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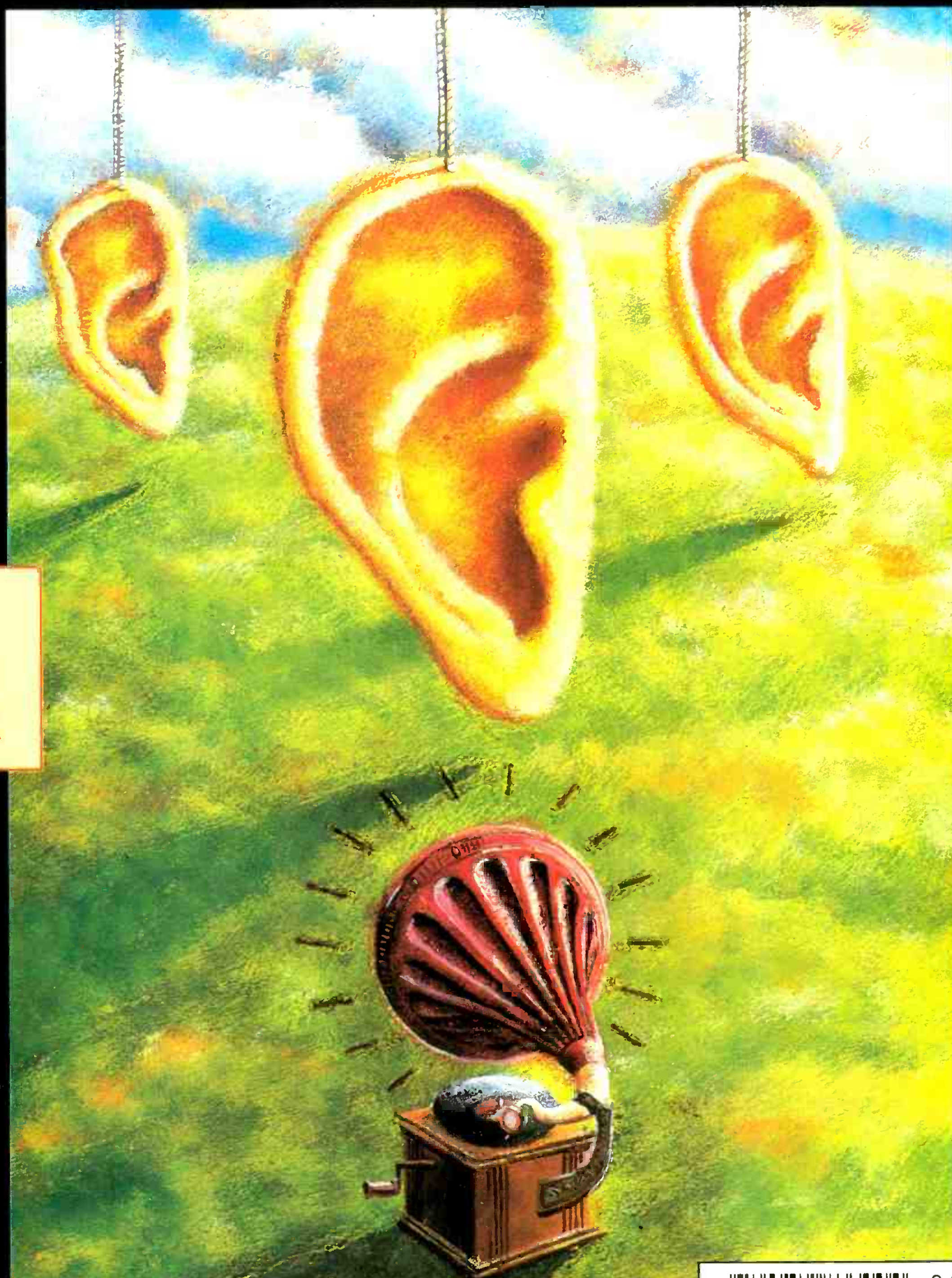

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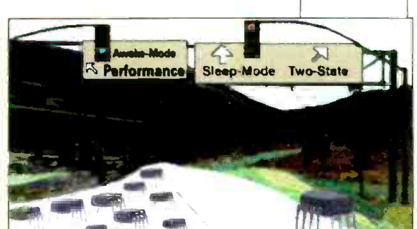

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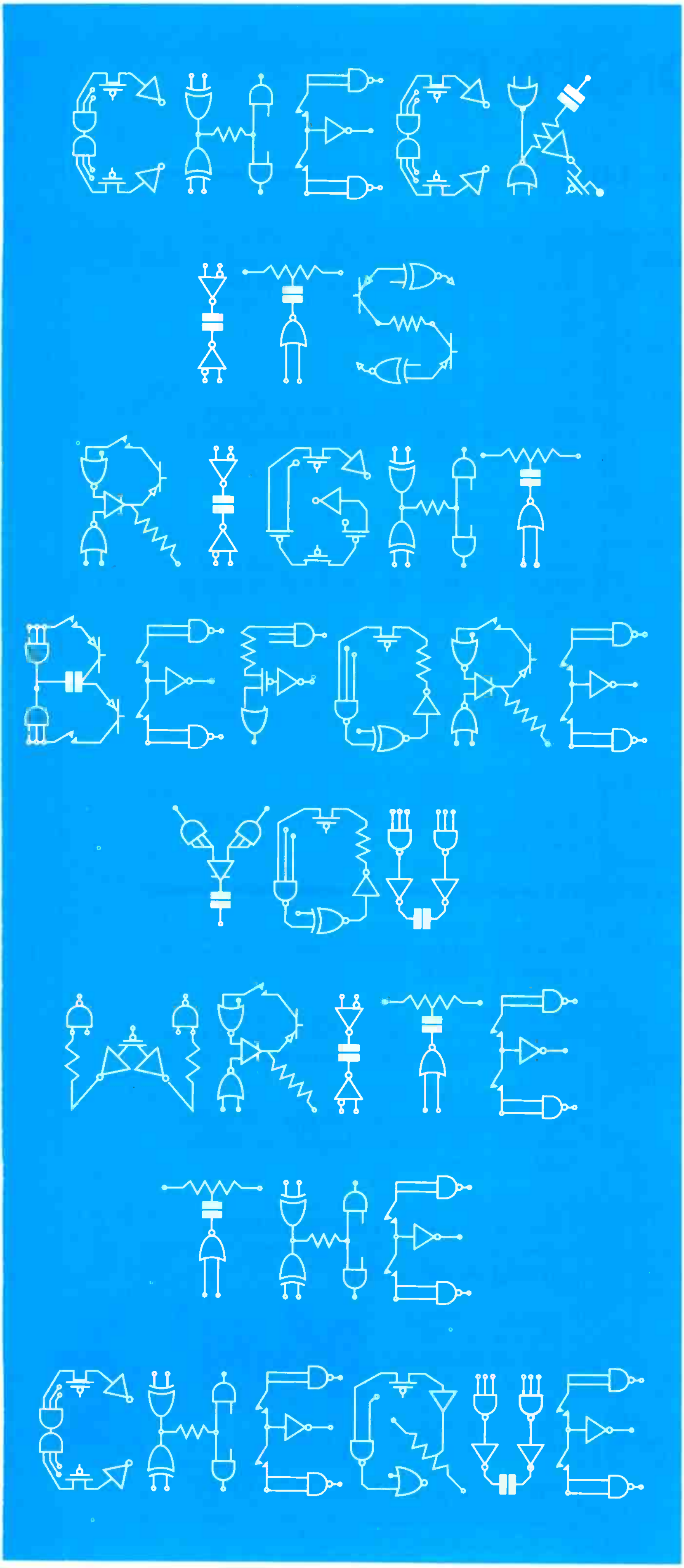

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Spoiling the picture

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The time has come to stop fooling about with analogue television standards. There can be no long term future in analogue transmission and thus there should be no further money invested in any technical television standard which is not based on all-digital coding. In short, there must be no more financial support for either Dmac or the Eureka Eu95 system. All further effort must be directed towards digital European TV.

There are plenty of reasons for going digital. Programme media such as video, music, telecommunications and computing sources already store their information in digital form. The major developments at the electronic component level will be in digital systems of higher complexity and performance. Software based customisation readily resolves the present problems of differing international standards. The spectral efficiency of compressed digital coding and its immunity to co-channel interference makes it the only sensible way ahead.

This last point should not be underestimated. The scarcity of frequency resource for both terrestrial and satellite broadcasting is already acute; digital processing makes frequency reuse highly practical.

National Telecommunications – the old IBA, Crawley Court engineering facility – has already demonstrated high quality digital television in a standard terrestrial channel space. An American competition run by the FCC for new US TV standards has already produced similarly spectacular achievements. It seems likely that when the FCC announces a replacement for the crusty old NTSC standard towards the end of next year, it will be digital. The principle is already here and mass manufacturing technology just around the corner.

So why does Europe persist in going for an analogue based mac system? The short answer is European politics. An even shorter one would be "the French". Their premier electronics company, Thomson Consumer Electronics, lost some £270 million in 1990 and, having bet heavily on mac based HDTV, is pulling every single string it can in Brussels to place mac impositions on Europe's broadcasters aided in no short measure by Philips. The public view from this lobby maintains that mac is here (the French are selling wide screen TVs at £3000 a time but are said to be disappointed with the sales) while digital isn't, and thus mac is the only logical choice for Europe.

The public's desire for widescreen HDTV is an entirely different issue – one suspects that any nation which can happily watch programme material derived from a VHS video recorder can hardly be ready for high definition TV – yet the debate does touch on digital standards. Because digital televisions can be software reconfigurable, a single set can receive, decode and display digital transmissions of any standard, aspect ratio or size.

The time has surely arrived when we should all stick two digits up at those who would look no further than national politics.

Frank Ogden

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Newstrade: IPC Marketforce, 071 261-6745.

Subscriptions: Quadrant Subscription Services, Oakfield House, Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 0444 441212. Please notify a change of address. Subscription rates 1 year (normal rate) £30 UK and £35 outside UK.

USA: $116.00 airmail. Reed Business Publishing (USA), Subscriptions office, 205 E. 42nd Street, NY 10117.

Overseas advertising agents: France and Belgium: Pierre Mussard, 18-20 Place de la Madeleine,Paris 75008. United States of America: Ray Barnes, Reed Business Publishing Ltd. 205 E. 42nd Street, NY 10117.Telephone (212) 867-2080. Telex 23827.

USA mailing agents: Mercury Airfreight International Ltd Inc, 10(b) Englehard Ave, Avenel NJ 07001. 2nd class postage paid at Rahway NJ Postmaster. Send address changes to above.
 ISSN 0959 8330

REGULARS

UPDATE

Desktop computing: riscy battle hots up

The Ace (Advanced Computing Environment) group established last year to establish a standard for risc-based desktop computing has been hit by yet another body blow. In late April, Compaq, once the leader of the pack among IBM PC compatible makers, pulled out. Its departure followed that of French computer maker Bull, which effectively withdrew from the group around Christmas and began a courtship with rivals which culminated in a technology and investment deal with IBM.

Standards in the computer business can be arrived at in one of two ways. In one, several interested parties can agree on a minimum set of requirements to which all will adhere; in the other, a product will be sold in such large quantities that it becomes established as a *de facto* standard.

Ace was an attempt by some industry leading lights to arrive at a standard for risc-based desktop computing via the first route, that of prior agreement of ground rules. The approach was in marked contrast to the means by which the IBM PC established itself as an industry standard – a fact which sceptics of Ace were quick to point out.

Principal members were computer makers DEC and Compaq, microprocessor designer Mips Computer Systems, and software producers Microsoft and the Santa Cruz Operation. A set of standards was agreed to allow computer makers to build machines around the Mips R4000 microprocessor.

These would run Windows NT, a new operating system to be developed by Microsoft which would transfer the Windows front-end GUI (graphical user interface) familiar to millions of PC users onto Mips-based single and multi-processor machines. They would also run SCO's version of Unix, and both operating systems would run on Intel-designed processors, providing an upgrade path to risc machines for millions of PC users.

"It will end in tears", the sceptics said at the time: too many diverse interests, two operating systems and two microprocessors to weld into one cohesive standard. Following the withdrawals of Compaq and Bull, they now say "It has ended in tears".

High end shake up

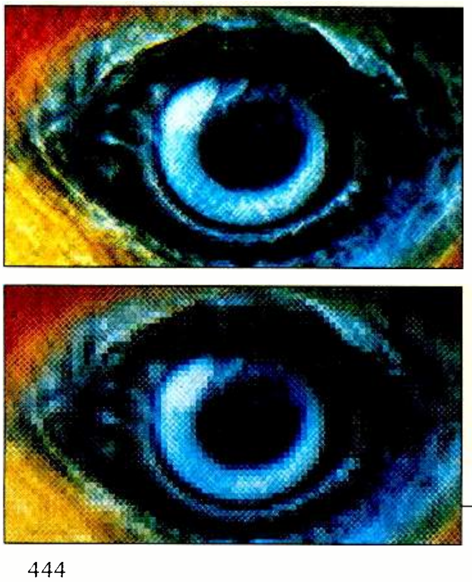
In February, DEC announced its Alpha microprocessor as a high-end alternative to the Mips device and wheeled out two customers who had agreed to build it into their computers. One of these was Kubota Pacific, then Mips' largest shareholder, who promised to build graphics workstations to compete in a market dominated by Silicon Graphics, a big Mips customer.

Some weeks later Silicon Graphics and Mips agreed to merge. Analysts interpreted this at the time as a desperate defensive move by SG to secure its hardware source. Not surprisingly, many of the Ace bashers were to be found among rival companies. In the risc computing arena they number Sun Microsystems and other Sparc vendors, Hewlett Packard – which has its own PA-risc processor and which bought workstation maker Apollo in 1989 – and IBM, still the world's largest computer maker although it lost billions in 1991.

Sun Microsystems is similar to Mips in many ways. Neither company actually makes its microprocessor; instead each has licensed its basic architecture to chip makers with varying agreements as to how it can be made and to whom it can be sold. Sun's Sparcstations, fitted with Sun's own design of Sparc processor, run a version of Unix which, bundled with a graphical front end, has recently been renamed Solaris 1.

By 1990, Sun had a 36% share of the market for technical workstations – a market that it now leads. It also had a market share

Robert Millar, Chairman, President and CEO of Mips Computer Systems, is bullish about the company's link with Microsoft.



Eye fidelity: Two different image qualities, due to much more than just the night before and morning after a hard night's squawking by the hornbill in question. The bright-eyed version is an 8 x magnification, of an original picture, produced by a lossless technique from Iterated Systems (0734 880261) called fractal transform resolution enhancement (FTRE).

High resolution enhancement is provided by expanding the pixel resolution of digitised images. Excellent detail and high quality are generated by the transform, which builds losslessly on the original image data, maintaining texture and crisp images.

The technique allows users to zoom in indefinitely without getting caught out by "blocky pixelation" (lower picture). As far as FTRE goes – *the eyes have it.*

of some 71% for risc-based workstations where its competitors largely used Motorola's 68000 processors. Since then, IBM and Hewlett-Packard have brought out the RS/6000 and the PA-risc architectures respectively. In terms of performance, both outmatch Sun's processors. A Sun Vice President, Anil Gadre, admitted to journalists recently: "If performance were all that mattered, our business would have shrunk to zero starting last year".

In fact Sun remains profitable and has some big allies. ICL, now owned by Fujitsu, is committed to using Sparc processors and has had early access to the Viking, a single-chip Sparc implementation built by Texas Instruments and due for launch this summer. Sun has also recently announced that it will license its own silicon designs, as opposed to the basic architecture, to "level the playing field for clone makers".

Sun's software division SunSoft is also due to announce its Solaris 2 operating system, which, like Windows NT, will run on Intel-designed microprocessors.

Ultimate price performance?

Hewlett Packard's PA-risc powered workstations vie with IBM's RS/6000's for primacy in desktop performance. HP launched its Series 700 range of computers in January, claiming the ultimate in price performance and promising a machine rated at 100 Specmarks before the end of the year. In April IBM responded by announcing that just such a machine, the Powerserver 970, based on the RS/6000, would be available from June.

Both companies have their own form of Unix, HP's called HP-UX and IBM's called AIX, although HP is committed to installing the Open Software Foundation's OSF/1 on its machines.

Both companies also see the importance of partnerships. Apart from Bull, IBM allows Wang to resell its risc machines and is working with Apple and Motorola to build a PC powered by its risc technology. HP also courted Bull and has licensed its PA-risc technology to semiconductor makers Hitachi, Samsung and Oki.

So, at the moment in the Risc workstation market, Sun has the installed base; Ace has Microsoft; and IBM and HP slug it out for performance and have their own particular advantages: IBM its sheer size, HP (via Apollo) long experience in the technical workstation market.

Mips' CEO, Robert Millar, said recently: "The key alliance in Ace is that between Mips and Microsoft." As long as the world's leading software house is committed to porting operating systems to a processor architecture, Ace has a chance.

David d'Arcy, *Electronics Weekly*

The potential – and problems – of radio-based TV

Mount a small, low-power microwave or millimetre-wave transmitter on the Blackpool Tower and you could provide a large part of the town with instant multi-channel television coverage. A mm-wave band would give over 100,000 people in thousands of homes immediate access to thirty or more TV channels.

Contrast this with the difficulty, time and cost involved in digging up the streets to provide a cable distribution system of equivalent capacity. Where the radio system could start making returns on the investment almost from day one, the cable installation might take years to reach equivalent financial performance.

Of course there are snags. A radio multipoint video distribution system (MVDS – see note) would have some very difficult frequency allocations to work with. In the UK these come down in practice to two bands: 12 and 40GHz. The first is already heavily occupied, primarily by the DBS service, making sharing necessary.

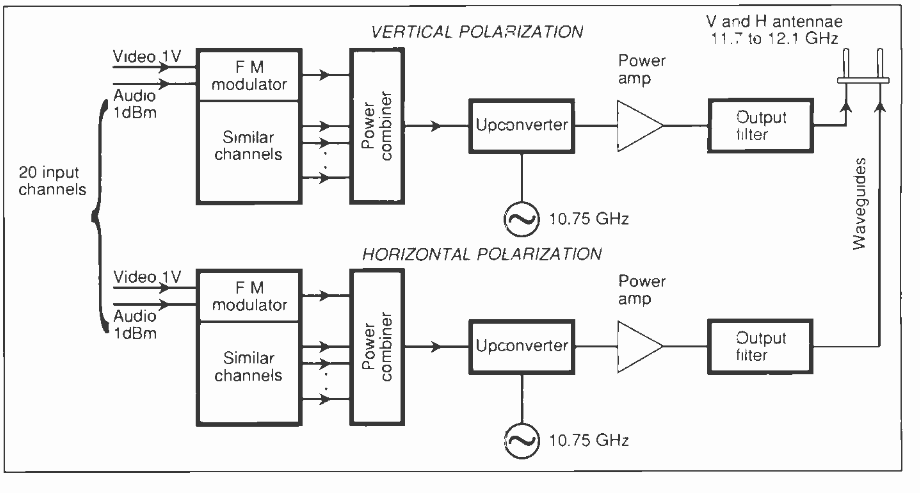
The second (40.5-42.5GHz) – which Cept recommends for Europe and the DTI sees as the ultimate goal for the UK – presents some other problems. In Blackpool rain is not an unknown phenomenon and 40GHz propagation is particularly vulnerable to rain attenuation. Also, technology, manufacturing and operational experience is almost non-existent in this mm-wave band.

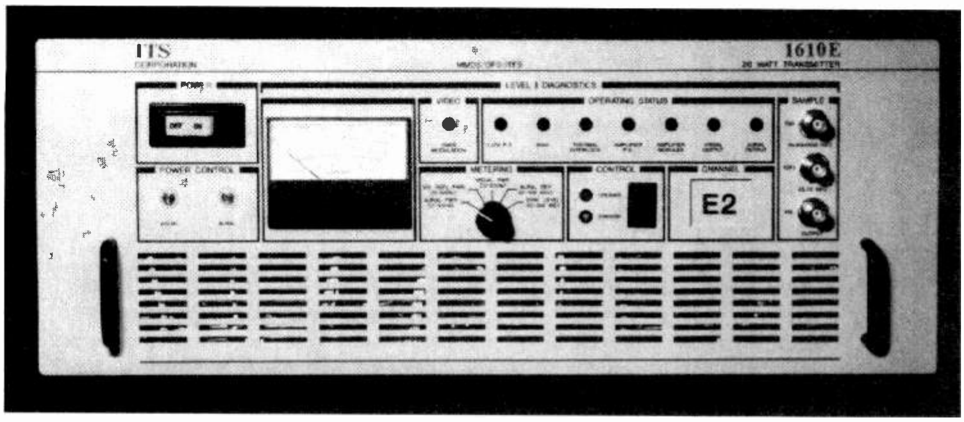
The Blackpool scenario – mentioned in passing at a recent IEE colloquium – illustrates several aspects of MVDS which are currently exercising the minds of British electronics engineers, administrators and business people. This kind of system is seen principally as a stepping-stone, supplement or alternative to cable TV distribution, though it has other uses as well. It does not yet exist in the UK, but the DTI has set up all the official machinery to allow potential operators to make system proposals and apply for licences.

Blackpool happens to be one of several urban areas which at present have no cable TV franchise in operation. Under current legislation a company would have to get a licence for local delivery of television services from the Independent Television Commission (ITC). This could be for cable or MVDS, or a combination of the two.

In 1988 Touche Ross Management Consultants reported to the Government that: "MVDS is a feasible, and potentially competitive, means of delivering additional TV channels". The opportunities for using this radio method in the UK do, in fact, seem very open. Cable TV is currently available to only 22% of homes in Britain – 2.2 million "homes passed" by cables – and less than half a million of these are actually connected. This means that the potential for

Fig. 1 Outline of transmitter for Marconi Electronics' 12GHz MVDS proposal. Use of V and H polarisation allows interleaving of frequency-overlapped 26MHz MVDS channels between five DBS channels. Cross-polar spacing between MVDS channels is 14.75MHz.





future local delivery services is considerable.

It is not just a matter of connections. The potential capacity of TV channels is enormous. Present broadband cable systems with bandwidths of 550MHz allow up to 45 channels, while the newer ones with 860MHz bandwidths can provide up to 70 channels. In the future, the possibility of 1GHz cable systems and digital compression techniques giving up to an eight-fold multiplication of channels could swell this capacity to gigantic proportions.

Scope for MVDS

In such a wide-open situation, and with the free-market "technology-neutral" approach of the 1990 Broadcasting Act, it looks as though there is plenty of scope for MVDS in the various roles mentioned above.

At the IEE colloquium ("MVDS: the way forward"), Don Hayter of the ITC and Ken Yard of the DTI said its main use will probably be in outer suburban areas, small towns or villages, with systems carrying about 30 channels. At 40GHz, transmitters would service areas with a radius of up to 3.5km. Possible interference between adjoining franchise areas would inhibit frequency re-use at distances less than 40-50km, or in some situations 20-30km.

If a 12GHz scheme were adopted, the rain attenuation which afflicts the 40GHz bandwidth would be avoided, and transmitter coverage would increase to about 6km radius for stations in town centres or 10km in other situations. Chris Wildey of Marconi Electronics described such a proposal, which he said the DTI had accepted in principle.

This would interleave around 12 orthogonally-polarised MVDS channels (**Fig.1**) with the existing five UK satellite broadcasting channels in the band 11.7-12.1GHz. Future expansion to 19 or 20 channels might be possible, he said, if the UK managed to secure a further five DBS allocations in the 12.1-12.5GHz band. Both

The 20W transmitter and modulator used in the Irish nationwide MVDS system. Manufactured by ITS Corporation, its 2.5GHz RF output stage uses GaAs fet power devices.

the 40GHz and the 12GHz proposals use FM in 26MHz bandwidth channels to carry pal or D2-mac signals.The number of homes that would actually obtain satisfactory TV service within the coverage area of a MVDS transmitter would, of course, depend on various propagation and topographic factors. Hayter and Yard thought a 40GHz system would deliver a satisfactory service to 40-70% of homes. Wildey said that a 12GHz system could be expected to do this for 60-80% of the homes.

Systems in place

Although the UK is hampered by frequency allocation difficulties, it does at least have the advantage of being able to look at the experience of other countries which have already installed MVDS. Worldwide, the five bands which are either in use or proposed are around frequencies of 2.5, 12, 17/18, 29 and 40GHz.

Very close to home, for example, since 1989, the Republic of Ireland has been pushing ahead with a nationwide MVDS scheme which aims to cover the 64% of homes with no access to cable TV. Ireland has been able to use the much more tractable 2.5GHz band because it is not already occupied by communications as it is in the UK.

At the colloquium, Patrick Keys of Off Air Electronics (Co Dublin) revealed that, in the 30 service areas licensed by the Irish Department of Communications three years ago, 10 main transmitters were now in operation. By the end of this year, 16 should be working. Service areas range between 24 and 48km radius.

The 2.5-2.68GHz band contains 22 channels, sub-divided into two separate groups of interleaved channels. Each transmitter site can provide 11 pal television channels. Use of AM/VSB avoids the need for demodulation and remodulation at the receiver and allows direct interfacing with existing TV sets.

For its part, Ireland has had the advantage of drawing on the established 2.5GHz technology and operational experience of the American educational MVDS scheme, Instructional Television Fixed Service (IFTS). This has been running in cities and suburbs since the 1960s.

According to Keys, the Irish systems have had no significant reception problems as a result of rain, snow, ice or man-made interference, except for some fading on over-water paths and slight interference from microwave ovens. Ghosting has been less serious than with UHF television. At 30km from 50W transmitters, C/N ratios are better than 50dB.

But 2.5GHz is not 40GHz and in the UK the prospect of using the official Cept/DTI millimetric waveband does not look at all good at present. Two research groups, at Essex University and Rutherford Appleton Laboratory, are still doing propagation studies and experiments for the DTI and ITC and have yet to produce firm results. Several speakers at the colloquium questioned the practicality of this frequency band. "It's very optimistic to think the 40GHz proposal is actually usable", said one. Another wondered if this frequency "is a mistake if it's intended for cable television". Yet another suggested it was "a blind alley".

Representatives of the DTI's Radiocommunications Agency seemed distinctly uncomfortable about it. They expressed doubts but at the same time, as civil servants, had to defend the Government's 1989 decision to adopt this frequency band. Ken Yard, for example, suggested that a 40GHz system will at least "get you by" until a full cable system becomes available.

In contrast, Chris Wildey was full of optimism about the 12GHz proposal he had been working on.

It was, he declared, "the only currently viable frequency band for rapid implementation of MVDS, leading to future widespread cabled services". An important problem here, though, as he acknowledged, is the control of possible interference between MVDS and 12GHz DBS and television outside broadcast services.

Tom Ivall

Note: MVDS also stands for the earlier term "microwave video distribution system" and embraces "millimetric microwave multipoint video distribution system" (M3VDS).

REGULARS

RESEARCH NOTES

Giant stereo plays sonic booms

Passengers travelling on Concorde are occasionally subjected to minor tremors as bits of the aircraft rudders fall off. But that is nothing compared to the effects of sonic booms at ground level, especially it seems *when you are indoors*.

To measure the effects of sonic booms on people and buildings, scientists at Georgia (USA) Tech Research Institute have built a small house in what seems to be a nice quiet residential area. Quiet, that is, until you step outside and came face to face with the biggest ghetto-blaster in the world.

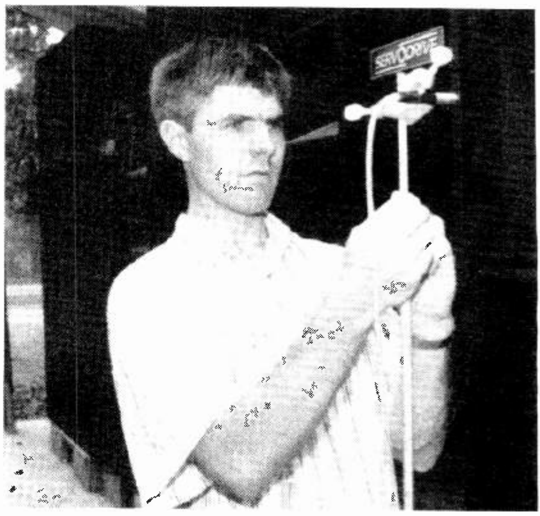
The monstrous 8 x 20 x 15ft stereo speaker system is not part of a new hate-thy-neighbour campaign. It is in fact part of a twelve month study into whether the US can develop its own fleet of supersonic civilian aircraft.

Much sonic boom research was done back in the sixties and early seventies when Concorde first entered service, but most of it concerned human reactions out of doors.

Georgia Tech researcher Krishan Ahuja says there is some evidence that people find sonic booms more objectionable when they are indoors because of the associated vibrations of objects and buildings.

Indoor research has tended to involve placing the boom-generating sound sources actually inside the building, an entirely artificial situation when the source of real sonic booms is invariably external.

Continued...



Georgia Tech research engineer Clarke Stevens sets up a microphone to measure noise emitted from the large speaker system during sonic boom experiments.

Tuning the infra-red

Researchers at the US National Institute of Standards and Technology have made a significant step forward in infra-red detection, using a combination of superconductivity and receiver techniques more familiar to microwave engineers. Nist physicists Donald McDonald, Erich Grossman and Joseph Sauvageau have opened up the possibility of making tunable devices that can scan the infra-red wavebands with the same ease now possible at radio frequencies.

Most infra-red detectors are based on semiconductors that are physically large in relation to the wavelength being received. The size ensures a large capture of the signal. But largeness does not make for particularly great sensitivity. (Imagine an RF diode 1km across and you have a good scale model of many of today's infra-red detectors!)

McDonald and his team realised that the answer to a highly sensitive infra-red detector lay in making a device smaller than the received wavelength and then cooling it in liquid helium. Superconducting niobium proved to be the ideal material, becoming more sensitive as size of the detector is reduced. The reason for the effect is quite simply that an incoming photon of radiation makes a proportionately greater electronic perturbation, the smaller the device it hits – a fact well known to all workers in the field of sub-micron electronics.

There is only one snag in making micron-sized infra-red detectors – they may be sensitive, but they have such a tiny capture area that not much radiation impinges on them. Here Donald McDonald and his team had an inspiration – do what is done with a physically small radio set: fit an antenna.

To achieve the signal-concentrating function, the team designed a spiral antenna 60μm wide, made out of gold and deposited directly on the niobium using standard lithographic techniques. McDonald admits that the idea of using an antenna at infra-red wavelengths is not entirely original, as nature apparently got there first. Some insects have spiral antennae of approximately the right dimensions which probably act as infra-red amplifiers.

The researchers are hopeful of outdoing nature and building a complete detection system of unparalleled sensitivity, combining the gold antenna and the niobium detector with a squid amplifier, all in a cryogenically-cooled package. They also plan to take this infra-red technology even further into familiar RF territory by adding a local oscillator – the idea being to create a highly sensitive narrow-band superheterodyne receiver.

As yet, there are still many questions surrounding the performance of such a system, not least its bandwidth and sensitivity.

But prospects are good for applications such as astronomy where tunability and ultimate sensitivity are vital. McDonald says that an infra-red telescope based on the new technology could well be used for detecting molecules in interstellar gas clouds or, nearer home, for identifying pollutants in our own atmosphere.

In the Georgia Tech experiments the artificial sonic booms – with various different wave-shapes – will be generated by the speaker system outdoors and experienced by volunteers seated both inside and outside the test house.

The speaker for the experiments sits on a concrete pad covered by a metal roof and shrouded in plastic when not in use, to protect it from the weather. It took two years for an outside company to produce and is undergoing final refinements at Georgia Tech.

Researchers plan to use noise produced in the 3-4000Hz frequency range.

During the research trials, a computer will generate truck, helicopter, aircraft and sonic boom noises through the speaker in random order while subjects inside and outside the house pass time reading or engaging in other activities. After each experiment, the subjects will answer questions to show which types of noise they found most unsettling.

About 150 people of all ages with perfect hearing will be chosen for the study, and will participate for two to three one-hour trials. Their hearing will be tested before and after each session.

Ahuja foresees additional research applications for the giant speaker. It might be used to broadcast sound into the sea or the sky so scientists could study how sound travels under water, as well as how turbulence in the sky affects sound propagation. But please… not in any back yard!

Changing attraction of Jupiter

Nasa may still be struggling to unfurl the jammed high-gain antenna of its Galileo space probe now *en route* to Jupiter, but valuable data on the planet is already being returned from the joint European/US Ulysses mission.

Ulysses recently circumnavigated Jupiter using the giant planet's gravitational attraction to fling it out of the plane of the Solar System, towards the Sun.

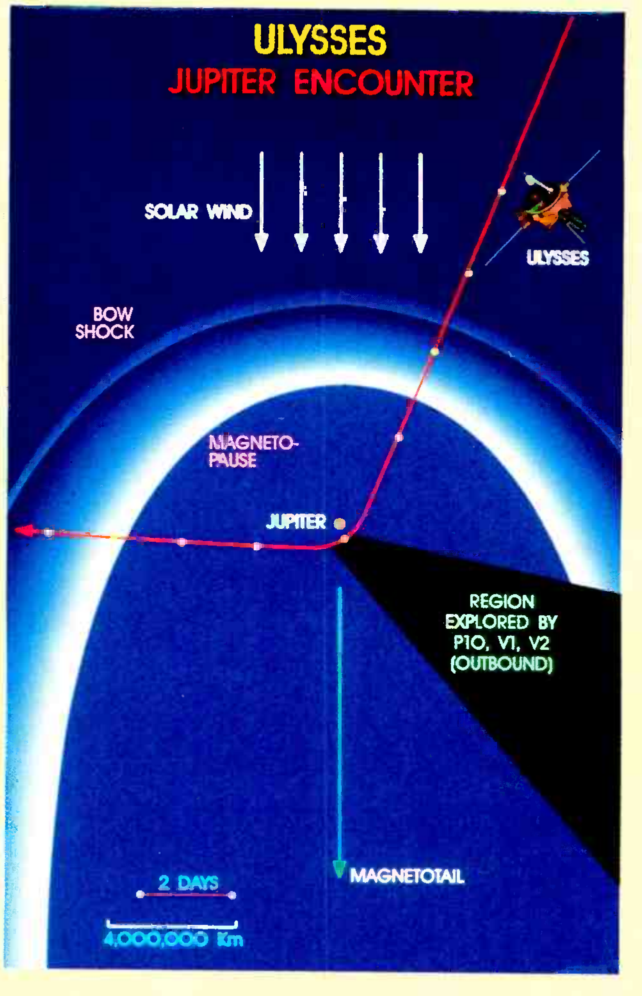
The slingshot operation was not without its hazards. The planet is surrounded by radiation belts in much the same way as the Earth is ringed by the van Allen belts, though the Jovian radiation belts are much more intense. They pose an obvious threat to electronic systems, so Ulysses was carefully routed to minimise the amount of radiation it received. Also, during the closest approach on February 8, two very sensitive instruments, a low-energy ion detector and a gamma ray burst detector, were switched off for 24 hours.

Measurements made by Ulysses, when compared to those of earlier missions, suggest that Jupiter's magnetic field is contracting and expanding. The magnetopause is now about seven million kilometres from the planet on the sunward side – twice the distance it was when the last Voyager spacecraft visited Jupiter in 1979 – but in the same position it was when Pioneer-10 flew by in 1973. (Pioneer-10, incidentally, has just celebrated its 20th year in space and is now the most distant man-made object in the Universe.)

Current speculation is that these variations are a result of alterations in the density of the solar wind, the stream of charged particles from the Sun. A reduction in the force of the wind may allow the magnetic field to expand. Ulysses' observations have revealed a cyclic variation in the strength of Jupiter's magnetic field with a period of 10h – the length of the Jovian day.

Ulysses also began to detect radio emissions when it was still a considerable distance from the planet. It used the directional capabilities of its radio and plasma wave experiments to search for the still-unknown sources of these emissions and also the X-ray emissions from within the magnetosphere. Possible candidates include the auroral zonesand a strange ring of plasma that surrounds the mysterious moon, Io.

With all these fascinating unanswered questions, it will be a desperate disappointment for space scientists if the problems of the Galileo probe cannot be righted before it reaches Jupiter in 1995.



Ulysses left the Jovian system via the dusk segment of the magnetosphere, a region not traversed during previous encounters by the Pioneer 10 (P10) and Voyager 1 and 2 spacecraft. Dots show seven day intervals.

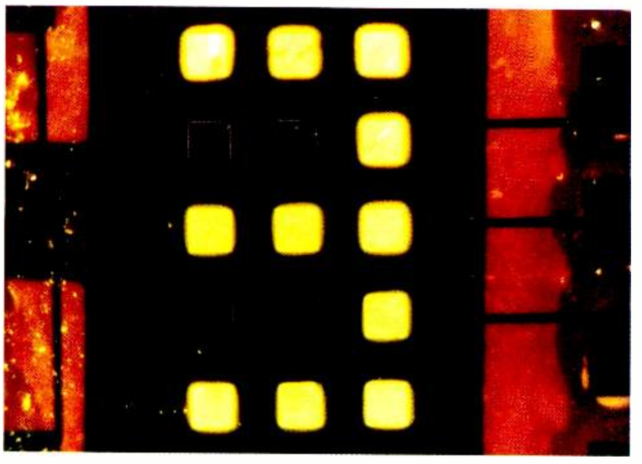
Will polymer diodes challenge colour LCD?

Semiconducting polymers have made enormous strides over the last half decade or so with development of totally organic transistors and all-polymer leds. Their attraction – apart from potential cheapness – is that polymers, unlike inorganic semiconductors, can be configured in almost endless variety. Polymers can also be manufactured in a variety of sizes and shapes without the constraints of wafer size.

The latest development in semiconducting polymers is a family of conjugated co-polymers developed from poly(*p*-phenylenevinylene) or PPV. This family, which can imitate the electronic properties of III-V quantum well structures, can be chemically "tuned" to provide an almost infinite range of properties, including variation in colour of light emission when configured as electro-luminescent diodes.

Now organic chemist Andrew Holmes and colleagues from Cambridge University Chemical Laboratory, along with physicist Richard Friend and researchers at Cavendish have reported (*Nature*, Vol 356, No.6364) that the luminous efficiency of devices that include co-polymers can be made comparable with inorganic devices in the blue-green part of the spectrum. At slightly lesser efficiency, emission has been demonstrated over the range of blue-green to orange-red, almost the whole of the visible spectrum.

The team emphasises that a great deal of work remains to be done, not least in respect of the metallic electrodes used to inject electrons into the polymers. Nevertheless, the potential of a range of fast responding electroluminescent materials that can be manufactured easily and robustly must surely be enormous. The team says that variations at the molecular level should enable the production of multi-colour displays that will be more than competitive with LCDs.



Light emission form a 5 x 3 pixel display showing the number 3. The individual pixels are 1mm^2 and are driven at around 0.5mA to produce a level of brightness appropriate for displays under ambient lighting (500cd/m^2). Efficiencies, measured as photons out per electrons injected are in the range 1.5% – higher than for commercial inorganic III-V green leds.

The ULF spike that could save thousands of lives

The Turkish earthquakes of a few months ago have provided yet more impetus for discovery of a reliable way of forecasting these natural disasters. Yet, in spite of considerable efforts, the best earthquake predictors still seem to be neurotic dogs and electrodes attached to house-plants. But there has been tantalising evidence for some time that big quakes are preceded by bursts of low frequency radio waves.

Evidence has come from a variety of satellite and ground station measurements in the ELF and VLF ranges, ie from 300Hz to 30kHz. No reliable connection has been established between natural emissions on these frequencies and subsequent earthquakes. But new findings presented at a recent meeting of the American Geophysical Union suggest that more dependable RF prediction of big quakes might be possible if we target the ULF range, especially the band between 0.01Hz and 10Hz.

Antony Fraser-Smith, an electrical engineer at Stanford University's Space, Telecommunications and Radioscience Laboratory, reports a strong precursor signal to the Loma Prieta earthquake in 1989, found while he was involved in a navy-sponsored study to determine how natural noises interfere with satellite communications. He believes the ULF band could be a better predictor of quakes because signals in this range can penetrate 10 to 20miles of earth without significant attenuation.

Fraser-Smith's equipment was set up in the Santa Cruz mountains to monitor ULF radio waves produced primarily by the solar wind as it distorts the Earth's magnetic field. Twelve days before the Loma Prieta Earthquake, the detector went wild, indicating a signal 20 to 30 times bigger than anything previously recorded. Since that time analysis of chart recordings made in the republic of Georgia (formerly the USSR) has revealed a similar ULF spike that preceded the big Spitak-Armenia quake in 1988 that claimed more than 30,000 lives.

Clearly two swallows don't make a summer. Even so, scientists from Greece, France, China and Japan all presented findings at the American Geophysical Union meeting, supporting the contention that ULF signals may precede major earthquakes.

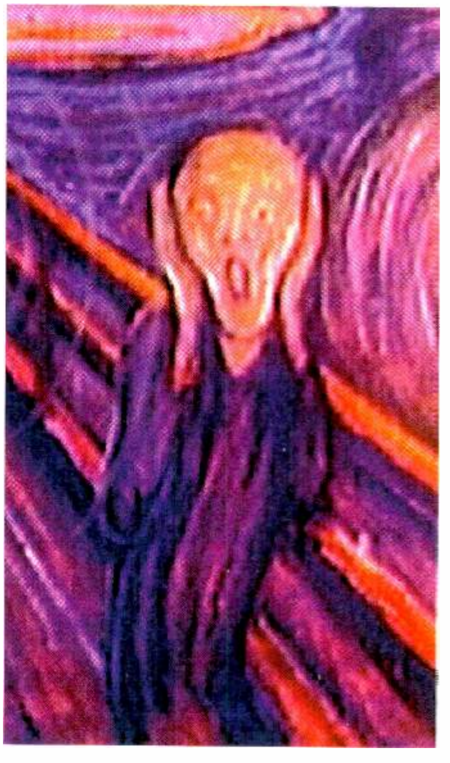
No-one knows what causes the signals, though numerous theories abound. These include the suggestion that ground-water carries currents as it pushes through rocks and fissures in the earth. Another explanation proposes that when rocks of different conductivities slide past one another they generate electricity.

But while theorists pursue an explanation, Antony Fraser-Smith, with the help of the US Geological Survey, has taken the practical step of setting up two ULF detectors in California. Dubbed the "Watched Pot" by geologists, the equipment could soon put the theory to the test. Parkfield, the equipment site, is already overdue for a "biggie" quake of magnitude greater than 6.0.

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

HIGH FIDELITY, LOW FEEDBACK, 200 WATTS

Valve power amplifiers could never use the values of open-loop gain which even the most basic transistor configurations routinely achieve. Phase shift in the high impedance circuitry and reactance in the output transformer associated with valves creates havoc in a very closed loop. However, this lack of feedback may actually enhance sound quality by suppressing transient intermodulation products. Andrew Hefley uses the lessons of valve design to produce a highly linear, low gain, wide band transistor power amplifier.



Most power audio designs over the last twenty years have used large amounts of negative feedback to ensure low closed-loop harmonic distortion. The main contributors to this type of distortion are the output devices. Other sources derive from the drivers, or devices preceding the output devices and the high voltage gain stage, also referred to as the transconductance stage.

A side effect of a high open-loop gain is that wide bandwidth is very difficult to achieve with two gain stages, therefore most of these high gain designs have an open-loop bandwidth of less than a few kilohertz. This means there is a need for even more than necessary gain mid-band (referring to the audio band) to get a low distortion number at 20kHz in the closed-loop condition.

One approach is to use lower open-loop voltage gain sections and increase the open-loop bandwidth to 20kHz. Better still might be to install multiple output devices to keep the current excursion low for each device so as to stay within a linear range of gain. Although a popular approach, it is a costly solution. A new pair of complementary Motorola bipolar power output transistors, the *2SC3281* and *2SA1302*, exhibit a high degree of linearity over their specified current range; enabling them to replace multiple transistors. The amplifier circuits which follow make use of this. Other attributes include a wide open-loop bandwidth with less than 25dB of negative feedback.

Low feedback, low distortion

A circuit diagram of a 100W amplifier is shown in **Fig. 1**. This design begins with the *MPS8099* and *MPS8599* complementary pair for the input stage. These devices are arranged as dual differentials. A 24V zener with 10kΩ resistors supplies approximately 2.25mA of current to each pair. This scheme supplies enough current to achieve the bandwidth necessary in the first stage while keeping the bias current low enough for an acceptable amount of input voltage offset error.

The second stage of voltage gain, sometimes referred to as the transconductance stage, is made up of a darlington pair. The input devices of the darlington pairs are the *MPSW06* and *MPSW56* respectively. Idling current is approximately 3.5mA which in turn sets the idling currents for the *2SC3298B* and *2SA1306B* at approximately 50mA or 2.4W in each of these devices. A small heat sink is required to keep case temperatures down. The transconductance stage is loaded by both the output stage and a pair of 2.7kΩ resistors which set the voltage gain of this stage.

Looking at the input stage and the second stage respectively, their gains are approximately 18.5dB and 34dB, producing an overall gain of about 52dB. Compensation networks are used on the outputs of both stages to provide adequate gain and phase margin for the closed-loop condition. The closed-loop gain of the amplifier is set at 27.8dB giving an input sensitivity of one volt.

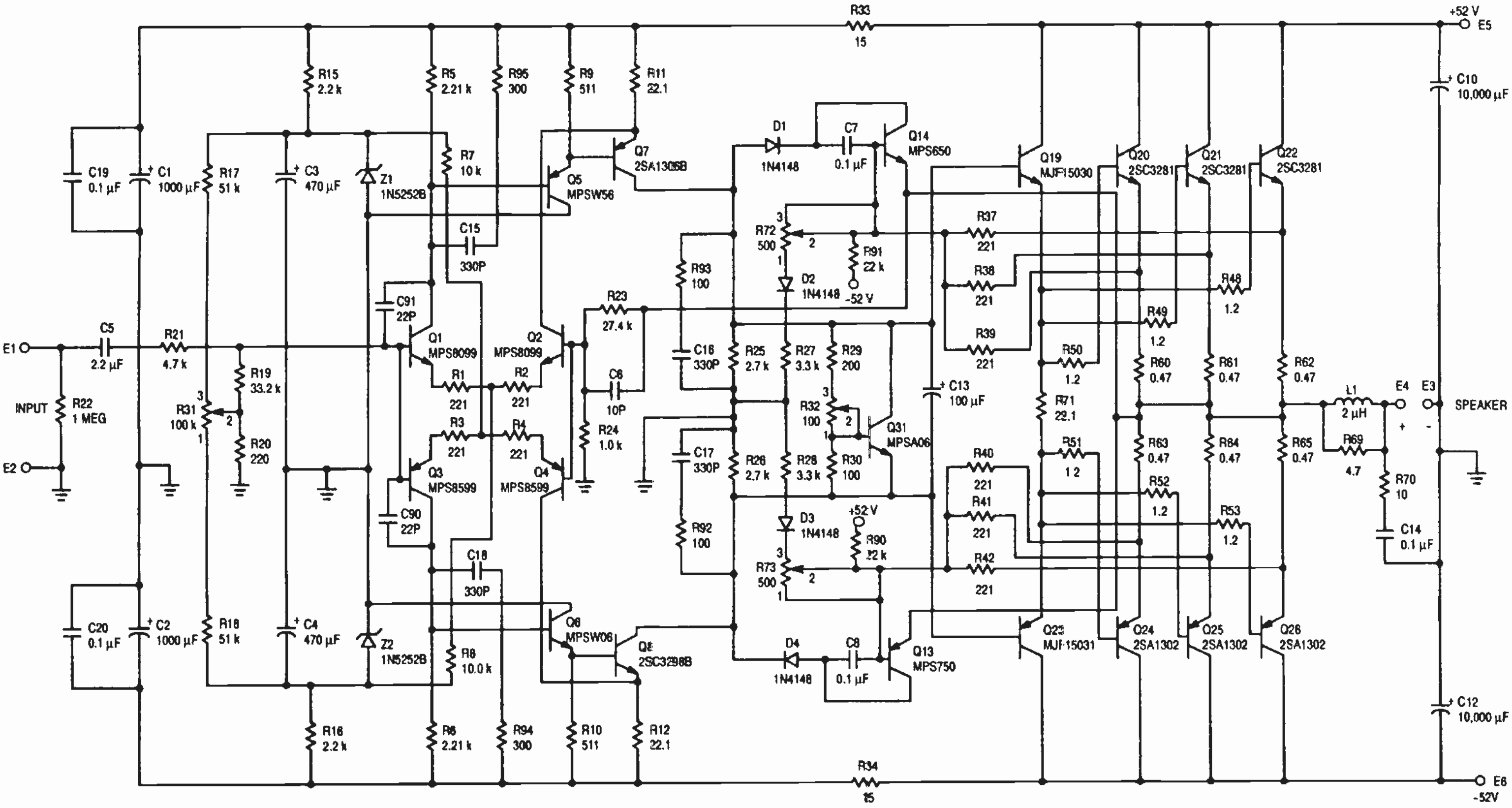


Fig. 1. 100W amplifier schematic. Open-loop gain is about 52dB which restricts feedback to less than 25dB. Like valve amplifiers, which exhibit similarly low levels of feedback, this contributes to the amplifier's low transient intermodulation distortion and its resulting 'clean' sound. Measured THD is around 0.02% at full power.

The output stage is a complementary darlington configuration. This stage uses three *2SC3281* NPN devices and three *2SA1302* PNP devices connected in parallel. These are driven by the complementary pair consisting of an *MJF15030* and *MJF15031*. The output devices are rated at 15A and 200V with power dissipation ratings of 150W. The drivers are 8A, 150V transistors with power dissipation ratings of 36W. The voltage ratings are adequate to handle the 100V nominal supply voltage. The *MPS650* and the *MPS750* devices are used for current limit protection. Both devices are rated at 2A producing excellent saturation and gain characteristics at 100mA.

Doubling the power

The circuit diagram of a 200W version of the amplifier is shown in Fig. 2. The 200W gain stages are very similar to the 100W amplifier with a few minor exceptions. Due to higher power supply voltages, a cascode configuration was used for the input stage. The level shifter portion of the cascode is tied to the 33V zener supplies which are used for the input current resistors.

One change is the use of paralleled pre-drivers, or transconductance stage transistors. This accommodates the increased current needed to bring the dual 2.7kΩ load and increased base current requirement of the output stage to the higher supply voltage. The benefit is a small increase in the voltage slew rate and an increase in the open-loop gain by approximately 4dB. This matches the extra gain needed for full power at one volt input.

Another change is the use of a cascode output stage. By effectively doubling the number of output devices without increasing the voltage seen in operation, secondary breakdown will not be a concern. The outside, or slave-devices, are driven by a series resistive divider network tied to the output of the amplifier. This divider network forces the string of output devices to share the voltage and power delivered to the load.

Output transistors

Areas of design importance include breakdown voltage, power dissipation, safe operating area (SOA), current gain linearity, and fT. At present, there are a limited number of complementary devices rated at 100V in plastic packages. Additionally, very few of these devices have good current gain linearity in the region beyond one amp. Most exhibit second breakdown points that usually fall between 20 and 40V. They may be classified as 150W transistors, but operate efficiently only up to 40V; then their power handling capability drops off rapidly. The output devices selected are the *2SC3281* and *2SA1302* NPN and PNP transistors rated at 200V, 15A. Power dissipation ratings are 150W in the TO-3 PBL package, a high power plastic encapsulation with an isolated mounting hole.

Setting the current limits of an output stage in an audio amplifier is not an easy task. Calculation may be used to find a starting point, but the actual results must be determined through experimentation.

There are two limitations to consider when dealing with the power handling ability of a transistor: the average junction temperature and second breakdown. In a class AB output stage, the output devices are not really in a 50% duty cycle situation. The bias current needs to be added to the calculated current that the load may present to the device. At higher frequencies the peak power can be considerably higher than the average power. At low frequencies the duration that one side of the output stage may endure during a load condition may be several hundred milliseconds, which nearly constitutes a DC condition.

The second breakdown of a transistor can severely limit the power dissipation capability of that device. When the supply voltages of an amplifier are greater than 100V, the output devices are pushing their limits. By configuring the output devices in a series-parallel configuration, rather than configuring them all in parallel, one can obtain an increase in output power from a set of devices. The load lines of the 200W amplifier indicate a constant power curve of 750W, which is twice that of the 100W amplifier. There are twice the number of output devices connected in a series-parallel configuration resulting in the same 125W criteria for each as in the 100W amplifier. Since these devices do not see more than 80V each in operation, second breakdown is of no concern. Although the devices can be operated in a series connected output stage, its operation is similar to a bridge configuration but

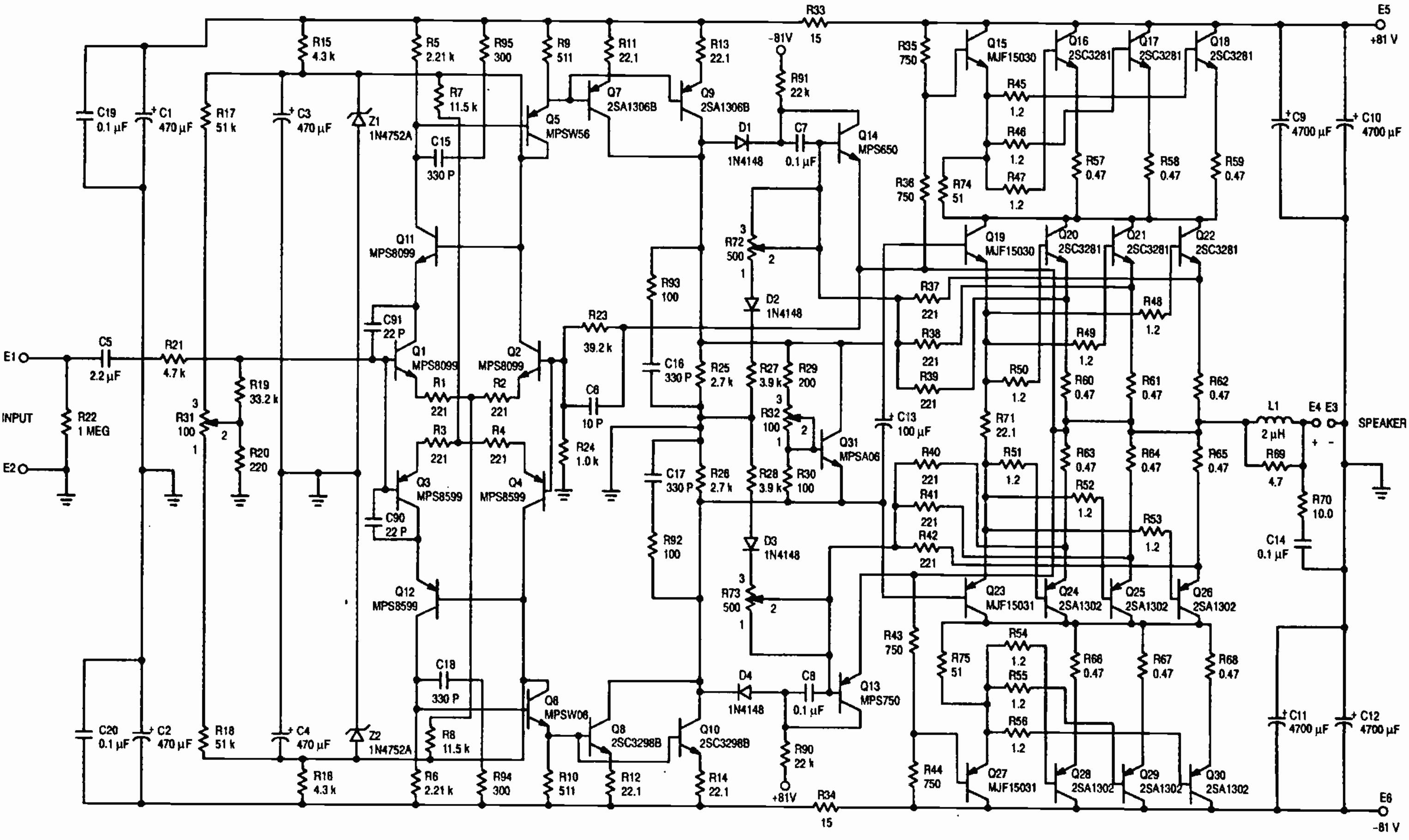


Fig. 2. The 200W amplifier version uses an output transistor tree to keep the peak voltage excursion across the output devices to around 80V, making the design almost immune to second breakdown failure. Note also the cascode tree at the input stage limiting the voltage swing across individual devices. There is virtually no distortion penalty in using the circuit arrangement.

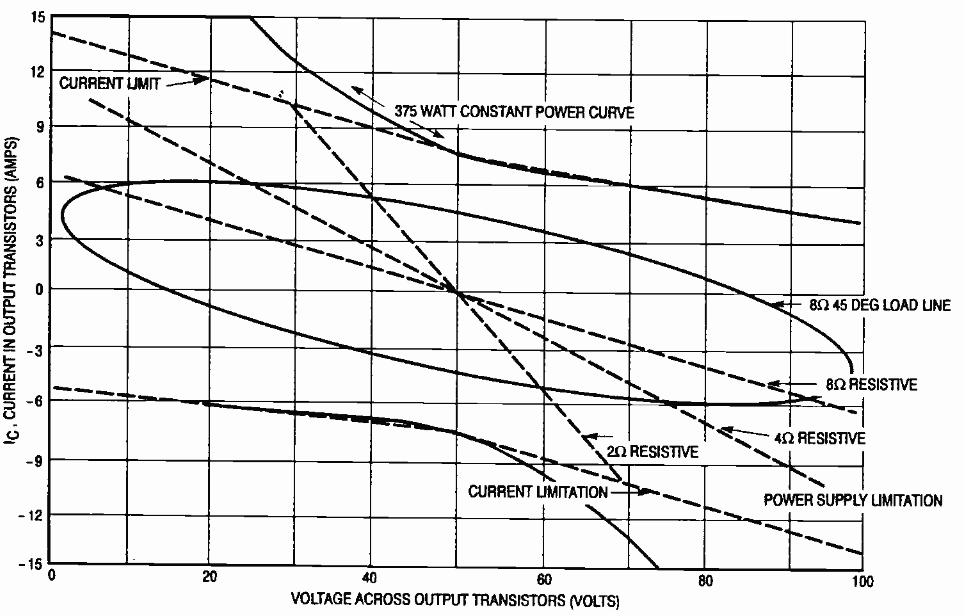
the low impedance performance is diminished. This amplifier does current limit when driving a 2Ω load where the 100W amplifier does not.

Heatsink requirements

Choosing the right heatsink is important. Regardless of the requirements of size, shape, form factor and cosmetics, the bottom line is heat transfer. The heatsink chosen for these two amplifiers is a standard aluminium tree shaped extrusion weighing about 3kg/metre. It has a surface area of 80cm2 per linear cm and a convection heat transfer rating of 1.8°C/W per 8cm piece. The use of two 18cm extrusions in the 100W amplifier gives the output stage the ability to dissipate about 90W while keeping a temperature rise of less than 35°C above room temperature (TA = 25°C). Tests run have shown that with the 100W amplifier running at full power, the heatsinks remained below 60°C. The addition of a fan blowing across the heatsinks will allow the amplifier to operate at a 4Ω load continuously. ■

Further reading

Motorola application note AN1308.



This diagram shows the load conditions for the entire output stage of the 100W amplifier. The load line for an 8Ω, 45 degree load indicates the need for 6A of collector current dropping to 4A of collector current when the collector voltage drops to 50V. The diagram also shows the peak currents at lower resistive load conditions. Most loudspeakers present a load impedance of less than 8Ω; however, most don't present a reactive load of less than 8Ω. A 90° reactive 8Ω load is handled easily and resistive loads as low as 2Ω do not present a problem. The constant power curve shown at 375W is drawn along the line where the current limits were set for this amplifier. If this power is shared between three devices, they will each be required to dissipate 125W. Detailed design attention must be paid to output device heatsinking.

Programmable digital sine wave generator

It was Fourier who first described how square waves are made up from sine waves of odd-multiple frequency components; the reverse of this theorem produces sine waves from square waves. A high frequency square wave put through a lowpass filter can produce a sine wave. The result of Fourier's transform can be demonstrated in a single chip. Gordon Lindsay*.

Gordon Lindsay is an applications engineer with Sierra Semiconductor

The single chip SC11313 programmable sine wave generator incorporates a square wave generator and a filter designed with a cut-off frequency below all higher harmonics of the square wave signal. Typically a fourth order filter provides sufficient attenuation. The lower order harmonics, particularly the third, may require further filtering. The Sierra chip uses a fifth order bandpass filter combined with a notch filter to attenuate the third harmonic of the required sine wave signal.

The chip is partitioned into a square tone generator and filter section. The sixteen LSBs of a 32-bit control word are used to set the square wave frequency using a programmable divider from a master clock. The actual frequency of the square tone is equal to the master clock divided by twice the programmed value: the divide by two stage ensures a symmetrical output for the square wave tone. The tone is then passed through an anti-aliasing filter to attenuate very high frequency harmonics. This is available at an external pin where it may be additionally filtered or used to provide timing for other circuits.

Twelve bits of the 32-bit control word program the bandpass filter by setting the sample clock frequency F_c for the switched capacitor bank. Again, a divide-by-two state is added to the filter clock generator to ensure a symmetrical output. This means that the master clock frequency can be divided by an even number from 4 to 8190.

The filter is designed to have the centre of its passband at $F_c/51$ and then a notch at $3F_c/51$. The minimum attenuation outside the passband is 50dB. The four MSBs of the 32-bit control word set the attenuation of the output signal. The maximum attenuation is 15dB. The sinewave output is thus produced at a frequency of $F_c/51$, from DC up to a maximum of 8kHz. An uncommitted op-amp with an output impedance of 600Ω is available on the chip to use as a smoothing filter for the output signal. Switched capacitor filters often exhibit staircases on the output waveform because of the sampling action used by the filter: this is easily removed by passing the waveform through a low pass analogue filter.

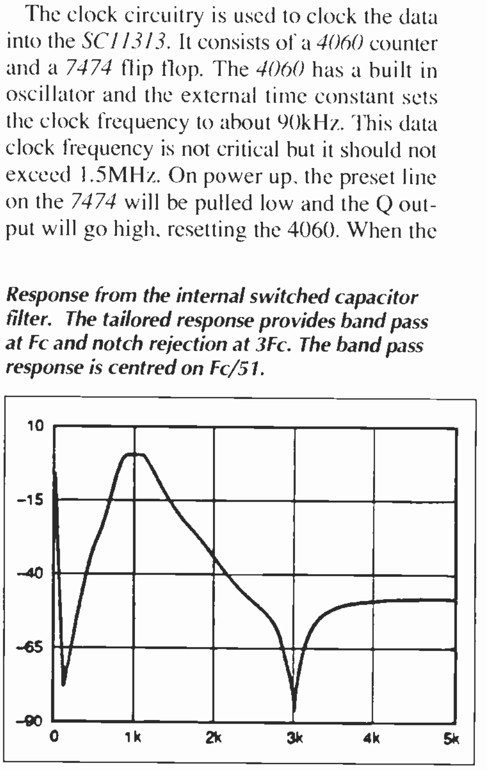
The resulting sine wave enjoys crystal stability. The *SC11313* has an amplitude variation of only ±0.1dB from DC to 3.5kHz increasing to ±1dB up to 8kHz. Gain error is kept to ±1dB over the complete frequency range. Total harmonic distortion is below 30dB even with a 10V pk-pk output signal. Third harmonic is less than –45dB.

Practical design

To use the *SC11313* a simple evaluation circuit can be made. The board has three distinct sections. The data registers, the clock and the analogue circuits. The data registers consist of four *74LS165* shift registers. These are asynchronously loaded with the data from the switches by pressing the load button. The shift registers are looped back so that the data is held in them even when it is sent to the *SC11313*.

The clock circuitry is used to clock the data into the *SC11313*. It consists of a *4060* counter and a *7474* flip flop. The *4060* has a built in oscillator and the external time constant sets the clock frequency to about 90kHz. This data clock frequency is not critical but it should not exceed 1.5MHz. On power up, the preset line on the *7474* will be pulled low and the Q output will go high, resetting the 4060. When the

Response from the internal switched capacitor filter. The tailored response provides band pass at Fc and notch rejection at 3Fc. The band pass response is centred on Fc/51.

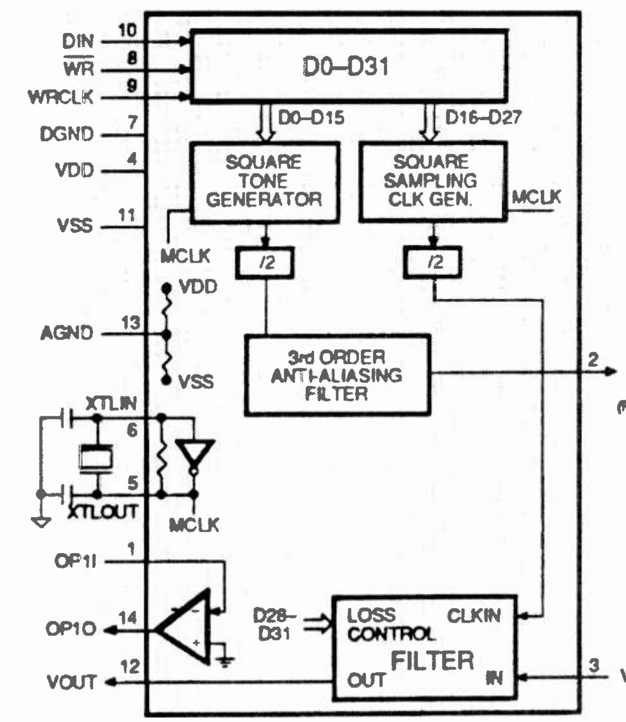


program button is pressed, the Q output will go low and the counter will run. The Q output of the *7474* also provides the /WR signal for the *SC11313*. After 31 clock pulses the D-type will be clocked forcing the Q output high and resetting the *4060* again.

The *SC11313* is connected to a standard 3.58MHz colour burst crystal. This provides the clock signal. The output sine wave signal is passed through the uncommitted op-amp connected as a 2nd order lowpass filter.

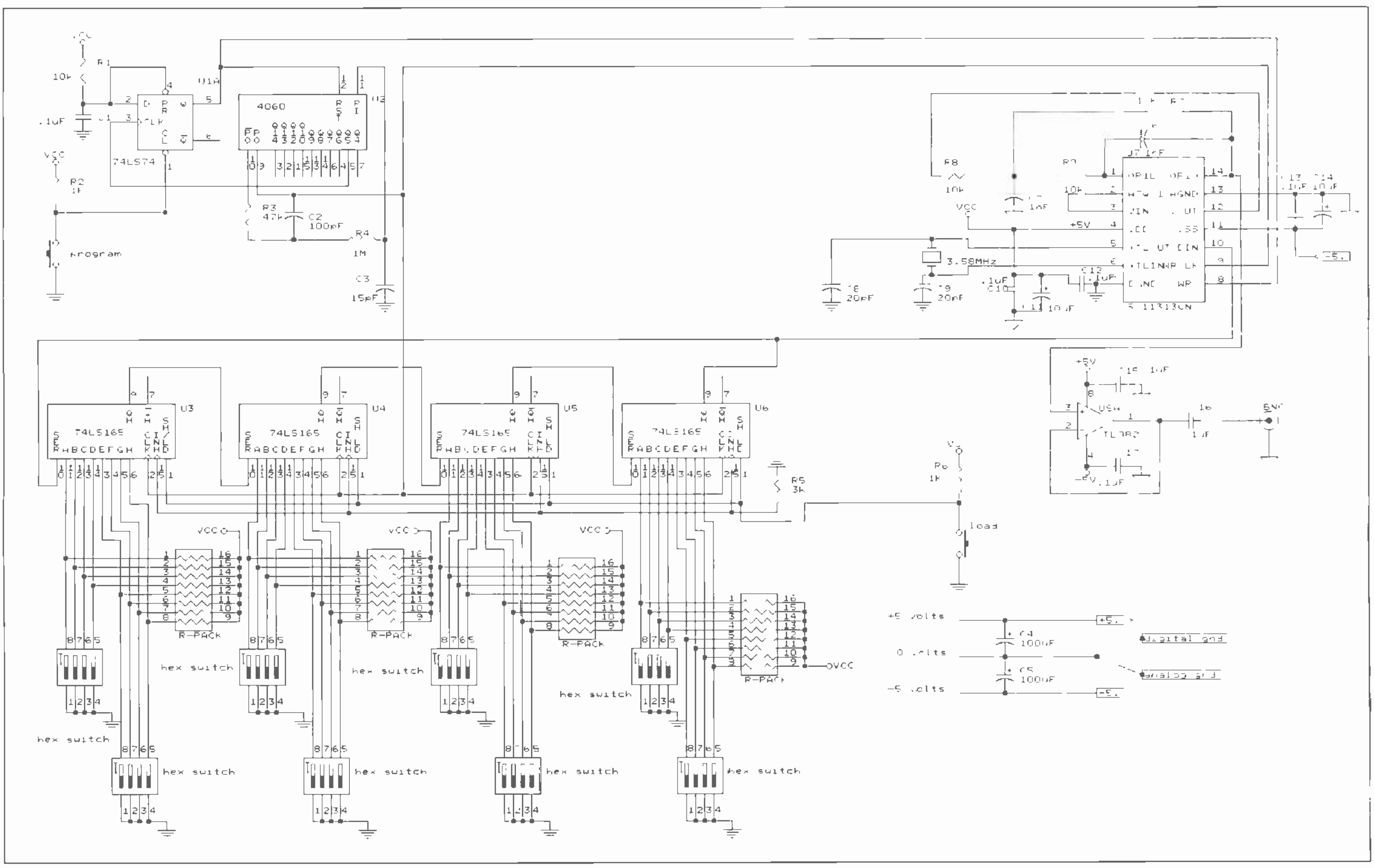
It is recommended that the power supplies for the digital and analogue sections are separated. For decoupling, 0.1μF on each of the digital circuits should be sufficient, however the analogue circuits should be de-coupled as shown in the schematic using both 0.1μF and 10μF capacitors.

Sierra Semiconductor Ltd, Terminal 3, 3B2, Stonehill Green, Westlea, Swindon, Wilts. SN5 7HB. Phone: 0793-618492



Typical application circuit. The extra logic shown in the diagram merely provides the means of serially programming the dividers in the SC11313.

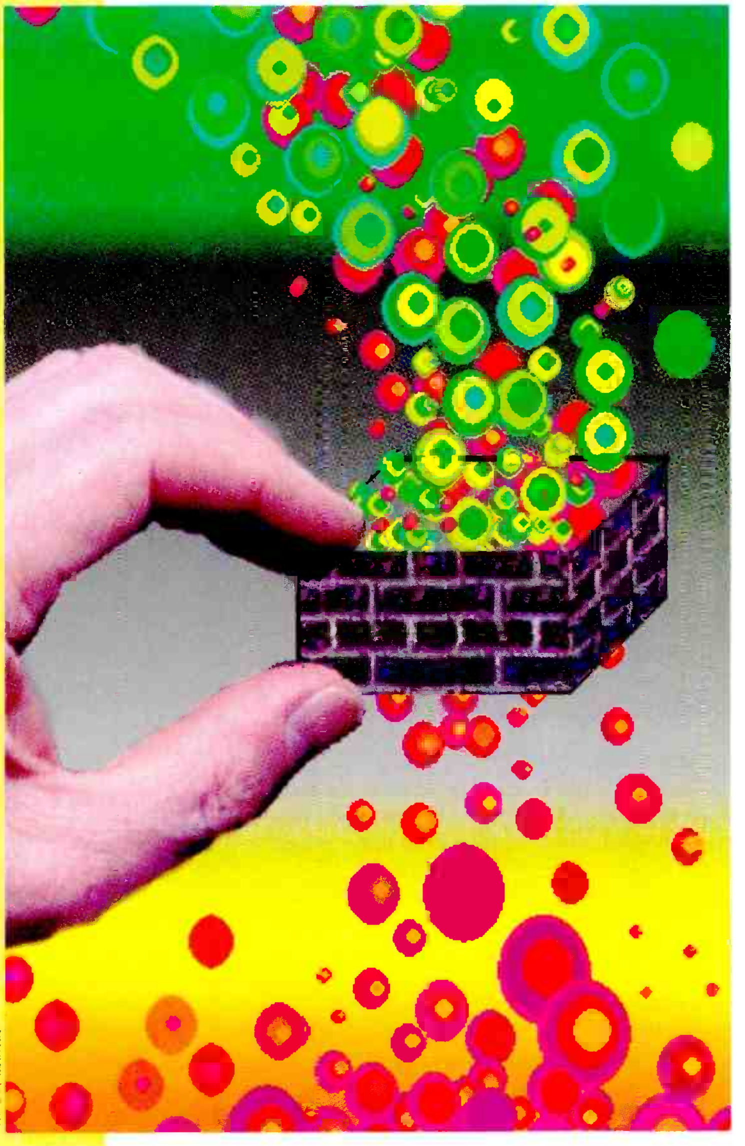
Block diagram of the SC11313CN. The frequency of the tone generator, determined by programmable division from a master clock, would normally be linked to the similarly programmed filter to produce sinewaves from the input square wave.



CIRCLE NO. 116 ON REPLY CARD

Building bricks into brick wall filters

Analogue to digital conversion normally calls for steep filters to remove alias components. The sharp cutoff characteristic is usually achieved by elliptic designs. The frequency dependent negative resistance elements normally found in low frequency elliptic filters can become too involved for a novice filter designer. Dr Bashir Al-Hashimi introduces a simplified approach to the design of elliptic filters.



TRACEY MARTIN

Analogue to digital conversion involves sampling a signal, a process that can accurately represent a continuous time varying signal with a bandwidth no higher than half the sampling frequency. This bandwidth limit is called the Nyquist frequency. For example, in a digital audio broadcast system with a sampling frequency of 32kHz, the Nyquist frequency is 16kHz. When frequencies above the Nyquist frequency are sampled, the sampled components "fold-back" below the Nyquist frequency. This means that signals beyond the Nyquist frequency generate unwanted signals within the passband.

These unwanted signals are commonly called alias components. An anti-aliasing lowpass filters is used to remove any signals that cause aliasing i.e band limiting the input signal to below Nyquist frequency.

Although anti-aliasing filters have different forms and characteristics, the scope of this article is limited to elliptic filters. Such filters can provide the near ideal lowpass characteristic, the brickwall response, which is needed to maximise bandwidth and minimise aliasing. The problem is shown graphically in **Fig. 1**.

There are a number of methods which can be used to design low frequency active elliptic filters. The frequency dependent negative resistance (FDNR) technique is the popular choice, since this approach, which was first introduced by Bruton[1], involves the simulation of the low sensitivity ladder LC filters. An investigation was made to find out if the FDNR design process could be simplified.

Conventional FDNR lowpass filter design

This usually involves the following steps:

1) Choosing a normalised minimum inductor LC filter that meets the required specifications;

2) Transforming the LC filter to its dual circuit;

3) Frequency and impedance scaling;

4) FDNR transformation and magnitude scaling.

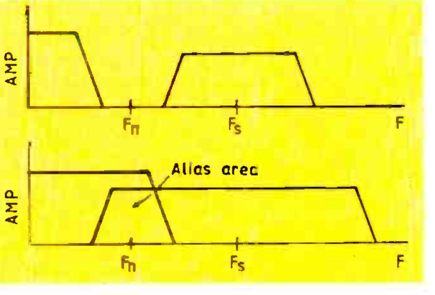


Fig. 1. When frequencies above the Nyquist frequency (the value of which is half the sample frequency) are sampled, the sampled components "fold-back" below the Nyquist frequency. This means that signals beyond the Nyquist frequency generate unwanted signals within the passband. These unwanted signals are commonly called alias components. An anti-aliasing lowpass filter is used to remove any signals that cause aliasing i.e band limiting the input signal to below Nyquist frequency. The area of spectrum occupied by alias frequencies is equal to ± the bandwidth of the sampled signal about the sampling frequency. An anti-alias filter must remove these.

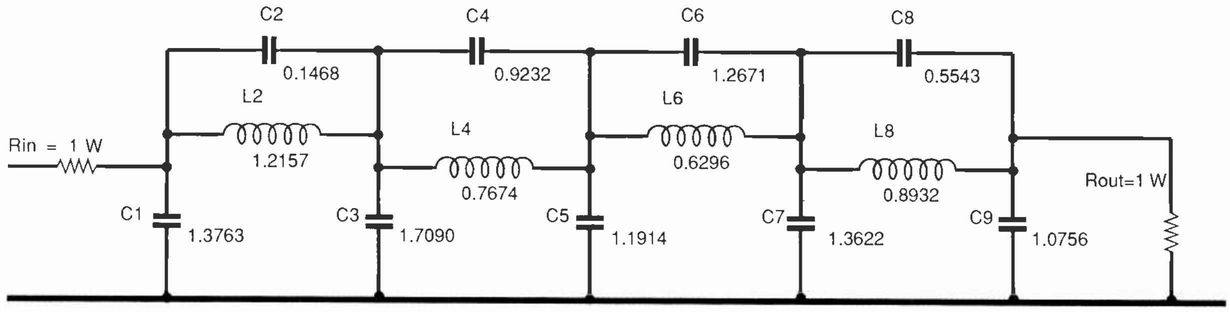


Fig. 2 shows the normalised required filter in LC form used as the model for an active version created with frequency dependent negative resistors. The LC version is easy to scale. The author claims the same facility with the active implementation.

Performing the above steps usually proves time consuming and error prone unless computer programs are used. However, such programs are either not easily or cheaply available, or are too involved for a novice filter designer. It was found that, while step 1 is essential, steps 2,3 and 4 can be minimised.

New design method

The new simplified design method starts by choosing the required LC lowpass filter. This filter can be obtained from filter tables which are readily available in text books[2,3].

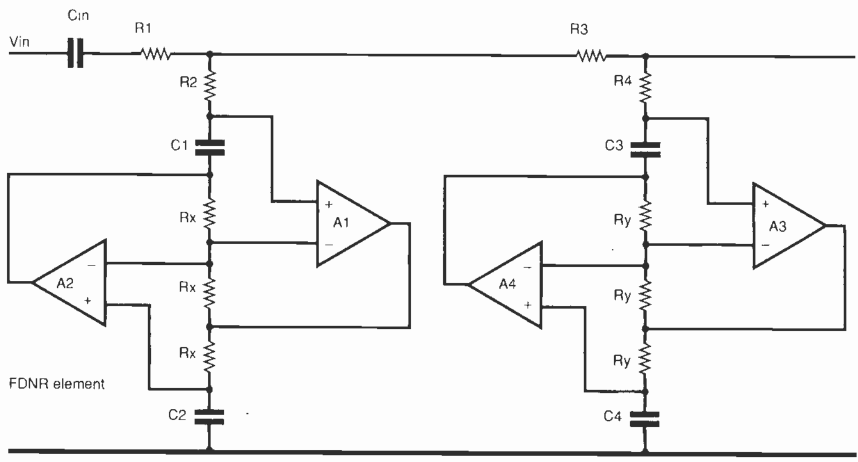
Consider as an example, a filter with the following specifications:

Passband ripple <1dB
Passband edge= 3.6kHz
Stop band edge= 4kHz
Stop band attenuation >60dB

The tables[2] indicate that a 9th-order filter will be needed. **Fig. 2** shows the normalised required filter. It is equally terminated (i.e. R_{in}, R_{out}=1Ω) and is chosen in preference to the singly terminated filter (i.e R_{in}=0Ω, R_{out}=1Ω) since the former has lower sensitivity to component variations, essential when designing filters to tight specifications. The filter of Fig. 2 is normalised to cutoff frequency of one rad/s and impedance of 1Ω. To obtain filters with practical component values and the required cutoff frequency, a frequency and impedance scaling must be used. This represents the next step of the new design method.

Table 1. 3.6kHz LC filter, all capacitors in nF and inductors in Henries

$R_{in} = R_{out} = 10k\Omega$	
C_1	6.085
C_2	0.649
C_3	7.555
C_4	4.081
C_5	5.267
C_6	5.602
C_7	6.022
C_8	2.451
C_9	4.755
L_2	0.538
L_4	0.339
L_6	0.278
L_8	0.395



Cin = Cout = 10n
C1 = C2 = C3 - - - - C8 = 10n
R1 = 6.085k, R2 = 0.649k, R3 = 7.555k, R4 = 4.082k
R5 = 5.267k, R6 = 5.602k, R7 = 6.022k, R8 = 2.45k
R9 = 4.755k
Rx = 5.38k, Ry = 3.39k, Rz = 2.78k, Rw = 3.95k

The scaling is done using the equation:

$$R' = R \times Z$$

$$L' = \frac{L \times Z}{2\pi F_c}$$

$$C' = \frac{C}{2\pi F_c Z}$$

where Z is the impedance scalar, F_c is the required cutoff frequency, and the prime values are both frequency and impedance scaled.

In our example, F_c=3.6kHz and Z (which can be set arbitrarily) was chosen to be 10kΩ, because this value of Z leads to capacitors in

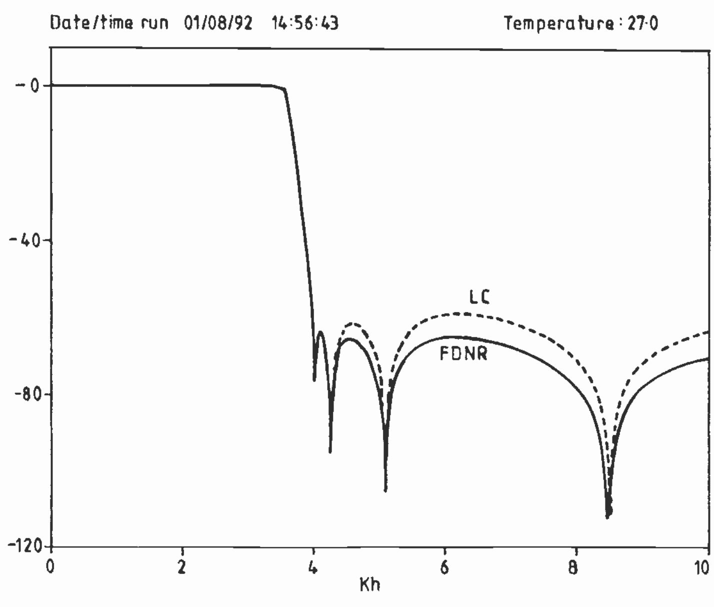


Fig. 4. Pspice simulation of the 3.6kHz filter implemented in both its LC and FDNR forms for comparison.

the FDNR filters with preferred values. Using equation 1, the denormalised filter values are given in **Table 1**. Once the denormalised LC filter is obtained, it is now possible to go directly to the FDNR filter as shown in **Fig. 3**, without the need to go through the FDNR transformation and the magnitude scaling (step 4 of the conventional method).

The resistors R_{in} and R_{out} of Fig. 2 are changed to the capacitors C_{in} and C_{out} of Fig. 3 respectively, the capacitors (C_1 - C_9) are changed to the resistors (R_1 - R_9), and the inductors are changed to the FDNR elements. The FDNR element consists of two op-amps, three equal resistors and equal two capacitors (see **Fig. 3**). The component values of the FDNR filter are found as follows:

Resistor values (R_1-R_9) are the capacitor values (C_1-C_9) in nF x 1(kΩ);

Resistor values of the FDNR element are the inductor values in Henries x 10(kΩ);

C_{in} and C_{out} are always 10nF. This capacitor value was chosen because of practical convenience and to yield filters with practical resistor values working at audio frequencies. For filters working above these frequencies, see reference [4] in relation to the choice of capacitors.

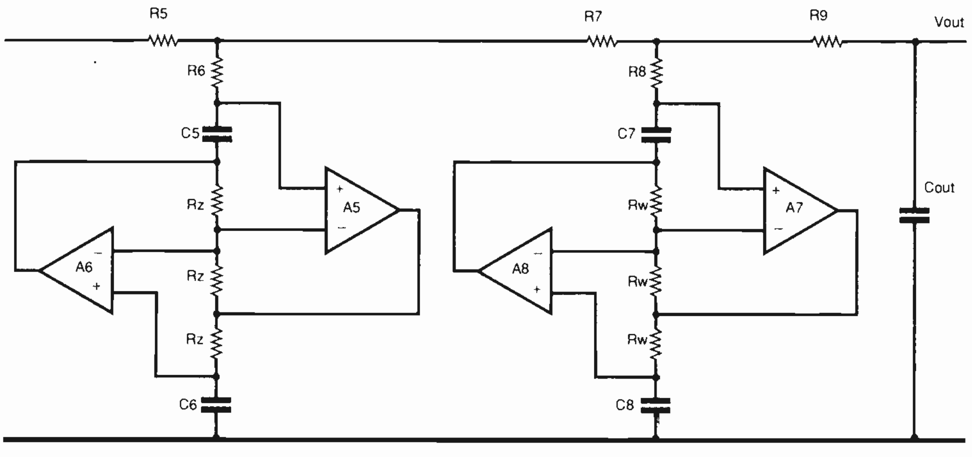


Fig. 3. 9th order FDNR 3.6kHz lowpass filter implementing the LC design of Fig. 2 in active form.

Table 2. FDNR filter table

LC filter	multiplier	FDNR filter
C_1-C_9(nF)	1	R_1-R_9 (kΩ)
R_{in}	Fixed	C_{in} (10nF)
R_{out}	Fixed	C_{out} (10nF)
L_n (Henry)	10	R_n* (kΩ)

*These are the FDNR elements resistor equal values, taken the capacitor values to be 10nF.

Capacitor values of the FDNR elements are taken to be 10nF.

Again this value is chosen because of practical convenience. However, if another capacitor value is preferred, then multiply all the FDNR element resistor values by the factor (old capacitor/new capacitor)^2, while leaving the other filter resistors unchanged (R_1-R_9), and C_{in}=C_{out}=10nF. These apparently arbitrary multipliers and component values have, in fact, been calculated or chosen to provide optimum designs in the range of 1kHz to 50kHz. We now have a FDNR look up table which can be used in a similar way to the LC filter look up tables. The FDNR filter table is given in **Table 2.**

Practical realisation

Examination of the circuit in Fig.3 shows that the non-inverting inputs of amplifiers A_1,A_3,A_5 and A_7 do not have a DC path to ground. Such a path is needed in order to supply the amplifiers with the required input bias current. A solution to this problem is to connect a resistor across each of the input and terminating capacitors (C_{in} and C_{out}). This resistor needs to be sufficiently large so as not to interfere with the filter performance: for audio filters 1MΩ is acceptable. Being an equally terminated filter a 6dB loss is incurred.

For practical circuits, input and output buffers can make up the 6dB loss if necessary. The FDNR filter of Fig. 3 can be built using dual op-amps (*TL072*) and 1% resistors and capacitors.

Figure 4 is a Pspice simulation of the LC and the 3.6kHz FDNR filters which verify the new design method. Models of the *TL072* op-amps supplied by Texas Instruments were used in the simulation. **Figure 5** shows the measured response of the 3.6kHz FDNR filter. The notch frequency of the *LC* filter is given by:

$$F_{\text{notch}} = \frac{1}{2\pi\sqrt{LC}} \qquad (2)$$

where *L* and *C* comprise the tuned circuit. The notch frequency of the FDNR filter is given by:

$$F_{\text{notch}} = \frac{1}{2\pi\sqrt{R_n RC^2}} \qquad (3)$$

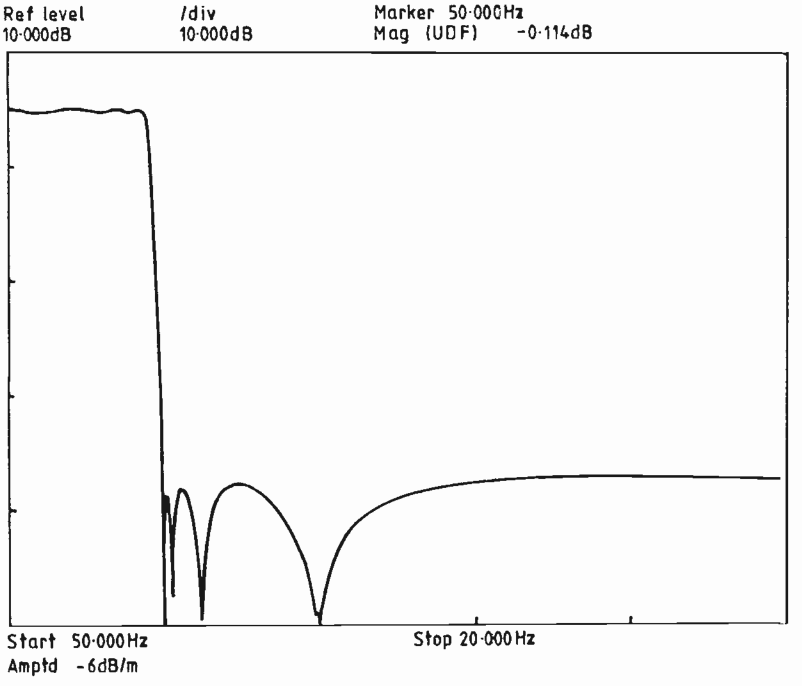


Fig. 5. Pspice simulation of LC characteristic against the active FDNR form together with the measured response of the active filter (Fig. 6).

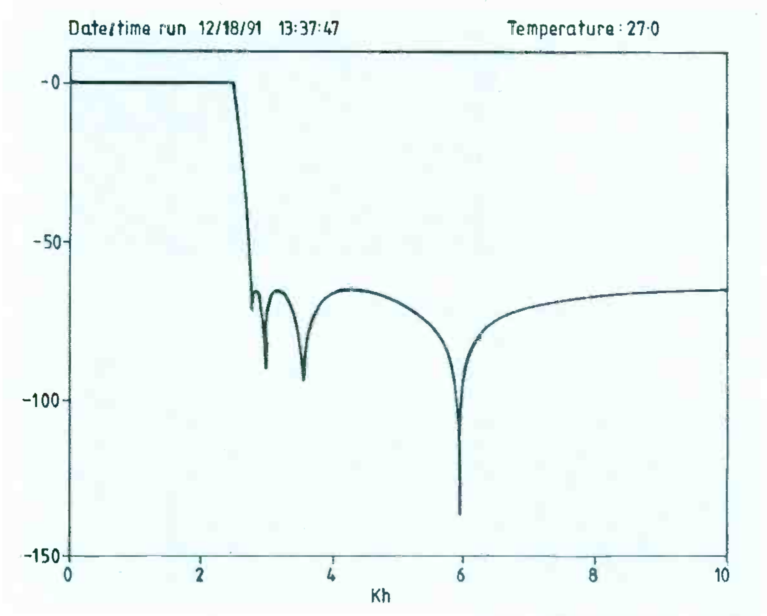


Fig. 6. Pspice simulation showing a calculated response of an FDNR filter scaled for 2.5kHz using the method described by the author.

Table 3 Notch frequencies of the 3.6kHz filter

LC Simulated	FDNR	Measured FDNR
F2(kHz) 8.517	8.376	8.20
F4(kHz) 4.279	4.266	4.251
F6(kHz) 4.033	4.024	4.022
F8(kHz) 5.115	5.110	5.110

where R_n is the series resistor with the FDNR element and C and R are the capacitors and the resistors of the FDNR element (Fig. 3). **Table 3** gives the notch frequencies of the LC parent filter together with the predicted and the measured 3.6kHz FDNR filters. This shows a good agreement between the theoretical and the practical response. The small difference in the notch frequencies between the practical and the simulated designs is explained in reference [5].

Filter scaling

Once a filter has been designed, it often needs scaling in frequency and passive filters do that easily. Traditionally, the same can't be said of scaling FDNR active filters. Using the new design method, scaling the active filter can be easily achieved by multiplying all the resistors – including the FDNR elements – by the factor $(F_c/F_c`)$, where $F_c$ is the existing cutoff frequency, and $F_c`$ is the new cutoff frequency; all existing capacitor values remain unchanged.

In our example, if we require to scale the filter to a cutoff frequency of 2.5kHz with the same characteristics, then we need to multiply all the resistor values by $(F_c/F_c`)=1.44$. **Figure 6** shows a Pspice simulation of this filter. The notch frequencies of this filter are found by dividing the notch frequencies of the 3.6kHz filter by 1.44. Experience and computer simulations have shown that common use op-amps such as the *TL072* or the *NE5532* can be used to design lowpass elliptic filters with cut-off frequency up to 20kHz. Beyond this frequency the open loop gain of such amplifiers is not high enough which is essential for the correct operation of the FDNR element[5]. ■

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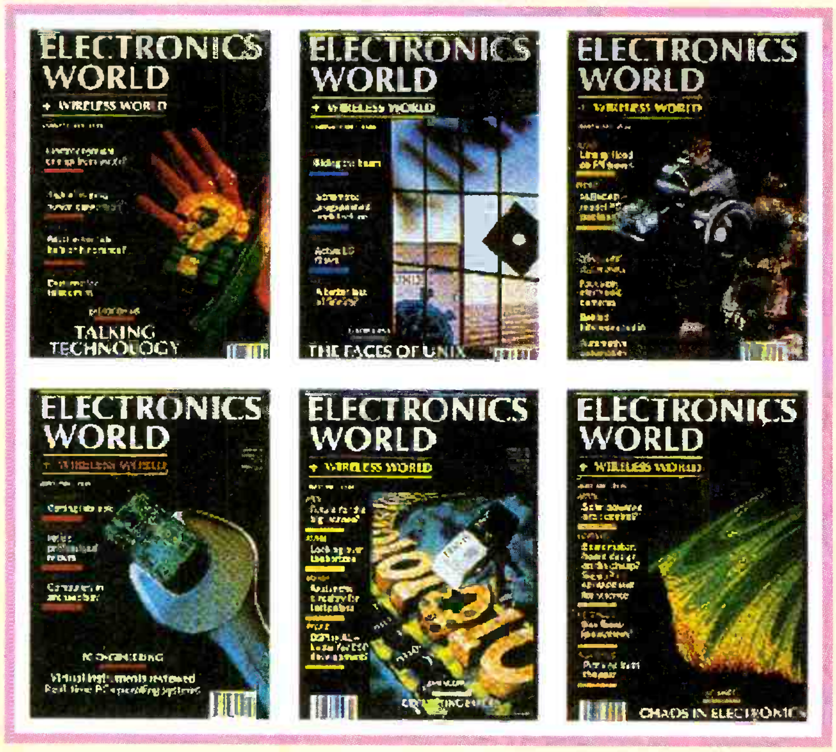
5. Al-Hashimi B.M. & Holden, A.G. *On the Practical Implementation of High Performance Active Audio Filters Using the FDNR Concept*, IEE Colloquium on Electronics Filters, London, May 1990

Dr Al-Hashimi works for Matthey Electronics as a filter design specialist.



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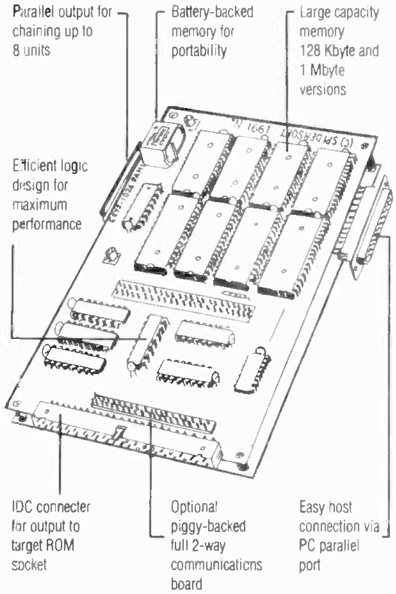

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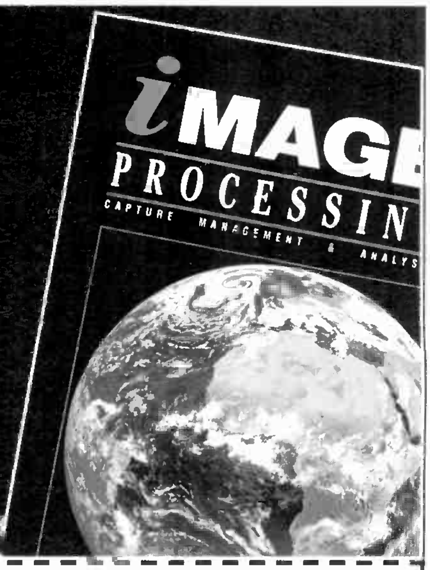

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Those who would audition their audio equipment for the effects of dielectric material in the coupling capacitors and oxygen in their speaker cables have long been ridiculed by the rest of us who merely use our hi-fi to listen to the music. Ultra-sensitive audio measuring instrumentation seems to provide evidence that the golden ears brigade might be right after all. Audio consultant Ben Duncan reports.

PROOF FOR THE GOLDEN EARS HYPOTHESIS?

Audio Precision, a US company founded by ex-Tektronix engineers, presently makes some of the world's most advanced audio test equipment.

Highly dependent on complex digital signal processing to produce its results, AP's recent software update has introduced challenging new tests for analogue audio, using DSP sampling at up to 192kHz. This can eke error signals out of the noise through steep but fast settling filtering with a degree of precision not previously possible.

One of the new tests enables individual harmonics up to the 10th to be plotted against frequency, down to some 0.00006% (60ppm). The other prods the DUT with a tone, typically 1kHz, which it cancels to <–130dB, and plots the resulting spectra. This shows both harmonics and intermod products. The sensitivity of the new tests is producing seemingly objective evidence of effects previously held to be utter rubbish by hard line objectivists.

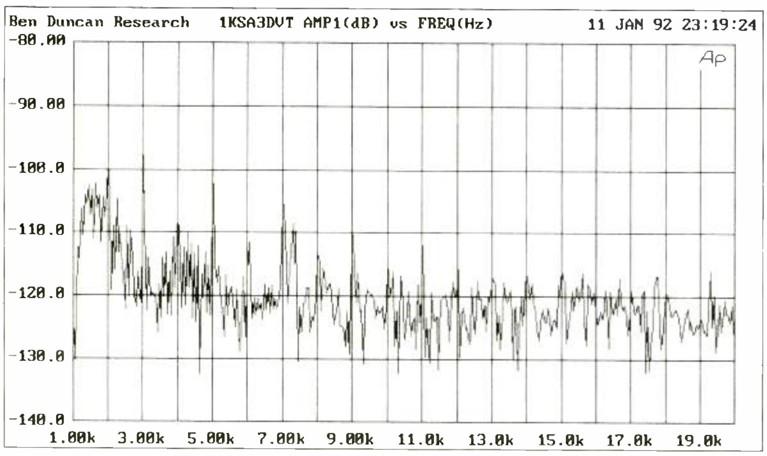

Fig. 1 Before the capacitor change. The output spectra of a Rauch DVT-50s professional power amp at 13dB below clip into 8Ω. Under these conditions, odd harmonics dominate.

Figure 1 shows the output spectra of a Rauch *DVT-50s* professional power amp at 13dB below clip into 8Ω. Under these conditions, odd harmonics dominate. In **Fig. 2**, the 100μF DC blocking capacitor in the grounding arm of the amplifier negative feedback loop has been changed for back-to-back elcaps totalling 165μF. The sonic benefits of doing this have long been recognised. All other conditions remain the same.

Figure. 2 clearly documents differences that corroborate the audible change. Looking carefully, one can see that the reversible elcap increases all the even harmonics up to the 8th, making them almost dominant. It also changes the residue so the odd harmonics above the 9th slope off monotonically. The full effect of

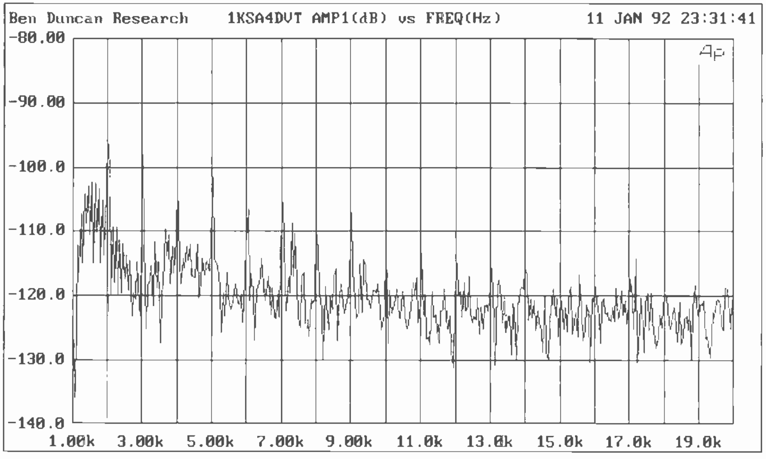

Fig. 2. And afterwards... the 100µF DC blocking capacitor in the grounding arm of the amplifier negative feedback loop has been changed for back-to-back elcaps totalling 165µF. Looking carefully, one can see that the reversible elcap increases all the even harmonics up to the 8th, making them almost dominant. It also changes the residue so the odd harmonics above the 9th slope off monotonically.

this change can be more fully appreciated by overlaying psychoacoustic weighting[1], but the principle is clear enough. **Figure 3** shows how a traditional THD+N vs frequency measurement misses the point; THD in the modified unit (upper plot) appears to be unchanged below 1kHz and slightly higher above 1kHz, leading to false conclusions. The new tests are so sensitive that the effects of changing and upgrading individual components can be seen.

Reference

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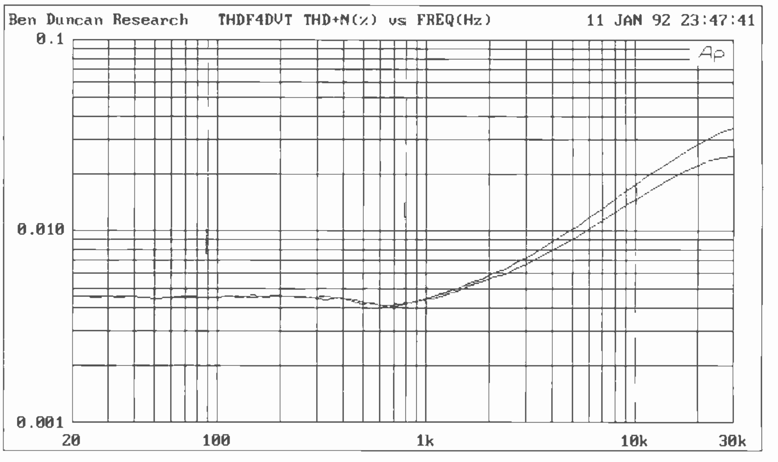


Fig. 3 shows how a traditional THD+N vs frequency measurement misses the point: THD in the modified unit (upper plot) appears to be unchanged below 1kHz and slightly higher above 1kHz. Even so, general opinion holds that the modified amplifier with the apparently higher THD levels sounds better.

Sorting out the plumbing with directional couplers

Which way is the power travelling along the line? Couplers will provide the answer. Dick Manton examines this most useful piece of RF plumbing.

Directional couplers are found in most transmission systems, where they act as samplers of transmitted power or give indications of mismatched loads or antennas. On a transmission line, a mismatched termination causes waves to be reflected back to the transmitter[1], resulting in a loss of available transmitter power and increased voltage on the line. But a directional coupler used to monitor forward and reverse powers can minimise losses and voltage problems.

Most amateur radio transmitters incorporate them in some form or another to ensure that maximum power is transmitted. At the other end of the power range, in high power-television transmitting stations, voltages derived from precision couplers are used to trigger executive circuits. The circuits ensure that economically-designed antennas do not become overloaded with power under transmitter fault conditions and they alert the operator if the antenna voltage reflection coefficient exceeds 1-2%. Greater levels of reflection coefficient than this may result in all viewers receiving perceptible ghost images on their screens.

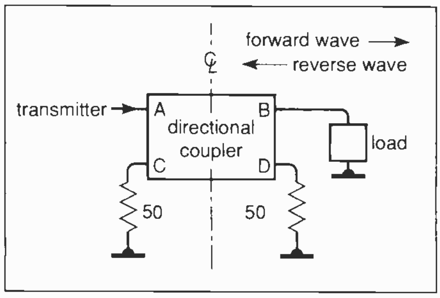


Fig. 1. Directional coupler showing the four ports.

The sampling systems

Directional couplers can be used for more than just monitoring power flow, and may take many forms – depending on frequency bands, power levels and required bandwidths.

But all wide-band versions operate by sampling the voltage V and current I (or the corresponding fields E and H) in a transmission line.

Equal co-phased currents I_E and I_H are then derived from these samples and are used so that, when power flows from A to B (**Fig. 2**), they add in load C and subtract load D. Reversing direction of power flow changes the direction or sign of either the voltage or the current (but not both). Derived currents then add in the load on port D and subtract the load on port C.

The current transformer in **Fig. 3** is suitable

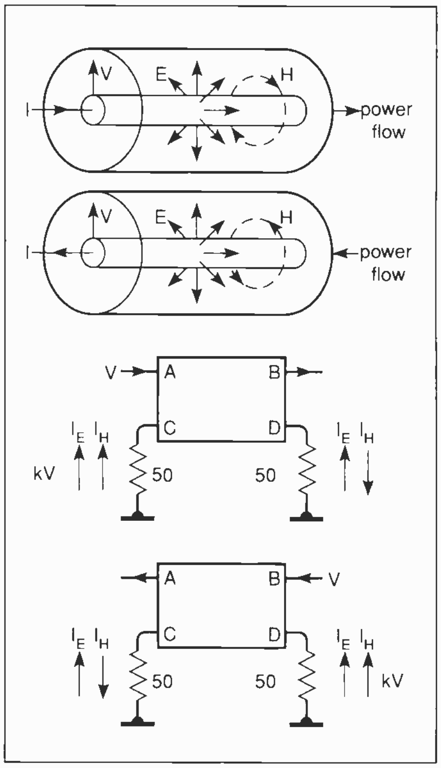


Fig. 2. Currents derived from the E and H fields within transmission lines rearranged to add or subtract in loads C and D.

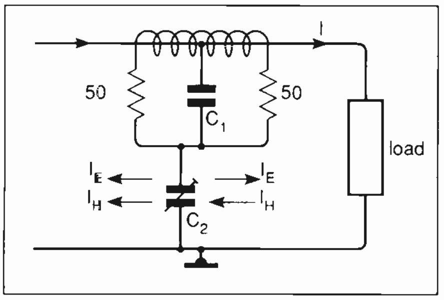


Fig. 3. Low-frequency directional coupler, using a current transformer.

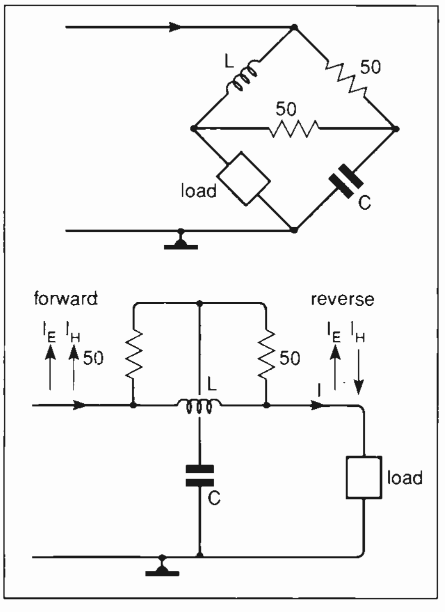


Fig. 4. Familiar form of the Maxwell bridge, and drawn as a directional coupler.

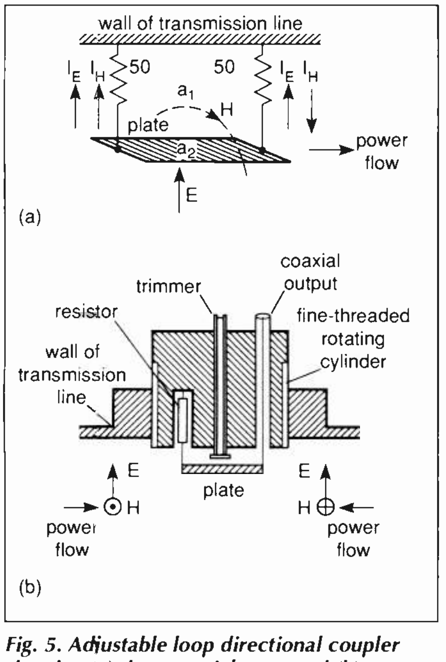


Fig. 5. Adjustable loop directional coupler showing (a) the essential parts and (b) practical form.

for use at frequencies below 3MHz. Current is sampled by a current transformer, and voltage is sampled by capacitor C_1. Loads R_C and R_D are the forward and reverse power loads respectively and a variable capacitor C_2 acts as a potential divider to allow the level of I_E to be adjusted. A further control may be needed to adjust relative phases.

Maxwell bridge

Figure 4 shows a Maxwell bridge in its usual configuration and then redrawn as a directional coupler. Voltage is again sampled by a capacitor C, and the voltage developed across inductor L is a measure of current.

The coupler is suitable for use at frequencies below 30MHz. In a 50Ω system the theoretical condition for perfect directivity and input impedance at all frequencies is that $L/C = 50^2$. Coupling factor k, which increases with frequency, is given by $k = 20\log_{10}(m/\sqrt{(1+m^2)})$ where $m = 1/100\pi fC$.

Adjustable loop

The essential parts of an adjustable loop coupler, suitable for examining the fields within large diameter coaxial transmission lines up to frequencies of 1000MHz, are shown in elevation in **Fig. 5a**; a practical version is shown in **Fig. 5b**.

A small metal plate, parallel to the screening wall of the transmission line, together with resistors R_C and R_D form a rectangular loop of area a_1, which couples with the magnetic field H. Electric field E couples into area a_2 of the plate. Suitable choice of areas a_1 and a_2 will result in the system having directional properties. The loop, supported on a cylinder (Fig. 5b), can be rotated on a fine screw thread to increase the E coupling by penetration and also to vary the H coupling by changing the angle of the loop relative to the direction of the H field.

Change of angle gives coarse adjustment of directivity and a trimming capacitor at the centre of the loop provides fine adjustment.

One load is built into the coupler, so that it is inaccessible, and a coaxial line is brought out to provide the coupled output. Separate couplers are required for forward and reverse indications.

The coupling factor of this type of directional coupler is rarely made greater than about –26dB (k = 0.05) because increased penetration would lead to significant mismatch on the line. k increases almost linearly with frequency. Subject to the built-in load having negligible reactance, directivity of the coupler remains good at all frequencies, once it has been set. (Both versions of UCL's directional loop antennas[1] are actually free-space directional couplers and bear close relationships to the above.)

Coupled transmission lines

Any two transmission lines, running parallel so that their E and H fields can interact, can be made to form a directional coupler. The easiest form to describe is the symmetrical arrangement, **Fig. 6a**, where two similar coaxial lines are made to share the same outer sheath over a portion of their length d.

The resulting directional coupler is useful at all frequencies between 3MHz and 1000MHz. Main features are:

• Coupled forward power is delivered to the coupled port closest to the input;

• Coupling factor increases almost linearly with frequency when the coupled electrical length d is very short compared with the wavelength λ_L within the coupled section.

Coupler components

A typical directional coupler has four ports *A*, *B*, *C*, and *D* (**Fig. 1**). Most power flows from the transmitter on port *A* to the load on port *B*. But a fraction of this forward travelling power is delivered to a matched load on port *C*, and no power is delivered to the load on port *D*.

The circuit is symmetrical about the vertical centre line, so the same fraction of any power flowing from *B* to *A* in the reverse direction is transferred to the matched load on port *D*.

For convenience, assume a 50Ω system and that a matched load implies a 50Ω resistance with zero reactance. Insertion of a perfect directional coupler into a transmission system will not disturb the impedance match of the system. By measuring voltages V_C and V_D across loads *C* and *D* the ratio of reverse-to-forward voltage or power can be calculated. If the coupling ratio is known, the actual forward and backward powers can also be calculated.

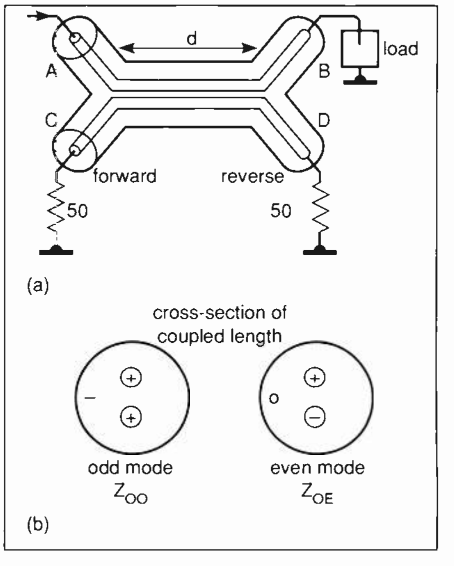


Fig. 6a. Coupled transmission line directional coupler; (b) cross-section of the coupled length driven in odd and even modes.

Definitions

Coupling ratio: Assuming all ports are appropriately matched (**Fig. 1**), the voltage coupling factor k is defined as V_C/V_A, usually expressed in dB as $20\log_{10}(V_C/V_A)$. It varies with frequency.

Directivity: In an imperfect directional coupler, even if there is no actual reverse power, a fraction of forward power may be delivered to the reverse power load on port *D*. Directivity, in dB, is defined as $20\log_{10}(V_D/V_C)$. Directivities of some types of directional couplers are fairly constant with frequency, but all are seriously affected by any mismatch or change of impedance of the load on port *C* (the negative sign of the coupling factor or directivity in dB is often omitted).

Sometimes the loads *C* and *D* on a directional coupler are replaced by appropriately matched detector circuits and meters to give indications of the levels of forward and reverse power, or circuitry is provided to take action if the power levels stray outside predetermined limits. The whole device is then referred to as a reflectometer.

Otherwise the coupling factor varies cyclically with frequency. The coupling is zero when electrical length is an integral number of half-wavelengths long and reaches a maximum when it is an odd number of quarter-wavelengths long. For optimum constancy of coupling factor with frequency, the coupler is usually designed to be a quarter-wavelength long at mid-band;

• Provided the relationship between cross-sectional dimensions obeys certain rules, the input match and directivity remain perfect at all frequencies, until these dimensions become comparable with a half wavelength.

The rules to be obeyed concern the characteristic impedances, Z_{OO} and Z_{OE}, of transmission lines having the cross-sections of the coupled section, but driven in either odd or even modes (**Fig. 6b**).

If k_{max} is the maximum coupling factor obtained when $d = \lambda_L/4$, the values of characteristic impedance for a 50Ω system are:

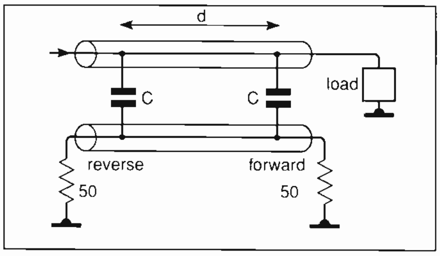


Fig. 7. A pair of capacitively-coupled transmission lines forming a narrow-band directional coupler.

$$Z_{OO} = 25\sqrt{\frac{1+k_{max}}{1-k_{max}}} \text{ and}$$

$$Z_{OI} = 100\sqrt{\frac{1-k_{max}}{1+k_{max}}}$$

ie $Z_{OO}Z_{OE} = 50^2$. Expression for the coupling factor at any frequency is:

$$k = \sqrt{\frac{k_{max}^2 \sin^2\theta}{1-k_{max}^2\cos^2\theta}}$$

where $\theta = 360°d/\lambda_L$ and λ_L = the wavelength within the transmission line. In any coupler much shorter than $\lambda_L/4$, k is almost proportional to frequency. When the required k_{max} is less than -30dB (0.032), an acceptable directional coupler can be formed by simply cutting away portions of the outer conductors of two coaxial transmission lines and then bonding the two lines together.

Capacitively coupled transmission lines

A simple directional coupler, operating reasonably well over a small band of frequencies, can be formed by connecting capacitors between the inner conductors of two lines at two separate points (**Fig. 7**). Expressions for the values of *C* and *d* are:

$d = \tan^{-1}m/360°$ line wavelengths

and

$C = 10^4/\pi fm$ pF

where

$$m = \sqrt{\frac{1-k^2}{k^2}}$$

and f = the frequency in MHz. A –20dB coupler of this type has a directivity better than –30dB and an input reflection coefficient better than 0.3% within a ±2% bandwidth.

Adjusting directivity

Directivity of a directional coupler should be set and fixed at manufacture – any subsequent unauthorised adjustments may require a complicated procedure to restore the directivity.

If directivity is only –20dB (10% voltage) and measured reflection coefficient is 25%, the real reflection coefficient could be anywhere between 5% and 35%. If a load is available to take the test power and is known to have a negligible reflection coefficient, adjustment of the directional coupler is simple; it can be adjusted to give a zero reverse power reading.

If the reflection coefficient of the test load is finite but unknown, it is still possible to eliminate it by using a swept-frequency method. A signal generator and the directional coupler should be connected to the load through a length of transmission line *d* which has the correct impedance and which is, preferably, several wavelengths long at the mid-band frequency of the coupler (**Fig. 8a**). Amplitude of the reverse voltage should then be measured as the frequency is swept through a band of frequencies which is greater than δf_0, where δf_0 = (150 x *velocity factor*)/*d* MHz.

Provided the reflection coefficient of the load is small and is practically constant over the swept range, the resulting plot of reflected voltage against frequency will generally complete a full cycle after δf_0 MHz. **Figure 8b** shows the resultant amplitude when the error introduced by the coupler's finite directivity adds vectorially to the reflection from the load as the latter changes its phase with frequency.

It is not immediately obvious whether the constant component or the ripple results from the reflection coefficient of the load. However, a slight change in the reflection coefficient, brought about by unscrewing a connector slightly or by changing the load, will resolve the ambiguity and the directivity can be adjusted.

A more advanced method of eliminating the reflection coefficient of the load makes use of a vector voltmeter. Two measurements are made at the same frequency, one through a short length of transmission line and the other through the same length plus an extra quarter-wave. In one case the two complex voltage vectors add and in the other they subtract. So it is relatively easy to separate them on a complex diagram. ■

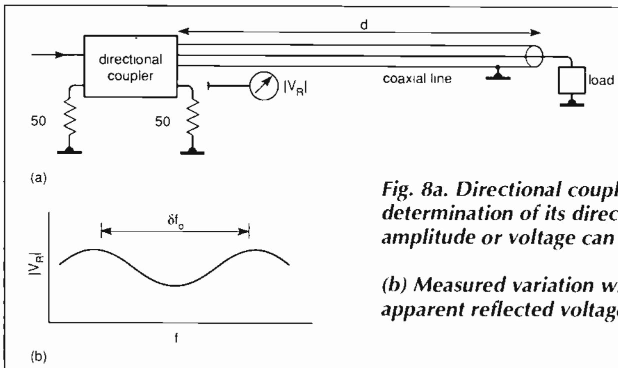


Fig. 8a. Directional coupler arranged for determination of its directivity when only amplitude or voltage can be measured;

(b) Measured variation with frequency of apparent reflected voltage .

References

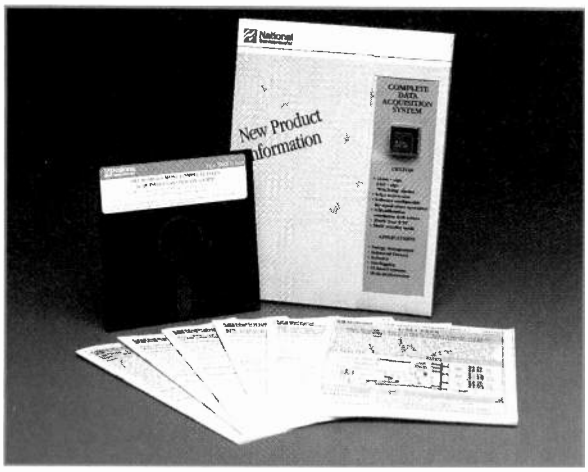
1. Manton, Dick. "*Impedance transformation with standard 50Ω coaxial feeder*", *EW + WW*, October, 1991, pp. 820-822.

Next month: Hybrids and couplers in power applications.

Connoisseurs of semiconductor data books will appreciate that data conversion products feature some of the most convoluted and arcane specifications to be found anywhere in the component world.

Easier access to the analogue world

Data acquisition sounds easy but most fully functional systems are highly complex. Until now, that is. Simon Prutton of National Semiconductor explains.



Digital processing, be it performed by a conventional microprocessor or DSP device, is often frustrated by the analogue nature of the real world. If the signal you want to measure is liable to change during the conversion then a sample & hold amplifier becomes essential; such changes can produce large errors if they are allowed to reach the ADC during a conversion cycle.

Anyone who has chosen a sample & hold in the past will know that chip specifications provide a fairly impressive list of definitions: droop, aperture time, aperture delay, etc... Next, to measure more than one signal, you are going to need a multiplexer – certainly simpler to select than the ADC and S&H but a significant step in the chain of data acquisition.

Finally the last analogue block required in any A/D system is a voltage reference. On top of the definition for these basic blocks, you may need to review the power supply board to see if you have generated all of the voltages needed by your ADC – and we haven't even begun to look at the host processor interface. It is not surprising that designers search for simpler solutions.

The commonest alternative is to buy a complete board that you can plug into the motherboard. In general, this approach reduces design time although such boards don't come cheap. What is required is an integrated solution, putting all functions on a single chip.

National Semiconductor calls this solution the *LM12458*, a complete 12-bit plus sign, monolithic data acquisition system running from a single +5V power supply.

A brief look at **Table 1** and the block-diagram in **Fig. 1** gives a fair idea of the devices's potential. The A/D at the core of the device uses an innovative self-calibrating architecture which, unlike most 12-bit ADCs, means 12-bit performance can be guaranteed: the *LM12458* is specified over –40°C to +85°C as standard.

The incorporation of various smart features have minimised the need for the host-processor intervention. For instance, the internal multiplexer can be configured either as eight single-ended inputs, four differential inputs or any combination of the two. Another example is the way in which the user can change the resolution of the output from full 12-bit to 8-bit in order to optimise the system throughput. Most importantly, these facilities and just about everything else on the chip can be programmed in real time on the fly.

However, the real power resides in the digital circuits. The *LM12458* incorporates an 8 x 48-bit instruction RAM which may be programmed to specify which channels should be accessed, whether to do a 12- or 8-bit conversion or a simple comparison versus two user-set limits (watchdog mode - see **Fig. 2**) and how fast to do this conversion.

A 16-bit programmable timer and sequencer carry out the instructions, and a programmable length fifo (32 word maximum) stores the output data until the user wishes to access it

Table 1. Potential of the LM12458

12-bit + sign Self Calibrating Core		
Operating Modes:		*12-bit + sign (13.5us max)* *8-bit + sign (4.2us max)* *Watchdog Mode (2.2us max)*
Accuracy/Linearity typical Unadj' Error		*±1 LSB*
8-Channel Mux with Differential Input Capability		
Built-in Sample & Hold and Voltage Reference		
On-board 8-Instruction Set-Up RAM & 16-bit Timer		
On-Board 32-word FIFO for Conversion Results		
Power Dissipation from +5V	*Active* *Sleep*	*30mW* *50uW*

Inst	Get Inst	Check Cal & Pause	Run Acq' Delay & Get Limit 1	Perform 1st Comparison	Get Limit 2	Perform Conversion or 2nd Comparison	Total Cycles	Total Time
1	1	1	9			44	55	11.0
2	1	1	9			44	55	11.0
3	1	1	9			44	55	11.0
4	1	1	2			21	25	5.0
5	1	1	2	1	1	5	15	3.0
6	1	1	3			21	26	5.2
7	1	1	3	1	1	5	16	3.2
						TOTALS:	247	49.4us

Also need to allow time to transfer data from the FIFO. 30 read cycles are required which adds 5 clock cycles per instruction set (assuming a 200ns CS).

Therefore, TOTAL CYCLES REQUIRED PER INSTRUCTION SET = 252 cycle
= 50.4us at 5MHz

Therefore, to meet the 50Hz throughput, the delay between finishing one execution of the instruction set and starting another must be 20ms - 50.4us = 19.9496 ms

1 unit of timer delay = 6.4us
Therefore, timer coefficient should be 19.9496ms/6.4us = 3117.

This gives a 50.002 Hz throughput.

Listed above are the number of clock-cycles required by the LM12458's sequencer to carry out all of the required instructions. The first two steps involve retrieving the relevant instruction from the on-board RAM and then checking to see if a re-calibration has been called for. Note that the next step (the programmed setting for the acquisition delay) has been incremented by 1 cycle for the high impedance sensor which is accessed in instructions 6 and 7 - the assumption being that the clock is running at 5MHz and therefore one clock cycle = 200ns as required in the specification. These steps can be facilitated through the use of the free design software discussed in the text.

Table 2. Sequencer steps, starting with an internal flag check

through the microprocessor interface. All of this plus the self-calibration and other features helps to reduce design and debug time through the effective integration of components. The whole thing fits into a 44-pin PLCC running from a single 5V supply. The device dissipates 30mW when fully active.

A practical application

To get a better understanding of how the *LM12458* works it is worth looking at a design example. **Figure 3** shows a furnace which requires precise temperature profiling and control to function correctly. Three temperature sensors are included – *T1*, *T2* and *T3* – and these are required to give temperature feedback with full 12-bit resolution. Two further sensors are required (this time just to 8-bit precision) to provide feedback of pressure (P) and gas-flow (F) through the furnace but since these parameters can cause a major system failure, an alarm function should also be implemented to indicate critically high or low values.

Additional design criteria include a throughput on all sensors of approximately 50Hz i.e. all sensors should be checked once every 20ms; the sensors should be digitised as close to simultaneously as is possible; the gas-flow sensor has a high source impedance and so requires an additional 200ns acquisition time before conversion. **Figure 4** shows five of the *LM12458*'s inputs in use to convert the analogue sensor outputs to digital information; seven of the chip's instructions are used to sequence through each measurement in turn – including the two watchdog commands to indicate alarm conditions. Given that five results have to be stored each time the instruction set is executed, the 32 word FIFO can be written to six times before it needs to send data out to the host processor. The FIFO interrupt register indicates to the host when 30 conversion results have been stored there.

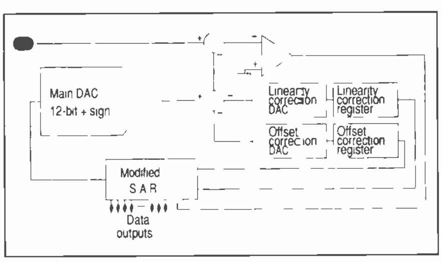
Table 2 shows each step taken by the sequencer in this particular application – starting with a check of some internal flags and culminating in either a conversion stored in

The ins and outs of self calibrating ADCs

Most 12-bit A/D converters are laser-trimmed during the manufacturing process to ensure sufficient accuracy. Unfortunately trimming is only done at room temperature – so the accuracy and linearity only holds for that temperature.

National Semiconductor sees the solution in a self-calibrating ADC. In this type of converter, extra circuits are integrated onto the chip to calibrate the ADC, the advantage being that the converter can be recalibrated during normal operation. Thus, if a large ambient temperature change is encountered, the host system can request the A/D to re-calibrate itself to compensate for any temperature induced changes.

The diagram displays a simplified representation of the *LM12458*'s core – basically a variation of the work-horse A/D architecture known as successive approximation. The main source of offset errors in a successive approximation ADC is its comparator: the first half of the self-calibration routine involves trimming this offset using the offset correction DAC which is driven by the modified successive approximation register. When the optimum value is determined, that data is stored in the offset correction register. Similarly, the linearity of the device can be trimmed and additional data stored in the linearity correction registers such that, when a 12-bit conversion is performed, the values stored in these registers are called-up to ensure that a 12-bit precise output is forthcoming.



The values in these registers may be changed at any time by issuing a re-calibrate command and ensuring 12-bit performance over the –40 to +85°C temperature range.

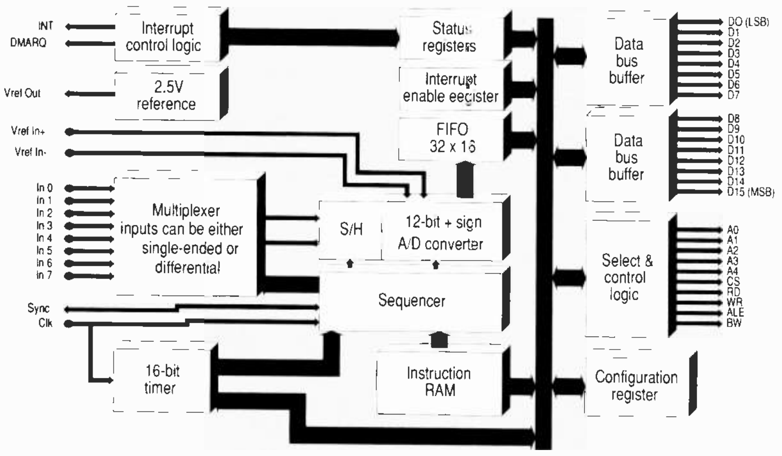


Fig. 1 gives an idea of the device's potential. The ADC at the core of the device uses an innovative self-calibrating architecture which, unlike most 12-bit ADCs, means 12-bit performance can be guaranteed.

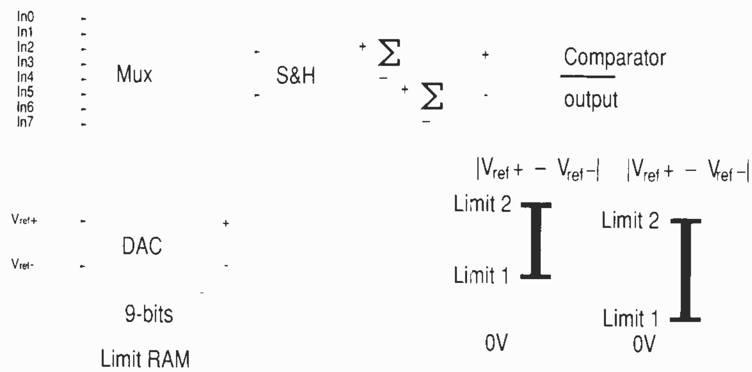


Fig. 2. The LM12458 can be programmed into watchdog mode. In this mode, the user can set eight programmable voltage windows (one for each input) and then set up the device to scan the inputs checking that the signal at each input remains within the defined window limits.

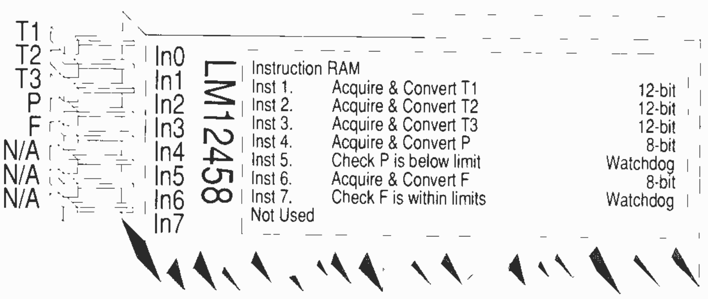
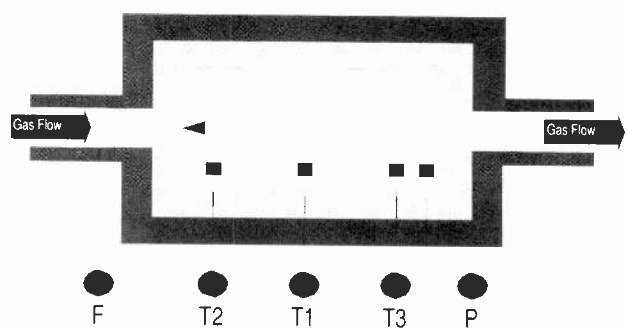


Fig. 4. Allocating the chip's resources in the furnace example. Five of the LM12458's inputs are used to convert the analogue sensor outputs to digital information; seven of the chip's instructions are used to sequence through each measurement in turn – including the two watchdog commands to indicate alarm conditions.

the FIFO or a limit check when operating in the watchdog mode. Without dwelling too much on the details of each step, suffice to say that the individual clock cycles can then be added up to calculate how many cycles are required to execute the whole sequence of instructions.

It is interesting to note how the different commands take a different number of clock cycles: the 12-bit + sign measurements need nine clock cycles to acquire the signal to the required precision while the 8-bit acquisition time is usually two clock cycles but can easily be extended to three clock cycles to acquire the signal from the sensor with a high source impedance. Also, since the device uses a successive approximation architecture to perform the conversions, an 8-bit conversion can be carried-out significantly faster than a 12-bit conversion. Finally, Table 2 also shows how the *LM12458* operates in its watchdog mode, retrieving the two user programmed limits from internal ram before comparing the incoming signal to each in turn. This provides a guarantee that neither of the defined alarm conditions has been violated.

As shown, the total number of clock-cycles required to loop around all seven instructions is 247 but we also have to allow for the data to be transferred from the internal FIFO by the host. 30 read cycles will be required which adds five clock cycles per instruction set (assuming a 200ns chip select and 5MHz system clock). Therefore the total number of clock cycles is increased to 252 which is equivalent to 50.4μs on the 5MHz clock. To



Required measurements

Absolute Temperature at T1	*12-bit + sign resolution*
Relative temperature at T2	*12-bit + sign resolution*
Relative Temperature at T3	*12-bit + sign resolution*
Pressure (P) in the Chamber	*8-bit + sign resolution*
Gas-Flow (F) through the Chamber	*8-bit + sign resolution*

Alarm Conditions

High or low Gas-Flow through the Chamber
High Pressure in the Chamber

Additional Design Criteria

Required throughput is 50Hz (=20ms execution time for whole loop)
Measurements to be taken as close to simultaneously as possible
Gas-flow sensor has a high output impedance so an additional 200ns acquisition time will be required before conversion.

Fig. 3. Single chip furnace control. This application requires precise temperature profiling and control to function correctly. Three temperature sensors are included – T1, T2 and T3 – and these are required to give temperature feedback with full 12-bit resolution. Two further sensors are required (this time just to 8-bit precision) to provide feedback of pressure (P) and gas-flow (F) through the furnace but since these parameters can cause a major system failure, an alarm function should also be implemented to indicate critically high or low values.

meet the 50Hz throughput target, a delay of 19.9496ms is required between the execution of each instruction set and the internal 16-bit timer can be used to generate this. One unit of timer delay is equivalent to 6.4µs so the timer coefficient should be:

19.9496ms/6.4µs = 3117

This value actually gives a throughput of 50.002 Hz.

Try it in your application

Most of these steps can be greatly simplified by the use of design software written by National Semiconductor to aid in the design of *LM12458*-based systems. A disk containing this programme will be despatched to you free of charge together with a sample of the *LM12458* if you complete the relevant post-card in the magazine. Additionally, should you to wish to further investigate this device, an evaluation board (*LM12458EVAL*) is available for purchase which, when plugged into an 8-bit slot in an IBM compatible personal computer, can be used to exercise the *LM12458* under software control.

The *LM12458* is huge step forward in data-acquisition but it is actually just the first of several in capturing analogue signals for processing in the digital domain. Watch this space for more developments. ■

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Alltech Stoddart receiver type 17/27A – .01-32Mc/s – £2500.
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Tektronix oscilloscopes type 2215A – 60Mc/s – c/w book & probe – £400.
Tektronix monitor type 604 – £100.
Wiltron 560 network scaler + 2 heads + book – £1000.
Marconi TF2330 or TF2330A wave analysers – £100–£150.
HP5006A Signature Analyser £250 + book.
HP10783A numeric display. £150.
HP239A oscillator – £250.
Alltech 7009 hot–cold standard noise generator.
HP 3763A error detector. £250.
Cushman CE-15 spectrum analyser – LED Readout – 1000Mc/s £650.
Tektronix 5L4N spectrum analyser – 0–100kc/s £500.
HP1742A 100Mc/s oscilloscope. £250.
HP1741A 100Mc/s oscilloscope. £250.
Tektronix 7104–7A29–7A24–7B15–7B10 – £2K.
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.
HP1815B T.D.R. sampler + 1817A head – 1104A trigger + 1106B TD mount £500.
Texscan AL-51A spectrum analyser – 4–1000Mc/s – £750.
Efratom Atomic FX standard FRT – FRK – .1–1–5–10Mc/s. £3K.
Muirhead fax receivers K649 – TR4 – solid state – speed – 60–90–120–240–auto – IC 288-576 – auto. £250 with book.
Racal 4D recorder – £350.
HP8350A sweep oscilloscope mainframe + HP11869A RF Pl adaptor – £2.5K.
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.
Tektronix 7L12 analyser – 1Mc/s–1.8GHz. £1500.
Clark Scam Heavy Duty 40' Telescopic Pneumatic Masts – retracted 7'8" – head load 40lbs – with or without supporting legs & erection kit – in bag + handbook – £200-£500.
Clark Scam Heavy Duty 70' Telescopic Pneumatic Masts – retracted 13'5" – head load 90lbs – with or without legs + erection kit + handbook – £500-£800.

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CIRCLE NO. 126 ON REPLY CARD

Real vocal point

We were fascinated to read Andy Wright's article "Analogue data storage: speaking of the future?" (*EW + WW*, Feb. 92).

The reason for our interest is that at PA Communications we have been building and selling PCM-based announcers since the mid 80s. Our systems are to be found in companies and authorities, from the Atomic Energy Authority to the RAF.

We are writing to take mild exception to the sentence used in the introduction to the article: "Phone callers on hold could be greeted with a real voice on the end of the line thanks to a new technique for storing analogue data".

We rather thought our systems (and those of our imitators) had been doing this without benefit of these fascinating devices for some years.

I suspect that it all depends on what one means by "real voice", but I venture to suggest that an *ISD1016* with its 35dB Sinad figure would not sound "more real" than a PCM-codec system – it all depends on the noise and distortion spectra.

This is in no way to decry the achievements of ISD - it seems a wonderful device, and well worth further investigation. But we continue to offer real voice announcers to the discerning, and yield to no others our right to claim this!

Ben Mullett
PA Communications
Milton Keynes

Playing chicken with power lines

Your provocative "Power politics: playing with lives?" (*EW + WW*, April 1992), raises two important questions. Since a little knowledge is a dangerous thing, why is more urgent and reliable research not being performed? Secondly, why do we the consumers, not try to help ourselves?. Can I not, figuratively, clad myself in metal armour to keep out harmful rays and radiations?

I can well imagine my younger and more agile neighbours placing looped and earthed galvanised iron sheeting in their attics. Similarly, new buildings could easily include their iron reinforcing rods as part of a suitably earthed and inter-connected iron mesh of galvanised wire, (chicken wire), in walls and roofs.

The use of twisted conductors in electrically shielded conduits would also improve our safety without adding to building costs.

Rather than confronting the utilities I would suggest a politically co-operative approach in dealing with non-ionising radiation. Thus industry, government and academe could all work well together to further the objectives of the metal interests together with our own. Yes, Minister?

Peter Hirschmann
Haifa
Israel

Power politic

Alasdair Philips' article "Power Politics:Playing with Children's Lives?" (*EW + WW*, April 1992) coincided with the publication of the NRPB Advisory Group on Non-ionising Radiation's report on "Electromagnetic Fields and the Risk of Cancer".

The group, which is under the chairmanship of Sir Richard Doll of the Imperial Cancer Fund Cancer Studies Unit, concluded: "The epidemiological findings that have been reviewed provide no firm evidence of the existence of a carcinogenic hazard, from exposure of paternal gonads, the fetus, children, or adults to the extremely low frequency electromagnetic fields that might be associated with residence near major sources of electricity supply, the use of electrical appliances, or work in the electrical, electronic and telecommunications industries."

Similar conclusions have been reached by the Science Advisory Board of the US Environmental Protection Agency.

I suggest that your readers should take note of the views of these expert and independent groups and discount the less informed interpretations of Philips' article.

D E Jeffers
The National Grid Company

Fluxgate magnetometer

Following Richard Noble's article "Fluxgate Magnetometry" (*EW + WW*, September 91) and our letter, December 91, Terry Arnold and myself wish to thank Noble and correspondent Gote Flodquist for their subsequent letters in February and March this year.

They point out that 47μT is the magnitude of the total vector for the Earth's magnetic field, inclined at 67° and that the horizontal component, as measured by our tangent magnetometer, would be 18.4μT (cos 47). This could account for the apparent "two times error".

I have since modified our tangent magnetometer by mounting the compass needle between a pair of bearings (escapement bearings salvaged from an old clock) so that the device can be inclined at 67° to north; about twice as much current is required to deflect the needle by 45° than when horizontal. So, taking imperfect balance of the needle, bearing friction and alignment errors into account, these results do agree with the comments made by Noble and Flodquist.

G Pickworth
Kettering

Model answer

I read Ben Duncan's article "A sound Model for Audio Design?" (EW + WW April 1992) with amazement. My impression is that Duncan knows quite a lot about the software he is using, but hasn't a clue about how to model a real circuit. At best, his models are wildly extravagant, and at worst they are simply wrong.

Fairly early in the article, Duncan states "... the performance (of emitter or source followers) is critically dependent on being hung from a low supply impedance".

Actually, emitter or source followers have excellent PSRR of the order of many tens of dB. So their supply impedance doesn't really matter as long as the collector or source voltage neither falls so low that the device saturates nor rises so high that over dissipation or second breakdown becomes a problem. It is thus necessary only to check the supply voltage at the trough of ripple waveform at minimum AC line voltage and maximum load, and at the peak of the ripple waveform at maximum AC and minimum load.

Next, Duncan tries to model his power supply, consisting of a mains transformer, bridge rectifier and reservoir capacitor. It is of course difficult to consider the circuit as having any meaningful small-signal impedance looking back at its output terminals; but without doubt the circuit he chooses to represent the power supply (his Fig.1) is simply wrong.

The transformer inductances (with series winding resistance and diode incremental resistance) do not appear across the output, in parallel with the reservoir capacitance.

If Fig. 1 were correct, the diode bridge resistance could be set to zero together with transformer winding resistance; at very low frequencies the "looking back" impedance would be inductive.

On the other hand, elementary circuit theory with the same perfect transformer and diodes suggests the output impedance to be resistive and of magnitude approximately $1/(4fC_{res})$ – the peak output always being $\sqrt{2}V_{RMS} - I_{out}/(2fC_{res})$.

But Mr Duncan, your Figs. 1 and 2 are wrong – you have a good computer but haven't used it properly. I note you are aware of your model's shortcomings. May I suggest that if you are worried about supply impedance at high frequencies, you take it to be that of your reservoir capacitor alone. This

will be true for much of the time, and will give a conservative figure for the duration of the diodes' conduction time. AC mains should also be modelled as having a source impedance equivalent to a series L-R network (250μH and 0.2Ω should do). Forget about parallel capacitance which vary so much from one place to another that it is pointless to give them values - fortunately their effect is negligible anyway.

Forget, too, about transformer magnetising inductance – it simply has no effect on output (unlike leakage inductance, which was not taken into account).

I can offer no constructive comments about the voltage regulator analysis, except that I think Duncan is wasting his time. It is very easy to set up three-terminal regulators like the *LM317* so that they work, and circuit optimisation is then best carried out by playing with the circuit itself which will always be better than any computer model.

I can also not see why anyone should want to simulate a short-circuit at the end of a 30m output cable. The stress on the output transistors will be far greater if the short occurs at the amplifier end of the cable– just as likely to happen in practice – so the cable can be neglected (which would have been done far more accurately as a transmission line rather than a three-stage RLC network).

I approve of the ability to see that the output devices operate always within their safe operating area – a really valuable use for the computer and I quite agree that simulation here can pay dividends. It is no use trying to prove a design for real if it keeps failing.

Finally I'm puzzled by the reference to beryllium dust. Neither beryllium nor beryllia (beryllium oxide ceramic) is used in any audio devices I have come across.

I think Duncan is trying to avoid a danger that just does not exist.

Peter F Vaughan
North Devon

Golden memories...

Alisdair Philips' article "Power Politics: Playing with lives?" (*EW + WW*, April) returns us to the killing fields about which you have devoted a good deal of space.

Whatever the merits of the case, I do protest at sloppy editing. Whether you wish to be persuasive, or merely informative, *EW + WW* readers deserve better than you gave to a topic that you obviously believe important.

Philips makes it hard to verify his sources. For example, his first reference is to Segal *et al*, but he does not quote the Journal, date or page numbers. There were other examples of incomplete citations and annoying inconsistencies.

It seems that *EW + WW* staff leave the office without looking at the final made-up copy. Please take a look at some early issues of *EW + WW*; they were not glossy, but were impeccably produced.

H W Shipton
Washington University

...fatally flawed

I was surprised to read (Letters, "Forty year delay" *EW + WW* April) about JM Chapman's vitalised valve-voltmeter, because I well remember the design.The balancing-circuitry was used in commercial instruments, such as the Marconi valve-voltmeter, which carried their quaint number: test-fixture *1041*.

It is unfortunate that the correction to the circuit, in *Wireless World*, February, 1951, p.53 - was missed.

T J Wynn
Newport

Sculpting the quantum world

I would like to comment on George Ho-Yow's response (Letters, *EW + WW*, April 1992) to Dr Millar's article "Scratching the surface of electromagnetism" (*EW + WW*, December 91).

The magnetic lines of force represent the space derivatives of the scalar quantity (gradient) of the magnetic field surrounding the bar magnet. Hence the lines do not "move" as suggested by Ho-Yow and in reality do not exist, but are analogous to the contour lines on an ordnance survey map. They represent the strength of the magnetic field at equipotential points.

But nobody understands the fundamental nature of magnetism, and the underlying reality is buried in the mathematics.

Kinetic energy of a body is defined as the capacity it has by virtue of its motion for doing work against a resistance. Or kinetic energy may be defined as the work which must be done on that body in bringing it from a state of rest to a velocity *v*.

In fact *energy* is an invention of scientists to allow the state of a system to be quantified. Whether the quantity of energy relates to energy levels in an atom (to use solid-state physics jargon) or the kinetic energy of an express train, is not important. The energy of a system is an abstract entity that does not exist in observable form. It is simply a measure of the ability of a system to do work. Therefore kinetic energy itself cannot be affected by gravity.

As regards the nature of an electron, the current perception is that it is *something* that sits outside of the nucleus possessing a certain amount of energy and some other properties. Electrons, as well as other particles, have an inherent wave particle duality which has resulted in one of the most fundamental mysteries of physics.

Returning to Ho-Yow's question on the nature of magnetism the accepted explanation regards permanent magnets as magnetic dipoles which exert a couple on each other. It is this couple that reveals the presence of a magnetic field. Many people find this explanation to be extremely unsatisfactory as regards the underlying reality of the magnetic force.

If all particles are considered to be steady-state structures in the quantum field then the fundamental forces of nature must inevitably be the result of a disturbance in the vacuum. Nobody has ever "seen" an atom and all we perceive are experimental results which are interpreted to be a nucleus surrounded by an electron cloud.

The strong nuclear force was hypothesised as the mechanism for binding the protons of the nucleus together against their natural tendency to repel one another. By using the vacuum curvature analogy it is much easier to picture the interlocking of vibrational nodes which are strong enough to overcome the dual repulsion effect. The atomic structure then becomes a group of interacting vacuum vibrations that exist as a stable entity, with electrons confined to null node points outside the nucleus.

In this alternative picture of quantum reality, all matter and all forces are the result of vacuum fluctuations which obey a fundamental set of rules that are only partially defined by the subsets of classical EM relativity and quantum theory.

The apparent creation or exchange of particles in accelerator experiments must then be due to interference between the standing wave structures we perceive as particles.

For example, production of a temporary standing wave structure in the vacuum caused by interaction of two electrons would apparently result in the exchange of a photon. In fact the photon is a fleeting interaction between two quantum field fluctuations.

Our experimental apparatus detects the temporary standing wave, caused by an overlap in the two interacting fields, and interprets this structure as a photon.

Clearly, if gravity can be described by geometric means as the bending of space-time (ie the vacuum) then magnetism, electric charge, etc may also be a localised manifestation of the same underlying mechanism, albeit of a different flavour and with different properties. Hence, the old ideas of action at a distance can more readily be rationalised with the notion of particles as carriers of the force. In this alternative approach the vacuum disturbance explains the presence of a force whereas the messenger particles of quantum theory are explained by interference between interacting stable vacuum structures ie particles.

If this theory is correct, then the accelerators of high energy physics are simply equivalent to a sculptor's chisel. The use of the same tools in an identical manner will always provide the same result consistently. Thus the plethora of particles that have been "discovered" in recent years may be of artificial origin, in as much as they represent a momentary interaction between two or more stable vacuum structures.

At present, modern physics can only offer the unsatisfactory conclusion that there is no underlying reality. The idea of vacuum fluctuations as the basis for all particles and fundamental forces provides a model of reality that can be more readily pictured than the abstract world of probability waves and wave functions. More importantly, it could prevent our headlong flight into accepting the "many worlds" interpretation of quantum physics and its inevitable links with Eastern mysticism.

George Overton
The Anti-Gravity Society
Hants

What a caution!

I was delighted that Jan Chapman (Letters, "Forty year delay", *EW + WW*, April) had retrieved from the attic the simple valve voltmeter described in *Wireless World*, December 1950. Correcting the value of R_{13} made it work at last in the 1990s.

With the wisdom of hindsight many of us now wish that we had waited for the corrections before building the latest circuit offerings from *EW + WW*. In the case of this voltmeter the corrected value of R_{13} was published on page 53 of the February 1951 issue.

Do I now build the AC magnetometer described in p.281-283 of the April issue – or would it be better to wait?

Roy Fursey
Dorset

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CIRCLE NO. 128 ON REPLY CARD

By summer next year the US should have selected a standard for high definition television (HDTV) broadcasting which will almost certainly be digital. What happens in the US could well affect European television standards. Aubrey Harris takes a good look at the competing systems

TRACEY MARTIN

PUTTING THE RIGHT NUMBERS INTO HDTV

Six systems have been submitted to the US's Advanced Television Test Centre for evaluation as a future HDTV standard. One is intended as an interim standard for extended definition only; one is based on transmitting signals in analogue form, and the remaining four use digital processing and digital transmission formats.

One of the main considerations is that the Federal Communications Commission has ruled that any system must operate, without interference, inside the existing American standard 6MHz-wide TV channels.

There are 68 such channels, twelve in the VHF band and 56 UHF, on which existing stations transmit in the NTSC format. In any one region, not all are occupied because interference considerations allow only every other VHF channel and every sixth UHF channel to be assigned. The new HDTV service will be carried on these "taboo" unused channels.

HDTV terminology

High definition television is a term generally used to describe systems able to transmit and display pictures with far higher detail and clarity than is possible with current TV broadcasting. Systems (**Tables 1** and **2**) may use a thousand or more scanning lines on the screen (possibly twice the number of a pal system), a wider screen and produce an improved signal-to-noise ratio with higher quality sound.

Higher resolution at the receiver is the prime aim, and this is achieved by increasing the active scanning lines, providing more resolution in the screen's vertical dimension, and using wider video bandwidth to match the improved resolution across the screen.

Factors determining transmission bandwidth are the picture aspect ratio (displayed image width to height), whether interlaced or progressive scanning is used, frame rate and the square of the number of lines in each frame.

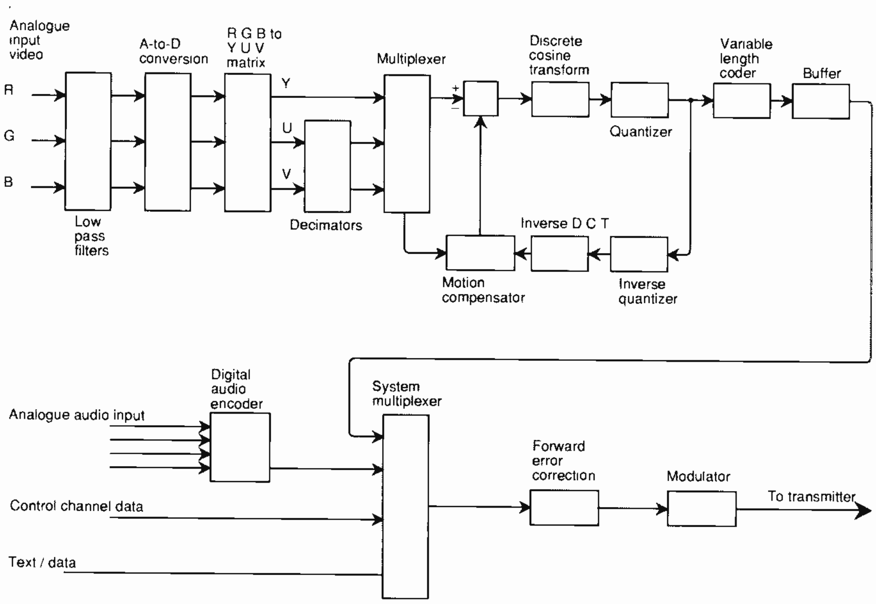


Fig. 1. Simplified block diagram, although specific for the Digicipher system, illustrates many elements common to other systems.

This last factor sets high bandwidth requirements for HDTV, since if the number of lines is doubled, the required bandwidth is quadrupled. Progressive, rather than interlaced, scanning also doubles the bandwidth requirement.

The present UK 625-line 50-field/s pal interlaced system with an aspect ratio of 4 by 3 requires a video bandwidth of 5.5MHz.

Doubling the number of lines would require a bandwidth of 22MHz, and conformance to the internationally agreed HDTV wide-picture aspect ratio of 16 by 9 (1.78:1) would bump that up to 39MHz. Progressive scanning would push the bandwidth up to 78MHz (almost ten times a single pal channel) and could not be accommodated in the present broadcast spectrum allocations.

Clearly bandwidth reduction techniques are needed. Broadcast TV uses interlaced scanning to reduce bandwidth, so that odd-numbered lines are transmitted in one field and even-numbered lines in the next, one twenty-fifth (or one thirtieth) of a second later. The criticism of interlacing is inter-line flicker and undesirable motion artifacts.

Bandwidth reduction

Exploitation of psycho-visual and statistical attributes of images, combined with various signal processing techniques, can significantly reduce the information/bandwidth needed.

For example, the eye does not need a high degree of colour information in an image as long as the full detail is presented by the brightness (luminance) signal. Pal, NTSC and secam use this phenomenon to reduce the transmitted chrominance bandwidth by 50% or more.

Detail is a main attraction of HDTV, and systems must be able to reproduce fine detail; but not usually over the whole of a frame. The portion of a typical – and even demanding – image requiring high frequency spatial response is usually small. In any case fine detail cannot be observed while an image is in motion.

In addition most TV images have a high degree of correlation between adjacent picture elements (pixels), both vertically and horizontally, and temporally between adjacent frames. Significant portions of pictures remain static over many frames and, even where motion is present, areas frequently remain unchanged for relatively lengthy periods: "talking head" pictures for example.

Digital electronics allows complex processing of signals (eg filtering) – cumbersome if not impossible in analogue form – and it is these advantages that make possible the compression of HDTV signals into the narrow channels available. Raw digitised HDTV data involves data rates of hundreds of megabits per second Compression reduces this by factors of 30-40. Combined with modulation techniques which allow multiple bits per hertz of bandwidth, the reduction makes terrestrial transmission of HDTV possible.

Digital proposals: Atva-Digicipher

Atva-Digicipher (**Fig. 1**) operates with an analogue RGB input at 1050-lines, interlaced 2:1, and a 59.94Hz field rate. Twice the NTSC line rate was chosen to allow easy conversion of baseband signals between the two systems. After digitising, compression and multiplexing

Table 1. Comparing technical characteristics of digital HDTV

	Digicipher	ADTV	DSC	Atva-P
Scan Lines/frame	1050	1050	787.5	787.5
Frames/s	29.97	29.97	59.94	59.94
Scan format	2:1 interlace	2:1 interlace	Progressive	Progressive
Line scanning frequency	31,468.5kHz	31,468.5kHz	47,203kHz	47,203kHz
Active luminance pixel format	1408 x 960	1440 x 960	1280 x 720	1280 x 720
Active chrominance pixel format	352 x 480	720 x 480	640 x 360	(unspecified)
Luminance bandwidth	21.5MHz	24.5MHz	34MHz	34MHz
Chrominance bandwidth	5.4MHz	12.25MHz	17MHz	34MHz
Video data rate	12.59/17.47 Mb/s	17.73 Mb/s	8.6/17.1 Mb/s	15.64 Mb/s
Sync data	-	-	292/544kb/s	-
Audio bandwidth	20kHz	23kHz	20kHz	20kHz
Audio sampling rate	48kHz	48kHz	47.203kHz	47.203kHz
Audio data rate	503kb/s	512kb/s	500kb/s	500kb/s
Auxiliary data channel	126kb/s	256kb/s	413kb/s	126kb/s
Error correction burden	6.17 Mb/s	4.96 Mb/s	1.3/2.4 Mb/s	3.042 Mb/s
Control data rate	126kb/s	40kb/s	40kb/s	126kb/s
Transmitted data rate	19.51/24.39 Mb/s	23.46 Mb/s	11.1/21.0 Mb/s	19.43 Mb/s

Note: All systems operate with SMPTE 240M colorimetry and with an aspect ratio of 16 by 9 (1.78:1). The Atva-P system is currently in design revision and some technical details are subject to modification.

Table 2. System names and proponents.

System Name	Proponents	Lines/frame	Scan format
Digicipher	American Television Alliance (ATVA) General Instrument Corp., Massachusetts Institute of Technology	1050	2:1 interlaced
Advanced digital television (ADTV)	American Television Research Consortium (ATRC) National Broadcasting Company, Philips, Thomson Consumer Electronics, David Sarnoff Research Center, Compression Laboratories Incorporated	1050	2:1 interlaced
Digital spectrum compatible (DSC-HDTV)	Zenith/AT&T Zenith Electronics Corp. and American Telegraph & Telephone Co.	787.5	Progressive
ATVA-progressive (ATVA-P)	American Television Alliance (ATVA) General Instrument Corp., Massachusetts Institute of Technology	787.5	Progressive

DIGITAL BENEFITS

Digital signalling formats for HDTV systems have many advantages:

- Regeneration of the digital bit stream at various processing points in the transmission path renders the signal virtually immune to degradations which plague current analogue TV systems.
- Multiple CD-quality sound channels are easily incorporated.
- Additional bit streams for text, data, and control can be interleaved with the vision and sound data.
- Multipath (ghost) images can be cancelled, and ignition noise and other impulse interference can be greatly reduced or eliminated at the receiver.
- Interfacing with future digital data communication and recording protocols should be easy.

the video data with four digital 20kHz audio channels, a data/text channel and a control channel, the data transmission rate is 24.39Mbit/s.

Analogue signals from programme origination equipment are low-pass filtered, A-to-D converted and fed through an RGB-to-YUV matrix to provide a luminance and two colour-difference signals. Resolution of the chrominance signals is reduced horizontally by a factor of four and vertically halved – the latter accomplished simply by discarding each alternate chrominance field.

Luminance and chrominance signals are sequentially multiplexed and then follow either a path to a motion-compensator stage, or after combination with the motion compensated product to a discrete cosine transform (DCT) stage. Coefficients in the DCT frequency array are significant for the DC (grey level) component or low frequencies, and can be low or zero for the higher frequencies in many images.

Coefficients of the frequency array are read out serially in a zig-zag fashion, so that they become arranged in ascending frequency and (usually) decreasing amplitude (**Fig. B**, see "Disrete cosine transform", p.494). "Run-length coding" assigns code words specifying the value of each non-zero coefficient and the number of zero value terms between it and the succeeding non-zero coefficient. The result is that a considerable reduction in data is needed to describe the majority of pixel blocks.

In a quantising process, weighting factors adapt the output under particular signal conditions to give priority to higher magnitude (low frequency) coefficients.

Motion estimation/compensation

The other path from the multiplexer goes to a motion-compensator for compression in the temporal domain.

In many television images the difference in picture content from frame to frame is small, particularly where there is little motion. Even with quite fast movements, where there is not a great deal of detail, frame-to-frame correlation is still high. So the difference in picture content between successive frames can be determined by comparing the current frame with the previous reconstituted (inverse-transformed and inverse-quantised) frame. A differential pulse code signal is generated and subtracted from the incoming current frame before DCT. Only the estimated differences between frames need be transform coded, rather than each individual frame.

The data rate is further reduced by assigning short code words to certain values that statistically occur more often than others in the data stream. Longer code words need only be used

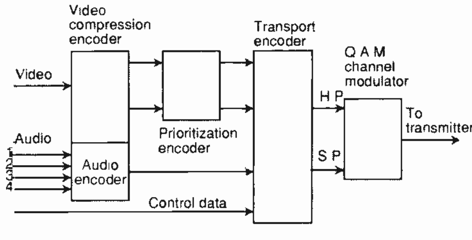


Fig. 2. In ADTV, approximately 20% of data goes into the HP stream, radiated by the transmitter at a higher power level.

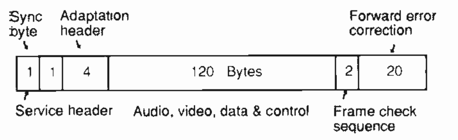


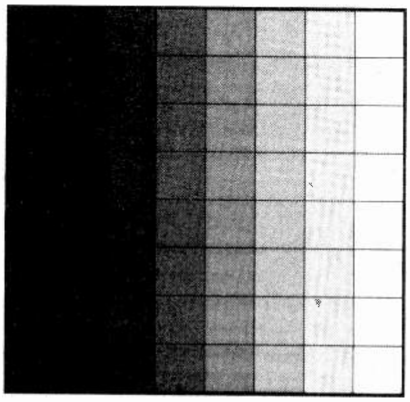
Fig. 3. ADTV's transport cell is similar to ATM data packets.

Why 29.97 and 59.94 ?

The surprisingly odd frame and field rates developed for NTSC in the early 1950s are also being proposed for US HDTV. They will probably remain because of the benefits of having a simple integer relationship between NTSC and HDTV.

Frequency of the NTSC sub-carrier was chosen, for minimum picture visibility, to be an odd-multiple of half of the then-current 15,750kHz monochrome line scan frequency. Development tests were made using a sub-carrier frequency of 3.583125MHz, the 455th-multiple of half the line rate. But at the time certain monochrome TV sets showed an objectionable beat between the sub-carrier and the 4.5MHz sound carrier. Experiment showed that making the beat signal frequency an odd-multiple of half the line scan rate would reduce visibility of the beat, so the sound carrier frequency should be an even-multiple of the line rate.

Line scan frequency was changed to 4.5MHz/286 = 15,734.264kHz; the sub-carrier frequency became 15,734.264 kHz x 0.5 x 455 = 3.579545MHz; and (still maintaining 525 lines per frame) the frame rate had to drop to 15,734.264kHz/525 = 29.97Hz. The field rate became 59.94Hz.



0	12.5	25	37.5	50	62.5	75	87.5
0	12.5	25	37.5	50	62.5	75	87.5
0	12.5	25	37.5	50	62.5	75	87.5
0	12.5	25	37.5	50	62.5	75	87.5
0	12.5	25	37.5	50	62.5	75	87.5
0	12.5	25	37.5	50	62.5	75	87.5
0	12.5	25	37.5	50	62.5	75	87.5
0	12.5	25	37.5	50	62.5	75	87.5

Grey scale DC value

43.5	-40	0	-4.1	0	-1.1	0	0
0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0
0	0	0	0	0	0	0	0

Fig. A. Brightness (left), amplitudes (middle) and frequency coefficients (right) for gradual increase in brightness. Horizontal frequency terms increase from left to right; vertical frequency terms increase from top to bottom.

DISCRETE COSINE TRANSFORM

Discrete cosine transform (DCT) is not in itself a compression. Its primary function is to convert an array – representing amplitude levels of a given number of spatial elements in an image area – to an array of the same number of frequency coefficients. Conversion forms the basis for subsequent reduction of the number of terms and amount of data necessary to define the image.

Analysing and dividing TV pictures into 64-pixel blocks often shows that very little high frequency information is contained within each block. Processing pictures in the frequency domain instead of the spatial domain allows advantage to be taken of this, and sometimes the low value HF components can be ignored.

Typically, spatial arrays of picture elements, representing intensity levels of portions of a video image in blocks eight pixels wide by eight pixels high, are sampled in turn and subject to DCT until the whole of the image frame has been transformed.

The transform turns the luminance or chrominance intensity block into a two-dimensional array of sixty-four frequency coefficients, representing amplitudes of the spatial frequency components of the original block (**Fig. A**). The zero-frequency (DC) term, defining the grey level of the whole image block, is placed at the upper left side of the block. Other coefficients relating to the increasing frequencies in the horizontal dimension of the spatial array are arranged from left to right in the transformed array. Those representing values of increasing vertical frequency are arranged from top to bottom.

Coefficients in the transformed (frequency) array are often read out serially in a zig-zag manner (**Fig. B**) so that, following the DC term, the second coefficient gives the amplitude of the lowest frequency in the horizontal dimension of the array. The next coefficient is the amplitude of the lowest frequency in the vertical dimension, and succeeding values reveal magnitudes of incrementally higher frequencies.

The final value, the 64th coefficient, indicates amplitude of the highest frequency in both the horizontal and vertical directions.

Dependent on the total number of pixels in the image, a 64-pixel block covers an area of the order of less than one twenty-thousandth of the full picture. For many TV pictures a large number of 8 x 8 pixel block images will be made up of areas of similar brightness or similar chrominance value and there is a high degree of horizontal and vertical correlation between adjacent pixels. For example, in pictures containing the human face, clothing or outdoor scenes, many transformed blocks are described completely by the DC level.

Other blocks contain low frequency coefficients and low magnitude higher frequency coefficients. High amplitude, high frequency terms only occur where there are marked, sudden changes in intensity in the televised image. When typical TV pictures are analysed, the number of areas in the image containing these kinds of transitions is surprisingly small. Therefore the number of data bits required to define such pixel blocks is drastically lower in the frequency domain than in the spatial domain.

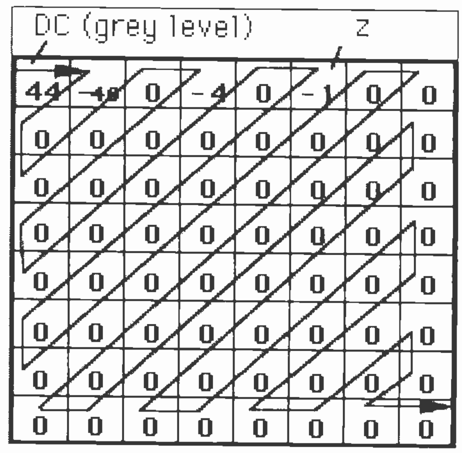
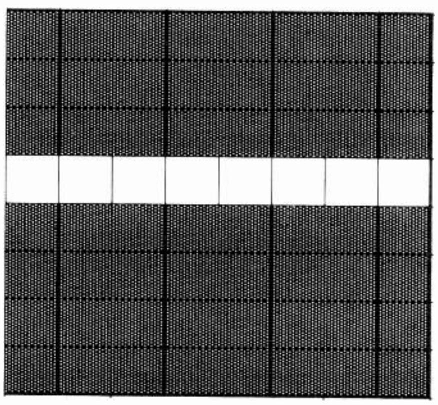


Fig. B. Zig-zag serial readout



10	10	10	10	10	10	10	10
10	10	10	10	10	10	10	10
10	10	10	10	10	10	10	10
100	100	100	100	100	100	100	100
10	10	10	10	10	10	10	10
10	10	10	10	10	10	10	10
10	10	10	10	10	10	10	10
10	10	10	10	10	10	10	10

42.5	0	0	0	0	0	0	0
6.2	0	0	0	0	0	0	0
-29	0	0	0	0	0	0	0
-18	0	0	0	0	0	0	0
22.5	0	0	0	0	0	0	0
26.5	0	0	0	0	0	0	0
-12	0	0	0	0	0	0	0
-31	0	0	0	0	0	0	0

Fig. C. DCT horizontal white line, and amplitudes and frequency coefficients.

for less frequent values. Huffman coding in a variable length coding stage (VLC) uses the minimum number of bits (compared to other methods) to represent given data.

Changes in TV pictures with time and the resulting response of the coding and quantisation processes, means the bit rate in the data stream, after the VLC, is not uniform. A first-in, first-out buffer regulates the varying data rate and provides a constant-output bit rate into the system multiplexer, where data streams from the audio, control and data/teletext sub-systems are appended.

Forward error correction minimises the effects of transmission channel errors on integrity of the received signal, and helps eliminate car ignition and other impulse noise pulses of up to 3μs duration.

The Digicipher quadrature amplitude modulation (qam) system operates at two different data rates. A lower mode (16-qam) at 19.51Mbit/s aims to provide error-free reception at a carrier-to-noise ratio (CNR) as low as 12.5dB. At the higher rate of 24.39Mbit/s (32-qam), higher video quality is received but a CNR of 16.5dB is required for error-free reception.

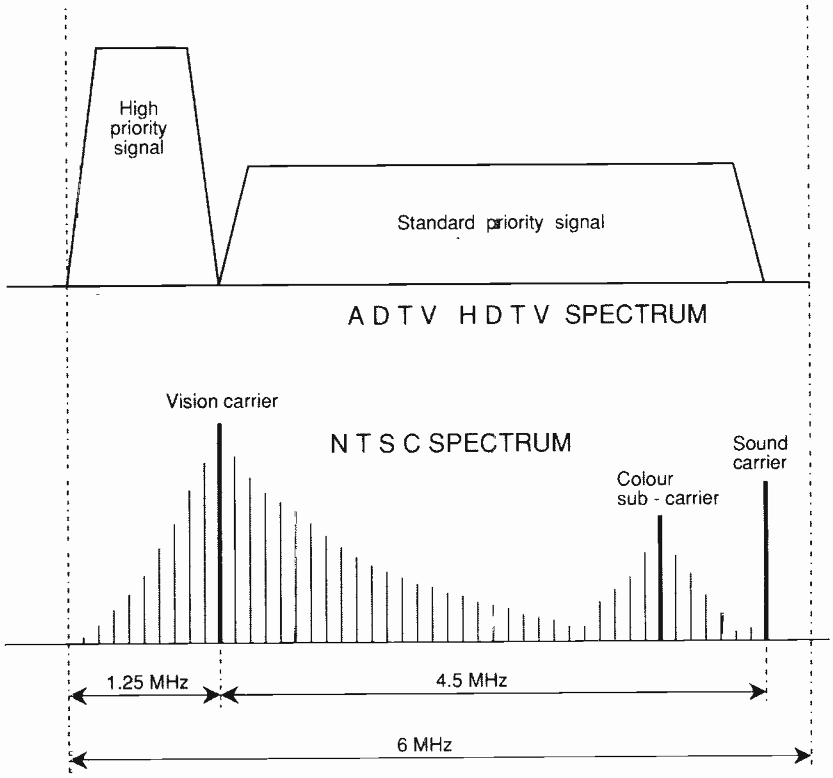


Fig. 4. Qam transmitter output filters and qam receivers have minimum response at the NTSC vision and sound carrier frequencies, reducing likelihood of mutual interference.

ADTV - advanced digital television

Baseband input for the ADTV system is 1050-lines, 2:1 interlaced, 59.94 fields/s. Source coding is based on the ISO MPEG (Moving Picture Experts' Group) draft specification for transport of moving images over communication data networks. ADTV has called its modified MPEG, handling the more stringent requirements of HDTV, "MPEG++".

Where the encoding system needs to process a lot of new data (in scenes containing much motion and/or fine detail) prioritisation separates information into data streams reflecting their importance to overall system operation. Data critical to the basic integrity of received pictures – grey scale levels, audio signals, data cell headers and motion descriptors – are assigned high priority (HP); low frequency coefficients and the higher frequency (fine detail) coefficients form the standard priority (SP) data stream. Assignment states are adaptive, and SP data may transcend to the HP stream when HP loading is light.

The two streams are formatted into separate 148-byte data transport cells (**Fig. 3**) similar in structure to asynchronous transfer mode (ATM) data packets on data communications networks. The ADTV cells are readily transcodeable for transmission on broadband integrated services digital networks (ISDN).

Both data streams are quadrature amplitude modulated onto separate carriers contained within a 6MHz band (**Fig. 4**). The HP channel is 960kHz wide; the SP channel occupies 3.84MHz and is filtered to have minimum power at the NTSC carrier frequencies.

ADTV HDTV receivers would have similar functioning filters so that a co-channel NTSC station would not interfere with HDTV reception.

HP would be radiated 5dB higher than the SP, so that where transmission is impaired with the CNR of the SP channel dropping below threshold but with the HP signal still above threshold, the received service would degrade gracefully. Quality would decline (less detail being temporarily displayed) but there would not be a complete loss of signal.

Digital spectrum compatible – DSC

Baseband/camera origination signals to the digital spectrum compatible (DSC) HDTV system operate at 787.5 lines/frame, 59.94 frames/s, progressively scanned. Line scan rate is 47,203Hz, three times NTSC, allowing easy conversion between the two systems.

Progressive scanning eliminates inter-line flicker and motion artifacts – but video bandwith is doubled.

Practicable video compression limits dictate that DSC must operate with only 787.5 lines (1.5 times NTSC) rather than doubling the NTSC rate to 1050 lines. DCT coding is used for spatial compression.

For transmission, compressed data are formulated into frames corresponding to NTSC time parameters. Coded data are embedded in two fields with field and segment sync, test data and parity check signals.

Data are formed into two streams. High priority data is radiated at higher power, providing a more robust received-signal and having a 7dB lower CNR threshold advantage over the low priority stream. The technique, similar to other proposed HDTV systems, counters a frequent criticism of digital transmission systems: that minor decreases in carrier signal level can cause error rates to rise abruptly to the noise threshold, giving little margin against a sudden loss of received signal.

A further claimed advantage is that DSC transmitter power is 14.5dB less than for an NTSC transmitter to cover a similar service area. Less risk of interference to other stations operating in the vicinity, both NTSC and HDTV, could be the result.

Atva-progressive

American Television Alliance's Atva-p operates with the same baseband standards as DSC. It is the only system out of the four proposals for digital HDTV not to use DCT for frequency derivation. Instead, following motion compensation, YUV components are individually analysed by a two-dimensional set of band-pass filters and down-sampled (sub-band coding). Resultant coefficients, relating to discrete fractions of the spatial frequency spectrum in the sampled block, are then weighted and coded to exploit the attributes (and deficiencies) of the human eye.

Atva-p, the final system tested by the ATTC, is undergoing design revision and many of its technical features have yet to be disclosed ■

MC4 PUTS THE CAP ON CIRCUIT ANALYSIS

In building MicroCap-IV, Spectrum Software has added a wealth of features to MicroCap-III. Ben Duncan finds the latest version pushes PC-based simulation into new realms

For somewhere around £2500, Spectrum Software's newly launched MicroCap-IV (MC4) handles Spice-type entry, emulates the more useful Spice facilities and lays claim to run-speeds up to four times faster than other Spice-based simulators. With that kind of performance it is probably the fastest analogue simulator running in dos: while working in the extended dos mode it can handle up to 10,000 nodes.

But these major additions aside, it is worthwhile looking at improvements in the crucial details that determine everyday ease of use.

Many of the criticisms made of MC3[1] were tackled in later versions of the program and these improvements have been carried forward into MC4. Text editing is aided by toggling to INSERT mode and the dos environmental variable SET MC4DATA=[PATH] in the command line, allows work to be contained in any number of sub-directories, all hung off the data directory, which can contain just libraries.

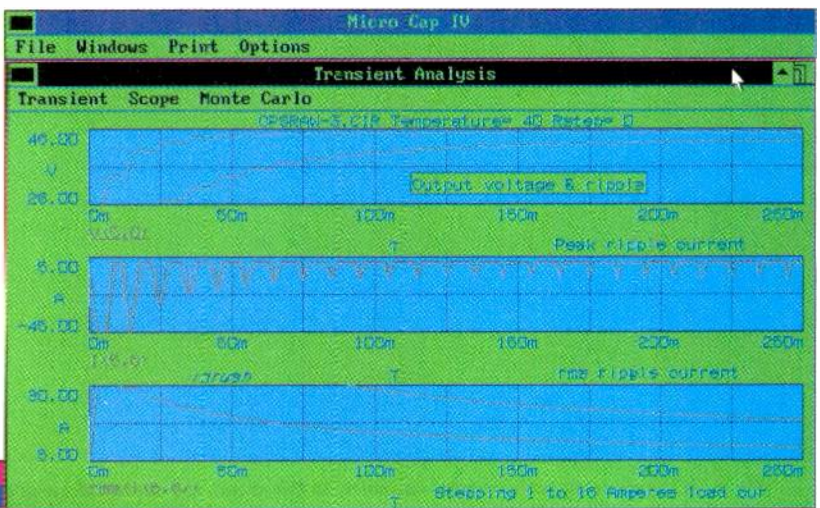


Fig. 1. Transient analysis of a power supply's output and ripple voltage, and peak and RMS current. Running computation of RMS is a new feature of MC4, as are the boxed and italicised graph labels, and the ability to have more than two full-width graphs on the screen. Each graph can contain six or more waveforms. Intelligent auto-scaling can be used to make plots fill their graph.

Fig. 2. Circuit of the power supply plotted in Fig. 1. Note the expressions and R_{step} dummy resistor used to step two components by separate ratios. Log stepping is now supported. The 50Hz signal sources on the left provide a first order approximation of bi-phase AC from the transformer's secondary.

Colour can now be used to distinguish nodes from text, and connectivity can be highlighted to aid beginners. Screen colours are selected globally – there are no longer separate sub-menus for each analysis mode – and colour mixing is simplified using three faders to combine the video primaries. Overall, screen graphics are much finer, allowing a bigger schematic and more information to be packed in at the 1:1 zoom level.

Prologue (set-up) screens for analysis can be printed in plain text. MC4 annotation of analysis graphs is improved, and text can be placed relative to a graph's inflexion, so it stays with the graph if the scale is changed. Bold, italic and

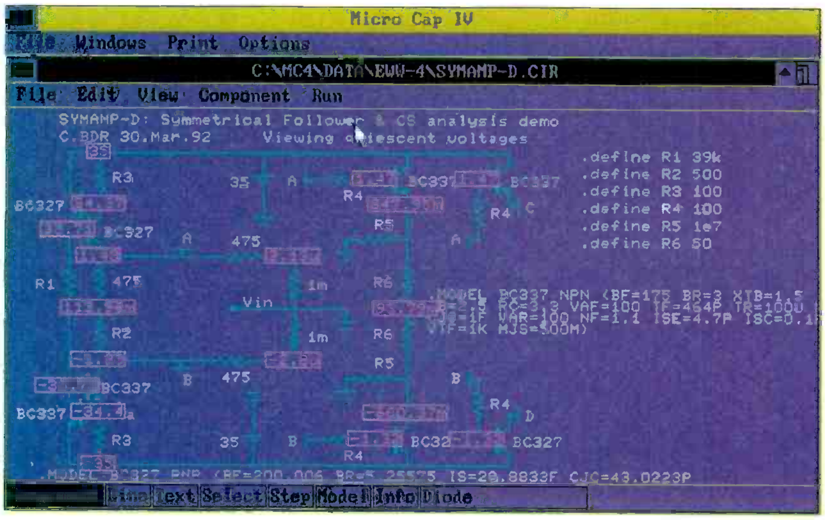


Fig. 3. To examine a symmetrical follower circuit, the transient analysis option OPERATING POINT ONLY *was used. On returning to the schematic, quiescent (or end-of-run) nodal voltages can be viewed, as seen here. Note the DC conditions are slightly askew, as happens in real circuits. Node numbers can be viewed at the same time, if desired.*

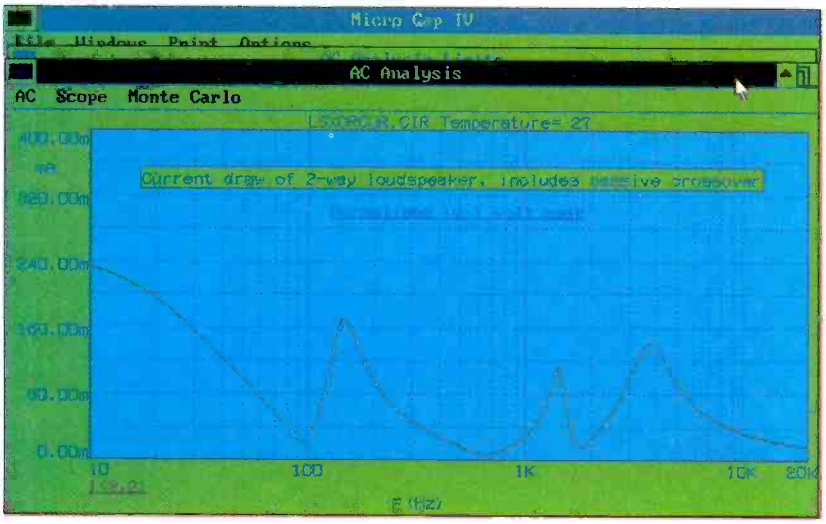


Fig. 4. AC analysis supports original variables. In this graph, the current draw of a two-way speaker is plotted against swept frequency, showing the regions of peak draw and worst case demand for a waveform of fixed magnitude. Previously, an exponentially swept sine wave had to be employed in transient analysis, and plotting took much longer. Drive level is normalised to 1V (pk). At 100V drive (corresponding to a nominal 600W into this speaker) current would be exactly 100 times greater, ie about 17A at 135Hz.

SYSTEM REQUIREMENTS

Any PC or PS/2
>dos 3.3
640k ram
1.2Mbyte or 720k 5.25in drive required to load
Executable and library files supplied occupy approx 3.3Mbyte (extended dos) or 2.5Mbyte (standard mode)
Coprocessor mandatory for 8086, 286, 386 CPUs
8086+8087, upwards.
80386/486 users can opt for extended version, using up to 16Mbyte extended ram, to increase all-round capacity and handle circuits up to 10,000 components or nodes (otherwise circuits limited to 100-150 components).
Up to 1Mbyte expanded memory to buffer schematics and analysis waveforms.
Auto graphics recognition and manual command line option encompasses all common video modes, including SVGA.
Output to dot matrix, laser, HP/HI plotters, or .PLT files.

underlined text as well as an outline box can be selected (**Fig. 1**) to clarity presentation, and for better identification, a title block option automatically puts the file name and date onto hardcopy. For large schematics, the page number and number of pages into which it is split appear on the copy, while schematics can also be autoscaled to fit the page. For those with 24 pin dot-matrix printers, the results are sharper.

Better multiple circuit handling

Opening of multiple circuit files has been vastly improved through TILE mode. Unlike CASCADE, where just the edges of all but the current circuit are seen behind the front window (unless it is scrolled to one side), TILE can display three circuits on the screen in equal, vertical strips. If four circuits are loaded, TILE displays them as four rectangles – like a TV broadcast control monitor.

Screen zoom now extends to 8:1, though only about 40% of the area is useable unless working in the extended dos mode. Throughout the program, every window now has a menu to define its position and size.

Schematic entry and libraries

Libraries now contain an impressive range of semiconductors, with some European and Japanese parts making a first appearance. But power engineers will still lament the non-inclusion of the thyristor/triac/UJT family, though Spectrum has outlined a cogent modelling method, using switches, in a past newsletter[2].

Simple clicking and dragging of a line, around a corner if necessary, makes for a faster line drawing, while the pull-down menu containing components has a "hot-list" which can hold the eight parts most needed. Parts include new virtual components – notably an arrow which can be used to show signal flow – invisible to the net-listing. For a sequence of numbered parts (eg R_1, R_2...), MC4 increments the number automatically as the next part is called up.

Semiconductors and cores drawn from libraries have to have a .MODEL statement appended to the screen. The statement (which lists all parameters different from Spice default values) is written automatically by selecting the part and clicking on the .MODEL command, so parts can be adjusted locally with the text editor specific to the file being worked on. A disadvantage of the approach is added screen clutter, especially if many different parts are in use. But .INCLUDE does allow the model lists to be displaced to a text file for the program to call on during analysis.

The number of parts can seem bewildering but it is easy enough to pick the few parts wanted from the libraries supplied, and copy them to a custom library.

New to MC4 are 100% Spice-compatible GaAs fets, and Level-III op-amp models (after Boyle) – the latter being far easier to enter than makers' Spice lists and promising similar accuracy. Any number of items can appear in a library list –100 op-amps consume about 38k – and to keep libraries tidy, a PACK command re-orders the listings in alphanumeric order, ie. 1-9, A-Z. The component editor contains approximately 80 types of component shapes and definitions, including common macros, with space for 40 or more DIY entries. Custom shapes can be created for original components and macros by the shape editor.

Analysis

DC analysis is little different to MC3, except in the range of expressions that can be plotted.

The separate Fourier analysis sub-routine has disappeared, while the AC and TA (transient analysis) routines have been considerably enhanced: more variables have been added, and

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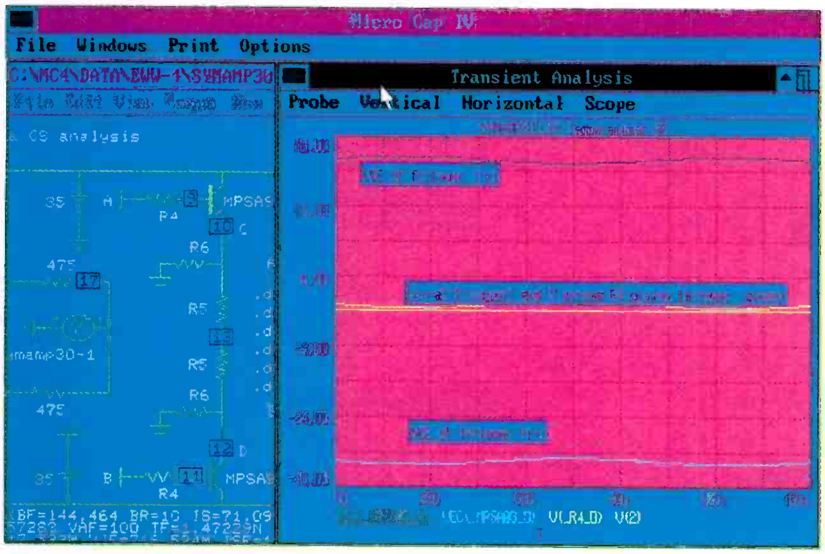

Fig. 5. Similar circuit to Fig. 3, seen in PROBE *mode. Click the mouse on any node or part to see the waveform. In this plot, voltage has been selected, and probing on the C-E portion of the upper transistor has plotted V_{ce}. If current had been selected, I_c would have been plotted instead. The graph is progressively autoscaled as more waveforms are added, up to the ceiling of six. In the figure, the voltages across R_4 and at R_2 are scaled too small, but the scope function is available to magnify them. The circuit segment in the left window is scrollable.*

Fig. 6. MC4's auxiliary MODEL *program. After entering data into the upper right table, pressing F10 prompts an optimisation routine to fit a curve to data (lower left), which can immediately be checked against the data sheet. More interpolation data and iterations will improve the fit of the h_{fe} vs I_c curve, at the high current corner (IKR). Many parametric curves show less inflexion and typically reach 10% curve-fit accuracy (indicated in top left box) with only six or so data points and two iterations.*

cardinal electrical quantities now include charge (C), and the Flux (X), B-field and H-field associated with cores. A much broader range of quantities related to common device terminals includes I_b, I_c, V_{be}, and V_{gs} and the charge on a mosfet gate.

There are more mathematical operators and signal processing functions too. Fourier functions (FFT, IFT, cross-spectrum, auto-spectrum, cross- and auto-correlation, plus coherence), are supported, as well as average, RMS and sum (running integral) quantities. Most functions, operators and expressions can be used to define both X and Y fields in TA (**Fig. 7**).

Figure 1 illustrates the RMS option, used to establish the heating effect of ripple current against load current, in a power supply where impulsive current draw is forestalled by an NTC thermistor. **Figure 2** shows the circuit and the expression used simultaneously to step the load resistor and an NTC resistor between their respective endpoints. The Polysource artefact in MC3 has been replaced by a dummy resistor, R_{step}, to which the stepping ratios are tied.

Analysis prologue screens are considerably simplified, notably for TA, by omitting fiddly settings governing speed vs accuracy. Some settings are now optimised automatically; others (like the number of iterations) are controlled globally.

Stepping of component values and other parameters can now be logarithmic, as well as linear, enabling broader yet detailed sweeps of many decades – if there is some uncertainty when the relevant parameter begins to "bite", for example.

TA and AC runs can be autoscaled simply by entering AUTO in the X and Y range columns. MC4 makes an initial run on a generic scale, then replots with the full-screen scale. Five or more separate graphs can be chosen for simultaneous plotting and display. The penalty is, naturally, the loss of detailed scales – though SCOPE overcomes this. Alternatively, up to ten plots can be made to run on one graph/five on two graphs, etc. The scale is the union of the biggest span, so at first the option appears only to be useful if the waveforms are the same order of magnitude – in digital simulation, for example. But the expression *100 (say) could be included in the Y entry for 100x magnification.

After a TA run, node voltages can be viewed simultaneously by returning to the schematic – a handy feature last seen in MC2 (**Fig. 3**).Potentials can be either static operating point (by selecting OPERATING POINT ONLY in the TA prologue), or instantaneous voltages at cessation of the run; or SCOPE will home in on portions of the graph. Routines to pick out peak and valley have improved SCOPE – giving a lot of ancillary numeric data, according to cursor position.

AC analysis handles more expressions than previously and is no longer limited to the regular magnitude/phase/group-

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MC4 CONTENTS

Components: R, C, L, transformers, cores, transmission lines. Excepting *R*, all types can include complex non-linear elements; eg during TA, inductance can be modified by an arbitrary function of any valid time-domain variable. Tol (%) and tempco (ppm/°C) may be appended to all types. Level-III (Boyle) op-amps and Spice-compatible GaAs fets. Libraries supplied expanded to *998* BJTs (Motorola and NSC), *1114* jfets and mosfets (Hitachi, IR and NSC), *210* diodes (Motorola), *540* op-amps (LT, NSC, PMI, Texas) and *260* ferrite cores (Philips, Toko). Nearly all parameters can have tolerances appended.

Other parts: As MC3 with addition of exponential pulse and FM waveform sources, Spice sub circuits, and an expanded range of analogue behavioural building blocks, eg hysteresis, multiplication and VCO macros.

Expressions: Used to define parts and analysis dimensions. Operators and functions include common math, Boolean, Relational, Complex and Signal Processing.

On-line utilities: Help, with interactive access from any location. Paste board. Scientific calculator. Title block with auto-dating.

Utilities via dos: Model utility computes parameters for semiconductors and cores, based on optimised piecewise approximation from tabular entry of data points. Interactive screen graphs provide validative feedback. Completed models are exported to MC4. CONVERT utility converts MC2 and MC3 circuit files and libraries to MC4 format. TOSPICE utility converts MC4 schematics to Spice format. Spice files of circuits and waveforms can be imported directly.

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In the new version V2.40, full consideration has been given to allow designers to continue using their existing schematic capture package as a front end to BoardMaker. Even powerful facilities such as Top Down Modification, Component renumber and Back Annotation have been accomodated to provide overall design integrity between your schematic package and BoardMaker. Equally, powerful features are included to ensure that users who do not have schematic capture software can still take full advantage of BoardMaker's net capabilities.

BoardMaker V2.40 is a remarkable £295.00 (ex. carriage & VAT) and includes 3 months FREE software updates and full telephone technical support.

AUTOROUTER

BoardRouter is a new integrated gridless autoroute module which overcomes the limitations normally associated with autorouting. **YOU** specify the track width, via size and design rules for individual nets, BoardRouter then routes the board based on these settings in the same way you would route it yourself manually.

This ability allows you to autoroute mixed technology designs (SMD, analogue, digital, power switching etc)in **ONE PASS** while respecting **ALL** design rules.

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You can freely pre-route any tracks manually using BoardMaker prior to autorouting. Whilst autorouting you can pan and zoom to inspect the routes placed, interrupt it, manually modify the layout and resume autorouting.

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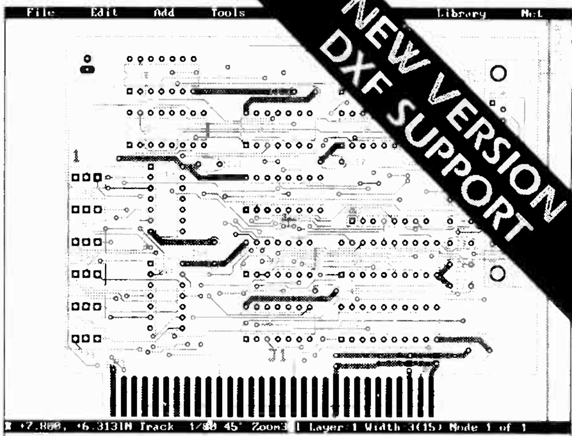
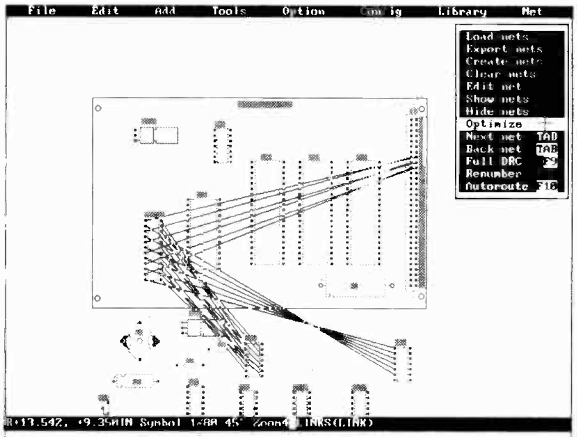
NEW VERSION DXF SUPPORT



Full analogue, digital and SM support - ground and power planes - 45 degree, arced and any angle tracks with full net-based Design Rule Checking.



Optimized placement by displaying ratsnest per component. Lines indicate the unrouted nets.

HIGHLIGHTS

- Net list import from OrCAD, Schema etc.
- Graphical and manual netlist entry
- Top down modification for ECOs
- Forward and back annotation
- Component renumber
- Effortless manual routing
- Fully re-entrant gridless autorouting
- Simultaneously routes up to eight layers
- Powerful component placement tools
- Copper fill
- Curved tracks
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DIFFERENT FAMILY

Spectrum Software's MicroCap family has always differed from basic Spice systems, by working from a schematic and plotting results during (not after) a run.

Over 10,000 copies of MC1, 2 and 3 have been shipped to date.

MicroCap-IV reads Spice 2G files directly, and emulates the more useful Spice characteristics, while building on the intuitive, graph and schematic-based friendliness of MC3. Like later versions of MC3, it is written in C. A Mac version of MC4 is expected in the 3rd quarter of this year. Spectrum says that MicroCap-III will remain available and continues to be developed if demand warrants it.

delay/noise set (**Fig. 4**). Bode plots and Nichols' charts are included, but the implicit "virtual" AC driving source is no longer supported. Instead, a sine generator, or any other source, has to be supplied. MC4 reads the peak magnitude, ignoring source resistance and frequency, etc.

Spice aid

Spice's dependent sources are now covered, aiding entry of SUBCKTS Spice netlists. Presently, these are readily available for numerous AD, PMI, SSM and Texas op-amps.

After entering (or importing) the Spice text listing, SUBCKT provides rapid access by linking the netlist to a suitable component shape, as in ordinary macros. MC4's level III op-amp routines may be preferred, as they are simpler to enter and give the same level of detail

Command statements have been extended: the .DEFINE, .MUTUAL and .MODEL statements common to MC3, and the regular Spice command statements, are now joined by .INCLUDE, .PARAMETERS, and .PINS. MUTUAL couples inductors; PARAMETERS passes parameters to macro circuits, and PINS defines external nodes connected to the macro's shape. INCLUDE causes statements to be read from an external text file and so stops model statements from cluttering the schematic.

Spice's IC fixes node voltages during a DC operating point calculation; NODESET does not, and both are helpful for tricky simulations. Other Spice statements include AC, DC, OP, TEMP etc but these rather clumsy commands are used solely when emulating Spice.

Perhaps the most important option is PROBE, allowing a circuit to run blindly (as in Spice). Saved as a file, the circuit can be recalled at any time, and by pointing the mouse at any node or nodes, the relevant waveform displayed dynamically (**Fig. 5**). Assuming comprehensive modelling by a manufacturer, PROBE has potential for augmenting service and training handbooks.

Semiconductor model development

The MODEL utility derives the esoteric parameters for semiconductors and magnetic cores from information commonly presented in data sheets. Its value can be gauged by the fact that the Ebers-Moll and Gummel-Poon parameters needed for simulating bi-polar transistors are still only readily available from only a handful of transistor makers, notably Zetex in the UK and PMI/SSM (USA). Excepting op-amps, the analogous level I, II and III Spice parameters needed for other devices are not recognised by most semiconductor manufacturers.

MODEL is accessed via dos and has the same windows format as the main program. Device parameters are stored in libraries for the 3000 or so parts supplied, grouped according to manufacturer – but more can be created, in model libraries dedicated to (for example) specific makers, projects or periods.

Once inside a given library, a pull-down window selects the type of new part to be entered. The main screen (**Fig. 6**) principally comprises (upper right) a box to enter subsidiary data points, (lower left) a graph of these, and (lower right), the numeric parameters derived from the data entered.

To derive a model, find a curve of the initial parameter – eg I_c vs V_{be} for bipolar transistors – then pick some data points and enter I_c and V_{be} in the table. Up to 100 points can be entered, but accuracy will often be within an acceptable 10% or so with only 2 to 5 points defined (according to inflexion) – providing these are reasonably spaced on a log scale.

After the first "page" of data is entered, scroll on to the next. E-M, G-P and Spice parameters are derived from a relatively small number of graphs of common measurements, so only half a dozen or so data-sheet parameters need entering to generate 40 or more of the esoteric parameters. On entry completion, MODEL computes the parameters automatically following INITIALISE and OPTIMISE. Using Powell's direction set[3], the modelling utility adjusts the curve to fit the numeric data entered. If the fit is not good enough, try again (**Fig. 6**); the number of iterations needed and the present curve-fit accuracy is displayed. In the example the error is still quite high at 16%, and more points will need to be added to track the h_{fe}'s high-current corner, with its strong inflexion.

Model also handles finding, sorting, duplicating and deleting part models and merging of libraries, and the completed component model can be exported into any MC4 library, or exported into MC4 or elsewhere, as a Spice library file. Entire libraries can be exported into MC4.

Fine tuning

Monte-Carlo statistics gathering, building on MC3, allows entering of both LOT and DEV tolerances for semiconductors and cores – the tolerances indicating group and individual variations respectively. For example, if a number of transistors are labelled *Q1*, LOT=X% can be set to vary all their h_{fe}s by X% in tandem. Adding a DEV=5% statement would also vary them individually by up to +5%.

For simpler parts (*R*, *C*, *L* etc), LOT is used on its own. The collating function used to create the 3-D statistical data[1] is limited to the defined X, Y window. MC4 highlights this area on the graph and typically it is set up to capture the -3dB frequency, rise-time, or peak gain.

Complex... but not daunting

To complete this review within a month of Micro-Cap IV's worldwide release, I was working with Beta versions of the software. Even so I met surprisingly few bugs and idiosyncrasies for such an ambitious program. Having seen how rapidly MC3 was refined in succeeding versions, I think we can be reasonably confident that the released versions will be fine. Spectrum points out that the small number of explicit shortcomings – no noise analysis for op-amps, and a limited diode library – will be conquered later in the year.

A program of this size and complexity could easily seem daunting – MicroCap IV aims to upstage Spice simulators costing over £6000 – but the cogent manuals, interactive help screens, extensive and informative error messages and robust analysis go to build on the friendly nature of its predecessor.

SUPPLIER DETAILS

Micro-Cap IV is produced by Spectrum Software, 1021 S.Wolfe Rd, Sunnyvale, CA 94086, USA. US cost is $2495. Distributed in UK by Datech, Sidcup, Kent. Tel: 081 308 1800. Cost is expected to be around £2500 + VAT. An evaluation version will cost $250 in the US; UK price not yet available. UK users receive a quarterly cae newsletter, dealing with modelling and technical tips. Program updates are free unless features have been added, in which case there is a small fee.

References

1. Micro-Cap III under test, *EW + WW*, July 1990.
2. SCR modelling, Spectrum CAE Newsletter, June 1990.
3. Numerical recipes in C, The Art of Scientific Computing, Cambridge University Press, 2nd ed, 1990.

CIRCLE NO. 133 ON REPLY CARD

CIRCLE NO. 120 ON REPLY CARD

CIRCLE NO. 121 ON REPLY CARD

PROFESSIONAL PCB DESIGN AT A BUDGET PRICE

Can Seetrax's Ranger1 entry-level PCB design software really offer a fully-fledged PCB design system for £100? John Anderson has no doubts.

SYSTEM REQUIREMENTS
PC compatible
640K memory
EGA or VGA screen
Mouse
Roller ball or tablet pointing device
Full parallel port for dongle
Outputs to pen plotter, photoplotter and dot matrix printer

Over 2000 sales of Ranger 1 point to the strong following won by this low-cost but powerful package. The Seetrax product is surely one of the best value, highest performance PCB design packages available.

All aspects of its work produce results with a professional style worthy of products costing ten times as much (apart from the menu system looking weak compared with its GUI).

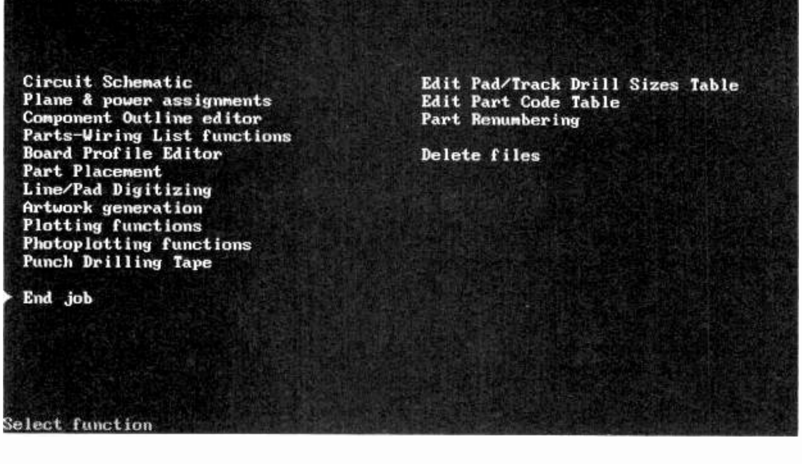
Indeed comparison with other products at the budget end of the market is rather difficult as Ranger 1 offers facilities such as project control, autorouting and photoplotter support that other packages either do not have or supply as expensive add-ons.

Starting first with one of the package's weaker attributes, operation is controlled through selection from simple text list menus. Bearing in mind the slick graphics user interface that characterises the rest of the product, this seems rather poor – though to be fair, with mouse selection, it is a fairly quick and simple process.

The three-option main menu presented at start-up is the job control menu, allowing an old job to be selected or a new one generated. No editing takes place in this process, but the approach does represent an excellent solution to the file management problem of PCB work – keeping tabs on the large numbers of schematic, netlist, PCB, wiring and associated files and their back-ups.

Having selected a job, a more complex menu shows all the possible options for schematic editing through to artwork plotting. Selecting CIRCUIT SCHEMATIC brings up the next menu allowing editing of the schematic and components to be used on the schematic.

Ranger's main job menu looks unattractive compared with the rest of its GUI.



UPGRADES

Ranger 1 has a simple upgrade path to the Ranger 2. But why Ranger 2 when Ranger 1 is so good? The listed benefits are 100% rip-up and retry autorouter, file transfer to dxf files (Autocad), auto-placement and gate and pin swap. But these benefits also show that the main functionality of Ranger 1 is comparable to the more expensive product in terms of its fundamental attributes such as layers, libraries and interlinked schematic and PCB layout. A slightly annoying omission from Ranger 1, available in Ranger 2, is the area copper fill. With Ranger 1 it can only be achieved by laying down a series of fat tracks.

Graphics editor

Like other PCB design products, Ranger's graphic editor appears in different guises when editing schematic, components or artwork. But the editor is undoubtedly a good one, with functions activated by moving the mouse onto one of the simple vertical and horizontal menus.

Once perfected, the approach is very effective, though inadvertently moving the mouse from the schematic on the menu area may trigger commands unintentionally. For example, touching PAGE with the graphics cursor will require an annoying keyboard input. Zoom and pan are quick, with a very good redraw speed even with quite complex diagrams. But exit from the graphics editor is only possible with enforced save; there is no reference in the index to exit without save, so the only way to abandon work appears to be CONTROL-ALT-DELETE!

No integral help system is available, and this will be missed by first-time users as the manual is not well organised and a context sensitive help would have eased the initial steep learning curve. However a single example is supplied on disc, and there is a helpful design section in the manual which goes through the processes step by step.

Editing artwork

The artwork editor is basically the same as the schematic editor, with the one quirk that if a small-pitch grid is defined then at some zooms the grid only covers part of the screen. But the same neat row and column menu system is used although some of the shortened action prompt names are slightly cryptic. Up to 15 layers can be used, with power planes assigned to any layer.

One drawback is that all data, netlists, component lists, outlines etc are stored in binary format. Data can be output pre-formatted in ascii to a printer, but this limits the ability to edit or review job data outside Ranger.

Autorouting and libraries

Ranger's autorouter is supplied at extra cost. At £50 it is not going to break the bank and even at such a low cost the product is a good router, with separate power (wide track) routing, memory single sided routes, and orthogonal routing ie horizontal and vertical routes through vias as necessary. Operation mode is selected at a set-up menu, controlling layers for routing clearances, routing resolution and cost factors.

The router attempts to find a route between the source pad and the target pad and allocates costs for each move possibility. Costs of vias, 45° angles etc are evaluated for each movement and the minimum cost path chosen once the target is reached.

Algorithms used are channel routing – a direct routing method used to achieve high completion speed and minimal deviations – followed by the well known Lee algorithm used to achieve maximum completion.

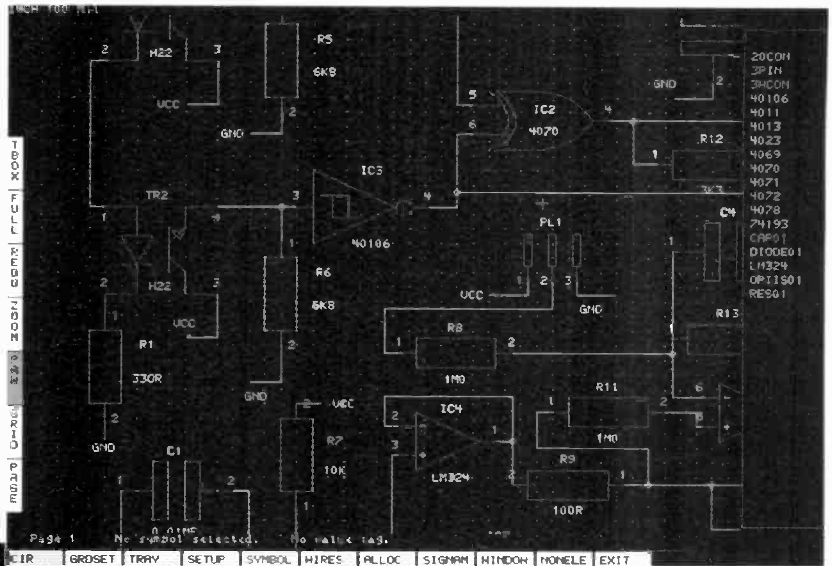
Reasonable libraries include device data, schematic and artwork footprints, and graphical aspects of a device can be edited or created in a similar graphical environment to that used for schematic capture or artwork generation.

Emulating more expensive products

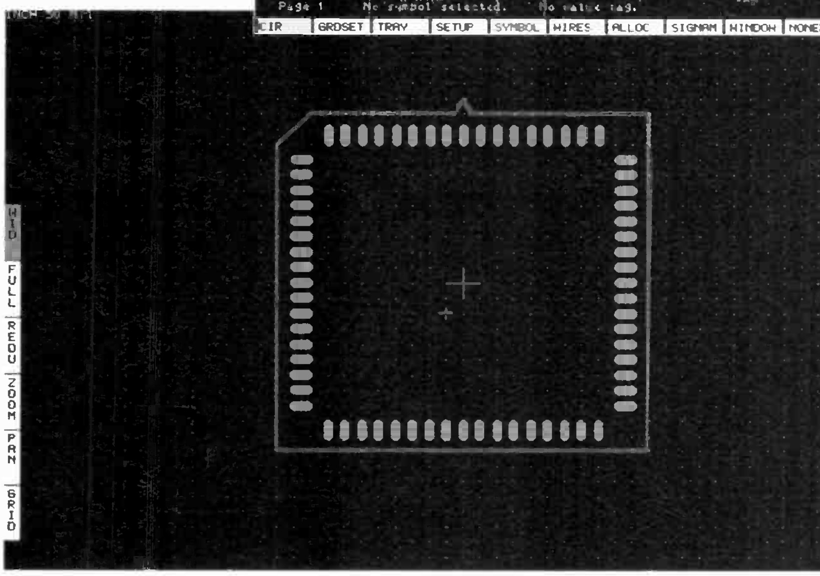
Design rule check, based on the simple criterion of maximum allowable distance, is rather weak, as clearance distance depends on whether a pad is drilled. It can work quite efficiently, through a textual screen outside the graphical editor: when the clearances are OK the filter completes within a few seconds. But when errors are found, the program slows to a crawl and a moderately-sized PCB takes several minutes to complete.

In a neat touch, errors are all logged onto the artwork as green arrowheads, allowing location and rectification to be completed quickly. A similar filter is used to check the schematic data for unused pins, producing a list of single nodes which can be output to the printer.

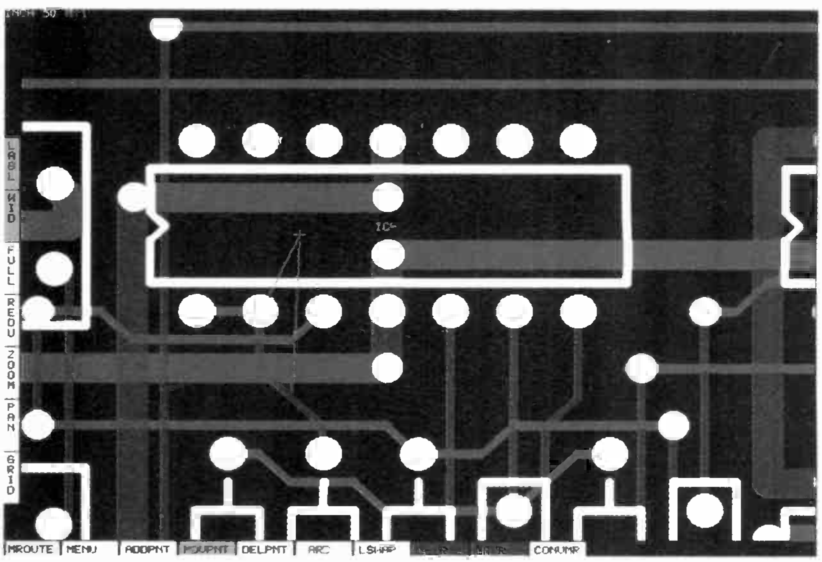
One very impressive facility of Ranger – usually available



Schematic capture – note the list of components used at the right hand side of the screen



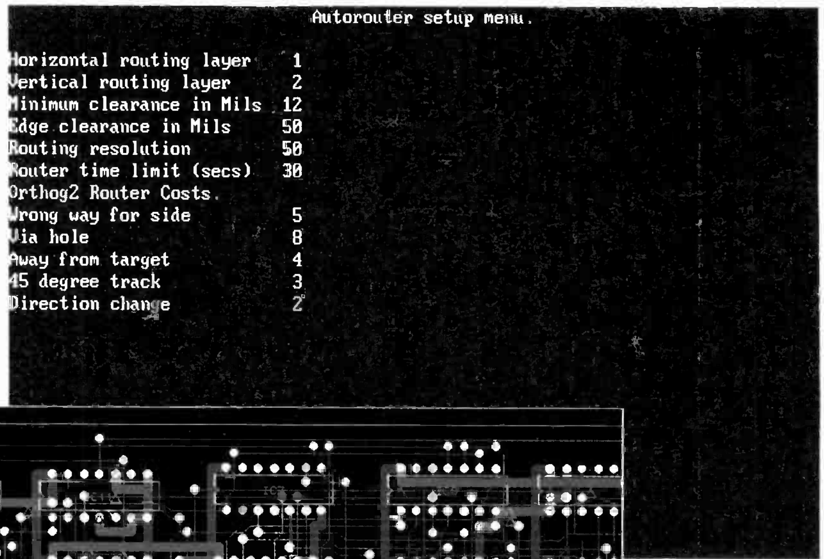
Editing the master outline library, in this case a PCC68 pin package



PCB layout manual editing is easy to achieve.

only on much more expensive products such as Pads – is back annotation, a scheme to update wiring and schematic data from the layout. It is established at a text editing session, where devices are listed as they are currently assigned, and the user can then re-code them as required. Another boon is that device prefixes can be set up as desired. So if all transistors are to be referred to as *Uxx* rather than *Txx*, this can be installed into the database of the system.

The job-based concept includes some job-specific file manipulations, including delete and copy. Since a job comprises a large number of files, the whole job methodology should ensure consistency of version control.

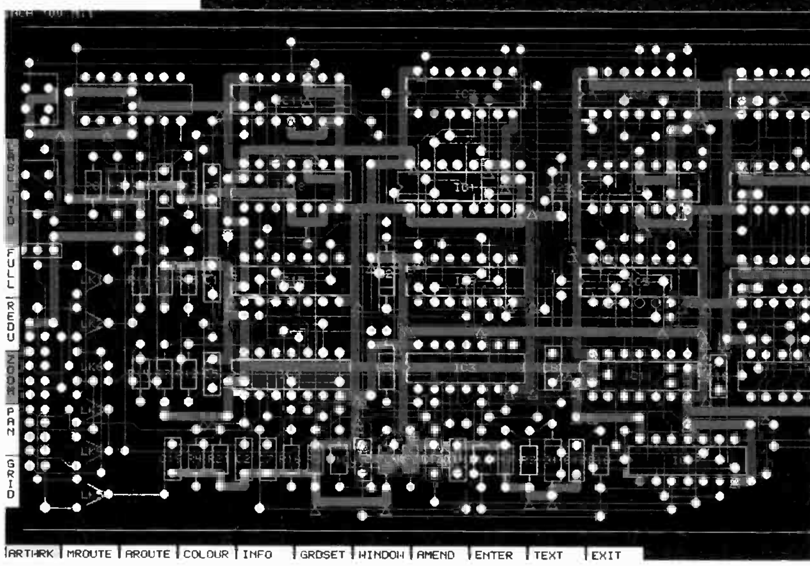


Autorouter set-up menu showing the user-definable routing costs.

Best available

Other packages in the same price bracket are simply outclassed in almost all departments by this exceptional product. Weak points are few, and each part of the PCB design process has been carefully considered from a user standpoint.

As an entry level product, it is hard to beat. But for many users, where designs are small enough not to run out of space on the PC, Ranger 1 might be the only product they will need to buy. ■



Artwork editing after design rule check. Green triangles indicate clearance errors.

SUPPLIER DETAILS
Ranger 1 Interactive: a PCB design system £100. Ranger 1 Autorouter: artwork router £50
Upgrade Ranger 2 £999 Supplied by Seetrax CAE, Hinton Daubnay, Lovedean, Portsmouth PO8 0SG.
Tel: 0705 591037

DESIGN BRIEF

DC accurate filter plays anti-alias role

Ian Hickman shows how a five-pole switched capacitor filter built around an eight-pin dip can produce a zero DC offset with no 1/f noise contribution.

Digital signal processing chips offer a wide choice of speeds, number-of-bits and architectures – especially at audio and video frequencies. But before a signal can be processed with a DSP it must be digitised, preferably after lowpass filtering.

It may be that in a particular application, frequency components in the signal are not expected at or above the Nyquist rate. But there is always the possibility that extraneous interference could enter the system, and so only the foolhardy will dispense with an anti-alias filter.

Passive LC types used to be the main filter choice – latterly displaced by active RC filters doing exactly the same job (subject to dynamic range limitations) much more cheaply. But like LC filters, they are still not easily variable or programmable.

Switched capacitor filters overcome this practical difficulty, but their time-discrete nature means aliasing can be a problem. Clock frequency is fifty or a hundred times the filter's low-pass cut-off frequency, so a simple single-pole RC roll-off ahead of the filter often suffices, with another after to suppress clock frequency hash in the output. Even where a variable clock frequency is used to provide a programmable cut-off, a fixed RC may still be enough if the cut-off frequency variation is only an octave or two – particularly if the following A-to-D converter only uses eight bits.

But aliasing problems are avoided entirely if a time-continuous filter is used, such as the 8th order/4th order *Max274/275*, with devices available in IC form requiring no external capacitors.

Cut-off frequency and response type (Butterworth, Bessel, Chebychev etc) are programmed by means of external resistors and frequencies down to 1kHz or lower are realisable with manageable resistor values. But the cut-off frequency cannot be varied once set,

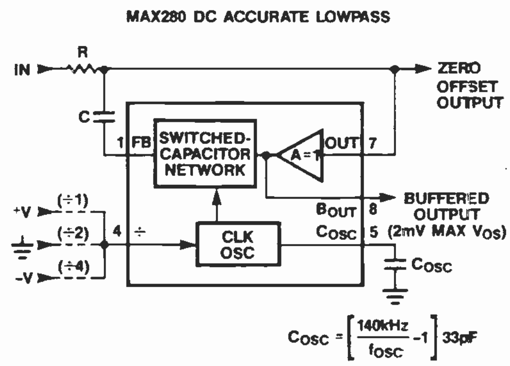


Fig. 1. Connecting up the Max280 to act as a capacitance multiplier, with C appearing ever greater with progress up the stopband.

though a limited choice of corner frequencies can (rather cumbersomely) be accommodated by using analogue switches to select different sets of resistors.

Good compromise

An interesting alternative filter type can be built with the "DC accurate" *Max280* plus a few passive components – producing a sort of halfway house between pure time-continuous filters and clock-tunable filters. The combination results in a five pole low pass filter with choice of approximations to Butterworth, Bessel or Elliptic characteristics. Since the *RC* passive single pole is located right at the filter's input, it also serves as the anti-alias filter, providing 43dB of attenuation at the Nyquist frequency.

In an unusual filter arrangement (**Fig. 1**) the "earth" end of the RC's capacitor goes not to ground but to a pin labelled FB (feedback). If it were grounded, the stop-band response would show the usual 6dB per octave roll-off. But the chip actually acts as a capacitance multiplier so that *C* appears ever greater as we move higher up the stop band. The result is a fifth order 30dB/octave roll-off.

Exactly why this works is not entirely clear. Filter cut-off frequency is set by the clock frequency, from a free-running internal oscillator. The oscillator may be over-ridden by an external clock applied to the C_{osc} terminal, pin five (11) on the eight- (16-) pin dip package, and swinging close to the V+ and V– rails.

Without additional C_{osc}, the internal clock runs at 140kHz nominal but as this can vary by as much as ±25% over the full range of supply voltages, it is as well to add stabilisation.

To check clock frequency with a 'scope, turn sensitivity up to maximum and *hold the probe near* to the C_{osc} pin – even the 11pF or so of a x10 probe can pull the frequency down 20% if connected directly.

With no additional C_{osc}, the filter's –3dB point will be a little over 1kHz with the divider ratio pin connected to V+. Connecting to ground or V–, divides the internal clock F_{osc} by two or four, lowering the cut-off frequency by one or two octaves.

An external C_{osc} can give an even lower filter cut off frequency but for a cut-off fre-

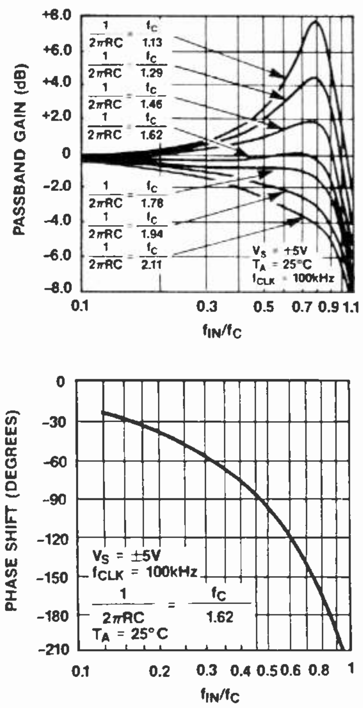


Fig. 2. Typical operating characteristics: (a) passband gain vs input frequency; (b) phase shift showing that this characteristic has already reached 180° at 0.85 of the 3dB cut-off frequency f_c

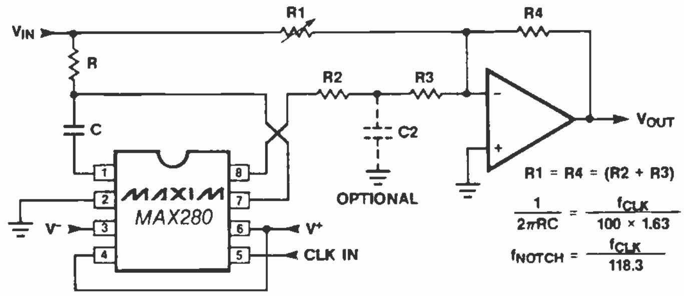


Fig. 3. The Max280/LTC1062 used to create a notch. The input signal can be summed with the filter's output to create the notch.

quency higher than 1kHz, an external clock of up to 4MHz may be used.

Filter response shape in the region of the passband/stopband transition is determined by the relation between the clock frequency applied to the *SC* network and the time constant of the passive *RC*. Being a fifth-order network, the phase shift has already reached 180° at about 0.85 of the filter's 3dB cut-off frequency f_c, (**Fig. 2**) and a notch filter can readily be implemented (**Fig. 3**).

Trying out the circuit with R = 39k, C = 6n2, and R_1 to R_4 all 100k, gives a nice deep notch at 890Hz.

Checking with a scope shows the internal clock to be running at 105kHz – which makes sense.

Far below the notch frequency, circuit gain is x2. Well above – where the path through the filter is dead – it is x1.

Targeting fastest response

As already noted, when used as a lowpass filter the response type is set by the CR time constant relative to the clock frequency – giving approximations to a Butterworth or Bessel response, **Fig. 4**. But where the fastest possible rate of cut-off is required in the stopband,

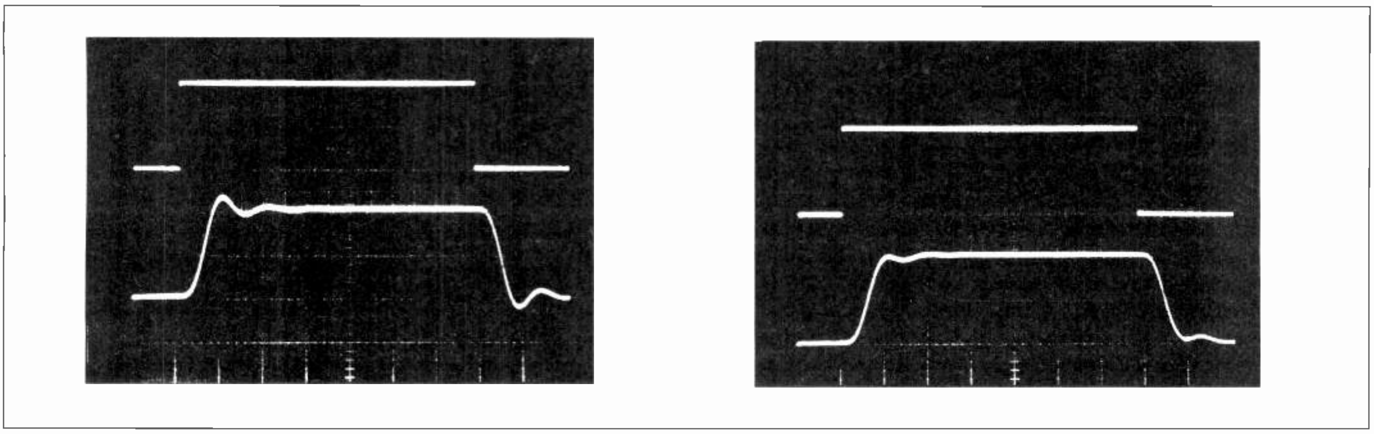
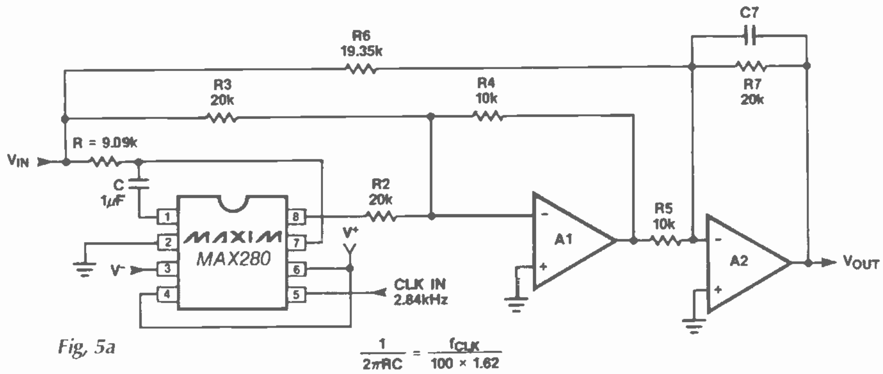
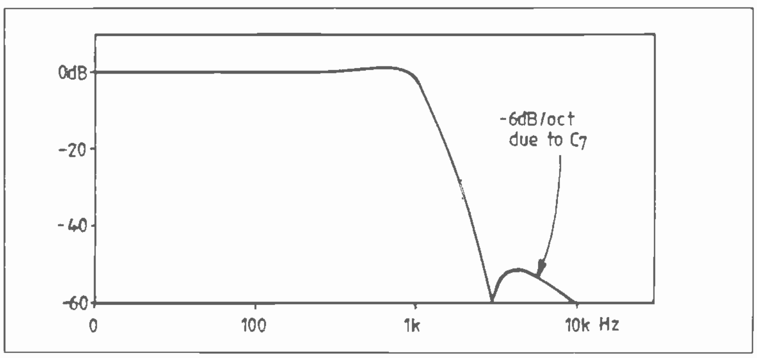


Fig. 4. Using a lowpass filter to give an approximation to (a) a Butterworth and (b) a Bessel step response.



Fig, 5a



Fig, 5b

Fig. 5. Modifying the lowpass filter circuit, 60Hz notch (a) with R = 39k, C = 5n6, C_7 = 1n, $R_{2,3,6,7}$ = 100k and $R_{4,5}$ = 47k gives the response shown in (b).

a response with a finite zero is the aim. Modifying the values in **Fig. 5a** to R = 39k; C = 5n6; C_7 = 1n; $R_{2,3,6,7}$ = 100k; $R_{4,5}$ = 47k gives the response shown in **Fig. 5b**.

If the output from the basic filter (Fig. 2) is fed back to its input via an inverting amplifier, there will be zero phase shift at 85% of the cut-off frequency and an oscillator should result. Using a pair of diodes for amplitude control (**Fig. 6**) results in a very convincing looking sine-wave. Measuring with a total harmonic distortion meter indicates 2% – which sounds disappointing – but using the 'scope to look at the residual shows it to consist almost entirely of switching hash.

Switching the THD meter's bandwidth from 80kHz to 20kHz gives the more respectable figure of 0.18% THD, virtually pure third harmonic.

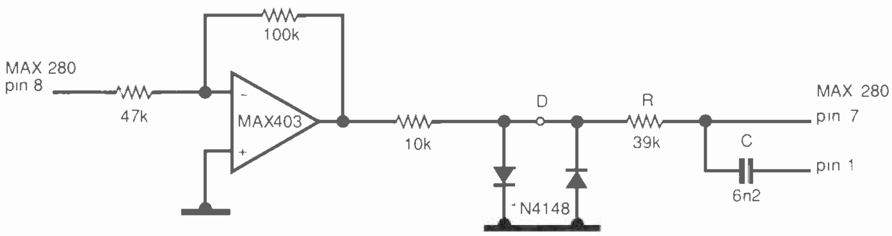


Fig. 6. Turning a filter into an oscillator. Feeding the output from the basic filter to its input via an inverting amplifier, using a pair of diodes for amplitude control gives a good sine-wave. Switching the THD meter's bandwidth gives a "virtually pure" third harmonic.

Practical considerations

A number of practical points arise when applying the *Max280*.

If it is only the AC component of the signal that is of interest, the output can be taken from the buffered low impedance output at pin eight. But if the DC component is also important, there may be an offset of up to 2mV. In this case, the DC accurate "output" at pin seven should be connected directly via R to the filter's input; the buffer's typical input bias current of 2pA is not likely to drop a significant voltage across R.

The DC accurate output should still be buffered before feeding to, for example, an A-to-D converter, since the pin-seven to pin-one path is part of the filter, and capacitive loading of even as little as 30pF at pin seven may affect the filter response. A passive RC post filter is also recommended to suppress the 10mV pk-pk (typical) clock feed-through hash. At the other end of the spectrum, the filter contributes no low frequency or 1/f noise, since in the active circuitry it would have to pass from pin one to the output via a passive CR highpass filter.

For critical filtering applications, two *Max280s* may be cascaded to provide a 10th order DC accurate filter. ■

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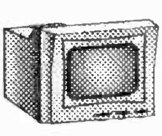

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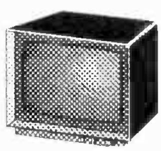

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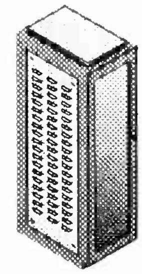

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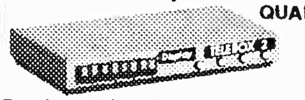

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Circuits, Systems & Standards

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

Design method yields low-noise, wide-range crystal oscillators

Traditionally, designing crystal oscillators with wide tuning range and acceptable noise performance has required time-consuming trial-and-error methods. But there is a straightforward design procedure for achieving a tuning range of hundreds of ppm without compromising noise performance. By characterising the tuning range of the frequency-determining network of a crystal oscillator, the effects that determine oscillator performance can be accurately quantified and included in the design.

This extends the traditional approach which relies on a crystal's equivalent-circuit model – a model that does not characterise the crystal in difficult designs. The equivalent-circuit approach simply fails to accommodate such relevant effects as off-resonance lossiness, crystal-model element variation with frequency and spurious modes. These effects are usually lumped into a vague term called "crystal pullability".

Here we will concentrate on feedback-type oscillators using crystals configured in the series-resonant mode (**Fig. 1a**), though the design techniques are applicable to other topologies. In the block diagram, $H(f)$ is a low-bandpass filter that selects the desired harmonic. G_2 is a buffer that provides a low-impedance termination for the frequency-determining network, and G_1 is a non-linear gain block providing variable gain or the amplitude limiting necessary for stable oscillation. Gain of G_1 decreases monotonically as a function of signal level.

When the circuit turns on (and signal levels are low), G_1 varies to provide an open-loop gain greater than three to ensure that the circuit oscillates. Once the circuit is oscillating and signal levels attain a steady state condition, G_1's gain is approximately unity.

The frequency-determining network contains the crystal and associated tuning elements and the circuit block is a one-port network, ideally described as $Z_f(f)$. For optimum performance, minimise undesirable admittances, such as those attributable to stray capacitance and varactor-isolating resistors. The dotted ground connection (carrying negligible current) from the frequency-determining network is an example of a low-admittance stray network path.

For the circuit to oscillate, open-loop phase shift must

Crystal oscillator design
It is easy to use one of the commonly used circuit arrangements to throw together a crystal oscillator – and the result may well be adequate for an undemanding application. But if low noise and/or wide tuning range is a requirement, a slightly more scientific approach is called for. This article will provide food for thought for anyone needing to design a crystal oscillator.
IH

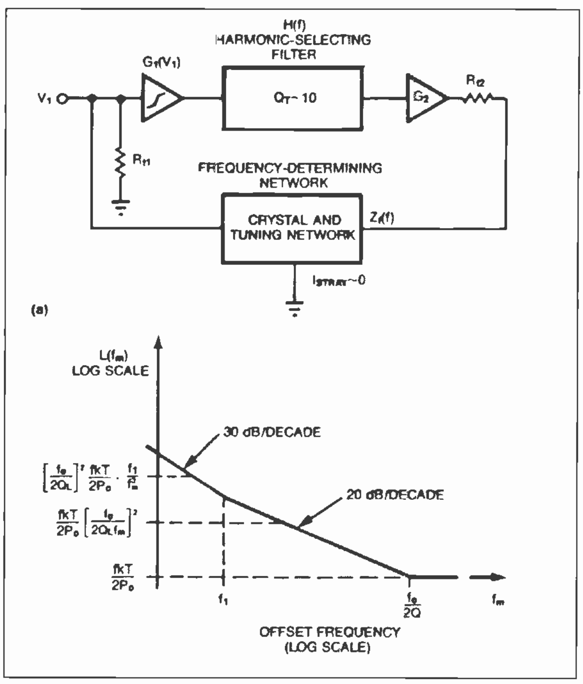


Fig. 1. Circuit including two frequency-dependent blocks, so a number of harmonic-selecting filter- and frequency-determining-network phase combinations will provide an overall loop phase shift of 0° – a necessary condition for oscillation.

equal 0°. Because the circuit in Fig. 1 contains two frequency-dependent blocks, a number of harmonic-selecting filter and frequency-determining network phase combinations will provide an overall loop phase shift of 0°. A 0° phase shift in each network is best for two reasons – maximum loop gain and minimum phase noise. The design method shown here uses 0° phase shift as a condition.

Concentrating on the crystal loop – the closed path around R_{t1}, the frequency-determining network, and R_{t2} – it is this that partially determines the oscillator's phase noise and completely determines the frequency tuning. The frequency-determining aspect of the loop is true because everything outside the crystal loop is independent of frequency over the narrow frequency range of interest. In the model, G_1's input impedance is infinite, and G_2's output impedance equals zero. If ISTRAY is negligible, the frequency-determining network can be considered to be a one-port impedance $Z_f(f)$. Therefore, the relevant equations for oscillation and noise become

$$I_m\left[Z_f(f)\right]=0 \qquad 1$$

$$Q_L(f)=\frac{f_0}{2(R_{t1}+R_{t2}+R_e\left[Z_f(f)\right]}\cdot\frac{\mathrm{d}I_m\left[Z_f(f)\right]}{\mathrm{d}f} \qquad 2$$

where $Q_L(f)$ is the Q of the frequency-determining network. This parameter is also known as the "loaded Q" of the crystal. The dominant parameter under designer control, $Q_L(f)$, determines the oscillator's noise level. According to Leason's equation, the single-sided phase noise equals:

$$L(f_m)=\frac{FkT}{2P_0}\left[1+\left(\frac{f_0}{f_m 2Q_L}\right)^2\right]+\left[1+\frac{f_1}{f_m}\right] \qquad 3$$

$$\cong\frac{FkT}{2P_0}\left[\frac{f_0}{f_m 2Q_L}\right]^2 \text{for } f_1<f_m<\frac{f_0}{2Q_L}$$

$$\cong\frac{FkT}{2P_0}\left[\frac{f_0}{2Q_L}\right]^2\left[\frac{f_1}{f_m^3}\right] \text{for } f_m<f_1<\frac{f_0}{2Q_L}$$

where f_0=oscillation frequency, $f_m \approx f-f_0$=offset frequency, f_1=(1/f) noise corner ($<f_0/2Q_L$)=resonator half bandwidth, kT=thermal noise floor=–174dBm/Hz, F=noise figure of the circuit, P_0=output power, and Q_L=loaded Q.

Figure 1b illustrates this phase-noise characteristic. It is worth noting that the crystal itself can contribute noise exceeding that predicted by Leason's equation. Noise is generally attributable to contaminants in the crystal. Proper crystal manufacturing techniques (in a clean room) and cold- or resistive-weld sealing techniques minimise the problem.

Inspection of equations 2 and 3 suggests that for minimum phase noise over the entire tuning range, $Z_f(f)$ must have an imaginary part with a large slope and a real part that remains small.

Ideally, oscillator loop gain remains constant over the tuning range, ensuring oscillator start-up and minimal AM noise over the tuning range. Constant loop gain can be realised by selecting $Z_f(f)$ such that it has a small and fairly constant real part over the tuning range. A constant Q_L is also desirable because it provides consistent noise performance over the tuning range and also simplifies design of the tuning network. For constant loop gain and optimal noise performance over the tuning range, a $Z_f(f)$ is needed that has a large slope, an imaginary part that is fairly linear, and a real part that is small and fairly constant.

One final requirement must be observed for stable oscillation – the imaginary part of $Z_f(f)$ must equal zero at only one frequency over the entire frequency range in which sufficient loop gain exists for oscillation. To meet this requirement, a crystal must have a monotonic reactance-vs-frequency characteristic.

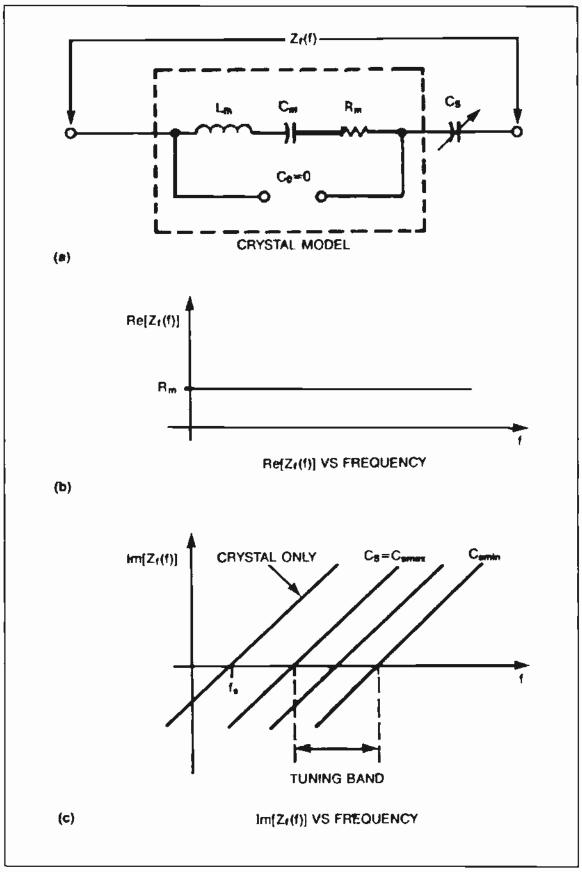


Fig. 2. Because a series tuning capacitor can only raise the oscillation frequency above the series-resonant frequency of the crystal, oscillators using this configuration are only tunable to frequencies above their crystal's series-resonant frequency.

Frequency determining network

To attack the design problem, first consider an ideal (C_0=0) crystal connected in series with a variable tuning capacitor (**Fig. 2**). As Fig. 2 illustrates, this network meets all the requirements. The variable tuning capacitance C_S shifts the crystal reactance down by a variable amount that is essentially independent of frequency, over a small fractional frequency range. Oscillation occurs at the point where the shifted reactance curve crosses zero. A straightforward analysis yields the equation:

$$\frac{\Delta f}{f_0}=\frac{C_m}{2C_{s\min}}\left[1-\frac{1}{C_R}\right] \qquad 4$$

where Δf is the tuning range and C_R is the tuning capacitance ratio (C_{smax}/C_{smin}).

Clearly, a series tuning capacitor can only shift the oscillation frequency to a value above the series-resonant frequency of the crystal. Therefore, most VCXOs (voltage-controlled crystal oscillators) using this configuration are tuneable to a frequency above the series resonant frequency of their crystals. Tuning frequency can be lowered by adding an inductor in series with the tuning capacitor. Although adding inductance helps avoid the spurious modes that typically occur at frequencies above the crystal's series-resonant frequency, the addition introduces susceptibility to magnetic-pickup problems.

Consider a network (**Fig. 3**) where the ideal crystal's C_0 is greater than 0pF. Clearly, this circuit fails to meet two of the previously cited requirements – the real part of Z_X is not constantly low, and the imaginary part is not linear. The effects of C_0 can be removed by adding an appropriate inductor L_0 to form a parallel-resonant circuit. Unfortunately, L_0's value cannot be selected based on measured or specified values of C_0, because the crystal model is not adequate for difficult applications.

The best design approach is to measure $R_e|Z_X(f)|$ and $I_m[Z_X(f)]$ over the entire tuning range. Then mathematically (or physically) add enough parallel inductance to satisfy the design requirements. Although, in theory, it is possible to meet, precisely, the crystal's requirements, in practice the crystal could deviate substantially from any modelled performance. But the approach does allow quick evaluation of crystal prototypes in environments that mirror the actual application – a real time saver.

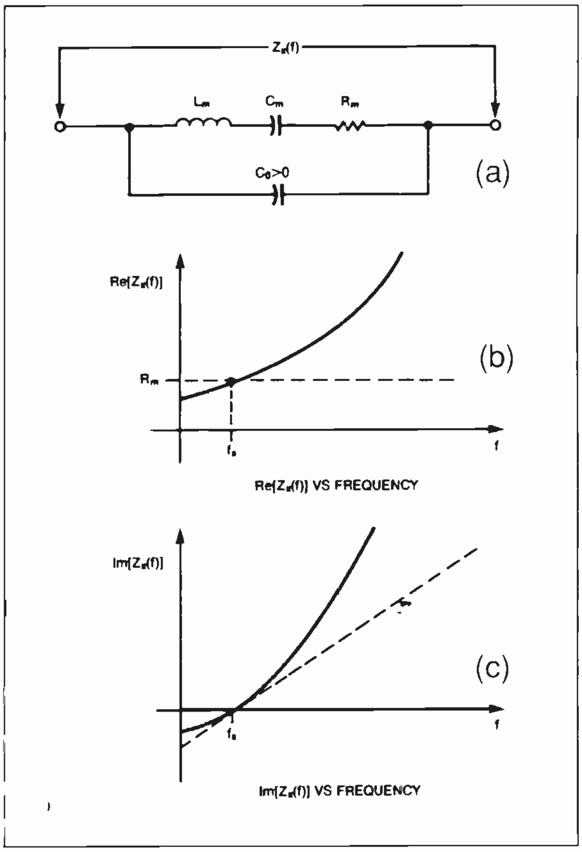


Fig. 3. Tuning range is limited (and oscillator noise is inconsistent) when the ideal crystal's C_0 is greater than 0pF, because the real part of Z_X is not constantly low and the imaginary part is not linear.

80MHz VCXO

An 80MHz VCXO for use in a phase-locked loop must accommodate a ±15ppm absolute error in its reference frequency. Typically, low-cost crystals specify a ±5ppm absolute frequency error, a ±16ppm aging error over ten years, and a ±10ppm variation as a function of temperature (over an industrial operating range). To accommodate all these error sources, the oscillator must have a ±45ppm tuning-range capability – that is, Δf=90 ppm. This is considered a very wide tuning range for a low-noise, 80MHz VCXO.

In some cases, two devices could be used to develop the required overall tuning capacitance – a mechanical trimmer capacitor for crystal-frequency error and aging, and a varactor to satisfy the remaining tuning requirement. In this example, however, the design uses a varactor for the entire 90ppm tuning range. Equation 4 shows that a wide tuning range requires a large varactor capacitance ratio (C_R), a small C_{smin} and an appropriately large crystal motional capacitance C_M. Increasing C_M may decrease the unloaded-crystal Q, so some trade-offs must be made to realise the optimum combination.

To obtain a large varactor C_R, use a hyper-abrupt diode. For a hyper-abrupt VHF varactor like the *BB105*, C_R≈5.5 and C_{smin}=2.2pF. From equation 4, therefore, C_M must be at least 0.5fF to ensure Δ=90ppm. The mean frequency (80MHz by design in this case) over the tuning range is:

$$\bar{f} = f_0\left[1+\frac{C_m}{4C_{smin}}\left(1+\frac{1}{C_r}\right)\right] \qquad 5$$

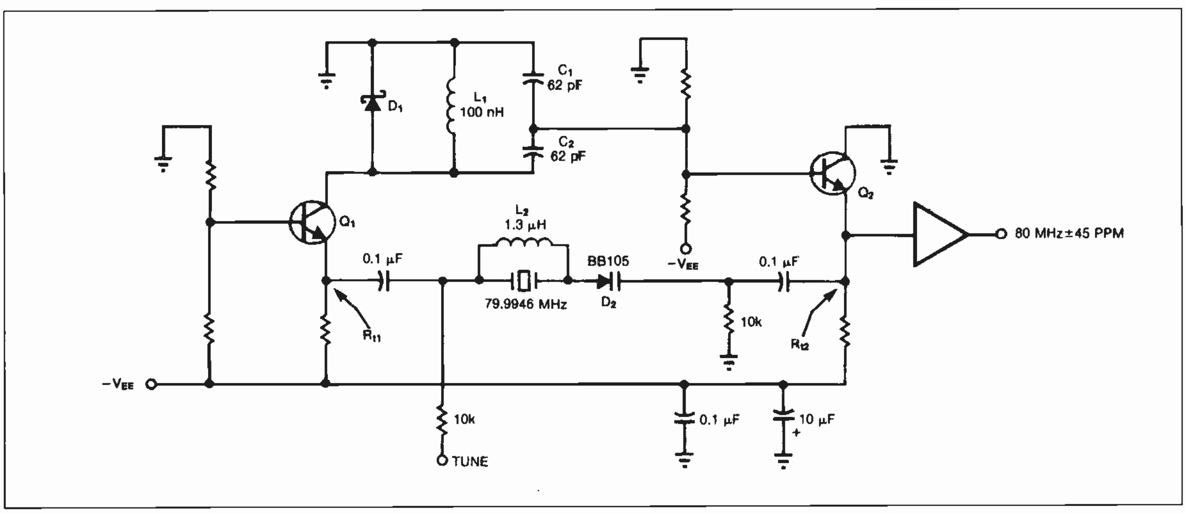


Fig. 4. C_1, C_2 and L_1 combine to set the desired crystal harmonic frequency in this two-transistor feedback oscillator. The network comprising the crystal, L_2 and D_2 determines the oscillator frequency.

From equation 5, the crystal's series-resonant frequency comes out to 79.99462MHz, or 67ppm below 80MHz. The crystal must have no spurious modes or aberrant characteristics over the tuning range – between 22 and 11 ppm above the series resonant frequency.

This design is based on the properties of an ideal crystal. In the real-world, a real crystal must be measured to determine the tuning-band measurements of the frequency-determining network. That data is then used to select the correct varactor and a crystal with the proper frequency.

To add detail to the theory, **Fig. 4** illustrates a transistor feedback oscillator that effectively implements the block diagram of Fig. 1. L_1, C_1, and C_2 combine to set the desired crystal harmonic frequency. D_1, isolated from the crystal, provides amplitude limiting. D_1's impedance decreases with signal level, a drop which reduces Q_1's voltage gain and thereby implements the G_1 variable gain block of Fig.1. The network, comprising the crystal, L_2, and D_2, determines the oscillator's frequency. Q_2 isolates the tank from the crystal and provides a low-impedance termination for the crystal's closed loop, composed of R_{t1}, the crystal, D_2, and R_{t2}. Q_1 supplies the voltage gain necessary for oscillation and also provides a low impedance termination for the crystal loop.

To start, adjust the oscillator's frequency-determining network to run at 0° phase; substitute an AC-coupled resistor whose value is equal to the typical value of $R_e[Z_f(f)]$ over the tuning range. Then select C_1 and/or C_2 to establish the oscillator's desired operating frequency – 80MHz in this case – and complete the design by replacing the resistor with the frequency-determining network.

When physically adding the frequency-determining network, stray admittances should be minimised. This can be done by removing the ground plane from beneath the frequency-determining network and using large impedance values for the varactor's bias resistors. Note that some nodal impedances in the frequency-determining network are very high – 900Ω in this case. To maintain this high impedance at high frequencies, pay strict attention to the physical layout.

At this point, the oscillator's tuning range should be very close to the design goal. In addition, noise performance will be fairly constant over the tuning range.

Several mechanisms will act to limit attempts to achieve greater and greater oscillator tuning ranges. Spurious operating modes are inevitable with attempts to tune further away from the crystal's series-resonant frequency. Other crystal characteristics can also cause problems.

Crystal reactance, for example, may become non-monotonic with frequency and thus cause unstable tuning. Maintaining high impedance at high frequencies is another problem area. The crystal's high-impedance node has a design impedance of:

$$Z(f) \cong R_{t1} + R_m + j\frac{f - f_5}{\pi f_0^2 C_m}$$

Stray capacitance and the varactor-bias ports will eventually limit attempts to increase the design impedance. In addition, accuracy of measurements of the frequency-determining network decreases as attempts are made to develop very high impedance levels.

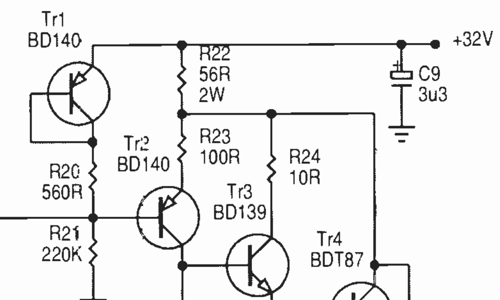
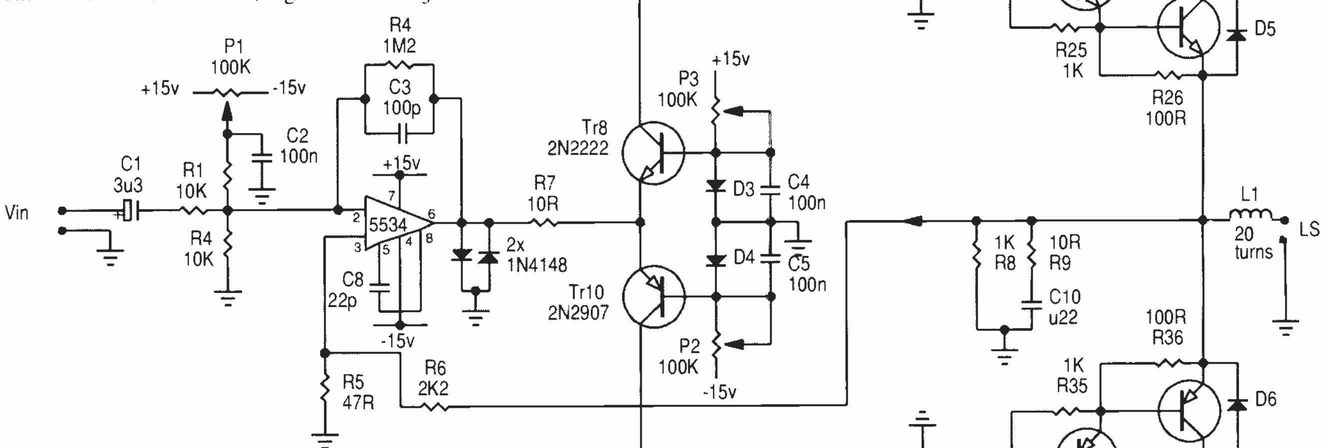
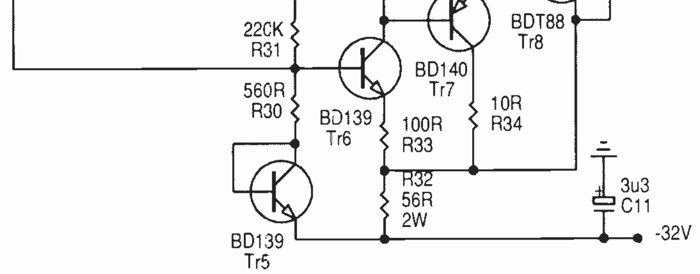
For example, using an *HP3677A* vector network analyser and an *HP356P77A* s-parameter test set, measurements will have adequate accuracy for impedances between 0.5 and 1000Ω. ■

Tim L Hillstrom, Hewlett-Packard Co

50W "Blomley"

Stirred into action by the letter from W Groome in the January issue, Hans Hartsuiker presents his own design, which is based on the Peter Blomley amplifier from 1971. Mr Hartsuiker built this one in 1984 and has been pleased with its performance on Quad *ESL-63* electrostatic speakers. It is based on the original proposition that Class-B switching should be done early, rather than in the output stage.

To take advantage of advancing op-amp design and to reduce the component count, the low-noise *5534* is the gain stage, R_4 and $C3$ determining its gain and frequency response; the two diodes protect $TR_{9,10}$ during clipping. Transistors $Tr_{9,10}$, which must be switching types, compose the phase splitter to drive the output triplets formed by $Tr_{2,3,4}$ – "a supertransistor" with the load in the collector which, together with Tr_3, takes the form of a

Amplifier employing the ideas put forward by Peter Blomley 20 years ago, in which phase splitting is carried out in an early stage. Output transistors are permanently conducting.

current mirror. Inductor L_1 and the filter at the output reduce the possibility of RF into or out of the amplifier.

Points in the design's favour are that no emitter followers are at the output, with capacitive loads in mind; that all the bipolar transistors are current amplifiers; and that the output-stage devices are always on. Against, there is the onset of hard clipping; the fact that this is an inverting amplifier; and overall gain is about 48, chosen for stability; feedback factor about 200.

To set up, adjust $P_{1,3}$ to maximum; set the output voltage to zero by P_1; set output current to 60mA by P_3 (33mV across R_{22}); adjust P_2 to the point where the output current starts to increase.

As regards construction , the usual precautions should be observed and the need to mount $D_{3,4}$, $Tr_{3,4,7,8}$ on the same heat sink.

Hans Hartsuiker
Eindhoven
The Netherlands.

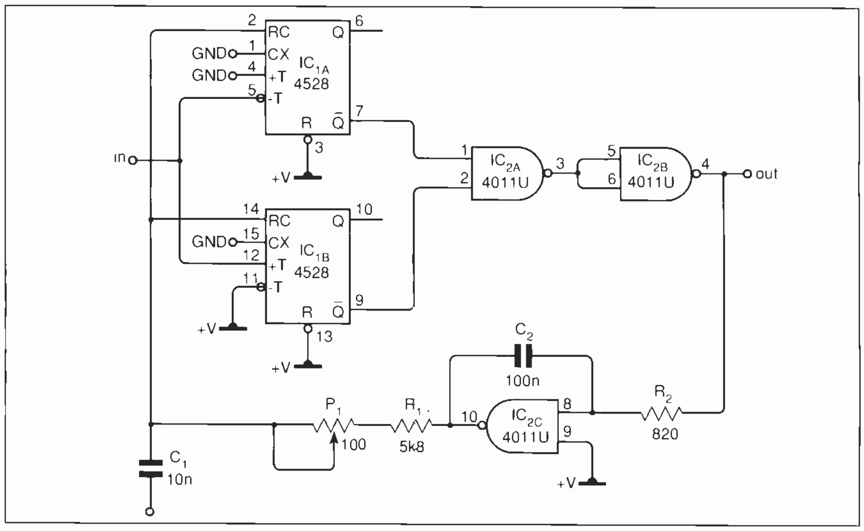
Frequency doubler

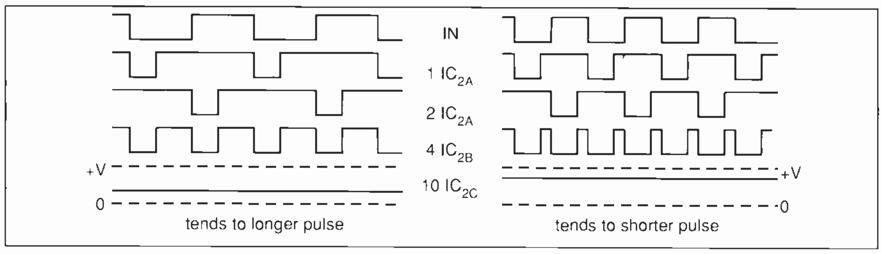
This circuit will provide frequency doubling over a wide range of frequencies.

Two monostable flip-flops contained in a *4528* use the same timing circuit – P_1, R_1 and C_1 in the circuit diagram. They trigger on opposite edges of the input waveform and, working with the Nand IC_{2a}, give an output pulse on each transition of the input. Low time constants from the *4528* allow a large frequency range.

Adding the inverter IC_{2b} and integrator IC_{2c} as shown permits adjustment of the duty cycle by P_1 to give symmetry. With component values as shown, the input can vary between 300Hz and 8700Hz, and ±20% without affecting output mark:space ratio once it has been adjusted.

W Dijkstra
Waalre
Netherlands





Frequency doubler working over a wide range (300-8700Hz in the case shown) and with adjustable mark:space ratio, which is constant over ±20% frequency change.

Signals in chaos

Using a bipolar transistor as a feedback element as in **Fig. 1**, in conjunction with a few passive components, produces a logarithmic function. Since the circuit is unstable, it is usable in a "chaotic" signal generator to give an output which, although not white noise, has a similar nature when heard.

Figure 2 shows such an arrangement. The output from the Fig. 1 circuit is delayed by the phase shift of the filter and fed back to the input, whereupon a continuous oscillation is set up. Varying R_5 alters the characteristic of the basic block; some settings give a periodic waveform at X, but most give a non-periodic output. Fine adjustments are possible by $R_{6,7}$, which may be omitted altogether, if desired. When set to give a "chaotic" output, the circuit is well-behaved in a chaotic sort of way and gives a continuous signal. Its behaviour becomes clear if X and Y signals are applied to the X/Y plates of an oscilloscope.

Figure 3 is a simpler circuit, but the inductor values need selection by trial and error.

D Ayers
Buxton
Derbyshire

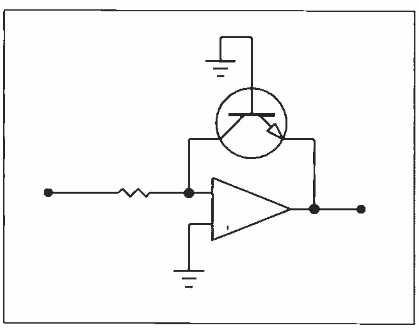


Fig. 1. Circuit often described as a logarithmic amplifier, which is inherently unstable.

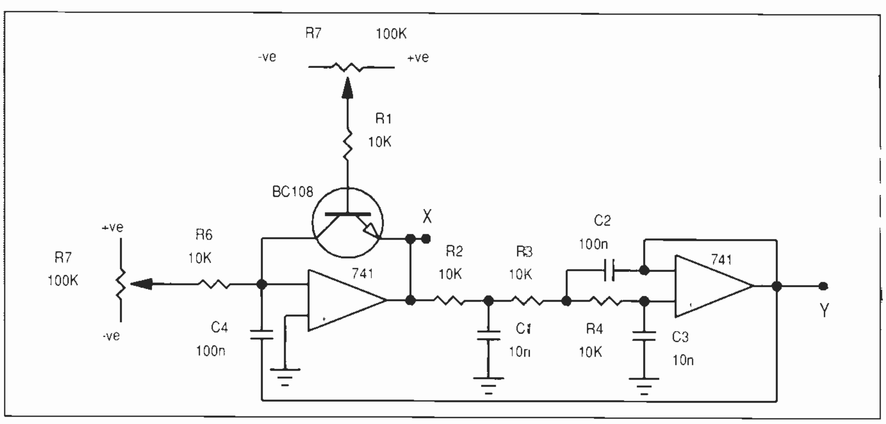


Fig. 2. Chaotic signal generator uses Fig. 1 circuit and phase-shifted feedback.

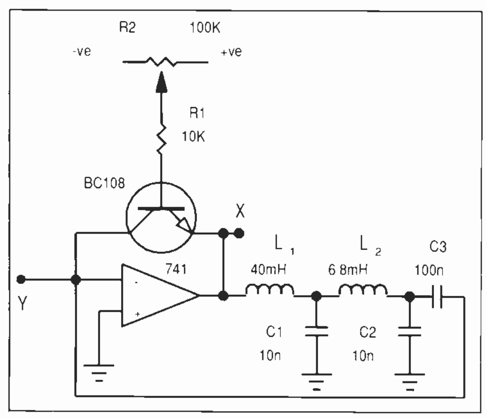


Fig. 3. Simpler circuit gives same result as Fig. 2, but some inductor tweaking might be needed.

Constant-gain tuned filter

In the classical state-variable filter, a summing amplifier is followed by an integrator train, the input, output, integral and differential of the output being summed at separately weighted inputs to give the required response. In the circuit of **Fig. 1**, a variable element is common to both input and output terms, so that the ratio of the two is fixed at the resonant frequency at any circuit Q.

Frequency-determining components are able to take a wide range of values, since they are isolated from the rest of the circuit and do not interact with its Q value.

High-pass and low-pass characteristics are at A_1 and A_3 outputs respectively; if a non-inverting, high-input-impedance amplifier with a gain of 4 is connected to the junction of the T network, a high-Q notch is produced.

Figure 2 shows a second-order Butterworth high/low-pass filter, based on the previous circuit but with a fixed-gain summing amplifier. Both characteristics are provided and a band-pass filter of maximum gain $1/\sqrt{2}$ is obtained from A_2. To make a notch filter, feed either of the A_1 inputs to a buffer amplifier of gain $1+\sqrt{2}$. The summing amplifier gain is accurately set using preferred-value resistors; for example, 3.6kΩ and 5.1kΩ give a gain of $\sqrt{2}$ within less than 0.2%.

John D Yewen
Leighton Buzzard
Bedfordshire

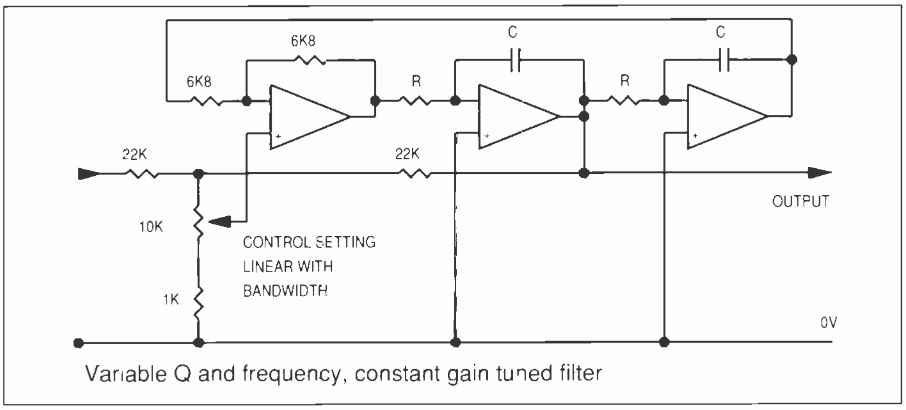


Fig.1. Constant-gain tuned filter with a wide frequency range and variable Q. High and low-pass characteristics are available and a band-pass type results from a further amplifier.

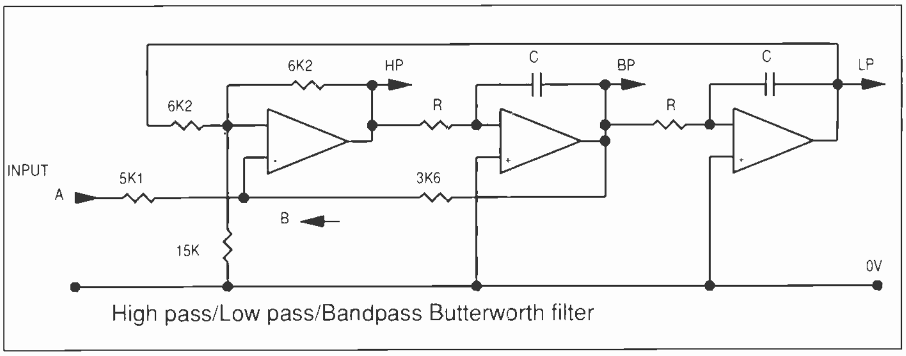


Fig.2. Variation of the Fig.1 circuit to give a second-order Butterworth high and low-pass, band-pass and notch filter outputs.

RS-232C monitor without power supply

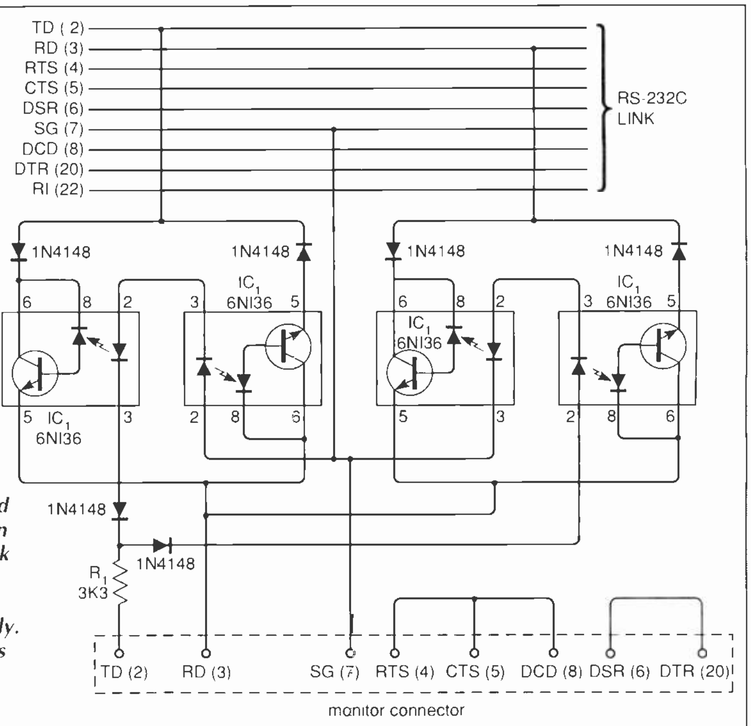
With a little work on the RS-232C serial port found on many PCs, one can monitor received and transmitted data.

Two bidirectional switches form the circuit, each composed of two opto-couplers connected in reverse sense and two diodes to connect the data to the correct switches. One pair sees the TD line of the port, while the other looks at the RD line. The TD monitor is activated when pin two of the monitor connector (TD) is low, the second pair when pin two goes high to monitor RD. Since the opto-coupler leds are in series with each other and with R_1, current from pin two of the monitor connector is only about 2mA, so no power supply is needed. *6N136* couplers were chosen for low drive-current needs.

The monitor connector is an equivalent to a null modem and a microcomputer set to the same baud rate as the link will monitor the TD or RD lines. If baud rate of the monitor is at least twice that of the RS-232C link, full-duplex monitoring is possible.

Frantisek Michele
Brno
Czechoslovakia

Circuit to monitor transmitted and received data from an RS-232C link without the need for a power supply. Circuit takes 2mA from monitor.



ACTIVE

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

Digital colourspace converter/corrector. Three parallel 12-bit colour components can be processed at up to 40MHz by the *TMC2272* from TRW LSI Products Inc. The device is said to be the first monolithic 36-bit colourspace converter/corrector, which converts moving camera, video, monitor and PC graphics and images in real time. It converts any set of three colour components into any other set with nine user-programmed coefficients stored on-chip, which may be varied to allow pixel-by-pixel correction. Ambar Components Ltd. 0844 261144.

Digital trim for op-amps. *Max528/9* are multi-D-to-As, in which eight converters and eight buffers with their serial control interface are collected on one chip. Three operating modes include unbuffered, in which output buffers are bypassed in the presence of high-impedances to reduce power and errors; full buffered, in which buffers are in circuit to provide +5mA/-2mA load currents; and half buffered for single-supply working and when +5mA output is needed. A shutdown pin drops supply current to 50μA with no loss of data. Maxim Integrated Products Ltd. 0734 845255.

Discrete active devices

Dual pin diode. Diodes in the *Bar64* series from Siemens are available in single and four-dual versions. They have a reverse voltage rating of 200V and are meant for RF attenuation and switching at frequencies between 1GHz and 2GHz. Total capacitance is 0.23pF with reverse 20V at 1MHz, and forward resistance at 100MHz is 0.85Ω. Charge carrier lifetime is 1.55μs. Siemens plc. 0932 752323.

Digital signal processor

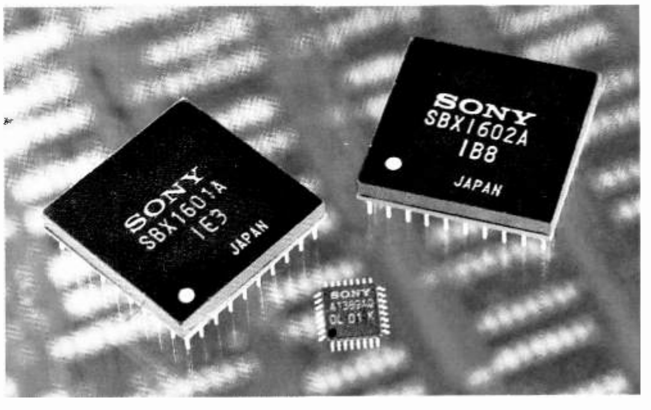
FIR filters. Harris's *HSP43168* dual finite impulse-response filter performs multiple functions to outperform asic designs in flexibility. It contains two 8-tap FIR filters, configuration control and storage of up to 256 programmable coefficients. Available operating modes include high, low and band-pass, complex, 2D convolution, interpolation and decimation. Further, it can be used as a building block in adaptive and polyphase filtering. Harris Semiconductor UK Ltd. 0276 686886.

Linear integrated circuits

Low-voltage regulators. Toko's *TK114/5* are regulators with a voltage drop of 0.2V and are meant for battery-powered equipment that lives most of the time in a state of suspended animation, taking only 0.1μA when the control terminal is at off. They are available in ten voltage ratings of 2-5.5V and output currents of 180mA, although best under 100mA. Dissipation is up to 200mW (114) and 600mW (115). Cirkit Distribution Ltd. 0992 444111.

Low-power op-amp. *LTC1047* is a dual chopper-stabilised op-amp that takes only 60μA supply current per amplifier. Offset voltage is 10μV and drifts 50nV/°C maximum. No other components are needed, since sample-and-hold Cs are on-chip. Common-mode and power-supply rejection ratios are 110dB. Since it is pin-compatible with standard dual op-amps, it can be used in place of practically any precision device. Linear Technology UK Ltd. 0276 677676.

Digital video interface. Three chips from Hakuto allow existing coaxial cable to handle digital data at 200-300Mb/s. *SBX1601A* (encoder), *SBX1602A* (decoder) and *CXA-1389AQ* (cable driver) can also be used for strip lines on boards, twisted pairs for board-to-board transmission and optical



fibre for distances. The encoder converts 10-bit parallel data and clock to serial emitter balance pair output, with PLL lock detection and sync. word generation; the cable driver is a video amplifier with one input and three 75Ω outputs; and the decoder converts back to 10-bit parallel form. Hakuto International (UK) Ltd. 0992 769090.

Logic building blocks

5GHz dividers. GPS's range of fixed-modulus dividers now includes the SP8900 series, with ratios of 2,4,8,10 or 16 at up to 5.5GHz (A grade) or 5GHz (B grade). Input sensitivity is -2dBm and, since the devices are in the HE bipolar process, power needed is only 350mW from one 5V supply. The process also confers a noise floor of -144dBc/Hz — much less than GaAs techniques. B grade 1000-off price is £18; A Grade £33. GEC Plessey Semiconductors. 0793 518000.

Memory chips

3V eeproms. Two eeproms from Hitachi, *HN58V257* (256K) and *HN58V1001* (1M), are guaranteed to work with a 3V power supply. Both are byte-wide devices in cmos and accept 2.7-5.5V, taking 10mA and 15mA respectively, with access times of 250ns and 200ns. A new i/o circuit allows direct interfacing with other cmos devices. Hitachi Europe Ltd. 0628 585000.

Burst-mode eprom. Am27HB010 is claimed by AMD to be the fastest burst-mode eprom in *captivity*, having a random access time of 50ns and burst access time of 15ns. It affords unlimited sequential access; burst mode can be entered without regard to page or word boundaries up to 128kbytes. Kudos Thame Ltd. 0734 351010.

Microprocessors and controllers

Televison controller. Zilog's *Z86127* is a low cost, 8-bit digital TV controller, meant for consumer use, which has a 1.5μs instruction pointer with stop and halt modes. A mask-programmed rom facilitates volume production of custom-programmed devices. It supports on-screen display of 8 rows by 20 characters in 5 by 7 or 11 by 15 resolution. PWM ports provide for voltage synthesiser tuning and control of audio and picture levels. Gothic Crellon Ltd. 0734 788878.

Microcontroller with largest memory. Hitachi's *H8/500* series of 16-bit cmos single-chip microcomputers has a new member — the *H8/536*, which possesses 62Kbyte of rom/prom/eprom and 2Kbyte of ram, claimed to be the largest on-chip memory in mass production. There are several timers for ease of measurement and control, three channels with two comparator outputs and input capture register, an 8-bit timer for event counting, a three-channel PWM timer and a watchdog timer. An eight-channel A-to-D converter is on-chip, as is a data transfer controller. Hitachi Europe Ltd. 0628 585000.

4-bit microcontroller. Toshiba's *TLCS47E* series of 4-bit single-chip microcontrollers are small in both size and cost. They are compatible with the *TLCS47/470* families, the first being the *TMP47C241*, which runs at 4.2MHz, has 2Kbyte of rom, 128 by 4 bits of ram, 21 i/o pins, a 5-bit led driver, 22-bit interval timer, two 12-bit counters, watchdog timer and A-to-D converter. Toshiba Electronics (UK) Ltd. 0276 694600.

Mixed-signal ICs.

Voice messaging by PC. A send and receive fax chipset from Sierra will also digitise telephone voice signal and feed it to a PC in data form, the PC providing a greeting to the caller. Voice data is 8-bit μ-law at 9600 or 4800 sample/s and the chipset behaves as fax, data modem or answering machine; the normal AT command set is expanded to handle the voice system. Using other modem chips, a quattro modem with fax and the voice messaging can be built. Demo boards are available from Amega. Amega Electronics Ltd. 0256 843166

Hall-effect IC. From Seiko Instruments, the *S-8143A* cmos Hall-effect IC drives TTL and cmos directly. It is controlled by a small magnet of 100gauss for low-to-high and 400gauss for high-to-low output transitions. Supply requirement is 4.5V to 16V at "low current" and the device is said to be ideal for non-contact motor switching. Amega Electronics Ltd. 0256 843166

Dual high-side driver. *LCT1155* is a mosfet dual driver delivering up to 12V gate drive from a 5V supply. Both channels are short-circuit-protected, gate drive being rapidly removed in the event of increased drain current and held off until the input is recycled. Its internal voltage multiplier needs no extra components and the chip takes only 8μA from the supply, dropping to 8μA when inputs are at off. Linear Technology Ltd. 0276 677676.

Continuous-time active filters. Two low-noise, continuous-time active filters from Maxim, *Max275/274*, contain two and four independent second-order sections respectively, which are configurable as fourth or eighth-order filters, with centre frequencies from 100 to 300kHz. Sections will implement Butterworth, Bessel or Tchebyshev responses, centre frequencies with programmable Qs. There is no clock, so no clock noise or aliasing. Noise floor is 120μV RMS and dynamic range 92dB. Frequency setting is guaranteed to within 1%. Maxim Integrated Products Ltd. 0734 845255.

Shrunken chips. Since PCs are being made smaller by the day, Sierra decided to match the shrinking process by condensing its modem chipsets into plastic quad flatpacks. For example, modem advance controllers, normally in 68-pin PLCCs, are now in 80-pin PQFPs, saving more than 65% in area. "A V42bis modem with fax occupies the same size as two Christmas stamps side by side", says Sierra. Sierra Semiconductor Ltd. 0793 618492.

Optical devices

Sensitive CCD imager. Twice the sensitivity of the *ICX039AKA-6* interline transfer CCD imager is offered by the Sony *ICX039BNA-6*. It possesses the features of its predecessor in filtering, shutter and the hole accumulation diode sensor, and also has a floating diffusion amplifier in the output stage which allows a reduction in lens size without loss of brightness and therefore a smaller camera.Sony Components. 0784 466660.

Power semiconductors

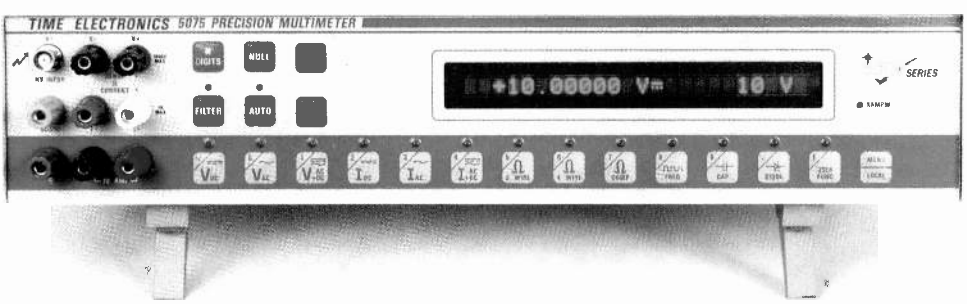
Low-power regulators. High-current, low drop-out micropower voltage regulators from National Semiconductor provide 250mA output as against 100mA for earlier versions. *LP2952/3/4* give output voltage accuracy within ±1%, stabilisation ±0.2% and regulation ±0.2%, with "A"-grade devices offer ±0.5%, ±0.1% and ±0.16%. Pin-strapping produces a 5V output or, with external resistors, between 1.24V and 29V. There is a low-battery flag and a "snap-on/off" feature to allow operation only when input voltage is high enough. Jermyn Distribution. 0732 740100.

Power mosfet driver. Design of low-voltage, switched-mode conrtrollers using n-channel power mosfets is considerably eased by the Linear Technology *LT1158* half-bridge driver, which has synchronously controlled high-side and low-side drivers working from 0 to 100kHz. One pin controls two n-channel power mosfets in totem-pole, adaptive non-overlap gate drives, eliminating mosfet matching. A 3000pF capacitive load can be switched in 150ns; 10000pF in 250ns. There is protection for output shorts and under/overvoltage. Micro Call Ltd. 0844 261939.

Isolated power mosfet. An n-channel, enhancement mode, power mosfet from Toshiba, the *2SK1365*, is meant for high speeds and high currents and is encapsulated in epoxy resin with no exposed metal surfaces. It is rated up to 1kV and is therefore suitable for 240V AC switched PSUs. Current rating is 7A and on resistance 1.5Ω. Toshiba Electronics (UK) Ltd. 0276 694600.

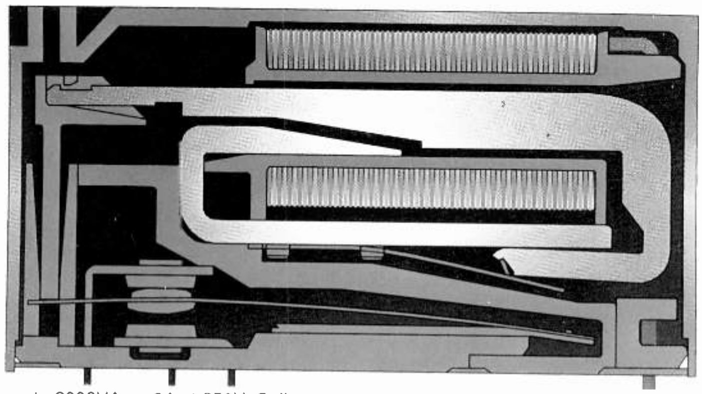
PWM controller. *UC3856*, a current-mode PWM controller by Unitrode, is an improved version of the earlier UC3846 with higher speed and better accuracy, reduced input-to-output delays and reduced noise sensitivity. It offers a low-impedance TTL-compatible sync. output with a tri-state function when used as a sync. input, fast 1.5A peak output for rapid power fet switching and 2kV minimum ESD protection on all pins. Unitrode (UK) Ltd. 081 318 1431.

Multimeter. 5075 is claimed to be the first seven digit multimeter able to measure nV to 10kV; pA to 30A; nΩ to GΩ, capacitance from 1pF to 400μF at 0.2% accuracy, frequency and temperature – to give 80 ranges for under £2000. 1-3-10 range sequence means measurements need never be less than 30% full scale. Digital auto-dynamic filter tracks the input, allowing large changes to be followed immediately while noise is filtered out. Fluorescent display can be set to different modes including dual display where two parameters eg AC volts and frequency can be displayed simultaneously. Time Electronics Ltd, 0732 355993.



Switches and relays

Miniature relay. *FEME M15* is the first of a family of efficient miniature PCB-mounting relays by Carlo Gavazzi, having only two ferromagnetic parts. It is smaller than usual (29 by 10 by 15mm), but is provided with a coil in the range 1.5-110V DC and switches up to 2000VA, or 8A at 250V. Coil power consumption is as small as 100mW, due to the fact that the armature is operated by a push rod from the coil centre to give a 50% power saving. Carlo Gavazzi Electromatic, 0252 29324.



PASSIVE

Passive components

Ceramic discs. Very small ceramic disc capacitors from Murata in the *DD003* range are rated at 100V in values between 1pF and 22,000pF, with a 47,000pF type in a 12V rating. NP0, SL, 2B4 and 2F4 dielectrics are available. ESD Mercator, 0493 844911.

Tantalum chip capacitors. Matsuo type *269* capacitors are equipped with an internal fuse to eliminate resin decomposition caused by shorts or thermal overload. They come in capacitance values of between 1μF and 68μF and in ratings of 10-35V, all with a tolerance of ±20%. Operating temperature is from -55 C to 125 C. Murata Electronics UK Ltd. 0992 444111.

Connectors and cabling

Filtered connector. MDMT multi-pin D connectors from ITT Cannon incorporate transverse monolithic filters to eliminate noise between 10MHz and 1GHz. They also use a twisted pin form having seven-point contact to withstand vibration. The one-piece aluminium shell is sealed with silicone-based rubber and insulated by glass-filled diallylphthalate. Four standard capacitance values allow matching to most filter requirements. ITT Cannon, 0256 473171.

Dual-readout PCB sockets. A 0.05in pitch PCB socket from Methode will take the latest double-sided SIMM/SIP memory and SIP sub-circuit modules. Vertically mounted sockets come with up to 100 contacts for single-in-line modules. Methode Electronics Europe. 0535 603282.

Coax. terminations. A novel type of termination for coaxial cables from Verospeed uses crimping to effect a mechanically sound connection between the braid and terminal. Tab ferrules, either surface mounted or with tapered legs to afford an interference fit for wave soldering and with one or two legs, are crimped, the coax. inner forming the other terminal itself. A variety of forms are available for popular cable types. Verospeed, 0703 641111.

Displays

Dual bicolour leds. Series *550-3007-220* displays contain two 5mm leds which each have a red and a green chip, so that each led shows two bits of information individually or as four bits simultaneously. Temperature range is -55°C to 100°C. Dialight Corporation, 0638 665161.

Instrumentation

Phase-angle voltmeter. Two phase-angle voltmeters from Inertial Aerosystems are digital and microprocessor-controlled, but are totally fuss-free: Models *4001C-1/2* offer measurement at the touch of one button. They operate from 10Hz to 25kHz and 26Hz to 54kHz respectively and need nothing extra for phase-angle measurement, synchro/resolver testing, in-phase and quad voltage measurement, amplifier test and the measurement of impedance angle. Phase accuracy is within ±0.5 and voltage measurement within ±2% FSD over the frequency range. Model *4001C-2* also provides for ratiometric measurement for direct indication of amplifier gain, line loss or transformer turns ratio. Inertial Aerosystems Ltd, 02518 2442.

6GHz oscilloscope. The new *TDS 820* from Tektronix is a 6GHz, two-channel digitiser with 0.4ps timing resolution and graphical user interface, priced at £14,986, said by Tek to be the lowest ever cost for an instrument of this type. The TDS 820 is meant to fill the gap between existing high-level digitising 'scopes and lab.-type research equipment, in which high-speed logic lies. It possesses a 14-bit A-to-D converter and digital error correction and includes two delay lines with trigger pick-off and a hold-off control. Trigger bandwidth is 2GHz or 8GHz without the delay lines. As is usual with Tek instruments, it is not possible to describe it fully here. Tektronix UK Ltd, 0628 486000.

Small oscilloscopes. Should anyone need to carry an oscilloscope around in a briefcase, one of the *LBO-315* range would probably fill the bill. Using 95mm tubes and either 2kV or 12kV acceleration, depending on the model, they have bandwidths of 60MHz and 20MHz. Y sensitivity is 5mV/6.35mm division, with a five times multiplier. All the usual functions are offered, with a delayed sweep on the 60MHz models. Power is from the mains, external DC or by the NiCad pack supplied. Thurlby-Thandar Ltd, 0480 412451.

Interfaces

RS232-to-current-loop converter. Amplicon's *CM200* is a plug-in unit to convert serial data from RS232 to 20mA current-loop form. It works in full duplex up to 9600baud in links up to 8.5km, is mains-powered via a supplied adaptor and uses opto-isolated current-loop transmitters and receivers. The RS232 interface can be either DCE or DTE and the loop active or passive, modes being switchable. Amplicon Liveline Ltd, 0273 608331.

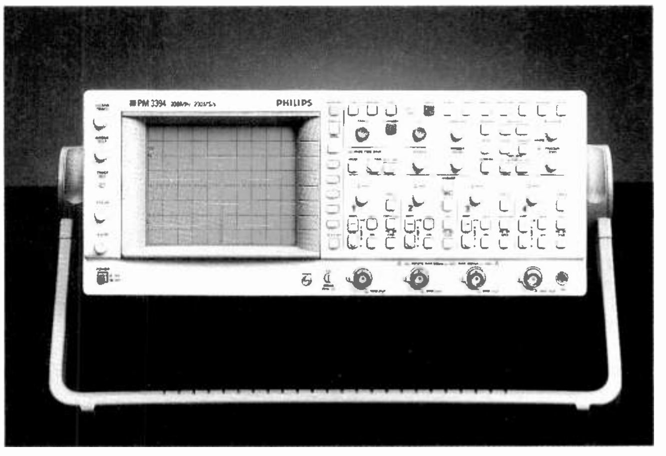
Literature

Hitachi catalogue. Five sections include research and development, information systems and electronics, power and industrial systems, consumer products and materials in 84 pages. "An encyclopaedia of the state of the art", says Hitachi. Hitachi Denshi (UK) Ltd, 081-202 4311.

Interference shielding. IVC's latest catalogue describes a range of RFI/EMI shielding and ESD protective coatings. It provides an introduction to vacuum coating, including sputter coating and the Elamet aluminium deposition process for a variety of substrates. Applications include transparent RFI/EMI shielding and reflective coatings. The company also has available data sheets giving detailed information. Inco Vacuum Coatings, 021-511 1115.

Wire and cable. Carol Cable is an American company making wires and cables and is perhaps the only company of its type so comprehensively in control of its affairs that it doesn't even buy its copper — it mines it. Verospeed now has the UK handling of the company's products and includes them in its new catalogue. Verospeed, 0703 641111.

Analogue/digital oscilloscopes. While preserving the familiar operation of an analogue instrument, the Philips *PM3394* series also offers the advantages of a 200Msample/s digital storage type. The range includes 100MHz and 200MHz versions, costing between about £3500 and £6000,



and there is a "Math+" facility for integrating, differentiating and FFT analysis. All the usual functions of a high-level instrument are present and include a "touch hold and measure" button on the probe and an 8K memory, expandable to 32K, which holds 200, 500-sample traces. Philips Test & Measurement, 0923 240511.

Power supplies

Custom power supplies. Power supplies often receive scant consideration from equipment designers and consequently have to be fitted in wherever possible. To meet this lack of respect for the prime mover, Astec has produced the Rack Power System — a range of 19in units with a large variety of panels, outputs, connections and metering, built to customers' specifications. Linear and switching types are available. Astec Standard Power Europe, 0246 455946.

Programmable power. General-purpose, single-output supplies from Coutant-Lambdaare said to be useful for IEEE-488 application. Twenty models in the *LLS-GPIB* series of units work in the range 28-800W, offering constant voltage/constant current operation up to 120V or 100A. Remote sensing avoids lead loss and the output displays read the remote value. If the outputs vary outside set window limits, an alarm signal is sent to the host controller that normally programs the unit; a keypad is provided for manual operation, when needed. Coutant-Lambda Ltd, 0271 863781.

UPS with battery test. *UPS9000* series uninterruptible power supplies by Fiskar offer the facility of either manually or automatically testing battery condition. Every 30 days, a 25s discharge test is carried out automatically, during which the UPS works normally; anything untoward cancels the test for another 18h. In the manual test, batteries discharge to 1.6V per cell and the time is recorded. After this, the rectifier takes over again. Fiskars Electronics Ltd, 0734 772599.

Laboratory power supply. Incorporating both GPIB and addressable RS232 (ARC) interfaces, the *TSX35-10P* programmable power supply is a 35V, 10A unit, the ARC system allowing up to 32 supplies to be daisy-chained together for individual addressing and control from one RS232 computer port. In this way, voltage set, current set, OVP set, on/off and voltage/current read are all under remote control. The regulator in the *TSX35-10P* is a combined switcher pre-regulator and linear post-regulator. The former reduces input-to-output capacitance and, therefore, common-mode noise, while the latter gives low output noise and good regulation. Thurlby-Thandar Ltd, 0480 412451.

Radio communications products

Tx design kit. To assist designers realise low-power transmitters that meet the required standards, Quantelec has produced design kits based on the RF Monolithics RO2021 single-port quartz saw resonator for 418MHz. Since layout can be critical, camera-ready artwork is included, together with a parts list and help with assembly, alignment and tuning. The unit's 75kHz bandwidth makes the transmitters ideally suitaed to use in a wide variety of applications – such as car alarms. Quantelec, 0993 776488.

Transducers and sensors

Pressure transmitter. In the *634* series of transmitters from Dwyer, a new device provides for field adjustment of range from 0-100bar to 0-400bar. Other models have ranges down to 0-1bar with accuracies within ±2% of full scale, two-wire operation and 4-20mA output. The Bourdon tubes are beryllium copper and the housing is in nylon, filled with mineral and glass. Dwyer Instruments Ltd, 0734 753808.

Photosensor. A cylindrical photosensor with a range of up to 3m, the *MF18* from Matsushita is available in 5V to 264V AC/DC working versions and provides either n-p-n/p-n-p or 100mA fet output transistors. It works at temperatures of -25°C to 55°C and is available with a right-angle mirror. Matsushita Automation Controls 0908 231555.

Data communications products

Optical-fibre link. A microwave package with 18GHz bandwidth, the

Lasertron QLINK from Lambda, is meant for EMI/EMP-immune Ku-band radar, remote antennas and other high-speed analogue links. The 18GHz laser transmitter converts microwave signal into optical signal for transmission on a 9μm fibre, the receiver performing the reverse function. Typical link loss is less than 40dB. The diode laser is kept at a constant temperature for stability and there is automatic transmitter power control. Lambda Photometrics Ltd. 0582 764334.

Development and evaluation

Microcontroller emulation. Picmaster development system is a real-time emulation tool for the Microchip PIC 8-bit risc microcontrollers. It runs on a PC in Windows 3 at up to 20MHz. Real-time traces can be captured and displayed without stopping emulation, an unlimited number of breakpoints are settable anywhere in memory and registers may be displayed and modified. Arizona Microchip Technology. 0628 850303.

COMPUTER

Computer peripherals

Windows graphics board. Pixelworks have a graphics board intended to accelerate Windows 3.0 and 3.1. AT-based Whirlwin generates non-interlaced 1280 by 1024 resolution when used with Pixelworks's Windows 3 driver and supports 1024 by 768 in applications compatible with the IBM AI graphics interface, such as AutoCad and Microstation. It is claimed to be up to 385 times faster than Super VGA for raster operations and up to 30 times faster for line drawings. GST Ltd. 0531 631163.

Software

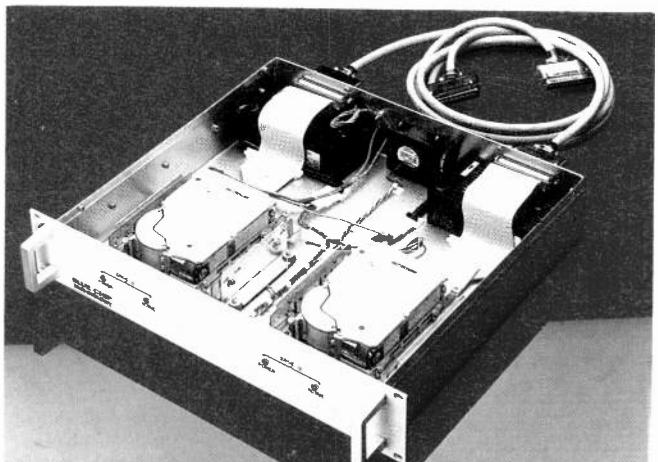
Lossy lines by PSpice. PSpice, the circuit simulator by MicroSim, now supports transmission-line modelling, adopting the distributed-model approach. Particularly significant in high-speed circuitry, one can simulate behaviour caused by attenuation and dispersion. The distributed model is of a complete line with distributed C.L.R and conductance, the behaviour being obtained using pulse responses. A speed increase is thereby obtained over the lumped model, which simulates segments of a line separated; a tendency to oscillate at transitions is also avoided. The company's Design Center suite of programs has also been extended to support the synthesis of passive, as well as active, filters. MicroSim Corporation, (USA)800 245-3022.

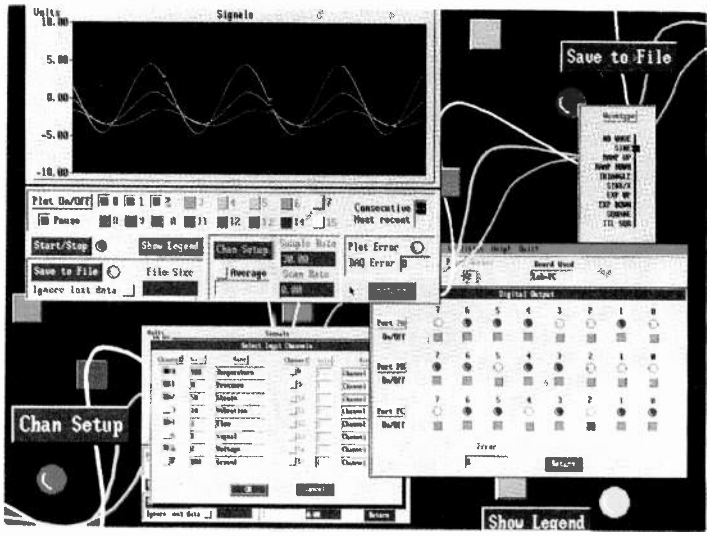
Satellite television. *Satmaster* is a PC program (Dos 3.0 or higher) intended for the design of satellite TV receiving systems. It finds dish angles from any global position, calculates a full link budget, including dish size, and plots beam width and lobe patterns to give an idea of possible interference. There is an on-screen guide and results can be printed. Swift Television Publications, 0793 750620.

All-Spice. Zetex is to provide a Spice model for every new device the company introduces. Already, 126 models exist for the more popular transistors, Darlingtons and variable-capacitance diodes, including some second-sourced devices, which Zetex consider are currently incorrectly modelled. Spice models are available free to equipment designers. Zetex plc. 061 627 4963.

Testing asic designs. Fast is a software package that enables ES2 to accept asic design from virtually anyone's design tool. It comprises a set of tools that allows standard cad packages to gain access to all the data needed for test and verification of standard cell asic designs and also allows access to new cad suites. The software goes further than existing layout-checking and test interfaces, which leave further work to be done before manufacture can start. European Silicon Structures, 0344 525252.

Rack-mount disk drives. Two 5.25in or 3.5in hard disk drives in one 19in rack are now offered by Blue Chip in the Racknet range. Combined capacity of 1070Mbyte is possible and the drives can be specified to run Compsurf. Independent 40W power supplies with short-circuit protection confer a degree of immunity to drive failure. Power and "active" indicators are on the panel of the 2U unit. Blue Chip Technology, 0244 520222.





Data acquisition. DaqWare from National Instruments is data acquisition software for NI's multifunction boards for PC XT/AT/eisa and PS/2 computers. It is dos-based and interactive, with an intuitive menu, requires no programming and will handle strip-charting, temperature measurement, waveform generation and streaming to disk in binary, ascii or Lotus 1-2-3. The package was developed using LabWindows, so that users can obtain DaqWare source code and use LabWindows to modify it. National Instruments UK Ltd, 0800 289877 (free).

PC process control. A combination of Sigma II software and DM 4000 "smart" indicators allows central PC control of up to 99 temperature measuring points. An operator can communicate with individual indicators to upload, download or edit. Sigma II exploits VGA graphics to provide real-time or historical XY or bar graphs, mimic displays, alarm summaries and up to five reports and four simultaneous logs. Four displays can be run on one screen. All data is captured or transmitted vis a communications port. Status Instruments Ltd. 0684 296818.

Schmitt optocouplers. *TCSS* optocouplers by Telefunken provide active-high or active-low Schmitt trigger outputs. A light source with a wavelength of 950nm and a detector face each other in a plastic case, detecting anything that breaks the beam in the gap. Aperture sizes are 0.25mm, 0.5mm or 1mm and the units are mounted on a PCB or screwed in position. Telefunken Electronic, 0635 30905.

PCB fault diagnosis. On unpowered boards, impedance signatures are a viable method of diagnosing faults. *T6000* from Polar Instruments uses current-limited AC as a stimulus and compares the impedance of circuit nodes with a stored set of responses from a known good board under PC control. The instrument is happy with complex ICs and simpler circuits. In its stand-alone method of working, the *T6000* presents analogue waveforms from the board on screen, this often being enough for diagnosis. Further, it can compare these responses with those from a good board on its dual-channel display. Polar Instruments Ltd. 0481 53081.

REGULARS

APPLICATIONS

Filtering reference voltages

Burr-Brown *REF102* buried-zener 10V reference lays claim to better stability and five times lower noise than a band-gap reference. But noise is still around 600µV pk-pk at a noise bandwidth of 1MHz. Bulletin AB-003, Vol 1, No 1, points out that filters and buffers go some way towards reducing noise and its bandwidth, but not far enough in many applications.

Figure 1 shows a typical example – a single-pole filter and an op-amp buffer. One problem with this is that capacitor leakage current goes through R_1 and is variable with temperature, particularly in large capacitors needed for this job; the resulting DC error will drift.

The buffer, too, puts in its own noise contribution, over its full unity-gain bandwidth. So even if the filter output is silent, unacceptable noise still appears at the circuit output.

To solve both problems at a stroke, use the circuit of **Fig. 2**. The filter is now at the output of the buffer, where its –3dB point is $2\pi R_1C_1$ (reducing noise bandwidth by, say, 100 reduces noise by 10). The R_2C_2 arrangement maintains stability and R_2C_2 should equal $2R_1C_1$ to escape amplifier noise gain peaks. Resistor R_2 should be kept fairly low, since it takes bias current and could cause DC error and noise; R_1 should also be low, since it takes load current and its volts drop, increasing the required output swing. It should drop less than 1V full load.

Since the filter is now in the feedback loop, leakage current voltage drop across R_1 is divided by the loop gain, DC output impedance is very low and the voltage across C_2 is almost nothing, giving rise to negligible leakage current. When driving large capacitive loads, $(C_{LOAD}+C_1)R_1$ must be less than $0.5R_2C_2$.

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

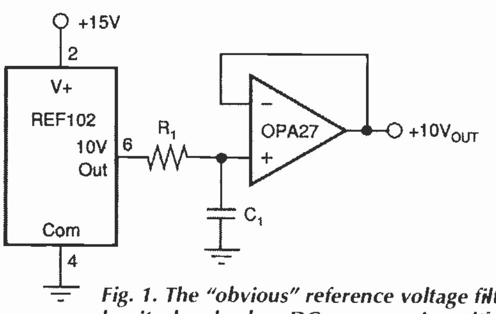


Fig. 1. The "obvious" reference voltage filter has its drawbacks – DC error varying with temperature and noise from the buffer.

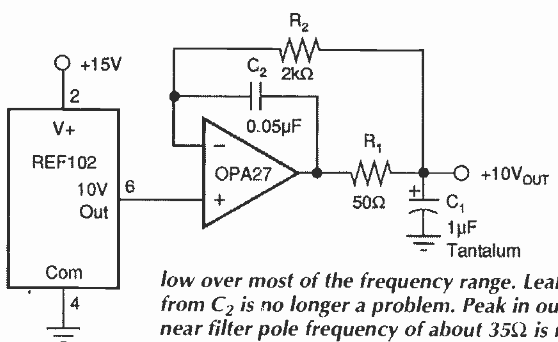
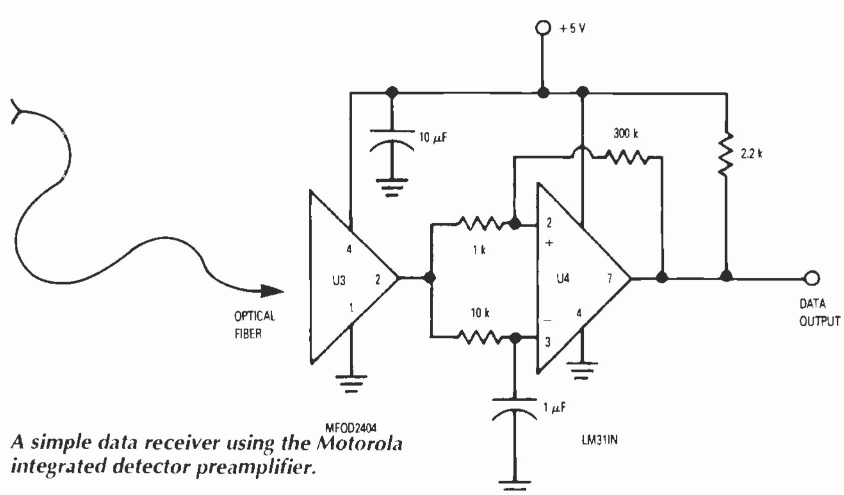


Fig. 2. An improved filter avoids both the problems of Fig. 1. The filter reduces noise from reference and op-amp. Output impedance is low over most of the frequency range. Leakage current from C_2 is no longer a problem. Peak in output impedance near filter pole frequency of about 35Ω is reduced by reducing R_1 and increasing C_1. The peak is $0.7R_1$.

Optical-fibre comms

Application Note AN846 from Motorola is one of those general ones that gives one an overview of a subject. In this case, the subject in question is optical fibres and their use in communications work. It covers the field from the physics of light itself and optical fibres to the use of semiconductors in light transmission and reception, ending with a description of a basic system. In comparison with some of the textbooks in current circulation, this paper by John Bliss and Joseph Slaughter is of much greater use to the practising engineer, even though it is now 18 months old.

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



A simple data receiver using the Motorola integrated detector preamplifier.

DC-to-DC converter transistors

Zetex Note AN81 describes use of its hi-rel range of E-line bipolar and mosfet transistors in DC-to-DC converters providing up to about 10W for small equipment such as fluorescent tubes and flash guns.

With high current capability, these devices will easily function in a single-transistor converter such as the 8W fluorescent lamp inverter in **Fig. 1**. Here, load comes well inside the "awkward" category, since it needs a high voltage to strike, but promptly becomes a low-impedance load taking up to about 3A for this size of lamp. Instead of an expensive TO220 type device, the Zetex E-line *ZTX652* handles the variable load, supplying the current without a high $V_{CE(sat)}$.

Figure 2 is the circuit of a minimal converter giving +12V and –5V for memory, interface and display circuits from a 5V supply. Two *ZTX449*s working at 25kHz have to supply up to 1.5A peak at $V_{CE(sat)}$ of 250mV to give decent efficiency from 5V.

Capacitors are another type of difficult load since, in a typical flashgun, for example, the voltage across them varies from 0 to 400V during the charging cycle. A flyback converter, isolating the load from the switching circuit is one answer (**Fig. 3**). The higher peak-switching-current needed in a flyback converter, compared to a forward converter, puts the high current and low saturation voltage of the transistors to good use. A shutdown sensor avoids overcharging of output C and increases battery life.

***Zetex plc**, Fields New Road, Chadderton, Oldham OL9 8NP. Telephone 061-627 5105.*

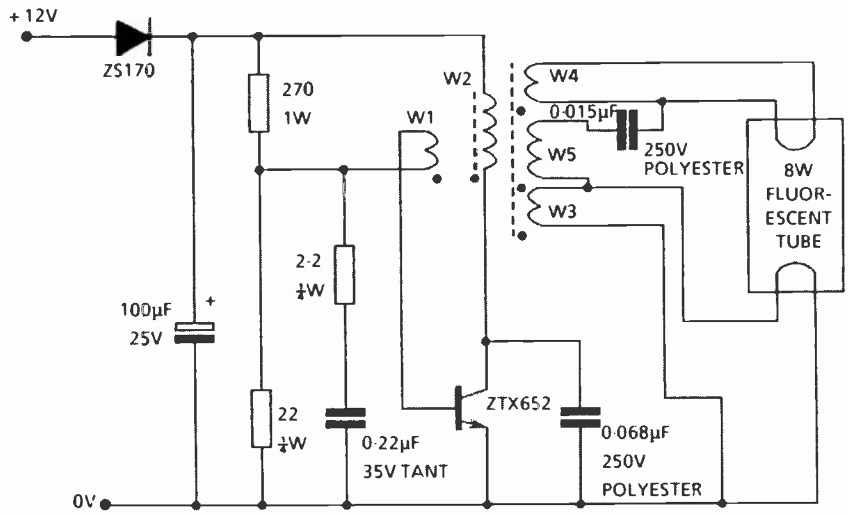


Fig. 1. Single-device 8W converter for a fluorescent tube, using Zetex E-line ZTX652 to pass high current at low saturation voltage.

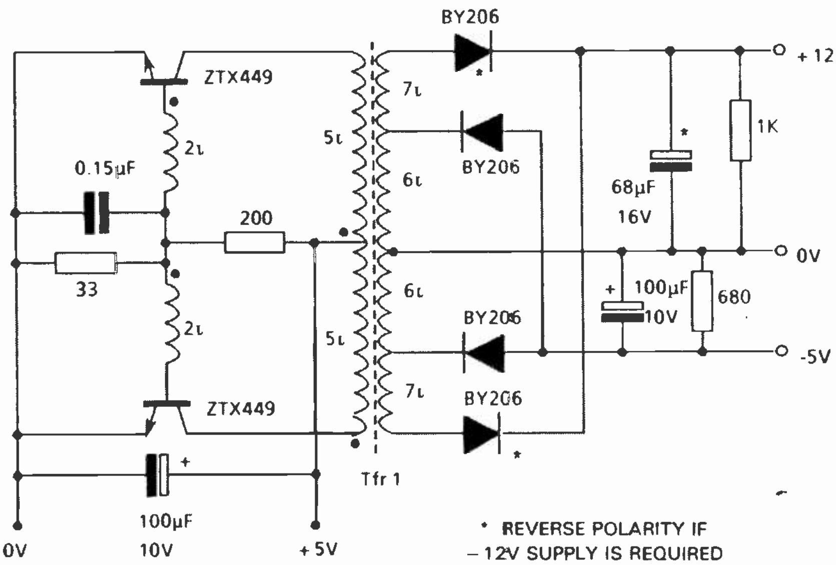


Fig. 2. Five-watt converter for memories and interfaces, powered by a 5V supply.

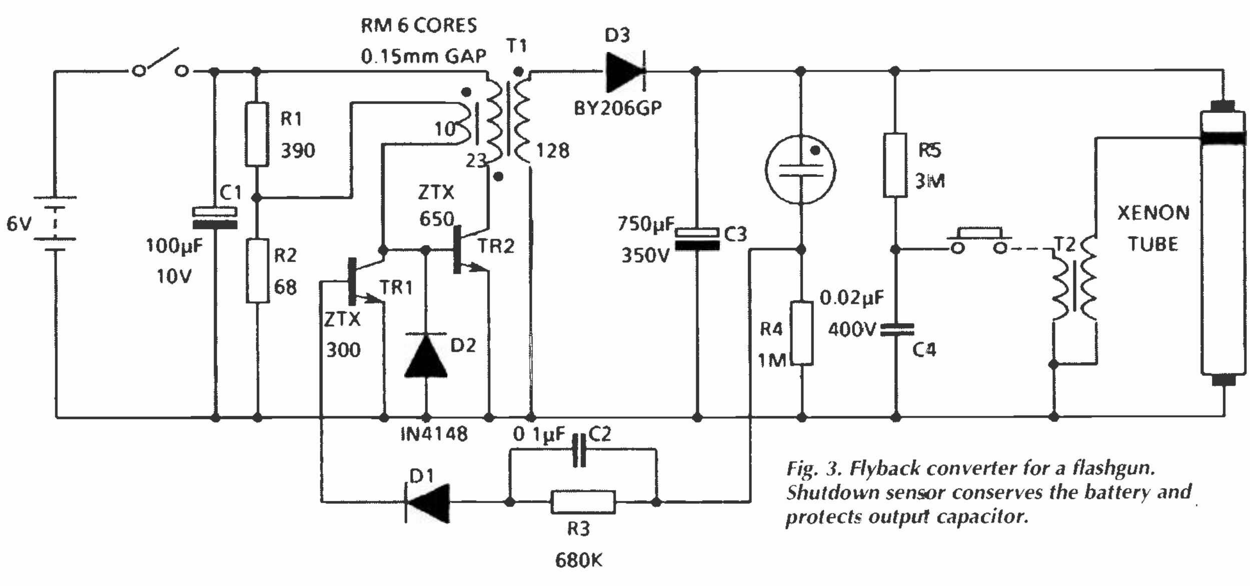


Fig. 3. Flyback converter for a flashgun. Shutdown sensor conserves the battery and protects output capacitor.

Range and direction by infrared

High-gain preamplifier *SL486* is intended to receive the output of an infrared diode and to feed the amplified and processed signal to a remote-control receiver. **Figure 1**, taken from GEC Plessey's latest consumer IC handbook, shows both a voltage regulator and a pulse stretcher. The latter is meant to increase the length of IR pulses, with widths of a few microseconds to about 2.5ms for use with microprocessor input.

Figure 2 shows how the *SL486* can be used to measure the direction and range of an infrared transmitter. AGC has a 68dB range and is taken to an output. Bursts of IR radiation are received by the *SL486*, producing an AGC voltage as shown in **Fig. 3(a)**, shaped to avoid the effect of noise on the received pulse. The AGC waveform has a 0-300mV "shelf" and sits on a DC level of 2V. The level varies slightly with background lighting and so is removed by an ambient sample-and-hold circuit to produce the signal shown at **(b)** for different ranges.

This is now subtracted from a 10ms wide sample pulse **(c)**, triggered by the AGC waveform, to produce the 10ms pulses **(d)** whose height corresponds to the shelf height of the AGC waveform and therefore represents range information. At **(e)** the range pulses are sampled and held to give the output of the unit – a DC level giving the range of the transmitter. Connecting two such circuits in parallel, but with the IR diodes mounted at 90° to each other gives differential signals, identical when the diodes point 45° either side of the transmitter and varying proportionally to allow a measurement of direction.

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

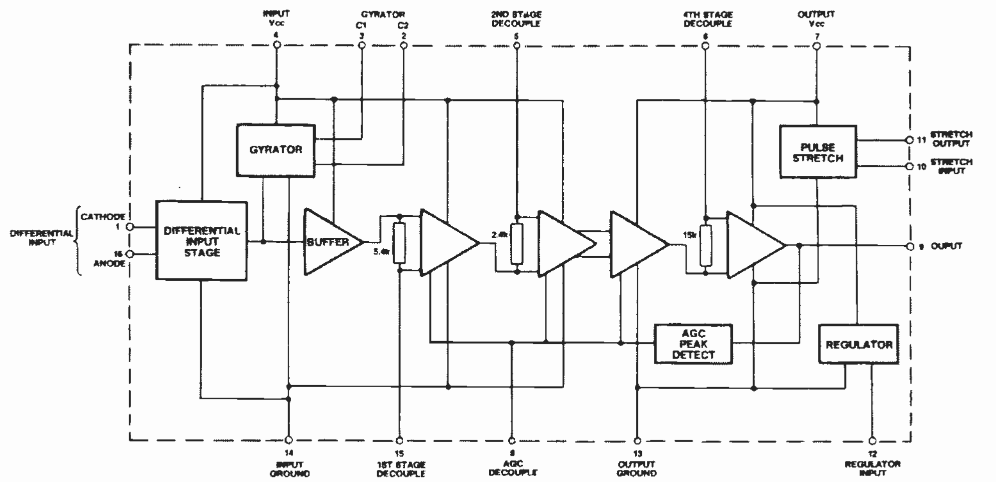


Fig. 1. GEC Plessey SL486 high-gain preamplifier for infrared diode signal processing for a remote-control receiver or microprocessor. The pulse stretcher lengthens pulses for input to micros and the AGC is shaped to reduce gain to noise during IR reception.

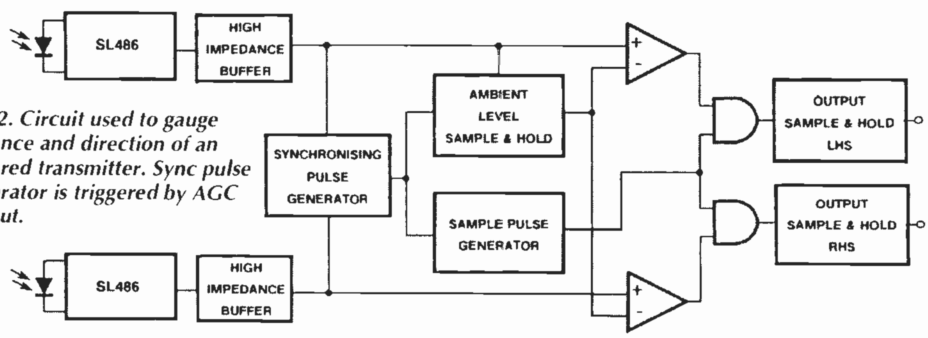


Fig. 2. Circuit used to gauge distance and direction of an infrared transmitter. Sync pulse generator is triggered by AGC output.

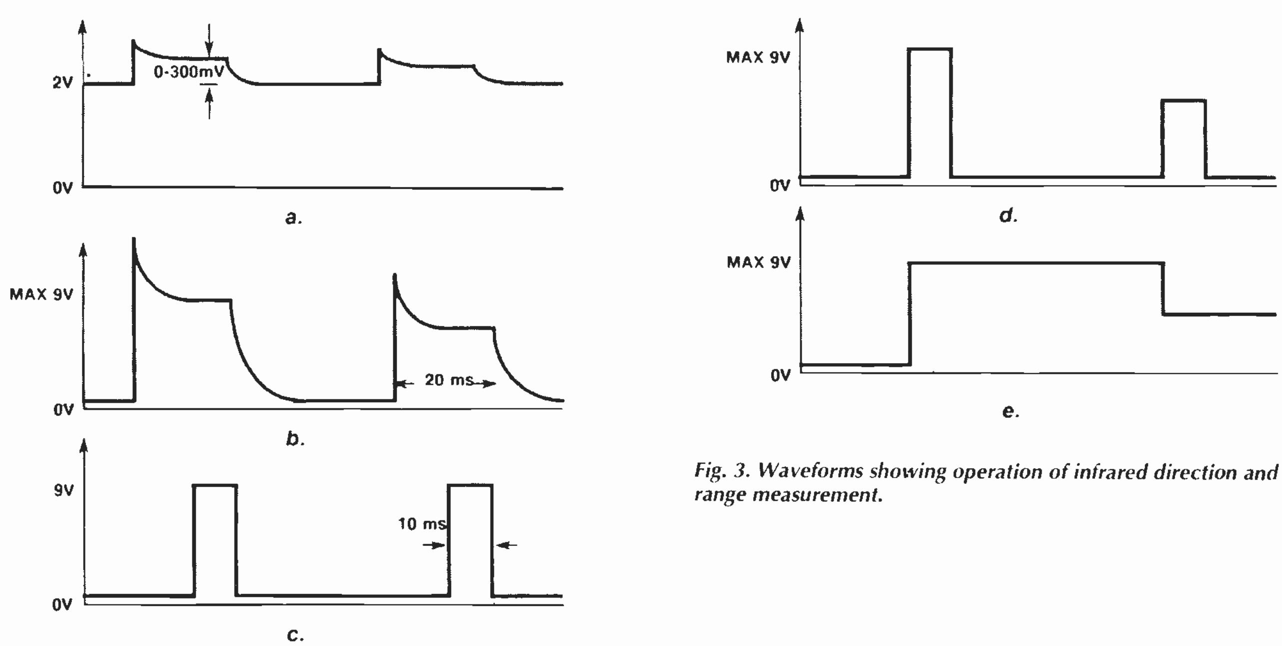


Fig. 3. Waveforms showing operation of infrared direction and range measurement.

£1 BARGAIN PACKS

In fact, cheaper than £1 because if you buy 10 you can choose one other and receive it free.

5 13A spurs provide a fused outlet to a ring main where devices such as a clock must not be switched off. Order Ref. 2.
4 In flex switches with neon on/off lights, saves leaving things switched on. Order Ref. 7.
2 6V 1A mains transformers upright mounting with fixing clamps. Order Ref. 9.
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12 30 watt reed switches, it's surprising what you can make with these – burglar alarms, secret switches, relay, etc. Order Ref. 13.
2 25 watt loudspeakers two unit crossovers. Order Ref. 22.
2 Nicad constant current chargers adapt to charge almost any nicad battery. Order Ref. 30.
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6 Safety cover for 13A sockets – prevent those inquisitive little fingers from getting nasty shocks. Order Ref. 149
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10 Keyboard switches – made for computers but have many other applications. Order Ref. 201.
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50 Leads with push-on ¼in tags – a must for hook ups – mains connections etc. Order Ref. 259.
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1 Mini 1 watt amp for record player attached to unit that will also change speed of record player motor. Order Ref. 268.
3 Mild steel boxes approx 3in x 3in x 1in deep – standard electrical. Order Ref. 283.
50 Mixed silicon diodes. Order Ref. 293.
1 6 digit mains operated counter, standard size but counts in even numbers. Order Ref. 28.
1 In-flight stereo unit. Has 2 most useful mini moving coil speakers. Ex BOAC. Order Ref. 29.
2 6V operated reed relays, one normally on, other normally closed. Order Ref. 48.
2 Plug in relays with 3 changeover contacts. Coil operated by 12V DC or 24V AC. Order Ref. 50.
1 12V pcb mounting relay. 2 changeover. Order Ref. ??
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4 Dolls house switches or use them for any other low voltage application. Order Ref. 57.
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— BARGAINS GALORE —

DON'T MISS THIS BARGAIN You have until the 31st July to buy a unit, which must have cost well over £60, for less than £10! It is a switch mode power supply by Astec. It has normal mains input and outputs +12v at 4A, +5v at 16A and −12v at 500mA. It is a really beautiful unit enclosed in a plated steel case. Was probably made to power an expensive computer, is still in maker's original packing, price until 31st July is only **£9.50**, Order Ref. 9.5P1. After 31st July, the price will be £15.

VARIAC – This infinitely variable unit gives any voltage from 0–230 AC at ½A. Obviously an invaluable piece of equipment which should be in every workshop and probably would be except that the usual price for this is £35 plus VAT. Now is your chance to buy one, brand new, at £15 including VAT, Order Ref. 15P42B.

SWITCH MODE POWER SUPPLY BARGAIN made by Astec, their model no. 51052. PCB mounted and wired up ready for use but not cased. Normal 230v AC input, outputs are +5v at 3A, +12v at ½A and −12v at 100mA. Secondary outputs, +5v at 20mA and +10v at 50mA. This also has a thermostatic overheat protection against overload. Expensive unit but yours for **£5**, Order Ref. 5P188.

HEAVY DUTY MAINS FLEX – 3-core 15A flex. Price is 10m for **£2.50**, Order Ref. 2.5P/3. You can have this in longer lengths, multiples of 10m up to 100m. Available in black or white, please state which. Also available, 3-core 20A flex, 10m for **£3**, Order Ref. 3P109 and 2-core 20A flex, 15m for **£3**, Order Ref. 3P110

OSCILLOSCOPE 301B developed for testing transmission lines, it makes and displays pulse echoes to find shorts and breaks in cable networks, this uses a 3" CRT to display the type of fault and a LCD to read out the distance from the fault. The instrument is powered by 12V of rechargeable nicads located in base, and it generates 1.5KV internally. It is housed in a high impact plastic case size approx. 9½" < 9½" × 5". Ex British Telecom in very good condition and working order, £49.50 plus £5 insured delivery.

WANT TO EXPERIMENT WITH DISC DRIVE You can have a double sided JAP-made model, brand new for only £10! Why, because it was made for a very compact unit and consequently uses a small non-standard disc (We stock these at £1 each). It requires 5V 300mA power supply (we can supply). The drive is nicely cased and has output and input leads, originally listed at over £100, so it's a real *bargain*.

LITHIUM BATTERIES 3.5v penlight size, 2 mounted on p.c.b. with diodes, other bits. Lithium batteries as you may know are virtually everlasting (until they are put in circuit of course) so they are ideal for alarms and similar devices that do not draw current but do rely on it always being available. 4 panels that is 8 batteries altogether £2, orde ref 2P258B.

POWER SUPPLY WITH EXTRAS output 12v 1amp, mains input is fused and filtered and 12v output is voltage regulated, very wel made on p.c.b, and also mounted on the board but easily removed are two 12v relays and a Piezosounder. Made for expensive equipment but never installed. price £3 order ref. 3P80B.

12 VOLT 1.9 AMP-HOUR rechargeable battery by Jap YUASHA brand new, charged ready for use £6.50 each. Solar charger to house this and keep it ready £29.50.

100 WATT MAINS TRANSFORMERS all normal primaries. 20-0-20 volt 2½A 30volt 3½A, 40volt 2½A and 50volt 2A all upright mounting, all £4 each, good quantities in stock.

PHILIPS 9" HIGH RESOLUTION MONITOR black and white in metal frame for easy mounting, brand new still in makers packing, offered at less than price of tube alone, only £15 plus £5 delivery — good discount for quantities.

16 CHARACTER 2 LINE DISPLAY screen size 85mm × 36mm. Alpha-numeric LCD dot matrix module with integral micro processor made by Epson their ref 16027AR brand £8 each, 10 for £70, 100 for £500.

INSULATION TESTER WITH MULTIMETER internally generates voltages which enable you to read insulation directly in megohms. The multimeter has four ranges. AC/DC volts, 3 ranges DC milliamps, 3 ranges resistance and 5 amp range. These instruments are EX British Telecom, but in very good condition, tested and gntd. OK, probably cost at least £50 each, yours for only £7.50 with leads, carrying case £2.00 extra.

BRUSHLESS D.C. 12V FAN tiny, only 60mm square, good air mover but causes no interference £8.00.

2MW LASER Helium Neon by PHILIPS, full spec, £30, power supply for this in kit form with case is £15.00, or in larger case to house tube as well £17.00. The larger unit, made up, tested and ready to use, complete with laser tube £69.00 plus £5 insured delivery.

MAINS 230V FAN best make "PAPST" 4½" square, metal blades £8.00

SOLAR CHARGER holds 4 AA nicads and recharges these in 8 hrs., in very neat plastic case £6.00.

SOLAR CELLS with terminals for joining in series for higher volts or parallel for extra current: 100mA £1, 400mA £2, 700mA £2.75, 1A £3 50.

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AIR SPACED TRIMMER CAPS 2–20 pf ideal for precision tuning uhf circuits 25p each, 10 for £2. 100 for £15.

1Khz. TONE GENERATOR this is PP3 battery operated and has a 1Khz output that can be continuous or interrupted at a rate variable by a panel mounted control. Constructed on a pcb and front panel size approx 105×50mm ex equipment but in as new condition £2 each.

MAINS ISOLATION TRANSFORMER stops you getting "to earth" shocks. 230V in and 230V out. 150 watt upright mounting £7.50.

MINI MONO AMP on pcb size 4" × 2" with front panel holding volume control and with spare hole for switch or tone control, output is 4 watt into 4 ohm speaker using 12V or 1 watt into 8 ohm using 9V. Brand new and perfect only £1 each or 12 for £10.

5 RPM 6OW MAINS DRIVEN MOTOR AND GEARBOX this has a 3in square mounting plate and is 4in deep. It is a shaded pole motor. Price £5

POWER SUPPLY UNITS mains in, dc out, based 4.5v 150mA regulated £1, 6v 200mA regulated £1, 6v 700mA £1, 12v 2A £6, 24v 200mA £2.

TORROIDAL MAINS TRANSFORMER with twin outputs, 3.3v 2 amps and 12v 1 amp, one use would be power supply, price £5.

AMSTRAD POWER UNIT 13.5v at 1.9A encased and with leads and output plug, normal mains input £5 each, 10 for £45.

AMSTRAD 3.5 FLOPPY DRIVE Reference FD9 brand new and perfect, £35.

ATARI 64XE COMPUTER at 65K this is quite powerful so suitable for home or business, unused and in perfect order but less PSU, only £19.50, Handbook £5 extra.

9" CATHODE RAY TUBE Philips M24/306W, which is not only high resolution but is also X Ray and implosion protected, regular price over £30, you can have them at £12 each. Tubes are guaranteed unused.

80 Watt MAINS TRANSFORMERS two available in good quality, both with normal primaries and upright mounting, one is 20V 4A the other 40V 2A only £3 each or 10 for £27 carriage paid.

PROJECT BOX size approx 8" × 4" × 4½" metal, sprayed grey, louvred ends for ventilation otherwise undrilled made for GPC so best quality, only £3 each or 10 for £27.

12V SOLENOID has good ½" pull or could push if modified, size approx 1½" long by 1" square, £1 each or 10 for £9.

WATER VALVE 230V operated with hose connections, ideal for auto plant spray or would control air or gas into tanks etc, £1 each or 10 for £9.

HANG UP PHONE won't clutter up your desk or workbench, current model, has push button dialling, last number recall, internal alarm etc., Ex B.T. in good condition and fully working ready to plug in. £5

HIGH VOLTAGE CAPS if you use these ask for our 1-20 Kv Capacitor list, we have over ¼ million in stock and might save you a lot of money.

ELECTRONIC BUMP & GO SPACESHIP sound and impact controlled responds to claps and shouts and reverses or diverts should it hit anything! Kit with really detailed instructions, will make ideal present for budding young electrician. Should be able to assemble but you may have to help with the soldering of the components on the PCB. Complete kit £8.95.

500V BRIDGE MEGGER developed for G.P.O. technicians the Ohmeter 18B is the modern equivalent of the bridge meggar. 9V battery operated it incorporates 500V generation for insulation testing and a null balance bridge for very accurate resistance measurement. Ex B.T. in quite good condition with data & tested. Yours for a fraction of original cost £45+ £5 insured delivery.

EXPERIMENTING WITH VALVES don't spend a fortune on a mains transformer we can supply one with standard mains input and secs. of 250-0-250V at 75 mA and 6.3V at 3 A. price £5.

15Watt 8ohm 8" SPEAKER & 3" TWEETER made for a discontinued high quality music centre, give real hi.fi. and for only £4 pair.

ULTRASONIC TRANSMITTER/RECEIVER with Piezo alarm, built into preformed case, is triggered by movement disturbing reflected signal, intended for burglar alarm, car alarm etc. has many extras, time delay, auto reset, secret off device etc. A £40 instrument yours for £10.

MOVEMENT ALARM goes off with slightest touch, ideal to protect car, cycle, doorway, window, stairway, etc. etc. Complete with piezo shrieker ready to use Only £2 (PP3 battery not supplied).

STEREO HEADPHONE extra lightweight with plug £2 each or 10 pairs for £18.

B.T. TELEPHONE LEAD 3m long and with B.T. flat plug ideal to make extension for phone, Fax, etc. 50p each, £40 per 100, £300 per 1000.

WATER PUMP very powerful and with twin outlets, mains operated, £10.00.

STUDIO 100 by Amstrad, the ultimate disco control panel, has four separately controlled and metered channels, twin cassettes, AM/FM radio, stereo audio amplifier, phono & C.O. inputs, etc, etc, etc, regular price over £400 we have a few still in maker's packing, brand new and guaranteed yours for £99.

ROTARY POSITION CONTROLLER for aerials, ventilators, dampers, rheostats, dampers or applications requiring 180 degrees clockwise and anti-clockwise movement. We have the Sauter MVE4 154 servo motor drive ref AR30W3S regular price over £70 brand new, £15 each.

12 VOLT 8 AMP MAINS TRANSFORMER £4, waterproof metal box for same, £4.

110 WATT SWITCHMODE POWER SUPPLY 230v mains operated, outputs of 38v 2½A and 5v 3A, we have a lot of these and need the space so you can have these at a fraction of their cost if you order before Dec 31 price is £6.

10 VA MAINS TRANSFORMERS all p.c.b. mounting, all £1 each, 10 for £9, 100 for £75, for output 12-0-12v order ref WA1, 20/0/20v order ref WA3, 18-0-18v not p.c.b. mounting but fully shrouded same price order ref WA4.

0-1 mA FULL VISION PANEL METER 2¾" square, scaled 0-100 but scale easily removed for re-writing £1 each. 10 for £9, 100 for £75.

PANEL AMP-METERS 80 × 70mm beautiful instruments £5 each, 10 amp order ref WA6, 5 amp order ref WA7.

Vu METER illuminate this from behind becomes on/off indicator as well, 1½" square 75p each, 10 for £6. 100 for £50.

EDGE-WISE PANEL METER ideal when short of panel space only 40 × 14mm, also have built-in led, 500 uA f.sdd, scaled 0-5, £1 each, 10 for £9, 100 for £75.

VIBRATING REED FREQUENCY PANEL METER 4" square, 55-65 Hz only £9 each.

P.C.B. DRILLS 12 assorted sizes between .75 and 1.5mm £1 the lot.

LOW PRICED FIELD TELEPHONES. Ex-GPO models, not quite so nice-looking but quite efficient, and have the big advantage that the ringing is done by means of a hand operated internal generator. This saves a lot of batteries. These 'phones have the normal type of rotary dial built in and can still be connected into a normal B.T. system. Tested, guaranteed in good order, price only £9.50 each. Order Ref 9P5.

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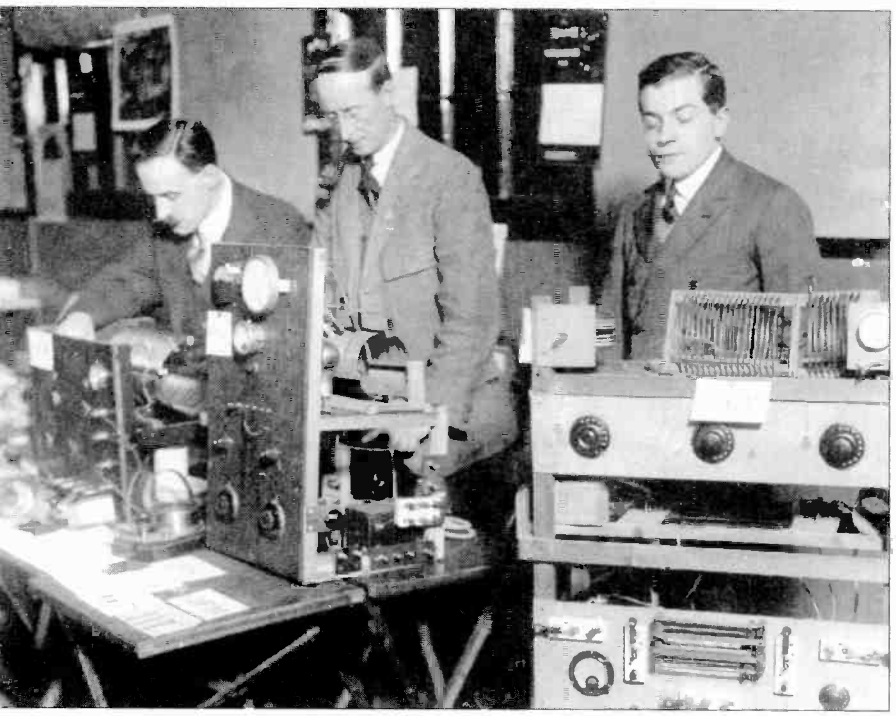
Tottenham Wireless Society's exhibition at Tottenham Town Hall in 1926. Identified are Messrs Wroth and Vickery with shortwave receivers.

WHEN AMATEURS

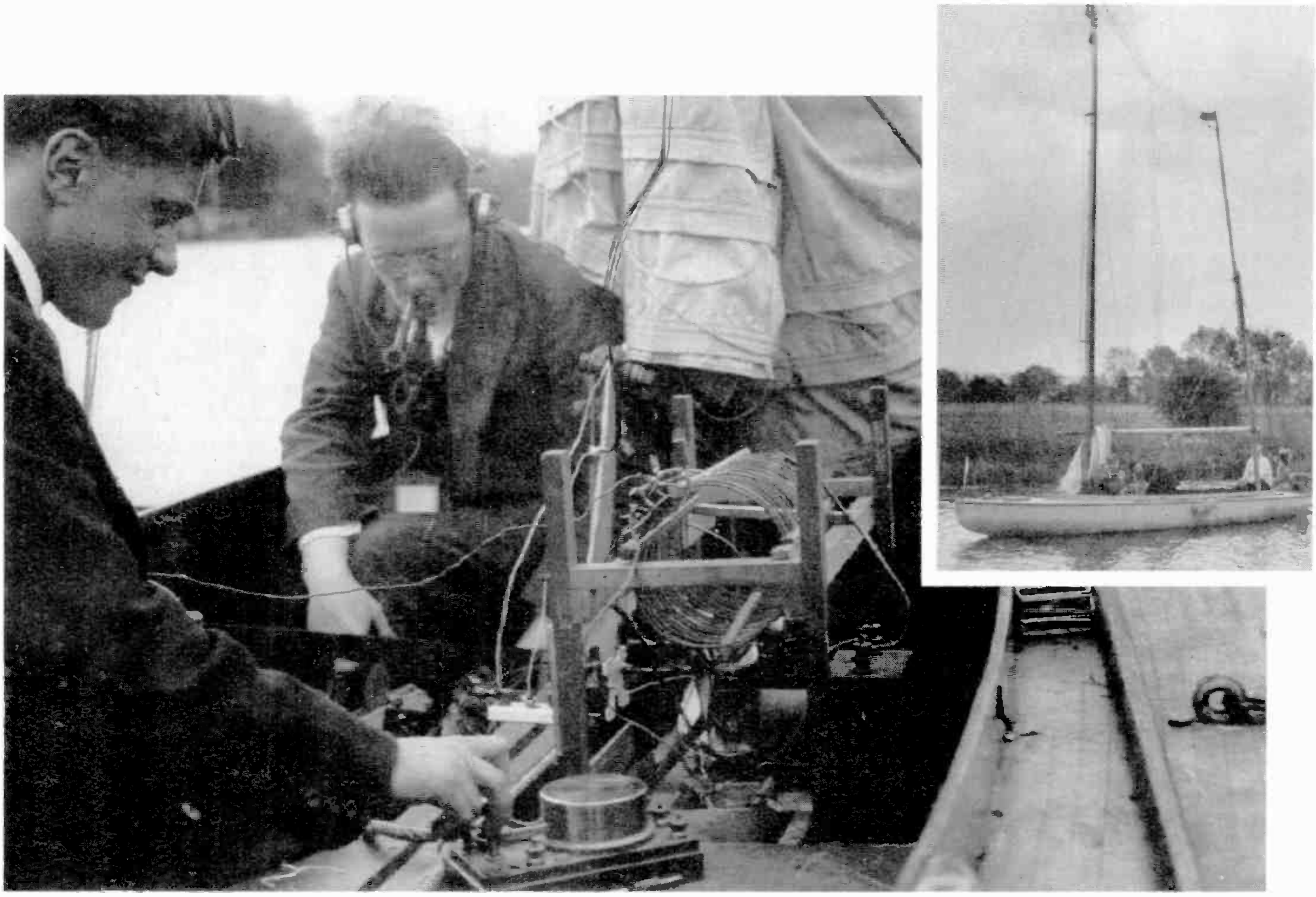
The 1920s was an extremely active decade for amateur wireless societies – for example, the Tottenham Wireless Society in particular missed no opportunity to keep *Wireless World* informed of its doings.

We know that in 1926 it held an exhibition at Tottenham Town Hall. In May the following year the society enjoyed a field day near Hoddesdon, Herts, which included a three-way telephonic link-up between a boat, a car and a fixed station. Other members went out by motorcycle to pick up signals.

Unlicensed amateur transmitters were at risk from Post Office patrol vans, although these purported to be for the purpose of locating oscillating receivers. Bearings were taken from three points within about a mile, using a rotating frame aerial. The van could then cruise down the suspect street, until direction-indicators pointed to the suspect house.

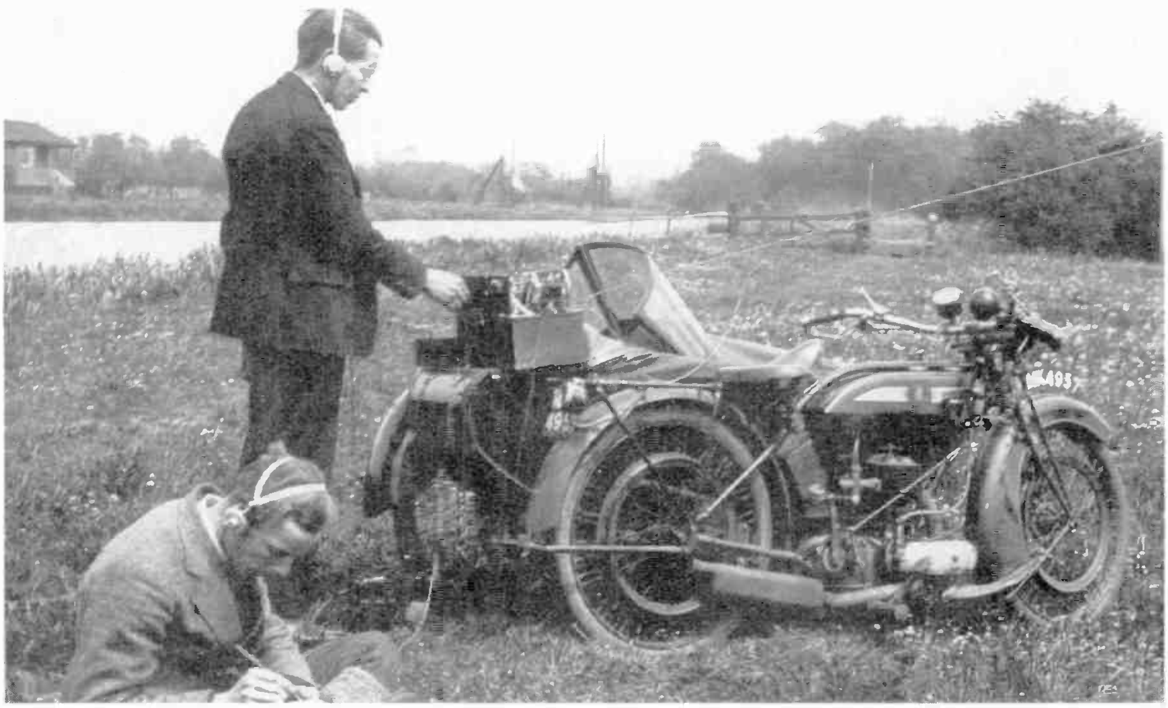


The shortwave transmission stand at the Tottenham Town Hall exhibition.



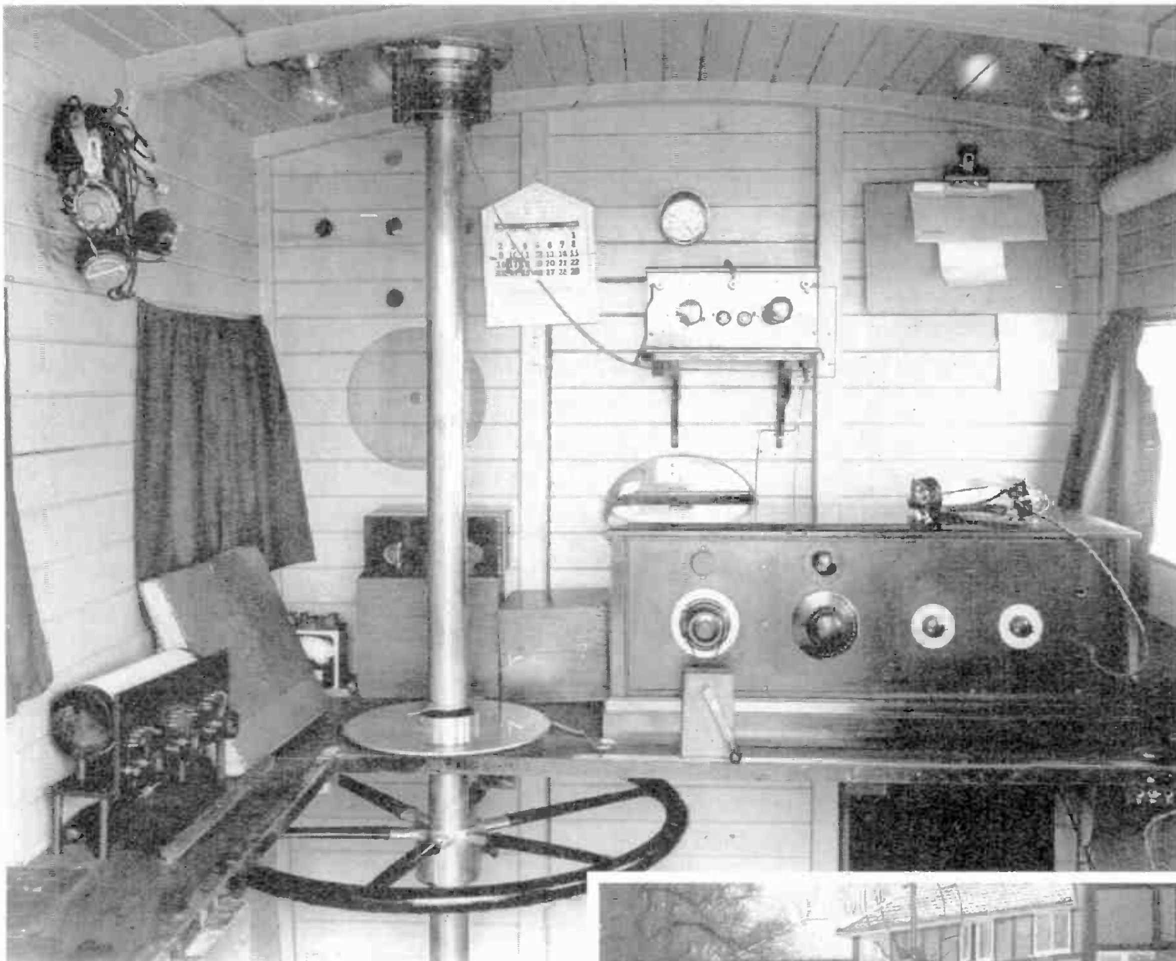
TALKED TO EVERYONE

A boat, with aerials rigged from the masts, was used in a three-way link-up during the Tottenham society's 1927 field day.



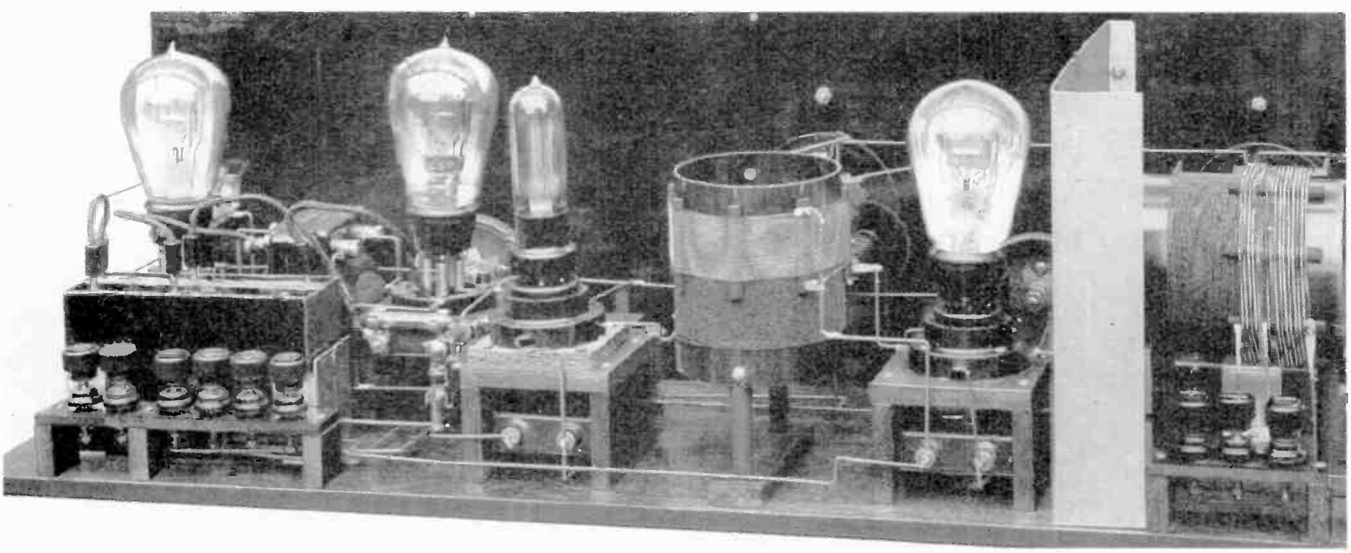
Other members went out by motorcycle to pick up signals.

On the right is a three-valve receiver-amplifier; its tuning capacitor dial can be seen at its left-hand end, above a small handle used to hold it secure against vibration. Above, below the clock, is a balancing tuner, to compensate for errors.



1927 Post Office patrol van, supposedly for the purpose of locating oscillating receivers, but also useful for tracking unlicensed transmitters. The frame aerial on the roof of the van was rotated by the large wheel within to take bearings.

One of the most popular home-build amateur receivers was the four-valve Wireless World Everyman 4. This picture, from the March 1927 issue, shows one constructed by a reader.



Wide screens – narrow benefits?

It is now ten years since the first closed-circuit demonstrations were given in Europe of NHK's 1125-line, 60Hz, 5:3 (later 16:9) wide-screen HDTV system. The first was in Dublin, followed a few months later by further demonstrations by Sony at IBC82 in Brighton. The pictures were truly spectacular and made a profound impression on all who saw them. If only we could receive such pictures in the home without having to use £100,000 large-screen projection displays.

Since the demonstrations, lively debate has been continuing as to which of three paths should be followed by broadcast television: is the domestic picture from 625-line pal good enough even for DBS?; should we aim to overcome specific defects of pal and introduce a system (such as mac) more suitable for future DBS and the next generation of larger, flatter screens?; or should we introduce a radically different HDTV system capable of providing pictures akin to those of widescreen cinema film, taking into account the possibility of using such a system for electronic video-cinema production and presentation?

In 1982, the issues seemed reasonably straightforward: 4:3 pal would continue for terrestrial broadcasting. Mac as a hybrid analogue-digital system seemed the logical choice for DBS and could be upgraded to widescreen enhanced-mac without affecting viewers with 4:3 mac displays. 1125-line HDTV required too much spectrum for broadcast transmission, except possibly for DBS above 20GHz but deserved consideration as a production standard for master tapes and electronic film-making.

But it was not long before the basic issues became blurred and the intervening years have raised more questions than answers.

The Japanese set out to prove that their non-compatible system could be digitally compressed in bandwidth and developed a series of muse compression systems that can dramatically reduce the bandwidth (at some loss in picture quality) while maintaining a 16:9 aspect ratio.

For technical and commercial reasons, the Europeans developed the HD-mac 1250-line system capable of fitting into a 12GHz FM satellite channel with 16:9 aspect ratio but fully compatible with 625-line 4:3 D-mac. An ambitious project with an all-digital HDTV objective has also been launched.

But the UK's attempt to establish first mac and then emac on BSB's Marco Polo has collapsed (for commercial rather than technical reasons) and is due to be taken out of service by BSkyB at the end of the year – though D-mac and D2-mac remain official European standards for DBS.

In the US, the FCC is following a "simulcast" track and has just about completed testing of six competing systems (see "Putting the right numbers into HDTV", this issue), including four fully-digital systems. The systems are all designed to fit into one or more standard FCC 6MHz channels without attempting any compatibility with existing 525-line NTSC but being suitable for terrestrial, satellite or cable distribution of 16:9 widescreen pictures.

Disruption without benefits

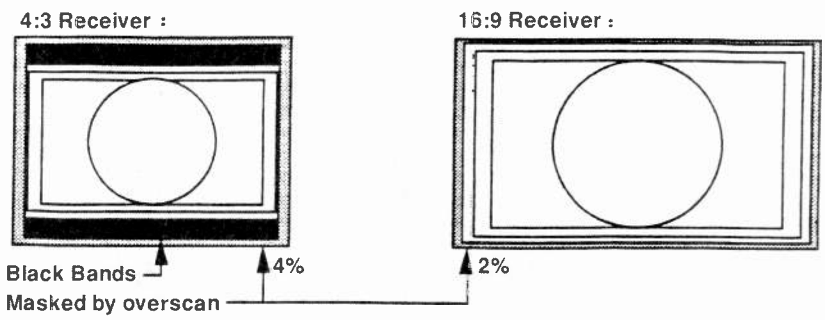
But within the North American broadcast industry there is a growing belief that HDTV or enhanced systems, far from benefitting the industry, are threatening to disrupt it.

Writing in the *SMPE Journal* earlier this year Stan Baron of NBC said: "For American broadcasters, HDTV has many portents of catastrophe. Current HDTV equipment costs are too high for terrestrial broadcasters to absorb. Furthermore, although the plan for terrestrial broadcasting calls for simultaneous broadcasting of NTSC and HDTV services, many of the current HDTV production equipment offerings are not compatible with current NTSC standards. Simultaneous broadcasting of the two services means that a single broadcast centre must provide two different services from the

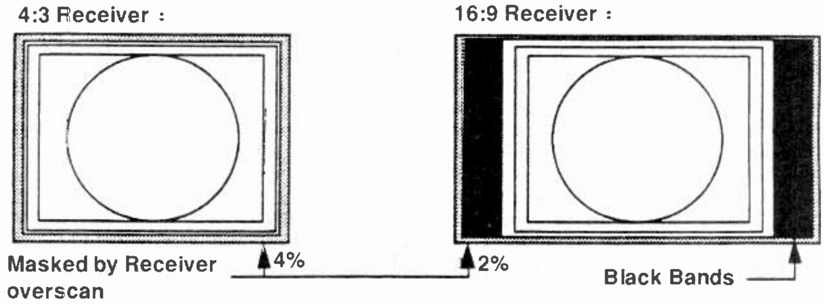
Table 1. Equidistant viewing experience

Screen diagonal	Viewing resolution*	Screen width	Viewing distance**
635mm (4:3)	330	510	6 to 8 x
790mm (16:9)	330	685	6 to 8 x
1070mm (16:9)	450	940	4 to 6 x
1650 (16:9)	700	1450	2 to 4 x

*In TV lines/picture height
** In multiples of picture height



How 16:9 pal-plus pictures appear on a conventional and a widescreen receiver (action area 5% border, safe area 10% border). Most receivers are adjusted for 4% overscan at the factory but this may not be maintained over a set's life.



4:3 pal pictures on a conventional and a widescreen receiver. (Source Gardiner, ITC) same programming sources..."

Where does all this leave pal-plus (see Tom Ivall's "Additional consumer benefits for pal plus?" *EW + WW*, March 1992, pp.182-3)?

For the vast majority of existing viewers, the only partially compatible system of widescreen transmission, although technically feasible and ingenious, might more aptly be renamed pal-minus. It will surely alienate viewers, particularly those with relatively small-screen displays (50mm or less diagonals) who will have to watch 16:9 pictures in letter-box format – not just widescreen films but run-of-the-mill studio productions with talking heads etc.

Industry pressure

There is little doubt that once pal-plus sets and production equipment are available, and the system becomes operational (if in fact it ever does in the UK), there will be strong pressure from the consumer-electronics industry to extend wide-screen transmissions to most programmes, allowing retailers to create a market for new, relatively high-cost widescreen sets.

It is no secret that UK broadcasters had hoped the pal-plus specification would have avoided the letter-box format in favour of side-panels, or a compromise between the two, but were forced to recognise that this would have introduced further complexities into an already complex system.

In the studios, pal-plus requires extensive re-equipping, including use of a new digital component-standard to cope with the additional luminance bandwidth and a new generation of videotape machines. Ideally, pal-plus also requires a new generation of CCD cameras – though attempts will probably be made to use anamorphic lenses which perform horizontal image compression though lose resolution.

All this expense to provide wide-screen pictures to an audience that has consistently voted in favour of pan-scan presentation of widescreen cinema film and believes that the major drawback to pal picture quality is that imposed by weak or multipath signals. Pal picture quality could be improved by using the BBC-developed Weston Clean pal system which, as Richard Storey has stressed on several occasions, "has progressed to the stage where complete transparency and freedom from cross-effects can be achieved using non-ideal separation filters. In other words it has moved out of the realms of mathematical possibility into the real world".

Clean-pal techniques, which compensate for basic errors in the theory of conventional luminance-chrominance sharing (mixed highs), could be applied to standard pal or indeed to wide-screen pal. But its application to 4:3 pal would not alienate existing viewers, while opening the way for a new generation of improved quality 625-line sets. Now the project seems to be in abeyance.

Can't see the difference?

More than a decade ago, a study by Dr Raymond Wilmotte showed that for domestic viewing there is no point in increasing the number of TV lines above about 600-700 for 660mm displays viewed at the usual distance of 4 to 6 times picture height. A more recent study has concluded that no difference between an NTSC (525-line) and an HDTV display would be discernible by the untrained viewer when shown on a 635mm diagonal as "normally" viewed in the home at six to eight times picture height. (See also **Table 1**, Stan Baron's comparison of equidistant viewing experience.)

As a result, how logical is it to provide a high-cost system based on the belief that viewers may want to sit close (three times picture height) to a large display while the option is available only at costs well beyond consumer budgets?

Pat Hawker

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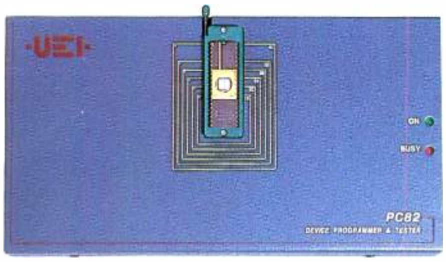

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