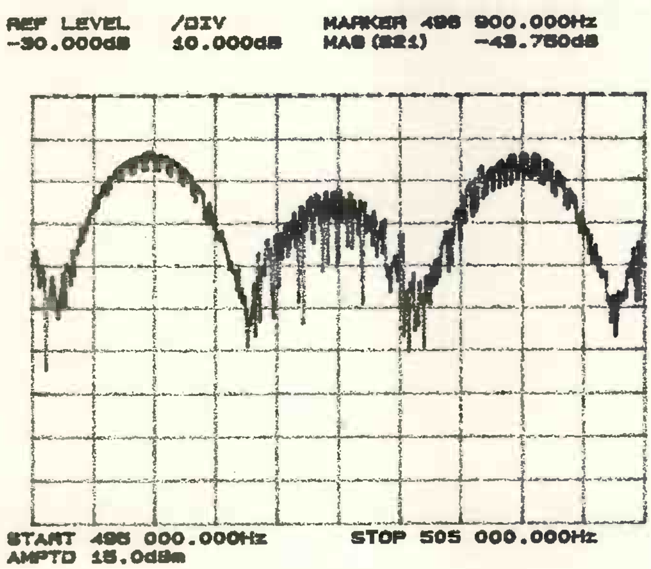


Fig. 6. SC filter frequency spectrum around the clock frequency (500kHz), with a 3kHz input signal applied to the filter.

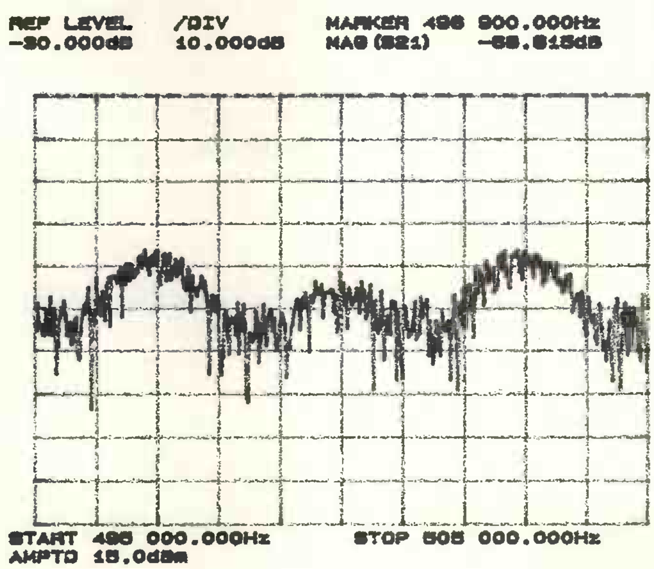


Fig. 7. Frequency spectrum of Fig. 6, after passing through a simple RC circuit with cut off frequency of 50kHz.

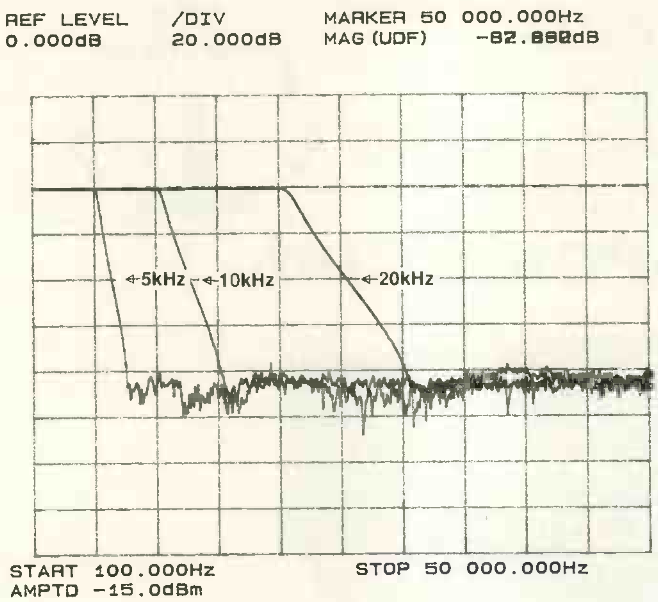


Fig. 9. Variable SC low pass filter response at 5kHz, 10kHz and 20kHz.

SC pros and cons

Advantages of SC filters include:
- Tunability:
- readily available (preconfigured, universal);
- easy to use

SC drawbacks include:
- Need additional circuitry (antialiasing and smoothing filters);
- external stable clock desirable
- THD and noise slightly inferior to continuous-time active filters.

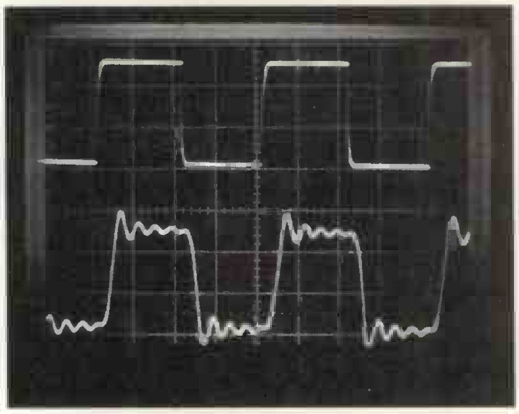


Fig. 10. Effect of non-linear delay versus frequency characteristics on a square waveform. Top trace, input 500Hz square waveform; bottom trace, output 5kHz elliptic filter v(X=0.5ms, Y=2v).

Switched capacitor resistor

A circuit that simulates the function of a resistor is shown in **Fig.** A. The switch is initially in the left hand position and the capacitor is charged to the input voltage V_I. The switch is now thrown to the right-hand and the capacitor is discharged after a determined time to some new voltage, V_2. The charge transferred is

$$Q = C_I(V_I - V_2)$$

If the switch is thrown back and forth at a clock rate F_{clk}, the average current flow is given by

$$i = C_I(V_I - V_2)F_{clk}$$

where F_{clk} is the switching rate or the clock frequency. From Ohm's law, the equivalent resistance of the switched capacitor is

$$R_{eq} = (V_I - V_2)/i = 1/(C_I F_{clk})$$

The switch is typically realised as mos switch driven by a non-overlapping two-phase clock (**Fig. B**).

A useful building block in filter design is an integrator (**Fig. C**), and the *SC* version of the integrator is shown in **Fig.** D.

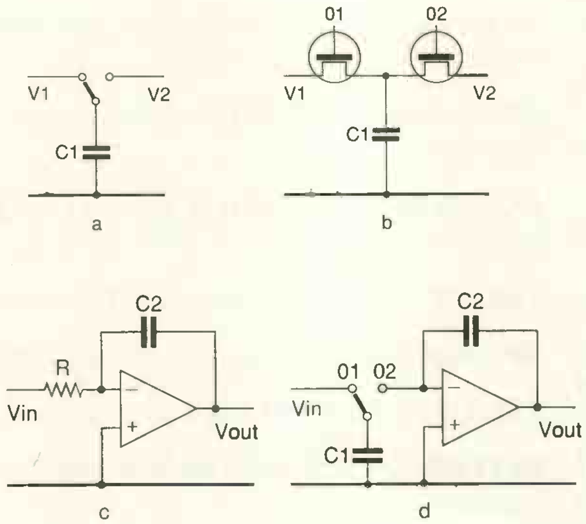
Transfer function of the op-amp integrator is

$$H(s) = -\frac{1}{sRC_2}$$

Substituing for R from the above

$$H(s) = -\frac{1}{s}\left(\frac{C_1}{C_2}\right)F_{clk}$$

The equation shows that the frequency response of the *SC* integrator can be varied by altering the clock frequency, F_{clk}.



is better than –70dB over the bandwidth 10Hz-100kHz: continuous-time *RC* active filters can achieve dynamic range and noise levels in excess of –90dB[4].

Acknowledgment
Thanks to Alan Holden for his help in preparation of this article.

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**Dr Al-Hashimi is a design engineer working for Matthey Electronics.*

MIGHTY FILTER POWER IN MINUSCULE PACKAGES

Using integrated filter packages has never been easier. Ian Hickman describes their application, and an audio circuit to test the response.

The Maxim devices *MAX291-MAX297* are 8th order lowpass switched capacitor filters available in 8-pin plastic dip, SO, cerdip and 16 pin wide SO packages, and even chip form. They cover a variety of filter types: Butterworth, Bessel, elliptic (min stopband attenuation A_s = 80dB from a stopband frequency F_s of 1.5 x the corner frequency F_o) and elliptic (A_S 60dB at 1.2 x F_o).

The corresponding type numbers are *MAX291/292/293/294* respectively, all at a ratio of clock to corner frequency of 100:1. The /295/296/297 are Butterworth, Bessel and elliptic (A_S 80dB) types, but use a 50:1 clock ratio, extending the maximum F_o to 50kHz against 25kHz for the others. All will accept an external clock frequency input, enabling the corner frequency to be determined accurately and to be changed at will. They may also be driven from an internal clock oscillator, with the frequency determined by a single external capacitor.

Although typical frequency response curves are given in the data sheets, an audio swept frequency source and detector were used to measure the responses independently, in the form of **Fig. 1a**. **Figure 1b** shows the result of applying the swept output direct to the detector.

The low amplitude at low frequencies is due to two separate effects. The first is that at low frequencies the output impedance of the internal current sources and the input impedance of the internal simple Darlington buffers in IC_2 are not infinitely large compared with the reactance of the 1.5nF capacitors.

The second effect is the rate of change of frequency, which at the start of the ramp is comparable to the actual output frequency itself, allowing the individual cycles of the frequency ramp to be seen. For measuring the filter responses, a much slower ramp would clearly be necessary, enabling the detector to follow rapid downward changes in level. The second effect would not then apply although the first still would (irrelevant since the filters

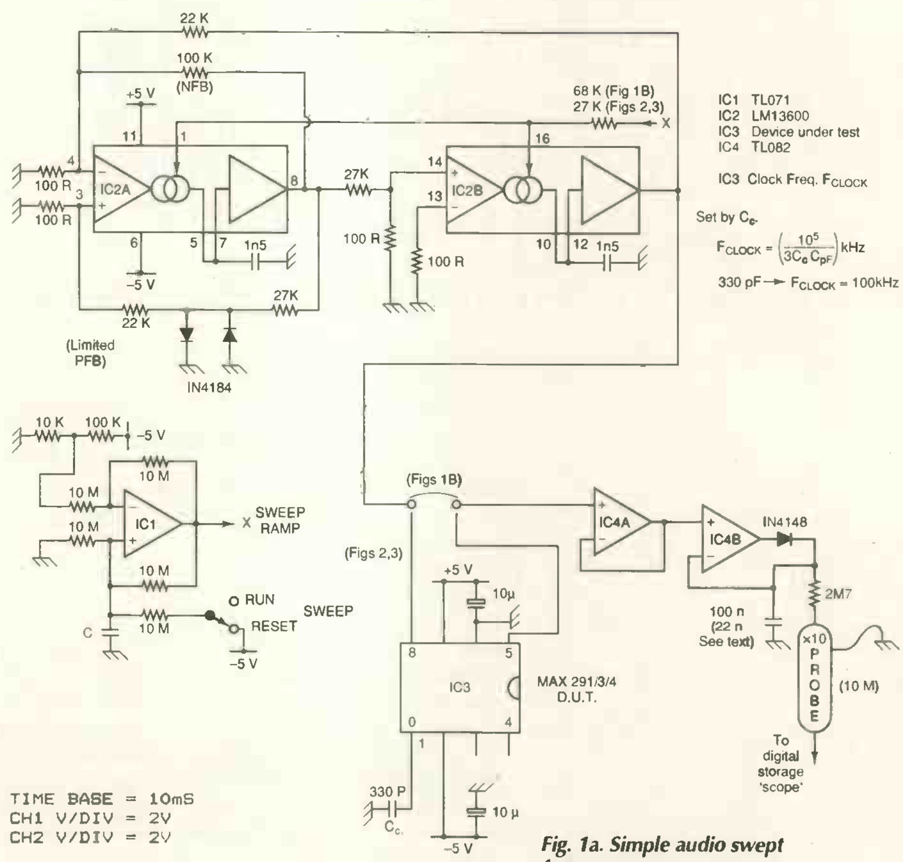


Fig. 1a. Simple audio swept frequency response measurement system. A Howland current pump is used to charge capacitor C, providing a linear sweep voltage at the output of op amp IC1. This is applied to the bias inputs of an LM13600 dual operational transconductance amplifier (OTA, IC2), used as a voltage-controlled state-variable-filter based sinewave oscillator. Its output is applied to the device under test, IC3, the output of the latter being detected by the ideal rectifier circuit IC4.

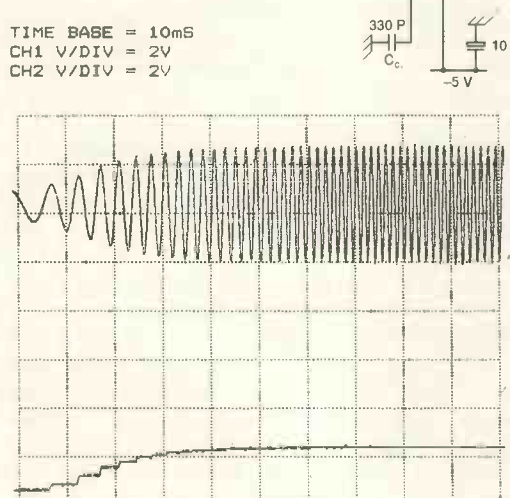


Fig. 1b. Using a small value of C, the swept oscillator output was applied direct to the detector circuit. The detected output (lower trace) follows faithfully the peak amplitude of the sweeper output (upper trace) over the partial scan shown, covering about 30Hz to 650Hz.

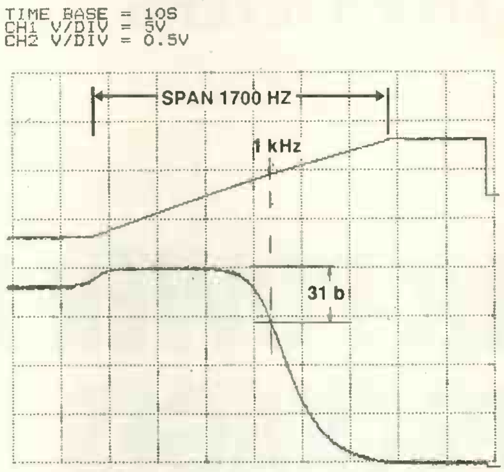


Fig. 2a. The ramp-voltage applied to the swept frequency oscillator (upper trace) and the detected voltage output from the MAX291 Butterworth 8 pole filter, set to F_o = 1kHz (lower trace).

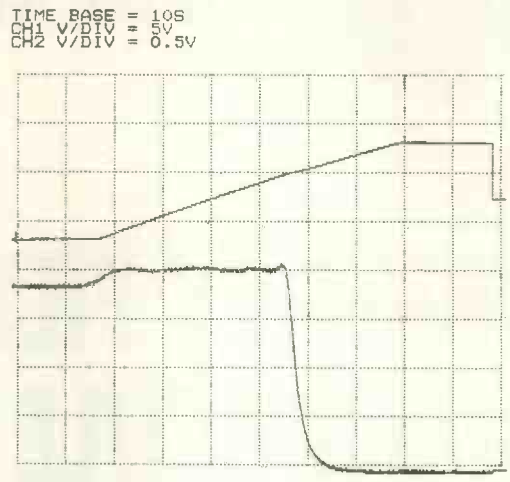


Fig. 2b. As 2a, but using the MAX293 elliptic filter with its 1.5:1 ratio of F_s to F_o.

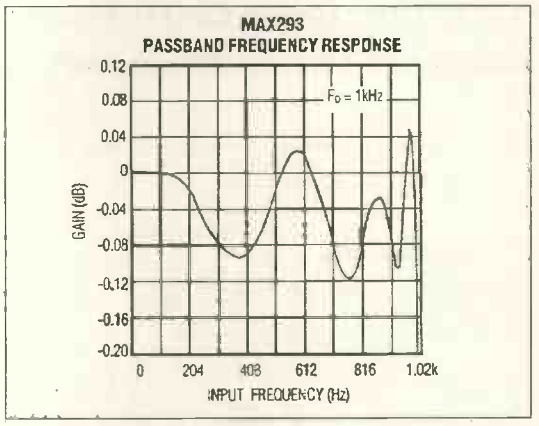


Fig. 2c. The manufacturer's frequency response data for the MAX293.

were operated at a cut-off frequency of 1kHz).

For testing the filters' frequency response, the value of *C* was raised from 1nF to 680nF, giving a sweep time of one minute. At this slow rate, the limited dot density of the digital storage oscilloscope resulted in a ragged meaningless depiction of the swept frequency test signal itself. **Fig. 2a** therefore shows the sweep voltage instead (upper trace), together with the detected output from the filter (lower trace, taken using the *MAX291* Butterworth filter).

The amplitude of the sinewave test signal settles rapidly to about $5V_{p\text{-}p}$ at the start of the sweep and remains constant over the whole sweep. The detected output starts to fall at the filter's corner frequency; as expected it is 3dB down at 1kHz. (The detected voltage is 2V, not 2.5V, due to the attenuation introduced at the trace two probe. This avoids overloading the digital storage oscilloscope's channel two A-to-D converter; the alternative – reducing the sensitivity from 0.5V/div to 1V/div – would have resulted in rather a small deflection.)

The 3dB attenuation at F_o and leisurely descent into the stopband, typical of the maximally flat Butterworth design, are clearly shown. Contrast this with the A_s = 80dB elliptic filter, **Fig. 2b,** which has dropped by 20dB from the passband level within a space of around 200Hz, agreeing with the maker's data, **Fig. 2c.**

With the linear detector of Fig. 1, Fig. 2b does not allow the detail of the stopband shown in Fig. 2c to be seen. Using a previous log amp circuit[1], detail up to around 80dB would be visible but this would still be insufficient to examine the stopband of this device adequately. The stopband detail of the *MAX294* could be seen, however. **Figure 3a** shows its performance with the Fig. 1 set-up.

The device's minimum stopband attenuation of 60dB is maintained while providing an F_s to F_o ratio of only 1.2:1. This plot was taken with the smoothing capacitor in IC_4's linear detector circuit reduced from 100nF to 22nF, enabling the detector to follow the very rapid cut-off of the filter at the given sweep speed. This means that increased ripple is observable on the detector output at low frequencies preceding the start of the sweep.

Figure 3b shows the same response with the original detector time-constant, demonstrating the distorted response caused by using an excessive post-detection filter time-constant – a point not lost upon anyone who used early spectrum analysers which did not incorporate interlocking of the sweep speed, span, IF bandwidth and post detector filter settings.

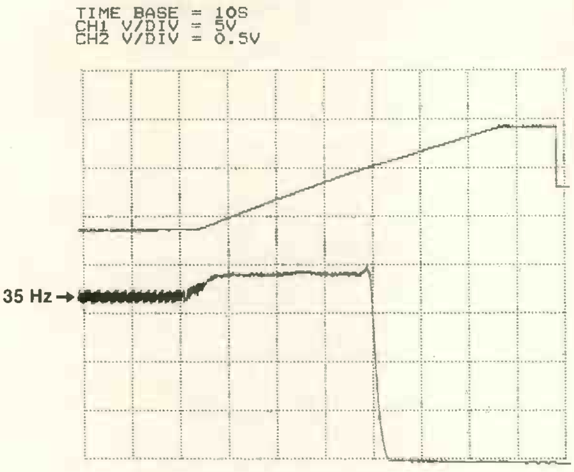
Of course an error-free measurement could have been taken using the original detector by reversing the polarity of the ramp to give a falling frequency test signal – at the expense of having a back-to-front frequency base.

Conversely, there would be no problem with the original arrangement when measuring a highpass filter, since the detector's response to increasing signals is very fast. The design of a detector with low output ripple but with fast response to both increasing and decreasing signal levels is an interesting exercise.

The maximally flat Butterworth response of Fig. 2a is of course peak free, but peaking can be expected in the elliptic responses. In Fig. 2b it appears to be about 1% at F_o, corresponding to +0.086dB. This is within the maker's tolerance, also measured at 1kHz, which is -0.17 to +0.12dB (typically +0.05dB).

With the faster cut-off offered by the *MAX294*, somewhat larger peaking (–0.17 to +0.26) is to be expected. Figure 3a shows it. Note that measurement accuracy is limited by a variety of factors other than the detector time-constant mentioned above. For instance, the distortion of the sinewave test signal produced by IC_2, measured at 1kHz, is as much as 0.6%. It consists almost entirely of third harmonic, which is thus only 44dB down on the fundamental.

Even assuming the level of the latter is exactly constant over the sweep, using a peak detector circuit, a 0.05dB change in level can be expected at 333Hz, at which point the third harmonic sails out of the filter's passband. Thus a very clean, constant amplitude test signal indeed would be necessary to test the filter's passband ripple accurately. It would also be necessary for basic measurements on a highpass filter, where the harmonic(s) of the test signal would sweep into the filter's passband whilst the fundamental was still way down in the stopband. All the filters in the range offer very low total harmonic distortion (THD), around –70dB.



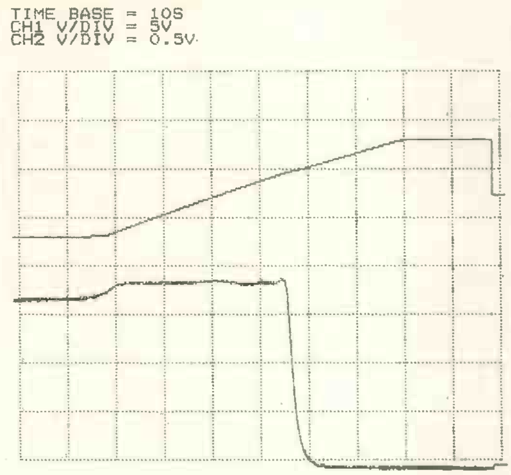


***Fig. 3a.** As Fig. 2a, but using the Max294 elliptic filter with its 1.2:1 F_s to F_o ratio, using a modified detector circuit.*
***Fig. 3b.** As 3a, but the detector circuit as in Fig. 1.*

Consequently the elliptic filters lend themselves very nicely to the construction of a digitally controlled audio oscillator as shown in **Fig. 4a**.

The *LS90* will divide by ten whilst giving a 50/50 mark/space ratio output. The F_{clock}/100 output of the second *LS90*, suitably level shifted, was applied to the *MAX294's* signal input, pin 8, and the clock input itself to pin 1. The MAX294 will operate on a single +5V rail (in which case the signal input should be biased at +2.5V) or, as here, on +5V and -5V rails.

Either way it will accept a standard 0 to +5V cmos clock input at up to 2.5MHz or, as it turns out in practice, a *74LSXX* input, though this is not stated in the data sheet. The *LS90* may be old hat, but it is nonetheless fast, so a clean clock drive and local decoupling were used to ensure no false counting due to glitches etc.

The attenuation of the *MAX294* at $3F_o$ is around 60dB. Given that the third harmonic component of the squarewave input to the device is 9.5dB down on the fundamental, the squarewave should be filtered into a passable sinewave with all harmonics 70dB or more down. This is comparable in level with the device's stated THD, so that although the *MAX293* could equally well be used in this application, its greater stopband attenuation would not in fact be exploited. The Butterworth *MAX291* also shows greater than 60dB attenuation at $3F_o$ relative to F_o: at $2F_o$ it is only just over 40dB relative, but of course the squarewave drive has no second harmonic. The *MAX291/293/294* are all equally suitable in this application.

Figure 4b shows a 1kHz sinewave output from the **Fig. 4a** circuit, lower trace; the 100kHz steps forming the waveform are clearly visible. At first sight, it looks very like the waveform out of a DDS direct digital synthesizer but there are one or two subtle differences. From a time point of view, the quantisation is always exactly 100 steps per cycle, whereas in a DDS it can be any number times (clock frequency divided by maximum accumulator count), the latter being typically 2^{32}.

Considering amplitude, the waveform is simply just not quantised. It is an example of a true peak measuring system where each step can take exactly the appropriate value for that point in a continuous sine wave. Figure 4b

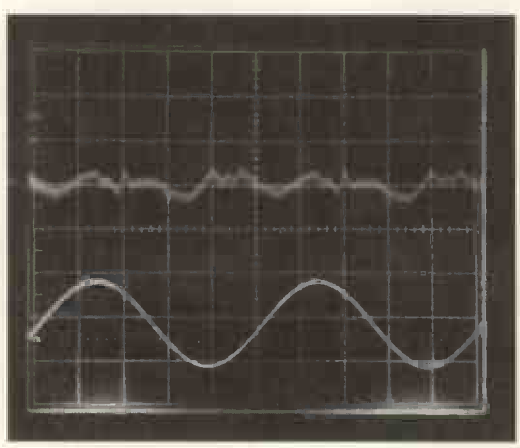
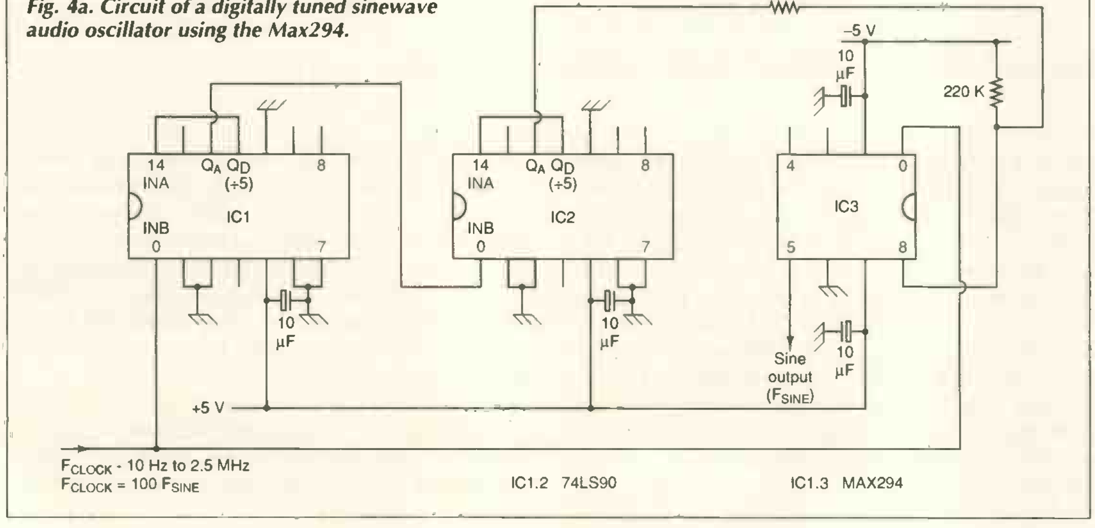
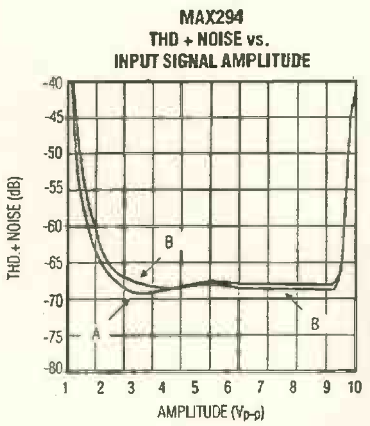


***Fig. 4b.** The circuit's output at 1kHz (lower trace) and the residual signal after filtering out the fundamental, representing the total harmonic distortion (upper trace).*

***Fig. 4a.** Circuit of a digitally tuned sinewave audio oscillator using the Max294.*





LABEL	f_{CLK} (Hz)	F_O (kHz)	INPUT FREQ. (Hz)	MEASMT BANDWIDTH
A	200k	2	200	30KHz
B	1M	10	1k	80KHz

Fig. 5a. THD + noise relative to the input signal amplitude for the MAX294

Pin Description

8-PIN	16-PIN	NAME	FUNCTION
	1, 2, 7, 8, 9, 10, 15, 16	N.C.	No Connect
1	3	CLK	Clock Input. Use internal or external clock.
2	4	V-	Negative Supply pin. Dual supplies: -2.375V to -5.500V. Single supplies: V- = 0V.
3	5	OP OUT	Uncommitted Op-Amp Output
4	6	OP IN-	Inverting Input to the uncommitted op amp. The noninverting op amp is internally tied to ground.
5	11	OUT	Filter Output
6	12	GND	Ground. In single-supply operation, GND must be biased to the mid-supply voltage level.
7	13	V+	Positive Supply pin. Dual supplies: +2.375V to +5.500V. Single supplies: +4.75V to +11.0V.
8	14	IN	Filter Input

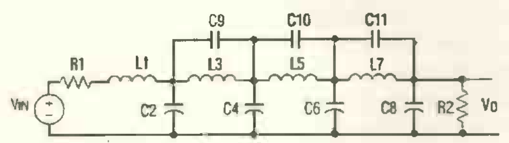
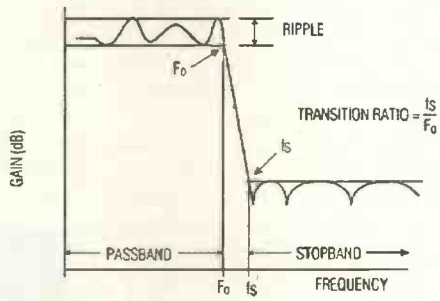


Fig. 5b. The MAX29X series filter structure emulates a passive eight pole lowpass filter. In the case of the elliptic types, this results in ripples in both the pass- and stopbands.



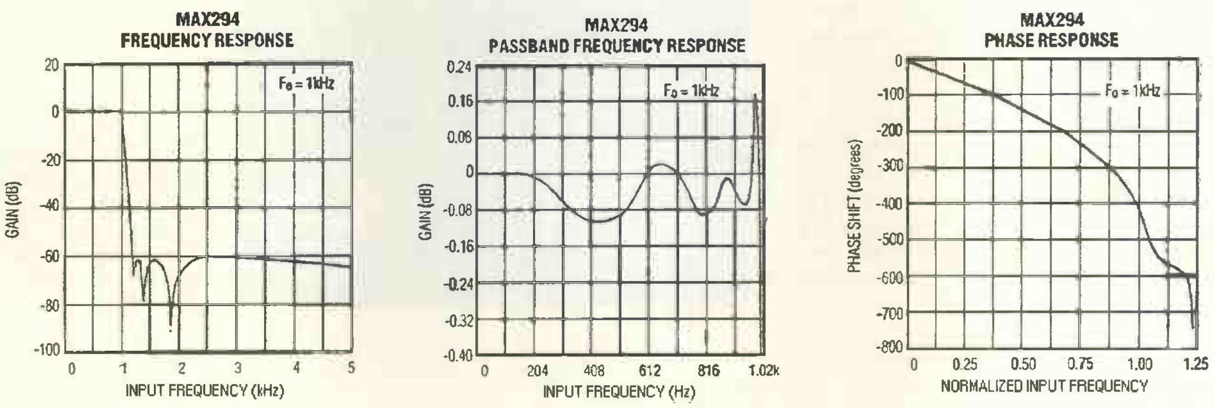


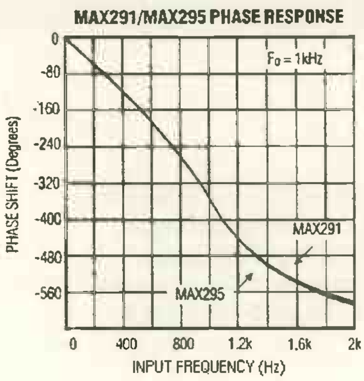
Fig. 5c. Passband and stopband performance for the MAX294 with a 100kHz clock (F_o = 1kHz).

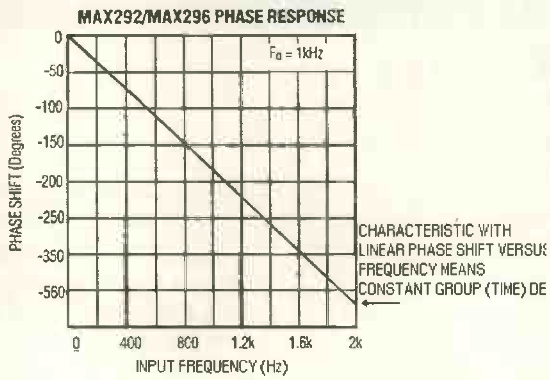
also shows residual THD (upper trace); and the monitor output of a meter on the 0.1% FSD range. Measured THD is 0.036% or 69dB down on the fundamental.

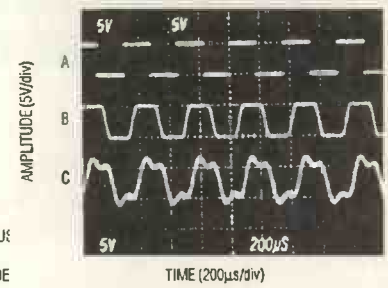
This agrees exactly with the manufacturer's data, **Fig. 5a**, which shows that the level of (THD + noise) relative to the signal is independent of the actual signal level over a quite a wide output range. The slight fuzziness of the THD trace is due to some 50Hz getting in to the experimental lash-up, not (as might be supposed) residual clock hash. The latter was suppressed by switching in the THD meter's 20kHz low pass filter; without this necessary precaution, the residual signal amounted to just over 1%.

Each of the *MAX29X* switched capacitor filters includes an uncommitted op amp which can be used for various purposes. It makes a handy anti-aliasing filter to precede the main switched capacitor section or can be used as a post-filter to reduce clock breakthrough in the output. Unfortunately, it cannot suppress it entirely, since it is part of the same very busy chip as the 8 pole switched capacitor filter section. Its use is illustrated in **Fig. 5e.**

Where a modest distortion figure of somewhere under 0.05% is adequate, an instrument based on the Fig. 4a circuit has certain attractive features. It can cover 0.1Hz to 25kHz with a constant amplitude output and much the







A: 3kHz INPUT SIGNAL
B: MAX292 BESSEL FILTER RESPONSE WITH Fo=10kHz
C: MAX 291 BUTTERWORTH FILTER RESPONSE WITH Fo=10kHZ

Fig. 5d. Comparison of the pulse response of the Bessel and Butterworth filter types.

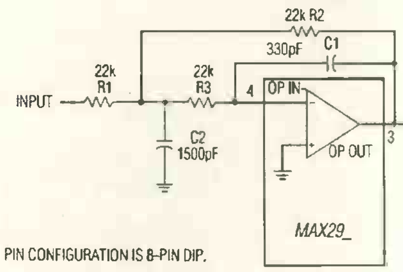


Fig. 5e. Use of the MAX29X's uncommitted op amp as an aliasing filter.

Table 1. Component values.

Corner frequency (Hz)	R_1 (kΩ)	R_2 (kΩ)	R_3 (kΩ)	C_1 (F)	C_2 (F)
100k	10	10	10	68p	330p
50k	20	20	20	68p	330p
25k	20	20	20	150p	680p
10k	22	22	22	330p	1.5n
1k	22	22	22	3.3p	15n
100	22	22	22	33n	150n
10	22	22	22	330n	1.5µ

Note: some approximations have been made in selecting preferred component values.

The passband error caused by a 2nd-order Butterworth can be calculated using the formula

$$\text{Gain error} = -10\log\left[1+\left(\frac{f}{f_c}\right)^4\right]\text{dB}$$

same THD over the whole range, given suitable post-filtering to suppress clock hash. The post filters need to be selected as appropriate, but with a clock frequency of 100 times the output frequency, each can cover a 20:1 frequency range or more. This means that only two or three are needed to cover the full 20Hz to 20kHz audio range, while four can cover the range 0.1Hz to 25kHz.

The clock can be fed to a counter with a 100ms gate time, providing near instantaneous digital readout of the output frequency down to 20Hz to a resolution of 0.1Hz, a feature which would require a 10s gate time in a conventional audio oscillator with digital read-out. If the clock is derived from a DDS chip, then the frequency can be set digitally, to crystal accuracy. The clock division ratio of 100 would reduce any phase-modulation spurs in the output of the DDS by 40dB: a necessary feature with many DDS devices.

The usual arrangement in a multipole active filter is to cascade a number of individual sections, each of which is solely responsible for one pole pair of the overall response. This can lead to substantial departures from the desired response, due to component tolerances in the individual two-pole sections, particularly the highest Q section(s).

Interestingly, the *MAX29X* series filters employ a design which emulates a passive ladder filter, **Fig. 5b**, so that any individual component tolerance error marginally affects the shape of the whole filter, rather than being concentrated on a particular peak. Ideally, the passband peaks and troughs are all equal, as are the stopband peaks. The actual typical performance (for the *MAX294*) is shown in **Fig. 5c**.

The Butterworth filter (with simple pre- and post-filters) provides a powerful anti-aliasing function to precede the A-to-D converter of a DSP (digital signal processor) system. The elliptic versions enable operation even closer to the Nyquist rate (half A/D's sampling frequency). The *MAX294* is suitable for 10 bit A-to-Ds and the max293 for 12 or 14 bit A-to-dDs. This assumes that the DSP system is interested only in the relative amplitudes of the frequency components of the input, and not in their relative phases. Where the latter is also important, to preserve the detailed shape of the input, the *MAX292* filter with its Bessel response is needed. Alias-free operation will then be possible only to a lower frequency; eg, one fifth of the Nyquist rate for a 10 bit system, since A_s = 60dB occurs at $5F_0$ for this device.

The Bessel filter with its constant group delay offers improved waveform fidelity over the Butterworth filter; this is graphically illustrated in **Fig. 5d**. The pulse response of the elliptic types would be even more horrendous than the Butterworth's.

The filtering function of a *MAX29X* would require a much greater expenditure of board space, power, money and number of chips if performed in DSP. These devices provide mighty filter power from their minuscule packages. ■

References

1. Design Brief "Logamps for radar and other uses", *EW + WW* April 1993 pp.314-317.

CFA – RIP?

Has the debate over the crossed field antenna at last reached a conclusion? Colin Davis presents the results of scientific testing on this electrically small antenna system.

Since the crossed-field-antenna (CFA) first appeared in 1989 [1] it has generated much debate about its performance claims and usefulness as an electrically small antenna. Now, following work carried specifically to investigate whether the CFA does operate as an efficient radiator as claimed, doubts about it seem overwhelming.

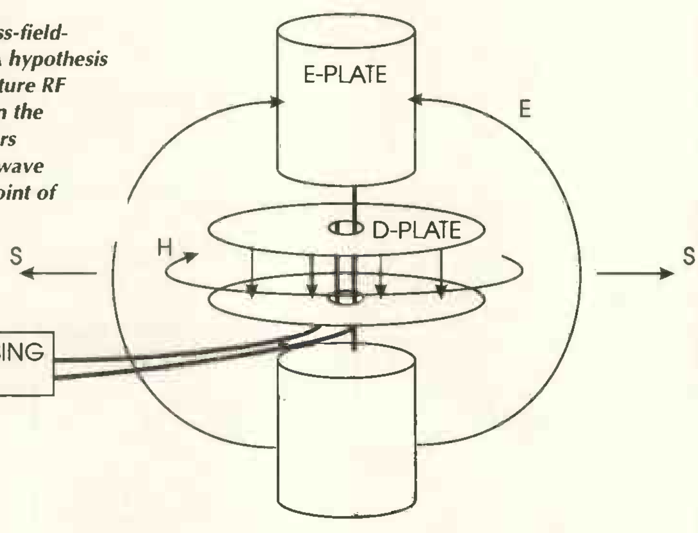
It is not possible to include all details of my year's work at Surrey University, studying the theory behind the antenna and the practical measurements made to support or refute the ideas. Nevertheless, my conclusion disputes the hypothesis behind the CFA and casts doubt on its performance claims.

Practical testing

Work was carried out on a crossed-field-antenna constructed to approximately half the linear dimensions of Maurice Hately's antenna, together with two reference antennas, enabling comparisons to be made with the CFA in operation. The reference antennas were simple half wave dipoles designed for use at 50MHz.

But one problem had to be solved before practical testing of the antenna could begin – power could not be successfully coupled to the CFA unless its input impedances were known so that suitable transformers could be designed to match to the 50Ω feeders being used. Little information is provided in available literature to indicate what the values might be. But a discussion with Mr Hately suggested that both CFA input impedances should be somewhere around 300Ω. Using this as a start point, 4:1 baluns were fitted to the CFA inputs to provide a reasonable match to the feeder cables and connected up a test rig. The phasing unit was of a similar design to Hately's and was used to provide the required transmitter signal

Transmit site antenna support at the University of Surrey, with CFA horizontally polarised. The support stands approx 4m high on the flat roof of a five storey building.

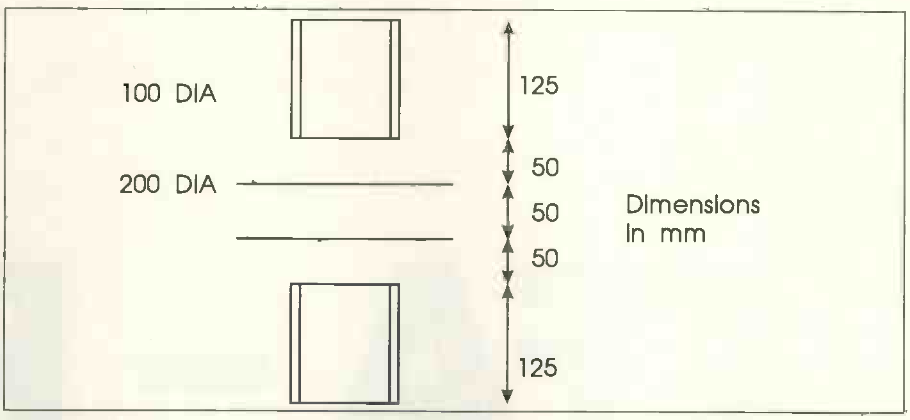


Barrel shaped cross-field-antenna. The CFA hypothesis states that quadrature RF currents flowing in the plates and cylinders synthesise an EM wave directly at their point of intersection.

splitting and 90° phasing of the two equal power outputs. In practice an RF trombone had to be included to provide precise adjustment of phase angle, and attenuators were included to help with equalising the power in both the CFA feeders.

No evidence has been found in past articles to indicate what was taking place within the two feeds to the CFA. The only results offered seem to be from measurements made in the transmitter feeder, so this aspect was of great interest because it would give results directly related to efficiency of the CFA.

The literature is also not clear on to what the 90° phase shift requirement referred: was it the incident waves travelling towards the CFA down the feeders; or the resultant signal at the CFA terminals due to forward and reflected waves? To cover both possibilities, both values were measured. Forward waves were measured with directional couplers, using a vector voltmeter which could display both voltage and the relative phases of two inputs (**Fig. 4**), and the resultants at the CFA inputs were measured using two high impedance probes.



Test CFA. The steel conductors were supported with insulating Perspex parts. The dimensions represent a 50% scaling on the author's original figures..

CFA theory of operation

An electromagnetic wave is comprised of two components, an electric field *E* and a magnetic field *H*, lying at right angles to each other and to the direction of propagation. Half the total power in the waves is said to exist in each of the components. The CFA aims to produce EM waves by synthesising the two fields in the correct orientation and phase from two distinct antenna structures, the E-plates and D-plates.

One form of the CFA, suggested by the inventors, is the barrel shaped device The two E-plates give rise to an electric field and the D-plates, which form a capacitor, generate a magnetic field around their circumference in accordance with Maxwell's wave equations. Vectors S denote the direction of wave propagation which should result if fields *E* and *H* coincide, in phase, in this vicinity. To obtain this in-phase condition, the signal applied to the D-plates must phase lag the E-plates feed by 90°. Also, to satisfy the condition that half the signal power is in each field component, the inventors suggest that half the signal power should be sent to the E-plates and half to the D-plates – the function of the phasing unit.

But the theory leads to an interesting observation: if the antenna is efficient and all the power sent to it is radiated when the above conditions are met, then the antenna inputs ought to appear purely resistive (since a perfect match would be made to the feeder cable); and if any phase change occurs between the signals incident on the antenna then some of the power will be reflected. So the input impedances will presumably appear reactive as well. In short, its characteristics should not necessarily be constant at a given frequency, as would be expected with a dipole for example.

CFA test results

Measurements were made on the CFA with 4:1 baluns on the inputs, with equal power in the forward wave feed signals and with the required 90° phase shift between them. Results are shown in **Table 1**.

The phase angle was varied by ±10° around the 90° point to allow for measurement errors. Even if the baluns on the antenna feeds were not of the correct ratio, an increase in radiated power and a dip in VSWR ought to be expected in the feeders when the phase angle was correctly adjusted for crossed-field operation. No such dip occurred.

In fact the VSWR remained constant indicating that the power radiated was not critically dependent on phase angle as suggested by Hately. The signal at the end of the antenna test range, received by a dipole in the same polarisation as the CFA, was –63dBm compared to –40dBm when the CFA was replaced by another dipole; a difference of 23dB. The figure means that the dipole was radiating 200 times as much power as the CFA, and clearly indicates that the CFA was not operating as an efficient antenna under these conditions, since that kind of difference is far too large to be accounted for by measurement errors.

Table 1. Measurements made on the CFA.

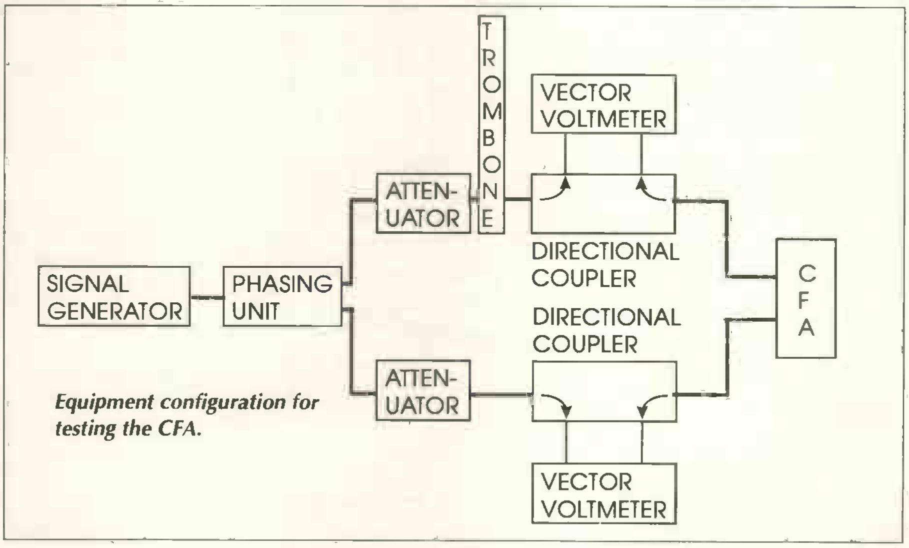
	D-plates	E-plates
Forward power (dBm)	3.5	3.5
Backward power (dBm)	–2.5	3.5
VSWR on feeder	3	-

For completeness, the resultant signals, measured by probes at the CFA inputs, were set up with equal amplitude (0.5V) and at 90° to one another; the received signal this time was, –65dBm. Again, sweeping the phase angle by ±10° did not improve the signal level.

The same tests were carried out on the CFA with a 1:1 balun fitted to the E-plates and a 4:1 balun on the D-plates. Forward wave measurement results are shown in **Table 2.**

Table 2. Forward wave measurements.

	D-plates	E-plates
Forward power (dBm)	5.5	5.5
Backward power (dBm)	2	3
VSWR on feeder	5	7



Equipment configuration for testing the CFA.

Table 3. Tests conducted in the two feeders.

	D-plates	E-plates
Forward power (dBm)	12.5	7
Backward power (dBm)	10.5	8
VSWR on feeder	9	17

Once again, no dip occurred in the feeder VSWRs as the phase angle was swept.

Literature on the crossed-field-antenna and articles by the CFA inventors have presented Smith charts and measurements made in the single cable connecting the transmitter to the phasing unit. But these have omitted to explain – or may be even consider – what was happening between the phasing box and the antenna. From the graphs given, it is clear that a good match had been obtained to the transmitter. But was all the power actually being radiated by the antenna?

To check this, the attenuators and RF trombone were removed from the test circuit, and the phasing unit was adjusted to give the minimum VSWR possible in the single feeder from the transmitter. In other words, the situation normally existing when measuring VSWR for an antenna system. Adjustment is quite critical, indicating a high Q system. Nonetheless, with steady hands a VSWR of 1.13 was achieved. Tests were again conducted in the two feeders to the CFA, yielding the results in **Table 3**.

The received signal at the end of the test range this time was –63dBm. Interestingly, more power returned from the E-plates than was incident, and it appears that energy was being coupled from the D-plates across to the E-plates. The results all suggest that most of the power from the transmitter was being dissipated by just heating up the antenna system.

Not all it is claimed

To sum up, the field strength measurements made at the receiver site showed that the CFA radiated signal levels were consistently 23dB below those made using a dipole (or worse), clearly showing that the CFA was not operating efficiently. The required feed phase angle of 90° was arranged for both the forward wave signals and the resultant signal at the plate terminals. In each case, no signal improvement or dip in VSWR was observed when sweeping the phase angle around the 90° point. The results demonstrated that the phase angle was unimportant to the operation of the antenna – at least under these conditions – contrary to Hately's suggestions.

Proving beyond any doubt that a theory is completely wrong, is extremely difficult. But it is possible to conduct experiments as scientifically and impartially as possible and to draw firm and well founded conclusions from them under the conditions tested. On this basis, the work conducted suggests the CFA is not all it claims to be – despite best efforts to make it work.

I rest my case.

Directional coupler design

Before work could begin in earnest on the crossed-field-antenna, directional couplers had to be constructed to monitor signals in the CFA feeders. The couplers were made of a small diecast box (dimensions 90 x 35 x 30mm) with BNC connectors for all the ports to enable quick and easy connection and disconnection when in use.

The main (through) transmission line of the coupler, was made of a piece of 0.5in copper pipe, chosen because it gave an impedance of almost exactly 50Ω when soldered between the two BNC connectors at either end of the box.

To tap off some power from the main line, another transmission line was set up along the side of the diecast box, resembling a microstrip line similar to those used in high frequency printed circuit board designs. For this, a piece of thin brass strip was cut to a length slightly longer than the distance between the back two BNC connectors. Its width was about 6mm.

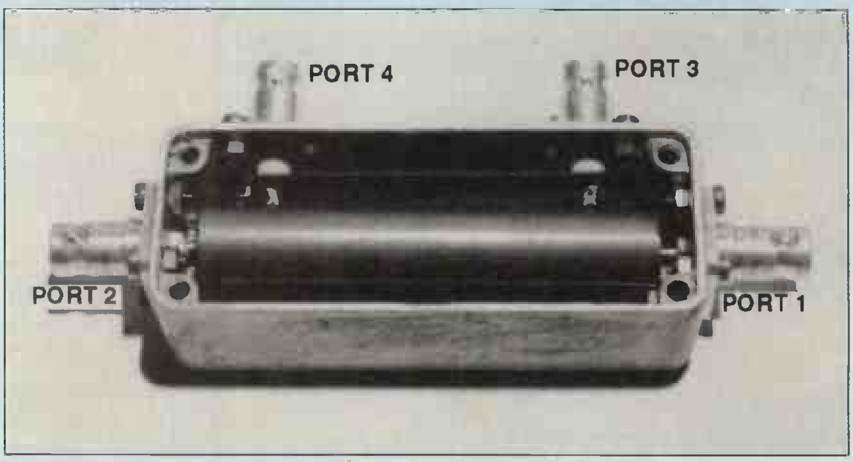
By drilling one small hole at either end, the strip could be fitted over the "inner" terminals protruding from the back of the BNC sockets. Then, by sliding it in and out parallel to the side of the box, the characteristic impedance of the secondary line could be adjusted.

A network analyser was used to find a position where the line impedance was 50Ω, after which the brass strip was fixed in place with solder. The amount of power coupled to the secondary line is dependent on the length, width and distance of the strip from the main copper line.

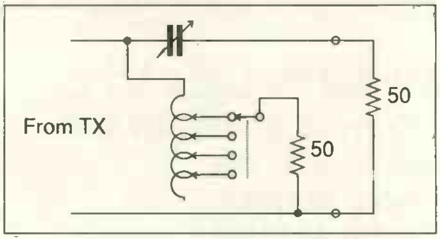
It so happens that these dimensions give a –40dB coupling at the test frequency of 50MHz. In other words, for a of 0dBW going in at port 1 and out of port 2, a signal of –40dBW would appear at port 3. Coupling is ideally from port 1 to port 3 and from port 2 to port 4. But if the characteristic impedances of the lines in the directional coupler are not exactly 50Ω then signal reflections will occur causing power to be coupled to and from the wrong ports. In practice, with standard measuring equipment, making directional couplers such as these work very well indeed is not difficult: one of the devices constructed achieved a directivity of 45dB. The ability to construct such devices satisfactorily, reduces the cost of undertaking work on the crossed-field-antenna considerably.

Anyone wishing to make a coupler should be aware that although expensive test equipment was available to set up these devices, it was quite a luxury and the same result could equally well be achieved using a transmitter, dummy load and an accurate VSWR meter to measure reflections.

Directional coupler main line can be seen between the two BNC connectors at either end of the box.



CFA phasing unit. The 50Ω resistors represent ideal input impedances of the CFA. If the CFA inputs are exactly 50Ω then the two phasing unit outputs can be set equal in amplitude together with a 90° phase shift.



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3. Bryan C Wells. "CFA Experiments". *EW + WW*, March 1990.
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5. Hately and Kabbery. GB Patent No. 2,215,524 A. Published 20-09-1989,

Colin Davis conducted his study of the CFA at the University of Surry as his final year project, part of a degree course in Electronic & Electrical Engineering. He graduated with a

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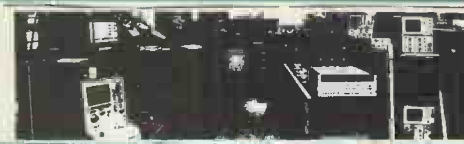

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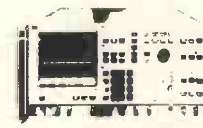

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This is a practical guide to real-time programming, the programs provided having been tested and proved. It is a distillation of the teaching of computer-assisted engineering at Humberside Polytechnic, at which Dr Hutchings is a senior lecturer.

Source code listings for the programs described in the book are available on disk.

CIRCUIT IDEAS

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Overcurrent protector

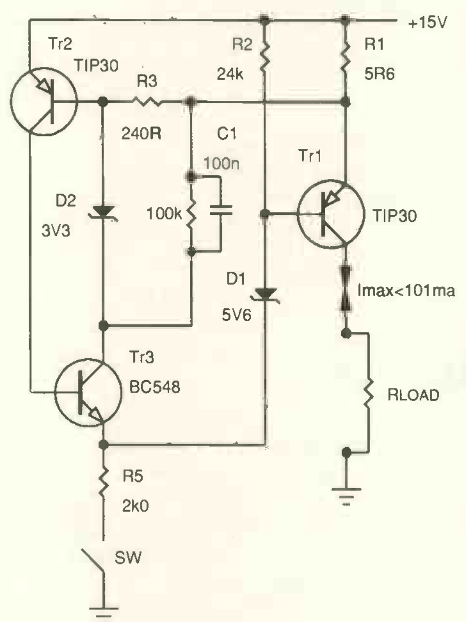
Placed between a power source and its load, this circuit protects the power source from excess current drain. It limits current surge at switch-on and acts as a circuit-breaker if a short or circuit malfunction causes current to exceed a preset limit in normal operation.

With the switch on, the series transistor Tr_1 is biased on via R_5 and the zener D_1, bias being set to give the required maximum output current of 100mA or less. The load receives virtually the entire rail voltage.

Excess current flowing through R_1 brings Tr_2, and therefore Tr_3, into conduction, diverting base current from Tr_1 and limiting output current for a short time, as in a switch-on surge. If the high current persists, C_1 charges up to the voltage of zener D_2, which connects Tr_3 collector to Tr_2 base and produces an avalanche effect. All the available bias current is now diverted from the output transistor and no output current passes.

The time to complete cut-off depends on the time constant C_1R_5.

N I Lavrantiev
Schiolkovo
Moscow Region Russia



Overcurrent protection circuit prevents power-supply damage caused by switch-on surges and acts as a circuit breaker in the presence of a short-circuit.

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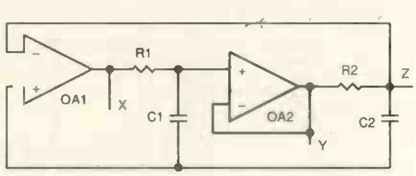
Function generator is digitally programmed

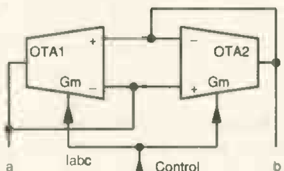
By means of a simple modification, the function generator put forward by R W J Barker in *Circuit Ideas* for June, 1991 becomes programmable from a digital input word.

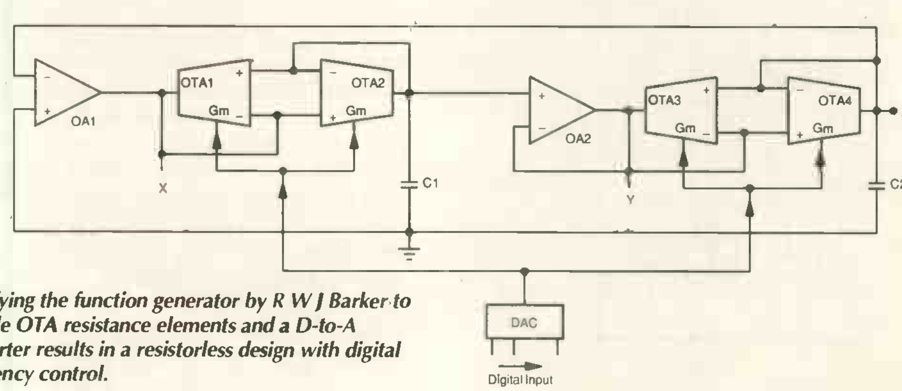
Figure 1 shows the original circuit, which is a ring oscillator producing approximate square, triangular and sine waves at *x,y* and *z* respectively, its frequency being determined by R_1C_1. In **Fig. 2**, the resistance seen between (a) and (b) is $R_{ab}=1/g_m$, where $g_m=I_{ABC}/2V_T$, the transconductance of the two transconductance amplifiers, I_{ABC} being the automatic bias control current and V_T the thermal voltage. Since the value of R_1 controls the frequency of oscillation, replacing R_1 with this circuit allows linear frequency control by variation of the input current.

Adding a digital-to-analogue converter, as shown in **Fig. 3**, produces a digitally-controlled, variable frequency function generator, which has been realised using *LM13600* operational transconductance amplifiers and *741* op-amps.

Muhammad Taher Abuelma'atti and Sulaiman Al-Gharbi Al-Sayed
King Fahd University of Petroleum and Minerals
Dhahran, Saudi Arabia







Modifying the function generator by R W J Barker to include OTA resistance elements and a D-to-A converter results in a resistorless design with digital frequency control.

Voltage-to-period converter

As in traditional designs, this converter relies on a ramp technique, but in this case the flyback is initiated in a different manner and jitter significantly reduced.

The current source supplies charging current to C_1, which ramps linearly in a positive direction. As the ramp voltage reaches V_{in}, the *LM311* comparator output goes positive, the edge being differentiated by C_2R_1. The resulting pulse turns Tr_2 on, blocking the comparator at the strobe input and maintaining the output condition for a time determined by the time constant of the *CR*. It also turns on Tr_1 to discharge C_1. Ramp time *T* is dependent only on the input voltage and the discharge time must only be long enough for full discharge of C_1.

The relationship between *T* and V_{in} is adjusted by varying the value of C_1 or current source output.

Viacheslav Shkarupin
Kiev
Ukraine

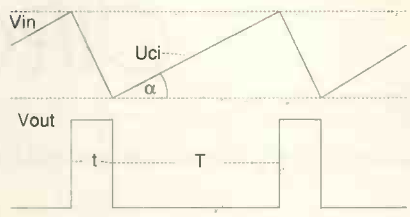


Fig. 1. Positive feedback from the comparator output to initiate flyback reduces jitter in this voltage-to-period converter.

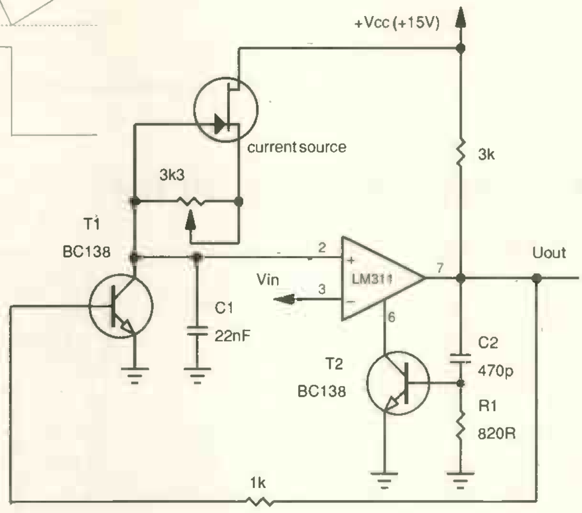


Fig. 2. Slope of ramp is dependent on the value of C1 and the current source. Period T bears a linear relationship to V_{in}.

Simple, high-gain amplifier

Two extra transistors in a cascode amplifier produce a much higher gain, a greater bandwidth and a reduced output impedance.

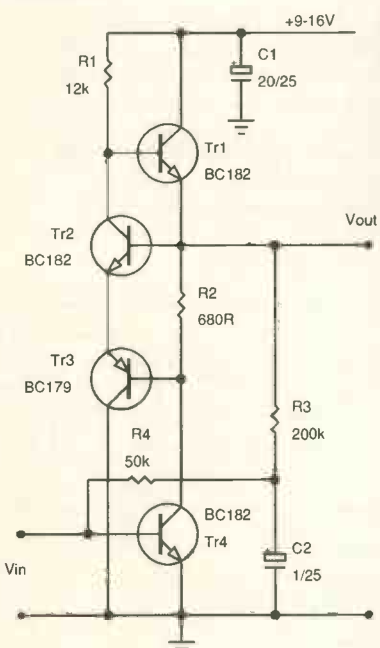
The load R_2 resistor of Tr_4, the input transistor, has virtually no voltage across it, because of its inclusion in the amplifier made up by $Tr_{1,2,3}$; current through it is therefore practically zero. Equivalent load, and therefore gain, of Tr_4 is accordingly extremely high.

Since the $Tr_{1,2,3}$ amplifier's frequency response is wide-band, the resulting amplifier exhibits a gain of over 60dB over a bandwidth of 850kHz into 50Ω, using *BC182* and *BC179* transistors. Feedback through R_2 gives an output impedance of only a few ohms. Capacitor C_2 across the bias resistors for the input stage removes AC feedback.

I have used the amplifier in the output of an IF amplifier, in which it gave a good match to a crystal detector.

G Mirsky
Akademtekh R&D Centre
Moscow
Russia

Three transistors in the load circuit of Tr4 produce high gain, wide band and low output impedance.



Precise power output stage

When a series regulator must both source and sink current, or if the standing current in an audio power output stage must be accurately set independently of temperature, then this circuit is one solution.

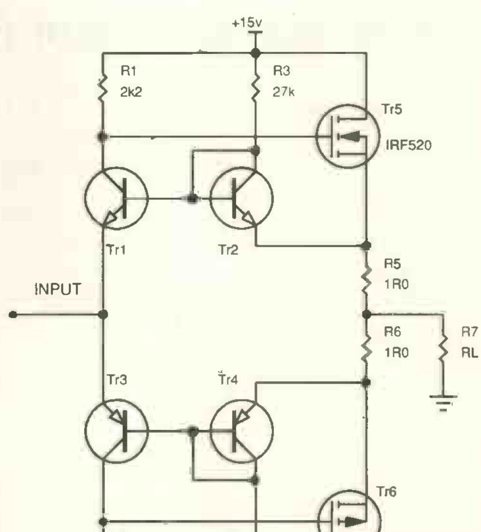
Since top and bottom circuits are identical, apart from polarity, the top half will be described. When quiescent, the current mirror $Tr_{1,2}$ has a voltage between the emitters which depends on standing currents according to $V=V_T\ln I_1/I_2$, where V_T is the temperature voltage kT/q of 26mV at room temperature. If $I_1=10I_2$, V=59mV at 25°C ambient and setting $R_{5,6}$ at 1Ω puts the standing current in Tr_5 at 59mA, independently of junction temperature, the values of $R_{5,6}$ and the ratio I_1/I_2 being adjusted to suit one's needs.

If $R_1=R_2$, small-signal input impedance is

$$R_{in}=\frac{1}{2}\frac{g_{fe}R_1R_L}{1+g_{fe}R_L}\approx\frac{R_1}{2},$$

when $g_{fs}R_L>1$, where g_{fs} is the transconduxctance of $Tr_{5,6}$, although replacing the resistors by current sources will increase that. Output impedance is $R_7/2$. Match the mirror pairs to avoid errors and to prevent possible thermal runaway.

Terence S Finnegan
Carlisle



Current mirrors in this power output stage, which sources and sinks current, allow accurate setting of standing current,

Near-field probes for EMC testing

Before spending money on having a new product assessed for its EMC, it might be advisable to check roughly on its noisiness while still in development. The diagrams show two probes for near-field "sniffing": an electrostatic probe and an electromagnetic type.

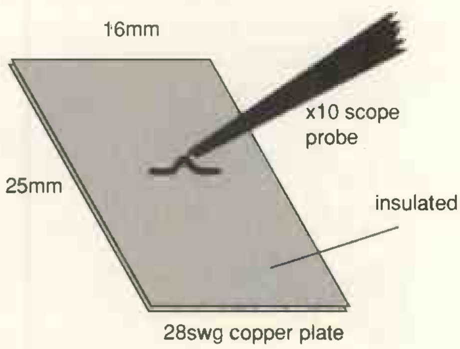
The former is a thin plate of copper or tinned steel measuring about 16 by 25mm and having a hoop of 20swg wire soldered to it so that an oscilloscope probe can clip onto it. The plate is insulated with tape, since it is used near live circuits. A 25mm length of wire carrying a 4Vpk-pk, 31kHz square wave gave a 5mV pk-pk oscilloscope deflection at a distance of 10mm. Holding the plate edge-on to a PCB track gives the best signal.

As a less expensive alternative to the Tektronix *Alternating Current Probe* with the jaws open which, as the table shows, worked reasonably well, my solution is 10 turns of enamelled copper wire at 20mm diameter. On signal transitions, this gives triangular spikes about 100ns wide, which trigger most oscilloscopes; loop currents of 2mA pk-pk are visible at 5mV/div. The 100Ω resistor gives a slightly under-damped response and a larger signal than with 50Ω.

The table shows measurements made with the detector coil at the centre of the source loop. If a spectrum analyser or a fast, sensitive oscilloscope is used, the number of turns can be reduced to give a truer spectral response.

C J D Catto
Elsworth
Cambridgeshire

Fig.1. Near-field electrostatic probe allows low-cost testing of prototype equipment for electromagnetic compatibility.



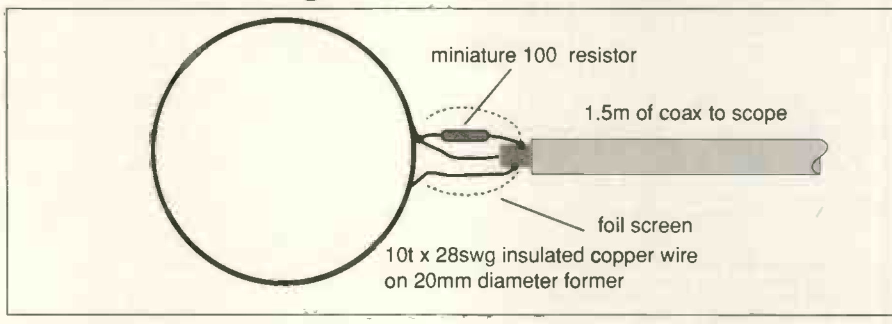


Fig.2. Electromagnetic probe.

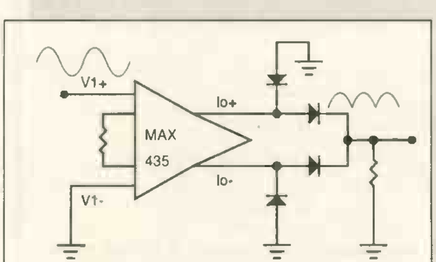
source detector	20mA in loop	40mm dia.	60mm dia.	decay time
Tek. P6021 probe (open)				
	switch to 2mA/mV	1.5mV	1.0mV	5µs
	switch to 10mA/mV	0.3mV	0.2mV	25µs
10t on 20mm dia with 100Ω		40mV	25mV	100ns spike

Fast full-wave rectifier

Loosely based on a design by Lidgey and Toumazou (*EW*[3], November 1987, p.1115), in which current mirrors sensed the supply current of op-amps, this circuit uses a *MAX435* wide-band, differential-output transconductance amplifier to give full-wave rectification of signals up to 250MHz. Output is $4Z_L/V_{in}$ for the *435*. You could also try the Burr-Brown *OPA660*, which offers 700MHz-plus operation.

Peter May

An up-date on a design by Lidgey and Toumazou, using a MAX435 or a Burr-Brown OPA660 for very high-speed rectification.



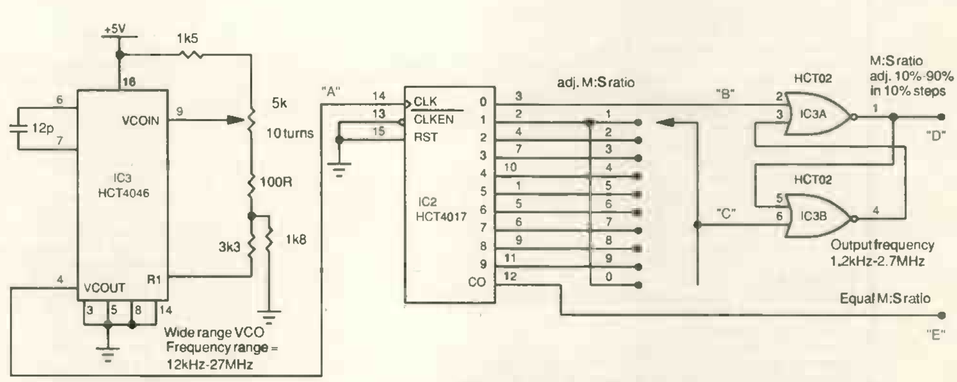
Independent m:s adjustment for wide-band pulse gen

This circuit delivers square waves and rectangular waves with a mark:space ratio of 10-90% at frequencies from 1.2kHz to 2.7MHz

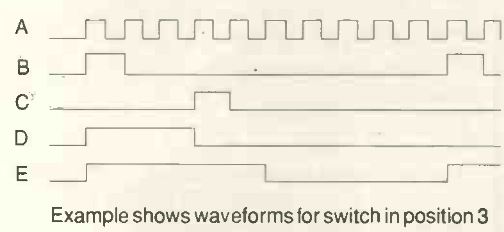
Frequency generation comes from the voltage-controlled oscillator IC_1 and associated components, the output of which is adjustable from 12kHz to 27MHz by means of the potentiometer.

Johnson counter IC_2 provides a Set pulse from the Q0 output to the SR flip-flop IC_3, the corresponding Reset pulse coming by way of the selector switch from outputs Q1-Q9, output frequencies being $^1/_{10}$ of the input from IC_1, as is the square waveform from carry output C0. ■

W Dijkstra
Waalre, The Netherlands



Three ICs form a 1.2kHz-2.7MHz pulse and square-wave generator having a mark-to-space ratio adjustable in 10% steps from 10% to 90%.



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HP Signal Generator type 8656A – 0.1-990Mc/s. AM/FM – £2000.
HP 8699B Sweep PI – 0.1-4GHz £750 – HP8690B Mainframe £250.
Racal/Dana 9301A-9302 RF Millivoltmeter – 1.5-2GHz – £250-£400.
Racal/Dana Counters 9915M – 9916 – 9917 – 9921 – £150 to £450. Fitted FX standards.
Racal/Dana Modulation Meter type 9009 – 8Mc/s – 1.5GHz – £250.
Racal – SG Brown Comprehensive Headset Tester (with artificial head) Z1A200/1 – £350.
Marconi AF Power Meter type 893B – £200.
Marconi RCL Bridge type TF2700 – £150.
Marconi/Saunders Signal Sources type – 6058B – 6070A – 6055B – 6059A – 6057B – 6056 – £250–£350. 400Mc/s to 18GHz.
Marconi TF1245 Circuit magnification meter + 1246 & 1247 Oscillators – £100-£300.
Marconi microwave 6600A sweep osc., mainframe with 6650 PI – 18-26.5GHz or 6651 PI – 26.5-40GHz – £1000 or PI only £600.
Marconi distortion meter type TF2331 – £150, TF2331A – £200.
Microwave Systems MOS/3600 Microwave frequency stabilizer – 1GHz to 40GHz £1k.
Tektronix Plug-Ins 7A13 – 7A14 – 7A18 – 7A24 – 7A26 – 7A11 – 7M11 – 7S11 – 7D10 – 7S12 – S1 – S2 – S6 – S52 – PG506 – SC504 – SG502 – SG503 – SG504 – DC503 – DC508 – DD501 – WR501 – DM501A – FG501A – TG501 – PG502 – DC505A – FG504 – P.O.R.
Alltech Stoddart receiver type 17/27A – .01–32Mc/s – £2500.
Alltech Stoddart receiver type 37/57 – 30-1000Mc/s – £2500.
Alltech Stoddart receiver type NM65T – 1 to 10GHz – £1500.
Gould J3B Test oscillator + manual – £200.
Infra-red Binoculars in fibre-glass carrying case – tested – £100. Infra-red AFV sights £100.
ACL Field intensity meter receiver type SR – 209 – 6. Plugs-ins from 5Mc/s to 4GHz – P.O.R.
Tektronix 491 spectrum analyser – 1.5GHz-40GHz – as new – £1000 or 10Mc/s 40GHz.
Tektronix Mainframes – 7603 – 7623A – 7633 – 7704A – 7844 – 7904 – TM501 – TM503 – TM506 – 7904 – 7834 – 7104.
Knott Polyskanner WM1001 + WM5001 + WM3002 + WM4001 – £500.
Alltech 136 Precision test RX + 13505 head 2 – 4GHz – £350.
SE Lab Eight Four – FM 4 Channel recorder – £200.
Alltech 757 Spectrum Analyser – 001 22GHz – Digital Storage + Readout – £3000.
Dranetz 606 Power line disturbance analyser – £250.
Precision Aneroid barometers – 900-1050Mb – mechanical digit readout with electronic indicator – battery powered. Housed in polished wood carrying box – tested – £100-£200-£250. 1, 2 or 3.
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HP141T+8552A or B IF-8555A RF – 10Mc/s–18GHz-A IF – £2400 or B IF – £2500.
HP141T+8552A or B IF-8556A RF – 20Hz–300kHz-A IF-A IF – £1200 or B IF – £1300.
HP8443A tracking generator/counter – 100kHz–110Mc/s – £500.
HP8445B tracking pre-selector DC–18GHz – £750.
HP ANZ UNITS AVAILABLE SEPARATELY – NEW COLOURS – TESTED.
HP141T mainframe – £550 – 8552A IF – £450 – 8552B IF – £550 – 8553B RF – 1kHz–110Mc/s – £550 – 8554B-RF – 100kHz–1250Mc/s – £650 – 8555A-RF – 10Mc/s–18GHz – £1550.
HP 3580A LF-spectrum analyser – 5kHz to 50kHz – LED readout – digital storage – £1600 with instruction manual – Internal rechargeable battery.
Tektronix 7D20 plug-in 2-channel programmable digitizer – 70 Mc/s – for 7000 mainframes – £500 – manual – £50.
Datron 1065 Auto Cal digital multimeter with instruction manual – £500.
Racal MA 259 FX standard. Output 100kc/s–1Mc/s–5Mc/s – internal NiCad battery – £150.
Aerial array on metal plate 9"×9" containing 4 aerials plus Narda detector – .100–11GHz. Using N type and SMA plugs & sockets – ex eqpt – £100.
EIP 451 microwave pulse counter 18GHz – £1000.
Marconi RF Power Amplifier TF2175 – 1.5Mc/s to 520Mc/s with book – £100.
Marconi 6155A Signal Source – 1 to 2 GHz – LED readout – £600.
Schlumberger 2741 Programmable Microwave Counter – 10Hz to 7.1GHz – £750.
Schlumberger 2720 Programmable Universal Counter 0 to 1250Mc/s – £600.
HP 2225CR Thinkjet Printer – £100.
TEK 576 Calibration Fixture – 067-0597-99 – £250.
HP 8006A Word Generator – £150.
HP 1645A Data Error Analyser – £150.
Texscan Rotary Attenuators – BNC/SMA 0-10-60-100DBS – £50-£150.
HP 809C Slotted Line Carriages – various frequencies to 18GHZ – £100 to £300.
HP 532-536-537 Frequency Meters – various frequencies – £150-£250.
Barr & Stroud variable filter EF3 0.1Hz–100kc/s + high pass + low pass – £150.
S.E. Lab SM215 Mk11 transfer standard voltmeter – 1000 volts.
Alltech Stoddart P7 programmer – £200.
H.P. 6941B multiprogrammer extender. £100.
Fluke Y2000 RTD selector + **Fluke 1120A** IEEE-488-translator + **Fluke 2180** RTD digital thermometer + 9 probes. £350 all three items.
H.P. 6181 DC current source. £150.
H.P. 59501A – HP-IB isolated D/A/power supply programmer.
H.P. 3438A digital multimeter.
H.P. 6177C DC current source. £150.
H.P. 6207B DC power supply.
H.P. 741B AC/DC differential voltmeter standard (old colour) £100.
H.P. 6209B DC power unit.
Fluke 80 high voltage divider.
Fluke 431C high voltage DC supply.
Tektronix M2 gated delay calibration fixture. 067-0712-00.
Tektronix precision DC divider calibration fixture. 067-0503-00.
Tektronix overdrive recovery calibration fixture. 067-0608-00.
Avo VCM163 valve tester+book £300.
H.P. 5011T logic trouble shooting kit. £150.
Marconi TF2163S attenuator – 1GHz. £200.
PPM 8000 programmable scanner.
Fluke 730A DC transfer standard.
B&K 4815 calibrator head.
B&K 4812 calibrator head.
Farnell power unit H60/50 – £400 tested.
H.P. FX doubler 938A or 940A – £300.
Racal/Dana 9300 RMS voltmeter – £250.
H.P. sweeper plug-ins – 86240A – 2–8.4GHz – 86260A – 12.4–18GHz – 86260AH03 – 10–15GHz – 86290B – 2–18.6GHz. 86245A 5.9–12.4GHz.
Telequipment CT71 curve tracer – £200.
H.P. 461A amplifier – 1kc–150Mc/s – old colour – £100.
H.P. 8750A storage normalizer.
Tektronix oscilloscopes type 2215A – 60Mc/s – c/w book & probe – £400.
Tektronix monitor type 604 – £100.
Marconi TF2330 or **TF2330A** wave analysers – £100–£150.
HP5006A Signature Analyser £250 + book.
HP10783A numeric display. £150.
HP 3763A error detector. £250.
Racal/Dana signal generator 9082 – 1.5–520Mc/s – £800.
Racal/Dana signal generator 9082H – 1.5–520Mc/s – £900.
Claude Lyons Compuline – line condition monitor – in case – LMP1+LCM1 £500.
Efratom Atomic FX standard FRT – FRK – .1–1–5–10Mc/s. £3K tested.
Racal 4D recorder – £350 – £450 in carrying bag as new.
HP8350A sweep oscillator mainframe + HP11869A RF PI adaptor – £1500.
Alltech – precision automatic noise figure indicator type 75 – £250.
Adret FX synthesizer 2230A – 1Mc/s. £250.
Tektronix – 7S12–7S14–7T11–7S11–S1–S52–S53.
Rotek 610 AC/DC calibrator. £2K + book.
Marconi TF2512 RF power meter – 10 or 30 watts – 50 ohms – £80.
Marconi multiplex tester type 2830.
Marconi digital simulator type 2828A.
Marconi channel access switch type 2831.
Marconi automatic distortion meter type TF2337A – £150.
Marconi mod meters type TF2304 – £250.
HP 5240A counter – 10Hz to 12.4GHz – £400.
HP 3763A error detector.
HP 8016A word generator.
HP 489A micro-wave amp – 1–2GHz.
HP 8565A spectrum analyser – .01–22GHz – £4k.
HP 5065A rubidium vapour FX standard – £5k.
Fluke 893A differential meters – £100 ea.
Systron Donner counter type 6054B – 20Mc/s–24GHz – LED readout – £1k.
Takeda Riken TR4120 tracking scope + TR1604P digital memory.
EG&G Parc model 4001 indicator + 4203 signal averager PI.
Systron Donner 6120 counter/timer A+B+C inputs – 18GHz – £1k.
Racal/Dana 9083 signal source – two tone – £250.
Systron Donner signal generator 1702 – synthesized to 1GHz – AM/FM.
Systron Donner microwave counter 6057 – 18GHz – Nixey tube – £600.
Racal/Dana synthesized signal generator 9081 – 520Mc/s – AM–FM. £600.
Farnell SSG520 synthesized signal generator – 520Mc/s – £500.
Farnell TTS520 test set – £500 – both £900.
Tektronix plug-ins – AM503 – PG501 – PG508 – PS503A.
Tektronix TM515 mainframe + TM5006 mainframe.
Cole power line monitor T1085 – £250.
Claude Lyons LCM1P line condition monitor – £250.
Rhodes & Schwarz power signal generator SLRD-280 – 2750Mc/s. £250–£600.
Rhodes & Schwarz vector analyser – ZPV+E1+E3 tuners – .3–2000Mc/s.
Bell & Howell TMA3000 tape motion analyser – £250.
Ball Efratom PTB-100 rubidium standard mounted in Tek PI.,
Ball Efratom rubidium standard PT2568-FRKL.
Trend Data tester type 100 – £150.
Farnell electronic load type RB1030-35.
Fairchild interference analyser model EMC-25 – 14kc/s–1GHz.
Fluke 1720A instrument controller+keyboard.
Marconi 2442 – microwave counter – 26.5GHz – £1500.
Racal/Dana counters – 9904 – 9905 – 9906 – 9915 – 9916 – 9917 – 9921 – 50Mc/s – 3GHz – £100–£450 – all fitted with FX standards.
B&K 7003 tape recorder – £300.
B&K 2425 voltmeter – £150.
B&K 4921+4149 outdoor microphone.
Wiltron sweeper mainframe 610D – £500.
HP3200B VHF oscillator – 10–500Mc/s – £200.
HP3747A selective level measuring set.
HP3586A selective level meter.
HP5345A electronic counter.
HP4815A RF vector impedance meter c/w probe. £500–£600.
Marconi TF2092 noise receiver. A, B or C plus filters.
Marconi TF2091 noise generator. A, B or C plus filters.
Tektronix oscilloscope 485 – 350Mc/s – £500.
HP180TR, HP182T mainframes £300–£500.
Bell & Howell CSM2000B recorders.
HP5345A automatic frequency convertor – .015–4GHz.
Fluke 8506A thermal RMS digital multimeter.
HP3581A wave analyser.
Philips panoramic receiver type PM7800 – 1 to 20GHz.
Marconi 6700A sweep oscillator+6730A – 1 to 2GHz.
Wiltron scaler network analyser 560+3 heads. £1k.
R&S signal generator SMS – 0.4–1040Mc/s – £1500.
HP8558B spectrum ANZ PI – .1–1500Mc/s – o/c – £1000. N/C – £1500 – **To fit HP180 series mainframe available – £100 to £500.**
HP8505A network ANZ + 8503A S parameter test set + 8501A normalizer – £4k.
HP8505A network ANZ + 8502A test set – £3k.
Racal/Dana 9087 signal generator – 1300Mc/s – £2k.
Racal/Dana VLF frequency standard equipment. Tracor receiver type 900A + difference meter type 527E+rubidium standard type 9475 – £2750.
Marconi 6960–6960A power meters with 6910 heads – 10Mc/s – 20GHz or 6912 – 30kHz–4.2GHz – £800–£1000.
HP8444A–HP8444A opt 59 tracking generator £1k–£2k.
B&K dual recorder type 2308.
HP8755A scaler ANZ with heads £1k.
Tektronix 475 – 200Mc/s oscilloscopes – £350 less attachments to £500 c/w manual, probes etc.
HP signal generators type 626 – 628 – frequency 10GHz–21GHz.
HP 432A–435A or **B–436A** – power meters+powerheads – 10Mc/s–40GHz – £200–£280.
HP3730B down convertor – £200.
Bradley oscilloscope calibrator type 192 – £600.
Spectrascope SD330A LF realtime ANZ – 20Hz–50kHz – LED readout – tested – £500.
HP8620A or **8620C** sweep generators – £250 to £1k with IEEE.
Barr & Stroud variable filter EF3 0.1Hz–100kc/s+high pass+low pass – £150.
Tektronix 7L12 analyser – .1Mc/s–1.8GHz – £1500 – **7L14 ANZ** – £2k.
Marconi TF2370 spectrum ANZ – 110Mc/s – £1200–£2k.
Marconi TF2370 spectrum ANZ+TK2373 FX extender 1250Mc/s+trk gen – £2.5k–£3k.
Racal receivers – RA17L–RA1217–RA1218–RA1772–RA1792 – P.O.R.
Systron Donner microwave counter 6057 – 18GHz – nixey tube – £600.
HP8614A signal gen 800Mc/s–2.4GHz old colour £200, new colour £400.
HP8616A signal gen 1.8GHz–4.5GHz old colour £200, new colour £400.

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

Frugal A-to-D. Intended, in the main, for battery-powered, portable equipment, Analog's *AD7883* 12-bit, sampling analogue-to-digital converter is powered by a 3-3.6V rail from which it uses 8mW in normal operation; in its power-saving mode, power consumption is 1mW. Signal:noise ratio is a minimum of 69dB and THD is –80dB. No external components are needed to use the device as a complete 12-bit data acquisition system when the reference is derived from the supply line. Analog Devices Ltd, 0932 232222.

Delta-sigma D-to-A. Crystal's *CS4303* delta-sigma digital-to-analogue converter for digital audio implements eight times interpolation and 64 times oversampled delta-sigma modulation to give a 107dB dynamic range up to 20kHz. The pass band is flat to within 0.0002dB to 21.8kHz and interchannel isolation is 115dB. An evaluation board is available. Sequoia Technology Ltd, 0734 311822.

Discrete active devices

Varactor diodes. Intended for use in voltage-controlled oscillators in mobile communications, the *BBY51* from Siemens is designed for the 900MHz band, while the *BBY52* is meant to operate between 1.5GHz and 2.5GHz. Series resistance of both is 0.5Ω at 1V and 1GHz. They are made as double diodes with a common cathode in SOT23 or as single diodes in SOD323. Siemens plc, 0932 752631.

Digital signal processor

Viterbi decoder. Qualcomm's Viterbi digital decoders are now obtainable in the UK from Chronos. *Q1601* decoders operate at 10Mbit/s and are full custom Viterbi systems in one chip which allow Rate 1/2 coding and 3-bit soft decision symbol inputs with V35 data descrambling, channel bit error rate, QPSK and OQPSK modems with no external circuitry. Coding gain is 5.2dB. Chronos Technology Ltd, 0989 85471.

1GOPS on a PC board. Allowing the development of multiprocessor systems with a processing power of a thousand Mflops, Loughborough's *QPC/C40* 200Mflops board has sites for four Texas Instruments *TMS320C40* DSP chips. An LSI asic and hardware links allow the *C40* modules to be interconnected in a number of topologies, including connection to other boards to form very large parallel processing systems. Loughborough Sound Images Ltd, 0509 231843.

Histogram chip. The *HSP48410* from Harris Semiconductor is a dedicated histogrammer and accumulating buffer. It has a 40MHz clock rate a 10-bit pixel resolution to analyse up to 1024 grey levels. On-chip memory is in 1K by 24bit form, with access by a 16 or 24-bit, three-state bus. The chip generates a histogram of grey levels in images up to 4096 by 4096 pixels and calculates the number of occurrences of each level for analysis or enhancement. Macro Group, 0628 604383.

Linear integrated circuits

Low-distortion op-amp. With a voltage noise of $0.9nV/\sqrt{Hz}$ at 1kHz and total harmonic distortion of –120dB, the *AD797* from Analog settles to 16 bit in 1.2µs. Maximum voltage offset is 60µV, drifting at 0.6µV/°C maximum. Analog Devices Ltd, 0932 232222.

15kV ESD protection. Replacing many discrete components, Harris's SP720AB/AP diode array IC uses high-speed SCR/diode structures to provide protection against 15kV of electrostatic discharge and overvoltage protection for up to 14 pins. The diodes clamp to one diode drop above the supply or a diode drop below ground, depending on polarity of the overvoltage. Harris Semiconductor (UK), 0276 686886.

Differential video amplifiers. Linear Technology has attacked the problem of obtaining a decent CMRR at high frequencies and introduced the *LT1187/1189* video amplifiers, which offer 100dB typical and 40/48dB at 10MHz. These DC-coupled devices offer a 50/35MHz bandwidth and slew at 165/220V/µs. Input offset voltage of 2/1mV, bias current of 200nA and input resistance of 100/30kΩ avoid the need for trimming in most circuitry. Settling time to 0.1% is 100ns/1µs and diff gain and phase are both very low. Linear Technology (UK) Ltd, 0276 677676.

Low-noise op-amps. *MAX410/412/414* single/dual/quad op-amps combine low wide-band noise (2.4nV/ Hz), 28MHz bandwidth and a current requirement of 2.5mA per amplifier. Operating from supplies of ±2.4V to ±5V, slew rate is 4.5V/µs and minimum open-loop gain is 115dB. Maxim Integrated Products Ltd, 0734 845255.

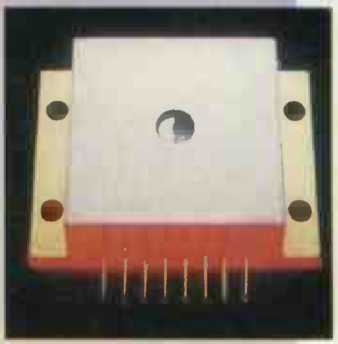
Video sync separator. *EL4581* from Elantec is a video sync. separator for NTSC and PAL systems and is a pin-compatible, but improved replacement for the *LM1881*. It extracts timing information, including composite and vertical sync, burst and back-porch timing and odd/even fiels data from standard negative-going NTSC, pal and secam video at 0.5 to 2Vp-p. Precision 50% slicing reduces the effects of noise. Microelectronics Technology, 0844 278781.

1GHz amplifier/mixer. Philips says its *NE/SA600* low-noise amplifier and mixer IC is the first single-chip 1.2GHz amplifier/mixer IC. The preamplifier has a 2dB noise figure at 900MHz and 16dB of gain, stabilised to within ±0.5dB over the –40 to 85°C temperature range. Input and output matching is carried out internally. Philips Semiconductors Ltd, 071 436 4144.

TV chipset. *TDA9160/8350* comprise a television chipset to decode pal, NTSC and secam and to drive the tube. It automatically identifies the television standard in use, extracts luminance and chrominance and generates all sync pulses for deflection, picture positioning and geometry. To decode all the pal, NTSC and secam standards, 3.6MHz and 4.4MHz crystals are needed, apart from which only a few passives and a baseband delay line are required. Either composite signals from television or recorder, or separate luminance and chrominance from an S-VHS recorder can be accepted. Philips Semiconductors, 071 436 4144.

Logic building blocks

Graphics LCD controller. The Seiko Epson *SED1335F* cmos graphic LCD controller generates all required signals and includes a character generator. It is configurable for the 6800 or 8080 processor families and is meant for use with medium-scale dot-matrix displays. Supplies must be between 2.7V and 5.5V and the unit draws 5mA when active, 0.05µA when in standby. Hawke Components Ltd, 0256 880800.



Laser diodes. High output-power density laser diodes in the *SLD320* series from Sony Semiconductor Europe emit nominal wavelengths of 790-830nm. Output powers in the series range from 500mW to 3W and optical density is high for all devices; a 1W component has a 100µm emission aperture. Operating current at 1W is 1.3A. Sony (UK) Ltd, 0784 466660.

Mixed-signal ICs.

1.5GHz synthesiser. *SP8861* by GEC Plessey is a low-power single-chip synthesiser with high input sensitivity for professional radio use. Only a loop amplifier is needed to form a complete 1.5GHz PLL synthesiser. It is programmable and has three independent buffers to store one reference divider word and two local-oscillator divider words for fast toggling. The reference source uses an external crystal. GEC Plessey Semiconductors, 0793 518510.

Datacom controller. Hitachi's *HD64570* serial communications adaptor with a built-in DMA controller provides full duplex operation at transfer rates of up to 12Mb/s. A two-channel multiprotocol serial communications interface supports a number of modes, including asynchronous, byte synchronous and bit synchronous modes such as HDLC and SDLC. Transmit and receive fifo buffers are each 32-bit

Please quote "Electronics World + Wireless World" when seeking further information

Crystal oscillators. Frequencies in the range 4-20MHz are offered by *HCD66* oven-controlled oscillators by HCD Research. They are intended for PCB mounting and measure 51 by 41 by 31mm high in a 5-pin case. AT or SC cut crystals can be fitte, characteristics with an SC type being ageing 10^{-10} per day, thermal stability $3 * 10^{-9}$ from –20°C to 70°C and phase noise down to $-160\mathrm{dBc}/\sqrt{\mathrm{Hz}}$. HCD Research Ltd, 0444 232967.

deep. Hitachi Europe Ltd, 0628 585000.

Modem chipset. A low-power chipset for a data, fax and voice modem from RCS, the *RC96V24AC*, is complete with its controller firmware. The set contains a Rockwell 9600b/s V.29 full-duplex data/fax/voice modem datapump and a C29microcontroller, together providing enhanced AT, fax class 1 and 2 and voice commands. As a data modem, the set supports V.23, V.22 bis, V.22A/B and V.21, plus Bell 103 and 212a standards. As fax, V.29, V.27ter and V.21 cahnnel 2 recommendations are supported. Voice commands use ADPCM for companding. RCS Microsystems Ltd, 081 979 2204.

Optical devices

0.5W laser diode. Sony's *SLD322XT* near-infrared laser diode emits a recommended 0.4W of optical power from a 50µm aperture in the band 790-830nm. It draws 650mA from 3V, the threshold being 150mA. Sony (UK) Ltd, 0784 466660.

Oscillators

Canny clock. At 32kHz, current drawn by Harris's *HA7210* oscillator chip is 5µA instead of the more usual 40µA. With the required crystal, the device will operate at frequencies between 10kHz and 10MHz, drawing 130µA at 10MHz. It can drive two cmos loads and has a disable mode. Harris Semiconductor (UK) 0276 686886.

Programmable logic arrays

State-machine proms. Using proms as state lookup tables in large state machines has had the disadvantage that no feedback terms have been present. Cypress now offer the *CY7C258/9* registered proms with internal state feedback of up to 2048 states and running at 83MHz. Bypassable i/o registers run from the same clock and add a pipeline feature. Ambar Components Ltd, 0844 261144.

Power semiconductors

Surge absorbers. Panasonics's *ZNR* type D transient and surge absorbers, in a wide range of voltages and currents are now obtainable from Abacus. Abacus Electronics Ltd, 0635 36222.

10A regulator. Solid State Devices has offers the *SVR117AHV* voltage regulator, which will supply 10A over a voltage range of 1.2-57V at a maximum input:output differential of 60V. It provides short-circuit and thermal protection. Britcomp Sales Ltd, 0372 377779.

70A mosfet. Harris's Megafet *RFP70N06* n-channel mosfet exhibits an on resistance of 14mΩ at 70 A and a 60V breakdown voltage. The low resistance is obtained by arranging several million power-handling cells in parallel to a density of 2.3 million cells per square inch, this method also resulting in 125ns switching speeds. Harris Semiconductor (UK), 0276 686886.

PWM controller chipset. Unitrode's *UCC3883* and *3885* provide primary and secondary PWM control for isolated switching regulators supplying light loads, such as in ISDN telecomms. *UCC3883* provides inrush-current limiting, high-impedance start and protection for the primary-side power switch, while the 3885 gives accurate secondary control by providing feedback to the switch. Macro Group, 0628 604383.

Audio power mosfets. Magnatec has a range of complementary lateral mosfet power transistors for use as high-power audio output devices. They are 8/16A, 160/200V device in single or double chip packages to give 125W for the single type or 250W for the double chip design. Magnatec, 0455 554711.

PASSIVE

Passive components

Electrolytic capacitors. *RE2* and *TE2* series electrolytics by Acal use improved aluminium foil to reduce size and increase reliability. In both radial and axial form, the components are guaranteed for 2000 hours at 85°C. and can be immersed in cleaning fluid for up to five minutes. Values available are in the range 0.1µF-22,000µF at 6.3-450V. Acal Electronics Ltd, 0344 727272.

Low-R chokes. Surface-mounted chokes in the *TDKACC* series pass a direct current of up to 3A and are wave-solderable for 10s. The range includes components with an impedance from 8Ω at 10MHz (100Ω at 100MHz) to 370Ω at 10MHz (150Ω at 100MHz), all with a resistance of 0.04Ω. Flint Distribution, 0530 510333.

Miniature electrolytics. Nichicon's *VS* series of electrolytics covers the 0.1µF-10,000µF range at between 6.3V and 400V at temperatures from –40°C to 85°C. Some of the 50V or less components are only 9mm long. Leakage is 3µA and ripple current 1.7A maximum. Nichicon (Europe) Ltd, 0276 685393.

7mm electrolytics. Nichicon's SP series of non-polarised electrolytic capacitors are only 7mm long, have a working voltage of 6.3-80V, a capacitance range of 0.1µF to 47µF, leakage current of 10µA and 75mA ripple. Operating temperature is –40 to 85°C and load life is 8000 hours. Nichicon (Europe) Ltd, 0276 685393.

Connectors and cabling

2mm connectors. In what 3M claims to be the widest range of 2mm board-mounted sockets and headers, the series 15 includes right-angle and straight sockets, low-profile, through-board-entry sockets and pin strip headers In 2-60 positions. Bodies are of glass-filled polyester or liquid-crystal polymer for higher temperatures. 3M United Kingdom plc, 0344 858000.

SCSI connectors. Connectors in Fujitsu's *FCN230R/240R* series are additions to the SCSI-II standard connector range. *FCN230R* is a family of pin contact connectors on a 1.27mm pitch that includes 50 and 68 pin versions with provision for EMI shielding and positive latch coupling, in straight and right-angle form. Round cable plug connectors can be used. *FCN240R* connectors are for flat ribbon cable and round cables. Fujitsu Microelectronics, 0628 76100.

Micro coax. connectors. Coaxial connectors in the Lynics MC series by the Japanese Emuden compan are 50% smaller than SMB connectors. They are rated up to 3GHz and are 50Ω types. All have a brass body, beryllium copper contacts and PTFC insulation. Westside Supples Ltd, 0243 542878.

Displays

LCDs. Three very slim, 200g LCD modules offer a resolution of 320 by 240 pixels, or one quarter of a VGA display, and are meant for the handheld computer and instrument market. *LMG691 0RPGR/1RPBC/2RPFC* provide blue on grey, blue on white and black on white displays, two of them being fitted with a cold-cathode fluorescent lamp for back-lighting. Hitachi Europe Ltd, 0628 585000.

LC displays. Toshiba's *TLX5171-C3M* and *TLX5171-C3B* dot-matrix LCDs are balck and white (*C3M*) and blue (*C3B*) modules with cold-cathode fluorescent backlighting. They are 320 by 240 units, 14.5mm thick and with a 121mm by 92.2mm viewing area. Toshiba Electronics (UK) Ltd 0276 694600.

Filters

Datacon filters. Filters in Matthey's OEM range, designed to satisfy CCIR requirements, are intended as anti-aliasing filters in video A-to-D and D-to-A converters. The range covers all CCIR 601 and Eureka 95 HDTV standards. Matthey Electronics, 0782 577588.

2GHz delay line. Specified for operation to 2GHz, the LDH family of delay lines are intended for optical-fibre interfaces, supercomputers and workstations. 21 models provide a choice from 0.1ns to 10ns, with tolerances of ±50ps at 0.5ns and ±0.2ns for 10ns devices. Murata Electronics (UK) Ltd, 0252 811666.

Hardware

Heatsinks. Heat "planes" by Enco are machined to register exactly with a PCB, using the same cad data as that used for the board itself, fed to CNC routers which also machine the bonding layer. Enco Industries Ltd, 05057 5151.

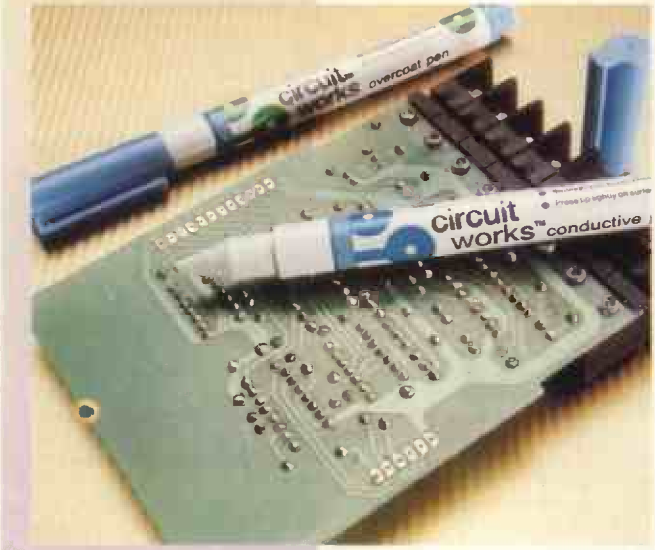
Solder mask. Loctite's Lite-Mask is a fast UV-radiation cured peelable mask to protect selected areas of PCBs in hot-air solder levelling, wave soldering and conformal coating. It cures in 20-30 seconds In ultraviolet light, has good adhesion and does not become brittle at elevated temperatures. Since it is thermosetting, it does not reflow. Loctite UK Ltd, 0707 331277.

Instrumentation

Memory tester. Taking less than two seconds to verify a 1Mbyte by 9bit simm, ABI's *RamMaster* compact offers high speed, flexibility and 100% test of all cells. Its microprocessor configures custom silicon prior to the test, so that the tester appears to be hardwired logic to the device, which can be a simm, sip, dram, sram or a PS/2 module. It identifies problem bits and gives voltage sensitivity, access time anomalies, pattern and temperature-related faults, and intermittent faults are also trapped. ABI Electronics Ltd, 0226 350145.

DSO plus. DataSys from Gould is a range of digital storage oscilloscopes with a number of extra features that turn it into a "data-acquisition and measurement" instrument. The basic oscilloscope usus a sampling rate of 100Msample/s with a repetitive equivalent time sample rate of 2.5Gsample/s. A colour LCD screen has 1000-times zoom facility, the overview and detail traces being viewed simultaneously. Its features are far too numerous to mention here, but there is computer interfacing and a hard-copy output, with a floppy disk option. Gould Electronics, 081-500 1000.

PCB repair. Circuit Works from RS Components is a pen-type dispenser to enable tracks to be drawn directly onto PCB material, including film and glass. A 1.3mm tip emits a silver-loaded polymer about 0.08mm thick, a smooth flow being enabled by a spring-loaded, non-clogging valve. The material dries in a few minutes and is heat-curable at 120-150°C to allow soldering and an increase in conductivity. RS Components Ltd, 0536 201234.



RF measurements. H-P's *HP4396A* is a 1.8GHz combined spectrum and vector network analyser with a built-in instrument controller as an option. Spectrum analysis accuracy is ±1dB, the sweep oscillator being a synthesized type. Instrument control is by means of HP IBASIC, which is a subset of HP Basic, with an external keyboard. There is a built-in floppy disk drive and a 7.5in colour screen. Hewlett-Packard Ltd, 0344 362867.

Vector signal analyser. H-P's *HP89410A* and *HP89440A* are signal analysers for work with burst, transient or vector-modulated signals. The former has one or two baseband channels of zero to 10MHz and the latter RF channels to 1.8GHz with one input. Facilities include vector spectrum analysis; frequency, amplitude and phase analysis; digital modulation analysis; and time-gated spectrum analysis. Hewlett-Packard Ltd, 0344 362867.

FFT analyser. From Hungary, the Pont PSA-100 audio spectrum analyser offers autocorrelation, cepstrum analysis and true RMS voltage measurement. Display amplitude accuracy is within 0.2dB, the dynamic range being 80dB to 25kHz. Battery-backed memory stores up to seven spectra. Printer output is provided. Manor Technology, 0794 40923.pinsapr93

Digital thermometer. A portable digital LCD thermometer from Maplin has a stainless steel probe and provides maximum/minimum alarms. Sensing speed is settable to 1s or, to conserve battery life, 10s. Indication is in either degrees Celsius or Fahrenheit. The probe is connected by a 1m wire to the unit. Maplin Electronics plc, 0702 554161.

Radio test set. In addition to a full set of programmable modulation and sweep facilities, Philips's *PM5330* Radio Test Generator has, as an option, the capability of testing RDS/ARI data systems. Up to 20 RDS messages are selectable, ten of them being programmable and down-loaded from a PC. There is an FM stereo mode with a choice of pre-emphasis time constants and the instrument also incorporates a 200MHz counter. IEEE-488 and RS 232 interfaces are optional. Philips Test & Measurement, 0923 240511.

Spectrum analysers. Advantest spectrum analysers *R3265* and *R3271*, now marketed by Rohde & Schwarz, have been provided with delayed-sweep triggering from an internal gate. Working from 100Hz to 26.5GHz between them, the instruments already possess external gated-sweep functions, but the new facility allows gate delay of between 300ns and 100ms to a resolution of 100ns, only those spectral components occurring during the gate time being displayed. Rohde & Schwarz Ltd, 0252 811377.

Multimeters. Four hand-held digital multimeters available from Saje, the 180 series, start with a basic 3.5-digit with data hold and AC, DC and resistance measurement, progresses through temperature, frequency and capacitance measurement with a bar graph, to a fully autoranging 4.5-digit instrument with all the previous functions. Saje Electronics, 0223 425440.

Function generator. Three further outputs are provided by the Thurlby Thandar 8550 in addition to the function generator: linear and logarithmic sweeps and a phase-locked generator. The instrument is microprocessor-based for accuracy and gives auto-calibration to within 1% on all functions and 0.1% continuous frequency accuracy. Sines, triangles and squares from the function generator cover the 0.01Hz-50MHz range, output voltage being from 10mV to 32V into open circuit. A GPIB interface is provided and 32 non-volatile button arrangements can be stored. Thurlby Thandar Instruments, 0480 412451.

Literature

DSP catalogue. Intelligent Instrumentation recently took over the ZP DSP cards and PC software from Burr-Brown and have now produced a catalogue of those and their own existing products in the DSP field, including DSPlay XL, a software DSP development package. Intelligent Instrumentation, 0923 896989.

Murata catalogue. Murata's 1993 catalogue is now available and includes data on filters, sensors, microwave devices, piezoelectrics, ceramic resonators and passive components. Murata Electronics (UK) Ltd, 0252 811666.

Data analysis. A brochure from National describes the analysis capabilities of *LabView* and *LabWindows* instrumentation software for the analysis of spectra, time, statistics and regression, digital filtering and numerical analysis. National Instruments UK, 0635 523545.

Power supplies

15W DC-to-DC converters. Calex single-output converters in the XC series need no external components — not even heat sinks or cooling air. They are mosfet switchers with high loop-gain current-mode control, working at 70kHz; efficiency is more than 80%. There is a built-in noise filter at input and output and RF radiation is "virtually eliminated". Transient and thermal limiting is incorporated. Calex Electronics Ltd, 0525 373178.

DC-to-DC converters. Single and dual output 5W converters from Conversion Devices, the 500UFR series, have a 4:1 input range, 75% efficiency, 500V DC i/o isolation, stabilisation of ±0.2% and regulation of ±0.5%. All models include continuous short-circuit protection with auto restart, reverse voltage protection and an input filter that is claimed to almost eliminate reflected ripple. Input ranges in the series are 9-36V or 18-72V and outputs are from 5V to 15V single and ±12V/±15V dual. Eurosource Electronics Ltd, 081 977 1105.

Radio communications products

Moderate mixer. Needing a local-oscillator input of only 1dBm, the Starved LO mixer by Synergy comes in a range of styles, including relay header, flatpack, TO and surface-mounting. Chronos Technolgy Ltd, 0989 85471.

RF switches. Alpha Industries GaAs fet MMIC switches and attenuators are surface-mounted devices in plastic SOIC packages and are meant for cellular telephone work. *AS002M2-12* is a single-pole, double-throw switch working up to 2.5GHz, with insertion loss 0.8dB, isolation 35dB and VSWR 1.3:1, all at 1GHz. Impedance is 50Ω and power consumption 50μA at -5V. Other switch formats and attenuators are available. Cirkit Distribution Ltd, 0992 444111.

Transducers and sensors

FLDT processor. Intended for PCB mounting, the SP100 signal processor board is for use with the Control Transducers family of fast linear displacement transducers. It has adjustable zero and span, with 0-20V or ±10V output. It also has a temperature compensation circuit to reduce errors caused by variations at the transducer, so that the temperature range is increased to –50 to 125°C. Accuracy is within ±0.15%, with ±0.1% as an option. Control Transducers, 0234 217704.

Industrial sensors. Honeywell's M18 ultrasonic proximity sensor allows position detection of almost any colour or material over a 130-500mm range. Noise immunity is afforded by the 215kHz carrier frequency. The company's *LL* series of high-temperature liquid-level sensors use a led, which reflects all its output internally when no liquid is present and allows light to escape when liquid covers the dome. Detection is very fast and operating temperature is –40 to 125°C. Verospeed, 0703 644555.

COMPUTER

Computer board level products

Smart PC I/o card. Possessing its own processor and memory, the *AS-1F* is a high-speed analogue and digital input/output card for PCs that acquires and processes data in real time, while the PC performs other tasks, such as providing a GUI. Eight analogue inputs acquire data at up to 170kHz using local ram with a recirculating DMA buffer. Four analogue and nine digital outputs are included, and six counter channels provide ADC sample timing and waveform generation. Data transfer is at 400kbyte/s. Pascal and C source code is supplied. Fairchild Ltd, 0703 559090.

A-to-D cards. With the Mbyte/s DMA transfer rate of EISA-bus computers, *AD* series analogue-to-digital converter cards from Adtek achieve a sampling rate of 10Msample/s at 12-bit resolution. Using the eight-channel, 2.6Msample/s *AD-830* card in a 33MHz 486 EISA-bus computer, continuous data throughput onto disk is shown to be 1Mbyte/s, 92% of the cpu processing power remaining available. Laplace Instruments Ltd, 0692 500777.

Computer systems

Rack-mounted PCs. A range of 19-in rack-mounted PCs from Sight Systems is available with a range of processors, memory and disk options, all in rugged units for industrial environments. Processors go from a 20MHz 286 to a 33MHz 486, up to Mb of ram, one or two disk drives, a choice of monitor and up to 100Mb hard disk. Prices start at £1,118. Sight Systems Ltd, 0273 439959.

Development and evaluation

8031 ICE. *NICE-31* is a series of in-circuit emulators for the *8031/51/552* family of microprocessors. The PC program supplied shows memory and register contents and lets the user set up memory mapping and define breakpoints. This latest version allows both assembler and C source debugging for IAR, Keil and Avocet C compilers. As an extra, a 16K by 48bit trace buffer can be used to set up a selection of pre- and post-trigger conditions. Computer Solutions Ltd, 0932 352744.

80C186 ICE. Great Western has the *CheckMate-C186*, a pocket-sized in-circuit emulator to integrate codeto an Intel *80C186/8xx*-based target at up to 20MHz. Features are PC AT or higher as host to a Paradigm Debug source-level debugger with high-

speed communications, overlay ram, hardware breakpoint and event system and trace memory. It will download a 256Kbyte .EXE file in less than six seconds. Great Western Instruments Ltd, 0272 860400.

Video for PCs. VideoBox from Iterated Systems is a low-cost fractal video system which decompresses and displays full-screen, full-motion video on a PC, no expansion cards being needed. It is to be available later this year as a software developers' toolkit and then in video databases and multimedia products. It decopresses and replays video at up to 30 frames per second in software alone. Iterated Systems Ltd, 0734 880261.

Low-cost ICE. The Raisonance *TINY-ICE* is an in-circuit emulator for *8031* and *80C31* microprocessors, costing only £300. It plugs into a PC and runs at 12MHz. An on-board monitor allows application development using up to 32K code space and 64K external data memory. Single-step and continuous emulation are supported and up to 800 breakpoints can be set in the code space in continuous emulation. Logicom Communications Ltd, 081 756 1284.

HP 68000 emulator support. A set of development tools from Microtec are for use with the latest version of H-P's in-circuit emulator for Motorola's 68000 family. Microtec's *XRAY/ICE* combination now realises the potential of the integrated debugger/emulator interface in the areas of C support and source-level access to real-time trace data. Microtec Research Ltd, 0256 57551.

Neural net starter kit. *NT6000*-series cards from Neural Technologies are entry-level network development cards for the PC. They are "plug-in-and-go" expansion cards with high-speed DSO, memory, digital/analogue i/o and software. Menu-driven software and a graphical display, together with an introductory book on neural networks enables a system to be established in a few hours. Neural Technologies Ltd, 0730 260256.

LabWindows C/C++/VBDOS libraries. National's *LabWindows* for DOS, version 2.2.1 instrumentaion software includes stand-alone libraries for Borland's *C++* and *Turbo C++* compilers and Microsoft's Visual Basic for dos compiler. *LabWindows 2.2.1* makes over 260 instrument drivers available to users of VBDOS. National Instruments UK, 0635 523545.

Software

Neural nets for Windows. *NeuDesk2* is a neural network development package from CRaG Systems, spreadsheet-driven and running under *Windows 3.0* or higher.The software automatically selects the network topology and training method and, once developed, the network can be run from within NeuDesk or from another application using the NeuRun runtime module, available separately. Input data is either manually entered, obtained from a number of different file formats or cut and pasted from a spreadsheet or database. CRaG Systems, 0635 873670.

PC data acquisition. Master Link PCI-20369S-1 software libraries for dos and *Windows* allow the development of gap-free, multi-megabyte data acquisition and analogue sampling at up to 1MHz. Master Link software aupports C, C++, QuickBasic, Visual Basic and Turbo Pascal. Intelligent Instrumentation, 0923 896989.

PCB cad. Tsien's *BoardMaker 2.5* cad package for printed-circuit board development generates thermal breakpoints for pads within power planes, and design-rule checks now ensure that nodes needing access to power planes have the necessary via for connection. It is also possible to examine the clearances of tracks and pads passing through a thermal plane. Speed gains in the new version include a three times improvement in the top-down modification process and a 12-fold improvement in netlist entry. Tsien (UK) Ltd, 0223 277777. ■

CIRCLE NO. 127 ON REPLY CARD

APPLICATIONS

Linear circuit active filters

Disadvantages of discrete-component active filters are well known: component tolerance, drift and some sensitivity to layout. For a well defined response shape at higher orders, these drawbacks can be so severe that a continuous filter is not feasible and a switched type is used instead.

But even switched-capacitor filters suffer from a number of limitations, in that Nyquist bandwidth is curtailed, switching noise can amount to several millivolts at the switching frequency, there is the aliasing problem and higher distortion.

For applications in which switched-capacitor filters are not to be used, Maxim has the *MAX274/275* integrated linear filter building blocks, which contain four op-amps and some very accurate, low-drift capacitors. For a second-order section, the IC and four external resistors form the low-pass or band-pass filter, and Butterworth, Bessel or Tchebysheff all-pass filters can be made. Maxim's 1993 Applications publication describes the use of the devices. It also presents some plots of the results of individual F_0 and Q errors of apparently

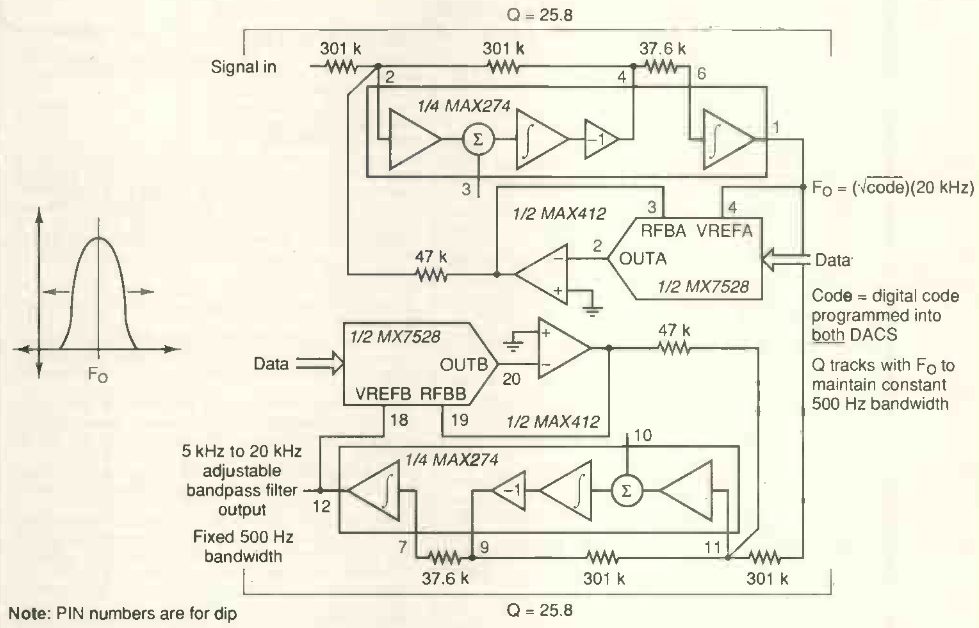
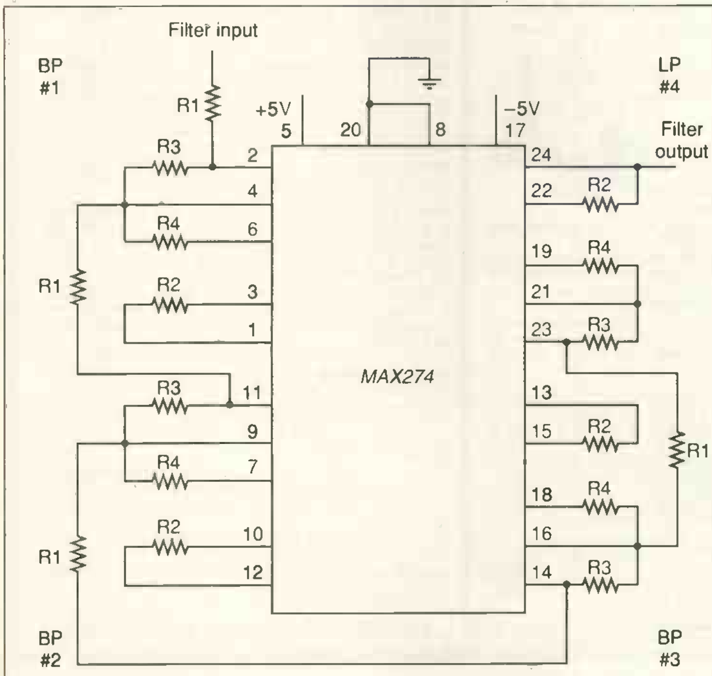


Fig.2. Digitally tunable band-pass filter with a centre frequency from 5kHz to 20kHz, depending on the digital input to the two D-to-A converters.



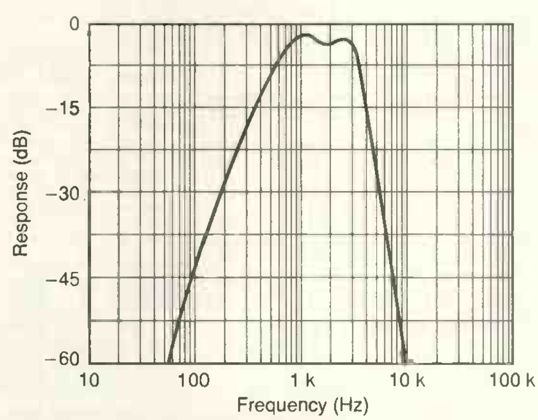


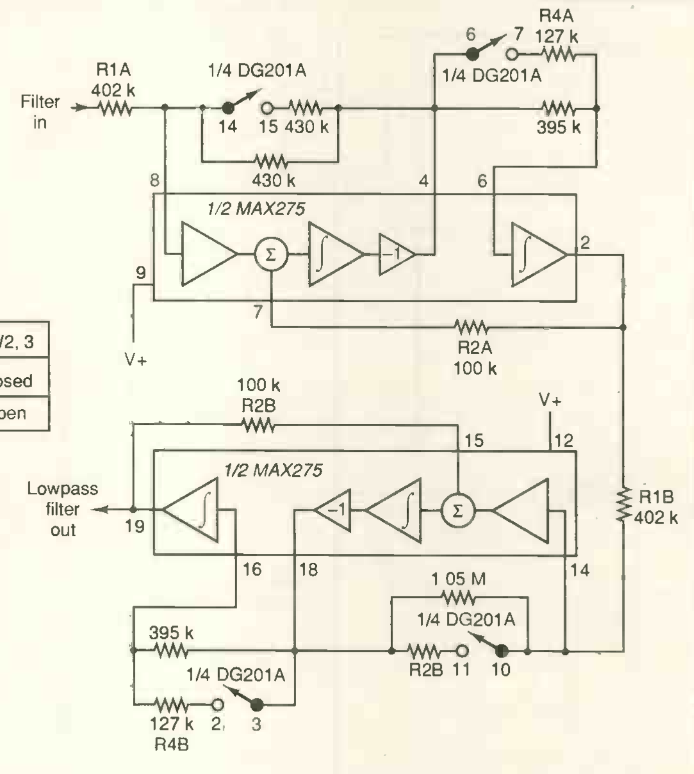
Fig.1. Maxim's MAX274/5 integrated linear filter sections offer an alternative to switched-capacitor filters at higher orders, low noise being one of the advantages of continuous filters. This C-message filter generates only 2.5µV of in-band noise and provides a 117dB dynamic range.

minor proportions in multiple-order filters; errors of ±5% in the four sections of an eighth-order Tchebysheff filter produce excessive pass-band peaking.

Figure 1 shows an application that exploits the -89dB sinad ratio of the *MAX274/5* – a C-message filter, which simulates the response of the human ear and which is used in telecomms for audio noise measurement. One *MAX274* does the job by cascading three second-order band-pass sections and a second-order low-pass section to produce the response shown. Operating from a 5V rail, the filter generates 2.5μVrms in-band noise, output swings of 5Vpk-pk providing 117dB of dynamic range.

Using a *MAX7528* dual D-to-A converter and a *MAX275*, the fourth-order band-pass filter in **Fig.** 2 is digitally tunable from 5kHz to 20kHz with a constant 500Hz bandwidth and constant gain. Centre frequency is proportional to the parallel code fed to the D-to-As and is $F_c=\sqrt{\text{code}}(20)$kHz, in which code is 16/256 to 256/256 – LSBs are not in use. For example, code FF(H) gives a centre frequency of 20kHz with Q_s of 25.8 to give a cascaded Q of 40.

On a similar theme, but rather simpler, is the circuit of **Fig. 3**, which uses a pair of switches to tune a fourth-order Butterworth low-pass anti-aliasing filter to cope with, for example, two conversion rates in a D-to-A converter. Resistor pairs R_{3a}, R_{4a}, R_{3b} and R_{4b} control the pole frequencies and Q of the two sections and are switched by a four-pole analogue switch to give 10kHz and 20kHz cut-off frequencies.



F_{3dB}	SW1, 4	SW2, 3
10 kHz	Closed	Closed
20 kHz	Open	Open

Fig.3. Low-pass filter with two settable frequencies, obtained by switching shunt R in and out by analogue switches. This one can be set to 10kHz and 20kHz, to filter the output of a D-to-A converting at one of the two frequencies

The application note goes on to describe two methods of DC offset removal, as might well be needed at the output of a number of cascaded amplifiers in a high-order filter.

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

Instrumentation amps are not always the best choice

Designers automatically choose the classic instrumentation amplifier to process sensor inputs, perhaps without considering that there might be a simpler and better way of going about it. Warren Schultz of Motorola discusses the subject in application note *AN1325*, beginning with the classic design and showing that there is indeed a simpler circuit that does a better job, in particular when the sensor is a pressure transducer.

An interface amplifier used for pressure sensors must provide gain from 100 to 250 or thereabouts, and convert the differential input to a single-ended input to the succeeding A-to-D converter if a microprocessor is involved. The half-supply common-mode voltage must also be translated to a DC of around 0.5V at zero pressure, so that the output swing is 0.5V-4.5V to lie within an A-to-D's 5V range.

Figure 1 is the classic design of instrumentation amplifier, in which gain, level shifting and differential-to-single conversion are taken care of, but single-supply operation is not. Modifications in **Fig.** 2 provide this, $U1_D$ providing a buffered offset voltage via R_3 to set up the voltage on the pot. wiper at the amplifier output for zero differential input. Choosing R_{10} at 240Ω gives a gain of 125, so that a 32mV input produces 4V at the output or, with offset at 0.75V, 0.75V-4.75V to work directly to a microprocessor A-to-D converter input. This is all very well, but resistor matching might be a problem and there is also the fact that there are two amplifiers in one feedback loop, which could lead to instability. In addition, the minimum output voltage of $U1_D$ forces the zero-pressure offset to 0.75V instead of a more normal 0.5V.

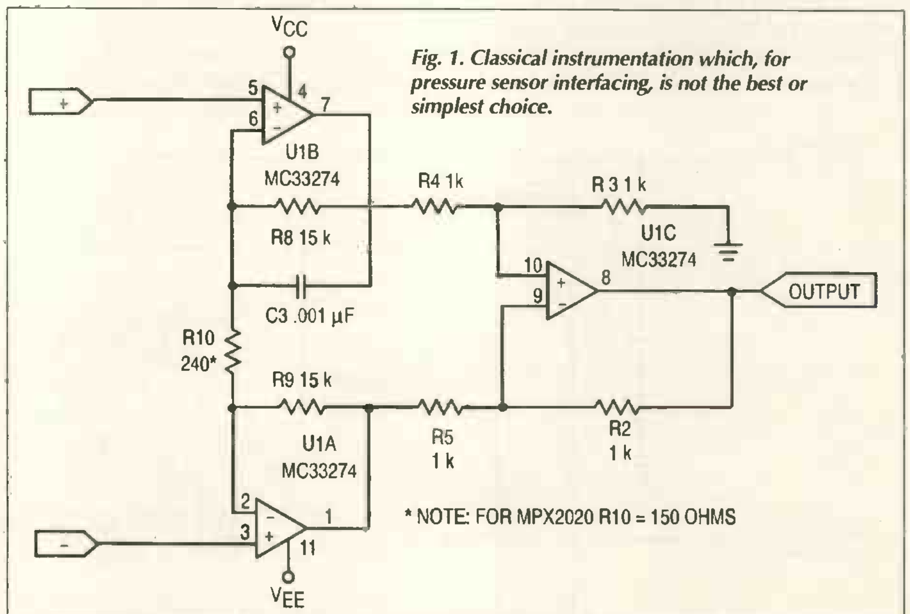


Fig. 1. Classical instrumentation which, for pressure sensor interfacing, is not the best or simplest choice.

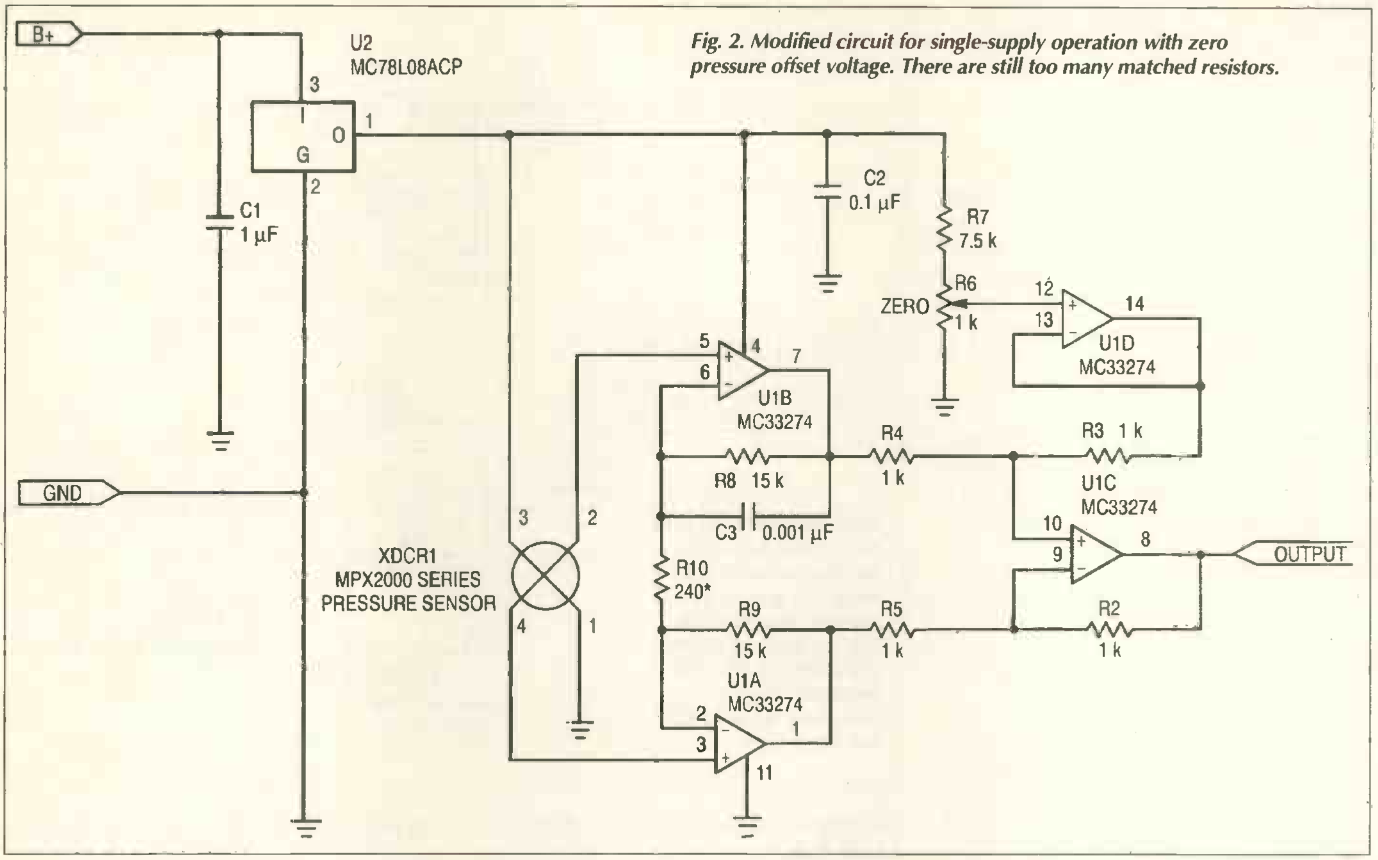


Fig. 2. Modified circuit for single-supply operation with zero pressure offset voltage. There are still too many matched resistors.

Figure 3 is a further step away from the traditional towards a simpler, but better interface amplifier, using one quad op-amp. Op-amp $U1_A$ in differential form is the gain element, $U1_B$ preventing feedback current through $R_{5,6}$ going into the sensor. Zero-pressure output at pin 1 of $U1_A$ of 4V is translated to the required zero-pressure voltage output by $U1_{C,D}$, zero being adjusted by R_9. Gain is $R_6/R_5(R_1/R_2+1)$, which gives 125, so that with an input of 32mV and the offset at 0.5V, the output swing is 0.5V-4.5V. This is a better design in that it is more stable, uses fewer resistors and will give a zero offset of 0.5V, but there is still the resistor-matching problem.

To realise an interface amplifier that is considerably simpler, smaller and cheaper, but that performs the same function as the classic design in a rather more elegant manner, the circuit shown in **Fig. 4** is the optimum; it uses one dual op-amp and a few resistors. In this case, the zero-pressure output voltage is exactly equal to the output voltage of the divider $R_{3,5}$ (the application note explains why!) and is independent of the sensor's common-mode voltage if $R_1/R_2 = R_6/R_4$, the value for R_6 including the resistance of the divider. Gain is again 125 and adjustment of the divider values to produce 0.5V again gives an output of 0.5V-4.5V.

It is pointed out that choosing 1% resistors

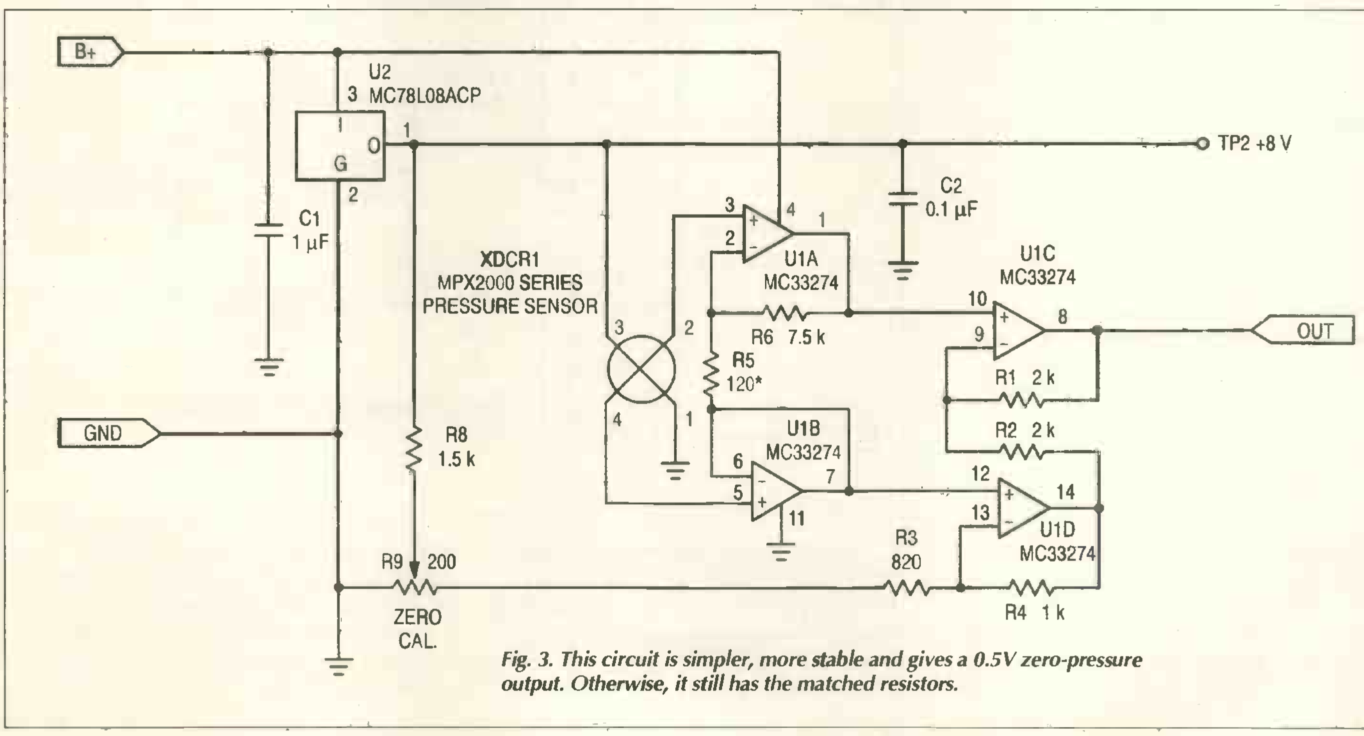


Fig. 3. This circuit is simpler, more stable and gives a 0.5V zero-pressure output. Otherwise, it still has the matched resistors.

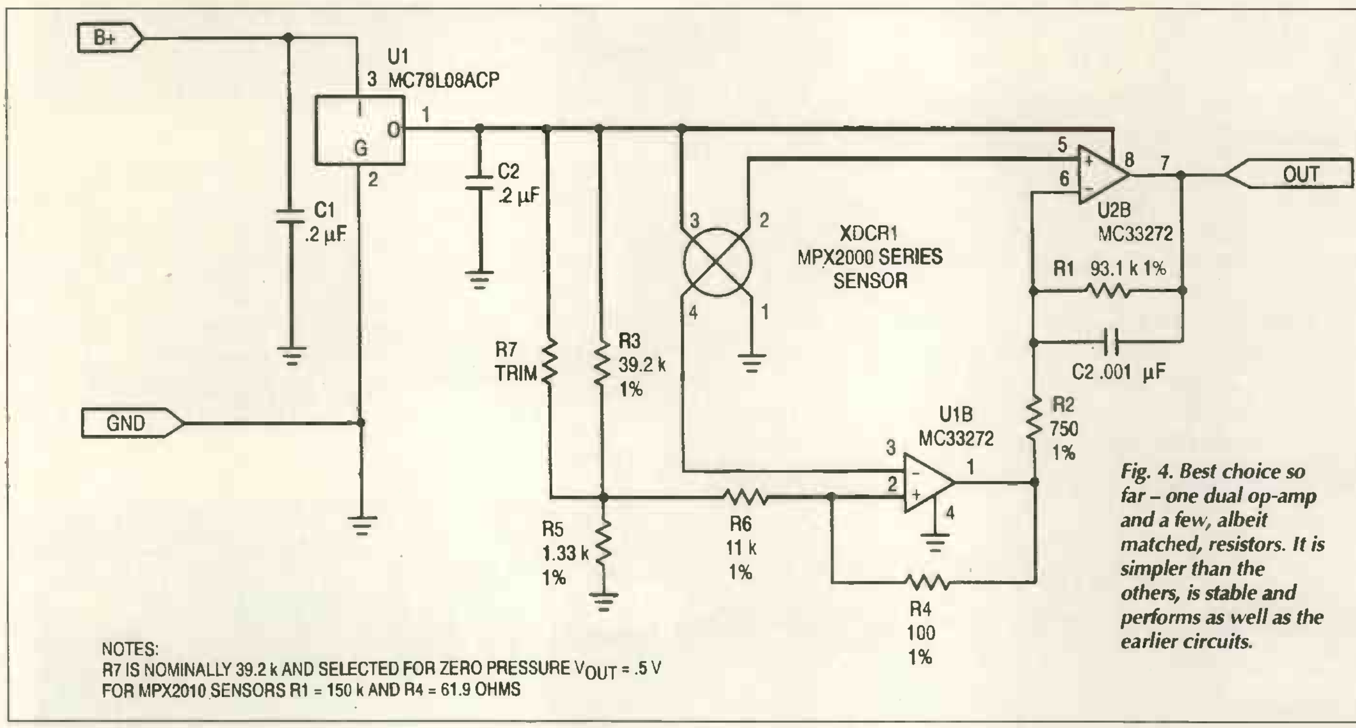


Fig. 4. Best choice so far – one dual op-amp and a few, albeit matched, resistors. It is simpler than the others, is stable and performs as well as the earlier circuits.

in place of the 5% type would probably mask any differences between the three circuits, all of them being capable of a pressure-to-voltage linearity to within ±5% from zero to 50°C,using theusing the Motorola *MPX2000* series sensors and an *MC33274* amplifier.

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

Two op-amps better than one for DC and wide-band

Classical op-amps and current- feedback op-amps each have their pros and cons, so it makes sense to let them do their own thing and combine the results. OR as Burr-Brown says in its application bulletin AB-007A, obtain the best of both worlds.

A classical op-amp performs well when you need a fairly low gain bandwidth compared with the op-amp's gain bandwidth product, but increasing the closed-loop gain lowers the amount of loop gain left for error reduction; it rolls off at 20dB/frequency decade anyway. At higher frequencies and greater required gains, therefore, errors accumulate.

On the other hand, current-feedback op-amps are happy at both low and high gain, since feedback sets both closed-loop and open-loop gain so that loop gain and dynamic performance are more or less unaffected by the closed-loop gain demanded. The trouble is that input voltage offset, offset drift and common-mode rejection are not a patch on those found in classical op-amps.

Combining a classical type such as the *OPA627* and a current-feedback *OPA603*, as shown in **Fig. 1**, gives a performance that improves on the individual characteristics of the two op-amps on their own.

Since the *OPA627* does not drive the load, its inherently good act at DC is preserved from the effects of thermal feedback when large loads are to be driven. Loads of 150Ω can be driven to ±10V with trouble from thermal feedback. As the *OPA603* adds gain to the output of the *OPA627*, the latter's slew rate goes up by he increased gain. As an example, in the amplifier shown, using a gain of 100, slew rate and full-power response of the *OPA627* increase from 40V/µs,600kHz to over 700V/µs,11MHz.

Settling time in the classical op-amp is retained at $T_s = n/2\pi b$, where b is amplifier unity-gain bandwidth and n is the number of time constants needed to settle to the required accuracy. Bandwidth of a classical op-amp increases with decreasing loop gain. Since, in the composite amplifier, the current-feedback device contributes a gain of 52, the *OPA627* is left to provide a gain of only two, so that its settling time is reduced to 330ns from the 6.9µs it would occupy were it to supply the gain of 100 alone. The only real point to watch is the bandwidth of the device chosen for A_2. If it is too small, phase shift could cause instability.

A dual op-amp is usable in this circuit if the boost in slew rate is unnecessary, but bandwidth and settling-time improvements are needed. An *OPA2107* dual type was chosen because the dynamic characteristics are well matched and the circuit's stability and transient response are best served by setting the gain of A_1 to twice that of A_2 and R_4 10kΩ. ■

$$R_3 = \frac{R_4}{\sqrt{R_2/(2R_1) - 1}}$$

Table 1. Settling time of classical, single op-amp, composite amplifier and cascaded type.

CONFIGURATION	SETTLING TIME TO 0.01%
Single Amplifier	20µs
Composite Amplifier	4.6µs
Cascaded Amplifier	4.1[1]µs

Note: (1) For cascaded amplifer stages, the combined settling time is the square root of the sum of the squares of the individual settling times.

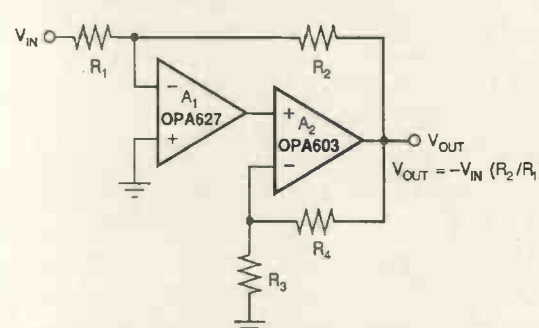


Fig. 1. Composite amplifier is better than the sum of its parts. DC and high-frequency performace are both enhanced and relatively unaffected by gain setting.

Burr-Brown International Ltd, 1 Millfield House, Woodshots Meadow, Watford, Hertfordshire WD1 8YX. Telephone 0923 33837.

INSTRUMENTS TO BUY

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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).

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MX1010F MX1100F

FG 506 FG513

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THE 180 SERIES

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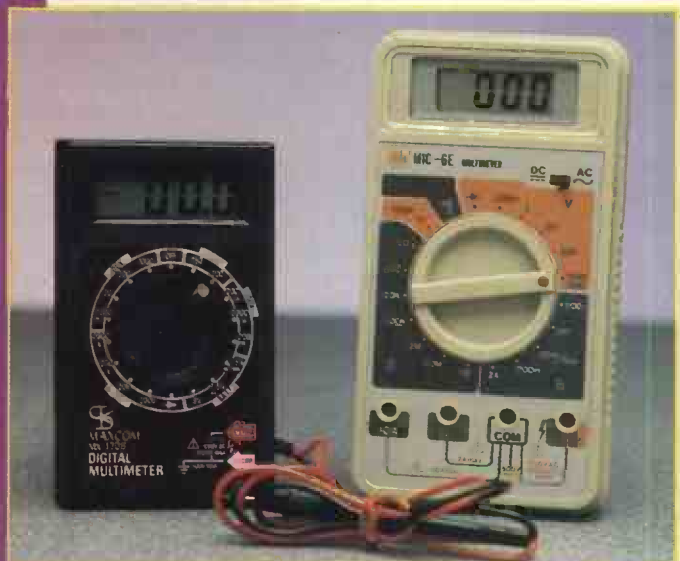
MIC-6E: $3^1/_2$ digit LCD, ACV, DCV, ACA, DCA, resistance, diode test, buzzer.
£33.50 plus VAT (£39.36).



MX170B MIC-6E

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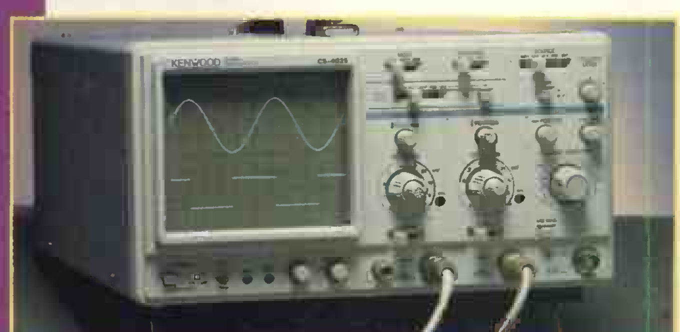
The CS4025 20MHz dual trace oscilloscope offers a comprehensive range of facilities including a high sensitivity vertical amplifier providing from 1mV to 5V/div in CH1, ALT, CHOP, ADD, CH2 modes with inverse polarity on CH2. The horizontal timebase offers a sweep range of 0.5s/div to 0.5µs/div plus x10 sweep expansion and X-Y mode. Triggering can be auto or normal from vert, CH1, CH2, line or external sources with coupling provided for AC, TV-F and TV-L. The CS4025 is supplied complete with matching probes for £295.00 plus VAT (£346.62).



20MHz 2-CH OSCILLOSCOPE

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Germany's imperial wireless system

The Marconi company is normally credited with the technology for long distance communications. But the German Imperial Wireless system was at least as impressive in its complexity and effectiveness. By George Pickworth.

Before the era of short wave beam systems, beginning around 1925, transoceanic radio communication was possible only with very long waves. Among the pioneers who developed their own particular techniques to establish successful commercial transoceanic radio links were Fessenden, Goldschmidt, Marconi, and Poulsen.

By the outbreak of the First World War, Germany was the only nation to have an imperial wireless system. Marconi's Ireland to Canada 1906 link had prompted the Italian government to commision the Marconi company to establish a very-long-wave link between Italy and its colonies in Africa, but Britain firmly rejected the idea of an empire wireless system until the Great War.

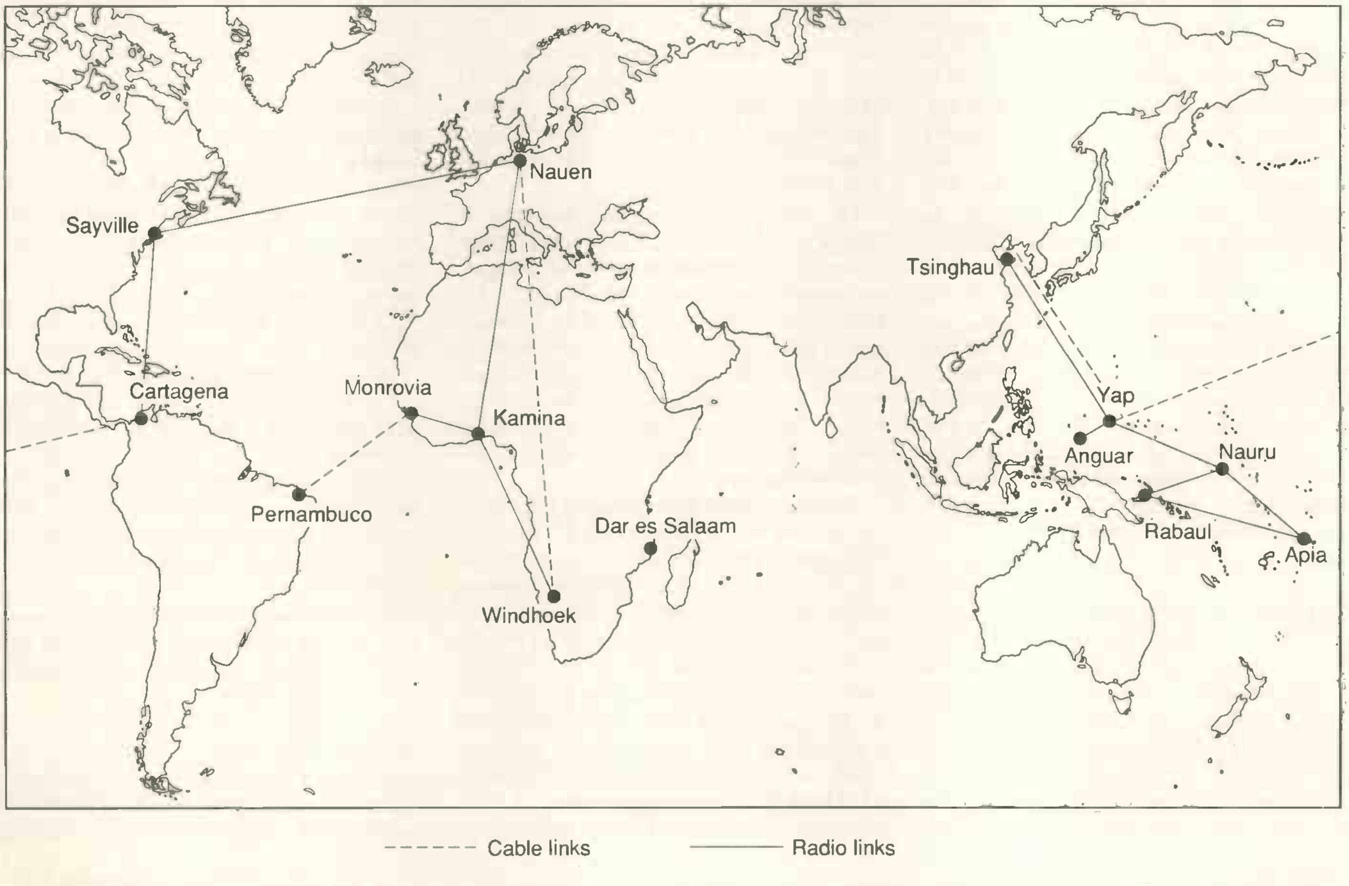
In 1906, when the German imperial system was conceived, radio communication was limited to a range of a few thousand kilometres. Greater range would have required relay stations. Germany had no territories in which to install such stations between its Pacific and Africa colonies, or German controlled stations in America.

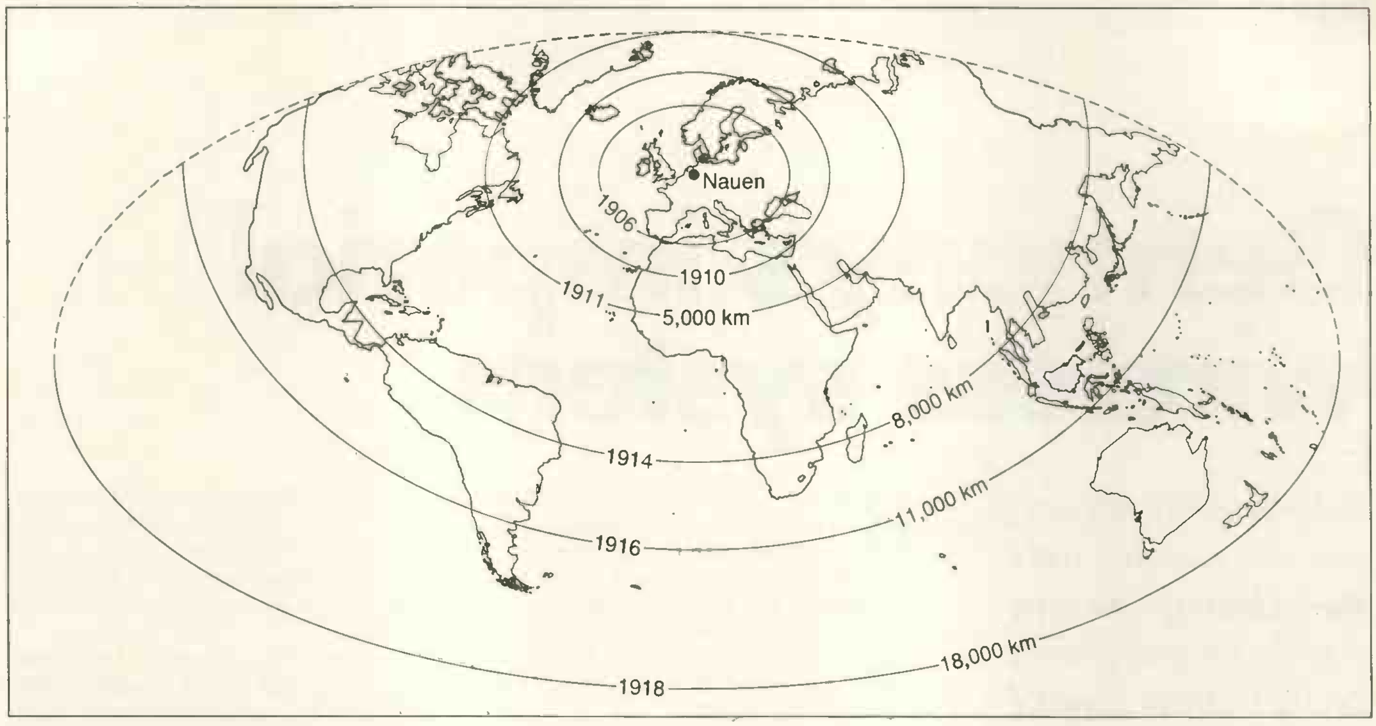
As an imperial wireless system independant of submarine cables was not possible in 1906, so, the most cost effective approach was to estblish strategic medium range stations throughout German Africa and Pacific colonies and link these to Berlin via a submarine cable complex.

A German controlled station was established at Sayville in Long Island, USA, primarily for communication with ships in the North Atlantic, and a second station was established at Cartagena, Columbia, connected to the Pacific and Atlantic submarine cable complex.

Radio frequency alternators were adopted

German Imperial Wireless System of 1914





Circle showing the rapid increase in range of the Nauen transmitter during 1906 to 1918.

for the central transmitter at Nauen and principal overseas stations whilst Wien quenched-spark transmitters were used for feeder stations. These were basically the same as those used by Telefunken maritime relay stations.

Vulnerable cables

Imminence of war made the vulnerability of submarine cables a matter for concern for Germany. So the range of the central transmitter at Nauen, near Berlin, was increased to make the system less cable-dependant for outgoing messages. Meanwhile, the range of the Sayville transmitter was increased to allow direct communication with Berlin, and a powerful transmitter was constructed at Kamina, Togoland, ultimately linking the German Africa colonies with Berlin. At the same time the Pacific stations were developed to the stage whereby radio communication was possible between all the German colonies and the German garison at Tsinghau, China – though cables remained vital to link the area with Berlin. Countries and islands of the British empire, together with mandated territories, were close enough for the whole of the empire to have been linked by radio. But the cost would have been high, and is probably why it was rejected. So, except for the North Atlantic links operated by Marconi and Fessenden, communication with the British empire was by submarine cable.

8000km range by 1914

At the outbreak of war, the range of the 24kHz Nauen transmitter had been increased to 8000km, enabling signals to be received reliably at Sayville and Kamina. Cartagena in Columbia, Windhoek, South West Africa, and Tsinghau were still in fringe areas. But power of the Sayville station was increased to maintain radio communication with Berlin and relay signals extended southwards to Cartagena from where smaller medium/long wave relay stations, were augmented to South America.

In 1914, the Kamina station also maintained direct communication with Berlin and relayed messages to and from Windhoek. Under good conditions direct communication was possible between Windhoek and Nauen – the only radio link extending to Southern Africa at the time. Relay stations connected Kamina with Douala, Camerouns, Dar es Salaam, Tanganyika, and Monrovia, Liberia, from where a German submarine cable connected with South Anerica.

The principal German radio station in the Pacific was at Yap in the Caroline Islands, connected by submarine cable to Tsinghau and America and ultimately to Germany. The Pacific radio network extended northwest from Yap to Tsingtau and eastwards to Nauru Island and finally Apia in the Samoa Islands. Branches extended to Anguar in the Palau Islands, and Rabaul, New Britain, and feeder stations linked the numerous islands with the principal stations.

Tsinghau garison could possibly have received messages from Nauen and relayed these to German Pacific colonies, but direct radio communication over the 8000km separting Nauen and Tsinghau does not seem to have been achieved (See map, first page of article).

Overseas signals were received at Geltow, near Berlin, which like Nauen, was connected to Berlin by land lines. Interestingly, the Lieben-Reisz valve, employed as a self excited RF oscillator, was used by Dr Meissner in 1913 to established an experimental radio telephony link between Berlin and Nauen.

At war

Britain's strategy to compensate for not having an empire radio link was to cut all German submarine cables immediately the First World War was declared. Allies then systematically destroyed the German colonial stations. Within the first year all had been captured. Kamina was of particular importance as it could relay messages to German ships in the South Atlantic and German agents in South America: it was blown up by the Germans before it could be captured. Shortly afterwards, Japanese troops captured the Tsingtau garison.

Nonetheless, radio communication was maintained between Berlin, Sayville in Long Island and Cartagena until 1916 when the US joined the allies in the war. From that year Nauen was Germany's only contact with the outside world. Britain belatedly established a chain of medium range relay stations extending to North America, the North and South Atlantic ocean, and, via the Indian Ocean, to Singapore without resorting to submarine cables. The system was completed in 1916.

Transmitters were typically 30kW and the relay employed a mixture of Marconi-synchronised spark, Poulsen quenched-arc and thermionic valve sets. Reasoning was that under war-time conditions these were more readily available and easier to install than the one-off high-power transmitters used on north Atlantic service.

With Germany isolated, Nauen assumed a new and important role in broadcasting messages to German ships and agents – particularly in South America.

A massive new antenna was constructed which increased range to 12,000km. By 1918, largely as the result of the development of active receivers, signals could be received virtually world-wide without large elaborate

SHAPING THE CARRIER

Wave trains radiated by early spark sets decayed too quickly for receiver resonance to be effective. Moreover, the first wave could shock the tuner into oscillation at any frequency to which it happened to be tuned.

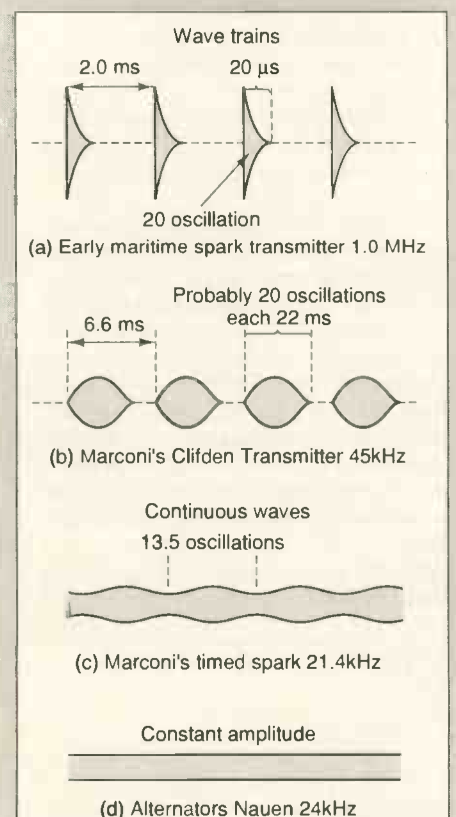
Trains radiated by Marconi's Clifden transmitter increased, and decreased in amplitude gently thus giving some degree of tuner resonance whilst at the same time modulating the transmission

With Marconi's timed spark transmitter, oscillations persisting in the antenna circuit were reinforced every 13.5 cycles to produce slightly undulating continuous waves

Radio frequency alternator type transmitters, of course, radiated continuous waves of constant amplitude.

Continuous waves, by virtue of resonance, progressively built up the amplitude of currents in the receiver tuner, so sensitivity was far greater than with wave trains. For example Fessenden's 1906 Brant Rock/Machrihanish link which employed alternators had less input power but longer range than Marconi's Clifden/Glace Bay link.

Moreover signalling speeds with CW could be much greater than with wave trains: there had to be sufficient trains present to make Morse characters identifiable.



(a) Early maritime spark transmitter 1.0 MHz

(b) Marconi's Clifden Transmitter 45kHz

(c) Marconi's timed spark 21.4kHz

(d) Alternators Nauen 24kHz

LONG WAVE SUPER STATIONS

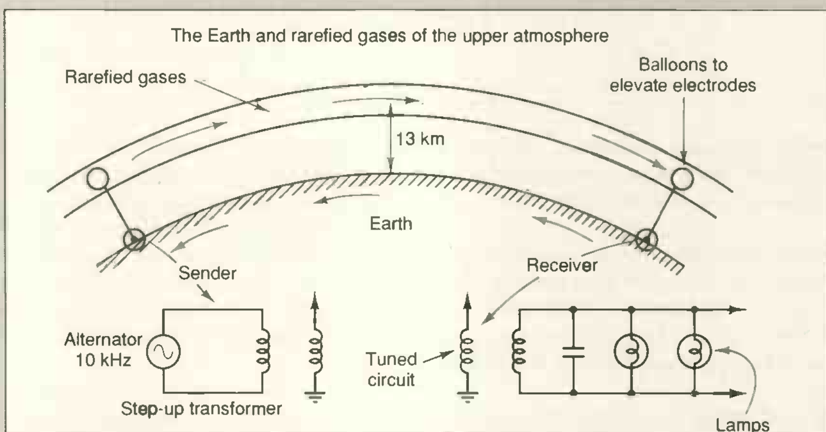
Marconi increased the wavelength of his experimental transatlantic transmissions from a few hundred metres to a thousand metres, and obtained a proportional increase in range.

Signal strength and range was much greater over a night-time transmision path than a daylight path – the night effect – but this was a phenomenon that became progressively less pronounced as wavelength increased. With waves longer than about 6.0km long (50kHz), transatlantic communication was possible for virtually 24 hours a day.

In 1902, Sir J J Thomson attemped to explain Marconi's first transatlantic signals by resuscitating Tesla's concept of a conductive layer in the upper atmosphere. Tesla was granted a patent for transmission of electric power via rarefied gases of the stratosphere in 1890, with the earth serving as the return path. (The drawing below is an illustration from Tesla's 1890 patent.) Heaviside and Kennelly also accepted that there was a conductive layer above the earth, but believed this to be caused by solar radiation – ultimately found to be correct.

Before 1925 there was no satisfactory explanation for night-effect. Now we know it is caused by the ionosphere's D-region absorbing medium length waves. It disappears at night to allow these waves to reach higher regions where ionisation persists after darkness and from where reflection occurs.

Maritime wavelengths, 300 and 600m (500kHz to 1.0MHz) were selected primarily because they allowed a quarter wave antenna to be suspended between the ship's masts. By the same token, German relay stations operated on shorter wavelengths because smaller antennas could be used, but range was influenced by the D-region.



antennas. Indeed, German agents often used frame antennas about 2.0m square.

System technology

Transoceanic communication with early passive receivers was possible because waves longer than about 6000m (50kHz) propagated within the cavity between the earth and ionosphere – the earth/ionisphere waveguide – with very little attenuation.

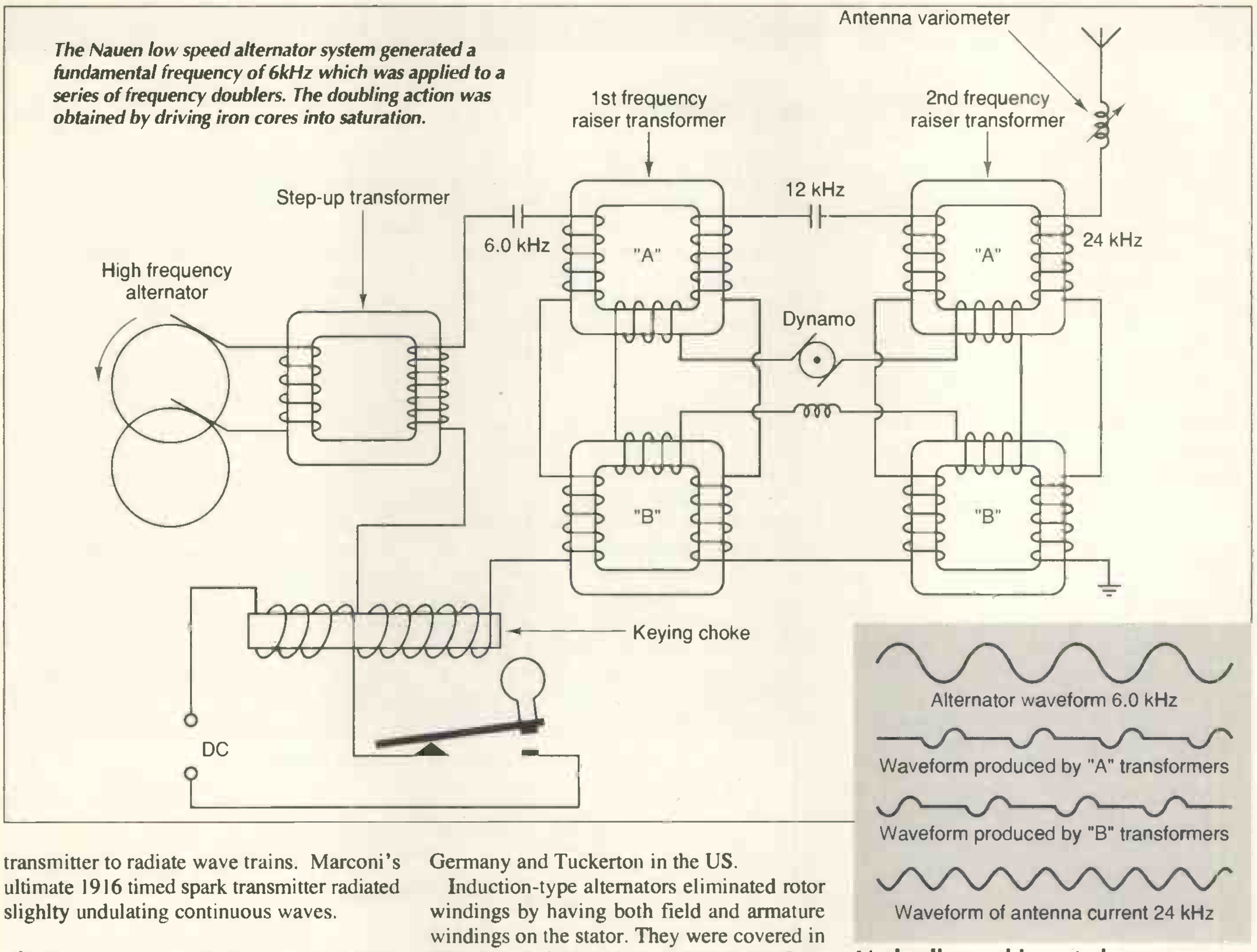
The effect is rather like sound pressure waves in a voice pipe, and the discovery brought about the era of the transoceanic very long wave 'Super Stations' (see box).

Frequency had to be less than about 50kHz to take advantage of the earth/ionosphere waveguide. Antenna size limited the lowest frequency to about 20kHz, so all transoceanic stations, were confined to a band about 30kHz wide. To avoid mutual interference, a high degree of selectivity was essential and this was only attainable with continuous waves (CW). By virtue of their resonance, CWs progressively built up the amplitude of the oscillatory current in the receiver tuner. Wave trains did not persist long enough for tuning by resonance to be effective.

Before the evolution of high power transmitter valves, there were three ways to generate high power CW: Poulsen's quenched arc, Marconi's timed spark and radio frequency alternators. The technology was not confined to radiation of CW but also extended to detectors. Most of those used for reception of wave trains were unsuitable for reception of CW.

Marconi's magnetic detector. Marconi's detector responded only to wave trains, but was rugged, reliable and – though less sensitive than some other detectors – not damaged by high voltages generated by the complementary, or nearby transmitter. So, it is understandable that Marconi continued to use the magnetic detector with his 1906 Clifden (Ireland/Glace Bay link, though this meant modifying bascially a CW quenched arc type

The Nauen low speed alternator system generated a fundamental frequency of 6kHz which was applied to a series of frequency doublers. The doubling action was obtained by driving iron cores into saturation.



transmitter to radiate wave trains. Marconi's ultimate 1916 timed spark transmitter radiated slighlty undulating continuous waves.

Alternator systems. Marconi concentrated on producing CW with spark systems. But other pioneers, including Alexanderson, Goldschimdt and AEG, directed their attention to overcoming problems inherent in RF alternators.

The most serious was that centrifugal force caused rotor windings to fly out of their slots. To attain a frequency of 100kHz, early Alexanderson alternators had a rotor with 600 rotor pole pieces and a drive speed of 20,000rev/min. Unfortunately speeds of this order were unsustainable and the machines normally operated below 10,000rev/min. An improved design, running at a lower speed, was employed by Fessenden for his Brant Rock (USA) to Machrihanish (Scotland) link. – see box, Long wave super stations.

Goldschimdt employed a fairly low speed alternator and increased output frequency by resonance circuits across the rotor and stator. This was the method used by Goldschimdt with his 48kHz link between Eilvese in Germany and Tuckerton in the US.

Induction-type alternators eliminated rotor windings by having both field and armature windings on the stator. They were covered in French and German patents by M M Cail-Herner and Guy. Guy's method was eventually patented by AEG Berlin and employed at Nauen and principal German overseas stations.

A further innovation was to employ relatively low frequency alternators and quadruple the output frequency by frequency raisers – and this was the approach adopted at Nauen. Frequency raisers were based on the non-linear saturation characteristics of the cores of a series of toroidal transformers, with a DC generator supplying the saturation current. At Nauen, the 6.0kHz ouput frequency of the 200kW alternators was increased to 24kHz. Keying was by saturation of the core of an inductor.

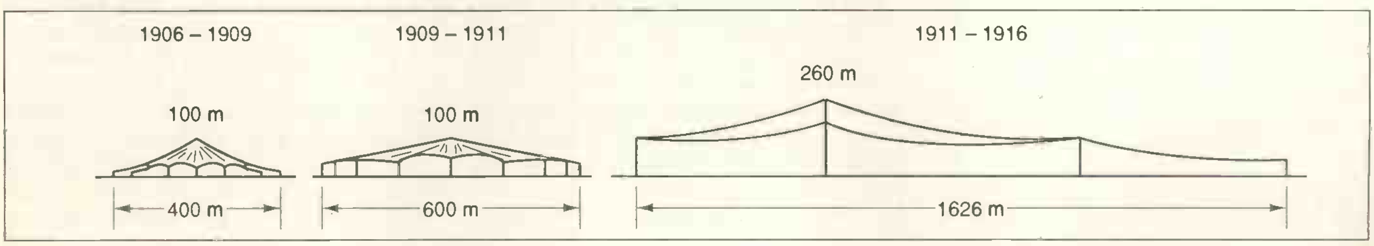
Evolution of the Nauen antennas. Each increase in size improved the range possible with simple passive receivers. The advent of thermionic valves removed the need for these impassive systems.

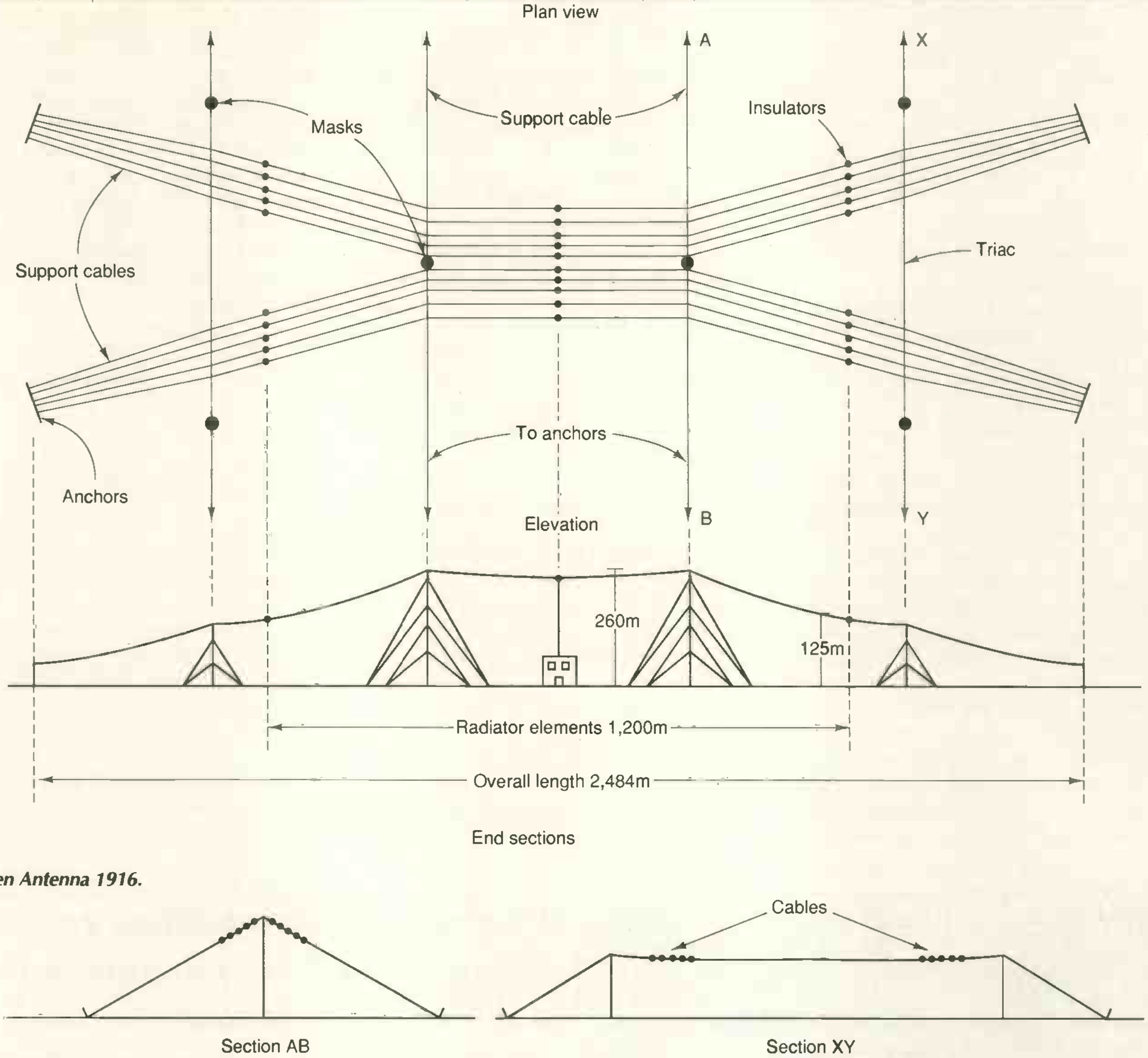
Umbrellas and inverted cones

Generating high power at frequencies less than 50kHz was relatively easy. But it was exceedingly difficult to radiate this energy efficiently – even the largest practical antenna structures were only a fraction of a wavelength long, so resonance by virtue of standing waves was out of the question.

One solution was to design the antenna system as a giant inductor/capacitor-tuned circuit by arranging the antenna as an inverted cone or "umbrella". The umbrella and earth formed the two plates of the capacitor, and the antenna coupling coil provided the inductance. Advantage of the configuration was that only a single mast was required as the guy wires could be integrated into the actual antenna. Nonetheless, even with the largest practical structures, radiation efficiency fell to an unacceptable level with frequencies less than about 20kHz.

Umbrella antennas were widely used with Telefunken medium wavelength maritime and





Nauen Antenna 1916.

relay stations and it was from these that the Nauen very-long-wave antennas evolved. The original mast at Nauen was 100m tall with a 400m diameter umbrella. But in 1909, diameter was increased to 600m, and in 1911, one side of the unbrella was lengthed to give an overall length of 1626m, with the height of the main mast increased to 260m.

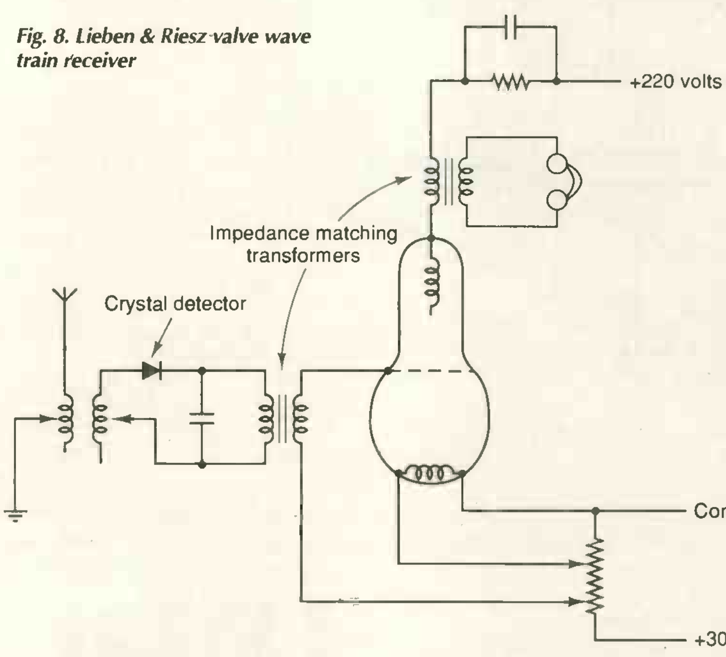
In 1916, the umbrella was replaced by a T-type antenna 1200m long and consisting of ten parallel wires, so increasing the capacity with earth. Overall length of the antenna was 2484m with – for some obscure reason – about half its length consisted of supporting cables. The two main masts were 260m tall.

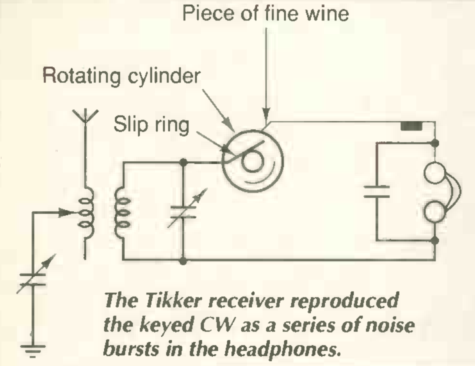
Wave train receivers.

Wave trains effectively modulated the transmission, and so, except for the magnetic detector, rectifier type detectors were the norm. Resultant DC pulses produced a tone in the headphones that corresponded to the repetition rate of the trains, typically 500 to 1000Hz

Galena crystal detectors, which conducted with a forward potential of only about 50mV were widely used with Telefunken equipment.

Fig. 8. Lieben & Riesz valve wave train receiver





The Tikker receiver reproduced the keyed CW as a series of noise bursts in the headphones.

Even so this potential was generally much greater than that of the RF current at the antenna terminals. Fortunately, damping of the tuned circuit was minimal. Headphones presented an impedance approaching 30kñ at 1.0kHz, so tuner resonance and transformation resulted in some voltage gain.

But, as early as 1913, the Telefunken company began to employ the Lieben-Reisz valve as an audio frequency amplifier, though unlike the Flemming valve, its characteristic curve made it unsuitable as a rectifier type detector.

Continuous wave receivers

An early method of resolving CW was to rectify the RF current with a crystal and then randomly interrupt the resultant DC so as to produce a sound in the headphones. Later, Poulsen and Pedserson eliminated the crystal, reducing the detector to a simple interrupter device known as the tikker; this was a fine wire lightly resting on a metal cylinder rotating at fairly high speed. Contact was intermittant so the RF current was chopped directly into on/off pulses.

Goldschimdt's "tone wheel" was essentially a commutator designed so that, at a certain speed, half wave rectification of RF current occured. But at that speed, the resultant DC gave no audible sound in the headphones. Speed was adjusted so that the tone wheel slipped in and out of synchronisation, alternately rectifying negative and positive going half waves. The tone corresponded to the rate at which the wheel slipped in and out of synchronisation.

Sensitivity of the tikker and the tone wheel was virtually the threshold of an audible sound produced by the headphones. For example, Baldwin's 4000Ω headphones were reported to gave an audible sound with a power input of 1.5×10^{-10}W. Nonetheless, the Lieben-Reisz valve could be used to boost audio output. Fessenden's heterodyne receiver, employing a variable frequency alternator as the local oscillator – was the most sensitive pre-triode receiver.

Triode valve regenerative and heterodyne receivers, which evolved during the First World War, could receive wave train or CW signals. Moreover, their sensitivity was infinitely greater than anything previously attainable, and it was this, rather than greater input power that dramatically increased the range of the Nauen transmitter.

The triode valve could have made the German imperial system independant of submarine cables. Unfortunately, it came too late and, ironically, the best valves were the French type, some of which were captured by the Germans during the war and developed for their own receivers. ■

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Thanks are expressed for the relevant literature provided by Deutsches Museum, Munchen

WHITE NOISE

by Hot Carrier

Too good not to be true

In the development lab of a company (that shall remain nameless to save red faces), the engineers were sweating over the RF power output spec of a small portable transmitter. The spec demanded 3dBW, but try as they might, with the devices available, they couldn't raise more than 1W.

The product launch was looming, and the marketing manager, who was aware of the 3dBW spec, insisted on knowing just how much they were actually getting "right now". On being told it was 0dBW, he exploded. Clearly in his eyes "zero" dBW was total failure – they hadn't even started.

Knowing he was the sort of manager to make ructions at board level, the chief engineer decided to repackage his plans accordingly.

At the all-important product review, the marketing manager complained bitterly that the lab had missed the target completely. However, the meeting congratulated the chief engineer warmly on being so close to the goal. Against the spec figure of 63dBμW, his engineers were currently achieving no less than 60dBuW.

Presentation, they say, is everything.

Restrictive practices

Input sub-octave filters in a new HF communications receiver, already overdue for delivery to customers, were causing a problem. The lower filters were all OK, but the top two, particularly the 22 to 30MHz filter, were too narrow and far too lossy. Valuable time had already been spent investigating without success, though it was obviously a design problem as all the production models showed the same symptoms.

Hot Carrier's instructions were to take a look at the situation and *not to spend more than a week on it* at the outside.

Checks showed there was nothing wrong with either the design of the filters (straight out of *Simplified Modern Filter Design* by Geffe), components used or the PCBs.

The problem was that the filters saw a fixed 1.4MHz highpass filter on one side, and I forget exactly what on the other. All of them were interconnected via banks of band-select relays and miles of PCB track. This meant the end capacitances of the filters were greatly in excess of the proper values. Reducing them – or in the case of the highest filter, removing them entirely – solved the problem completely: "Filter" problems often turn out to be problems with the termination).

Half way through the first afternoon, I passed on the good news to the manager, prompting a "Oh well, it can't have been very difficult then, can it?".

Had he been left to stew for the best part of a week before coming up with the answer I would have been the hero of the hour. So remember, don't work too hard.

Renascence of Wireless

"Glamorous" digital technology has been swamping analogue circuitry, especially in electronics degree courses. But wireless communications are back in the headlines and we are seeing a severe shortage of skilled RF engineers.

In wireless lans, European regulators are planning to define pan-continental standards for high capacity wireless local area networks. These will use short range radio links to connect PCs and workstations with fixed hard-wired data highways such as Ethernet. The frequency bands used are likely to be different from those in the US, owing to differences in prior allocations between the two continents.

Unfortunately, even limited standardisation does not seem possible in the case of microwave radio links used to implement the whole or part of a local subscriber's loop between the telephone and its local exchange. Operators in the deregulated market would naturally like an early pan-European standard to provide equipment manufacturers with economies of scale and thus bring equipment prices down. But organising a Europe-wide standard takes time, and the DTI is reluctant to go it alone in the meantime as happened with CT2. Of course, the cable TV companies with telephone franchises are not affected, they have an underground cable drop to each of their subs. But other operators such as British Rail Telecommunications – who are planning to use microwave radio tails for residential phone connections – would benefit from a standard that ensured fully compatible equipment, capable of interworking with kit from other manufacturers.

Where are the RF engineers who are going to sort out the mess?

Free software.

Japan's Miti (Ministry of International Trade and Industry), Tokyo, has announced sponsorship of a ten year collaborative programme between government and industry – the Real World computing project. Aim of the programme is to develop massively parallel computers, neural networks and optoelectronics. But this new initiative is planned to be an international effort, unlike the ill-fated all-Japanese Fifth Generation computer project. Less than a month earlier, MITI announced officially that it had closed the books on that particular project, admitting that its ten-year long research effort had failed to overtake the USA's lead. MITI says that Japan will make all the software resulting from the $400 million project available to all for free. Applications, presumably, direct to MITI.

Royal secret

Could electronics have saved the Queen the annoyance of reading her Christmas message in the papers before it had been broadcast? Bandwidth of the audio channel on the video tapes distributed to TV stations beforehand is doubtless adequate to support a digitised version of the speech – certainly this is the case if a Nicam facility is included. So audio could be digitised and an encrypted version recorded along with the video. The key could be distributed to authorised recipients such as broadcasting stations, by phone, telex or fax after the last pre-Christmas editions of the papers had gone to press. Only one trusted person need know the key. Indeed in a two key system, one key could be retained by the Queen herself. It would only remain to ensure that personnel at the recording session were frisked for concealed mini-dictation machines, oh and of course to sweep the room for radio-mike bugs: you can't trust anyone nowadays.

Balance of power

Russia's Ministry of Atomic Energy is reported to have had discussions with the US Dept of Energy about selling tonnes of uranium available since decommissioning its nuclear arms. The USA has 110 nuclear power stations, needing partial refuelling every year or so.

Curious. Surely the USA should have ample ex-weapon uranium stocks of its own.

Could low cost DSP signal the end for analogue audio?

New digital signal processing chips specially designed for low-cost home and automotive audio open up opportunities for better sound and new functions in mass-market products. Phil Atherton explains.

Apart from CD, Nicam and a little dat, domestic audio is still entirely analogue. But a recent drop in the cost of processing digital audio will shortly eliminate most analogue elements. Digitally driven loudspeakers, where a DSP chip directly controls the position of the cone, have already appeared (Philips) and a fully digital power amp could well be on the market within 12 to 15 months.

Motorola's *DSP56004* was recently launched specifically for domestic use. Based on the company's 24-bit DSP architecture first appearing in 1987, the chip is tailored to meet emerging demands in interfacing and performance required for high-volume, high-performance but low-cost consumer audio. Currently it costs around £22 in 1000-off quantities but this price will eventually fall to around £17, and lower still for larger quantities.

Obviously a device operating at tens of mips needs a high clock frequency. But fast square waves running around a circuit board create unwanted noise. To overcome this problem, Motorola has designed an on-chip low-jitter PLL for the *DSP56004* that allows slow-edge clock inputs as low as 9.8kHz to drive the chip at its rated 20mips. Future devices will have an even higher throughput.

To encourage as many people as possible to use the device, Motorola has incorporated on-chip emulation technology, allowing programming *in situ* at low cost. A complete hardware/software development system is currently around $3000, a price which although not at enthusiast level is nevertheless accessible to small companies. Competent engineers should be able to design their own development system provided they have the full interface specifications – we could well see the odd home produced DSP audio system. Passing of data to and from the host involves only a simple serial interface, **Fig. 1.**

Peripherals tailored to audio signal processing are built into the chip. An external memory interface can access static or dynamic ram with flexible address generation circuitry that accommodates nibble or byte-wide memories. Memory limits are 256Kbyte for sram or 4Mbyte for dram in single or multiple blocks. Data transfers to and from the core are in 24bit words. For accessing delay line taps in a reverberation algorithm, special addressing modes have been added to simplify the implementation.

The *DSP56004* has a built-in serial audio interface capable of receiving two stereo channels and transmitting three stereo channels via major data communication protocols, among them I²S, I²X, Sony and Matsushita. For professional and semi-professional applications, Motorola has also recently launched an AES-EBU transmitter/receiver (see box) that can

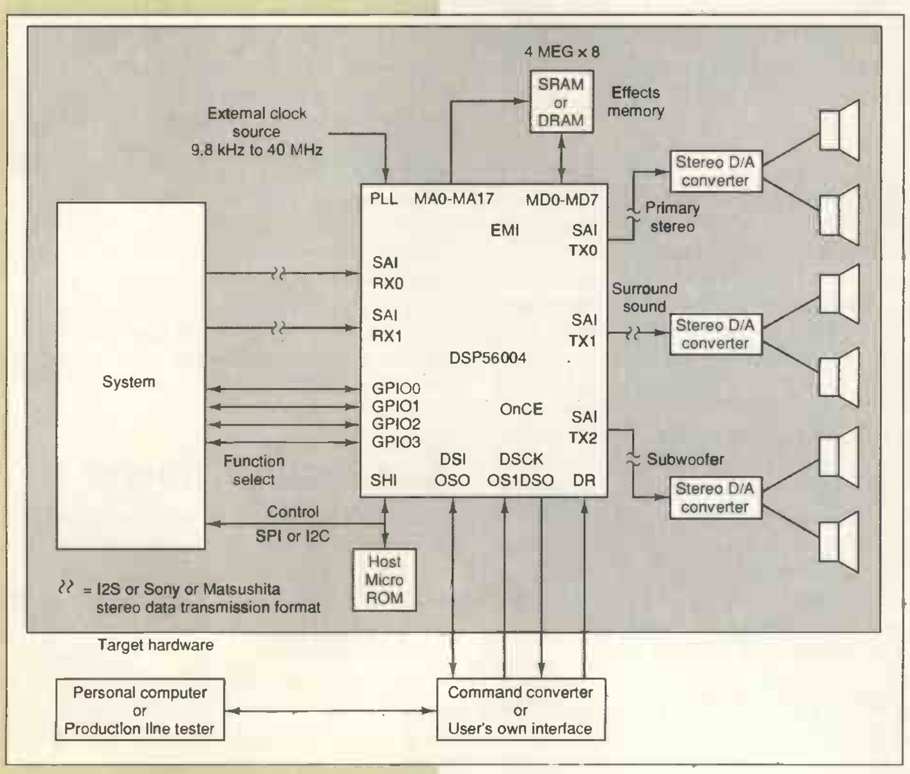


Fig. 1. Implementing the audio DSP chip is not as difficult as might be expected. The system shown represents components not directly related to audio processing such as video circuits.

combine with the *DSP56004* to provide studio performance at relatively low cost.

So that the processor is free to carry out its task without having to handle general operations such as control-signal decoding and updating, a serial host interface takes care of communication with microcontrollers through Motorola SPI or Philips I²C protocols. Up to four *DSP56004*s can be pin-programmed to respond to different addresses from the host, allowing them to be used on the same bus without contention.

TV and video

Use of Nicam and cinematic techniques such as Dolby Surround on video tapes and broadcast transmissions is prompting more and more consumers to demand high quality audio sound from their TVs. **Figure 2** shows how the processor might fit into a TV/video system to handle Nicam and produce four-channel surround sound plus monologue channel in high-quality audio. In conjunction with Motorola, Dolby Laboratories has produced an implementation of the Dolby *Pro Logic* surround-sound decoder in software for the *DSP56004*.

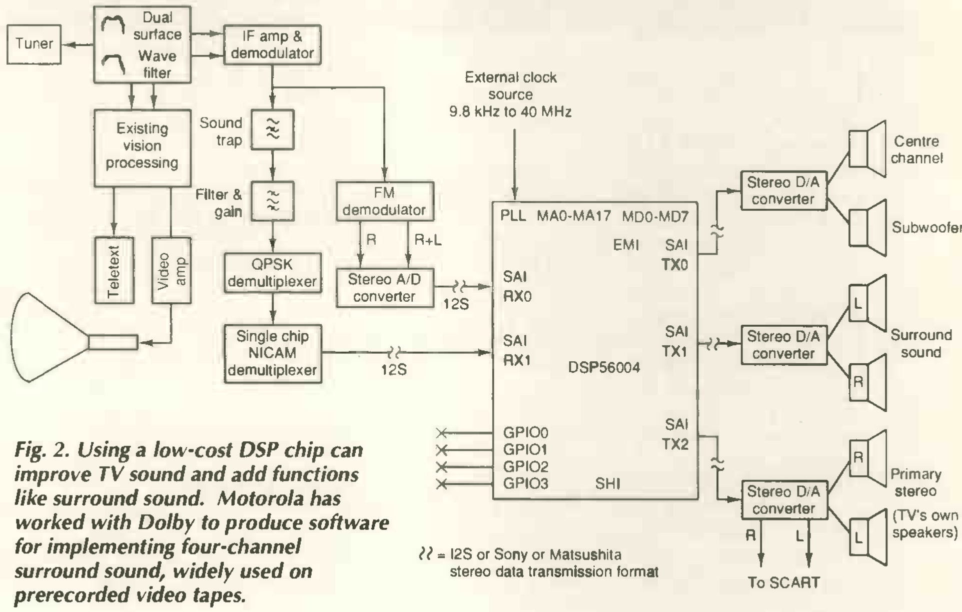


Fig. 2. Using a low-cost DSP chip can improve TV sound and add functions like surround sound. Motorola has worked with Dolby to produce software for implementing four-channel surround sound, widely used on prerecorded video tapes.

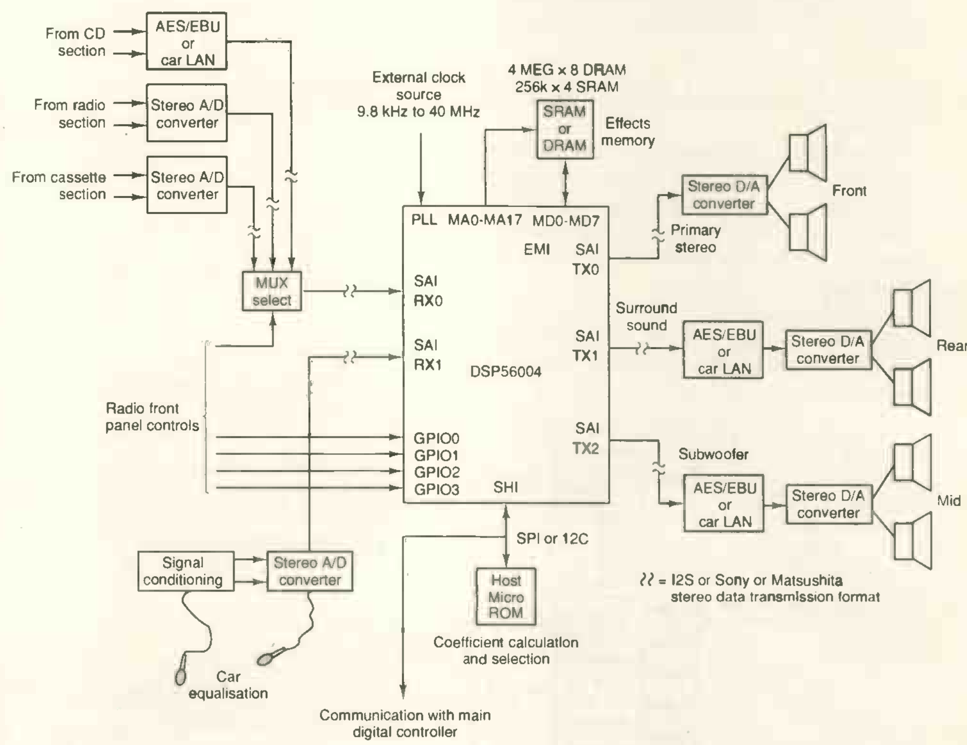
New receiver/transmitter for digital audio

Until now, transmitting and receiving digital audio in AES-EBU format has needed chip sets. But by concentrating on reducing data recovery PLL jitter, Motorola has managed to produce the first high-performance single-chip solution. It has independently-clocked transmitter and receiver sections together with four on-chip oscillators and on-chip PLL frequency and phase detectors.

The device, the *DSP56401*, is based on the same DSP technology as the *56004* but is intended for studio-type environments. But it could also find applications in top-end audio since it handles the simpler audio-only EIAJ-CP340 digital transmission standard used in CD players and rdat recorders. Besides audio, the device could also be used in conjunction with rdat recorders for mass storage for data.

The processor converts data in both directions between another dedicated DSP chip or serial audio data converter and one of the digital transmission standards. It has various levels of use from a simple interface for audio-only data to a professional system manipulating non-audio data in real time.

Although the chip is not yet priced for consumer applications, interesting applications arise for it in optical lan areas. In the automotive field for example it could act as a controller for an optical lan. The lan could not only route data around the car but also feed the rear loudspeakers, freeing them from the constraints of wire losses and distortions experienced in the electrically noisy environment. Even further in the future, devices like it could act as controllers for optical lans in the home.



Automotive

Improving the quality of in-car audio has long been a goal for manufacturers. The *DSP56004* allows automotive audio to be improved significantly while remaining part of the mass market. High-quality surround sound, acoustic equalisation and adaptive volume control are examples. Using adaptive volume control, output from the sound system is adjusted automatically to cater for changes in vehicle speed and passenger compartment noise.

Figure 3 illustrates how the *DSP56004* might be implemented. An important feature in this application area and for portable audio is physical size. The device is housed in 80-pin quad flat pack and small enough to fit easily into the confined space of a car radio.

Fig. 3. In automotive applications, DSP can not only improve sound quality but also offer features like automatic level adjustment and surround sound at low cost.

In the home

While not yet obsolete, traditional stereo amplifiers will eventually be replaced by audio-visual high quality amplifiers and home-entertainment switching centres.

Additionally, amplifiers capable of driving up to six speakers are appearing on the market to accommodate surround sound. With a DSP system **(Fig. 4)** the listener will not only be able to set the sound to emulate a predetermined venue but also to customise the

Audio converter with 107dB dynamic range

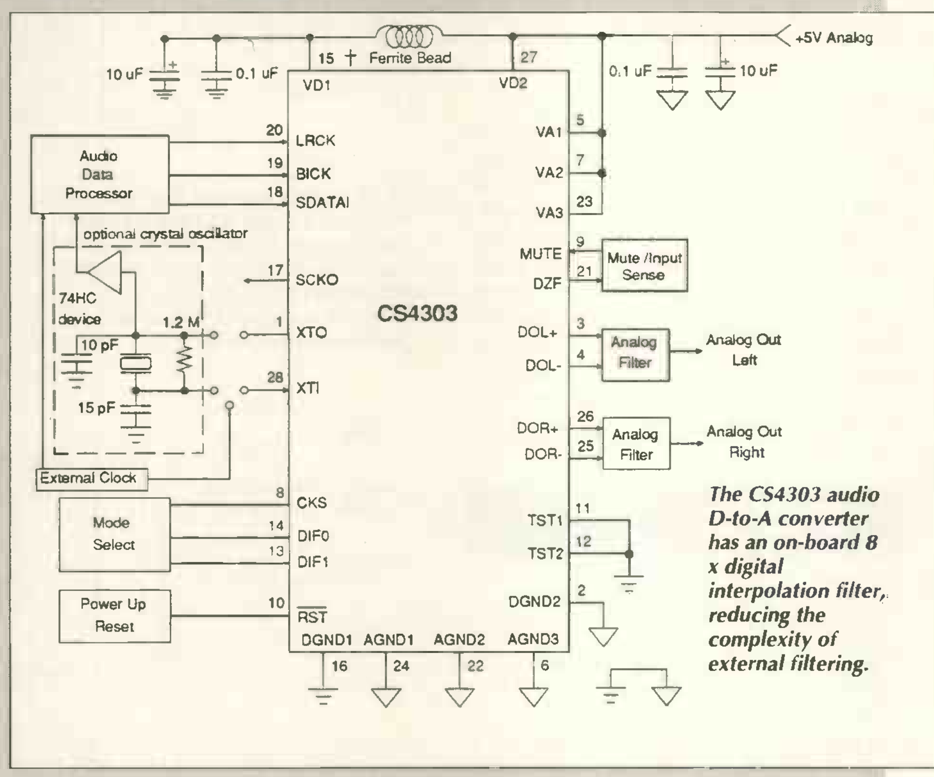
US semiconductor manufacturer Crystal has achieved a remarkably wide dynamic range of 107dB and a passband ripple of ±0.0002dB for its new D-to-A converter for professional digital audio.

The *CS4303* incorporates an 8 x interpolation filter together with a 64 x oversampled delta-sigma modulator that outputs a single-bit signal to an external low-pass filter. Input serial port is configurable providing four interface formats, and its master clock can be either 256 or 384 times the input word rate to accommodate various audio environments. Data at standard audio frequencies of 48, 44.1 or 32kHz is accepted.

Delta-sigma modulation avoids the linearity limitations of laser-trimmed resistive D-to-As and is now becoming universally accepted for high-quality audio. Its advantages include ideal differential linearity, no distortion mechanisms due to resistor mismatches and no linearity drift over time and temperature due to variations in resistor values.

Digital interpolation filter increases the sample rate by eight times to eliminate images of the baseband audio signal existing at multiples of the input sample rate. This reduces the complexity of analogue filtering since it can be based on out-of-band noise attenuation requirements rather than anti-image filtering.

A complementary device for receiving and decoding AES-EBU and S/PDIF data formats, the *CS8412*, already exists.



The CS4303 audio D-to-A converter has an on-board 8 x digital interpolation filter, reducing the complexity of external filtering.

response of each individual output to suit their own tastes.

The highly integrated, high-performance solutions which can now be achieved at significantly lower costs mean we can look forward to the benefits of greatly improved audio reproduction quality and sound effects in the near future.

■

Phil Atherton is an applications engineer with Motorola.

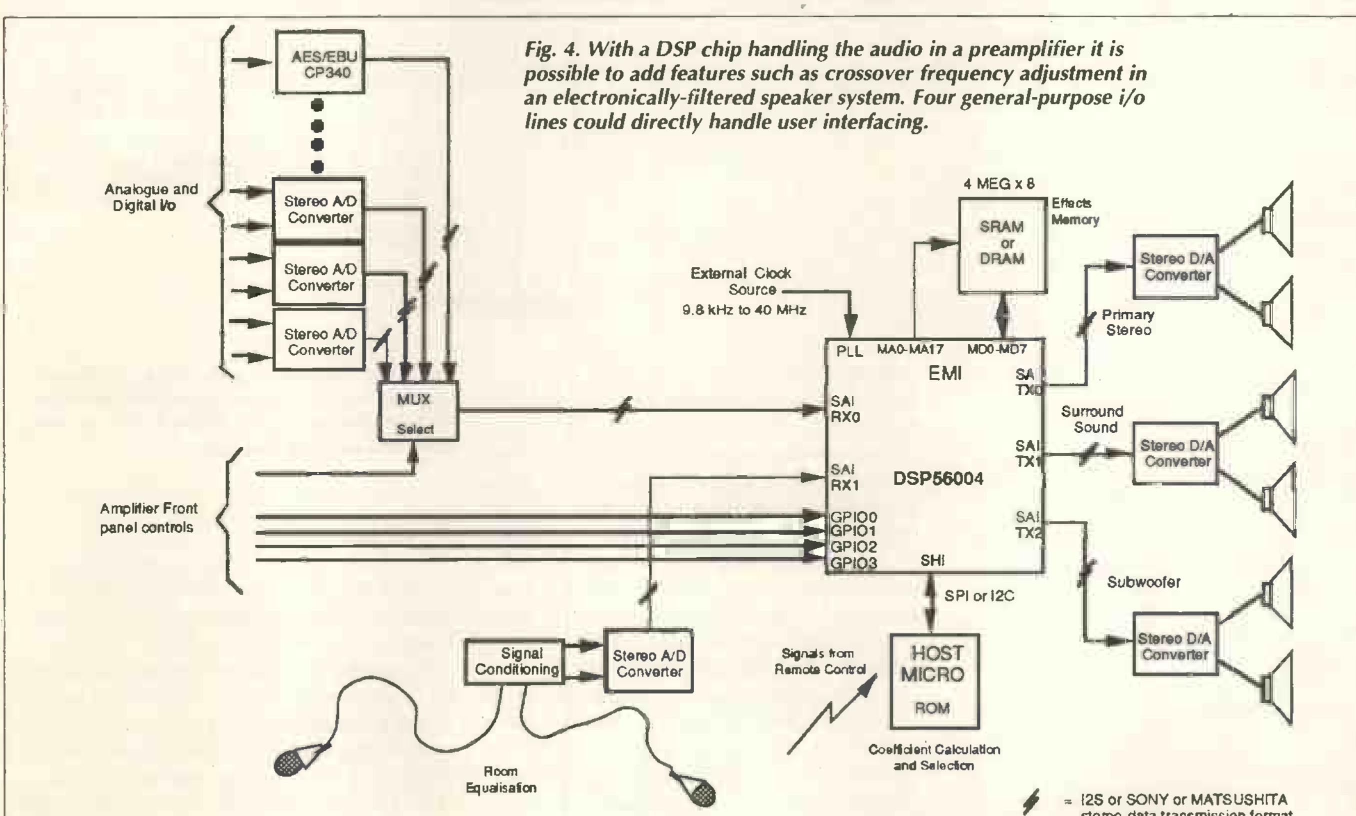


Fig. 4. With a DSP chip handling the audio in a preamplifier it is possible to add features such as crossover frequency adjustment in an electronically-filtered speaker system. Four general-purpose i/o lines could directly handle user interfacing.

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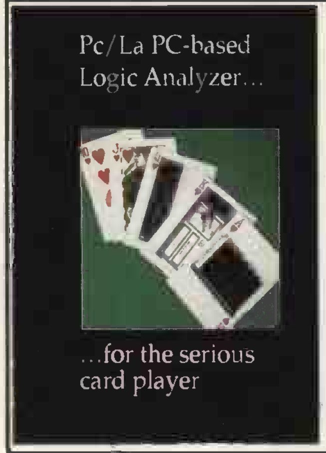

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IR Group, Europe's leading supplier of used instrumentation, has published the latest update of models available. With a range from power supplies to network analysers, most items are available on short delivery and come with a 12 month parts and labour warranty. For a detailed quotation call

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Cirkit have just published the 2nd Edition of the Toko RF Catalogue, featuring details of Tokos' extensive range of RF coils, inductors, filters and comms ICs.

The 128 page catalogue includes many new products such as; Surface mount high current inductors, surface mount multilayer inductors, helical filters at 2.5GHz and a new section of push button and tact switches.

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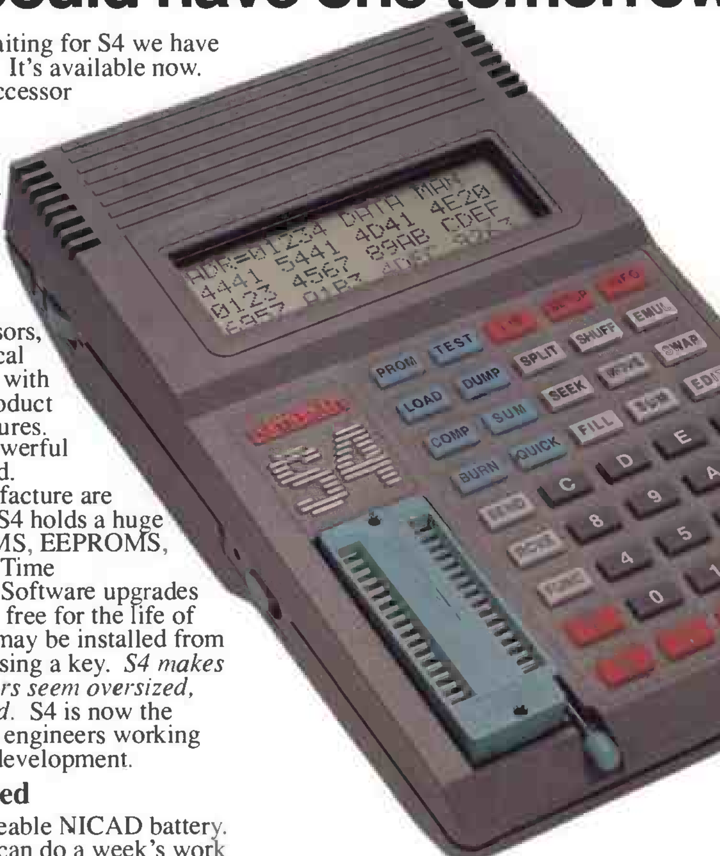
Dataman's new S4 programmer costs £495
You could have one tomorrow on approval*



If you've been waiting for S4 we have some good news. It's available now. S4 is the 1992 successor to Dataman's S3 programmer, which was launched in 1987. The range goes back through S2, in 1982, to the original Softy created in 1978. Like its predecessors, Softy4 is a practical and versatile tool with emulation and product development features. S4 is portable, powerful and self-contained. Design and manufacture are State of the Art. S4 holds a huge library of EPROMS, EEPROMS, FLASH and One Time Programmables. Software upgrades to the Library are free for the life of the product, and may be installed from a PROM by pressing a key. *S4 makes other programmers seem oversized, slow and outdated.* S4 is now the preferred tool for engineers working on microsystem development.

Battery Powered

S4 has a rechargeable NICAD battery. On average, you can do a week's work without recharging. On a single charge, up to a thousand PROMS can be programmed – and charging is fast: it only takes an hour. Normal operation can continue during the charging process.

Continuous Memory

Continuous Memory means never losing your Data, Configuration or Device Library. You can pick up S4 and carry on where you left off, even after a year on the shelf. If the NICAD battery loses all of its charge, RAM contents are preserved by the LITHIUM backup battery.

Remote Control

S4 can be operated via it's RS232 Serial Port. The standard D25 socket connects to your computer. Using batch files or a terminal program, all functions are available from your PC keyboard and screen.

Free Terminal Program

You could use any communications software to talk to S4. But the Terminal Driver program, which we include free, is the best choice. It has Help Screens to explain S4's functions and it sends and receives at up to 115200 baud – that's twelve times as fast as 9600 baud. At this speed a 64 kilobyte file downloads in 9 seconds. There is a memory resident (TSR) option too, which uses only 6k of your precious memory, and lets you "hot key" a file to S4. Standard *upload* and *download* formats include: ASCII, BINARY, INTELHEX, MOTOROLA and TEKHEX.

S4 loads its Library of programmables from a PROM in its socket, like a computer loads data from disk. Software upgrades are available free. Download the latest Device Library from our Bulletin Board.

Microsystem Development

With S4 you can develop and debug microsystems using Memory Emulation. This is an extension of ROM emulation, used for prototype development, especially useful for single-chip "piggy back" micros. When you unpack your S4 you will find an Emulator Lead with a 24/28/32 pin DIL plug and a Write Lead with a microhook. Plug the EMULead in place of your ROM. Hook the Write-Lead to your microprocessor's write-line. Download your assembled code into S4. Press the EMULate key and your prototype runs the program. S4 can look like ROM or RAM, up to 512K bytes, to your target system. Access-time depends on S4's RAM. We are currently shipping 85ns parts.

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Your microprocessor can *write* to S4 as well as *read*. If you put your *variables* and *stack* in S4's memory space, you can inspect and edit them. You can write a short monitor program to show your *internal registers*.
S4's memory emulation is an inexpensive alternative to a full MDS and it works with any microprocessor. Many engineers prefer it because their prototype runs the same code that their product will run in the real world.

Dimensions & Options

S4 measures 18 x 11 x 4 cm and weighs 520 grams. 128k x 8 (1MB) of user memory is standard, but upgrading to 512k x 8 is as easy as plugging in a 4MB low-power static CMOS RAM.
The stated price includes Charger, EMUlead, Write Lead, Library ROM, Terminal Driver Software with Utilities and carriage in U.K. but not VAT.

*Money-back Guarantee

We want you to buy an S4 and use it for up to 30 days. If it doesn't meet with your complete approval you will get your money back, immediately, no questions asked.

Call us with your credit card details. Stock permitting, we are willing send goods on 30 days sale-or-return to established U.K. companies on sight of a legitimate order.

Customer Support

Dataman's customer list reads like Who's Who In Electronics. Dataman provides support, information interchange, utilities and latest software for S4, S3, Omni-Pro and SDE Editor-Assembler on our Bulletin Board which can be reached at any time, day or night.

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