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The data path to multimedia

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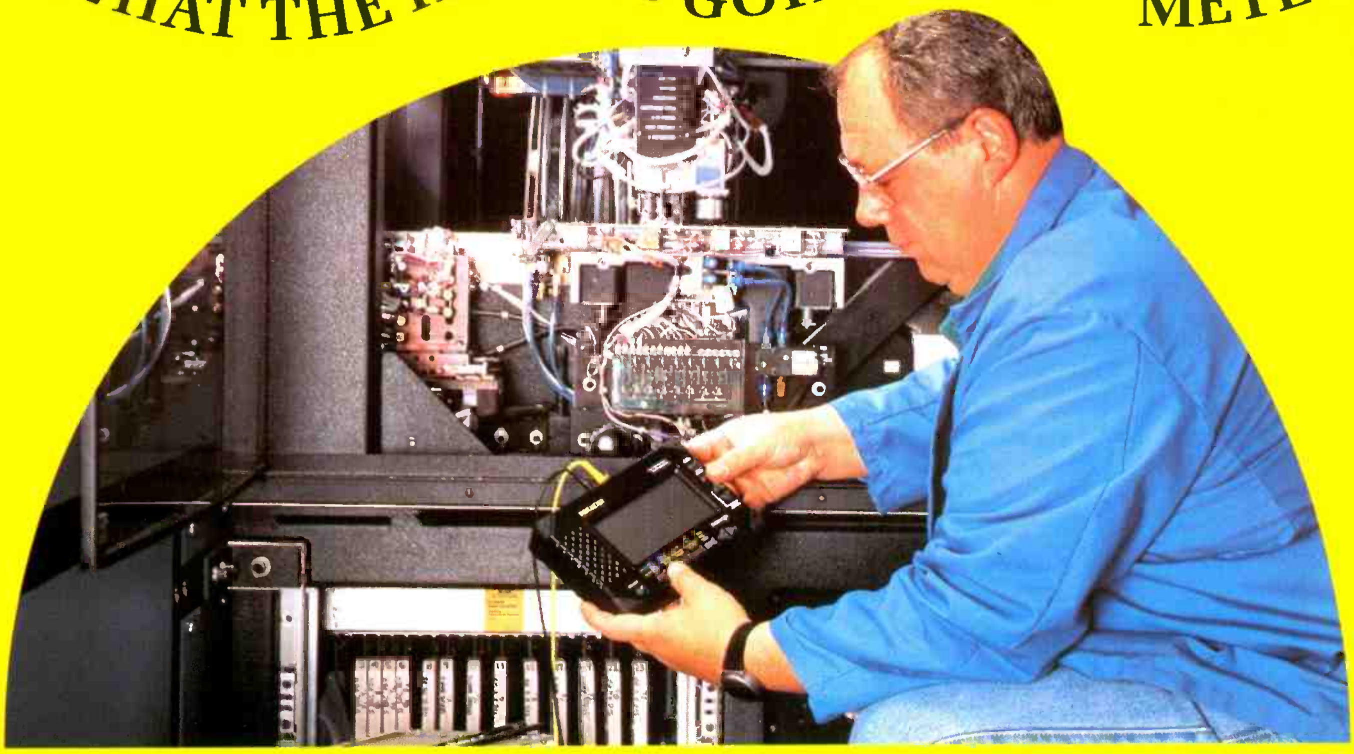

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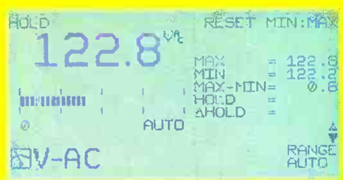
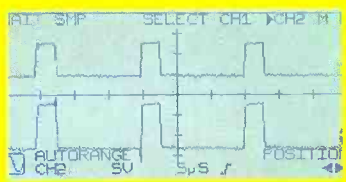
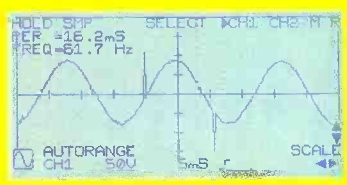


Large LCD Display gives clear indication including Max, Min and Max-Min values.



Dual channel operation allows relative timing of signals to be observed.



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CIRCLE NO. 100 ON REPLY CARD

CONTENTS

FEATURES

FREE: Crystal sample pack p659

REGULARS

In next month's issue: In next month's issue: Circuits for rms conversion. There only two types of circuit which will measure the power content of a signal directly. You can either amplify it and use the resulting output to heat a thermocouple, or you can process the signal with a root mean square algorithm. Danny Ayers tells all.

Plus a FREE pair of ZTX689B Zetex super-performance power transistors for every UK reader.

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REED BUSINESS PUBLISHING

Its all in the game

One often gets the impression that electronics for such things as business, defence, communication and scientific research leads development of the art. This may just about be true at present, but not for much longer. The drive increasingly comes from consumer leisure.

We are about to see real changes in our everyday experience coming from consumer electronics. And they will also affect the way in which we do business. Let me give an example of consumer technology affecting our own business, magazine publishing.

British Telecom and other PTTs are currently experimenting with video-on-demand, a service to deliver your personal viewing schedule directly to your home. At present, if you don't like the terrestrial or satellite offerings of an evening, you must rent a video from your local newsagent or wherever. Very soon, the PTT will be able to deliver your viewing choice in 1.5Mb/s digital form directly to your home via the same exchange line which operates the telephone. The viewer dials in a viewing request and back comes the film down the telephone line.

Compare this with the existing business equivalent. While a great improvement on the 14.2kbaud of a top quality analogue line, data at 64kb/s from ISDN seems primitive by comparison. Video compression and modem technology for consumer use is streets ahead of the commercial equivalent. Important for its acceptance, it will be cheap because it will have been designed for a mass market. Businesses of all sorts will jump at the chance to squirt high bandwidth signals about using consumer derived equipment. For instance, all those things which we as publishers may offer our readers on cd-rom could be made available via the telephone network. The magazine which you are now reading could soon be transmitted in its entirety by electronics. This might include the current issue plus the complete archive going back five years.

And for those people who are sceptical about the possibility of electronics ever replacing paper, I would recommend that they take a look at *Encarta*, Microsoft's encyclopaedia on cd-rom. While one may hate the Americanised content of this work, the hierarchal data access is without parallel. Everything on the cd-rom is content addressable and searchable and all subject texts include highlighted keywords for cross reference. These may be examined at will by double mouse click on the highlighted word. I assure you that if you had your magazines and other pictorial/textual matter delivered and updated regularly in electronic form, you would never look at a paper magazine again... except on a train or plane, perhaps.

But who wants to be tied to a telephone line? There are currently some 250 high power satellite transponders available for rental with operators bending over backwards to find suitable traffic. Ten seconds of transponder time could squirt this entire magazine, diagrams and all, to every corner of the country. The reader would plug his PC into a decoder connected to a standard domestic satellite receiver. We could then update your magazine once a month, once a week, once a day or once an hour...

All this goes full-circle to Arthur C Clarke's seminal article *Extra Terrestrial Relays*. Upheavals in magazine publishing will be just a single shot in the information revolution which is about to overwhelm us.

Frank Ogden.

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UPDATE

Intel and Hewlett-Packard in micro design pact

After two decades of sticking to the x86 architecture, Intel has announced it is devising a new microprocessor architecture in association with Hewlett-Packard.

Both companies claim that the reason for the link-up is a technological breakthrough that will dramatically increase microprocessor power. First fruits of the link are not expected until 1999 and Intel will carry on with its proposed development of Pentium successors P6 and P7.

Unlike its main micro rival, the IBM/Motorola/Apple consortium making PowerPC, Intel does not intend to set up a common microprocessor development centre. Instead, Intel and HP will design chip sets independently but with a common architecture.

It is expected that HP will concentrate on making high-end chips for mainframes, servers and workstations while Intel concentrates on chips for the lower-end, mainly PCs. Only Intel will sell the chips to other manufacturers but H-P will have the right to use them in its products.

The companies have been collaborating on the project since January and only announced it now because the teams of people had become so large that the project was becoming leaky.

It is generally thought that the reasons propelling Intel into the deal are a feeling that the x86 architecture is running out of steam. There is also a perception among big computer systems buyers that Intel is a provider of components for desk-top computing and not one for the heavy-duty end of the computer market.

The chips developed under the link will run all existing Intel and HP applications software. Neither company would say if applications will run under some form of hardware emulation, the emerging technique of performing the emulation function on-chip, or if they will be truly native to the architecture. Nor is it revealed whether the chips will be able to run other people's software, for instance PowerPC's. **David Manners**, *Electronics Weekly*.

PowerPC outsells Pentium

Apple Computers *PowerPC* machines have out sold their Pentium-based rivals since their launch in March.

According to market research firm Computer Intelligence Infocorp, Apple sold 66,300 *PowerPC* based systems in the first three months since its introduction in March 14. This compares with 40,900 systems sold in the same period based on Intel's *Pentium* microprocessor.

The sales figures were for the US market. Worldwide sales of the Power Macintosh were 200,900 units. The results, while encouraging for Apple, include a large initial demand for Power Macintosh models with at least 45,000 units sold within the first two weeks. Apple's goal is to sell at least one million PowerPC based systems within the first year of introduction.

The market research firm had some other good news for Apple. Profit margins on *PowerPC* based systems are more than double compared with *Pentium* models. This means that Apple can reinvest more money in new product development and marketing than competitors manufacturing Intel based systems

But despite the good sales figures, Apple's attempts to win over PC users with emulation technology running on its Power Macintosh systems has not worked. Almost all buyers of Power Macintosh models are Macintosh users. The failure to win PC customers is blamed on poor emulation technology. Apple and other companies are developing better emulation technology.

More to Harwell than nuclear reactors: AEA Technology recently opened its doors at the Culham and Didcot research centres to show of its commercial R&D capability. Its expertise in semiconductor processing and battery design are just a couple of examples where its work could be of direct relevance to the electronics industry.

The picture shows a batch of 100mm wafers about to be bombarded with protons at high energy which will deform the silicon structure to leave gaps. These gaps will later act as trapping centres or 'lifetime killers' which terminate the minority carriers in semiconductor junctions. This improves device switching time. By controlling the energy of the proton source, the gaps may be controlled in size and depth to produce an optimum substrate for bipolar switching transistors.

Other work at AEA Technology includes development of a lithium polymer battery system which allows the construction of a battery laminate just 200μm thick. This can be shaped and sized to any requirement.

World's first 64Mbit synchronous dram

Ahead of everyone, Fujitsu has announced a synchronous 64Mbit dram and says it will present engineering samples in August. Access time for the chip is just 6ns. The process used is 0.35μm.

The six-man Fujitsu design team, led by Yukinori Kodama, divided the column access path into three pipeline stages.

To increase operating speed they developed an address incrementing pipeline scheme which can concurrently access data at two consecutive addresses. Using this scheme, the area penalty is 1.5% more than that of conventional drams but it can operate at high clock rates: at 100MHz, the device allows data transfer rates of 100Mbyte/s.

Instead of the conventional two-bank structure, the designers have gone for four data banks allowing for greater independence of operating areas and, consequently, increased throughput.

Each bank has two cell array blocks which are separated by the word decoder. Each block has eight global bus lines which are connected to eight DQ pins.

The global bus lines are formed with the second metal layer and are laid out on the shunting area – they do not carry an area penalty. During a burst read/write operation, the separated cell array blocks are activated concurrently.

Bit data from each of the eight DQ pins at two consecutive addresses is read or written simultaneously. This 2-bit prefetch increases the bandwidth of a sequential data transfer.

The Fujitsu team recommends a terminated low voltage ttl interface which terminates a transfer line in the middle potential of the power supply voltage and the ground potential. The output circuit matches the impedance of the transfer line and therefore the interface responds to data transfer at 100MHz.

The chip size is 21.06 × 11.02mm and contains 140 million transistors and capacitors.

PDAs fail to deliver the goods

The market for Personal Digital Assistant (PDA) devices has failed to develop as quickly as industry analysts predicted, creating problems for computer firms that expected it to be the next growth market.

According to market research firm Link Resources, fewer than 49,000 PDAs were sold in the first quarter of 1994, and the rate for the second quarter is not expected to be much better.

In addition to the slow market, Motorola and IBM report manufacturing problems for their PDAs and Compaq Computer has delayed plans for its *Mobile Companions* PDA-like devices based on Microsoft's *WinPad* software interface.

Cordless videophone uses dect

UK researchers have demonstrated the feasibility of a cordless videophone, by transmitting compressed video pictures over the dect, the European digital cordless telephone protocol.

Paul Stein, head of radio comms research at Roke Manor Research, said that it is technically feasible to use a dect cordless phone, fitted with a solid state camera and screen, as a video telephone. Siemens, which owns Roke, is believed to be interested in commercialising the technology in its dect handsets.

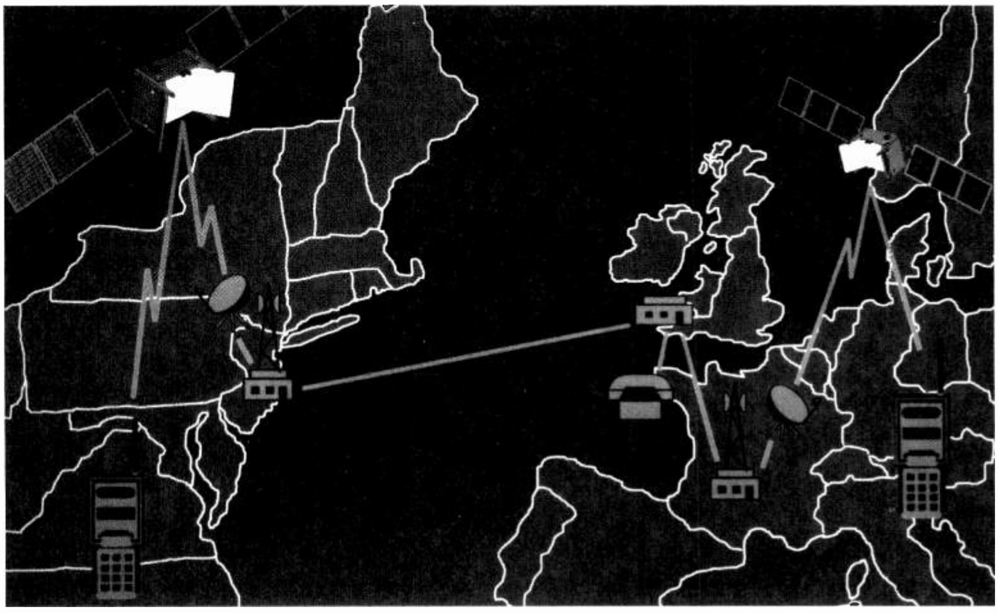
Roke's first demonstration system is a slow scan tv format which allows a cordless network of dect handsets to be used for security surveillance. The system uses a single timeslot in the dect 32kbit/s radio channel to transmit a compressed video picture running at two frames per second. A dect transceiver fitted with a camera transmits the video data via a standard wall-mounted base station to a second handset linked to a PC, which displays the picture.

Although dect's ADPCM coding scheme limits the picture quality achievable in a single 32kbit/s channel, Stein said that picture quality can be increased by using more than one time slot to transmit the video data.

Standard H.261 videoconferencing codecs rely on a minimum 64kbit/s data channel, which can be supported by the dect protocol using two 32kbit/s slots. Because time division multiple access (TDMA) radio carrier modulation is used, the relevant time slots need not be adjacent and can be multiplexed with other traffic on the cordless network.

Video data inevitably makes heavy use of the 12 available radio channels in each dect carrier and consequently there is a trade off between picture quality and network capacity. Stein, however, is confident that video telephony, running at 30 frames per second is realistic.

Siemens is said to be very interested in the possibilities of cordless videophones.



Callbox in the sky: US dominated consortium Globalstar, one of around half a dozen groups planning to offer satellite based mobile telephone services in 1998, claims that it has the simplest proposal. Forty eight low earth orbit satellites will act as radio basestations for a new breed of dual mode handsets which will work on either existing terrestrial cellular networks or the CDMA based Globalstar network. There is no complex routeing of calls in space, radio calls are quickly beamed back down to an earth station and onto the local PSTN. Globalstar's partners include cellular operator Vodafone and CDMA specialist Qualcomm.

Nynex backs video on demand

Nynex, one of the largest UK cable operators, will offer the first video-on-demand services on its broadband cable network by the end of the year, and it is investing £200m in the venture.

The cable company is buying a head-end system and set-top decoders from General Instruments in a contract worth at least £120m over the next six years. A Nynex spokesman said the value could be larger if the market for video-on-demand took off. "This gives us a potential for video-on-demand and we will be offering a kind of video-on-demand service late 1994, early 1995," said the spokesman.

Nynex is also buying £80m worth of switching and transmission systems from Nokia Telecom which, like the General Instruments equipment, will be deployed in the company's ten franchises in the north west of England.

This represents Nokia's largest contract with a UK cable company. It will supply it *DX200* digital switches which will support a total of 100,000 telephone lines.

General Instruments' MPEG-based ICFT 2000 set top box will allow users to select, order and pay for individual tv programmes.

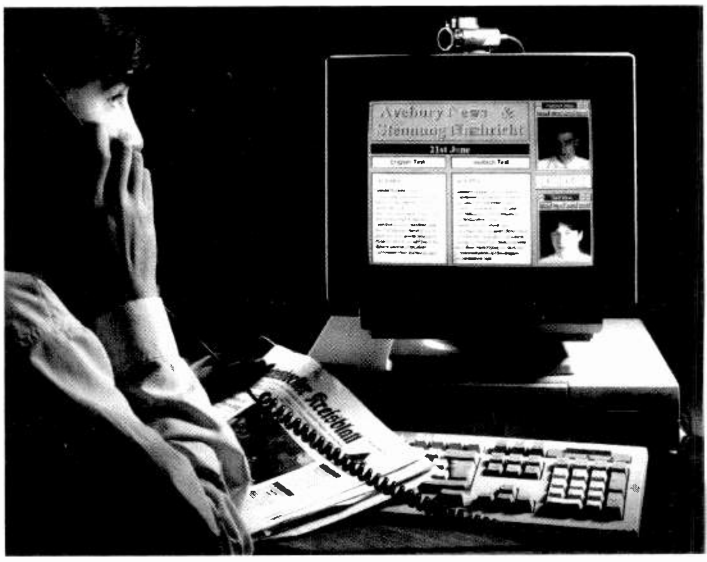
Last year Nynex executives said there were no plans for interactive TV services. The U-turn is possibly a response to BT, which is testing a service using ADSL technology on analogue telephone lines this year.
Richard Wilson, *Electronics Weekly.*

British police test remote control car stop

Three British police forces are testing radio technology which will enable them to remotely control and stop stolen vehicles. The police in Northumbria, Kent and London are testing an electronic device, called *DemonScan*, which once fitted in a car as an anti-theft device, allows remote control of the vehicle's speed, down to between 20 and 80 per cent of its full value, and eventually stopping it.

The system activates itself once a thief has bypassed the immobiliser system, causing it to send a signal which can be picked up by police vehicles fitted with a radio transceiver up to 1 km away. The police car can then follow it and gradually reduce its speed by remotely controlling the car's fuelling via its engine management system.

DemonScan, when perfected, will add £250 to vehicle price but will be free to the police force. The company behind *DemonScan*, is the Kent based Knightwatch, which developed the device under the 'Sold Secure – Partnership Against Car Theft' organisation, that provides testing and endorsement services for vehicle security devices.



School to school: ISDN was at the centre of a technology link-up demonstration between St John's School, Marlborough, and Bookholzberg School, Gandesrkesee in North Germany. Arranged by multimedia expert Digithurst, the demonstration is part of a European Union research programme. The system includes video conference, translation and co-working facilities. The latter lets two people work together remotely using the same computer screen. The system is designed for use in schools as a language learning aid, and also in the publishing industry.

Microsoft loses software patent battle

Microsoft and Stac Electronics have settled their differences on data compression for computer disks. Although the two companies now talk of cooperation, the agreement is a climb down by Microsoft, following a decision handed down in June by a judge in Los Angeles.

In February, Judge Edward Rafeedie held that Microsoft's *DoubleSpace* infringed Stac's patents and a jury awarded Stac $120 million damages. Microsoft took the module out of dos in the USA, and gave customers a token promising a new system when it was ready.

In granting the injunction Judge Rafeedie was at pains to reassure the 15 million people who have bought the Version 6 of dos which includes *DoubleSpace*, and the near 100 million others who use earlier versions of dos which have no compression. They can all continue to use their PCs.

Stac's two US patents, USP 4 701 745 and 5 016 009, claim the root idea of searching data for sequences of bytes which are identical to sequences already processed, and then storing only code which identifies the location of previously stored identical sequences.

Within weeks Microsoft and Stac Electronics settled a cross licensing agreement. Microsoft will now be able to use Stac's compression technology.

In return, Stac will be able to use the 'pre-load' technique which Microsoft developed for seamlessly strapping a compression system to the MS-DOS operating system on which 90% of all personal computers now rely.

Microsoft will pay Stac $1 million a month in royalties, for 43 months, and pay $39.9m for shares in Stac.
Barry Fox

Full colour led display

Integrated full-colour led based displays can now be built thanks to a breakthrough from Toshiba scientists.

They have developed a high-brightness blue-green led that can be constructed with conventional red leds on a single substrate to create full-colour displays. Leds have found favour in outdoor displays because of their high brightness, but full-colour displays could not be built because the blue primary colour was lacking.

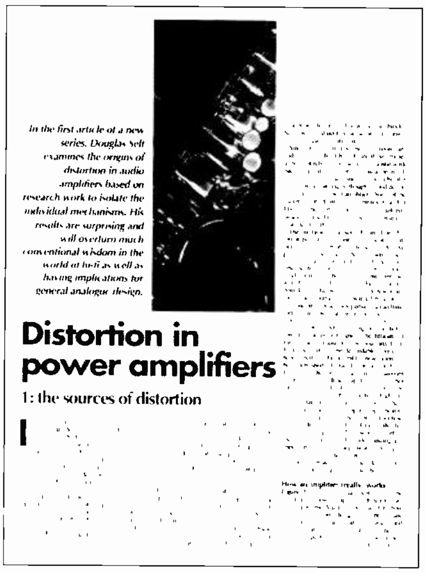
The Toshiba led structure employs p and n-type layers of zinc selenide (ZnSe) grown on a gallium arsenide substrate. The blue-green led produces a brightness of 2Cd at a wavelength of 500nm. Its spectra width is just 10nm. The structure is designed to ensure current flows horizontally across the p and n-type layers.

Self wins *EW+WW*/H-P writer's award

The winner of this year's Author Award Scheme sponsored by Hewlett-Packard is Douglas Self for his series of articles *Distortion in power amplifiers* which appeared between July 93 and March 94.

The judges, who were unanimous in their decision, were impressed both by the original observations presented in the series, and the rigorous investigations undertaken by Douglas Self used to support his assertions.

Douglas was presented with his prize, a Hewlett-Packard *HP54600A* 100MHz digital storage scope worth £2500, by Alan Grahame, UK director of Test and Measurement at the company's Bracknell headquarters.


In the first article of a new series, Douglas Self examines the origins of distortion in audio amplifiers based on research work to isolate the individual mechanisms. His results are surprising and will overturn much conventional wisdom in the world of hi-fi as well as having implications for general analogue design.

Distortion in power amplifiers

1: the sources of distortion

Calling rf designers: 94/95 Writer's Award features £4000 prize

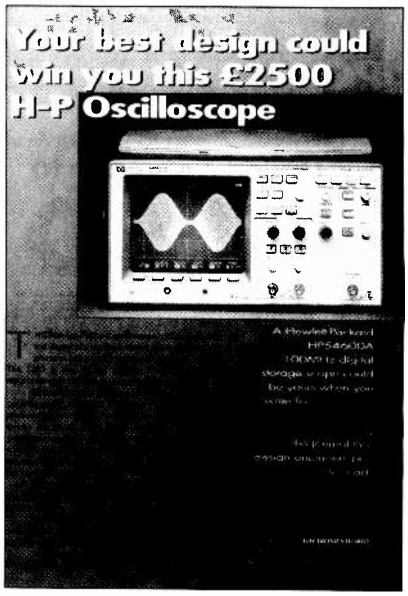
Following the success of last year's Writers Award, *Electronics World + Wireless World* and Hewlett-Packard will be launching a new scheme to run from October 1, 1994 to September 30, 1995.

The prize for the coming year's award is truly magnificent: a £4000 Hewlett-Packard HP8647A 1GHz programmable signal generator with HPIB interface, solid state programmable attenuator and built in AM-FM modulation capability.

Unlike last year's scheme, only articles which have an element of rf design will be eligible for consideration by the judging panel. It is hoped that this year's award will focus writer interest on rf engineering in line with the growing importance of radio frequency systems to an increasingly cordless world.

The aim of the award scheme is to locate freelance authors who can bring applied electronics design alive for other people. We want to commission articles on circuit design using the wealth of modern components and techniques now available but, for this year, with a focus on rf and microwaves.

Qualifying topics might include direct digital synthesis, microstrip design, application engineering for commercially available rf ICs and modules, receiver design, PLL, frequency generation and rf



measurement, wideband circuit design, spread spectrum systems, microstrip and planar aerials... The list will hopefully be endless.

All articles accepted for publication will be paid for – in the region of several hundred pounds for a typical design feature.

If you have a potential article to discuss, please call us on 081-652 3128, or write for further details to The Editor, *Electronics World*, Quadrant House, The Quadrant, Sutton SM2 5AS.

Rough ride for video CD

Video CDs are now going on sale amid total confusion and compatibility problems in the PC and video market.

Stickers on the covers of the first discs (*Pavarotti* and *Dinosaurs* from Castle Communications) explain that "video cd will play back on cd-i players with fmv cartridge, Amiga CD-32, 3DO, PCs with MPEG, Apple Mac with MPEG". Inside an insert slip promises that, "This disc can be viewed on any player which is MPEG compatible". Wisely Philips plays it safer and labels its first White Book CDs (movies such as *Indecent Proposal*) with talk of playback on cd-i players or any system that is "compatible with the video cd standard".

In theory a video cd should play back on any PC fitted with a cd-rom drive and MPEG video decoder board (eg Sigma Designs' *ReelMagic*). In practice it will often display pictures which lurch and jerk, with sound that spurts in bursts.

This may happen on fast, high spec PCs, hooked to fast, high spec rom drives. To add to the confusion, the same PC rom systems may deliver smooth motion and sound when playing MPEG game discs marketed by Sigma.

So far the manufacturers have failed to give a clear explanation for this.

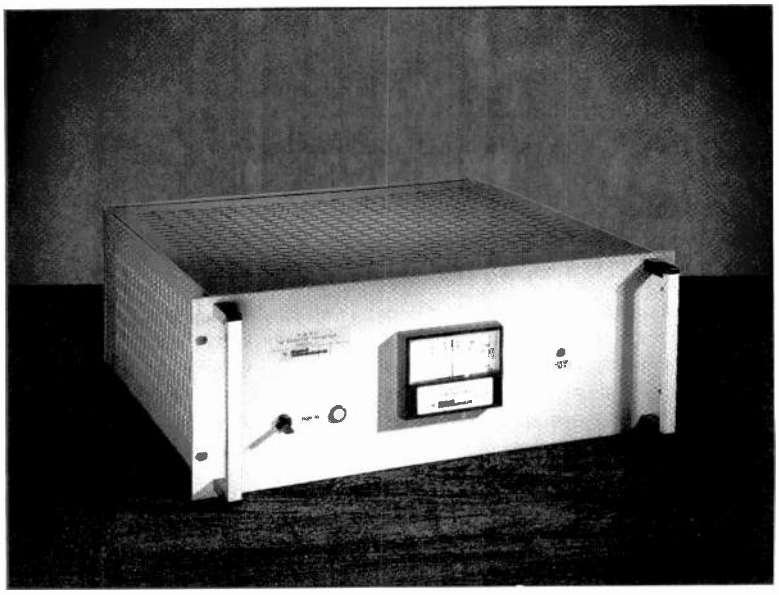
There are now three types of cd-rom disc.The MPEG CDs sold for use with the *ReelMagic* board do not follow the White Book standard. Instead they package the MPEG video as ISO 9660 rom files which any fast rom drive can read at full speed. This is why a rom drive and RM board may play RM's MPEG discs perfectly, while failing to play White Book video cds properly.

Sony has modified its 33A rom drive to work at full speed in 2/2 mode. The modification involved changing rom chips in the player which store the firmware. Sigma now bundles the Sony drive with an RM decoder as an upgrade kit.

This will be no consolation to all those PC users who have already bought what they believe to be fast drives, and now find that they are obsolete. It will be no consolation to PC users who now buy new drives, or multimedia PCs, before the manufacturers are forced to admit that they are obsolete. Even if manufacturers provide new firmware in replacement rom chips, retrofitting will be a clumsy and expensive operation.

The situation will not resolve until the 2/2 requirement is publicised and manufacturers start labelling rom drives as 'White Book' or 'video cd compatible'. No manufacturer with a warehouse full of White Book incompatible drives will start 'compatible' labelling any sooner than absolutely necessary. **B. F.**

TELEVISION TRANSMITTERS AND TRANSPOSERS



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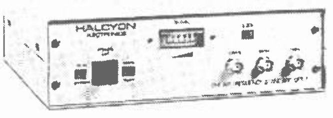

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RESEARCH NOTES

Jonathan Campbell

DSP flattens steel precisely...

In the arduous environment of a steel rolling mill, DSP can be used to track the distance between rollers up to 60mm apart to a resolution of 0.25µm.

DSP technology could find itself being used in the anything but quiet environment of a prototype steel rolling mill. Engineering research consultant Adrian March has developed a system that relies on dsp to process linear measurements in absolute terms rather than relative ones.

Normally coordinate measuring systems simply count the number of scale positions between one position and another. March's optical device determines the 'actual' position under the reading head so it can not be thrown off by dirt or power cuts – or steel sheet slamming into the mechanics. In the steel mill the system will need to measure roll separation of up to 60mm with a resolution of 0.25µm.

Measurement depends on reading a transmissive scale consisting of binary pseudo-random patterns. By reading a small length, absolute position can be determined.

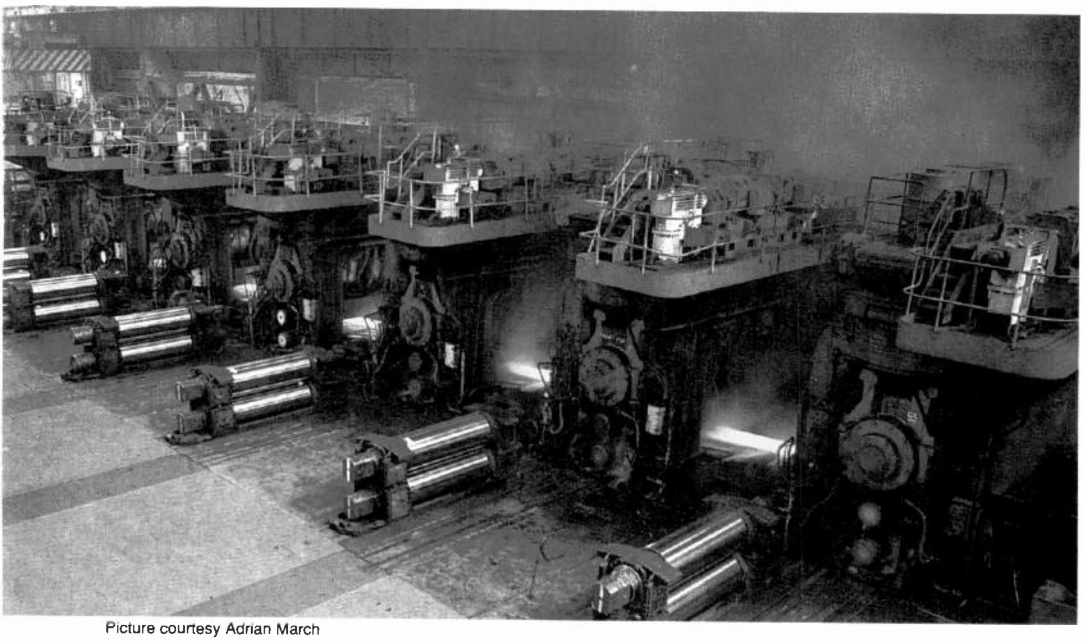
Imaging optics have been replaced by a lens that focuses light from a led into a parallel beam to shadow the scale on a standard linear charge-coupled device. The image from the sensor is read out at intervals between 0.5-5ms.

DSP interpretation consists first in recognising the position to the nearest ccd pixel of the primary pseudo-random number pattern. The process involves a mathematical correlation which establishes the position in which the best match is obtained against a reference number string. That done, an interpolation can be performed to a fraction of a scale division.

Any variations in scale accuracy are overcome by reading at least 256 scale divisions at all times. For highest accuracy a scale can be calibrated and a look-up table used by the processor to correct residual errors at all points along its length.

In theory, March says, the system can measure up to 24m and down to fractions of a mm, with claimed resolution of an Angstrom.

A second system, aimed at a different application, is due to go into production at the end of this year. It will measure a distance of 1mm with a resolution of 1nm.



Picture courtesy Adrian March

...and sharpens hearing

Fast developing dsp technology could hold out new hope for millions of partially deaf people who are unable to benefit from current hearing aid design, if work being carried out at Washington University comes to fruition.

Devices that work and people actually want to use would be the result, according to Washington's A. Maynard Engebretson who has been testing a dsp-based device with impressive results.

Professor Engebretson, who holds posts in both the Department of Otolaryngology and the department of Computer Science at Washington University, says (*IEEE Engineering in Medicine and Biology*, April/May 1994, pp.238-248) that today's hearing aids are simply not good enough. "It is not uncommon for a person to buy a hearing aid and then not use it because it does not help very much", he says.

But with dsp technology: "it would appear that the stage is set for a major advance in hearing aid development".

Engebretson likens the impaired ear to a flawed instrument such as a spectrum analyser with defective channels.

Part of the difficulty with hearing aid design is that the ear behaves differently for soft sounds near the hearing threshold than it does for loud sounds. There are also physiological and psychological characteristics of an impaired ear that have to be taken into account. The situation is complicated by the fact that hearing aid development requires a knowledge of acoustics, transducers, signal processing, auditory physiology and psychophysics as well as low power semiconductor technology.

Current hearing aid design using analogue circuitry is very close to its limits in terms of size constraints and fundamental transistor scaling considerations. Dynamic range of an op amp – the building block of an electronic filter – decreases as the three-halves power of feature size. With hearing aid dynamic range only just being acceptable now, further size reduction is not practical. But the elementary amplifier (inverter) of a digital processor requires only enough dynamic range to resolve two states.

Overall dynamic range is determined by the number of digital stages used.

In a digital processing structure, complexity increases as the inverse of the square of the feature size and switching energy decreases as the cube of feature size. So as feature size is reduced, greater dynamic range and complexity is possible with less power consumed.

Modern hearing aids still reflect their 1940s origins, though growing from single-transistor amplifiers to modern multichannel designs containing thousands of transistors. Multiple channels attempt to provide different compression characteristics for different frequency ranges and tiny potentiometers are often used to adjust the compression characteristics. Newer hearing aids may be electronically programmable, with parameters down-loaded from a computer-based fitting system and stored in digital registers.

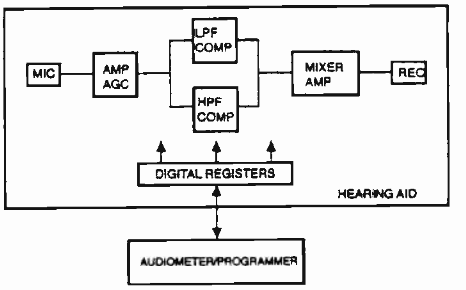
But tests on digital hearing aids carried out at the Central Institute for the Deaf at Washington University could radically alter hearing aid design. Functionality of the prototype devices includes four-channel compression, adaptive noise reduction and feedback cancellation. For example, feedback cancellation, using an adaptive fir filter, provided an additional hearing aid gain margin of between 10-15dB, meaning roughly an additional 20-30dB of patient hearing loss that can be accommodated by the aid. Partially deaf people, using the device adjusted to a gain of their own liking for various conditions, improved their word identification scores by 7% with this facility.

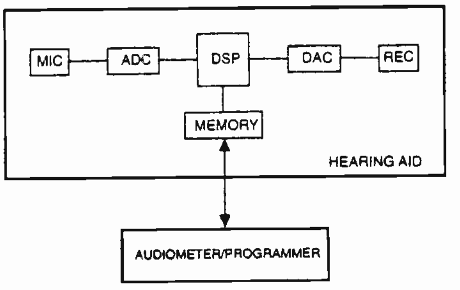
Adaptive noise reduction creates an estimate of the stationary components of signal and noise then uses an inverse filter to reduce hearing-aid gain at frequencies where the estimate is large. Engebretson says noise reduction was preferred by patients for both soft and loud speech, and word intelligibility scores was much improved for soft speech in the presence of machinery noise.

FIR filters were also used to experiment with what Engebretson refers to as level-dependent-spectral shaping (LDSS). Work was based on the fact that hearing-impaired people preferred different gains and frequency responses depending on signal level. Frequency responses of two channels were set to the person's preferred frequency responses when listening to soft and loud speech and the other two channels set at in-between, interpolated, values.

Though there was little difference in loud speech in noise conditions, soft speech in noise using LDSS produced significant improvements over a wide range of signal levels.

Much work still has to be done before dsp hearing aids become a reality. One question yet to be answered is: what exactly is required from a hearing aid?





Top – multichannel hearing aid pushing current hearing aid design to its limits. Bottom – dsp design could open up new possibilities for people with hearing impairments.

Engebretson summed up the current position as that we are technically at a point where we can create very sophisticated hearing aids. But the question is what exactly do we want to do?

"Building circuitry is just a matter of putting up the money and getting people together. The problem is that we don't have nice neat categories of hearing impairment. Research depends so much on the individual patient that it is difficult to produce general solutions".

Sharing the secret to scientific problem solving

How secret is secret? If you've got a PC, 599 computer-powered friends, access to the Internet communications network and eight months to spare you might just be able to find out. That's what it took Arjen Lenstra of Bellcore in Redbank, USA, Paul Leyland of Oxford University and colleagues at MIT to unzip one of the methods used by cryptographers to keep messages for their eyes only. But the record-breaking cooperative computing feat may have ramifications far outside the arcane world of cryptography.

What the code-breakers did was to find the two massive prime factor keys of a 129-digit code. In effect they had cracked the code used in the RSA method of encryption where two very large prime numbers are multiplied together to create a third gigantic number. The shear physical size of the numbers involved makes it very difficult to find those original factors from the product alone. Decoding encoded messages depends on knowing both factors as well as the large number that has been used to code the message.

The degree of security given by using a 129 digit product can be judged from the massive amount of computing power and time that was needed to factor it. But as far as cryptographers are concerned, the code *was* cracked, and indicates that they must move to using much bigger numbers, perhaps 400 digits long.

There is, however, a much broader lesson for the scientific community in terms of problem solving.

Dividing a task between 600 volunteers running 1600 machines varying in size between parallel supercomputers and humble PCs, is the biggest demonstration of distributed number-crunching so far attempted.

In the search for the factors, with such large numbers involved, the only way to progress was to use an algorithm to look at the relationships between pairs of numbers and see how close they come to a solution. All those relationships were then collected together and cross-referenced in a database.

Paul Leyland of Oxford University told *EW + WW* that finding the prime factors involved a search similar in magnitude to needing to find millions of needles in a galactic-sized haystack.

"We were using an algorithm that looked for interesting relationships between numbers. When we had enough interesting relationships, we put them together".

But there are an infinite number of relationships and only one relationship in 100 million is interesting. The team had decided they needed 8.2 million interesting bits of data to solve the puzzle, meaning they had to investigate 8.2 million x 100 million relationships.

Their method was to call on volunteers to link up to a central MIT computer, using electronic mail, and down-load the necessary code to enable them to do their own chunk of processing. The MIT computer also gave them instructions on what to do and a place in the haystack to start searching for their needles.

When their program had been processed, volunteer data was e-mailed back to MIT where it was batched up.

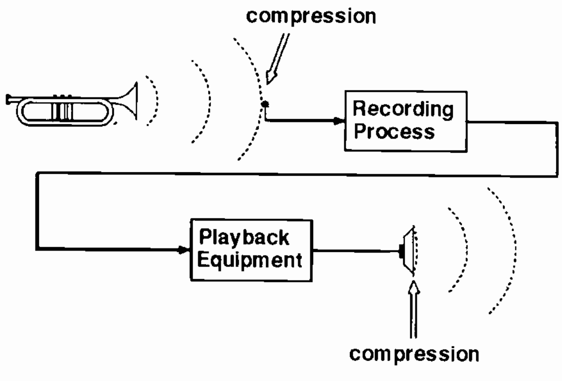
After eight months toil, two prime numbers of 64 and 65 digits, were squeezed out the other end.

"Similar techniques have been used over the past 5 years though ours was by the far the largest project to do this", says Paul Leyland.

But without doubt, the real achievement was the coordination of so much, and so diverse, computing power in the pursuit of a single scientific goal.

Polarity proposal that is music to the ears

Would the kettledrum sound different if struck from the inside rather than outside? It might seem the sort of problem to worry only the most dedicated of percussionists. In fact it is part of the much wider issue of acoustic polarity – can we hear when polarity of reproduced music has been reversed: when loudspeakers have been incorrectly wired up for example? The answer is a qualified yes, but you have to look hard to find out cases where it matters.



If compressions at the microphone were reproduced as rarefactions at the loudspeaker, the polarity of trumpet sound reaching our ear would have been inverted. But would we notice the difference?

The researchers who posed the kettledrum question, throwing down the gauntlet for avant garde musicians the world over, are Richard Greiner, professor of electrical and computer engineering at University of Wisconsin, and Douglas Melton, a sound and vibration specialist at Digisonix Inc. They were prompted (*Journal of the Audio Engineering Society*, Vol 42, No 4, pp.245-253) by the fact that polarity inversion does not distort the phase and amplitude relationships between frequency components of a signal, nor does it change the temporal shape of a transient signal. But it does present the ear with a fundamentally different signal – compressions are replaced by rarefactions and vice versa.

The eardrum at any instant is "pushed" instead of "pulled", an effect that audiophiles have long wondered if they should include on their hit list along with the need for reduced phaseshift, phase linearity and minimum phaseness in audio equipment.

Up to now, there seems to have been little attempt in recording practice to maintain original polarity – in the consumer market Greiner and Melton describe the situation as "total chaos". But with processing in the digital domain making it easier to control polarity, they decided that now was the time to determine the contribution that absolute acoustic polarity makes to the accuracy of reproduced sound.

Previous experiments have proved that when the polarity of a simple waveform is switched, listeners describe a change of pitch or timbre of the signal. So Greiner and Melton set up a listening room in such as way that this effect was apparent to almost everyone. Then with the help of 39 students at Wisconsin they set out to listen to a range of instruments played in various styles to see if the effect could still be heard.

Musical examples demonstrating large asymmetry in the time domain were selected, as this property was felt to be significant when a signal was inverted, with individual instruments highlighted in the pieces. The result was that the ability to hear the polarity inversion gave only a slight positive bias, with piano and classical guitar yielding significantly higher correct responses.

The masking factor seems to be that music is usually so complex that there is often too much going on to allow human concentration on subtle effects, and vibrato, tremolo and instrument filigree may obscure inversion.

The fact that the guitar and piano were most often picked out when inverted, instruments showing little asymmetry, suggests that is not the property most affecting inversion. More likely, they say, is that there are other psychoacoustic effects caused by the attack and decay properties of the signal that help the ear identify the correct acoustic polarity of the signal.

The authors conclude that though polarity inversion is not easily heard, it can be picked up in certain circumstances. It makes sense to keep track of polarity during the recording process to avoid any problems later. Then, when someone actually does play their kettledrum from the inside, we'll all be able to appreciate their efforts. ■

John Wilson has relinquished his post as *EW+WW* Research Notes creator – a function he has executed splendidly for the past eight years. This results from his promotion to a new and more demanding position at the BBC. Although this means the loss of one of our most valued contributors, we wish John every success in his new position.

Taking over from John is former *EW+WW* deputy editor Johnathan Campbell, who has a lust for scientific knowledge and a keen nose for a story. Of course we wish Johnathan every success too.

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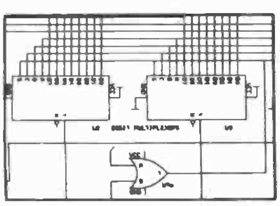

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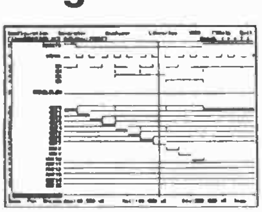
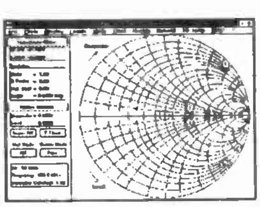
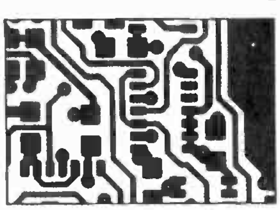


Modify the configuration and change component values until the required performance is achieved.



PCB Design



The design, complete with connectivity, can then be translated into the PCB. The connectivity and design rules can be checked automatically to ensure that the PCB matches the schematic.

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CIRCLE NO. 111 ON REPLY CARD

Teletext via your PC

Today's teletext broadcasts contain a wealth of information on diverse subjects from share prices to weather. Laurence Cook's teletext plug-in for the PC opens up the possibility of grabbing this information and printing it, processing it or turning it into a database.

Broadcast teletext services are a useful, regularly updated and essentially free source of information on a wide range of topics. Accessed via a tv screen however, teletext information cannot be printed or saved for future reference.

A solution to the transient nature of teletext is the subject of this article. Described here is a complete teletext adaptor that plugs into a single slot in your PC. It accepts a standard antenna input and appears to the PC as a standard serial communication port.

With suitable software, your PC becomes a teletext terminal with the ability to print or save pages to disk or ram. Built into the adaptor's tuner is an intermediate-frequency stage together with high-performance, factory pre-aligned video and audio demodulators. As a result, baseband video and audio signals can be made available via sockets at the rear of the PC. These can feed multimedia cards or other devices which might rely on such signals in the future.

System requirements
IBM PC-AT or compatible with DOS 3.3 or higher
640K ram minimum
Hercules, CGA, EGA, VGA or SVGA graphics adaptor
Free 8 or 16 bit slot for $^3/_4$-length card.

My complete teletext adaptor is a three-quarter length printed circuit board laid out entirely for conventional through-hole components. As such it lends itself to small quantity and one-offs. Details of the board are presented at the end of this article.

Design overview

As is clear from **Fig. 1**, the adaptor divides into distinct functional blocks.

Using a ready built module with baseband outputs for the tuner section save a lot of design and commissioning work. Made by Philips, the chosen module has a digitally synthesised front-end. Using an enclosed and decoupled module also avoids problems with emi emanating from inside the PC.

Control of the tuner is carried out via an I^2C bus interface, which is common in many modern TV sets. Synchronous clock and data lines of the I^2C bus are driven by simply toggling bits on a parallel port.

Since I had already developed PC terminal software for an external adaptor that connected to a serial communication port, I developed the teletext adaptor with the same interface. This has the advantage of making software easy to write. Jumpers allow setting the interface up as COM1 to COM4, with interrupts IRQ3 or IRQ4. More on this later.

Because the Philips tuner has varicap diodes it needs a stabilised 30V, but the PC only has 12 and 5V rails. A simple charge pump generates about 50V from the PC's 12V rail. This is then shunt stabilised to 30V. The 5 and 12V supplies are used by the adaptor's co-processor and analogue circuits respectively. They are first filtered by small inductors and decoupled capacitively.

Off-air video from the tuner module passes to the video interface. Here, recovery of the 6.9375MHz clock and data from up to 16 data lines per tv frame take place.

Within the video interface, data slicing is carried out by a Philips *SAA5231*. This device connects to a separate level shifter which converts the chip's output to ttl levels for connection to the digital block that follows. Off-air sync is extracted by the chip and conditioned as line and frame syncs before also passing to the digital block.

In earlier designs, level shifting was performed by two expensive high-speed comparators to minimise skew between clock and data. This design uses a much cheaper dual comparator. Initially I discounted the dual comparator because it had poorer switching times combined with a broader tolerance. This could have caused reception errors due to skew with jittery off-air signals.

A little thought and experimentation revealed however that differential delay between channels is most important. Since the chosen IC has both comparators on the same substrate, the actual skew is no worse than with two separate, expensive chips.

At the core of the logic block is a 10MHz Hitachi *HD64180S* microprocessor. This highly integrated device has a *Z80*-compatible instruction set but with additional instructions and enhanced microcode.

Integrated features of the processor include a high-speed multi-protocol serial communications interface and direct memory access controller.

These were key factors in my decision to use the device. Off-air data is clocked into the serial controller and shifted by the direct-memory-access controller into memory for processing. Even with the fast serial interface feeding the dma controller, the data capture leaves only 20ms for processing before the next packet of data arrives.

Pages requested by the host PC are processed by the *64180*, which maintains an output buffer in ram. It then offloads the buffer more slowly at 9600 baud via its on-chip uart directly into the PC communication-port uart. From there, the data passes to the ISA bus and onward to the terminal software.

Design details

Tuner module. I chose the Philips *FQ844* digitally-synthesised tv tuner. It has an integral IF strip and demodulates to baseband audio and video. Designed for European channels E21 to E69, the tuner covers the entire UK uhf tv band.

Tuner input is via a standard coaxial antenna socket which is an integral part of the casing housing the electronics. Since the casting is slightly too large to pass through the PC chassis slot, it needs filing. Make sure that absolutely all filings are accounted for before attempting to plug in the board.

Remaining connections are presented as pins on the edge of the case, which is intended for vertical mounting in a tv set. There is little room between slots in the PC so the tuner is laid flat on the pcb and connected via long wire-wrap type pins. Double-sided adhesive tape may be needed for anchoring the tuner to the board.

Audio from the tuner is fed via point TP_{25} and dc block C_{31} to phono socket P_4. This is off-air sound. When debugging the adaptor it's handy to have a monitor amplifier to hear whether the tuner has actually locked on to a given channel if the data side of things still isn't working.

Off-air video from the tuner feeds a phono socket for external use but its main purpose is driving the data slicer in the video interface.

PC communications. Signals of interest on the 31 way card edge connector of the PC ISA bus are shown in **Fig. 2**. These are the lower eight data bits, SD0 to SD7, the lower ten address bits, SA0 to SA9, address enable, i/o read and write signals and system reset. Power rails and ground are also used. Interfacing these straightforward signals to the chosen uart, a National Semiconductor *NS16550AF*, is simple.

In the original IBM PC specification only addresses for communication ports COM1 and COM2 were defined. However a de-facto standard has developed defining addresses for COM3 and COM4.

My design follows the four communication-port standard, jumpers $J_{101,102}$ setting the

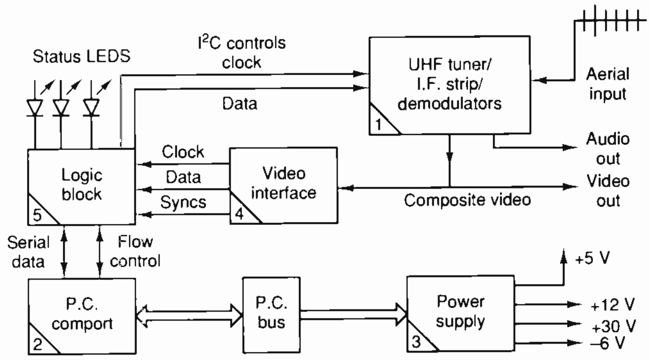
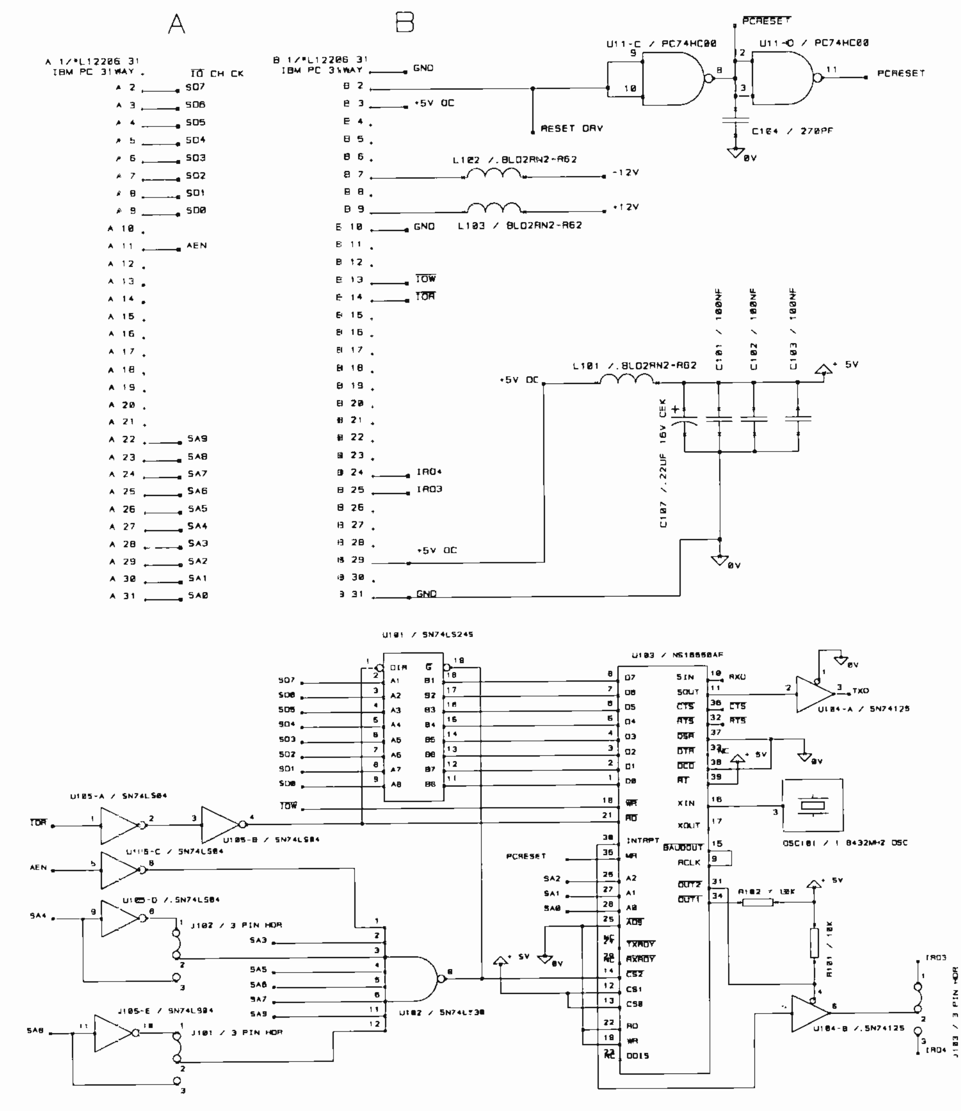


Fig. 1. Essentials of an off-air teletext decoder card for the PC. In addition to feeding teletext data to the PC, the adaptor also makes tv audio and video from the synthesized tuner available. No external power supplies are needed, but some power conditioning is required for comparators and the tuner varicap.

Fig. 2. To make hardware convenient and software writing easy, the teletext decoder is designed as a PC-slot card with an interface to any one of the four dos COM ports. The NS16550 universal asynchronous receiver/transmitter, or uart, interfaces to the PC slot using straightforward data, addressing and control signals.



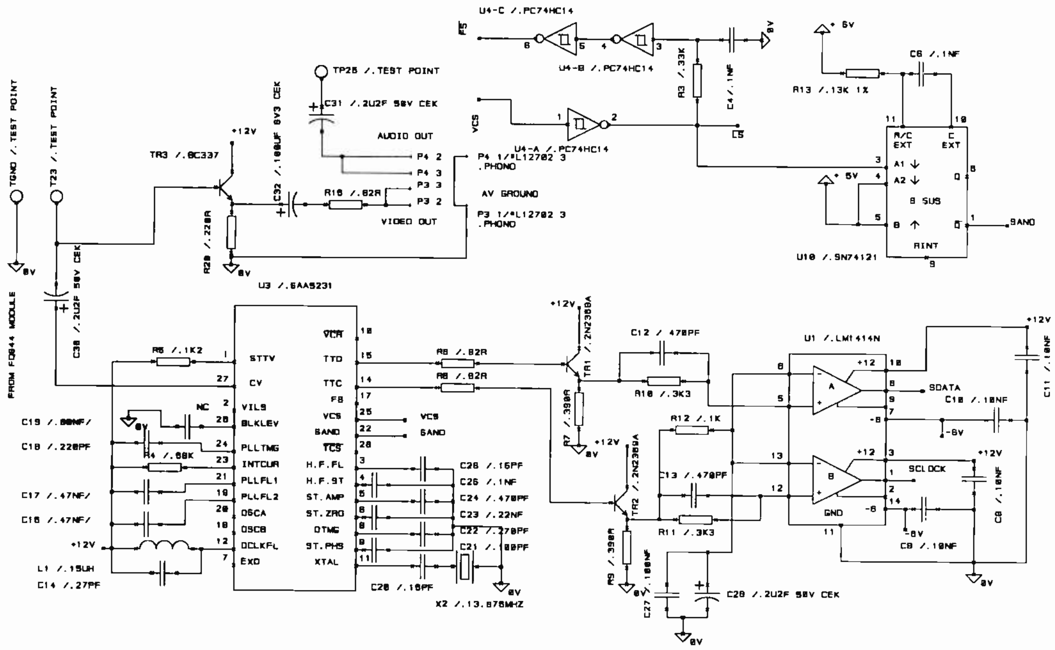


Fig. 3. Teletext information is separated from raw video by an SAA5231 data slicer. Off-air composite video appears at pin 27 while sync pulses are presented at pin 25.

required address according to **Table 1**. These jumpers select the true or complement address lines A4 and A8 for decoding together with remaining addresses from A3 to A9 in nand gate IC_{102}. This generates an active-low chip-select signal directly feeding uart IC_{103} and bi-directional buffer IC_{101}.

Buffering is needed to provide sufficient drive for the uart when it has to send data or status messages via the ISA bus. Should several cards be plugged into the bus, the uart would be overloaded if connected directly.

On the arrival of data, the uart alerts the PC by generating an interrupt signal. This is fed via IC_{104B} to ISA bus common-interrupt line IRQ3 or IRQ4, depending on the setting of jumper J_{103}. Recommended settings are shown in the table.

Serial data lines SIN and SOUT, together with handshake lines RTS and CTS, connected to the uart in the microprocessor. This provides out-of-band flow control between the microprocessor and the system, ensuring commands and data are not lost. All other uart serial port signals are ignored, being left unconnected or strapped to an appropriate level.

A separate oscillator module provides the 1.8432MHz ttl clock needed for the uart. My attempts to find a clever way of obtaining this clock from one of the other chips failed.

Video interfacing. Shown in **Fig. 3**, the video interface revolves around a Philips *SAA5231* bipolar data slicer. Off-air composite video appears at pin 27. The chip strips out sync pulses and presents them at pin 25. They are inverted by IC_{4A} and presented as active low line syncs to the *64180* uart's data-carrier detect (DCD) pin. This is configured as a general-purpose input pin.

Monostable IC_{10} produces an 8.5μs pulse every line sync which is fed back to the sandcastle input of IC_3, pin 22. This makes the chip respond as though it had other members of its family connected to it .

Timing components R_{13} and C_6 should be accurate for correct operation. In the data slicer, crystal X_2 oscillates at twice the broadcast data rate, which is exactly 444 times line rate. Inductor L_1 and C_{14} prevent the crystal from oscillating at harmonics of its fundamental.

The oscillator locks onto the data stream, and is divided by two to provide true recovered data and clock signals at pins 15 and 14 respectively. Since these signals are not at ttl levels IC_1 – a rather unusual dual comparator – is needed to convert them. It is powered from +12V and –6V, will accept inputs with a high dc offset, and produces complementary ttl outputs.

In this design, recovered clock from IC_3 is buffered by Tr_2 then integrated by $R_{12}/C_{27}/C_{28}$ so as to generate a reference level at pin 13 of the comparator. This reference is the mid-point of the signal.

After hf compensation provided by C_{13}, the signal is fed via R_{11} to pin 13 of the comparator. This makes dc offset on IC_3 irrelevant. As a result, a ttl version of the data clock appears with unity mark-space ratio at pin 1 of IC_1.

Recovered data follows a similar path through Tr_1 to arrive at pin 8 of IC_1. Delays through IC_1 are closely matched between each half, so the phase relationship between important clock and data signals is preserved.

Now, the clock and data signals are applied directly to the microprocessor serial interface clock and data pins.

Logic block. Although the heart of the system, this section is essentially simple. Referring to **Fig. 4**, IC_{14} is a watchdog chip made by Maxim, which serves as a longstop should the software run out of control. It also provides a clean power-on reset signal for the microprocessor; its output is or-ed with a reset signal from the ISA bus so pressing the PC reset button also resets the co-processor.

Inclusion of a watchdog reset timer, in conjunction with on-board non-volatile memory, IC_9, means that the teletext receiver can operate without supervision. It can survive power failure or static-induced interruptions and recover whatever configuration it was last loaded with.

Connected to the microprocessor are 32K of eprom and 32K of ram, decoded on-chip by /CS0 and /CS1 respectively. These occupy the full lower 64K physical address space. The third on-chip decode line, /CS2, is used to trigger the watchdog and is selected via the on-chip memory manager. No external decoding logic is involved in selecting memory devices.

Or gate IC_{13} generates memory and i/o read and write pulses. Access to the control port IC_{12} and associated circuitry by the read and write pulses is possible when A11 is true, /LATCH is low.

The lower three bits control the status leds, the next three control eeprom IC_9, and the top two bits produce I^2C clock and data signals. These are applied to the tuner module via points T_{13} and T_{14}. Gates $IC_{5B,C}$ direct the I^2C and eeprom data onto the microprocessor data bus from the latch when an IN instruction is executed. Thus with very little circuitry, the firmware is able to indicate its status, save and reload its configuration, and control the Philips tuner.

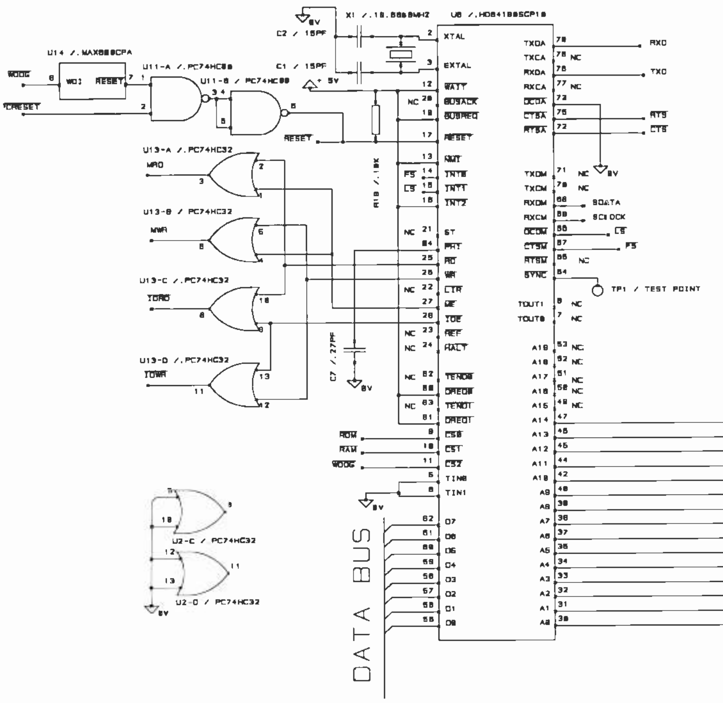
Connector P_1 is provided for future expansion, such as the control of auxiliary I^2C devices in commercial applications, or as a medium speed link to, say, text-to-speech converters. This could be a useful application of teletext for the blind.

Since the microprocessor is only available as a surface-mount plcc part, the board has been tracked to take an 84-pin through-hole socket. A bonus of the socket is that it allows the relatively expensive processor to be fitted last, minimising the chances of esd damage.

The processor needs to be a 10MHz part, specifically the *HD64180SCP10*, which can be clocked at up to 20MHz. In this design the clock is 19.6608MHz which divides down to allow the uart to operate at 9600baud exactly. In addition, the maximum data rate of the communication interface under dma transfer is a function of clock speed. This clock frequency is high enough to accommodate the broadcast data rate.

Table 1. COM-port jumper settings used on the PC teletext board

	IRQ	J101	J102
COM1	4	3	3
COM2	3	1	3
COM3	4	3	1
COM4	3	1	1



Power supply. Rails of +5V, –12V and +12V from the PC are decoupled by inductors L_{101}, L_{102} and L_{103} respectively.

Power for the data slicer IC_3, comparator IC_1, and tuner module is provided by the 12V rail. Feed to the tuner is further filtered by L_{13} and C_{15} before being applied to point T_{24} as a supply to the local oscillator. Since psu designs for PCs vary, noise levels on the supply rails are uncertain so heavy decoupling is advisable.

Inverter IC_{106} and associated components pump the +12V rail up to 50V or so, **Fig. 5**; Gate A of the IC is a free-running oscillator which drives the diode-capacitor chain, culminating in C_{114}. So many stages are needed because of losses due to output impedance of the gates.

Following the voltage multiplier, resistive current limiting followed by a shunt zener diode provides a highly stable tuning voltage for the front end. Note that IC_{106} must be the device specified for this circuit to work.

To power the dual comparator only, the –12V rail is dropped to –6V by R_{15} and ZD_1.

How it works...

Broadcast data is contained in up to 16 television lines per frame within the frame blanking period, **Fig. 6**. Each line comprises data at 6.9375Mbit/s with a logic one corresponding to 66% of white, and a logic zero to black, or thereabouts, **Fig. 7**.

Lines start with a stream of one and zero bits for clock run-in, followed immediately by a framing code to allow byte synchronisation within the receiver **Fig. 8**. Normally, framing is carried out by feeding the data into a serial-in-parallel-out shift register then examining the data with a wide nand gate and inverters. This results in a pulse when the desired pattern is present.

In this system, the microprocessor is simply programmed to look for the framing code as a sync byte. Having set up the dma controller to shift 43 bytes into ram, the processor simply waits while a line of data is moved into memory. Now the processor can sample both frame and line syncs, so it can detect what lines to expect data on, and when the last data line of that frame has occurred. At the appropriate time the processor begins to examine this data.

There is almost 20ms available for the processor to move those lines forming part of a requested page into an intermediate buffer area. In fact, up to six pages can be assembled simultaneously if specified by the terminal software. When the buffers are complete, they are transferred into the global output buffer. From there they are transferred via the uart link to the PC communication port.

In addition to the recognition and transfer process, all headers of the selected magazine are sent out in a sort of queue-jumping fashion, so as to maintain an accurate clock dis-

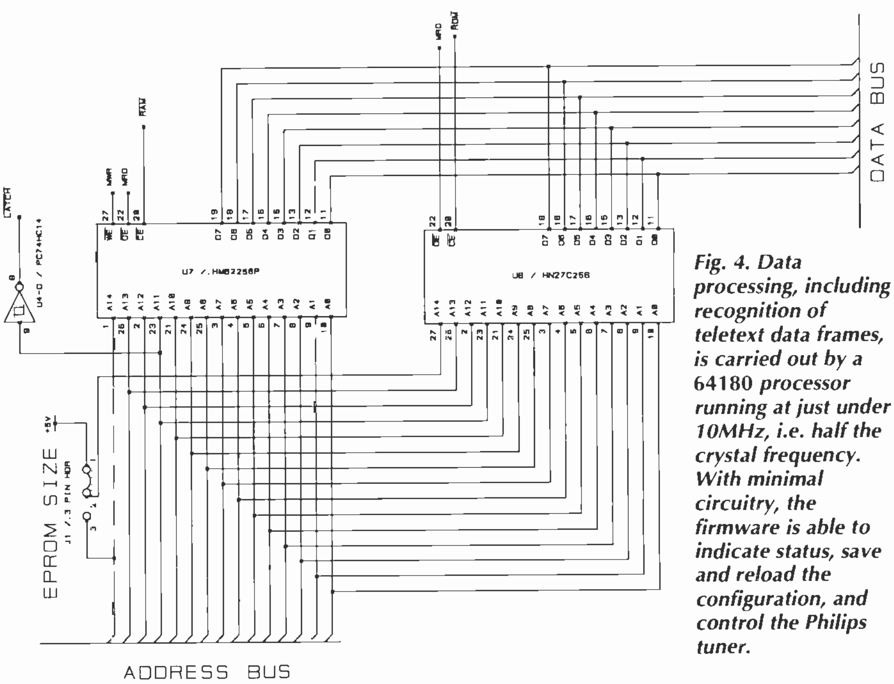
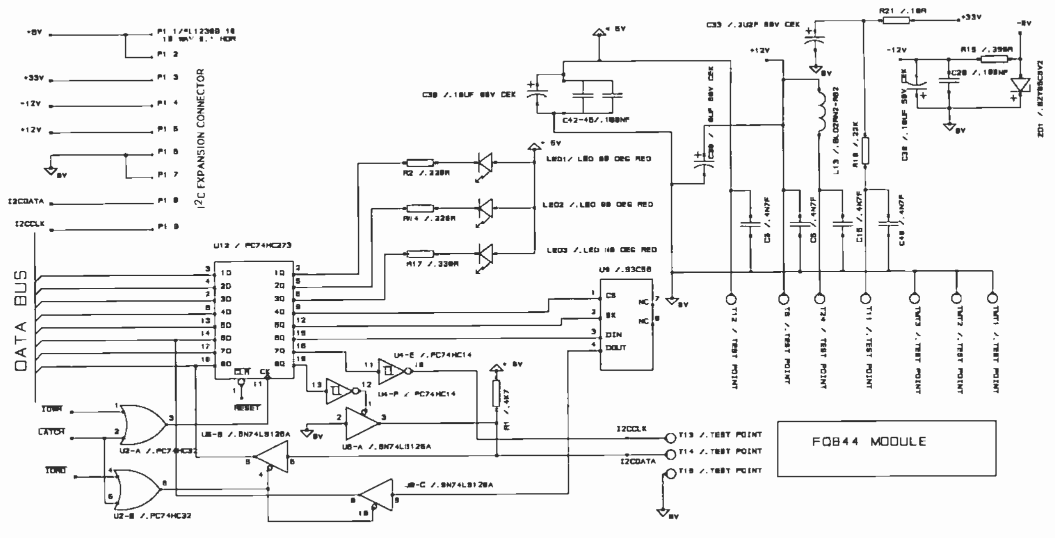


Fig. 4. Data processing, including recognition of teletext data frames, is carried out by a 64180 processor running at just under 10MHz, i.e. half the crystal frequency. With minimal circuitry, the firmware is able to indicate status, save and reload the configuration, and control the Philips tuner.



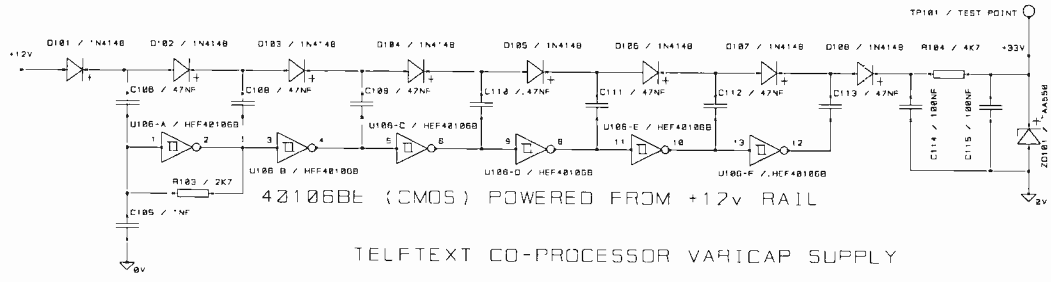


Fig. 5. Voltage multiplier used to supply the synthesized tuner's varicap. The free-running oscillator on the left feeds the multiplier chain, producing around 50V on the right. Following current limiting, a shunt zener produces a stable 30V varicap feed.

play on-screen. Headers are the top row of the display, the line with the date in it, while a magazine is the jargon for page hundreds. Empty page buffers are thus available for assembling the next transmission of the required pages, row by row.

Control of the co-processor is achieved over the serial link, using a well-established protocol. Operator page requests are converted into a format that the firmware understands by the terminal software. This protocol is available to programmers.

Addition of a tuner required the command set to be extended to select Channels E21 to E69, which was fairly easy to implement. Less easy was modifying the terminal software which in a commercial application had been produced without regard for such extensions.

In order to leave working software well alone, I chose to use a terminate-stay-resident (tsr) program to supplement the keyboard commands. The required channel, 0 to 9, is simply typed in as, say, Alt-C2 for BBC2, and so on.

Correspondence between E channel numbers and single-digit selection is achieved by extending the default configuration file to include local tv channel numbers. To invoke teletext operation, a batch file is run on the host PC which first loads the tsr then runs the terminal software. The tsr picks up the local channel numbers from the configuration file and uses them to send to the PC communication port as required. On quitting the program, the tsr is unloaded.

Commissioning

There is no substitute for a thorough visual inspection of all the components before powering-on. Time spent here is worthwhile. Further testing depends on what test equipment you have access to.

A PC card extender for example allows easy access to a powered board for measurements. Clearly the first thing to check is the power supply and look for activity on the crystals and oscillator module. Thereafter you should set up the communication so as not to clash with existing devices. I used COM3.

Next, connect a good quality tv aerial and run the batch program supplied, having first edited the default.cnf file to include local channels and the chosen PC communication port and IRQ settings. Type Alt-C1, to select BBC1, and 100, for page 100. You should be rewarded with page 100 on screen with headers rolling past. If not, then there is a fault.

The external audio and video outputs are a good starting point although the software will need to have selected a channel for this to work. You can look for activity on the I^2C bus when a channel is selected. A command from the PC causes LED_1 to flash. Off-air syncs cause LED_2 to flash, and data to the PC causes LED_3 to flash. ■

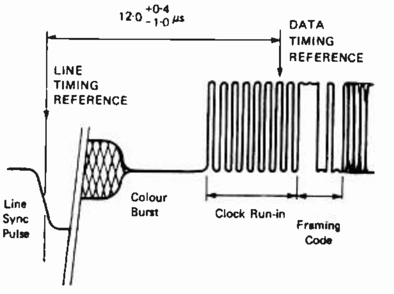


Fig. 8. Signalling the start of teletext data is a clock synchronisation run in comprising ones and zeroes. Immediately following is framing information that ensures that whatever is decoding the data starts reading it at the first bit of the first byte.

Reading

Figures 6, 7 and 8 are taken from the September 1976 *Broadcast Teletext Specification*, ISBN 0 563 17261 4.

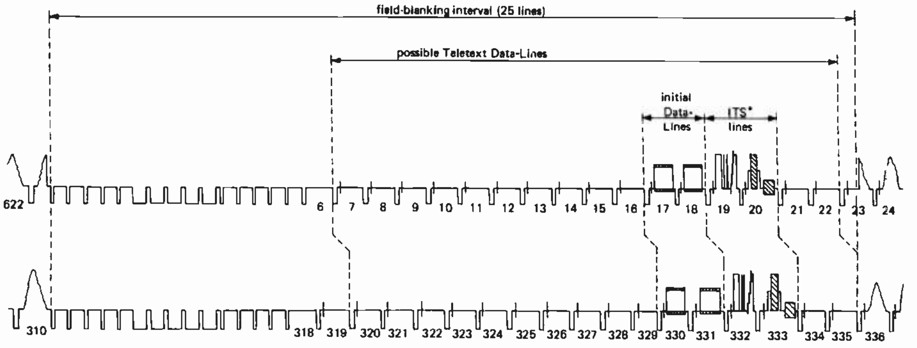


Fig. 6. Broadcast teletext data streams appear in up to 16 lines per tv frame, all within the blanking period. Data is transmitted at 6.9375Mbit/s.

Component availability

A double-sided circuit board, with gold plated edge connectors, together with the required eprom and terminal program, are available from Citifax, *EW+WW*, 9 Goose Cote Hill , Bolton BL7 9UQ, United Kingdom. Tel./fax +44-0 204 307293.

The package includes the teletext co-processor pcb, a full parts list, terminal program and programmed eprom for £65 inclusive. Overseas add £5 postage.

Please note that although this offer is made in good faith, EW+WW has no involvement in it and therefore cannot take responsibility for any problems that may arise relating to transactions between Citifax and readers.

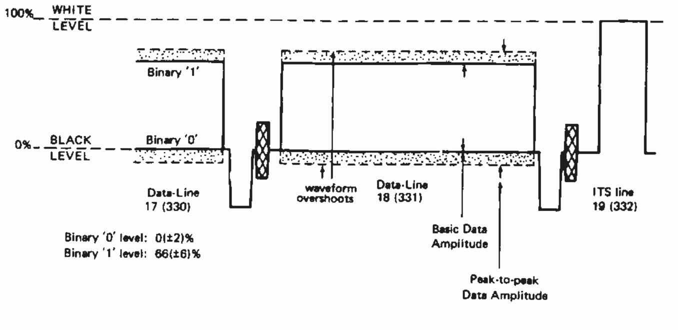
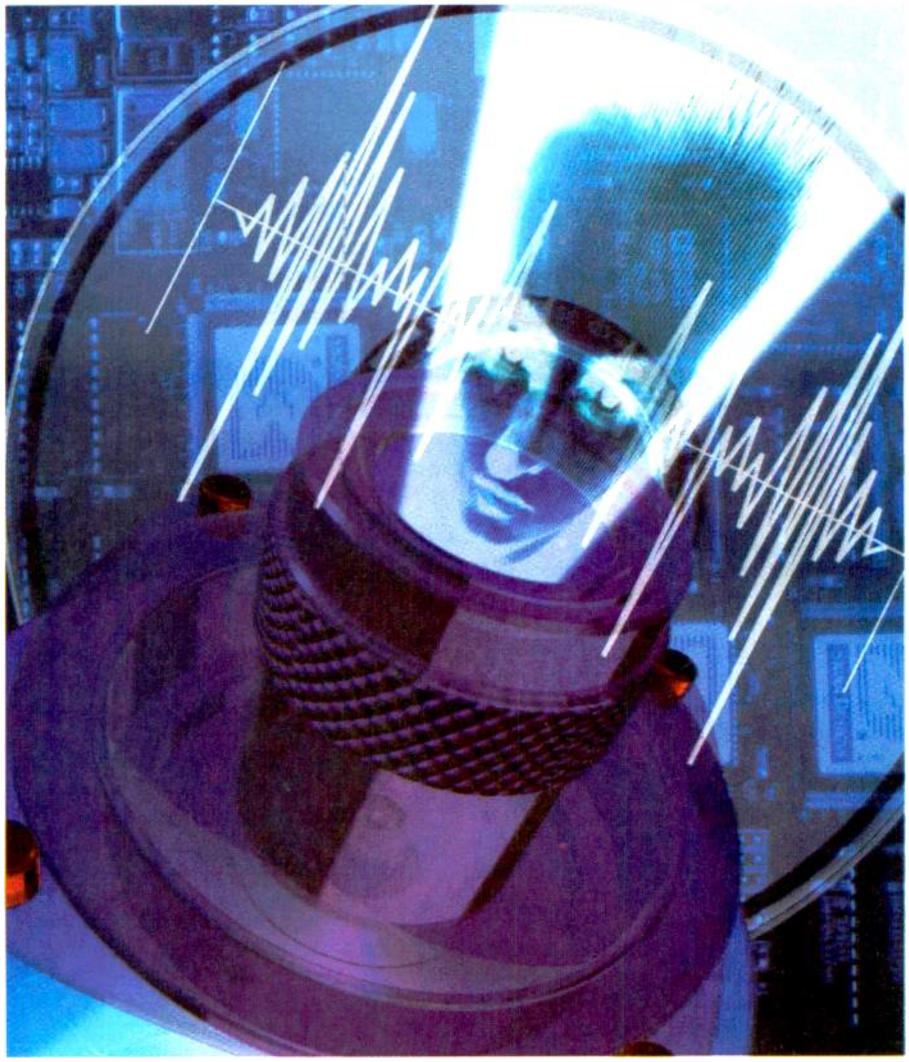


Fig. 7. In the teletext lines of a broadcast tv frame, a logic 0 corresponds to black level while 66% white represents a logic one.



VIDEO compact disc

One source predicts that video CD sales will exceed 16 million by 1998. With films on CD about to hit the shops, George Cole looks at events that led to current video-compression standards.

This autumn, Panasonic is launching a home audio system with a difference. Called *SC-VC10*, this system will include a new type of compact-disc machine with the ability to play conventional audio *and* video CDs – which store over an hour of digital full-motion video, or fmv.

Video CD is a cross-platform system whose software can be played on domestic players, such as Compact Disc Interactive equipment, computer games consoles such as Commodore's *Amiga CD32*, and desktop PCs. Consumer electronics companies hope video CD will boost sales of compact-disc players, while computer companies think that the ability to play full-length movies on your PC or games machine will prove appealing.

Video CD is not the first CD-based format to offer pictures or even video. The compact-disc Digital Audio Red Book standard, set by Sony and Philips in the late 1970s, **Fig. 1**, makes provision for a CD plus graphics, CD+G, system.

Around 97% of the data stored on an audio CD is for music and error correction. Sub-codes containing control information take up the remaining 3%. These sub-codes are called P, Q, R, S, T, U, V and W. The first two, P and Q, are used for features such as track indentification and track time, while R-W are set aside for graphics.

There is space for up to 1500 graphic elements, each with a display size of 288 by 192 pixels. The graphics look like a teletext picture and can display up to 16 colours from a palette of 4096. It takes around 2.5-10 seconds to draw each display, which can also be scrolled off the screen.

Ordinary CD decks ignore sub-codes R to W, but players equipped with a CD+G decoder can display them on any television screen. It was thought that many record companies would make use of the graphics to display song lyrics and artist information, but few bothered.

JVC launched CD+G decoders in the US, Germany and Japan, where they were used for karaoke titles. But few CDs featured graphics. WEA launched several CD+G discs, some of which can still be found in music shops. A few multimedia players however, among them CD-i decks, also include CD+G decoders.

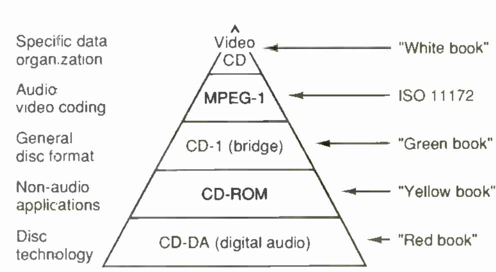


Fig. 1. Video CD technology pyramid.

In 1985, Sony and Philips launched the cd-rom, or Yellow Book standard, which enables a CD to store over 600M-byte of data. This data may be in the form of sound, text or pictures. Although a cd-rom can store over 250,000 pages of text, it can only hold a small amount of uncompressed digital video.

A video frame uses around 800K-byte of data and tv systems like PAL display 25 frames a second to create the illusion of motion. Simple mathematics shows that a CD could only store less than a minute of digital video. Another problem is that CD transfer rate is only 1.5M-bit/s, which is far too slow for displaying 25 frames a second.

In 1987, Philips announced Compact Disc Video, CDV, which stored analogue video and 16-bit pcm audio on the same disc. A 12cm CDV disc stored six minutes of video and sound, plus a further 20 minutes of audio. Philips hoped that CDV would appeal to music buyers, but the format flopped.

Practical CD video – almost

During the same year, researchers from RCA's David Sarnoff Laboratories in New Jersey astounded the world by displaying a CD which held over an hour of digital full-motion video. Called DVI, for Digital Video Interactive, the system which made it possible used powerful algorithms to reduce a 800K-byte frame to just 5K-byte. DVI works by comparing each frame and only coding the differences. It also analyses a frame for repetitive or redundant information. This might be large areas of the same colour, such as sea or sky.

DVI was a fine technical achievement, but the complex number crunching needed meant that it took a super computer over a day to compress an hour of video. What is more, playback DVI boards designed for desktop PCs initially cost thousands of dollars.

Moving pictures through the window

In 1988, the CD-i Green Book standard was set. Originally, the CD-i specification included motion video, which played video running at around 15 frames per second in a small window on the screen. At the time Philips said that it was not practical to include full-motion video.

A development in May of 1988 forced Philips to change its mind. Jointly, the International Standards Organisation and International Electrotechnical Commission set up the Moving Picture Experts Group, MPEG, to establish a world standard for digital video. The group included representatives from computer, consumer electronics and telecommunications companies.

In 1989, Philips announced that CD-i would include digital video based on the MPEG standard and hoped to launch the first CD-i players with full-motion video capability. But the first MPEG standard, MPEG-1, was not set until November 1992 – a year after CD-i was launched in North America. As a result, the first decks had space for a plug-in full-motion video cartridge to hold the MPEG chipset.

In January 1993, Nimbus Technology and Engineering demonstrated a system capable of storing over an hour of MPEG-1 video on a CD and offered it to music, film and video companies; there is more on this in the inset panel.

Born from karaoke

On 29 June 1993 however, JVC, Philips, Sony and Matsushita – owner of Panasonic, Technics and JVC – announced the video CD format. It was covered by a new standard known as the White Book. Video CD is based on another standard known as Karaoke CD, which is a sub-set of CD-i. Karaoke CD was developed by Philips and JVC.

Video CD is designed for linear video programmes, such as music videos, and stores up to 74 minutes of MPEG-1 video and audio on a CD. There are two specifications for PAL and NTSC systems, shown in the table, although all video CDs will play on all video CD decks.

Initially, the announcement stated that video CDs would play on a variety of player formats. These included CD-i decks with an add-on full-motion-video cartridge, dedicated video CD players, PCs with a MPEG card and cd-rom drive and modified audio CD decks with an add-on video CD decoder.

News that modified audio CD decks could be used caused confusion. It suggested that owners of existing CD decks could upgrade to video CD with an add-on decoder. But this is not the case.

Video CD uses a CD-ROM-XA 'bridge disc', which includes a header telling the player that it contains computer data. If you try playing an XA disc in most music CD decks, the audio output is muted to protect the speakers from damaging white noise. Philips has since confirmed that it will not market add-on decoder boxes.

In August 1993, the basic specifications for video CD were released, with news of two optional features. One of these is for display-

The other video CD

At the 1993 Midem music festival, Monmouth-based Nimbus Technology and Engineering created a stir by showing two compact discs holding over an hour of MPEG-1 video.

Playable on many conventional CD decks, the first of these discs held 79 minutes of video. The second held 135 minutes of full-motion video. Nimbus said that this too could be played on ordinary CD decks, albeit with a small adjustment to the laser system.

Both long playing CDs were made possible by a new mastering system developed by NTE and called video CD. The company planned to launch add-on MPEG-1 decoders priced at around £100. These would include a *CL-450* MPEG-1 chip set designed by C-Cube Microsystems and link directly into the digital output socket of a CD player.

NTE said that roughly one third of the world's 120 million CD decks had a digital output connection and that such a socket was becoming a standard feature on most new machines. By plugging the decoder into their tv set, users could watch video movies on their audio CD player.

Although the development caused excitement worldwide, Philips stemmed the fervour by claiming that Nimbus's long-playing discs broke the CD standard and would not play on all audio decks.

Nimbus decided to press ahead with its 79-minute disc – which was within the Red Book standard. The company offered an MPEG-1 compression service to film and video companies, charging $100 per minute of video material. It also sold a couple of its new mastering systems to China.

But the announcement of the 'new' video CD shocked NTE. Athough it claimed that many Hollywood film companies preferred its system, NTE's alternative became significantly less attractive. Earlier this year, NTE was still demonstrating its system to Hollywood, but there is no sign of any software company backing it.

Video CD software specifications

Parameter	Video system NTSC	PAL
Coding Method	MPEG-1	MPEG-1
Resolution (pixel)	352h x 240v	352h x 288v
Frame rate (Hz)	29.97	25
Pel aspect ratio	1.0950	0.9157
Bit rate (bit/sec)	1151.929k	1151.929k
	Audio	
Coding method	MPEG - 1	LAYER-II
Sampling rate (Hz)	44.1k	
Bit rate (bit/sec)	224k	
Emphasis	ON or OFF	
Mode	Dual channel or stereo	

ing still pictures, which can be in standard resolution, at 353 by 288 pixels, or high resolution with 704 by 480 pixels.

It is also possible to put control codes on video CDs for branching interactive programmes used for educational or training programmes and the like. In January 1994, the first video CD standard, known as version 1.1, was set.

Better than VHS

Video CD offers 'better than VHS' picture quality and titles can also include widescreen pictures and Dolby Surround sound. A number of electronics companies have shown video CD players, including Panasonic, Sony, Goldstar, Samsung and Fisher. These machines will run video and audio CDs, but not CD-i or cd-rom titles. They will also offer video-recorder type features such as still frame and picture search.

Home multimedia formats CD-i and 3DO also support video CD and Commodore has launched a £200 add-on decoder for its Amiga *CD32 system*. Atari is promising the same for its *Jaguar* games system, while US company ReelMagic has launched a MPEG-1 video card for PCs, priced £400. Video CD software supporters include Polygram, BMG, Warner, Paramount and MGM/UA.

But video CD has not been without its problems. One of these is the confusion caused by Philips, which has launched a number of movies on CD. The films are part of a deal with Paramount, announced in autumn 1993, which involves the launch of 50 Paramount movies on CD-i over the next two years. However, the first movies to appear were not video CD discs, but dedicated CD-i titles.

Philips says the Paramount deal was struck before the video CD announcement and there was no time to convert its video encoders. Philips has since switched its mastering system over to White Book encoding. This means that there are now two types of CD-i movies: those which only play on CD-i decks and those which play on CD-i and video CD machines.

Another problem has been caused by the difference between the CD-i and video CD standards. The Green Book standard defines a picture 384 pixels wide, compared with 352 pixels for the White Book. The result is that when a video CD is played in a CD-i deck, there are thin black bands at the sides of the picture.

Initial video CD titles also looked slightly distorted in CD-i players because of the difference between the 15MHz clock speed of CD-i and video CD, which was set at 13.5MHz. Philips has since adapted the CD-i full-motion video cartridge which now automatically changes the clock speed, depending on the type of disc being played.

Commercial constraints

Other problems have been commercial. Different frame rates, line numbers and colour encoding systems used for PAL and NTSC television formats mean that VHS tapes and laser discs bought in the US will not play in most PAL machines.

Film companies use these differences to negotiate separate deals in PAL and NTSC territories and stagger release dates. Most movies appear on video in the US months

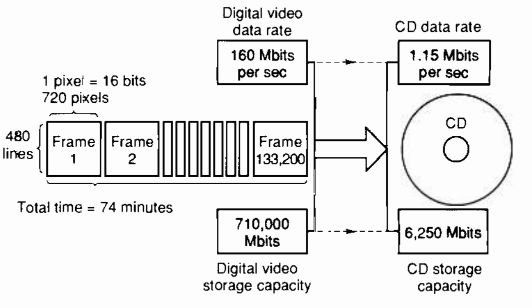


Fig. 2. The video encoding challenge – storing a 710G-bit film on a half-gigabit CD.

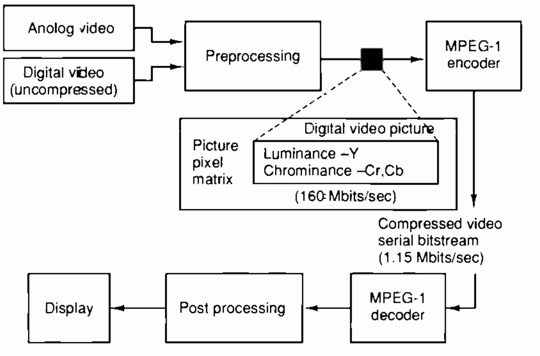


Fig. 3. Video encoding/decoding scheme for the MPEG-1 standard.

before they reach Europe. But video CD breaks this barrier and discs can be bought and used anywhere.

The only caveat to this is that discs designed for NTSC display a letterbox picture when played in a PAL machine. This is because PAL pictures have 288 lines, compared with 240 lines for NTSC; in PAL, these lines are repeated sequentially to produce a full display. As a result of the line difference, PAL pictures look stretched on an NTSC video CD player.

In practice, the letterbox effect is not too obtrusive and most people find it acceptable – especially if it means they can watch a movie long before its UK release date.

Film companies have insisted on separate mastering for PAL and NTSC titles. They want their stars to 'always look their best'. Earlier this year, Paramount began inserting blocking codes into its titles to stop NTSC discs being played in PAL machines. A compromise has now been reached. European viewers playing an NTSC video CD see a warning at the start of the disc stating that the title is designed for NTSC.

Panasonic's first UK video CD system, the *SC-VC10*, will cost around £800. Most movies on video CD will be on two or three discs, although carousel players, which store several or more discs are promised. Also promised are portable video CD decks with liquid-crystal video screens. Next year, Philips is expected to announce a new CD standard offering longer playing time and better picture quality.

MPEG video compression

There are several MPEG standards under development. MPEG-1, now known as ISO 11172, is optimised for storage media, such as CD. The idea was to develop a digital video system whose data rate was below the CD's 1.5M-bit/s ceiling.

The data rate of MPEG-1 video is approximately 1.2M-bit/s, with 224K-bit/s allocated for the audio. MPEG-1 video is also being used for asymmetrical digital subscriber loop, ADSL, systems used by telephone video-on-demand services. This enables full motion video to be sent down narrow-band copper telephone wires into homes. MPEG-1 picture quality is claimed to be better than VHS, and in some cases can approach Super VHS.

The MPEG-2 standard is designed for data rates of 2-15M-bit/s and aimed at digital broadcast services, including HDTV. A new generation of cd-roms will also use MPEG-2 to offer even better picture quality. MPEG-3 was originally designed for HDTV, but its work is now part of MPEG-2. MPEG-4 will be aimed at low bit-rate transmission systems, such as video phones, where the data rate is in tens of kilobits/sec.

Note that MPEG is concerned with the organisation and synchronisation of the output digital bit stream. It does not define the encoding algorithms used to compress the audio and video. This leaves companies free to develop their own algorithms and opens the way for improved compression. It also explains why some MPEG-1 pictures look much better than others.

Additionally, MPEG does not define what scrambling and encryption systems companies may wish to use, so there are likely to be many different conditional access systems on the market.

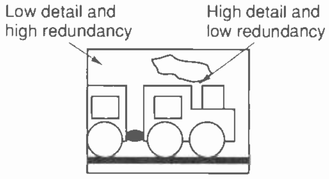


Fig. 4. In video compression, redundancy in individual pictures is exploited.

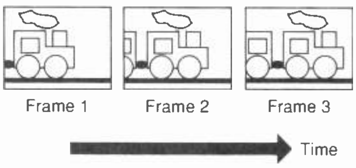


Fig. 5. There is also redundancy in sequences of pictures that can be removed to save memory.

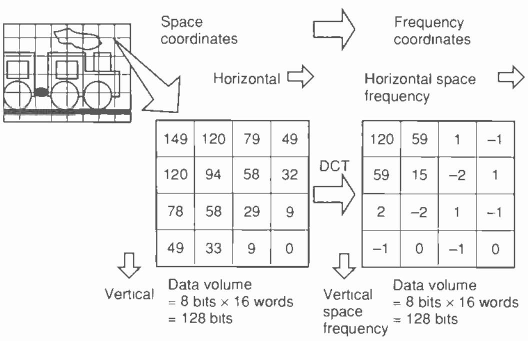


Fig. 6. The process used to sort out redundant data is known as discrete cosine transform, or dct.

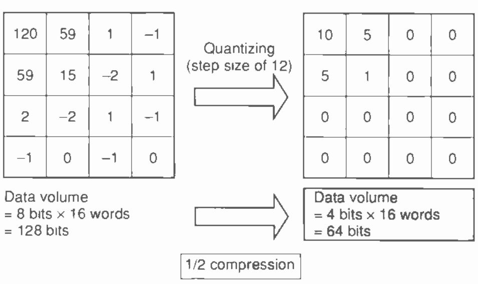


Fig. 7. Here, quantizing the dct components involves scaling down by a factor of 12.

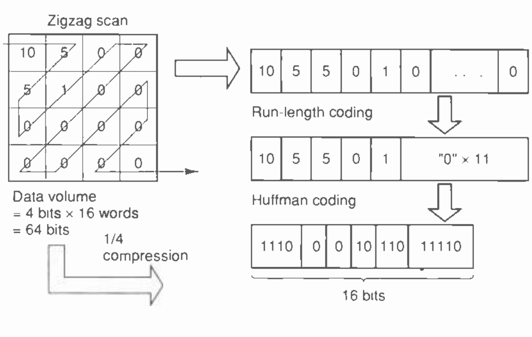


Fig. 8. Variable-length coding involves scanning the pattern in a zig-zag to produce a bit stream.

Panasonic predicts that by 1998, world sales of video CD players could be well over 16 million

Video companies find video CD attractive because it costs around $1 to press a CD – compared with $3 for duplicating a tape. Video CDs will cost around £16 for a single disc and £20 for a double-disc set. Panasonic says consumers will like video CD because the discs are non-wearing and offer fast access.

Panasonic predicts that total worldwide sales of video CD players will reach 300,000 this year, rising to 16.4 million by 1998. However, the company does not see video CD replacing the domestic video recorder, which offers longer playing times, masses of cheap software – and recordability. Indeed, Panasonic is a supporter of a new home digital vcr format called dvr. However, the company says that video CD should make audio-only CD players obsolete by the end of the decade.

MPEG technology breakdown

Storing over an hour of digital video on a CD is difficult, as **Fig. 2** implies.

Figure 3 shows the general encoding/decoding scheme. The process starts with a high-quality analogue or video source which is pre-processed to obtain a Y/C video signal. This is fed to the encoder where it is compressed then multiplexed with the audio stream.

Also added to the data stream are graphics, such as logos and copyright messages, and control codes for chapter selection, etc. In the player, the MPEG-1 decoder decompresses the audio and video. Post processing is used to produce a display output.

Currently, MPEG-1 encoders cost between £50,000 and £100,000, but prices are expected to fall rapidly. Encoding costs around £90 per minute, so a 90-minute movie can be mastered for around £8000 – which is relatively cheap. The video CD mastering process is shown in **Fig. 15**.

MPEG exploits redundancy within a frame and between each frame, **Figs 4, 5**. There are large areas of repetitive detail in the locomotive image. In addition, the short time interval between each frame of $^1/_{25}$s means that there is little change between successive frames.

The process used to sort out the redundant data is known as discrete cosine transform, dct. First, an eight-by-eight segment of pixels, called a block, is analysed and each pixel is assigned a number which represents its luminance value, **Fig. 6**. Discrete cosine transform is used to produce a set of numbers representing the spatial frequency. This is the speed at which the luminance value changes between adjacent pixels. The new set of numbers are known as output coefficients.

Within the dct matrix, the top left number, in this case 120, is most important. It represents the average luminance value for the area being analysed. The 120 looks high for the average value because in this example the values have been scaled down by a factor of 8 rather than 16, for simplicity's sake.

The small values of 1 and –1 indicate that there are no fast changes in brightness from pixel to pixel. Note that at this stage we have still not reduced the amount of data. It is still 128 bits, comprising 8 bits by 16 words.

During compression, the first stage involves the quantization of the dct values, **Fig. 7**. In this example, the numbers are scaled down by a factor of 12. This means we now need four, rather than eight, bits to code each word, halving the amount of data.

Variable length coding is used in the next stage, **Fig. 8**. Here, the output coefficients are scanned in a zig-zag fashion to produce a string of bits. Zig-zagging is used because the numbers furthest away from the average value are more likely to have the lowest luminance value. Visually, this makes them least important.

A long string of zeros is usually produced by this process. The next step is to use Huffman coding. The shortest length code is assigned to the most frequent data value and the longest length code for the least frequent data value.

A long string of zeros can also be represented by a small number of bits. This reduces the data by a factor of four. These processes are also applied to the colour signal data.

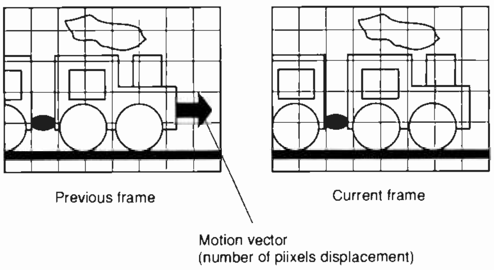


Fig. 9. Further compression is possible by analysing motion over successive frames.

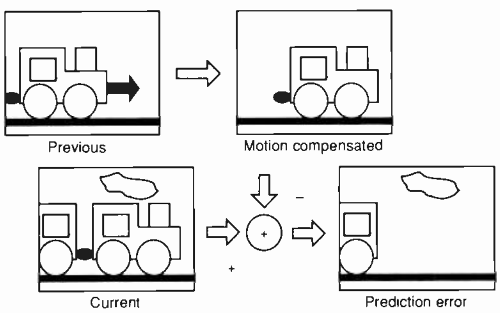


Fig. 10. The image of a moving train need only be stored once, together with a motion vector.

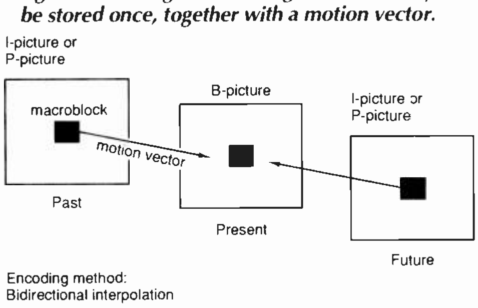


Fig. 11. Encoding a bidirectional picture involves interpolating from a non-compensated I reference frame and a predictve P frame.

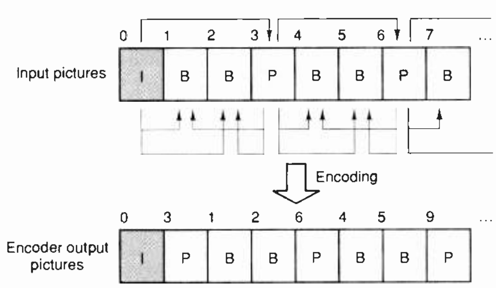


Fig. 12. Video CD picture organisation. Since B frames take longest to encode, output frames are shuffled by the encoder.

Fig. 13. Video CD decoding elements. I frames are sent via the top route while B and P frames travel the lower path.

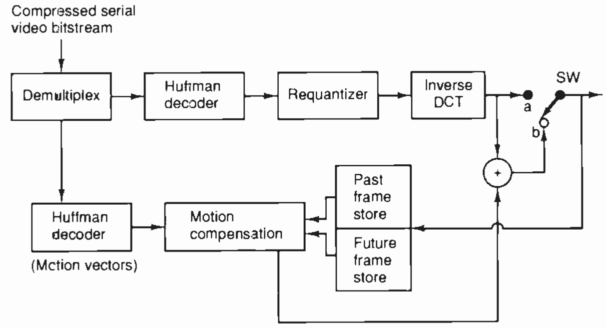
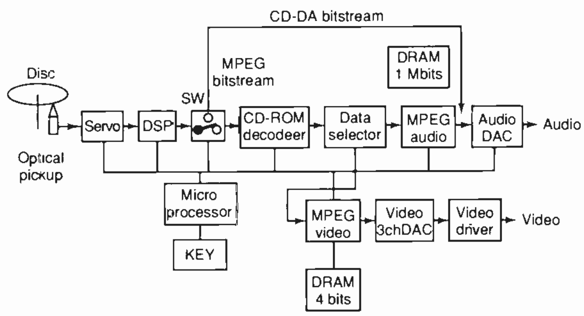
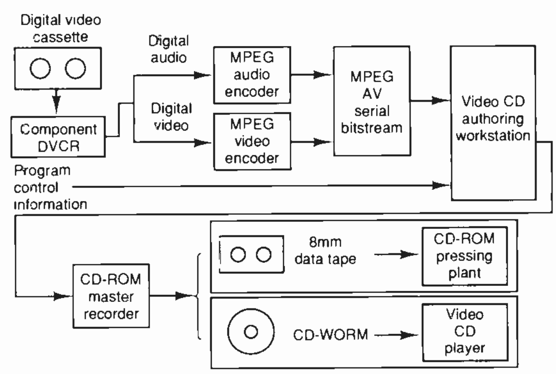


Fig 14. In the player, a switch routes data depending on whether the disc is audio or video.



Video CD mastering combines encoding and decoding technologies.



Another compression process, namely motion detection and compensation, analyses differences between frames. **Figure 9** shows a train moving across a screen. If detail of the engine has already been encoded in the previous frame the only further information needed is the number of pixels moved in the current frame. This is known as the displacement value or motion vector, and requires far less coding.

Motion vectors are produced by analysing the frames before and after the current frame. This works by storing the frames in memory and calculating the displacement values.

The current frame may also contain new information – part of the first carriage in this case – which is known as the prediction error, **Fig. 10**. This is coded using dct and added to the motion compensated frame.

There are three types of frame involved in the compression process, **Fig. 11**. The I frame is the reference frame and no motion compensation is applied to this. The P or predictive frames are produced by motion compensated process. Bi-directional or B frames are produced by an interpolation method which uses the motion vectors from an I and P frame or two P frames.

This process combines blocks of data known as macroblocks from both the I and P frames to create an average value. Macroblocks comprise four luminance and two colour blocks.

Figure 12 shows how the frames are organised by the encoder. The B frames produce the largest data compression, but take the longest to encode. For this reason, output frames are shuffled by the encoder to increase speed.

Video CD uses an MPEG audio format known as Layer-II. This has a bit rate of 224k-bit/s. It gives stereo or dual channel sound and represents a compression ratio of 6.3 when compared with CD audio.

MPEG uses a coding system similar to the Precision Adaptive Sub-band Coding, PASC, process used by the digital compact cassette. Audio is sampled at 44.1kHz, split into 32 frequency sub-bands, and analysed. Frequencies which are hidden or masked are discarded to reduce the amount of data.

Elements of the the video-CD decoder are shown in **Fig. 13**. I frames are sent via the top route, while B and P frames follow the bottom route. A switch detects each type of frame.

Figure 14 is a block diagram of a video CD player. The switch here detects whether the machine is playing an audio or video CD. ■

DESIGNING FOR NOISE IMMUNITY

In passing from its source to a data acquisition system, a low-level transducer signal inevitably picks up noise. There are ways of recovering signals from noise, as described in last month's issue, but here, Dave Robinson looks at a better alternative – building in noise immunity.

Under certain circumstances, it is possible to recover at least an approximation of data that has become corrupted by noise. However recovery techniques should be only used as a last resort.

Lord Rutherford once advised, "If your experiment needs statistics – think of a better one." In this instance the 'better experiment' relies on preventing the noise from corrupting the data in the first place.

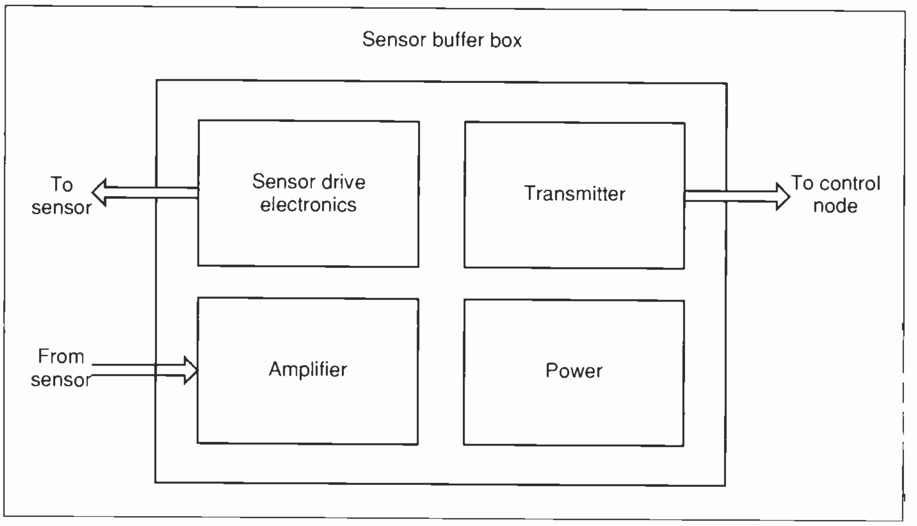
Sensor buffering

In most data acquisition systems, the most vulnerable element will be the sensor itself. Sensors often need some form of excitation, or drive. Noise at this node is often transferred to the output, possibly as a multiplicative distortion.

Every inch of extra cable – shielded or not – between your sensor and its processing electronics is a further reduction in signal integrity. One solution is to insert a sensor interface buffer as close to the sensor as physically possible. This interface should produce any sensor drive needed, amplify sensor output and convert the information into a robust transmittable form.

Fortunately modern electronics make the design of sensor buffers straightforward. They can be very compact, and can be designed to consume very little power, **Fig 1**.

Fig. 1. To ensure that a low-level signal from a remote sensor can be received at a distance with integrity, an excellent solution is to introduce a buffer subsystem as close to the sensor as possible.



Power for sensor buffers

Powering these buffers can be a problem; there are three methods that generally work well:

- battery power
- remote power and dc/dc converters
- local mains with isolated communication links.

Battery power offers many advantages. Each buffer unit can be totally isolated. Noise generation mechanisms such as earth loops are much easier to handle.

Using cmos devices, a small battery with regulation electronics can provide months of trouble free running without maintenance. This is, however also the major disadvantage. Batteries do expire, and at fairly unpredictable times. Two questions need answering–does the buffer enclosure facilitate battery changes and will the system function if one sensor becomes inoperative? Rechargeable batteries, perhaps charged from solar cells, can solve many problems.

If power has to be transferred from the central controller some distance from the sensor buffer, considerable noise can be picked up on the power lines. As zero volts often acts as the signal reference point, noise picked up on this line can prove troublesome.

One excellent solution is to use the transmitted power as input to an isolated output dc-to-dc converter within the sensor buffer unit, **Fig 2**. In this way, the isolated zero volts of the converter output and system zero volts need not be brought together until the signal is safely back into your central control system.

An alternative, ingenious and very popular method of powering remote sensors is the so called 4-20mA current loop. This forms a complete two wire power distribution and communication system. Power is provided on the loop by the control unit. The sensor circuitry at the far end essentially modulates the current flowing in the loop, according to the parameter being measured.

Since data is communicated by switching current in the loop between 4 and 20mA, there is always at least 4mA available on the loop for powering the transducer circuitry. This system works extremely well; it is fairly immune to noise pickup and is supported by most of the major manufacturers in the field.

Powering the buffer and the central control node from their own local mains supplies – which may be a fair distance apart – is problematical, **Fig 3**.

In my experience, this configuration has been the cause of a great deal of heartache – mainly caused by the varying earth potentials. Connecting the signal earths of the two systems results in what can be quite large and noisy currents flowing through the earth wire.

The easiest safe method of powering the remote nodes from their local mains is to use completely isolated data transfer between the nodes and the control node. Options here include, telemetry, optical-fibre links and opto-isolator with copper cabling.

Sensor drive signals

Where appropriate, drive signals depend on the type of sensor you are using.

A thermocouple for example, does not need any drive. Platinum resistance thermometers normally require a dc supply, as do a strain gauge bridges.

With dc-excited sensors similar to prts and strain gauges, it is worth considering the advantages to be gained from using an ac excitation signal instead. Frequency-selective amplifiers, lock in amplifiers, phase sensitive detectors and so called box-car integrators are electronic amplifiers designed to produce a dc value. Amplitude of this value is directly proportional to input signal whose frequency and phase are equal to the reference clock provided to the amplifier, **Fig 4**.

Suitable ac-to-dc converters can easily be made from switched-input amplifiers. Exciting the sensor with a known frequency modulates the sensor output signal with the same frequency. Any change in signal level anywhere else in the spectrum is entirely due to noise, and can therefore be ignored – a process that the 'lock-in' amplifier does superbly.

The lock in amplifier is driven from the same oscillator as is driving the sensor. It automatically follows any frequency drift in the excitation generator, allowing a quite crude oscillator to be used. This is better than using a straightforward filter to isolate the signal, as this would require a much more stable frequency standard. Note that the speed of your oscillator should be substantially faster than the maximum rate of change that you are expecting from your signal.

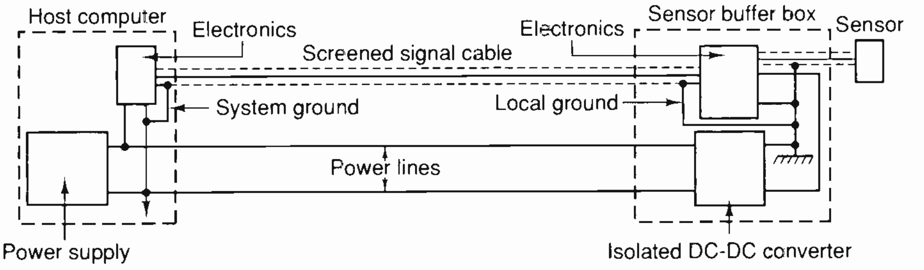


Fig. 2. Powering a sensor buffer from a host can cause grounding problems. One solution is to regenerate clean power supplies using isolated dc-to-dc converters within the sensor buffer unit.

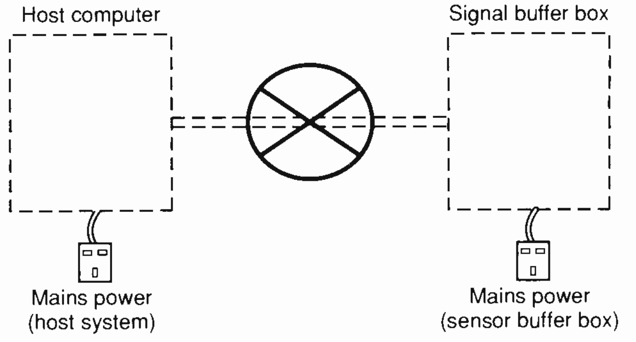


Fig. 3. When the host and remote signal buffer have independent mains-derived power supplies, the two are connected via their earths. If the two are also linked by a signal earth ground loops will inevitably cause problems.

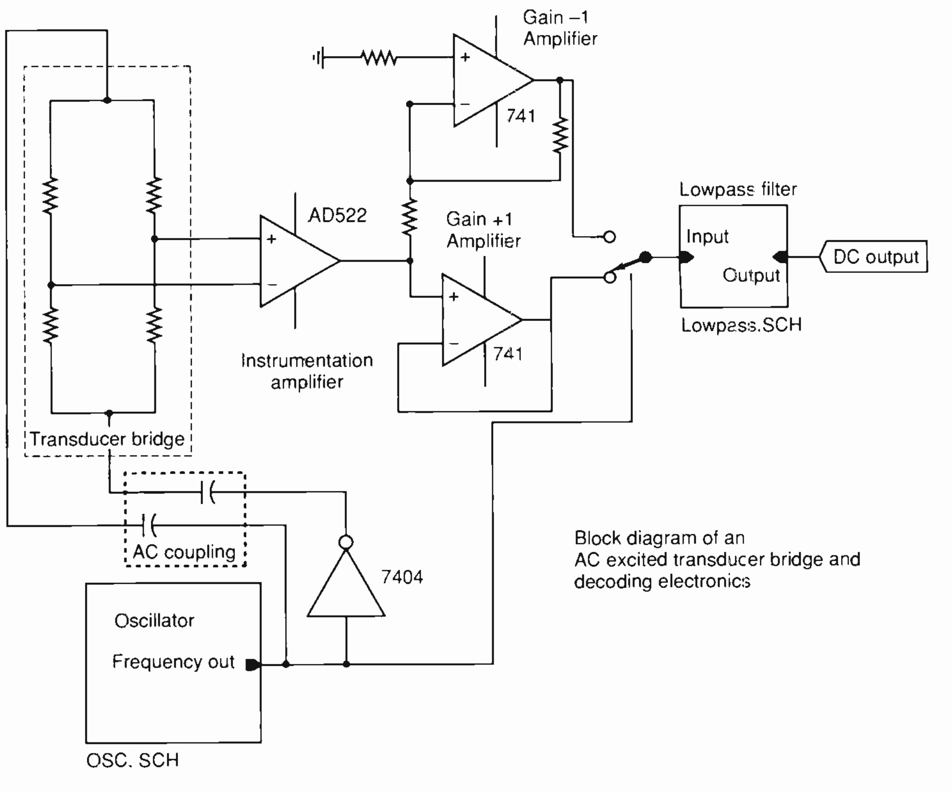


Fig. 4. With certain types of transducer, noise problems can be reduced by exciting the sensor with an ac signal. At the receiving end, a frequency-selective amplifier driven by the exciter clock makes sure that unwanted signals are rejected.

Data protocols

After amplifying and decoding sensor output, the buffer unit has to convert the signal into a format suitable for transfer to the central control node – a crucial task.

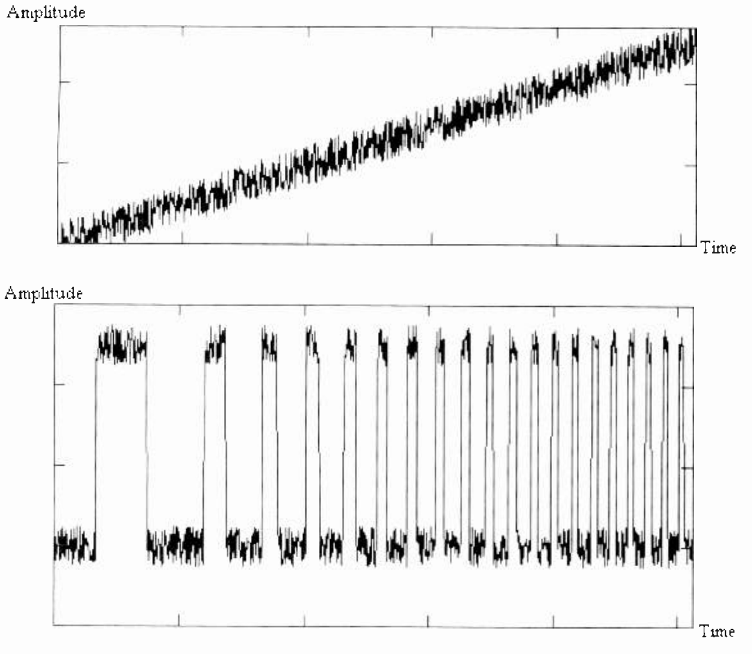


Fig. 5. In a purely analogue system, the ramp's detail is obscured by noise. Converting the clean ramp to frequency-modulated format reduces the effects of noise.

Transducer types

Transducer is a general name for electronic components that interact with the real world. They fall into two key categories, namely input and output transducers.

Output transducers take an electrical signal and convert it to some form of non-electrical output. Examples are loudspeakers, which convert an electric signal into an acoustic one, and solenoids, which convert between electricity and movement. Output transducers are usually called actuators.

Input transducers – the main subject of this article – take some form of non electrical input and produce an electrically measurable parameter change. Input transducers are called sensors. Sensors have been designed to use changes in virtual every conceivable electrical parameter. Although far from exhaustive, the following list outlines some of the more common types.

Change in charge:
Pyroelectric infrared detectors
Microphones

Change in current:
Geiger Müller detector
Photodiode
Photomultiplier
Smoke detectors

Change in voltage:
thermocouple
photodiode

Change in resistance:
strain gauges
thermistors
platinum resistance thermometers
light dependent resistor
position measuring potentiometer

Change in inductance:
position LVDT
metal detector

Change in capacitance:
accelerometers
microphones

Sensors fall into two categories. One is direct output sensors, which produce a measurable electrical effect representing the parameter being measured. A thermocouple is a simple example of this type of sensor. It produces an output voltage which is a direct function of the temperature difference between the so called hot and cold junctions.

The other category of sensor requires some form of drive or stimulus which is modified or modulated by the parameter being measured. Resistive sensors usually fall into this category. In almost all cases, the signals provided by the sensors are very low level, and need to be preprocessed prior to being digitised for digital processing.

What are the criteria? Obviously the transmission should be more immune to interference and noise pickup than simply transmitting the raw sensor data. The first option is to send the amplified analogue output down a wire. Assuming that noise picked up has a fixed amplitude, then there must be a value of gain for which the signal-to-noise ratio becomes acceptable.

Screening the cable helps to diminish noise pickup, as does using a differential format. Differential encoding produces two signals. Difference between the two signals is proportional to the amplified sensor signal. Provided that the conductors follow similar paths to the control node, then any pickup should be almost identical on both cables.

Feeding the resulting differential signals into a differential amplifier results in recovery of the original signal. Noise on the other hand, being common mode, is rejected. Again screening this double cable is a sensible precaution.

The multiplex disadvantage

Now all that remains is to interface the data into a data-acquisition board. This is true where one sensor is concerned, but consider what is needed if data is being acquired from more than one source.

In certain cases, a data-acquisition card with input multiplexer may be suitable. This is especially so if the data has been transferred using a low noise protocol such as current-loop technology.

Consider the following situation however, involving 16 remote sensors. These are interrogated at the same sample rate. This means that during any complete multiplexer cycle, each of the sensors is only being looked at for a sixteenth of the total time.

In other words, 93.75% of the data from the sensor is wasted. If some way of using this lost data could be found then the square-root law shows that signal-to-noise ratio could be increased substantially. From this argument you can see that the major drawback in the use of a simple signal multiplexer is that it succeeds in decreasing our potential signal-to-noise ratio by a factor proportional to the square root of the number of channels on the multiplexer. For the above simple example, the factor is four. So what is the solution? Until recently the only answer would have been to use frequency modulation.

Frequency modulation

There is much to commend frequency modulation. To implement it, each sensor buffer is fitted with a device known as a voltage-to-frequency converter. These are very accurate voltage-controlled oscillators. Their output amplitude remains fixed. Frequency of the output signal on the other hand is directly proportional to input voltage – in this case the amplified sensor output. Once frequency modulated, the signal is transmitted directly to the acquisition system.

Being essentially digital, the fm signal is easily transmitted via an optical fibre, ensuring no noise pickup and complete isolation between the buffer and control node. Even if the cheaper option of screened electrical cable is opted for, isolation can still be accomplished using a standard opto-isolator.

Noise immunity provided by this technique comes from two directions; the first can be seen from **Fig 5**. In the top diagram, the ramp illustrates what an amplified and noise-contaminated analogue signal might look like on arriving at the control node. Although general shape of the ramp is clear, any fine detail is lost in the noise.

A fragment of a frequency-modulated waveform being transmitted down exactly the same transmission cable is shown in the second diagram. Its statistical noise corruption is identical. Removing the noise from the first waveform is not a trivial operation, relying on the techniques covered in last month's issue. Removing it from the second however is very simple.

The second advantage of fm stems from the way that the frequency encoded information is demodulated. This can be implemented via several devices, among them frequency-to-voltage converters and digital frequency integrators.

Frequency-to-voltage converters are complementary to voltage-to-frequency converters. In essence the voltage-to-converter, transmission line and frequency-to-voltage converter are transparent. Any change in voltage at the input produces a proportional change in voltage at the output. Noise corruption affects the amplitude of the signal. Because the information travelling down the transmission line is encoded as a frequency change, the signal output is virtually identical to the signal input.

Another efficient way of reconstructing the data is via a digital frequency integrator. It has the advantage that it removes the need for any analogue to digital converters at the control node end. Incoming frequency modulation is simply allowed to increment a counter for a fixed period of time. The faster the frequency, the higher the count.

At the end of the integration period, the count reached is directly proportional to the average voltage at the input of the voltage-to-frequency converter. If this integration period is made equal to the sample period, the accumulated count is a highly suitable format for applying averaging to the incoming data. Because the technique is immune from noise pickup on the transmission line, it helps to remove the noise already corrupting your sensor reading before it reaches the sensor buffer. As the resulting count is already in digital form, its interface to the control computer becomes trivial.

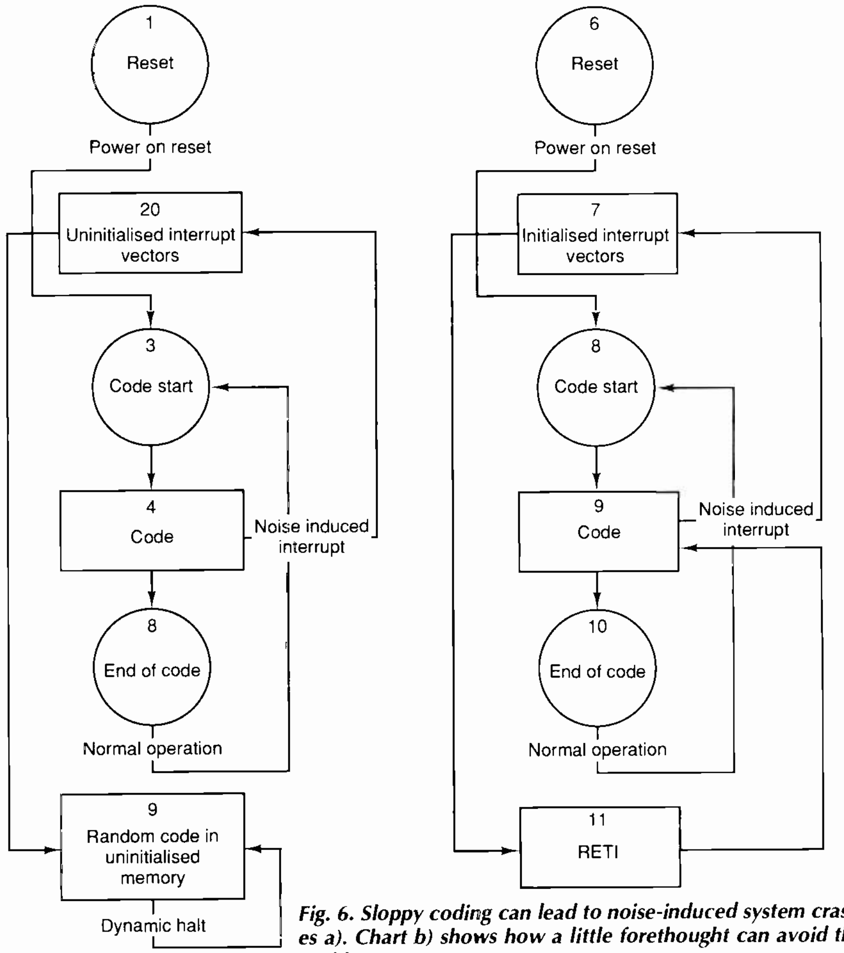


Fig. 6. Sloppy coding can lead to noise-induced system crashes a). Chart b) shows how a little forethought can avoid this problem.

Intelligent buffers

Powerful yet low power microcontrollers and microprocessors have considerably changed the nature of data acquisition. It is now possible to build in intelligence directly into the sensor buffer unit. This opens up options like applying sophisticated preprocessing to your sensor data locally. Preprocessing is application dependent but might well include removing noise from sensor input, sensor linearisation and temperature compensation.

Inserting intelligence at this position expands your options but it can also introduce pitfalls. To avoid them, make sure that any local power supplies are clean. If you have to receive the power from a remote source, then regenerate the rails via isolated dc-to-dc converters and use copious decoupling. Also use separate power planes to distribute power, as opposed to simple tracking.

On the logic-signal side, make sure that any unused interrupt lines and dma requests are strapped in their inactive states using pull up resistors. Do not leave them floating. Ensure that interrupt handling routines are included for all unused interrupt lines. These should comprise at least a return from interrupt instruction, even though you have tied the input line inactive, **Fig 6**. Problems caused by noise-induced interrupts to a non-existent interrupt handler can be very difficult to detect.

Make sure that the reset line is filtered against transients. Include hardware so that the processor can perform a cold reset to all distributed elements of the system under software control. This is essential for restarting a multi-sensor system synchronously.

Include hardware and software to perform a watchdog function. When running normally, the software outputs periodic pulses that are detected by hardware. If the software runs out of control after a noise induced excursion, the watchdog pulses stop and the hardware causes a reset, bringing the system back under control.

Fill all unused program memory with no-operation instructions. This avoids the possibility of having an accidental 'dynamic halt' or similar instruction due to random bit patterns left in the prom.

When writing your serial communications software, do not rely on serial information getting through even at RS232 levels; expect your messages to become corrupted. Include some form of error checking, and handshake between the sensor buffer box and control node. Also consider using a synchronous protocol. These often automatically handle cyclic redundancy checking, etc. Both Motorola and Hitachi produce cost-effective micros with built in support for this. ■

New maps for Smith's world

The Smith chart is the rf engineer's best friend. Ian White reviews Z-Match for Windows, the new Smith-chart program from Number One Systems.

The Smith chart is a visual way of designing and understanding impedance transformations in networks and transmission lines. Designed for pencil and paper it remains a powerful design tool; whatever you can draw on a paper chart, you can also draw on a computer screen.

The programming challenge is to keep all the paper chart's good features and then to add facilities unique to the computerised version. Although previous versions of Number One Systems' *Z-Match* software for dos came close to those objectives, *Z-Match for Windows* has made it.

On the screen

The main display is divided into two areas, the Smith chart on the right and the control and information panel on the left (**Fig. 1**). You draw on the chart in the usual Windows fashion using the mouse. Buttons on the left-hand panel control the main functions, so you seldom need to pull down the menus from the top of the screen. The non-linear scales that used to be the bane of the paper chart are replaced by precise numerical displays, linked to the position of the mouse cursor on the display.

The Smith chart has always been essentially a visual design aid (see the sidebar for an explanation of how the chart works). You draw lines and curves corresponding to your design ideas, and watch what happens to the circuit impedances and other parameters. Quite frequently you need to back-track to an earlier point and reconsider what to do next or even throw away the existing chart and start again with a clean sheet. The software makes good use of the Windows environment to provide these facilities and more.

A cross-shaped flashing cursor represents the current operating point for dragging and dropping around the display using the mouse. Cursor movement can be either free or constrained to follow an arc of constant R, X or VSWR away from its starting point, something you could not do on a paper chart without first drawing the relevant arc. Whenever you move the cursor, numerical displays of nine parameters are very rapidly updated once again, a facility unique to a computerised Smith chart.

Alternatively you can type in an explicit impedance and the cursor will go there. If you want to leave a permanent mark at any cursor position, simply double-click the mouse button.

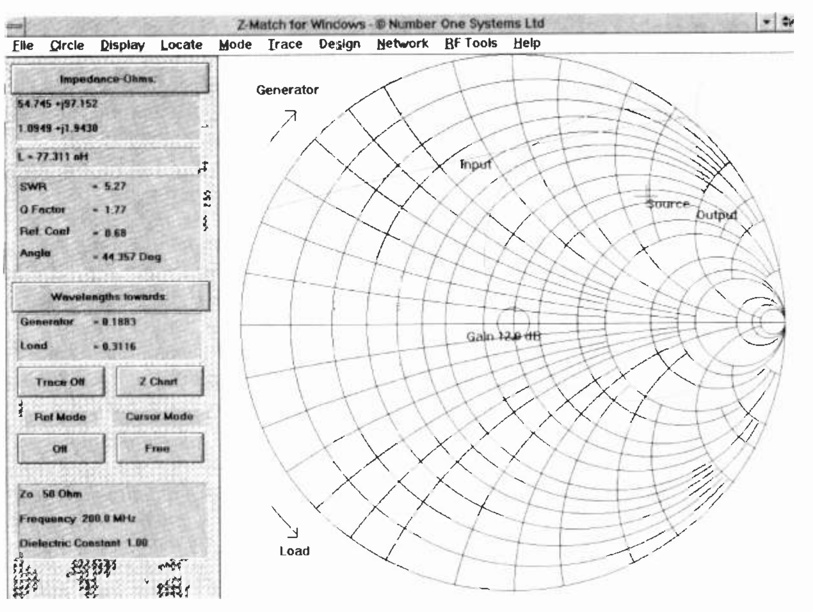


Fig. 1. Z-Match for Windows displays its power with a complicated Smith chart, showing stability and gain circles. Note the cross-shaped cursor over the point marked Source: the panel at the left of the screen displays a range of information about the cursor location, including its absolute and normalized impedance. Quick-change buttons control the most used features of the chart.

For back-tracking and erasing mistakes, *Z-Match for Windows* has an undo function which successively removes objects you had drawn on the chart and restores the cursor to its previous position.

The final requirements to replace the basic paper Smith chart are somewhere to file your work for later retrieval, a waste-basket and an endless supply of clean charts. Windows' familiar File menu provides these functions using Save, Save As, Open and New and of course you can always send a chart to the Windows system printer. Together, all these features provide a good simulation of the familiar paper-and-pencil environment.

Many other functions are accessible through the menu system. For example, several of the numerical displays make use of the system reference frequency (in MHz), the reference impedance Z_0 and the velocity factor of the transmission line, any of which can be changed using the Display - Parameters sub-menu. In addition, buttons in the left-hand panel provide shortcuts to the most useful charting functions.

The software makes two of the most difficult Smith chart operations simple. You can convert with a single click between impedance and admittance charts (see sidebar), allowing the addition of admittances in parallel just as easily as impedances in series. The whole chart flips, leaving the cursor exactly where it was on the screen while maintaining your trace of previous movements. Two clicks draws a unit conductance circle on an impedance chart, or a unit-resistance circle on an admittance chart.

To understand what it was like without these aids, imagine navigating using two road-maps, one showing only major roads and the other showing only minor roads. You could manage it with practice, but you wouldn't ever want to! These features of the software banish one of the most confusing aspects of the paper Smith chart.

The other difficult routine operation is impedance renormalization which occurs on moving from one system impedance to another. The software offers two ways of doing this. Either you can type in a new system reference impedance Z_0 (and optionally have the cursor position recalculated) or you can specify any cursor location as the new reference impedance; all subsequent cursor movements and displays are referenced to this location on the chart. It is easy to alternate between two different reference impedances.

Advanced functions

A regular use of the Smith chart is to design impedance matching networks, starting from some arbitrary impedance and usually ending at the reference impedance Z_0 in the centre of the chart. The technique is to work your way in towards the centre by adding reactances in series or in parallel.

Reactance in series is added by moving along a constant-resistance line on the impedance chart, while reactance in parallel is added by moving along a constant-conductance line on the admittance chart. In many practical cases the quickest route home is via a point further out towards the periphery of the chart. All this is straightforward in *Z-Match for Windows*, but the software also has a special network synthesis function which draws the circuit diagram of the network you have designed.

Smith chart refresher

Any impedance can be represented as $R \pm jX$ where R is the resistive or "real" component and X is the reactive component, positive for inductance and negative for capacitance. Although you can plot R and X on conventional rectangular coordinates, it can be difficult to represent the very low and very high values that occur in transmission-line calculations.

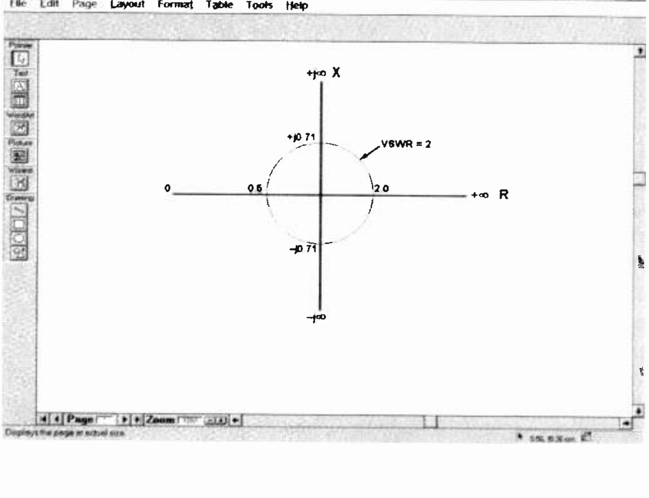
The problem of very low and high impedances is solved by two techniques: normalization and logarithmic axes. When working in a system that has a defined standard impedance Z_0, for example a transmission-line impedance of 50Ω, it is useful to normalize all impedances, so that R becomes R/Z_0 and X becomes X/Z_0. Thus a normalized impedance of 2+j1.5 represents 100+j75Ω in a 50Ω system.

Logarithmic resistance and reactance scales are possible because real-life rf resistances and reactances do not truly reach either zero or infinity.

Fig. A shows a logarithmic resistance-reactance plot using rectangular axes, with a single "zero-to-infinity" scale for resistance and two mirror-image scales for positive and negative reactances. The system reference impedance Z_0 falls in the centre of the normalized plot, at 1+j0.

What happens if you plot all impedances corresponding to a voltage standing wave ratio vswr of say 2.0 on this diagram? The two relevant points on the normalized resistance axis are 2.0+j0 and 0.5+j0, and these are equidistant from the origin because the scale is logarithmic. The two corresponding points on the normalized reactance axis are 1+j0.71 and 1–j0.71. If we scale the axes appropriately, these four points can be made to lie on a circle. Moving around this constant vswr circle, one complete revolution will pass through all possible combinations of resistance and reactance that produce a vswr of 2, equivalent to having travelled a half-wave along a transmission line.

Unfortunately the angular scale is not linear: 1+j0.71 represents a phase angle of 35° rather than 90° as it appears on Fig. A. The genius of Phillip Smith was to realise what happens when the straight-line reactance axes of Fig. A are bent around to form two half-circles whose "infinity" ends join together at the right of the chart. Figure A has become transformed to **Fig. B**, a Smith chart. Remarkably, the vswr circle remains a circle, but its phase-angle scale



A) One way to represent all practicable impedances R ± jX on two rectangular axes. All combinations of resistance and reactance that have a normalized voltage standing wave ratio of 2.0 will fall on the constant-VSWR circle. To transform this diagram into a Smith chart Figure B, bend the two reactance axes round so that all three "infinity" points meet at the right-hand end of the resistance axis.

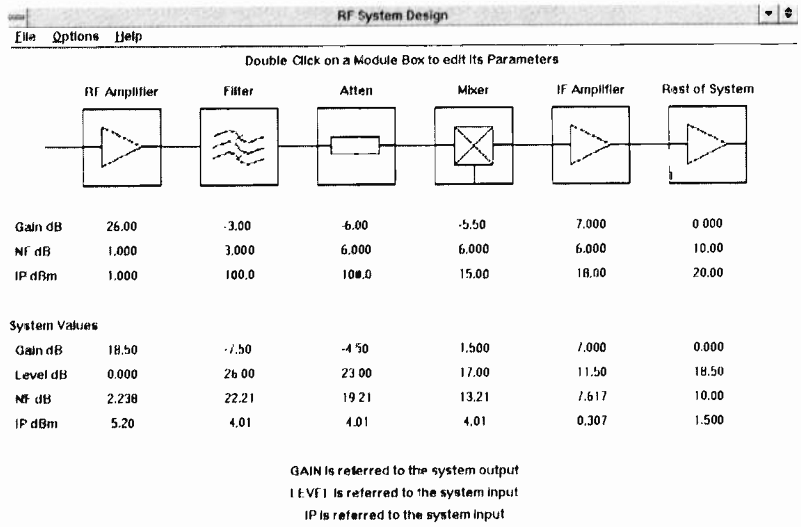
The resulting netlist can be saved in a file format compatible with Number One Systems' *Analyser* nodal analysis software. Similarly you can read in results files from *Analyser* and have *Z-Match* plot them in Smith chart form.

S-parameters are widely used in rf design, and the review software also handles them. Device data may be entered either manually or from files (the package contains the Motorola device library in industry-standard S2P format). It will interpolate frequency dependent file data to the desired frequency, and gives an immediate indication whether the device is stable at a selected frequency. It can then go on to calculate stability circles, source and load impedances, etc.

In Figure 1 the program has calculated and plotted the input and output stability circles (the stable area is towards the middle of the chart), a constant-gain circle for a user selected value of 12dB, and a suitable source impedance. You can then go on to match this impedance using the network synthesis function described earlier.

Off the map

The information panel beside the Smith chart can display its data in several alternative forms, avoiding the need for routine interconversions. In addition there is an RF Tools menu offering interconversions between power and voltage ratios, dB and dBm; and between VSWR, reflection coefficient and return loss. The same menu also offers reactance and resonance calculations, and a designer for simple transmission-line transformers to match any arbitrary impedance to Z_0.



	RF Amplifier	Filter	Atten	Mixer	IF Amplifier	Rest of System
Gain dB	26.00	-3.00	-6.00	-5.50	7.000	0 000
NF dB	1.000	3.000	6.000	6.000	6.000	10.00
IP dBm	1.000	100.0	100.0	15.00	18.00	20.00
System Values						
Gain dB	18.50	-7.50	-4 50	1.500	7.000	0.000
Level dB	0.000	26 00	23 00	17.00	11.50	18.50
NF dB	2.238	22.21	19 21	13.21	7.617	10.00
IP dBm	5.20	4.01	4.01	4.01	0.307	1.500

Fig. 2. The receiver noise figure and intermodulation calculator from the rf Tools menu. The top three rows of data refer to the individual modules in the block diagram drawn above. The bottom four rows show how the cascaded stages interact to produce the overall system performance. Double-click on any module box to alter its data, and the whole display is automatically recalculated.

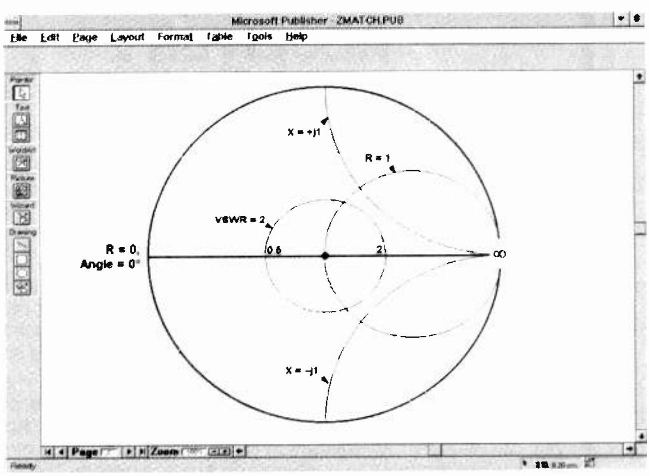
has now become linear. What is more, the entire Smith chart can be constructed from one horizontal straight line and a series of simple arcs and circles.

The horizontal diameter still represents pure resistance, ranging from zero to infinity with R=1 at the centre of the chart. Lines of constant resistance are circles, crossing the horizontal diameter at normalized values of R=0.5, 1, 2 and so on. R is zero all around the periphery of the chart, except at the "infinity" point where all the constant-R circles meet tangentially. Positive reactances lie above the horizontal diameter, negative reactances below. Lines of constant reactance are arcs of circles, mirror-imaged above and below the diameter and all converging on the "infinity" point. Figure B shows all these characteristics.

Thus you can plot any practicable impedance on a Smith chart, and watch how it changes as you travel on a constant-vswr circle along the transmission line. Anticlockwise is forwards towards the load, clockwise is backwards towards the generator, and a full revolution of the chart represents 180° along the line, in other words a half wavelength.

What happens if you add an inductance in series with an impedance already plotted? You move along a constant-resistance circle into the region of higher inductive reactance. How much has the phase angle changed as a result? Draw straight lines out from the centre of the circle through your origin and destination points, and measure the angle between them not forgetting that every 1° on the chart represents 0.5° along the transmission line.

Once you know how to read the map, there are yet more wonders in Smith's World. Thanks to its logarithmic scaling, the Smith chart works equally well as an admittance 1/Z chart, which opens the way to dealing with impedances added in parallel. Changes of characteristic impedance are handled very easily: you can either renormalize all the plotted impedances; or else take advantage of the fact that a constant-vswr circle can be centred anywhere on the chart and will still remain a true circle. This latter property opens yet further paths to the analysis of amplifier stability and noise figure, both of which can be represented using constant-value circles.



B) A basic Smith chart, showing the same constant-VSWR circle as Figure A. The horizontal resistance axis is still the same, and all other components consist of simple arcs and circles. To convert this impedance chart into an admittance chart, mirror-image it from right to left. The Smith chart is the rf design engineer's best friend. Once you know your way around, it provides a convenient and simple way of solving all manner of problems of impedance transformation, amplifier design and stability analysis. Here's how it works.

Although receiver noise and intermodulation calculations are not directly related to Smith charts, it is useful to have these facilities to hand. **Fig. 2** shows how the concept is implemented in *Z-Match for Windows*. Double-click on the box representing a module in the block diagram and a window opens, into which you can enter data relevant to that module. On closing the window, the software recomputes the noise and intermodulation performance of the whole system. Noise performance can be displayed as either noise figure (dB) or noise temperature (K). This Windows "applet" has its own File menu with facilities to save and retrieve designs, and to print out the block diagrams and results.

The package

The software comes with a thorough and well-written 125-page manual, and installation from the single program disk is quick and straightforward. The Motorola device library comes on a second disk, and need not be permanently installed.

Even if you start out knowing very little about Smith charts, the manual gives an extensive explanation of the concepts followed by worked examples to introduce you to the software implementation. Lack of experience may even be an advantage, because experienced Smith chart users may need to unlearn some cherished techniques that were actually geared to the inadequacies of the paper chart.

In spite of a price which may deter amateurs, *Z-Match for Windows* could easily make itself indispensable to any professional rf design engineer. ■

SYSTEM REQUIREMENTS

286, 386 or 486-based PC (at least 16MHz 386SX preferred)
Coprocessor is used if available
Graphics display and monitor (at least colour VGA preferred)
Mouse
Windows 3.0 or later, under DOS 3.0 or later
3.5in 1.44MB floppy disk drive
1MB hard disk space (program will run from floppy disk)

PRICE

Price £245.00 + VAT (upgrade from Z-Match for dos £50.00 + VAT)

SUPPLIER DETAILS

Number One Systems Ltd, Harding Way, Somersham Road, St Ives, Cambridgeshire PE17 4WR. Telephone 0480 461778, fax 0480 494042.

Please mention *Electronics World + Wireless World* when seeking further information.

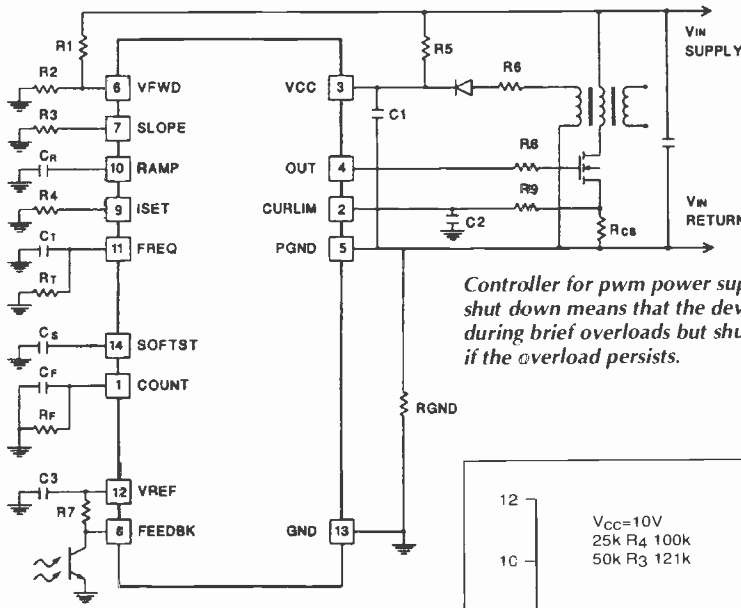
Low power PWM with programmable shutdown

Combined features of the *UCC1570* family of pulse width modulators are low power and high-speed operation. Typical start-up current is 85μA, running current is 1mA, and the device can drive a power mosfet at up to 500kHz. Peak gate drive current is 1A.

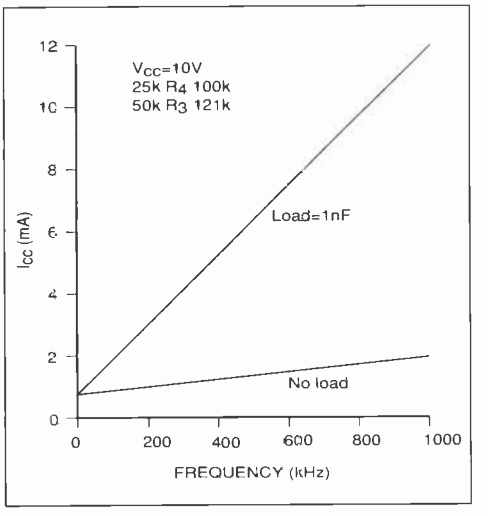
As this circuit diagram from the *1570* family preliminary data sheet shows, the device is intended for use in switching supplies. It is optimised for isolated designs with primary side control and a voltage mode feed-back loop.

Made in bicmos technology, the three versions in the range have a low start up current requirement. This allows efficient off-line starting with a bootstrapped low-voltage supply. Voltage feedforward provides fast and accurate response to wide line voltage variation without the noise sensitivity associated with current-mode control.

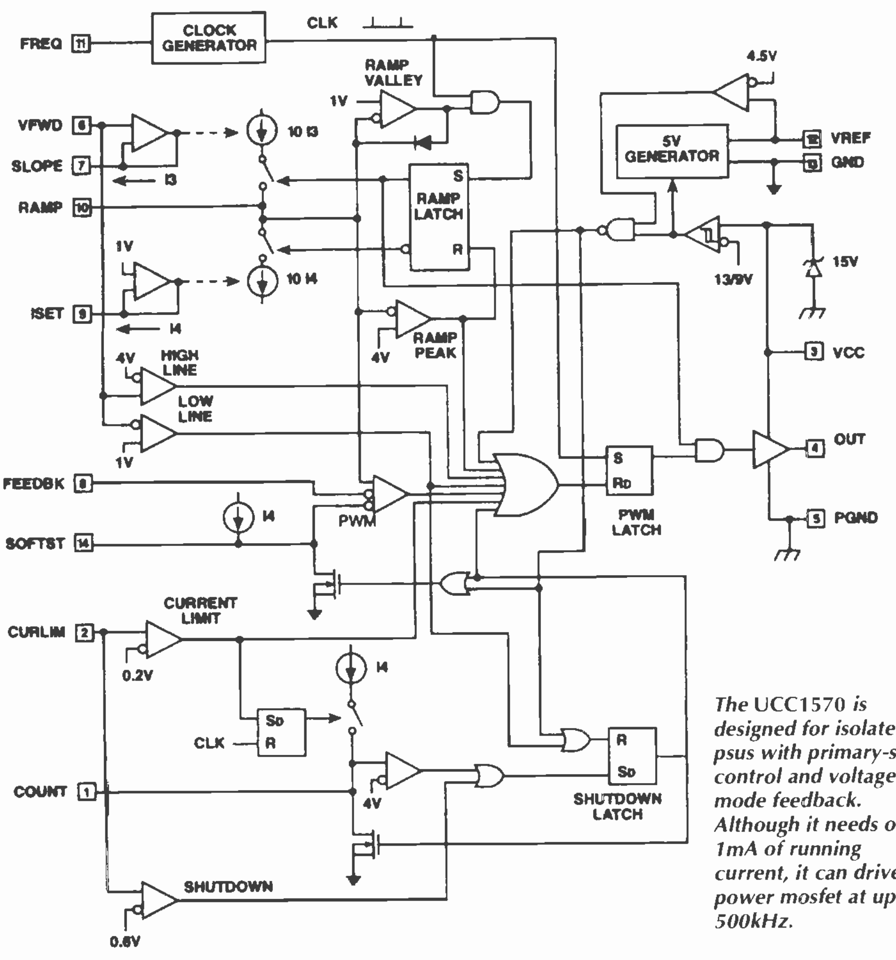
Fast current limiting is included with the ability to latch off after a programmable



Controller for pwm power supplies. Fault-counting shut down means that the device stays operational during brief overloads but shuts down completely if the overload persists.



Current versus frequency for the UCC1570 pwm controller while in operating mode shows that little current is needed to operate the device – even at high switching speeds and with high mosfet gate capacitance.



The UCC1570 is designed for isolated psus with primary-side control and voltage-mode feedback. Although it needs only 1mA of running current, it can drive a power mosfet at up to 500kHz.

number of repetitive faults has occurred. This enables the power supply to withstand a temporary overload, while still shutting down in the event of a permanent fault.

A built-in maximum duty-cycle clamp is programmable between 20% to 80%. Line voltage sensing is enhanced by the provision of a programmable window of allowable operation. Further facilities offered by the devices are an optocoupler interface and fault latch-off or automatic-restart options.

Notes on how to calculate the components shown are included in the data sheet.

Unitrode UK Ltd, *6 Creswell Park, Blackheath, London SE3 9RD. Tel. 081 318 1431, fax 081 318 2549.*

Video selection for broadcast industry

Signal routeing and buffering circuits for video feature in Maxim's *Analog Design Guide 3/8*.

In the oscillograph, input versus output is shown for a 180MHz video buffer with guaranteed 0.99V/V gain over its temperature range of -40 to 85°C. This buffer, the *MAX405*, has a differential phase and gain errors of 0.01° and 0.03%. In addition, it holds its gain with loads as low as 50Ω. Suitable for driving NTSC, PAL or SECAM, the device operates from ±5V supplies and has a slew rate of 650V/μs.

For the eight-by-four crosspoint switch shown, Maxim claims the highest integration and precision in the industry.

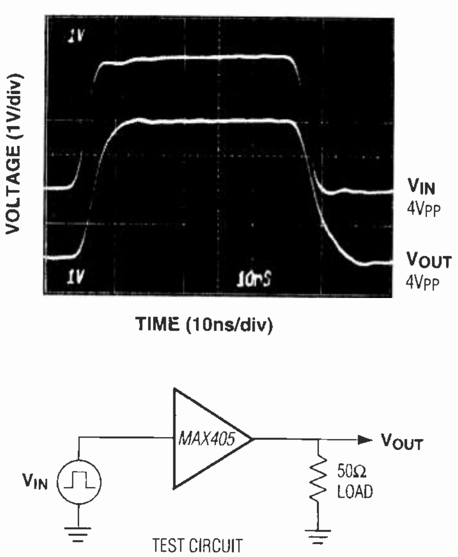
Operating at 100MHz and handling 'broadcast-quality' video, the *MAX458* and *459* switches incorporate a digital control interface and output buffers. In the *458*, these four buffers are unity gain while in the *459* they have a gain of 2. This extra gain is for driving 150Ω back-terminated cable directly, without needing feedback resistors. Phase error is 0.06° while differential gain error is 0.03%.

One of eight video channels is selectable via the *MAX440*. It has a 160MHz response, 250V/μs slew rate and can switch a pixel in 15ns. Differential gain and phase errors are 0.03° and 0.04% respectively. There are also four and two-channel versions of this switch.

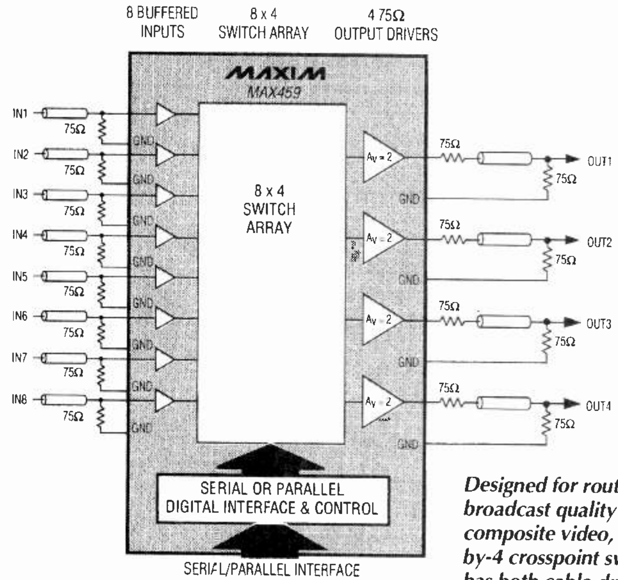
The final circuit is a four-pole, two-way switch for selecting rgb video and sync from two sources. Output swing, provided by buffers with a gain of 2, is ±2V with a 50Ω load and the device operates up to 100MHz.

Other video circuits in the note cover triple and quad 100MHz buffers with no external feedback and a 250MHz no-feedback video amplifier with a slew rate of 850V/μs.

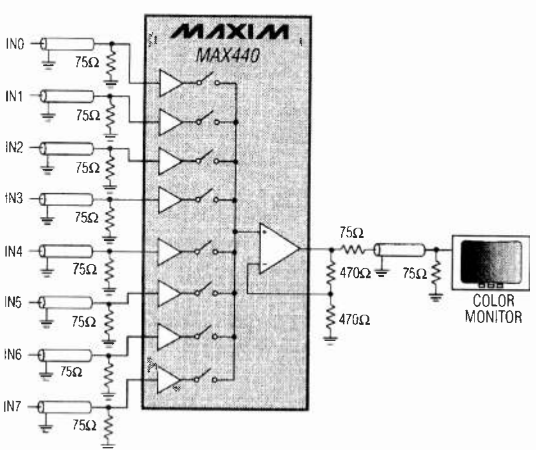
***Maxim Integrated Products**, 21C Horseshoe Park, Pangbourne, Reading RG8 7JW. Tel. 0734 845255, fax 0734 843863.*



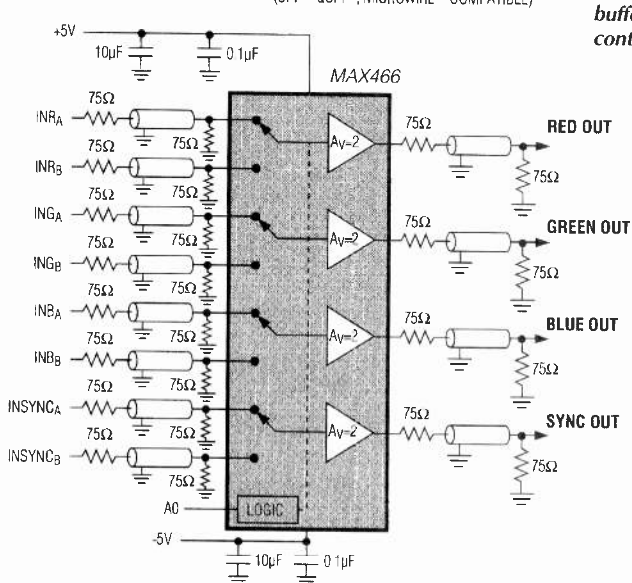
With a differential phase error of 0.01° and a gain error of 0.03%, the MAX405 buffers video up to 180MHz and provides 0.99V/V gain over its full temperature range.



Designed for routeing broadcast quality composite video, this 8-by-4 crosspoint switch has both cable-drive buffers and digital control interface.



Useful in applications such as selecting surveillance camera sources, the MAX440 eight-way switch has pin-selectable frequency compensation and operates to 160MHz.



The MAX466, intended for switching between two RGB-plus-sync sources, has 100MHz 6dB buffer amplifiers capable of driving ±2V into a 75Ω load.

Integrated frontend devices for DECT

Three ICs designed to simplify and miniaturise DECT-compatible equipment have been introduced by Motorola.

These application diagrams are from device data sheets for the new devices, namely the *MRFIC1801, 1803* and *1804*. The *1801* is a 1.8GHz antenna switch, the *1803* an up-converter, and the *1804* a combined LNA and down-mixer.

Also suitable for the Japan Personal Handy Phone standard, JPHP, and other personal communication systems, all three devices are produced in surface-mount packaging. They operate from 85° down to –30°C, rendering them suitable for all-weather use.

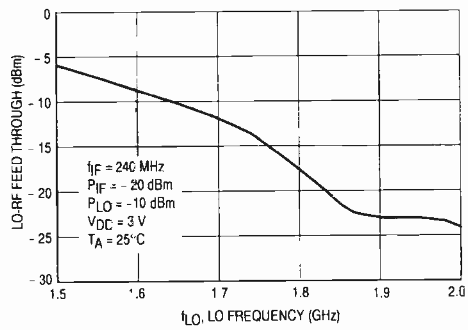
With the exception of the antenna switch, which can run from a rail up to 5.5V, operation is from a single 2.7-3.3V supply. Combined consumption of both converters is typically just over 100mW, while current needed for the antenna switch is 300μA in transmit mode, or 50μA when receiving.

Transmission insertion loss of the *1801* antenna switch is typically 0.75dB, while the 1dB compression point is quoted as 29dBm. Capable of operating at frequencies from 1.5 to 2.5GHz, the switch is activated via a simple on/off logic input and exhibits 22dB transmit to receive isolation.

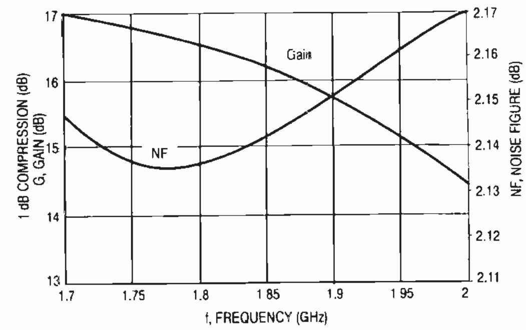
Incorporated in the *1803* up-converter are an active up-mixer, exciter amplifier and local-oscillator buffer amplifier. The device provides 10dB intermediate-to-radio-frequency conversion gain and has a usable frequency range of 1.7 to 2.5GHz.

Within the *1804* is an rf amplifier with a 2.3dB noise figure and providing 14dB gain. The mixer section provides 4dB of gain, with typically 13dB noise. Usable frequency range of the device is 1.8 to 1.925GHz.

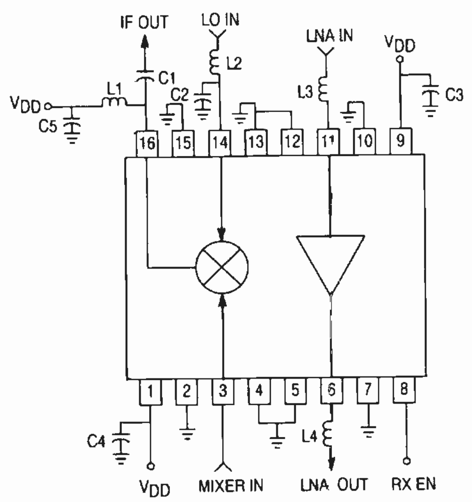
Motorola European Literature Centre, *88 Tanners Drive, Blakelands, Milton Keynes MK14 5BP. Tel. 0908 614614, fax 0908 618650.*



Local oscillator versus rf feedthrough for the 1803 1.8GHz up-converter.

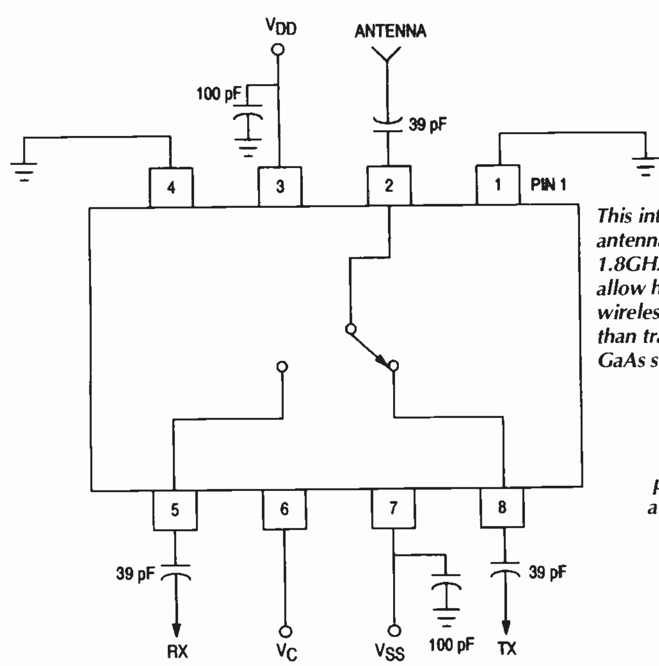


Gain and noise of the 1804 DECT low-noise amplifier versus frequency.



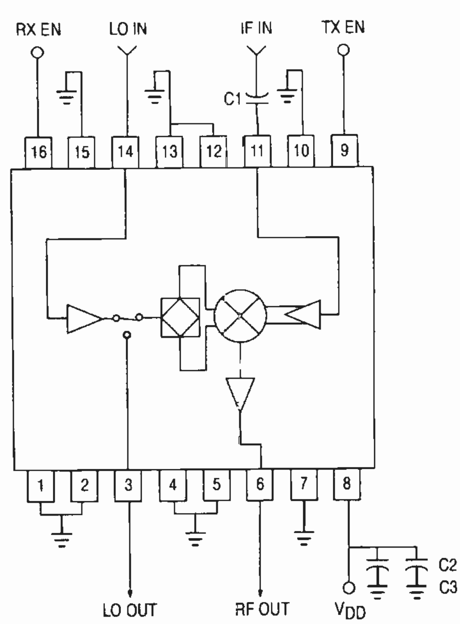
C1	12 pF (110 MHz) or 7.5 pF (240 MHz)
C2	0.8 pF
C3, C4	100 pF
C5	1000 pF
L1	82 nH (110 MHz) or 15 nH (240 MHz)
L2	8.2 nH
L3, L4	0.3 nH (Microstrip)

Complementary to the 1803 up-converter, the 8104 integrated DECT lna and down-mixer has a 0.9dB mixer input intercept point and needs only –5dBm of local-oscillator power. Components are for 110 and 240MHz IFs.



This integrated antenna switch for 1.8GHz is said to allow higher power wireless systems than traditional GaAs switches.

C1	470 pF
C2	100 pF
C3	0.1 μF (optional)



Combined within this up-converter for DECT and other personal communication systems are an active up-mixer, an exciter amplifier and local-oscillator buffer. No external baluns are needed and the IF, LO and RF ports are matched to 50Ω.

Micropower op-amp for 2.7V supplies

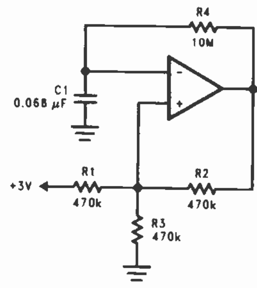
These circuits are designs based on a new op-amp from National Semiconductor. Called the *LMC6572*, this IC is optimised for low-voltage and battery-powered applications involving supplies down to 2.7V.

To obtain a rail-to-rail output swing at low supply voltages and relatively high loads, the unity-gain output buffer traditionally found in ICs has been abandoned. Instead, output is taken directly from an integrator. This is said to provide low output impedance combined with high gain.

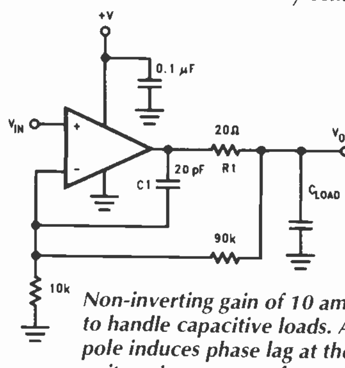
Feed-forward techniques are used to maintain stability over a wider range of operating conditions than is usual for micropower op-amps. Ultra-low-power op-amps are normally very slow. This device has a 0.22MHz gain-bandwidth product together with a typical slew rate of 90V/ms given a 2.7V supply.

There is a quad version of the *LM6572* dual op-amp called the *LM6574*. In the joint *6572/6574* data sheet, there is further information on how the devices are compensated for input capacitance and capacitive loading. There is also further text on pcb layout for high-impedance circuitry.

***National Semiconductor**, The Maple, Kembrey Park, Swindon, Wiltshire SN2 6UT, Tel. 0793 614141, fax 0793 697522*



Since each op-amp in the LM6572 consumes typically 40µA, a 1Hz oscillator such as this can run for long periods from two small dry cells.

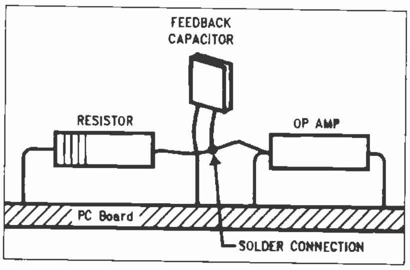


Non-inverting gain of 10 amplifier designed to handle capacitive loads. A feedback-loop pole induces phase lag at the op-amp's unity-gain crossover frequency. This results in oscillation or underdamped response when driving capacitive loads – a problem solved here by the addition of a few passive components.

Air improves op-amp guarding

Most op-amp application information prescribes making guard rings for high-impedance op-amp inputs from a ring of pcb copper. According to the *LM6572* data sheet, guarding is improved using air insulation.

To implement air insulation, the IC input pin concerned is bent away from the pcb. Components are then soldered directly to the pin. NS admits that some of the advantages of using a pcb are lost, but says that these can often be outweighed by the improvement in guarding performance



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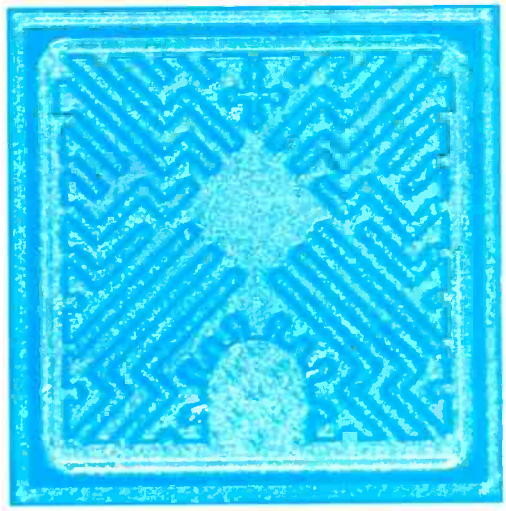

In the same issue: Circuits for rms conversion. There are only two types of circuit which will measure the power content of a signal directly. You can either amplify it and use the resulting output to heat a thermocouple, or you can process the signal with a root mean square algorithm. Danny Ayers tells all.

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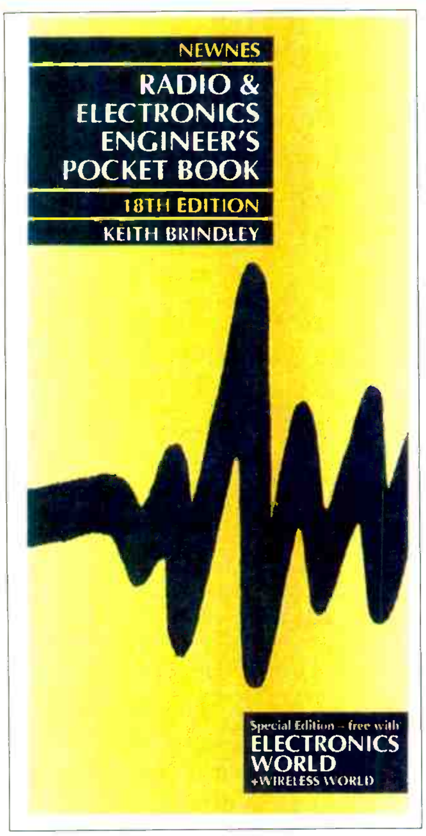

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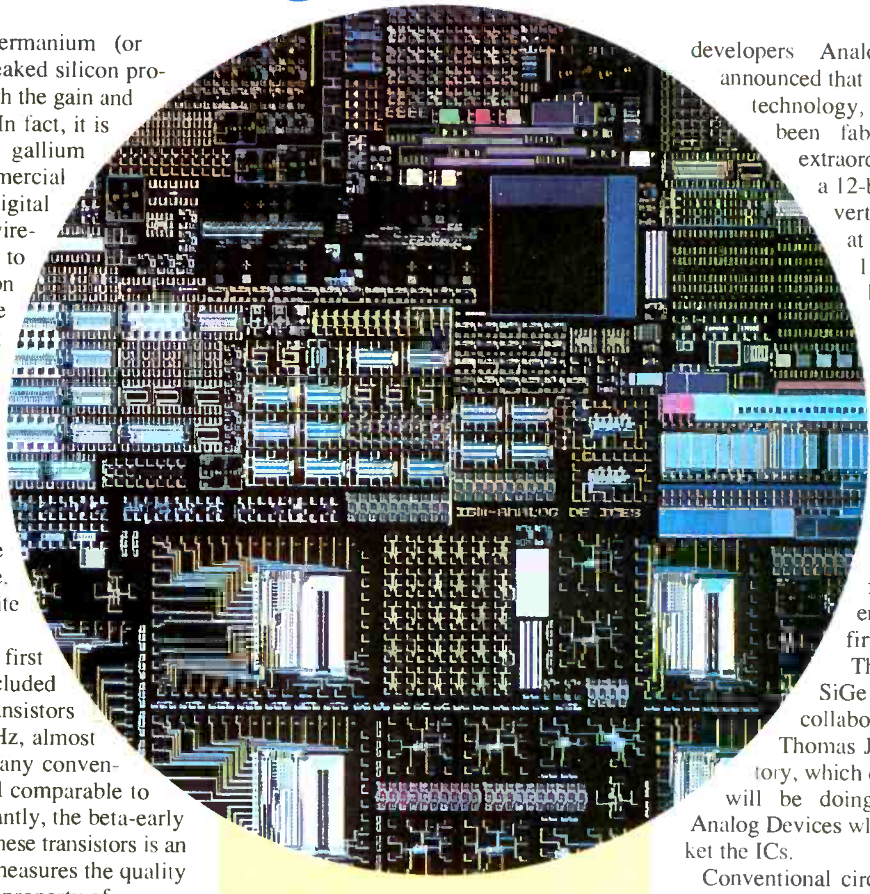
Imagine a semiconductor technology which delivers VLSI density from transistors with a transition frequency greater than 100GHz, yet costs virtually no more to process than standard cmos. The new silicon-germanium IC manufacturing process suggests that it might just deserve the hyperbole "the most significant development in electronics since cmos". Jon Mosely investigates.

Silicon-germanium: the integrated future?



In essence, silicon-germanium (or SiGe) is a slightly tweaked silicon process, that improves both the gain and the speed of transistors. In fact, it is fast enough to replace gallium arsenide in most commercial applications, including digital cellular telephones and wireless lans (frequencies up to 3GHz). Finally, as a silicon process, it can be made inexpensively on existing process lines and may easily be integrated with conventional circuits such as sub-micron vlsi cmos.

This means that a single chip cellphone, integrating DSP, mixed-signal and rf stage into one chip, is quite conceivable. Better still, it could be quite affordable.

More specifically, the first results published have included heterojunction bipolar transistors with f_T as high as 110GHz, almost twice those reported for any conventional silicon process and comparable to GaAs. Perhaps as importantly, the beta-early voltage product (βV_a) of these transistors is an impressive 48,000. This measures the quality of a current source, a key property of an analogue circuit; to be useful for precision tasks the product must reach at least into the thousands). In other words, the speed has not been achieved at the expense of having worthwhile devices.

To give an idea of just how radical this leap in performance is **Fig. 1** is a graph of the fastest Si transistors reported anywhere; and shows how much faster SiGe is.

The technology has been discussed for a few years, but became real when joint process developers Analog Devices and IBM announced that not only was SiGe a viable technology, but that a complex IC had been fabricated which delivered extraordinary results. The part was a 12-bit digital to analogue converter which could be clocked at 1GHz, and drew less than 1W. This is about two and a half times faster than the fastest commercially available silicon DAC.

> ***"This is about two and a half times faster than the fastest commercially available silicon DAC"***

The fact that a IC of such density (3000 devices) had been made, and reportedly with high yields, obviously made SiGe a lot more than a vague possibility or laboratory curiosity, and the further results that have emerged since seem to confirm its potential.

The commercialisation of SiGe technology results from a collaboration between IBM's Thomas J Watson research laboratory, which developed the process and will be doing the manufacture, and Analog Devices which designs and will market the ICs.

Conventional circuits are made with pure silicon/silicon junctions (homojunctions); adding a second material (in this case a silicon/germanium alloy) forms a heterojunction, with a natural electric field generated where the two different materials meet. By controlling the concentration of germanium, a graded junction can be built (**Fig. 2**) with a profile that affects the properties of the transistor, increasing both gain and speed. This bandgap engineering improves device performance. Germanium has a smaller bandgap than silicon; introducing a small amount of it reduces

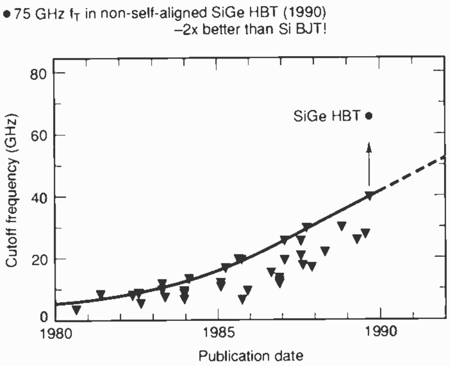


Fig. 1. Trends in cut off frequency. SiGe is a huge leap in speed over the fastest Si transistors reported.

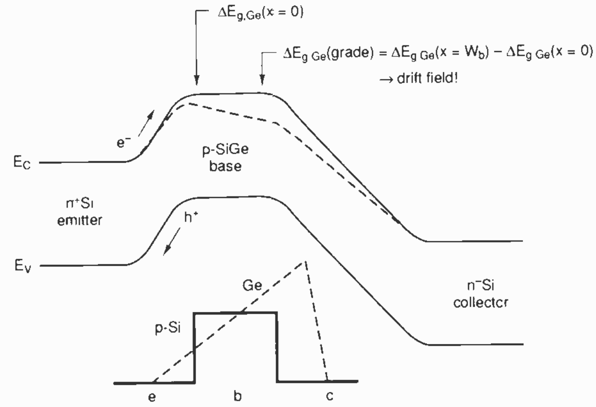


Fig. 2. SiGe HBT Band Diagram. The germanium is implanted across the base of the transistor, with concentration rising toward the collector. This graded junction causes a drift field which accelerates electrons through the transistor.

the transistor's base bandgap and increases electron injection, increasing gain, B. The increased gain allows a more heavily doped base, lowering the base resistance. Finally, the bandgap can be controlled to generate an electric field as great as 30,000V/cm across the base, accelerating the electrons and reducing their transit time through the transistor.

In effect, the heterojunction 'tilts' the circuit, forming a slope down which electrons can slide more quickly than before. This reduces the energy and time required to move from one side of the junction to the other. Some engineers have referred to the addition of germanium as 'electron grease' since it eases the movement of the particles.

The concept has been known since the 50s, but the difficult part has been in successfully fabricating a heterojunction. Although silicon and germanium share the same crystal structure, their lattice spacings differ: Ge's is 4% greater than that of Si. If the pure materials were simply deposited together, this mismatch would cause strain, forming defects and preventing circuits from operating. By analogy, if a nut and bolt are of different screw pitches it may be possible to force a match, but the threads will be stripped. This mismatch frustrated engineers trying to build SiGe heterojunction devices.

The answer was developed by Dr Bernard Myerson, IBM Research Fellow who leads the IBM SiGe team.

The approach uses a silicon-germanium alloy, which has a lattice spacing part-way between the two substances, reducing the mismatch and strain. As a result, the heterojunction can be made between the pure silicon and the alloy without defects, although the lattice is still under some stress.

The junction is built up with low temperature ultra high vacuum chemical vapour deposition (UHV/CVD), which uses a mixture of gases to leave a thin, highly controlled, layer of atoms on top of a silicon substrate. Conventionally CVD requires high tempera-

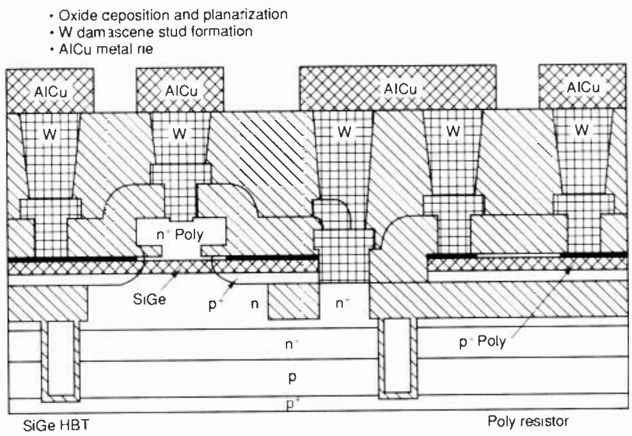
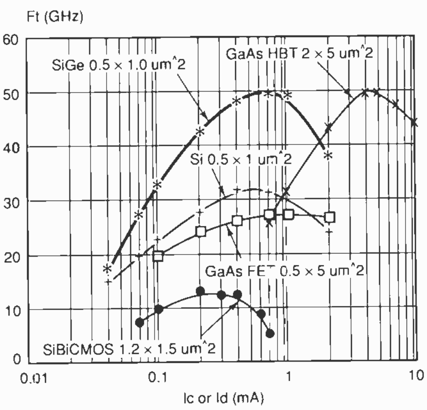


Fig. 3. The structure of an SiGe HBT



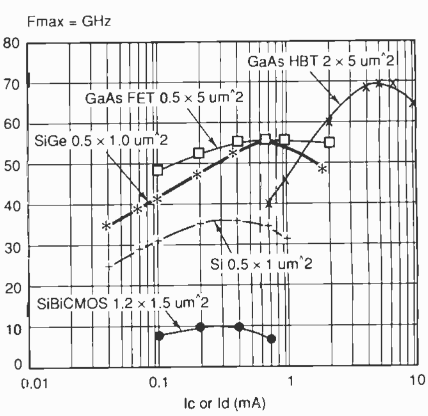


Fig. 4 and 5 show the transition frequency (FT) and maximum oscillation frequency (Fmax) of the SiGe HBTs and compares this to alternative processes.

Table 1. Test results for competing rf technologies

		SiGe HBT ADI/IBM	Si BJT IBM	AlGaAs/ GaAs HBT Hitachi	GaAs mesfet M/A-COM	Si BJT bicmos
Device minimum size (µm)		0.5 x 1	0.5 x 1	2 x 5	0.5 x 5	1.2 x 1.5
BV_{CEO}/BV_{DS} (V)		4	4	15	8	6
f_T (GHz)		50	32	50	30	13
F_{max} (GHz)		55	35	70	60	11
G_{max} (dB)	@2GHz	28	24	19	20	17
	@10GHz	16	11	13	13	1
NF min (dB)	@2GHz	0.5	-	1.5	0.3	-
	@10GHz	0.9	-	-	0.9	-
–1dB p_{out} (dB)		9	9	16	12	9
Added efficiency (%) @3V		70	-	60 @5V	70	40
$1/f_{corner}$ (kHz)		0.1-1	0.1-1	1-10	10,000	0.1-1

Source: Kerrmarec et al, IEEE MTT-S, San Diego May 23-25 1994

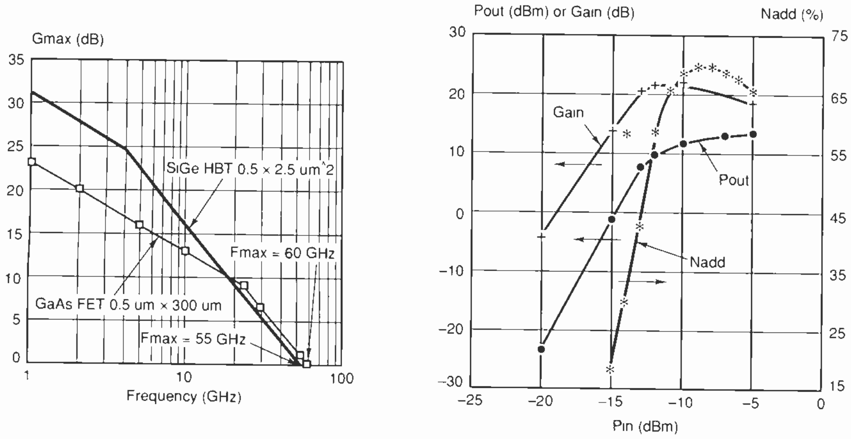


Fig. 6 and 7 show the gain and efficiency of the SiGe HBTs and compares this to alternative processes.

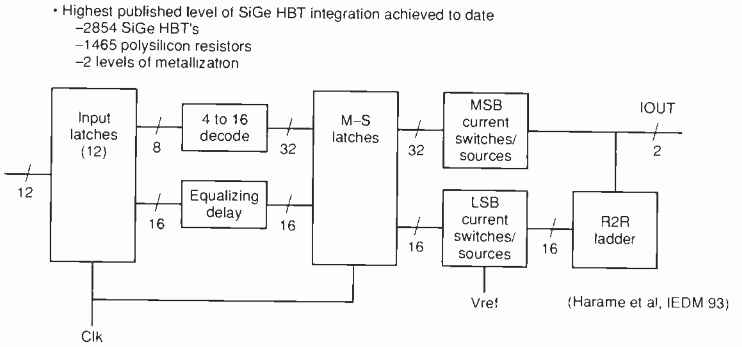


Fig. 8. 12-bit 1GHz digital to analogue converter: block diagram.

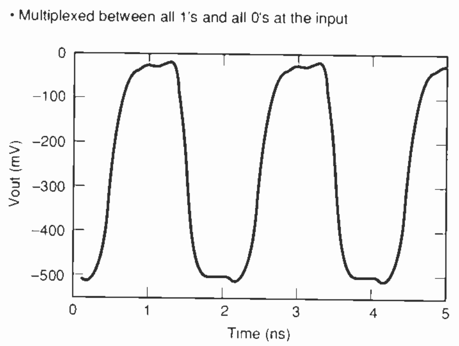


Fig. 9. 12-bit 1GHz digital to analogue converter: output waveform.

> **"The key fact of SiGe is not merely its speed, but that such speed can be achieved from a standard silicon process using existing equipment"**

tures, but this will not work for SiGe, as the strained material will fail. By developing their own innovative CVD process machinery, Dr Myerson and his team were able to perform the deposition at temperatures low enough for the circuit to endure.

Figure 3 shows the structure of a complete SiGe HBT. While researchers outside IBM (including Motorola, M/A-COM, NEC and Daimler Benz) are also working on SiGe, they are achieving fewer devices and lower yields. In addition, most are working on classic mesa structures, which are incompatible with conventional processing techniques.

The first devices were fabricated in 1989, and even those trial parts achieved speeds more than twice of the fastest comparable silicon device (f_T of 75GHz). More recent results have achieved even higher speeds – f_T up to 117GHz.

Figures 4 and **5** show the transition frequency and maximum oscillation frequency (f_{max}) of SiGe HBTs and compares this to alternative processes.

Not only are these fast transistors, they are also exceptional analogue transistors. Their βVa is at least an order of magnitude better than the fast silicon processes used today. The βVa of a low voltage, high speed silicon device would typically be under 1000 compared to the 48,000 achieved by SiGe.

The gain and noise performance are also good: 25dB gain at 2GHz with a corresponding noise figure of 0.5dB at 2GHz and low 1/f noise. Finally, the process seems to achieve good efficiency (critically important for portable and battery powered equipment), with a 70% power added efficiency at 900MHz and operation from either 3V or 1.5V supplies. (**Figs 6** and 7)

SiGe transistors are comparable in speed to gallium arsenide mesfets or HBTs. However, unlike GaAs, the SiGe devices have a very low 1/f corner frequency – between 100Hz and 1kHz – allowing them to be used for dc coupled amplifiers or zero IF down converters.

There are some aspects which are difficult to determine about a technology which has hardly moved out of the conference circuit. One concern is reliability. Due to the strain in the lattice, it is possible that devices will not cope well with radiation or high temperatures. However, this is only speculation – the experimental results have not been revealed.

Breakdown voltage – just 4V between collector/base – may be more problematic. This allows a maximum swing of just 8V. While adequate for low power signal processing or receiver circuits, it looks very low when discussing a power amplifier. Even for battery powered systems, the power amplifier still needs a good voltage swing and 8V is marginal. When you consider battery overvoltage or antenna mismatch, it is probable that an rf PA design would not be practical in SiGe.

Commercially, the key fact of SiGe is not merely its speed, but that such speed can be achieved from a standard silicon process using existing equipment and compatible conventional cmos. IBM will be manufacturing the new devices on a full-size 8in wafer line. As a result the volumes and economies of scale are immense.

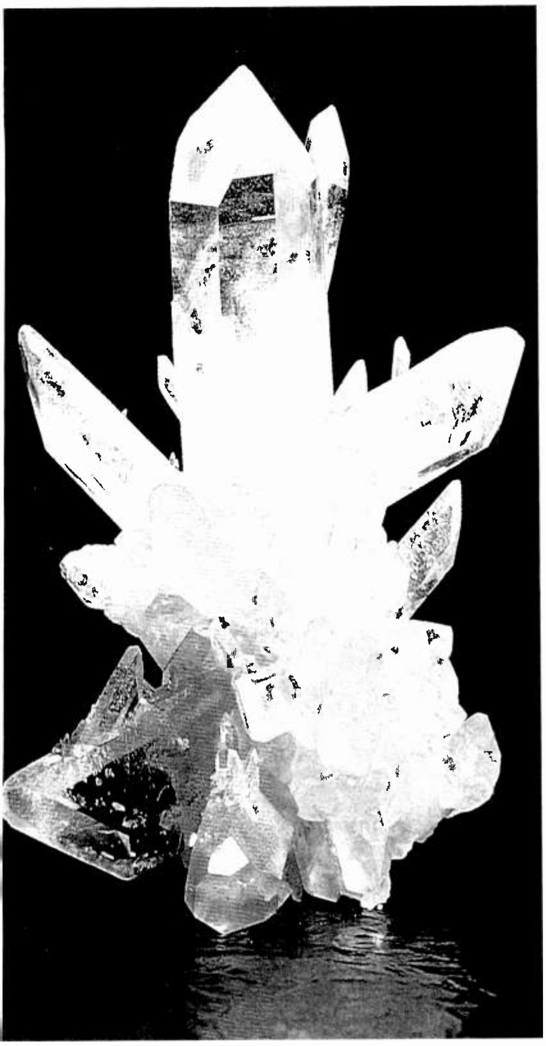
Because the process is based upon a silicon substrate, it is easy to combine it with conventional circuits; for example sub-micron cmos. This would allow single chip cellular radios using SiGe for the front end radio stage, and cmos for the complex digital and mixed-signal stages.

The first SiGe product will be a version of the dac, and this will be available at the end of the year. Apparently, it is now being sampled for use in a fibre to the kerb application in Eastern Europe; a number of channels are combined and digitised before being sent down a fibre-optic link at rates of 9Gbits/s. At the kerbside they are received and fed into the 12-bit dac, to produce a 750MHz bandwidth analogue signal. **Figure 8** shows the block diagram while **Fig. 9** gives the output waveform for toggling between 0 and full scale at a rate of 1GHz. Incidentally, the dac may be faster than 1GHz but, at the time these results were released, that was the maximum speed of the available test equipment!

Subsequent commercial products are likely to be highly integrated blocks for use in wireless systems, perhaps even complete rf transceivers in a chip. This would combine up- and down-converters, frequency synthesisers, mixers and amplifiers into a single block.

Silicon germanium has the potential to redefine high speed semiconductors. The only question is when it will be solid enough for an appearance in catalogues? ■

Applying crystals



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This round up of general-purpose crystal oscillator designs – both discrete component and logic-gate based – satisfies frequency generation requirements from kilohertz to 200MHz. By Martin Eccles.

Quartz crystals used for time keeping and frequency references combine three key features – extremely high Q, small physical size relative to alternative timing devices with the same performance, and excellent temperature stability.

Frequency stability of a crystal is limited in the short term by its temperature coefficient and in the long term by ageing. Quartz crystals with an AT cut generally provide the best temperature coefficient and usually have stability tolerances of ±0.0025% or ±0.005% from –55 to 105°C.

Between 1 and 200MHz, AT-cut crystals are normally chosen. They represent the best compromise between temperature stability and frequency accuracy and frequency pulling capability. Above around 27MHz however, AT crystals are normally only available for use in overtone mode. This is restrictive since overtone oscillators are more difficult to design and are prone to spurious responses, **Fig. 1**.

Recently, BT-cut crystals have been developed to overcome the overtone problem. These operate in fundamental mode up to 46MHz – almost double the frequency of traditional AT-cut alternatives. In addition, the BT-cut devices can operate in overtone mode.

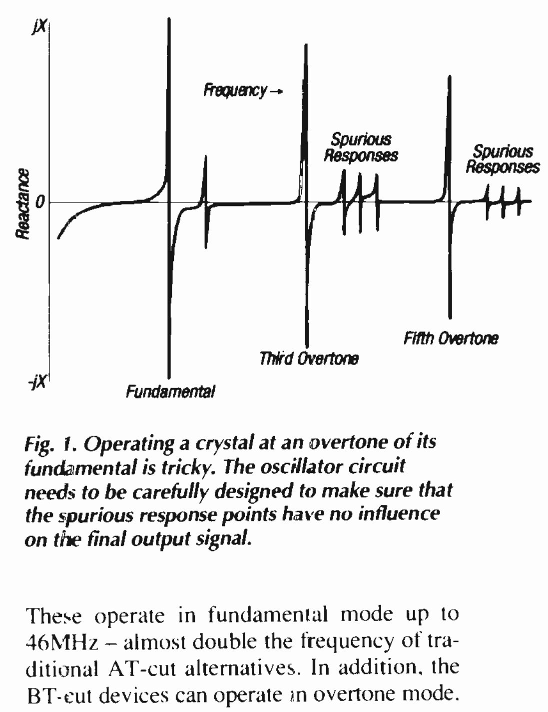

Fig. 1. Operating a crystal at an overtone of its fundamental is tricky. The oscillator circuit needs to be carefully designed to make sure that the spurious response points have no influence on the final output signal.

Load capacitance and frequency pulling

Near a crystal's resonant frequency, lowering capacitive loading on the device increases its output and frequency, **Fig. 2**. In addition to any external loading, the crystal has its own shunt capacitance, C_0, which is typically between 3 and 15pF. When making calculations, this capacitance should be added to that of the load.

Figure 3 shows how far the frequency of a typical crystal can be pulled by changing its load capacitance. The degree of pulling for a given configuration is obtained by;

$$\frac{C_1 10^6}{2(C_o + C_L)^2}$$

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Sample pack parts

Part	Description	Package	Cut	Use and typical circuit
A160K	15MHz	HC49/4H	AT	Microprocessor clocking, Fig 6
A123A	4194.304kHz	HC49	AT	Timers and RTCs, Fig. 14 lower
A103A	32.7680kHz	3X8	Tuning fork	Clocks/watches, Figs 13 & 14 lower
A166A	32MHz	HC49	AT 3rd o/t	Microprocessor clocking, Fig. 7

where C_1 is motional capacitance, C_L is load capacitance and C_0 is shunt capacitance. Typical values for motional and inherent shunt capacitances are shown in **Table 1**.

All crystals operate at series resonance. The term parallel resonant is often used to describe a crystal designed to handle a high load impedance across its terminals. Series resonance exists within the crystal, whereas parallel resonance exists only as a crystal measurement phenomenon.

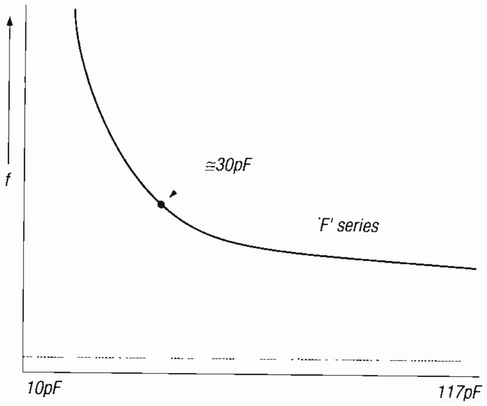


Fig. 2. Lowering capacitive loading on a crystal increases its output frequency.

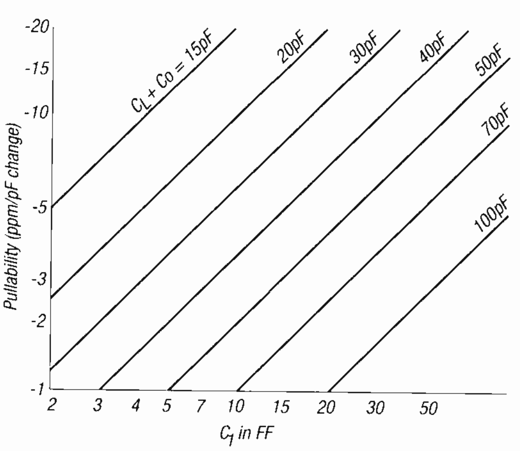


Fig. 3. When estimating how far a crystal's frequency can be pulled by changing its load capacitance, C_L, both package shunt capacitance, C_0, and motional capacitance C_1 need to be taken into account.

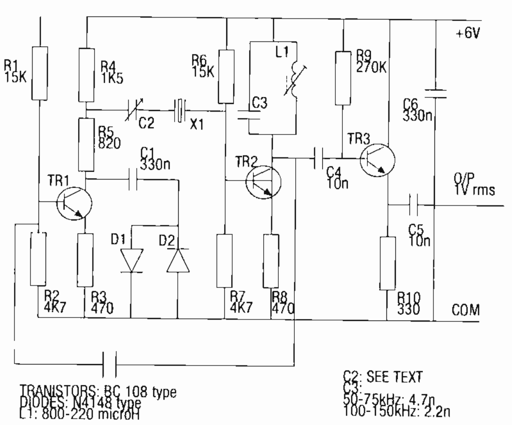


Fig. 4. At frequencies below 150kHz, equivalent series resistance of a crystal is high so oscillator circuits need to have a high gain.

Using quartz crystals

For a circuit to oscillate, it has to have both positive feedback and a loop gain greater than unity. Without any other frequency-sensitive elements in its oscillator circuit, a crystal will oscillate in its fundamental mode. Frequency dependent components need to be added to the circuit to force the crystal to oscillate at an overtone.

Closing an oscillator's feedback loop causes sine wave oscillations which gradually build up, clip, and approach a square wave as the circuit overloads. Crystal oscillators can usually provide either a sine or square wave. The driving signal is usually a square wave while the crystal output is always a sine wave. Either of these waveforms can form the output.

Below 150kHz. Because equivalent series resistance of a crystal is high at low frequencies, high amplifier gain is needed. In **Fig. 4**, which is optimised for 50kHz, the gain problem is solved using dual cascaded common-emitter stages.

Two diodes limit crystal drive level to avoid damage, while tuning in the middle transistor's collector adds selectivity. For parallel resonance, the adjustable capacitor provides calibration. For series calibration, replace the trimmer with 1nF and trim via the inductor.

150 to 550kHz. For this frequency range, DC and CT-cut crystals are normally used with a selective amplifier to suppress unwanted modes, **Fig. 5**. Initially, the crystal is shorted and frequency is adjusted to near the crystal frequency via the inductor. Afterwards, the inductor can be used for trimming.

Series-resonance is preferable but a parallel-resonant device can be used if the first capacitor is replaced by one with a value equal to the crystal load capacitance.

0.95 to 21MHz. Circuitry in **Fig. 6** is designed for high-stability AT-cut crystals calibrated at parallel resonance and operating in fundamental mode.

15 to 105MHz. At higher frequencies, crystals specially designed to oscillate at their overtones are mainly used. Examples of third and fifth overtone oscillators are shown in **Figs 7 & 8** respectively. The first example handles crystals from 15 to 63MHz while the second is suitable for 50 to 105MHz. As with Fig. 5, the crystal is initially shorted and the oscillator set to run as near as possible to the crystal frequency via the inductor. After removing the short, the inductor can be used for trimming.

By adding a tuned circuit at twice or three times the crystal frequency in the transistor's

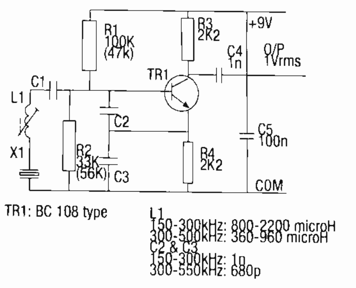


Fig. 5. Relative to circuitry needed for oscillators working at under 150kHz, this circuit for 150 to 550kHz is much simpler.

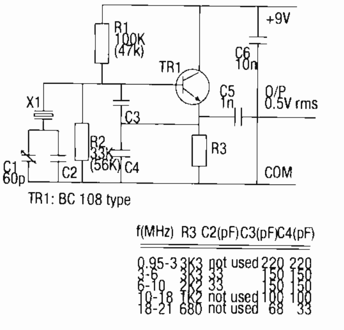


Fig. 6. Designed for use with high-stability AT-cut crystals, this oscillator operates between 0.95 and 21MHz.

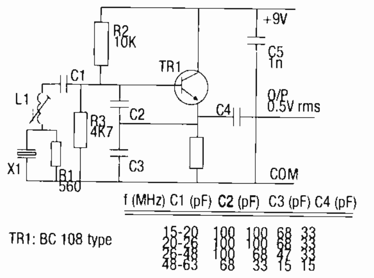


Fig. 7. Third-overtone crystal oscillator configuration for frequencies from 15 to 63MHz.

Table 1. Using motional and inherent shunt capacitances of a crystal, it is possible to work out how far the device's frequency can be pulled.

Frequency (MHz)	Vibration mode	C_1 (fF)	C_0 (pF)
1 to 2	fundamental	5-8	3
2-4	fundamental	6-12	3
4-6.5	fundamental	8-20	5
6.5-30	fundamental	16-25	6
21-150	3rd o/tone	1-2.5	6
60-150	5th o/tone	<0.7	6
85-210	7th o/tone	<0.4	6

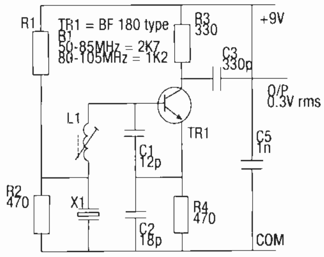


Fig. 8. By driving a crystal into its fifth overtone mode, this oscillator circuit extends useful frequency range to cover 50 to 105MHz.

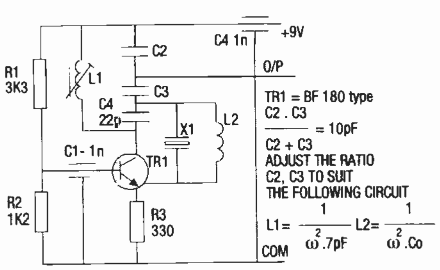


Fig. 9. Above 105MHz, stray capacitances make designing a reliable oscillator difficult. To help prevent oscillation not controlled by the crystal, this circuit uses a small inductor in parallel with the crystal to tune out its static capacitance.

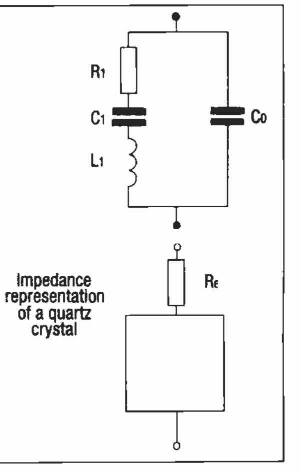
Crystal characteristics

A crystal's equivalent circuit is shown below. Series combination R_1, C_1 and L_1 represent the quartz while C_0 is shunt capacitance of the electrodes in parallel with capacitance of the can.

Reactance versus frequency is also shown. Resonant frequency F_s is given by;

$$F_s = \frac{1}{2\pi\sqrt{L_1 C_1}}$$

where F_s is series resonant frequency in hertz, L_1 is motional arm inductance in henries and C_1 is motional arm capacitance in farads.



Typical crystal parameters

Parameter	200kHz Fundamental	2MHz	30MHz 3rd o/tone	90MHz 5th o/tone
R_1	2kΩ	100Ω	20Ω	40Ω
L_1	27H	520mH	11mH	6mH
C_1	0.024pF	0.012pF	0.0026pF	0.0005pF
C_0	9pF	4pF	6pF	4pF
Q	18×10^3	18×10^3	18×10^3	18×10^3

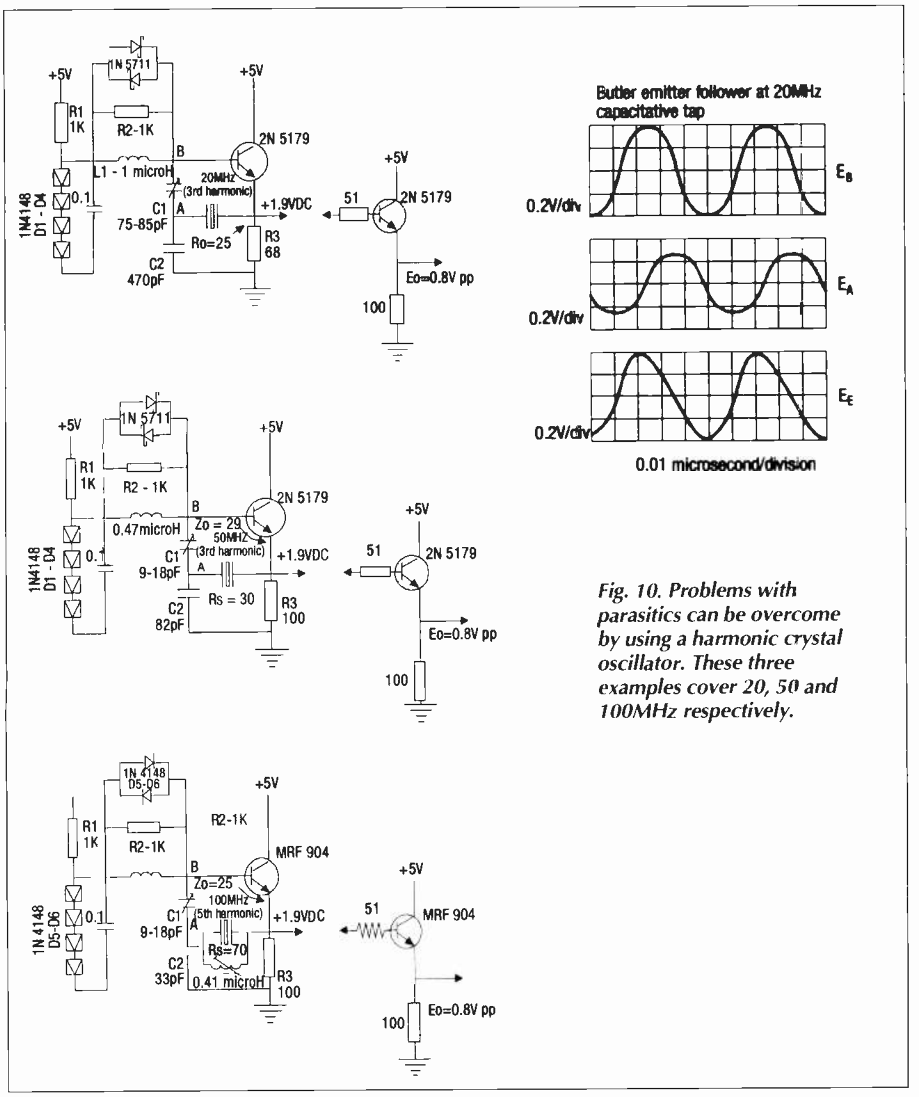


Fig. 10. Problems with parasitics can be overcome by using a harmonic crystal oscillator. These three examples cover 20, 50 and 100MHz respectively.

collector, it is possible to extract harmonics of the crystal frequency. This is a simple and economical way of making a vhf oscillator.

Above 105MHz. Reactance from stray circuit capacitances can make high-frequency oscillators difficult to design. In the solution shown in **Fig. 9**, a small inductor is added over the crystal to compensate for its inherent capacitance, which may cause undesirable effects. Tuning is carried out via the other inductor or by inserting a variable reactance in series with the crystal.

20-200MHz. Parasitics can be eliminated by using an oscillator relying on the crystal's harmonics. **Figure 10 a)** is such an example for operation at around 20MHz. In it, the crystal is tapped into the capacitive side of the *LC* tank.

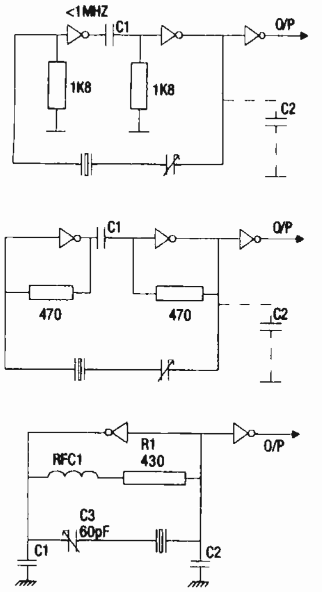
Four signal diodes provide base biasing. Emitter output driving the crystal is 25Ω while the crystal load is around 35Ω. Figures 10 **b,c)** show values needed to make the same circuit operate at 50 and 100MHz respectively.

Oscillators from logic gates

Designs for oscillators using standard ttl logic gates as their active components are shown in **Fig. 11**. With nand gates, unused inputs should be tied to the positive rail while with nor gates spare inputs should be grounded.

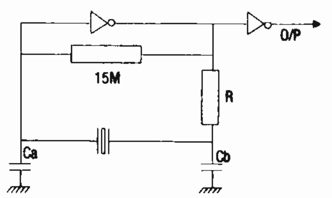
In Figs **11 a,b)**, for frequencies of less than 1MHz and 1-4MHz respectively, the fixed capacitors are best determined experimentally; the dotted component may not be needed at all. Figure **11 c)** is for frequencies from 4 to 14MHz.

Jitter can be a problem in ttl-based oscillators due to random phase shift within the IC. In addition, the relatively high crystal drive



f1(MHz)	C1 (pF)	C2 (pF)	RFC1 (microH)
4-6	330	220	100
6-9	330	180	68
9-11.5	220	150	39
11.5-14	330	120	22

Fig. 11. Three crystal oscillators based on ttl logic gates, covering <1MHz, 1-4MHz and 4-15MHz respectively. Many similar circuits have been produced in the past but not all work reliably and at the intended frequency.



Ca = 1.1 CL
Cb = 4 CL
R = -1/(11ω²CL² Re)
where
Re = Rl (1 + (Co/CL))

Fig. 13. Using cmos logic, a crystal oscillator can be made from only two gates. This design works up to about 3MHz.

level can affect long-term stability. A discrete component oscillator followed by a buffer such as those in **Fig. 12** provides better stability.

In the first buffer, the resistor in the supply rail decouples the oscillator stage. The complementary buffer has more active components but offers faster switching and improved capacitive load driving. Note that drive level from some of the preceding discrete component oscillators may need reducing to interface to these buffers.

Cmos oscillators

A typical cmos gate oscillator with the crystal forming part of a pi network is shown in **Fig. 13**. This design will work up to 3MHz but values found via the formulas on the diagram may need adjusting to take account of stray capacitance due to component layout. Resistor R is expressed in terms of crystal load capacitance C_L and its equivalent series resistance at parallel resonance R_e.

For operation above 3MHz, resistor R is omitted and the crystal connected directly between the inverter input and output. Low values of crystal load capacitance, of say 12pF, and/or a high supply voltage may be needed for reliable operation.

Reducing power

Timing, clock and calendar circuits often rely on battery-powered frequency references. As a result, power consumption of the frequency reference is important.

Two commonly used timing crystals, both implemented in oscillators designed around a single logic gate, are shown in **Fig. 14**. Given a 5V supply, the 4000 series cmos gate with a 32kHz watch crystal consumes only 30μA. For the same supply, the *HC* gate alternative increases current to 1300μA, mainly due to its higher switching frequency.

Even though the HC circuit appears much worse in terms of power consumption, it can still provide significant advantages in battery-powered applications – in addition to the benefits of its higher operating frequency.

Firstly, during tests based on a National Semiconductor *74HC4060* device, oscillation started at 1.1V. The traditional cmos *4060* needs 3V. And at 1.5V, the HC-based oscillator consumed less than 5μA, **Table 2**. As a bonus, the HC produced much faster edges than the 4000-series IC.

Table 2. Power consumption of cmos versus high-speed cmos crystal oscillators shows that the high-speed alternative is suitable for low-voltage battery applications. Test circuits for both sets of results incorporated a 32kHz watch crystal. Over the supply ranges measured, frequency change was 0.1Hz.

Supply (V)	Consumption (μA) *CD4060*	*74HC4060*
1.1	-	2.3
1.5	-	4.8
2	-	34.9
2.5	-	117
3	7	256
3.5	7	449
4	9	687.3
4.5	20	979
5	31	1325
5.5	46	1712
6	65	2148
8	180	-
10	362	-
12	603	-
14	895	-
16	1252	-

Frequency of a tuning fork watch crystal has a significant negative temperature coefficient. In a watch, this is of little consequence since human body temperature remains constant, but there are many applications where temperature coefficient does matter.

Using a more stable higher frequency crystal and dividing it down circumvents this

Inside the can

Crystal resonators rely on the piezoelectric effect of quartz. Applying stress to a quartz crystal produces an electric charge perpendicular to the stress. Conversely, applying an electric field causes the crystal to deflect mechanically.

In a quartz crystal resonator, a sliver of quartz is placed between two electrodes that apply the field. An ac signal applied to the electrodes causes the resonator to vibrate. When the frequency of the signal is near a mechanical resonance point of the quartz, amplitude of the vibrations becomes very large.

Resonant vibration of the quartz produces a sinusoidal electric field controlling the effective impedance between the electrodes. This impedance is strongly dependent on the excitation frequency and possesses an extremely high Q.

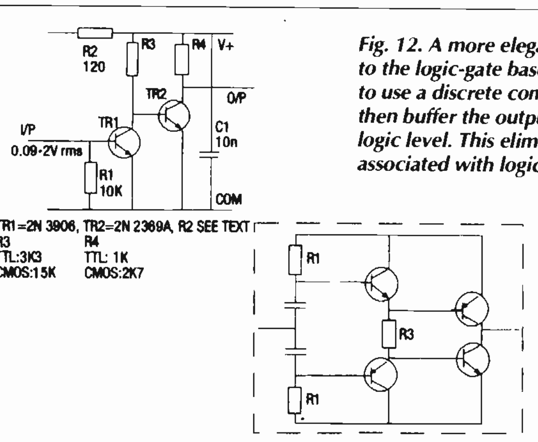
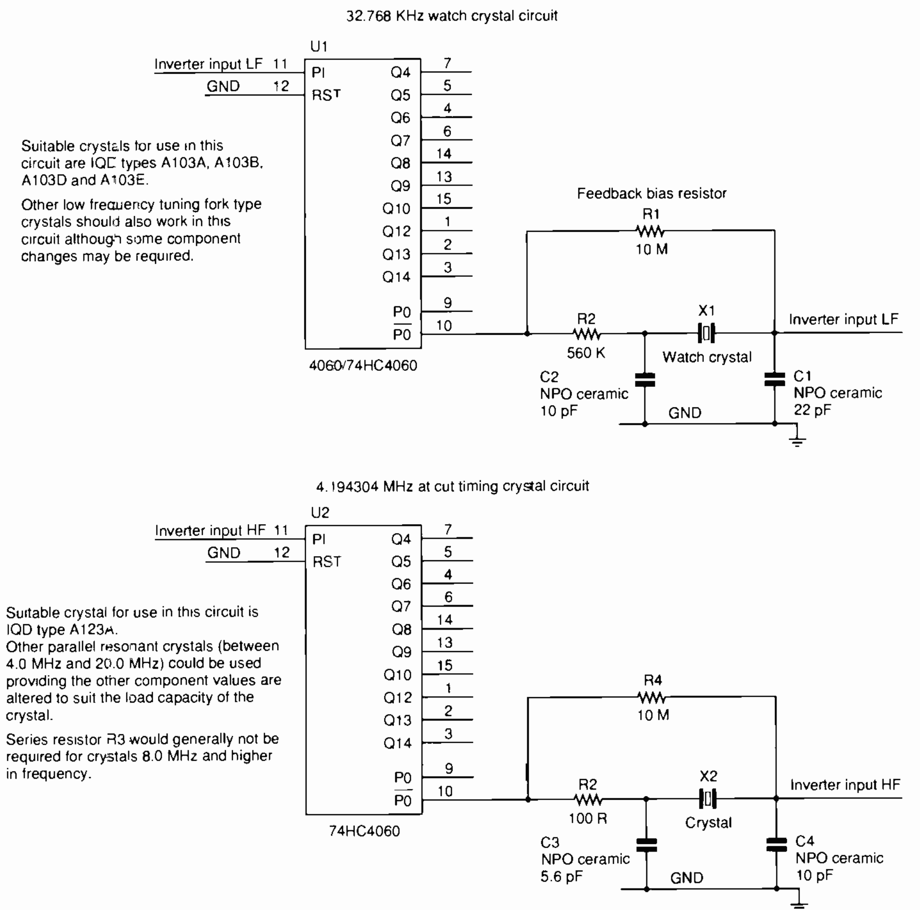


Fig. 12. A more elegant alternative to the logic-gate based oscillator is to use a discrete component circuit then buffer the output to produce logic level. This eliminates the jitter associated with logic-gate designs.

Fig. 14. Two timing circuits – one using a watch crystal and one a crystal with a frequency often used for real-time clocks. Although the high-speed cmos circuit consumes more power with a 5V supply, it can have benefits in battery circuits since it operates down to 1.5V. It is also significantly more stable.



problem. With a faster crystal, it is quite easy to achieve stability of ±10ppm over between 0 to 50°C.

To reduce current flowing around the circuit, low-power oscillators are designed with a relatively small crystal load of typically 12pF. Using a *74HC4060*, the circuit shown consumes about 2.4mA at 5V. It will start reliably with a 1.5V supply, at which point the circuit draws about 360μA.

In both circuits, NP0 or COG dielectric capacitors should be used in the network around the crystal because of their low temperature coefficient. ■

Crystal source

Crystals suitable for all the applications described in this article – and most other quartz crystal applications – can be obtained from IQD at Station Road, Crewkerne, Somerset TA18 7AR. Tel. 0460 74433, fax 0460 72578. IQD also supplied the information upon which this article is based.

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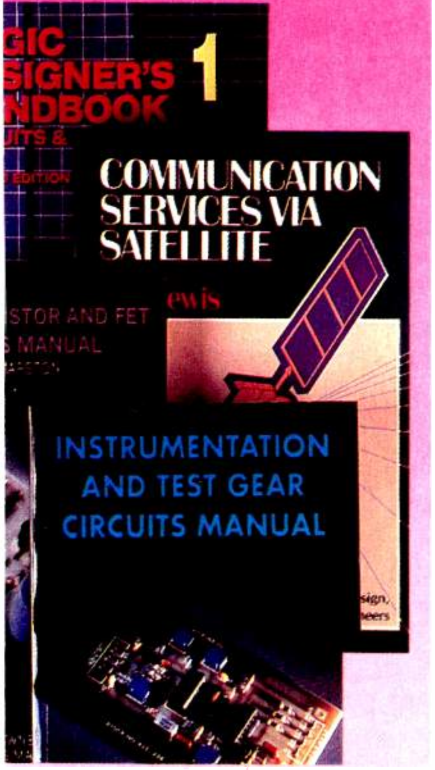

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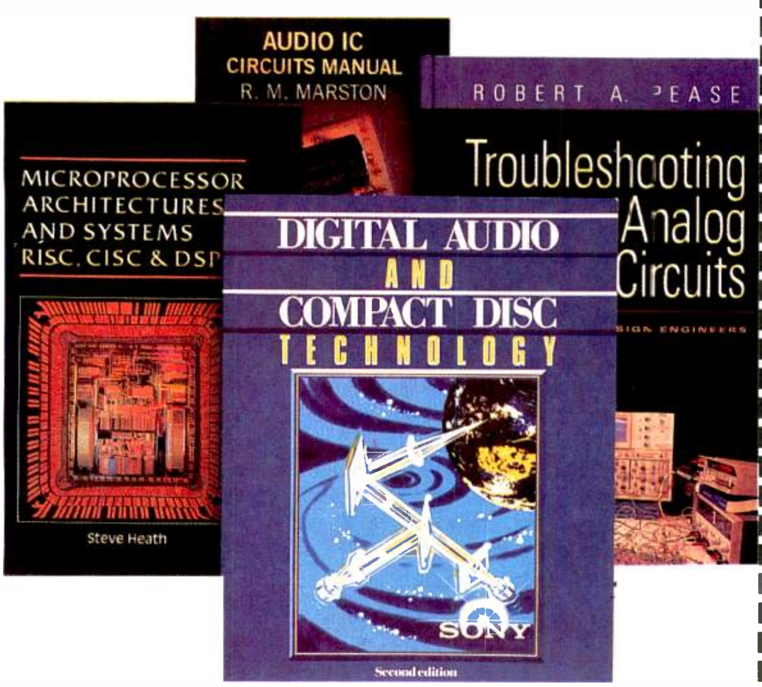

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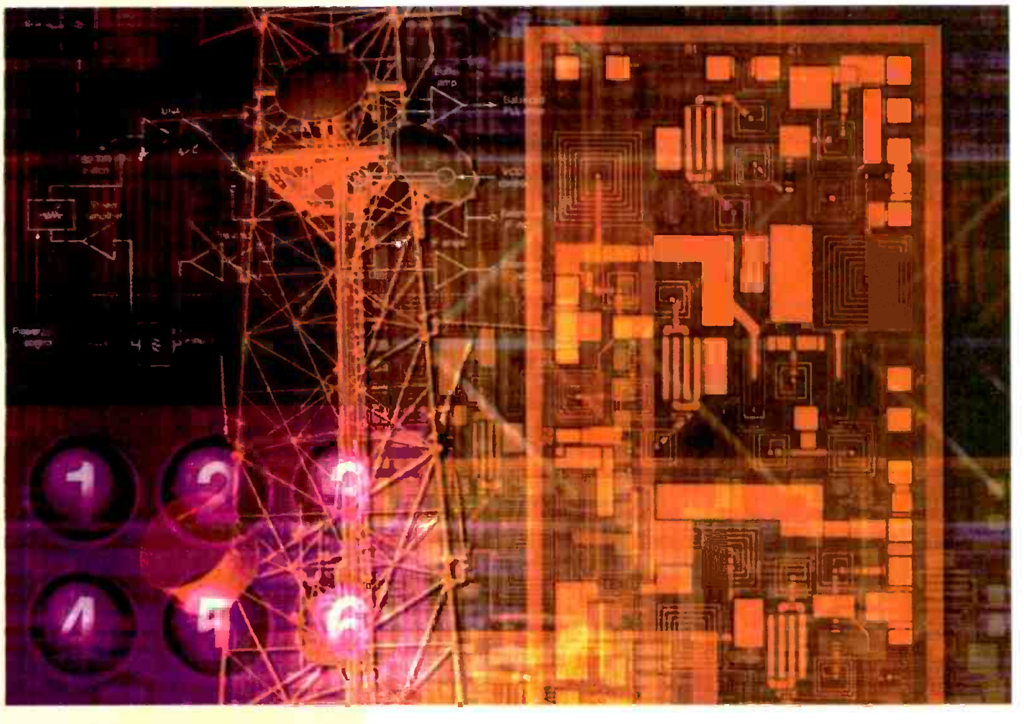
New wave MICROWAVES

5: oscillator and amplifier devices



The GaAs mesfet and its derivatives provide the active element of choice in low noise and medium power amplifier designs to 30GHz. However, new technologies look increasingly attractive. Most exploit the features of the heterostructure. Mike Hosking reports on the new wave of semiconductor devices.

*Mike Hosking is a lecturer in telecommunications and microwaves at the University of Portsmouth.

Gunn and impatt devices, once the main active components for microwave power generation, are now being relegated to millimetre wave and high power operation. The competition has come from mesfet technology, particularly high electron mobility transistor (hemt) and the pseudomorphic hemt or phemt.

These devices take over from the silicon BJT in the overlapping technology band between 2GHz and 4GHz and offer low noise, good efficiency and medium power levels up to the millimetre wave region and beyond (>30GHz). In recent years, a further contender has appeared on the technology stage: the GaAs heterojunction bipolar transistor (HBT). This possesses many of the advantages of the Si BJT: notably low phase noise, high power-added efficiency and a medium power output; but translated in frequency through the microwave and into the millimetre bands.

Frequency multiplication, as a means of signal generation, is not now widely used except when very low phase noise is needed. One particular technique of interest uses a device called a step recovery diode as a harmonic generator and is used in certain local oscillator and instrumentation applications.

Microwave transistors

Starting at the lower end of the microwave region, the silicon BJT dominates amplifier and oscillator technology. Low phase noise and optimal power-added efficiency are essential to PCS mobile phone equipment. The BJT can easily and reliably provide several watts of output power at greater than 30% efficiency in the 960MHz to 1400MHz bands. Bicmos, relevant to the signal processing stages of these systems, combines bipolar speed with cmos complexity.

Si BJT technology can also generate several hundred watts from a single transistor amplifier at around 1GHz. Solid state radars using hundreds of such high power modules, have been successfully integrated into antenna arrays for L-band operation (between 1GHz and 2GHz).

Construction of the microwave Si BJT, **Fig. 1a**, is not so different from its lower frequency counterpart and consists of a very highly doped n-type emitter, a p-type base region of small width ($\approx 1\mu m$) and a low resistance n-type collector contact. Although operation is possible at frequencies above 10GHz, for most practical purposes the BJT is technology limited to below about 4GHz. Electron transit time from the emitter-base junction to the collector sets the upper frequency limit, but the length of this region cannot be made arbitrarily narrow because of voltage breakdown and power handling.

As the mobility of electrons in silicon is greater than that of holes, microwave transistors are always n-p-n types with electrons as the majority carriers. The maximum frequency of oscillation is a function of the time constant created between collector-base capacitance and base resistance. Interdigital geometry for the base-emitter junction, **Fig. 1b**, allows an increase in current by the increase in emitter periphery, without increasing the capacitance.

The BJT exhibits low phase noise, generally many times better than that which can be achieved with fet technology. The phones used in PCS, typically have an rf subsystem as shown in **Fig. 2**, where the same voltage controlled oscillator is used to generate the data-modulated, up-converted carrier as well as the LO for the receiver. With the emphasis on spectrum conservation and a likely change from FM and FSK to high order (16 to 64 state) QAM modulation, phase noise performance is vital. This assures the Si BJT a place in the design of power amplifiers for these systems.

Mesfet devices

The GaAs mesfet dominates low noise amplifier and medium power amplifier applications from microwave to millimetric frequencies. It has revolutionised the performance of DBS receivers and is replacing the travelling wave tube in satellite transponders.

At microwave frequencies, the rectifying junction of the fet is a metal-semiconductor interface of Schottky barrier type, **Fig 3a**.

Electrons in the n-type GaAs diffuse into the metal to form a barrier potential and leave behind a region of high resistance containing few free carriers: the depletion region. Minority carriers play little part in the operation of the junction and, thus, there are virtually no charge storage effects. This allows the junction to switch from conducting to non-conducting states at microwave and millimetre wave frequencies.

Like conventional fets, a reverse bias applied between the gate and source will increase the height of the depletion region until it equals that of the channel, thus pinching-off the current between drain and source. Hence, this is a means of controlling output current and of applying modulation. The mesfet is a transit time device in that the cut-off frequency is determined by the saturation velocity of the electrons and by the time taken to traverse the gate length, *L*. Thus GaAs, with its higher mobility is preferred to silicon for microwave fets.

A 1μm gate length allows operation up to 10GHz and present lithography can achieve 0.1μm resolution, resulting in achievable operating frequencies of 60GHz and higher. At around 10GHz, a chip-only noise figure of less than 1dB could be expected, with a gain of about 12dB. As a power device, several hundred milliwatts with power-added efficiency greater than 25% could be provided at the same frequency.

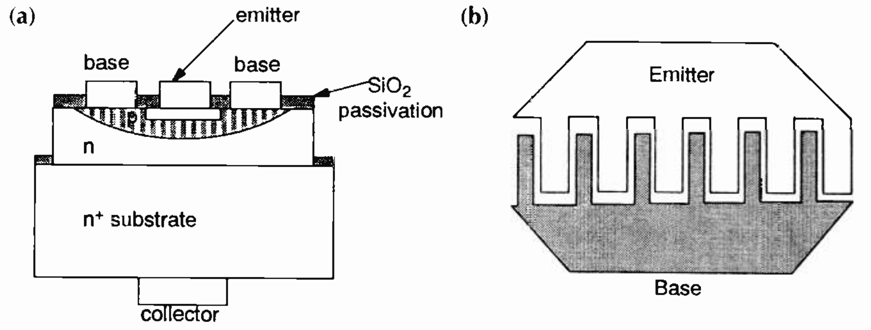


Fig. 1. Basic structure of the silicon bipolar transistor (a) with (b) the general type of interdigitated emitter which allows the current to be increased without a corresponding increase in capacitance.

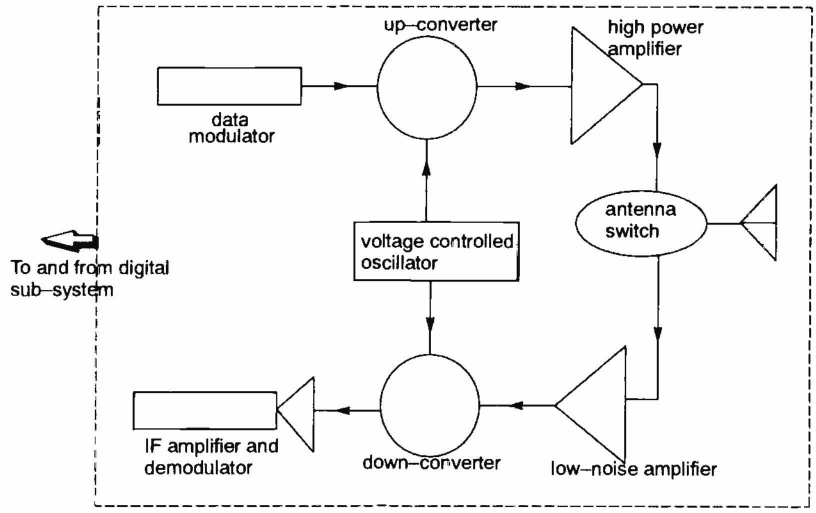
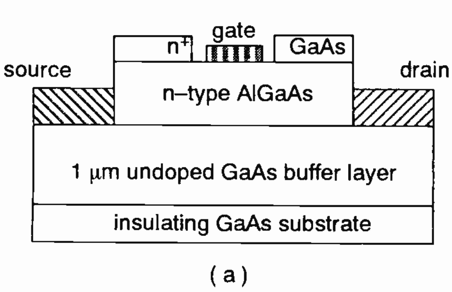


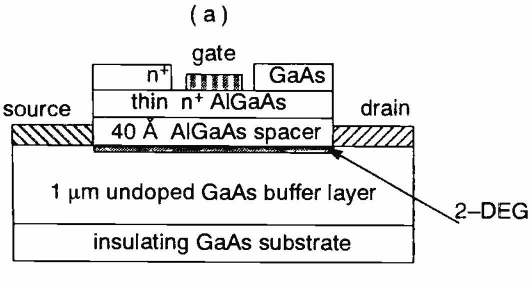
Fig. 2. Typical rf section of a PCS communicator. Different technologies are being used to produce highly optimised devices for each of the major components.

Heterojunctions

Two variants of the classic mesfet device, the hemt and phemt, extend the mesfet operating envelope. **Fig. 3b** shows the construction profile of the hemt for comparison with the standard mesfet. The device is still a FET with regard to its method of operation but, instead of the n-type GaAs channel, the hemt has a thinner (≈500Å) layer of highly doped, n-type AlGaAs grown onto an even thinner (≈40Å) undoped AlGaAs layer. This, in turn is grown onto the normal undoped GaAs buffer layer ≈1μm thick. The key operational difference is that the donor atoms lie in the AlGaAs layer which has an energy band gap of 1.72eV compared with the 1.42eV band gap of GaAs. This creates what is called a potential well of 0.3eV and, when electrons enter the gate channel from the source contact, they are drawn to this energy well and form a narrow sheet of charge. This sheet lies as a thin (<100Å) layer on the GaAs side of the junction, as shown, and is referred to as a two-dimensional electron gas (2-DEG).

In a mesfet, the current fills the whole width of the gate channel defined by the extent of the depletion region. Hence, electrons in the hemt are travelling in the undoped GaAs





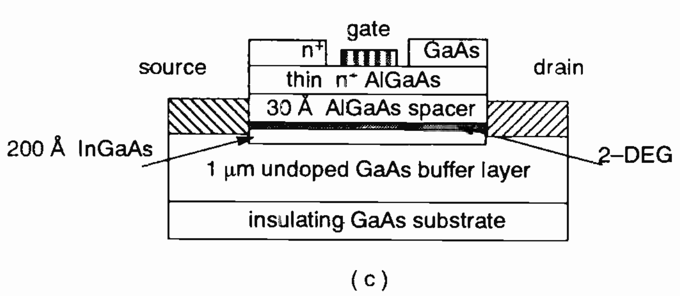


Fig. 3. Schematic layering of variations of the fet to produce the standard GaAs mesfet (a); the hemt (b); the phemt (c). The last two structures rely on the conduction channel formed in a high mobility two-dimensional electron gas cloud which confines the carriers.

which contains virtually no donor atoms in this thin electron gas layer. They experience no impurity scattering, with the result that mobility is increased three-fold over that of the GaAs mesfet with a corresponding improvement in performance. About 0.5dB noise figure at 10dB gain could be obtained at around 10GHz and cut-off frequencies well above 100GHz have been achieved. The hemt is also well suited to power applications: Texas Instruments, for example, offers a device delivering over 5W at about 40% efficiency around 10GHz using 0.5μm geometry.

Further improvements to performance can be obtained by modification to the device structure, **Fig. 3c**. For a sub-micron gate length, the current gain-bandwidth product (f_{max}) of the hemt is approximately given by:

$$f_{max} = \frac{V_t}{2\pi l} \text{ Hz}$$

where V_t is the transit velocity of electrons under the gate and l is the gate length. Thus, a material having a higher electron velocity will contribute directly to a higher frequency of operation.

InGaAs has a peak electron velocity over 30% greater than GaAs and, in the phemt, a thin (≈200Å) layer is introduced on top of the GaAs buffer layer. The two dimensional gas penetrates this higher velocity region and, in fact, is additionally confined by the 0.17eV energy band gap of this material. The structure is called 'pseudomorphic' because there is a slight mismatch between the lattice structure of InGaAs and the GaAs buffer layer. With careful control of the layer thickness, the strain of this mismatch can be accommodated by the InGaAs lattice which distorts to 'mirror' the GaAs structure. Hence, 'pseudomorphic'.

The result of these device modifications are incentives for a higher frequency of operation (f_{max}>200GHz) and lower noise figure (<1dB at 20GHz) with efficiencies around 30%. Phemts have been commercially available for several years, but development is continuous and performance increases steadily.

There is also development effort being spent on the manufacture of hemt devices using InP instead of GaAs to deliver higher frequencies and lower noise. Operation to 300GHz is predicted and Hughes in the USA has already space qualified a 1.2dB noise figure InP hemt at 60GHz. Cooling of fet devices to liquid nitrogen temperatures of 77K results in general noise figure improvement. However, the hemt and phemt structures perform exceptionally well and noise figures similar to those only obtainable at present with cooled masers are predicted from InP.

Bipolar heterojunctions

The latest in microwave transistors is the GaAs heterojunction bipolar device which offers hundreds of milliwatts at 40 to 50% efficiency at millimetric frequencies. The vertical topology of the HBT means that the emphasis on lithography can be relaxed; 2μm geometry is typical, c.f. fet structure.

Transit time is determined by layer thickness and doping: a simpler and higher-yield process. **Fig. 4** shows one structure of the HBT, with the epitaxial layer hierarchy grown upon the collector substrate. The base-collector junction is GaAs, but the main difference from the BJT lies in the base material, typically of n-type AlGaAs. The electron mobility in this material is greater than that of GaAs (and much greater than silicon) thereby increasing the frequency of operation.

Perhaps more importantly the base region of the HBT may be much more heavily doped than the emitter, thereby reducing base resistance and reducing the base-emitter capacitance. The wide bandgap of the emitter prevents hole injection from the base. Also, high current capability is possible as the whole of the emitter area conducts uniformly. Thus, all the desirable features of the Si BJT are translated to higher frequencies (4 to 40GHz at present).

The vertical process technology also circumvents the surface defect limitations of the fet resulting in superior phase noise performance. As an example, Class AB operation at 10GHz has produced in excess of 55% power added efficiency with greater than 10dB gain and a power output of greater than 300 mW.

Both mesfet and HBT devices are themselves compatible with monolithic technology, leading to integrated low noise or high power amplifiers with optimised performance.

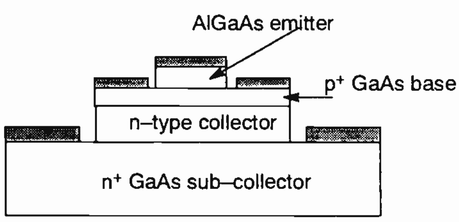


Fig. 4. Structure of the heteo-junction bipolar transistor. The key to its high frequency and low noise performance lies in the AlGaAs layer and the high doping possible of the base region.

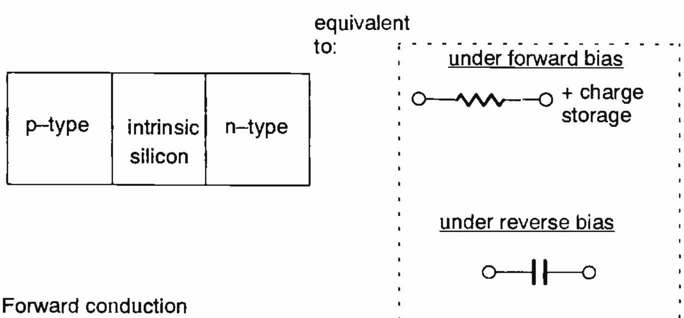


Fig. 5. Basic structure and equivalent circuit of the step recovery diode under different bias conditions. Unlike the normal p-i-n diode, there is no intermediate impedance state.

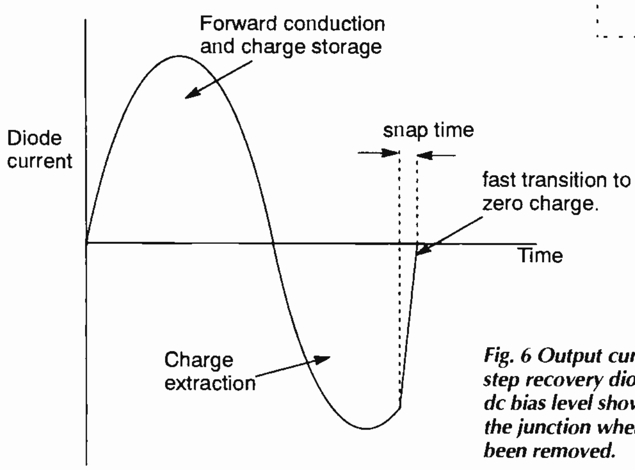


Fig. 6 Output current waveform of the step recovery diode superimposed on a dc bias level showing the snap action of the junction when all the charge has been removed.

Step recovery multiplication

Although largely replaced by fundamental oscillators, one method, used in instrumentation to generate marker frequencies and local oscillator signals, uses the step recovery diode (SRD or snap varactor). It requires simple circuit design and only a single frequency input; the output comprises pure harmonics of the input.

The efficiency of this process is proportional to $1/n$, where n is the harmonic number, whereas the normal varactor multiplier using its non-linear capacitance as the multiplying mechanism has an efficiency proportional to $1/n^2$. Thus, the comb generator is suited to high-order multiplication and can give outputs of 10's of milliwatts, or more, into the millimetre band at about 30GHz.

At frequencies below 10GHz, several watts of output are possible. Multiplication factors up to 10 are typical and efficiency, depending upon n and frequency, can exceed 65%.

The SRD itself is made from silicon and its doping profile is of the p-i-n form shown in **Fig. 5**, where the intrinsic layer is relatively thin and the exact grading of the p-i junction is crucial to performance.

Successful operation depends upon the charge storage properties of the diode and silicon is preferred because of its minority carrier lifetime. This is the time for which the carrier will remain in its free state before recombination: in Si this can be made to be similar to the period of the input rf cycle.

The diode of Fig. 5 has two extreme states under conditions of bias: forward bias will inject charge into the i-region and the SRD will appear as a low-value resistance. Reverse bias will remove charge leaving, essentially, a parallel-plate capacitance. However, for the duration of the carrier lifetime, there will be a transitory state of charge storage which is the basis of the SRD mechanism. **Fig. 6** shows the condition where the SRD is forward biased to just below its conduction threshold (0.6V >1.0V) and an input RF waveform is superimposed.

During the positive half cycle, the SRD is driven into conduction and the output current will essentially follow the rf input. The i-region fills with electrons and holes. As the input waveform goes negative, charge injection will cease: however, conduction will continue as the i-region charge can still be removed because it has been 'stored' due to the long carrier lifetime. There will come a point, though, in the negative cycle when all charge has been removed and the diode will change state to its high impedance (capacitive) state. The essence of the SRD principle is that this change of state occurs very rapidly: typically in tens of picoseconds and thus appears as a fast, transient spike.

Figure 7 shows a typical comb generator circuit using the SRD. The diode is inductively driven by the input waveform. In some high frequency designs the impulse inductor can be the actual package inductance. During the charge-extraction portion of the rf cycle, energy builds up in this inductance and then, when the diode snaps into its rapid impedance change, the energy in this transient shock excites the output circuit (usually a length of transmission line) into resonance. This produces a comb of harmonics of the input frequency each input cycle. The final output stage is often a bandpass filter, selecting a particular harmonic. The output frequency spectrum and stability matches that of the input source. If this is a low-noise, high quality oscillator, then so too will be the output. Input frequencies cannot be too low in frequency and typically range from 250 MHz to 10GHz, depending upon the required output range. ■

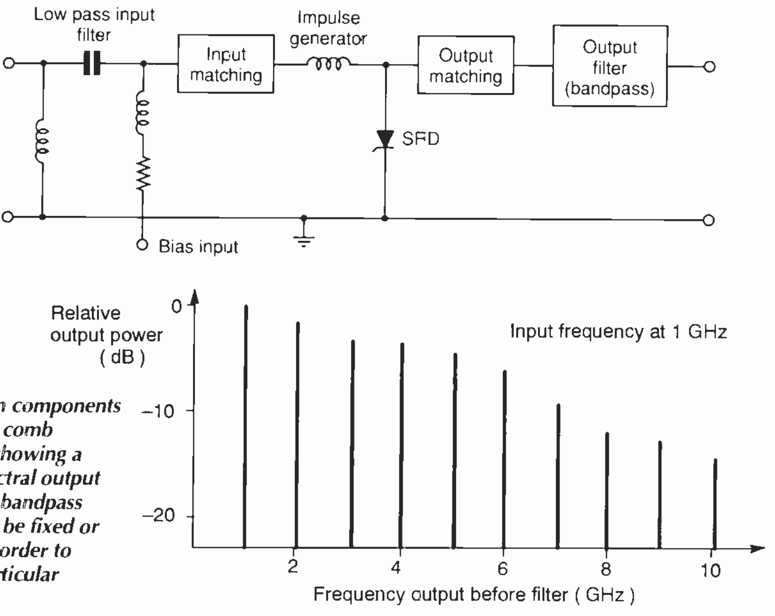


Fig. 7. Main components of the SRD comb generator showing a typical spectral output comb. The bandpass filter could be fixed or variable in order to select a particular harmonic.

Next month: circuits and methods for tuning and stabilising microwave sources.

The transformer ratio arm bridge

Every so often, it becomes worthwhile to reinvent the wheel. Ian Hickman speculates on a new type of instrumentation based on the age-old bridge measurement technique.

The bridge principle in electrical measurements has been extended far beyond Wheatstone's original application for the measurement of resistance at dc. It has been adapted for measuring inductance, capacitance and even frequency (e.g. Wien bridge).

Bridge methods have largely fallen into disuse, especially for rf measurements. The reason is that each measurement is taken at a spot frequency and involves adjusting two standards for balance. Investigating the variation of impedance or admittance of an unknown with frequency is tedious, involving a large number of spot measurements. It is so much easier to connect the unknown to a network analyser and take an s_{11} measurement covering the frequency range of interest. The answer can be viewed instantly as an angular vector plot versus frequency, or I and Q components, or as a Smith chart display, with the value corresponding to a marker at any desired spot frequency indicated on the screen as a numerical readout.

Often the unknown is not a simple two terminal impedance, but a two port pi network, as in **Fig. 1a**. Given three independent equations relating the three unknowns to three measured values, the various impedances or admittances can be calculated. So any three of the four s-parameter measurements in **Fig. 1b** should in theory suffice, permitting the evaluation of the three impedances.

In practice, if Z_1 and Z_3 are low and Z_2 high, the computation will involve the difference of two large quantities, magnifying any measurement errors and may lead to a large margin of error for the calculated value of Z_2. This is precisely where a particular type of rf bridge, the transformer-ratio-arm-bridge, scores over other methods. The trab was developed during WWII by Mr Gilbert Mayo of the BBC Research Department, and subsequently further developed and marketed by at least two companies in the U.K.

The transformer bridge

Figure 2a shows an rf trab in its simplest form. An rf test signal from a bridge source or signal generator is applied, via a stepdown transformer T_1, to a calibrated variable standard capacitor C_s and a conductance standard G_s which is effectively variable from $1/R_1$ down to zero by means of the non-inductive calibrated variable resistor R_v. The other ends of these two components are connected to one end of a centre-tapped symmetrical winding on T_2. The unknown is connected between the output of T_1 and the other end of the centre tapped winding on T_2. An output winding on T_2 is connected to a bridge detector, which can be any radio receiver covering the band of interest.

In use, C_s and G_s are adjusted until there is no detectable output at the receiver. Balance occurs when the parallel components C_x and G_x of the unknown are equal to the capacitance setting of C_s (marked on its dial) and the effective value of G_s (which is marked on the dial of R_v). Balance occurs as a result of the current flowing via the standards through one half of the balanced symmetrical winding of T_2 equalling the current flowing via the unknown through the other half of the winding. As the number of turns on each half winding are the same, there is no net magnetising force, and consequently no resultant flux on the core. There is thus no voltage induced in the output winding, but equally, there is no voltage appearing across either half of the centre-tapped winding either. Thus T_2 acts as a virtual earth, the significance of which will appear in a moment.

Note that a trab inherently measures Z_2 in terms of an admittance Y_2, i.e. equivalent components G_x in parallel with B_x. Given the known test frequency, these can be converted to the equivalent series components R_x and X_x using the formulae in **Fig. 2b**, or using a Smith Chart[1].

As shown in Fig. 2a, the range of the unknown that the bridge can handle is limited to C_x not exceeding the maximum value of C_s, and G_x not higher than $1/R_1$, with no capability at all for measuring either inductance, or negative conductance. This capability can be provided by the modified arrangement of Fig. 2b. Here, the terminals to which the unknown are connected are mounted upon two substantial blocks of metal, firmly bolted together with an insulating film between, to form a capacitance equal to $C_{s\,max}/2$. The dial of C_s is now calibrated in positive and negative values of capacitance, with zero occurring where C_s is set at $C_{s\,max}/2$; this arrangement was employed in the Wayne Kerr trabs.

Similarly, a resistor of value $2R_1$ can be added in parallel with the unknown terminals, allowing for negative values of conductance and permitting measurements on active devices. The shunt components of resistance (or conductance) and capacitance of the unknown could be read directly from the dials, at any test frequency. Inductance, on the other hand, was read as a negative capacitance. Knowing the test frequency enables the susceptance of the negative capacitance, and hence of the unknown inductive component, to be calculated, from

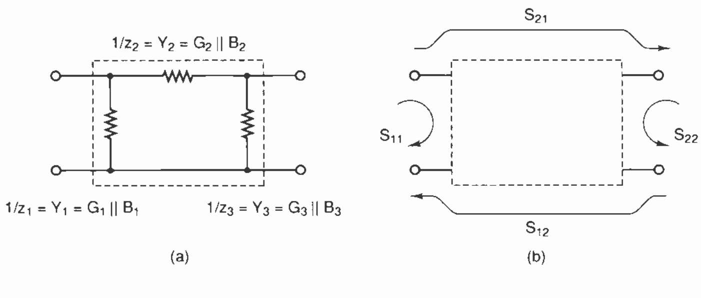


Fig. 1a) General two port pi network expressed in parallel components.
b) s-parameter measurement of two port pi network

which the value of inductance itself was readily derived.

Although shown with the source connected to T_1 and the detector to T_2, a trab can be used with the source and detector interchanged, where convenient.

Extended possibilities

Fig. 3 shows how the trab is inherently adapted for three terminal measurements, permitting the accurate measurement of Y_2 whatever the values of Y_1 and Y_3, by virtue of the bridge's neutral connection. It can be seen that admittance Y_1 shunts the source whilst Y_3 shunts the detector (T_2 is a virtual earth). Thus while Y_1 in particular may reduce bridge sensitivity slightly, neither Y_1 nor Y_3 affects the accuracy of the measurement of Y_2.

The operation of adaptors which extend the range of a bridge is of interest not only from the electrical point of view, but also because they illustrate the usefulness of the Binomial Theorem, which should really be called "The Engineer's Friend". The *LE305*, a well specified but low frequency bridge from Hatfield Instruments consisted of two non-inductive 100Ω resistors in a massive metal housing which are interconnected with the common and neutral terminals of the bridge and, by means of a substantial plug, with one of the multiplier terminals associated with T_2. The unknown to be measured was connected between the junction of the two resistors and the bridge neutral terminal, see **Fig. 4**. If Z_x is a resistance of 1Ω or less, the current i_1 will be determined virtually solely by R_2 (and the bridge drive voltage at the common terminal). The volt drop e will thus be directly proportional to Z_x, and hence so will the current i_2 delivered to T_2. If Z_x is zero, i_2 will be zero and bridge balance will be achieved.

As Z_x increases, so R_2 must be advanced, so its dial in fact now reads resistance directly instead of conductance. Similarly, if Z_x is an inductive reactance, the voltage across it and hence i_2 will both be *leading* and must be balanced by a capacitive current via C_s fed to the opposite side of T_2.

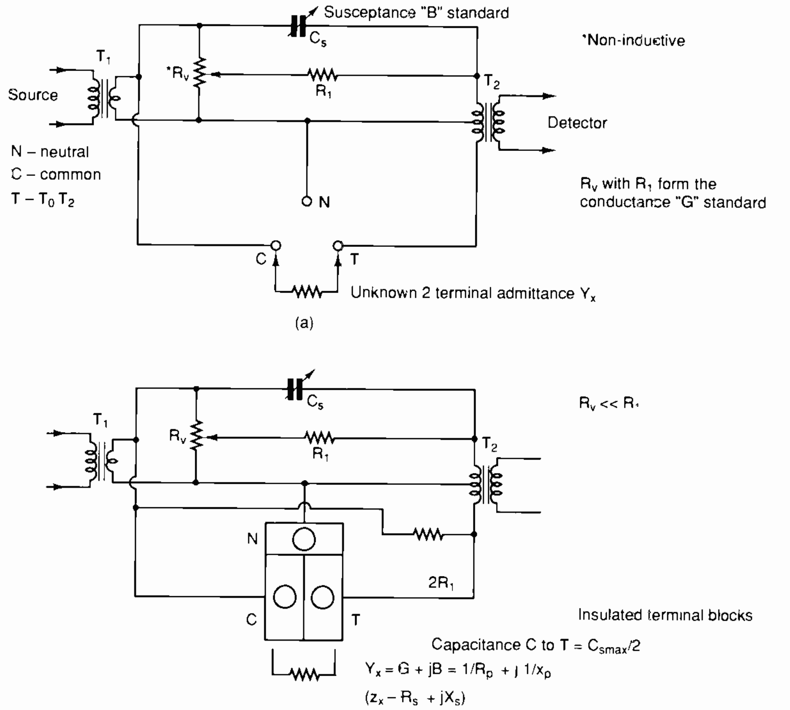
For an inductive Z_x, as the frequency increases, so will e in Fig. 4, and hence also i_2. Thus there will be no change in the setting of C_s for balance, so that with the low impedance adaptor in use, it is now *inductive* unknowns that become independent of frequency, even though they are measured by comparison with a capacitive standard. Furthermore, the unknown is now measured as series components r+jx, rather than shunt components G+jB. If Z_x is resistive, then the error due to the finite voltage e subtracting from the available bridge drive voltage reaches 2% for Z_x = 1Ω. (The 1% *increase* in mesh resistance 100Ω+1Ω results in a 1% *decrease* in i_1: furthermore, the 1Ω unknown is shunted by the second 100Ω resistor, so e is reduced overall by 2%, per the binomial expansion, ignoring second order terms.)

The reactive component of Z_x may be up to 3Ω, since the component of e due to it subtracts from the bridge drive voltage rms-wise, rather than directly.

Automated trab?

A bit of speculation. A self-balancing trab – if such a thing existed – could directly measure the parallel components of susceptance of the series element of a pi network. The drive signal could then be swept, to give an impedance versus frequency display, not unlike a network analyser. For instance, the variable conductance standard formed by R_1 and R_v in Fig. 2 could be replaced by a four quadrant multiplier, with the rf applied to the X port and the control signal to the Y port. The output would thus be adjustable in both amplitude and sign, avoiding the need for selecting one side of T_2 or the other. The bridge output from T_2 would be synchronously detected and the resultant fed back to the Y port of the multiplier to automatically achieve a balance with the conductance of the unknown.

The degree of balance achieved would depend upon the loop gain. The variable susceptance could be obtained from another multiplier instead of C_s, by feeding its X port from a quadrature version of the drive

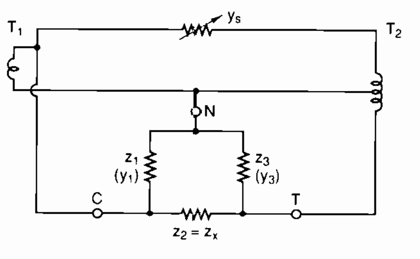


If the equivalent series components of the unknown are required, then

$$R_s = \frac{R_p X_p^2}{R_p^2 + X_p^2} \text{ and } X_s = \frac{R_p X_p}{R_p^2 + X_p^2}$$

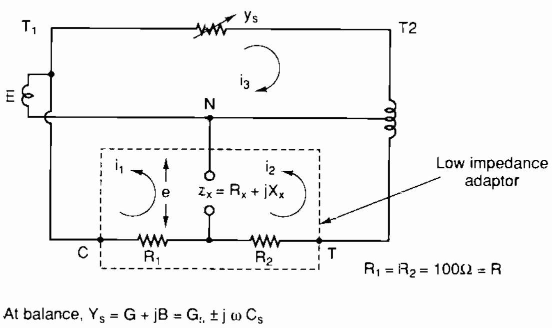
(b)

Fig. 2a) In its simplest form, a trab measures only capacitance and conductance (resistance).
b) With modifications, it can also measure negative capacitances (inductance) and negative conductances. Additional circuitry (not shown) is required to permit balancing of the bridge with C_s and R_v set to zero, before connecting the unknown.



$1/z_x = Y_x = Y_s$ (at balance)

Fig. 3. Circuit showing how a trab is inherently perfectly adapted for three terminal measurements. Admittance Y_1 shunts the source whilst Y_3 shunts the detector (T_2 is a virtual earth). Thus while they may reduce bridge sensitivity slightly, neither affects the accuracy of the measurement of Y_2.



At balance, $Y_s = G + jB = G_s \pm j\,\omega\,C_s$

Then $L_x = C_s R_2$ (independant of ω)
or $C_x = 1/(R^2 \omega^2 C_s)$
and $R_x = G_s R^2$

Fig. 4. Showing how a low impedance adaptor not only extends the measurement range of a trab, but also turns conductance into resistance, and capacitive susceptance into inductive reactance and vice versa.

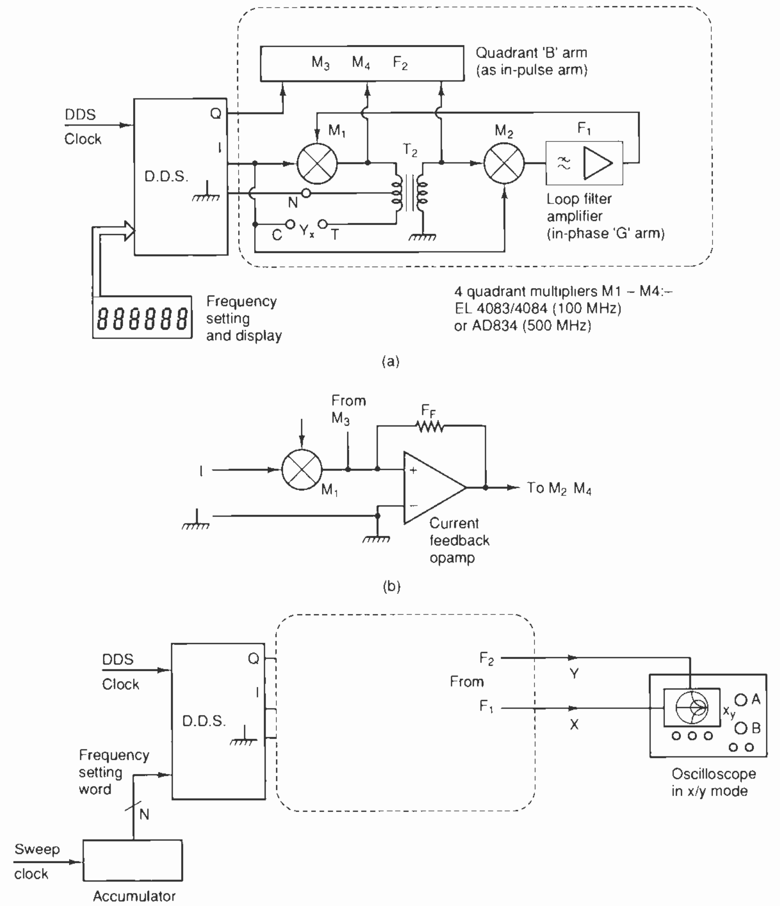


Fig. 5a) Trab with autobalance facilities
b) As the multipliers can implement rf outputs of both polarity, i.e. can act as both negative and positive conductance and susceptance standards, T_2 is redundant, and can be replaced by an op-amp virtual earth.
c) As a), with sweep facility and Smith Chart display.

signal. This requirement is easily met since sine and cosine outputs are available as standard from many direct digital synthesizer chips.

The arrangement might look something like **Fig. 5a**, where the synchronous detectors could be multipliers identical to those used as the G and B standards, fed through lengths of line to compensate the path-length through the bridge. As the multipliers can implement rf outputs of both polarity, i.e. can act as both negative and positive conductance and susceptance standards, T_2 is in fact redundant. It can be replaced by an current feedback op-amp virtual earth as in **Fig. 5b**: conveniently the multipliers provide current outputs. **Fig. 5c** shows how a swept measurement, with Smith Chart display of the unknown as a function of frequency, might be organised.

The only disadvantage of this 'active trab' arrangement is that a very low output impedance buffer is needed to drive the unknown. Any reduction of the I drive voltage due to loading by Z_1 (or indeed Z_2) will affect the conductance standard equally and thus cause no error with this arrangement, it will not correspondingly affect the susceptance standard arm.■

Reference

1. Ian Hickman, *Design Brief, EW+WW* Oct 93, p972.

Electronics help police to see the genetic picture

Britain's police are shortly set to make substantially more use of DNA fingerprinting in human identification. Tom Ivall explains how instrumentation is developing to make the process more practical.

Under the latest Criminal Justice Bill, soon to become law, police will be able to take DNA samples, without consent, from anyone arrested for a recordable offence. The resulting molecular 'fingerprints' will be stored in a national database.

DNA testing for forensic and other identification purposes has already become well established since professor Alec Jeffreys of Leicester University invented the technique in 1985. Now public and commercial laboratories provide it as a service – though in the UK, Cellmark Diagnostics, part of Zeneca, is the only commercial lab.

The US government meanwhile, takes and holds samples from all its military personnel. Legal challenges have been made to the technique. But its validity is now accepted, because every individual has certain inherited patterns within the DNA molecule which are, for practical purposes, unique – better than 1 in 10^{12} – except for identical twins.

Only the methodology is being refined, with electronics and computers automating some of the processes.

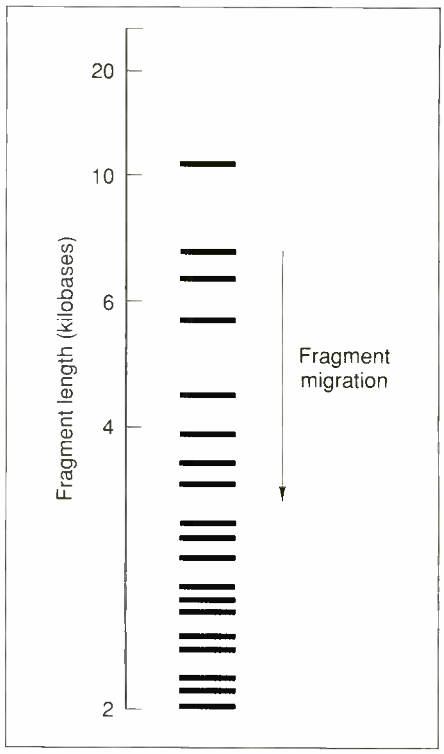
Obtaining a DNA fingerprint

The technology for DNA fingerprinting is based on electrophoresis. In addition, optical scanning, digitisation and data analysis are used to map and list both the numbers and sequences of the highly significant nitrogenous bases in the DNA molecule (see box).

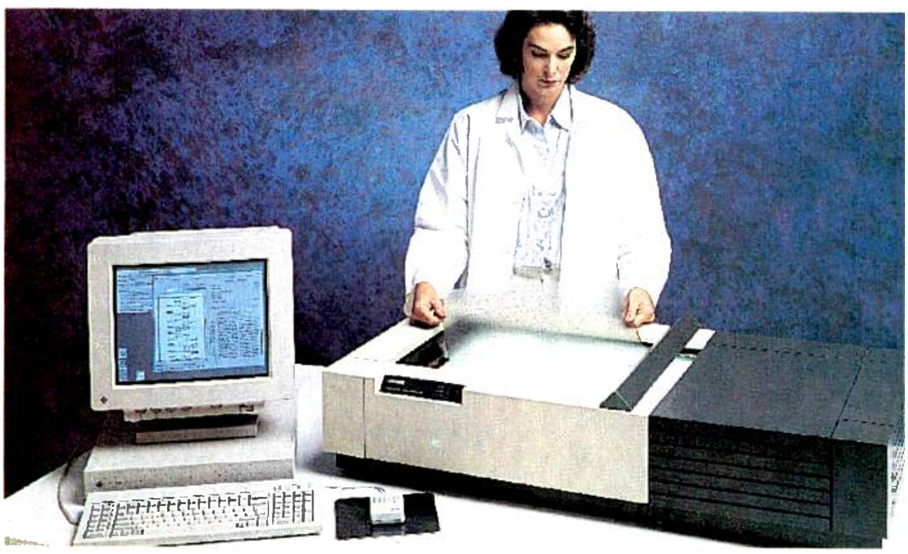
DNA fingerprints are images derived from a molecular pattern in the body's cells. The patterns are composed of short sequences of base molecules, typically 20 bases long, which keep on repeating consecutively along a DNA strand at particular locations.

Two factors make the pattern specific to an individual: the number of repeats at a given location and hence the overall length of the repeating section, normally in the range 4000 to 20,000 base pairs; and the particular locations on the DNA molecule at which these repetitions occur.

Both features vary enormously from person to person and are inherited – so are useful for paternity investigations – but are not part of the functional genes.



DNA fingerprint, resembling a bar code, is an autoradiograph taken from an electrophoresis separation. Each dark band represents probe-selected DNA fragments of a particular length, measured in number of base molecules. The pattern is specific to an individual.



Automated scanner made by Bio-Rad Laboratories. The GS Gene Reader scans autoradiograph films for DNA sequence analysis. Digitised images and a corresponding list of base letters are displayed on the workstation screen.

Key to detection

Detecting the molecular pattern is a little like probing an unknown lock with a known key to see if it fits. The probe is a short length of single-strand DNA with a known sequence of bases, derived by cloning from a human source. It will fit any complementary sequence on a single strand of the DNA being tested, so 'recognising' it in a process called hybridisation. The repeating sections detected are subsequently translated into an image by sorting the repeating sections into their different lengths. Then a purified solution of the DNA sample – commonly obtained from blood, saliva, hair roots or semen – is treated with one or more enzymes to cut the double-helix molecules at points on each side of the repeating section. The enzymes recognise particular, short, base-pair sequences of the DNA and bind to them, slicing the duplex DNA at or near these positions.

The resulting assortment of fragments is placed in an electrophoresis cell and separated according to length (see box). A continuous distribution of fragment lengths is the outcome, from about 200 base pairs (bp) to 20kbp or more, along a 'lane' in the electrophoresis gel. Only those fragments containing the wanted repeat sections have to be identified.

Sodium hydroxide is used to split the DNA into single strands, then the gel is dried and its distribution of fragments is transferred to a solid support – normally a nylon membrane – by a process called Southern blotting.

The membrane is exposed for several hours to a solution containing the probe, which is radioactively labelled. Binding occurs between the probe and only those lengths of separated-DNA-fragments which include the repeating sections.

X-ray film exposed to the membrane produces an autoradiograph (arg), and detected fragment lengths with their radioactive labelling stand out as a pattern of dark bands on the developed film. This is the fingerprint.

The process does not identify the actual locations of the repeating sections on the molecule. A different method detects repeat lengths at single, chosen molecule locations and produces only two bands on the arg. The first, multi-locus form is used mainly for paternity tests while the second, single-locus type, is used mainly for forensic work.

Structure of human DNA

DNA, or deoxyribonucleic acid, exists in most cells of living organisms – from single-celled bacteria to multicellular animals. In the nuclei of human cells, with proteins, it forms the substance of our 46 chromosomes, and carries the genes for the copying process when cells divide and proliferate to grow and repair the body.

Its form is polymeric, consisting of a long chain of smaller molecules of four distinct types called nucleotides. In a single human cell the total length is about 2m. Each nucleotide comprises three yet smaller molecules: a sugar ring, a phosphate group and a nitrogenous base formed from carbon, hydrogen, oxygen, nitrogen and phosphorus.

The four nucleotides have different nitrogenous bases – adenine (A), thymine (T), cytosine (C) and guanine (G) – and it is these that form the 'symbols' of the genetic and DNA fingerprinting information through their changing sequence along the structure.

The DNA molecule, the famous double helix can be likened to a twisted rope ladder. The two sides of the ladder are made up of alternating, covalently-bonded sugar and phosphate molecules from the nucleotides: the rungs are pairs of bases.

There are about 6×10^9 of these base-pairs in the DNA of a human cell. A complete 360° turn of the double helix is about ten base-pairs and 3.4nm long.

Base A always pairs with base T, and C with G, so the two halves of the ladder – single strands of the molecule – are complementary. The complementary bases are held together by relatively weak hydrogen bonds, allowing easy chemical separation of the two strands.

A single existing strand of DNA forms a template for a new, complementary strand to be synthesised onto it. This template also allows short segments of single-strand DNA from other sources to chemically bind to parts of it (hybridisation).

Called probes, such external segments with known base sequences are used in DNA fingerprinting to detect patterns of base sequences at different locations on the human molecule.

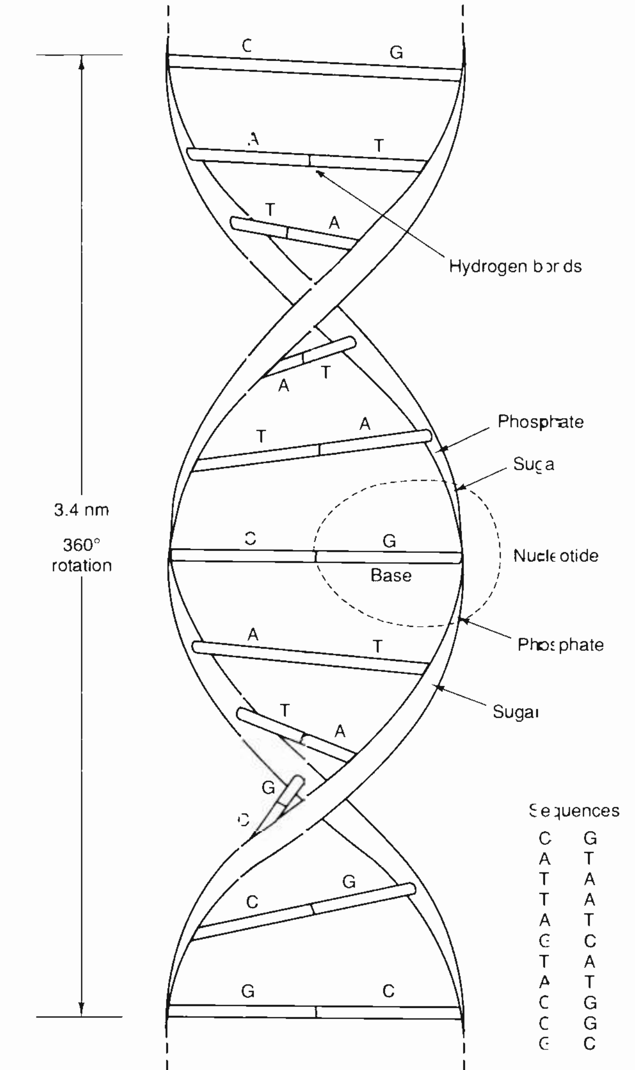


Fig. 2. Twisted ladder model of part of double helix DNA molecule, showing A-T and C-G pairing of bases in complementary strands.

Electronic interpretation

Electronic scanning of the electrophoresis bands speeds up the interpretation process, reduces subjective errors in reading and produces data suitable for computer processing and analysis.

Scanning can be used directly on the fingerprints themselves but also, with a different electrophoresis separation, in sequence analysis, to find the A-G-C-T base sequences (see box) in DNA samples and probes.

For fingerprint args the scanning system is essentially a digital densitometer. A video camera or linear ccd array scans the image and the resulting signals are digitised. One FBI system uses a quantisation of 512 by 480 pixels and 256 grey levels.

ARG bands are fuzzy and of varying density (not sharp as in the figure) so the information obtained is a profile of density against distance along the arg. From the peaks in this profile an associated pc derives the positions of the bands and calculates the corresponding DNA fragment lengths.

Software can also locate the lane boundaries, correct image distortion, control the digitisation, perform statistical analysis, display data and graphics on the pc screen and finally control disk storage and/or transfer of data to a database.

Sequence analysis

DNA sequence analysis sometimes requires scanning systems to work on arg films and sometimes directly on electrophoresis gels. The biochemical reactions for determining base sequence involve four sets of copies of the DNA, each a gradually lengthening fragment ending on one of A-G-C-T. A four-well electrophoresis cell separates the fragments on the same gel, producing a succession of bands in each lane, A bands in one, G in another, C and T. From the relative distances of these bands along their four lanes, a continuing sequence of bases can be read off such as GTATGCCAT...

Electrophoresis bands are then scanned, translated into letters and then data transferred into a computer for processing and storage. The reader uses a 1728-element ccd linear array to scan arg films with a resolution of 50μm. Variable speed scanning improves resolution where the bands are closely spaced, economising on computer memory when high resolution is not needed.

CCD signals are digitised and transferred to a Sun *Sparc* workstation where pattern recognition software identifies lanes, corrects errors and translates into base letters. The workstation screen displays a four-lane band pattern and alongside it a list of the corresponding sequence of base letters. Reading and processing at a rate of about 50 bases/min are possible.

Sequence-analysis instruments aim at high accuracy as well as speed, eliminating the arg stage and possible reading errors by sensing the electrophoresis gel directly. They also work in real time by detecting fragment lengths and translating them into base letters while electrophoresis is proceeding.

Reaction products are combined and electrophoresed in a single gel lane, sandwiched between glass plates, and each is labelled with a fluorescent material of distinct emission wavelength.

As the DNA fragments thus identified approach the end of the 25cm lane they are illuminated by an argon ion laser beam, exciting the fluorophores so they emit light of different colours, a wavelength for each DNA base. A lens collects the light and focuses it through an optical bandpass filter, one for each colour, onto a photomultiplier tube. As electrophoresis proceeds, the pmt signal, a succession of peaks, is a-to-d converted and the resulting data is processed and translated into a sequence of DNA base letters. The complete optical system mechanically traverses up to 24 such single lanes, scanning them in succession. ■

Electrophoresis

Electrophoresis separates colloidal particles by size. Here the particles are different length fragments of DNA and are made to migrate through an electrolyte solution by applying an electric field.

Each fragment carries a net charge, resulting from its own molecular composition and the ions in the electrolyte. The moving force on each fragment is proportional to the product of the net charge and the electric field strength.

Rate of migration, the electrophoretic mobility, is measured as velocity per unit field strength. It depends partly on a potential difference, determined by the-electrolyte, between the outside of the ion layer bonded to the fragment and parts of the electrolyte some distance away.

Other variables affecting it are viscosity and dielectric strength of the solution. Mobility is directly proportional to the net charge on the fragment and inversely proportional to its size – smaller fragments travel faster.

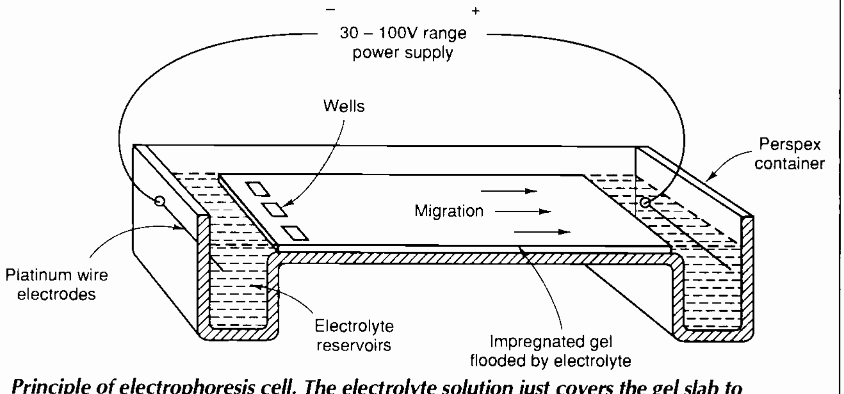
But free liquid does not allow stable separation of fragments. So the electrolyte is supported in a porous medium, a slab of gel impregnated with the liquid. Fragments migrate through the gel pores when a voltage is applied across the slab through electrolyte reservoirs. Resolution of size separation is increased, because the gel acts like a mechanical sieve, allowing the shorter fragments through easily but retarding the longer ones.

When the electric field is switched off the DNA fragments end up distributed according to length at various distances along the gel.

In commercial systems the gel, made from agarose or polyacrylamide, is about 1mm thick and mounted either horizontally or vertically. At one end is a well, into which the purified DNA solution is placed. The electrolyte is chosen for its hydrogen ion concentration (pH) and acts as a buffer solution, to keep the pH relatively constant. DNA fragments have a net negative charge and are therefore attracted towards the anode electrode.

Voltages range from 30V to 100V for DNA fingerprinting, or up to about 3000V for DNA sequence analysis. Time required for separation is normally several hours. Using a number of wells in the gel enables different DNA samples to be electrophoresed simultaneously for comparison, forming parallel electrophoresis lanes in the gel.

Resolution of larger DNA fragments (above 100kb) can be improved by periodically reversing the electric field and hence direction of migration. Typically the cyclic forward:reverse time ratio is 3:1 and times can be anything from seconds to hours. Programmable pulse generators are used for this purpose.



Principle of electrophoresis cell. The electrolyte solution just covers the gel slab to maintain a conductive path through it.

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HP180TR, HP182T mainframes £300-£500.
Fluke 8506A thermal RMS digital multimeter. £400.
Philips panoramic receiver type PM7900 – 1 to 20GHz – £400.
Marconi 6700A sweep oscillator + 6730A – 1 to 2GHz – £500.
HP8505A network ANZ + 8503A S parameter test set + **8501A** normalizer – £4k.
Racal/Dana VLF frequency standard equipment. Tracer receiver type 900A + difference meter type 527E + rubidium standard type 9475 – £2750.
HP signal generators type 626 – 628 – frequency 10GHz – 21GHz.
HP 432A – 435A or B – 436A – power meters + powerheads – Mc/s – 40GHz – £200-£1000.
Bradley oscilloscope calibrator type 192 – £600.
Barr & Stroud variable filter EF3 0.1Hz – 100Kc/s + high pass + low pass – £150.
Marconi TF2370 spectrum ANZ – 110Mc/s – £900.
Marconi TF2370 spectrum ANZ + TK2375 FX extender 1250Mc/s + 1st gen – £1.5k.
HP8614A signal generator 800Mc/s – 2.4GHz, new colour £400.
HP8616A signal gen 1.8GHz – 4.5GHz, new colour £400.
HP 3325A syn function gen 20Mc/s – £1500.
HP 3336A or B syn level generator – £500-£600.
HP 3586B or C selective level meter – £750-£1000.
HP 3575A gain phase meter 1Hz – 13Mc/s – £400.
HP 8671A syn microwave 2 – 6.2GHz – £2k.
HP 8683D S/G microwave 2.3 – 13GHz – opt 001 – 003 – £4.5k.
HP 8660 A-B-C syn S/G. AM + FM + 10Kc/s to 110Mc/s PI – 1Mc/s to 1300Mc/s – 1Mc/s to 2500Mc/s – £750-£2800.
HP 8640B S/G AM-FM 512Mc/s or 1024Mc/s. Opt 001 or 002 or 003 – £800-£1250.
HP 8656A S/G AM-FM 0.1 – 990Mc/s – £1500.
HP 8622B Sweep PI – 01 – 2.4GHz + ATT – £1750.
HP 8629A Sweep PI – 2 – 18GHz – £1000.
HP 86290B Sweep PI – 2 – 18GHz – £1250.
HP 86 Series PI's in stock – splitband from 10Mc/s – 18.6GHz – £250-£1k.
HP 8620C Mainframe – £250. IEEE – £500.
HP 8615A Programmable signal source – 1MHz – 50Mc/s – opt 002 – £1k.
HP 8601A Sweep generator .1 – 110Mc/s – £300.
HP 4261A LCR meter + 16038A test leads – £400.
HP 4271B LCR meter 1MHz digital meter + 16063A test adaptor – £850.
HP 4342A Q meter 22kHz – 70Mc/s 16462A + qty of 10 inductors – £850.
HP 3488A HP – IB switch control unit – £500 + control modules various – £175 each.
HP 3561A Dynamic signal ANZ – £3k.
HP 8160A 50Mc/s programmable pulse generator – £1400.
HP 853A MF ANZ + 8558B – 0.1 – 1500Mc/s – £2500.
HP 8349A Microwave Amp 2 – 20GHz Solid state – £1500
HP 3585A Analyser 20Hz – 40Mc/s – £4k.
HP 8569B Analyser .01 – 22GHz – £5k.
HP 3580A Analyser 5Hz – 50kHz – £1k.
HP 1980B Oscilloscope measurement system – £600.
HP 3455A Digital voltmeter – £500.
HP 3437A System voltmeter – £300.
HP 3581C Selective voltmeter – £500.
HP 5370A Universal time interval counter – £450.
HP 5335A Universal counter – 200Mc/s – £500.
HP 5328A Universal counter – 500Mc/s – £250.
HP 6034A System power supply – 0 – 60V – 0 – 10 amps – £500.
HP 3960A 3964A Instrumentation tape recorders – £300-£500.
HP 5150A Thermal printer – £250.
HP 1645A Data error analyser – £150.
HP 4437A Attenuator – £150.
HP 3717A 70Mc/s modulator – £400.
HP 3710A – 3715A – 3716A – 3702B – 3703B – 3705A – 3711A – 3791B – 3712A – 3793B microwave link analyser – P.O.R.
HP 3730A+B RF down converter – P.O.R.
HP 3552A Transmission test set – £400.
HP 3763A Error detector – £500.
HP 3764A Digital transmission analyser – £600.
HP 3770A Amp delay distortion analyser – £400.
HP 3780A Pattern generator detector – £400.
HP 3781A Pattern generator – £400.
HP 3781B Pattern generator (bell) – £300.
HP 3782A Error detector – £400.
HP 3782B Error detector (bell) – £300.
HP 3785A Jitter generator + receiver – £750-£1k.
HP 8006A Word generator – £100-£150.
HP 8016A Word generator – £250.
HP 8170A Logic pattern generator – £500.
HP 59401A Bus system analyser – £350.
HP 59500A Multiprogrammer HP – IB – £300.
Philips PM5390 RF syn – 0.1 – 1GHz – AM + FM – £1250.
Philips PM5519 Colour T.V. pattern generator – £250.
S.A. Spectral Dynamics SD345 spectrascope 111 – LF ANZ – £2500.
Tektronix R7912 Transient waveform digitizer – programmable – £400.
Tektronix 496 Analyser 1kHz – 1.8GHz – £3.5k.
Tektronix TR503 + TM503 tracking generator 0.1 – 1.8GHz – £1k – or TR502.
Tektronix 576 Curve tracer + adaptors – £900.
Tektronix 577 Curve tracer + adaptors – £900.
Tektronix 1502/1503 TDR cable test set – £1000.
Tektronix 7L5 LF analyser – 0 – 5Mc/s – £800. OPT 25 – £1000.
Tektronix AM503 Current probe + TM501 m/frame – £1000.
Tektronix SC501 – SC502 – SC503 – SC504 oscilloscopes – £75-£350.
Tektronix 465 – 465B – 475 – 2213A – 2215 – 2225 – 2235 – 2245 – 2246 – £250-£1000.
Kikusui 100Mc/s Oscilloscope COS6100M – £350.
Farnell PSG520 Signal generator – £400.
Nicolet 3091 LF oscilloscope – £1000.
Racal 1991 – 1992 – 1988 – 1300Mc/s counters – £500-£900.
Tek 2445 150Mc/s oscilloscope – £1400.
Fluke 80K-40 High voltage probe in case – BN – £100.
Racal Recorders – Store 4 – 4D – 7 – 14 channels in stock – £250 – £500.
Racal Store Horse Recorder & control – £400-£750 Tested.
EIP 545 microwave18GHz counter – £1200.
Fluke 510A AC ref standard – 400Hz – £200
Fluke 355A DC voltage standard – £300.
Schlumberger 5229 Oscilloscope – 500Mc/s – £500.
Solartron 1170 FX response ANZ – LED dislay – £280.
Wiltron 610D Sweep Generator + 6124C PI – **4** – 8GHz – £400.
Wiltron 610D Sweep Generator + 61084D PI – 1Mc/s – 1500Mc/s – £500.
Time Electronics 9814 Voltage calibrator – £750.
Time Electronics 9811 Programmable resistance – £600.
Time Electronics 2004 D.C. voltage standard – £1000.
HP 8699B Sweep PI YIG oscillator .01 – 4GHz – £300. 8690B MF – £250. Both £500.
Schlumberger 1250 Frequency response ANZ – £2500.
Dummy Loads & power att up to 2.5 kilowatts FX up to 18GHz – microwave parts new and ex equipt – relays – attenuators – switches – waveguides – Yigs – SMA – APC7 plugs – adaptors, etc.
B&K Items in stock – ask for list.
W&G Items in stock – ask for list.
Power Supplies Heavy duty + bench in stock – Farnell – HP – Weir – Thurlby – Racal etc. Ask for list.

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ACTIVE

Asics

Cost-effective asics. Although small asics are out of fashion, they can be much cheaper to develop. Accordingly, Sharp has a special deal on asics with up to 2000 gates, which will handle glue logic, pals and eplds in one chip. Minimum quantity is 5000 pieces and the typical once-only cost is around £10000, design time being 1-4 weeks. Hero Electronics Ltd. Tel., 0525 4055015; fax, 0525 402383.

A-to-d and d-to-a converters

Eighth-order 100kHz lp filter. *LTC1066-1* by Linear Technology is an eighth-order, pin-selectable, elliptic or linear-phase low-pass filter that offers up to 100kHz cut-off and 14-bit dc gain linearity. It has a 94dB signal:noise ratio and is suitable for use as an anti-aliasing or smoothing filter in 12 or 14-bit data acquisition. The device is clock-tunable from 10Hz to 100kHz; clock:cut-off frequency ratio is 50 and the input is sampled twice to reduce the risk of aliasing. Ripple in the pass band is ±0.15dB and there is 80dB of stop band attenuation at 2.3 times cut-off. Linear Technology (UK) Ltd. Tel., 0276 677676; fax, 0276 64851.

Delta-sigma modulator. Crystal Semiconductor offers the *CS5321* 24-bit, fourth-order, 256-times-oversampled delta-sigma modulator for seismic and general scientific measurement. Its dynamic range is 123dB with a –118dB distortion level at 256×, and the device dissipates 70mW per channel – modulator and filter. A switched-capacitor architecture minimises the effects of clock jitter without the use of vcos and plls. Crystal Semiconductor Corporation. Tel., 0101 512 442 7555; fax, 0101 5512 445 7581.

Discrete active devices

SM power mosfets. Motorola's *MiniMOS* surface-mounted, high-density, n- and p-channel and complementary power mosfets feature 200mΩ or less on-resistance, achieved by laying down six million cells per square inch. The body diode has a shorter and softer recovery time and less stored charge, generating less noise during switching when compared with previous mosfets. They are rated at 50, 30, 20 and 12V breakdown and have a gate/source voltage of 2.7V, at which voltage on-resistance is constant. Motorola Ltd. Tel., 0908 614614; fax, 0908 618650.

Linear integrated circuits

100MHz video switches. A series of two-channel, 100MHz video buffers from Maxim, the *MAX463/4/5/6*, are single-chip devices containing video switches that switch between two rgb sources or rgb+sync. in less than 20ns, directly driving ±2.5V into 75Ω back-terminated cables. *MAX463/464* buffers have fixed unity gain, the others a fixed gain of 2 for 75Ω back-terminated use. A shutdown mode reduces current and also tri-states the output to allow multiple switches to be in parallel. Cross-talk is better than 60dB. Maxim Integrated Products UK Ltd. Tel., 0734 845255; fax, 0734 843863.

First phemt amplifier. H-P says its *MGA-86576* is the first commercial low-noise amplifier to use a GaAs phemt, which is to say pseudomorphic high electron mobility transistor. Although similar in concept to silicon microwave ICs, the new device extends frequency to over 8GHz and reduces a front-end component count from two mesfets and 20 components to one *MGA-86576* and three components. Noise figure is 1.6dB, 4GHz gain 23dB. Hewlett-Packard Ltd. Tel. 0344 362277; fax, 0344 362269.

Subscriber loop circuits. New ICs from AT&T for subscriber loop use include the *T7256*, which offers all the functions of a Basic Rate ISDN termination, and a family of subscriber loop interface circuits (*ATTL7554/7561/7564*) form a versatile solution for short-loop analogue use. The four slics provide independent adjustment of open-loop voltage, feed resistance and current limit and the ISDN interface performs all network termination functions without a microprocessor. AT&T Microelectronics. Tel., 0732 742999; fax, 0732 741221.

V.32bis chipset. The *CL-MD1414BA* three-chip set from Cirrus Logic is claimed to be the first to offer a high-performance data, fax and voice modem and a sound card, needing only an external 32kbyte sram. All popular standards are supported, including V.32bis, CCITT v 32, V.22bis, Bell 212A and 103 for data and CCITT V.17, V.29, V.27ter and V.21 ch2 for fax. Users may dial from a keyboard and use a headset for voice. Cirrus Logic Inc. Tel ,(USA) 0101 510 623 8300; fax, 0101 510 226 2240.

'Fastest' op-amp. Analog claims the industry speed record for its *AD8001* 800MHz op-amp, which achieves this performance on 5mA from ±5V. It is meant mainly for video application providing 0.1dB gain flatness to 100MHz, 0.01% differential gain error at a gain of 2 into 150Ω, 1200V/μs slewing and a 10ns settling time for a 2V step to within 0.1%. Worst harmonic component at 20MHz is at –60dB and voltage noise at 10kHz is 1.8nV/√Hz. The device will drive up to six 75Ω cables. Analog Devices Ltd. Tel, 0932 253320; fax, 0932 247401.

Zero-drift op-amp. The *LTC1152* low-power op-amp by Linear takes rail-to-rail input and produces rail-to-rail output swings, even into heavy loads. It is unity-gain stable into 2000pF with no extra components and one extra C enables it to drive unlimited capacitive loads. Offset voltage is 1μV, offset drift 10nV/°C, cmrr and power supply rejection are 130dB and 120dB and open-loop gain 130dB. Low frequency noise is kept to 2μV pk-pk and GB product is 1MHz. Linear Technology (UK) Ltd. Tel., 0276 677676; fax, 0276 64851.

Gilbert-cell array. Harris's *HFA3101* is a low-cost, silicon Gilbert-cell array intended in the main for rf mixing and amplification use up to 2.5GHz. Power gain-bandwidth product is 5GHz and the transistor f_T is 10GHz. Harris claims for it low cost, easier design and lower power consumption than discrete components and GaAs chips. Transistor noise figure is 2.5dB into 50Ω. Spice models and rf-specific scattering parameters are available. Harris Semiconductor UK. Tel., 0276 686886; fax, 0276 682323.

Logic building blocks

Fast bus switch. Cypress's *CYBUS3384* bus switch is a Fast Cmos Technology (FCT) bus switch for bidirectional data transfer between multiple buses or between 5V and 3.3V devices. Propagation delay is less than 250ns. FCT devices are also available as FCT-T types with ttl output, or as FCT2-T with 25Ω terminating resistors to further reduce ground bounce. Ambar Components Ltd. Tel., 0844 261144; fax, 0844 261789.

Fast changeover camera. Pearpoint's *P328* dual-sensor ccd-based camera can switch between colour and intensified-image modes in less than 1.5s, also switching rapidly between zones with greatly differing illumination levels. The changeover block has the two sensors back-to-back on the same axis, a gain limiter ensuring that the intensified-image sensor is not exposed to high light levels to prolong its life. The image intensifier has a gain of 90, so that real-time video is produced in levels down to 10^{-3}lux, the eht unit being built in. PAL, NTSC and S-Video signals are available. Pearpoint Ltd. Tel., 0420 489901; fax, 0420 477597

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Clock recovery chip. *AD802-155BR* from Analog is a clock recovery and data retiming device at rates of up to 155.52Mbit/s. It uses frequency lock for acquisition and tracking and phase lock for close, low-jitter tracking, needing only a single external capacitor. Lock is maintained through a 240-bit transitionless data run. The device tracks the SONET OC-3 standard jitter-tolerance mask and its jitter bandwidth is 130kHz. Analog Devices Ltd. Tel, 0932 253320; fax, 0932 247401.

Memory chips

Flash/eeprom. SST's *Superflash* is a 128K by 8 cmos memory chip combining the reliability of an eeprom with the small size of the Superflash memory cell. The new technique uses thick oxide tunnelling to increase endurance and eliminate the over-erasing associated with the thin tunnel oxide process. Erasing is by sector, taking less than 10ms to erase and rewrite a page sector and there is software and hardware write protection. No 12V supply is needed. Ambar Components Ltd. Tel., 0844 261144; fax, 0844 261789.

Microprocessors and controllers

Improved PIC16C54. An improved version of Microchip's *PIC16C54* 8-bit microcontroller uses a 0.9μm double-layer metal wafer process and is powered by a single lithium battery. It is a risc-like device operating at up to 20MHz and with a fuse configurator to select *RC* timing circuits and crystal/resonator clock options. Memory is 512 words of eprom for program and 25byte of static ram for data. Peripherals include an 8-bit clock/counter with a programmable prescaler, watchdog timer and 12i/o lines with individual directional control. Current drain in sleep mode is 4μA. Arizona Microchip Technology Ltd. Tel., 0628 850303; fax, 0628 850178.

Risc i486 replacement. The *Rab2it* chipset from NEC enables the i486 processor to be replaced by a risc chip, providing a link between the NEC *VR4X00* processors and the PC bus and between processor and conventional or synchronous dram. Currently available are the i/o controller and a memory controller; the third device – to provide a direct connection to Rambus dram – will be made available later this year. The 3.3V or 5V i/o controller handles dma and interrupt requests from PC AT systems and can be used with *VR 4200/4400PC/4400SC* processors. NEC Electronics (UK) Ltd. Tel., 0908 691133; fax, 0908 670290.

Mixed-signal ICs

Forward-thinking teletext chip. *MV1817* is a GEC Plessey teletext decoder that captures and stores in dram up to 256 pages for rapid viewing. It stores all linked pages in the same magazine, ten pages either side of the current one and all page numbers ending in zero. In the German TOP system, it captures all pages on an index page and will work with the Advanced Header transmissions in Austria and Belgium. Automatic television tuning and video recorder control are also carried out. GEC Plessey Semiconductors Ltd. Tel., 0793 518510; fax, 0793 518582.

100Mbyte/s Fibre Channel. AT&T is to sample two ICs that meet the Fibre Channel physical interface for 100Mbyte/s data transfer in networked computer groups and for connecting peripherals. *DA204/5* chips use less than 5W from 5V and provide the retiming, transmit and receive functions for 1Gbit/s products now under development by Ancor and Finisar in America. AT&T Microelectronics. Tel., 0732 742999; fax, 0732 741221.

Overvoltage protector. Harris's *SP721* is a low-cost discharge/overvoltage protector, capable of handling six inputs, including microprocessor i/o and buses, from voltages up to 15kV. It switches in 6ns and takes up to 2A of peak current, triggering when the voltage on positive or negative lines increases by more than one diode drop and returning to normal operation after the transient disappears. Normally, the device appears as 3pF input capacitance and draws 1nA leakage. Voltage range is 4.5-30V dc. Harris Semiconductor UK. Tel., 0276 686886; fax, 0276 682323.

Game sound generator. For the games market, Yamaha offers the *YMZ280* eight-channel sound generator, which takes a 4-bit adpcm input or 8/16-bit linear pcm to produce the eight channels. It samples at up to 44.1kHz maximum or 172Hz minimum, an external address space 16Mbyte being available for waveform data in rom or sram. Also on offer is the *YAC513* d-to-a converter and the *YSS225* effects processor, which samples up to 48kHz and has a delay time of 1.5s. Polar Electronics. Tel., 0525 377093; fax, 0525 378367.

Optical devices

Laser diode modulation head. The *AVX-SRB* module by Avtech is a bias-T modulation head, which combines the dc output of a laser diode driver with an rf modulation input in the 10MHz-1.5GHz range to drive 3-pin and 4-pin laser diodes, which fit into a socket in the head. A current of 0-500mA is applied to the laser diode, maximum rf input power being 100mW for 125mA pk-pk current swing about the dc bias current. The module mates with heat sinks and coolers and takes laser diode drivers by Seastar. Lyons Instruments Ltd. Tel., 0992 768888; fax, 0992 788000.

Multi-chip leds. Multi-chip leds in the *EBT* range can be used as direct replacements for panel-mounting incandescent lamps. The devices consist of several led chips bonded directly onto a substrate, conferring better resistance to shock, no surge, longer life and lower heat when compared to filament lamps. Devices with 4, 6 or 8 chips are produced in red, yellow or green with a choice of series or parallel connections. EAO-Highland Electronics Ltd. Tel., 0444 236000; fax, 0444 236641.

Oscillators

Surface-mounted oscillators. Contained in a plastic package measuring 13 by 9.8 by 1.7mm, IQD's *IQXO-50* series of surface-mount oscillators work in the 1-70MHz range, with frequency stabilities of ±100ppm or ±50ppm, including load, supply and temperature (0-70°C) variations. All are described in a new Data Book, now available. IQD Ltd. Tel., 0460 77155; fax, 0460 72578.

Power semiconductors

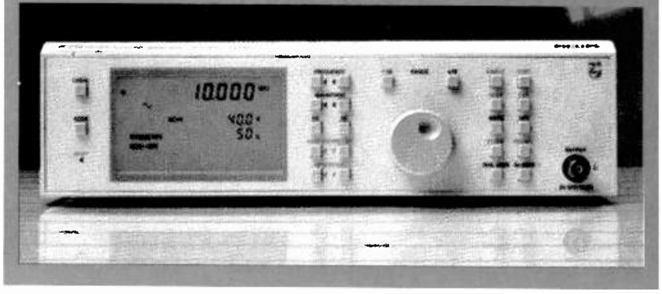
Isolated Hexfets. Low gate-charge Hexfets from IR are now in a fully isolated TO-220 FullPak, which guarantees 2.5kV rms isolation, its thermal resistance comparing well with other methods of isolation. First FullPak parts are 400-600V types with the same current and resistance characteristics as Hexfets, but having a gate charge reduced from 63nC to 39nC, Miller capacitance from 120pF to 18pF and input *C* down to 1100pF. International Rectifier. Tel., 0883 713215; fax, 0883 714234.

Pulse transformer/mosfet interface. *HV400MJ/883* takes its input directly from a pulse transformer and drives a power mosfet at switching rates up to 300kHz, or about three times the usual frequency obtained without the use of fairly complicated discrete circuitry. Harris's IC is a 6A device for use as a power switch building block for high and low-side switches, secondary-side regulators and synchronous rectifiers, driving capacitive loads in the 5-100nF range. Two pins allow independent control of rise and fall times. Harris Semiconductor UK. Tel., 0276 686886; fax, 0276 682323.

High-current IGBT. The *MG600Q1US41* from Toshiba is a 1200V, 600A insulated-gate bipolar transistor with only 20nH inductance between positive and negative terminals, its gates being protected by integrated back-to-back zeners. Saturation voltage with a resistive load is 3V and current fall time 200ns. A signal-collector terminal allows simple desaturation detection. Toshiba Electronics (UK) Ltd. Tel., 0276 694600; fax, 0276 691583.

Micropower regulator. Cherry's *CS-8101* linear voltage regulator takes only 70μA of quiescent power (50μV in sleep mode), drops out at 5.2V to give a 5V output and provides a microprocessor reset pin, keyed from the output, valid down to 1V output voltage. The device is protected against esd, overvoltage, shorts, thermal runaway and reversed battery, and is pin-compatible with the National Semiconductor *LP2950/51*. Clere Electronics Ltd. Tel., 0635 298574; fax, 0635 297717.

Function generator. With a frequency range of 0.1mHz-10MHz, Fluke's *PM5138A* synthesised function generator provides up to 40V pk-pk protected output into either 50Ω or 600Ω. Seven standard waveforms are produced and an arbitrary one that can be created on a PC using Fluke's *AnyWave* software, and down-loaded on the optional IEEE488 or RS232 interface. Waveforms can also be captured on a digital oscilloscope and directly transferred to the *PM5138A*. Modulation includes am, fm, psk, burst, gating and linear or log. sweep. Fluke UK Ltd. Tel., 0923 240511; fax, 0923 225067.



PASSIVE

Passive components

Ceramic resonators. Ceramic resonators provided with built-in capacitors, *ZTS* components from Integrity Technology Corp. cover the 2-12MHz frequency range. Capacitors are either 30pF connected for ttl use or 100pF for cmos. Frequency tolerance is ±0.3% per year. Integrity Technology Corporation. Tel., (USA) 0101408 262-8640; fax, 0101408 262-1680.

Improved electrolytics. Nichicon's *WT* range of surface-mounted electrolytic capacitors now handle up to 60mA ripple current at 100μF and 10-16V operating voltage. Load life is 32000h at 55°C, range is 0.1-100μF and working voltages 1-50V. The range measures 5.5mm high and diameters are 3-6.3mm. Nichicon (Europe) Ltd. Tel., 0276 685393; fax, 0276 686531.

Ceramic resonators. In five package styles, ceramic resonators by Panasonic cover the 3.58-35MHz frequency range. There are two ranges of leaded components, all with a height off-board of 5mm and three of chips, two of which stand 2.5mm high and the third 1.8mm. All are based on PCM ceramics and can be supplied as general-purpose types or with integral capacitors. Tolerance is down to ±0.3% and ten-year drift to –0.1%. Panasonic Industrial (Europe) Ltd. Tel., 0344 863444; fax, 0344 861656.

SM ceramic C sample kits. Eight surface-mounting sample kits for Philips's *0603*, *0805* and *1206* standard and microwave capacitors are now available, containing representative samples with rated voltages of 63V for the microwave types and 63/20V for the standards. A variety of terminations and capacitor characteristics is presented. Gothic Crellon Ltd. Tel., 0734 788878; fax, 0734 776095.

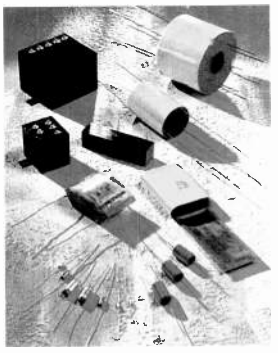
Dielectric filter. AVX announces the *PDFC* series of dielectric filters meant for use in telecomms, particularly in the DECT sector. Frequency range is 1.8-2GHz, insertion loss 3dB and, for compatibility with the newest equipment, size is 6.5 by 5.5 by 3mm. Filters to provide lower insertion loss and improved stop-band attenuation are available to order. AVX Ltd. Tel., 0252 336868; fax, 0252 346643.

Connectors and cabling

Sealed BNC. From Bulgin, a range of sealed connectors that incorporate 50Ω or 75Ω bnc inserts and that can be assembled into any of the five *Mini Buccaneer* bodies to provide sealing to IP66, BS5490:1977. Dust caps preserve the connectors' invulnerability to dust and moisture when the connectors are separated. A brochure of Buccaneer sealed products is available. Gothic Crellon Ltd. Tel., 0734 788878; fax, 0734 776095.

PCMCIA memory. Fujitsu's *FCN 560H* PCMCIA memory cards connectors are single and double types for mounting 68-pole memory boards on PCs, conforming to JEIDA/PCMCIA I and II standards. They come with straight or right-angled pins, with or without an ejector and with polarisation. Resistance of the gold-plated pins is 40mΩ maximum. Electrospeed. Tel., 0703 644555; fax, 0703 610282.

Eurocard connectors. The range of Siemens DIN41612 connectors is now available from Quiller. They are in two parts and come with contact ratings up to 5A and rf types, meeting all major standards. Also available is a range of press-fit connectors with a highly compliant press-fit zone to give good connection with low insertion



Custom capacitors. *Suflex Capacitors* has a new custom design and manufacture service for prototype and short runs as well as production quantities. Plastic-film components can be produced in polypropylene, polycarbonate, polyester and polystyrene in values from picofarads to microfarads and with working voltages up to 30kV. Customers specify size, shape, encapsulation and fixing method. Suflex Capacitors Ltd. Tel., 0633 615511; fax, 0633 613583.

and extraction forces. Through-hole and wire-wrap versions are made and the connectors meet IEC 352-S. Quiller Switches Ltd. Tel., 0202 417744; fax, 0202 421255.

Shielded ribbon connectors. High-density mini delta ribbon (mdr) board-mounted plug connectors from 3M in the *101 series* are for internal and external i/o use where small size, emi/rfi shielding and electrostatic discharge before mating are needed. The *101* is a right-angled through-hole-mounted plug in staggered four-row form on a 0.05in by 0.075in matrix and is a miniature version of the standard D ribbon or Centronics connectors, being rated at 500 insertions/ withdrawals. It is available in 20, 26, 36 and 50-way versions 3M United Kingdom plc. Tel., 0344 858000; fax, 0344 858758.

Filters

Coaxial filters. Three diameters of coaxial filter by *Atlantic Microwave* cover the 30-5000MHz frequency range in band-pass or low-pass form. Depending on the number of sections, they offer a 70dB rejection with low insertion loss and 1.5:1 vswr. Band-pass units with 2-12 sections can be specified at 50Ω, 75Ω or 100Ω and with pass bands of 1-100% of centre frequency. Unit size is 9.5-19mm diameter and 50-450mm in length. Atlantic Microwave Ltd. Tel , 0376 550220; fax, 0376 552145.

Hardware

Instrument cases. *Serpac* general-purpose cases are intended for almost anything from test instrument enclosured to computer interfaces, having a precision fit, a textured finish and prices down to £1.50. These ABS boxes are in two types: the *i* series with a recessed top for membrane keypads and the standard series with a flat top. OKW Enclosures Ltd. Tel , 0489 538858; fax, 0489 583836.

Pocket boxes. These boxes by OKW feature IP65 sealing, a recessed top to take membrane keypads and a battery compartment to avoid the use of the less reliable battery clips often found in this type of case. They are in flame-retardant ABS and come in three sizes from 85 by 46 by 16mm to 120 by 65 by 22 mm. Internal pcb mountings are provided and belt clips, emc protection and silk-screening are offered. OKW Enclosures Ltd. Tel., 0489 538858; fax, 0489 583836.

Compact keyboard. Providing the full recommended 3mm travel, Cherry's *G84-4100* low-profile, compact keyboard is said by the company to be the smallest standard keyboard available, its enclosure measuring 23.5mm by 281mm by 131.5mm. There is an enclosed version with tilt feet and a model supplied without the housing. Mechanical keyswitches do not deteriorate with time and temperature and keycaps are available for most languages Cherry Electrical Products Ltd. Tel., 0582 763100; fax, 0582 768883.

Instrumentation

High-current probe. A current probe for use with multimeters and oscilloscopes, the *PR1000-1* from Lem Heme, uses Hall effect to measure currents up to 1000A peak at frequencies up to 10MHz to an accuracy of ±1% of reading ±0.5A. Output is 1mV/A. Lem Heme Ltd. Tel., 0695 20535; fax, 0695 50279.

Disturbance analysers. Avo's *PDA2* power disturbance analyser will now show pulse direction as well as amplitude and duration, using a current sensor kit. By placing the analyser at several points, the user can determine whether, for example, noise pulses come from the equipment or from the mains. The two-channel instrument evaluates disturbances on single and three-phase, 400Hz and dc supplies, results being shown on an lcd or built-in plotter. An RS232 port enables remote control and data downloading. Avo International Ltd. Tel., 0304 202620; fax, 0304 207342.

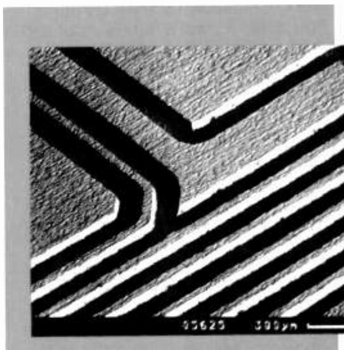
Panel meters. The 30-model *Select* range of panel meters from Sifam is in five groups containing voltage and current , meters, process meters, rate and frequency monitors, batch counters and temperature indicators. Display is by red or green leds and the user can set the instrument up and protect the setting by a password or two concealed buttons. Sifam Ltd. Tel., 0803 613822; fax, 0803 613926.

IR monitor. New from the Raytek range of non-contact infrared temperature measuring instrumentation, the *Thermalert 30* is a universal monitor to work with the Raytel T series of sensing heads in the –50 to 3000°C range. Features include variable emissivity and peak/valley hold for sampling temperature of objects on a conveyor belt while ignoring the belt temperature. Objects down to 1mm can be measured as well as those at long distance and laser sighting heads are on offer. Calex Electronics Ltd. Tel., 0525 373178; fax, 0525 851319.

True-rms oscilloscope. TTi's *Scopemaster* range of lcd oscilloscopes now includes the *SM630*, which offers five measurement functions, including the indication of true-rms quantities. It functions as digital storage oscilloscope, digital multimeter, data logger, counter/timer and serial data analyser. Bandwidth of the dso is 20MHz, with a maximum sampling rate of 20Msample/s, roll mode down to 200s/div. and repetitive sampling to 2.5ns resolution. The 3000-count dmm provides ac and dc current and voltage scales, *R*, continuity and diode test and the logger stores up to 1000 readings. Thurlby Thandar Instruments Ltd. Tel., 0480 412451; fax, 0480 450409.

Literature

Pressure sensors. Motorola has revised its data book on microprocessor-compatible pressure sensor devices, which contains data sheets and application notes describing interfaces to users' circuitry, together with information on evaluation boards. Motorola Inc. Tel., 0908 614614; fax, 0908 618650.



Production equipment
Laser PCB machining. Tracks CAD Systems has a laboratory pcb machining tool, the *LPKF*, which can cut seven conductor paths between two IC pads, leaving 40μm isolation channels, cutting with a resolution of 1μm to an accuracy of 2μm. It cuts standard FR4 or copper-coated ceramic material, producing clean, square edges. Services needed are power, compressed air and water – no clean room necessary. Tracks CAD Systems Ltd. Tel., 0344 55046; fax, 0344 860547.

NEW PRODUCTS CLASSIFIED

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

Materials

UV coating. A new type of ultraviolet conformal coating has been introduced by Intertronics which cures as well in shadows as it does in direct uv light. *Dymax Darc Cure* cures in a few seconds throughout the entire coating by means of a post-cure reaction with air, without the use of aqueous or chemical accelerators. The coatings are free of solvent and ozone-depleting chemicals and can be inspected under black light. Intertronics Ltd. Tel., 0865 842842; fax, 0865 842172.

Power supplies

Low-profile transformers. Clairtronic's new low-profile encapsulated power transformers cover the 3-50VA range from the now harmonised mains input of 230V. All produce two outputs, which may be connected in serial or parallel and there is one-time thermal fusing. Materials used are flame-retardant to UL 94VO. Clairtronic Ltd. Tel., 0753 692022; fax, 0753 535096.

600mA dc-to-dc converter. From a two-cell input, Linear's *LT1302* converter supplies 5V at up to 600mA at 87% efficiency. Quiescent current is 200μA, reduced to 15μA at logic-controlled shutdown. The internal n-p-n power switch can handle over 2A and switch at 400kHz, so that small components can be used. Micro Call Ltd. Tel., 0844 261939; fax, 0844 261678.

Efficient step-up converters. *MAX856/7/8/9* are current-limited, step-up dc-to-dc converters claimed by Maxim to offer the world's best combination of efficiency (85% at 100mA), low quiescent current (25μA) and low shutdown current (1μA). *MAX856/8* accept a 0.8-6V input and produce a pin-selected 3.3V or 5V output, while *MAX857/9* give adjustable output between 2.7V and 6V. Maxim Integrated Products UK Ltd. Tel., 0734 845255; fax, 0734 843863.

120W UPS. *BI-UPS*, which is short for built-in uninterruptible power supplies, is a range of 120W ac supplies from Magnum that are able to supply 17in colour monitors in battery backup mode. The units protect themselves and other equipment against a whole range of supply line disturbances in compliance with UL1449, and give backup for 30min during blackouts. If the ac supply drops below 82% of nominal, the ups switches to battery backup. Magnum Power Solutions Ltd. Tel., 0236 433325; fax, 0236 427366.

Dc-to-dc converters. A new range of converters from Schroff includes an extended pluggable type for use with SELV circuits. There four in/out ranges from 8.5-18V to 80-160V, with 5V, 12V, 15V and 24V outputs, single, dual or triple at 25W, 60W or 120W. Overvoltage and short-circuit protection is provided on inputs and outputs. Schroff UK Ltd. Tel., 0442 240471; fax, 0442 213508.

Wide-range dc-to-dc converters. Single-output *UWR* and dual-output *BWR* 10W and 20W converters take an input range of 4.6-13.2V in an operating temperature range of −25°C to 105°C. Outputs are 3.3, 5, 12, 15 and ±15V and sizes are 2 by 1 by 0.375in in the 10W type and 2 by 2 by 0.45in for the 20W version. The pwm techniques adopted confer 84% efficiency and 200μs transient response. There is overvoltage shutdown and i/o isolation of 500V dc. Datel (UK) Ltd. Tel., 0256 880444; fax, 0256 880706.

Radio communications products

RF cable. SMC's range of cables includes the UR67, which gives a loss at 150MHz of 9.3dB/100m and at 2000MHz 53dB. *1OF-SFB* exhibits 3.8dB loss at 150MHz and 16.8dB at 2000MHz. South Midlands Communications Ltd. Tel., 0703 255111; fax, 0703 263507.

Switches and relays

Automotive relays. *SAM* relays from Selectronic are compact and offer 15V dc 20A, 24V dc 15A switching current for use in vehicle control systems. The standard *SAM* single-pole device measures 17 by 12 by 16.1mm, the two-pole type comprising two of those in one package. Ambient temperature range is −25°C to 80°C. Selectronic Ltd. Tel., 0993 778000; fax, 0993 772512.

Transducers and sensors

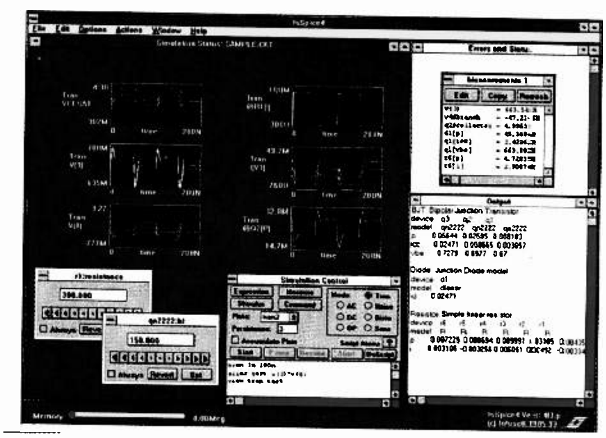
Load washer. Control Tranducers has the *Model LW-M* load washer, temperature-compensated and hermetically sealed to measure bolt stresses, axial thrust, static, dynamic and clamping forces. Ranges are 300-150,000kg and comes in sizes from 34mm diameter, 26mm thick. Creep is ±0.2% fsd over four hours, repeatability ±0.1%, non-linearity and hysteresis better than ±0.2%. With a bridge impedance of 350Ω, sensitivity is 2mV/V. Control Transducers. Tel., 0234 217704; fax, 0234 217083.

COMPUTER

Computer board-level products

Single-board computer. The *Octagon PC-425* is a single-board computer for embedding in systems needing only limited i/o. It uses a 386/25 cpu and 387 co-processor (486 at the moment on special offer). Features include dos 5.0 in rom, four silicon disks giving 2.2Mbyte, keyboard connector and speaker, COM1 and COM2 ports with jumpered RS232, RS422 and opto-isolated RS485 interfaces and a rack for opto-isolated i/o modules accepting digital and analogue modules. Gothic Crellon Ltd. Tel., 0734 788878; fax, 0734 776095.

Data acquisition boards. Two more in the *200 series* of data acquisition boards are introduced by Amplicon Liveline. *PC234* has four channels of bipolar analogue voltage output at 16bit resolution, a four-wire voltage sensing circuit on each channel eliminating load and cable resistance errors. It provides four-quadrant d/a multiplication. *PC224* gives 16 bipolar channels with an independent ground line taken to source. Both give sequential or simultaneous update on all channels and features included are internal clock, external clock source, three programmable counters and an external trigger. Programs can be written in Quick Basic, Turbo Pascal and C/C++. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 0273 570215.



Software

Interactive Spice. *ICAP/4 Virtual Circuit Design Lab* from Intusoft is a circuit simulation system that is completely interactive, integrating schematic entry, the new *IsSPICE4* simulator, libraries and graphical waveform analysis. It runs under Windows 3.1X, NT and 4.0. When *IsSPICE4* is started, the user may examine the design by running different analyses, changing circuit values and measuring the results, without having to close and restart the simulator and alter the Spice input netlist. A cross-probing tool allows the display of waveforms on the circuit diagram and the interactive stimulus mode enables the user to sweep any number of component or parameter values to compare circuit performance. Technology Services Ltd, Tel. 0638 561460, fax 0638 561721.

Data communications

Single-cable video, audio and data. Stand-alone or rack-mounted optical-fibre transmitters and receivers by Fiber Options are intended for cctv, carrying two-way video, audio and data by fm on one 50m or 62.5m cable. A typical application would be in remote camera control, where pan/tilt-zoom, windscreen wiper and heater are controllable by RS232, RS422 or TTL levels and the audio and video go to the control centre on the same link. Auriga (Europe) plc. Tel., 0908 274200; fax, 0908 378998.

Development and evaluation

83CL782 development. Ashling Microsystems has a new development system for the Philips *83CL782* low-power microcontroller family, based on an 8051 core and used in cordless telephones, DECT, pagers and portable GPS. The Ashling system provides real-time, non-intrusive emulation of all devices and package styles in the family and source-level debugging under Windows 3.1 is available for all the 8051 languages. Ashling Microsystems Ltd. Tel., 010353-61-334466; fax, 010353-61-334477.

Embedded bios v2.0. The General Software Inc. Embedded bios is a rom bios development kit with full feature support for the Intel 386EX processor in embedded designs. The new kit has support for PCMCIA, Intel Flash, Intel System Management Mode (SMM), synchronous serial i/o, watchdog timer, power management and the on-chip 386EX processor devices. There is a disassembler and debugger for hardware drivers and v2.0 also includes standard desktop PC AT bios functions such as Post, cmos support, setup screen, protected mode, PC AT keyboard support, serial and parallel i/o, floppy and IDE drives, clock and video functions. Great Western Instruments Ltd. Tel., 0272 860400; fax, 0272 860401.

PowerPC compilers for Power Mac. Motorola's *PowerPC C, C++ and Fortran microprocessor compilers* are to be ported to Apple's Power Macintosh range of computers, where they will be compatible with the Macintosh *Programmers' Workshop* development environment. There are also plans to do the same for the Metrowerks *CodeWarrior* environment, so that users will be able to use Motorola's compilers without having to adopt a new development environment. Both projects are to be completed this year. Motorola Inc. Tel., 0908 614614; fax, 0908 618650. ■

Digital filters simplified

With DSP replacing analogue equivalents in almost all areas of electronics, digital filters are becoming increasingly important. Bob Lawlor* looks at how a field-programmable gate array can simplify the design of both infinite and finite-impulse response filters.

There are two types of digital filter, one with finite-impulse response, FIR, the other with infinite-impulse response, IIR. A digital FIR filter is a discrete-time filter. Any sample of its output sequence is equal to a weighted sum of a finite number of past and present samples of the input sequence.

Mathematically an n-tap FIR filter is expressed as,

$$y(k) = \sum_{j=0}^{n-1} c_j x(k-j)$$

where x(k) and y(k) represent the input and output sequences respectively and represents the filter coefficients.

Figure 1 shows an FIR filter block diagram. The τ-boxes represent unit sample delays. Sampling frequency depends on the application. For example an audio system might warrant up to 40k samples per second while the standard sampling frequency for video is 13.5M samples per second. How many bits there are in each sample also depends on the application, hi-fi audio being typically sixteen bits and video eight bits.

Filter coefficients, c_0, c_1, etc. multiply or weight the delay outputs. Resulting product terms are summed together to form the filtered output data sequence, y(k). The name finite-impulse response is derived from the fact that an impulse injected into the input will cause the output to become active, i.e. non-zero, for only a finite duration.

A key advantage of FIR filters is that they can be designed to have a linear phase characteristic, **Fig. 2(a)**. This means that if the input signal to the FIR filter has more than one frequency component – and most real signals have a wide spread of frequencies – then in passing through the filter, all frequency components undergo the same delay.

For example, if the two frequency components of Fig. **2(b)** each undergo delay T, the phase change of the higher frequency component is proportionally greater than that of the lower frequency component. If on the other hand the phase characteristic is non-linear, then different frequencies undergo different delays, causing so-called phase distortion.

The ear is extremely sensitive to phase distortion which would occur if the audio signal was processed by a filter having a non-linear phase characteristic. In a composite video signal, the colour information is coded in the phase of the subcarrier, so phase distortion appears as colour distortion.

IIR digital filters

A digital infinite-impulse response filter is a discrete-time filter. Any sample of its output sequence is equal to a weighted sum of a finite number of past and present samples of the input sequence, and also of the past output samples.

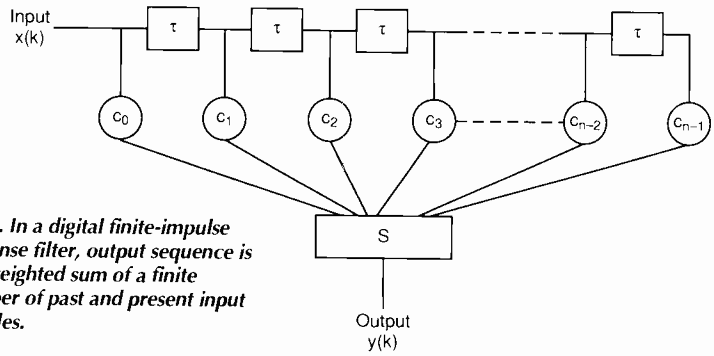


Fig. 1. In a digital finite-impulse response filter, output sequence is the weighted sum of a finite number of past and present input samples.

**Bob Lawlor was the winner of our December 1992 competition, sponsored by Xilinix. He was a senior R&D engineer with Sony, and has recently established his own DSP consultancy.*

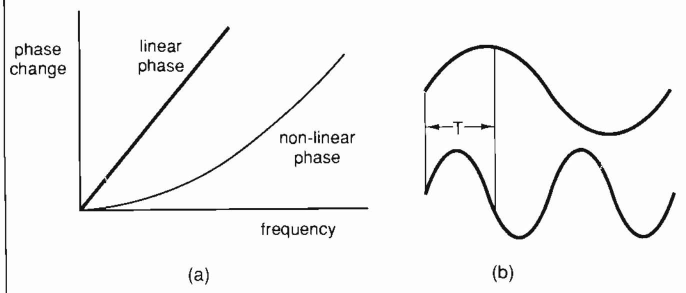


Fig. 2. Finite-impulse response filters can produce a linear phase characteristic (a). If two frequency components each undergo delay T, as in (b), phase change of the higher component is proportionally greater than the lower.

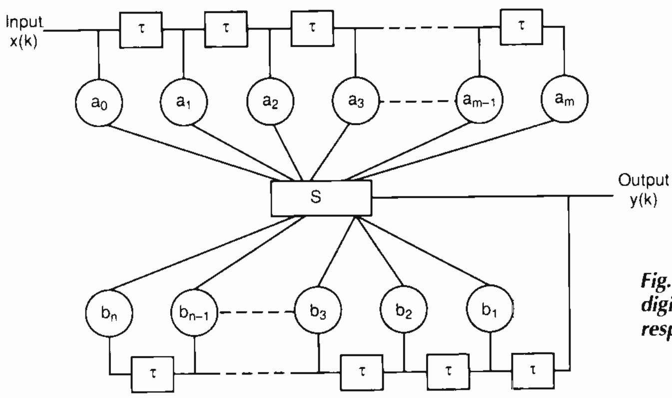


Fig. 3. Elements of a digital finite-impulse response filter.

Mathematically this can be expressed as,

$$y(k) = \sum_{j=0}^{m} a_j x(k-j) - \sum_{j=0}^{n} b_j y(k-j)$$

where again x(k) and y(k) represent the input and output sequences respectively. There are now two sets of filter coefficients, namely the feed forward coefficients, a_j, and the feedback coefficients, b_j.

The name infinite-impulse response comes from the fact that an impulse injected into the input can cause the output to remain active for an infinite duration. Inclusion of previous output samples in the present output sample computation is called feedback. For this reason IIR filters are often referred to as recursive filters and FIR filters as non-recursive.

An advantage of IIR filters is that they generally require less hardware than their FIR counterparts. For most applications however, this advantage is outweighed by their non-linear phase characteristic.

Video application

In decoding composite video to component RGB, an FIR bandpass filter centred on the subcarrier frequency, and a video field rate FIR comb filter combine to produce very high quality separation of the luminance and chrominance data. Subcarrier frequency is 4.43MHz for PAL, digital video is sampled at 13.5MHz and the video field rate is 50Hz.

This results in a much clearer picture than is possible with conventional TV sets, **Fig 6**. This type of video decoding is used in the professional video industry, for example in the Sony *BVX100* digital decoder.

Each of the FIR filters and the subtracter can be implemented readily using a single Xilinx FPGA. The 50Hz FIR comb filter would need two external field stores, but so also would other implementations.

Since the Xilinx FPGAs are so easily reprogrammable the FIR filter transfer functions can be changed on a field by field basis. This is very useful in adaptive signal processing applications.

Implementing the FIR filter

In Fig 1, values of coefficients c_0, c_1...c_{n-1}, determine the filter characteristics[1]. Being readily reprogrammed, the Xilinx FPGAs makes changing the resolution to suit the application easy. A prototype filter for example could be implemented with relatively low resolution. At the pre-production stage, resolution could then be increased to enhance overall system quality. Existing asic general-purpose FIR filters are mostly fixed in terms of resolution.

In order to control round-off error, the multiplier outputs must be kept at full resolution. If for example the coefficients are n-bit numbers, then the multiplier outputs should be 2n-bit numbers. The final rounding to n bits will be done at the summer output.

In most cases the filter response needed is symmetrical so that $c_0=c_{n-1}$, $c_1=c_n$, etc. This means that by adding the relevant delay out-

EXAMPLE 1: AUDIO FIR FILTER APPLICATION

A popular technique used for both audio and video data compression is sub-band coding. This involves splitting the audio spectrum into bands, **Figs 4, 5** using a number of FIR bandpass filters. Next the different filter outputs are quantized with different resolutions, i.e. bits per sample. Lower frequency bands are generally given a higher number of bits per sample than higher frequency bands.

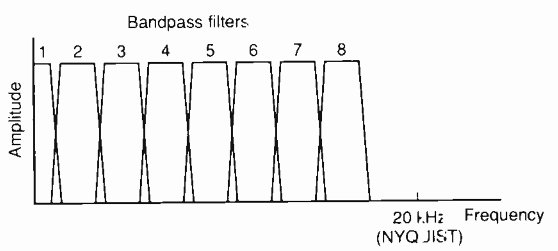


Fig. 4. Spectral segmentation FIR bandpass responses for audio sub-band coding

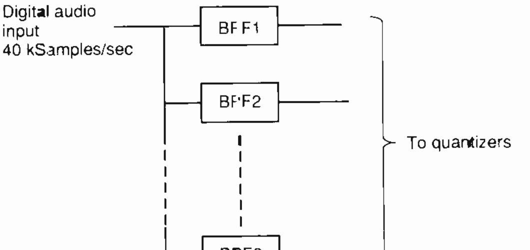


Fig. 5. finite-impulse response bandpass filters used at audio sub-band coder front end. These filters can be implemented in Xilinx FPGAs.

puts before the coefficient multiplication stage, the number of multipliers required is reduced from $2n+1$ to $n+1$.

Also, using general purpose multipliers to realise the required filter response as indicated in Fig. 1 can introduce redundancy. By using a shift-and-add technique, a Xilinx FPGA can remove this redundancy allowing a much more efficient implementation of the required FIR filter.

Shifting-and-adding

Consider an 8-bit two's complement number:

$$s\ b_7\ b_6\ b_5\ b_4\ b_3\ b_2\ b_1$$

This consists of a sign-bit, s, and seven other bits b_{7-1}. To multiply this number by a half, it is necessary only to shift it one place to the right, duplicating the sign-bit:

$$s\ s\ b_7\ b_6\ b_5\ b_4\ b_3\ b_2\ b_1$$

If the original number is thought of as having a range of −128 (10000000_2) to +127 (01111111_2) in integer steps, then the shifted number has a possible range of −64 (110000000_2) to +63.5 (001111111_2) in half integer steps.

By shifting the original number two places to the right, you effectively multiply it by a quarter. The sum of these two shifted numbers is therefore equal to the original number multiplied by three quarters. This process of shifting and adding can be continued to implement the product of any two digital numbers.

There is no requirement for CLBs in the shifting process because it is inherent in the routing. As a result, each multiplication can be implemented with adders alone. Extra delays in the form of D-type flip-flops may be needed between the addition stages, depending on the coefficient value and the sampling frequency.

Clearly, the coefficient value determines the number of addition stages necessary to implement the product. Each logic one in the coefficient binary representation represents an adder input in hardware terms. The fewer the number of ones, the lower the hardware requirement. This is why 1/2 (01000000_2) and 1/4 (00100000_2), assuming a range of −1 to +127/128, are convenient coefficients.

A coefficient like 01111111_2 is not as difficult as it may appear because it can be implemented as its two's complement inverse (10000001_2). The resulting product is then inverted before being fed to the final summer.

One way to guarantee convenient coefficients is as follows. Firstly design the filter using a standard filter design technique such as the Remez Exchange Algorithm[1]. This will produce 'not-necessarily-convenient' coefficients in terms of ease of implementation by a shift-and-add technique. It does however minimise the deviation from the desired ideal filter specification. Secondly, carrying out a discrete space search around these coefficients to produce a convenient set of coefficients. These keep deviation from the desired ideal filter specification within a specified tolerance.

The greater the number of coefficients, the more difficult it becomes to find a set of convenient coefficients. An easy way around this using a Xilinx FPGA is to implement the desired filter as a cascade connection of smaller filters, i.e. with fewer taps. When two or more FIR filters are cascaded, the resulting transfer function is equal to the product of the individual transfer functions.

For example, the bandpass filter transfer function of **Fig 7(c)** could be implemented as a cascade connection of a lowpass filter, Fig **7(a)** and a highpass filter, Fig **7(b)**. The relatively undemanding transfer functions of the low- and highpass filters make it easy to find convenient coefficients for their shift-and-add implementation in the FPGA.

If a sharper bandpass transfer function were required, the above filter could simply be cascaded with an identical version of itself, Fig **7(d)**. As the individual filters in the cascade chain are relatively simple, a large number of them can be incorporated into a single Xilinx FPGA device, yielding a complex overall transfer function. ■

Glossary

ASIC application specific IC.
BPF bandpass filter
CLB configurable logic block, a functional element from which the user's logic is constructed.
DAT digital audio tape
DCC digital compact cassette
DSP digital signal processing
FIR finite-impulse response
FPGA field programmable gate array
HPF highpass filter
IIR infite impulse filter
LPF lowpass filter
PAL phase alternate line
RGB red green blue

Reference

1. Rabiner, L. R. and Gold, B., *Theory and Application of Digital Signal Processing*, Prentice-Hall.

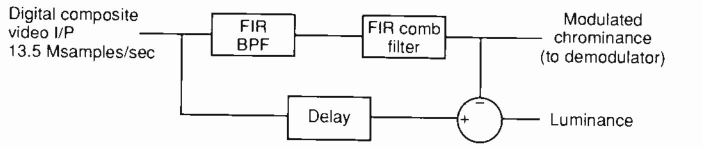


Fig. 6. Decoder for high quality video filtering provides a clearer picture than is possible with conventional methods.

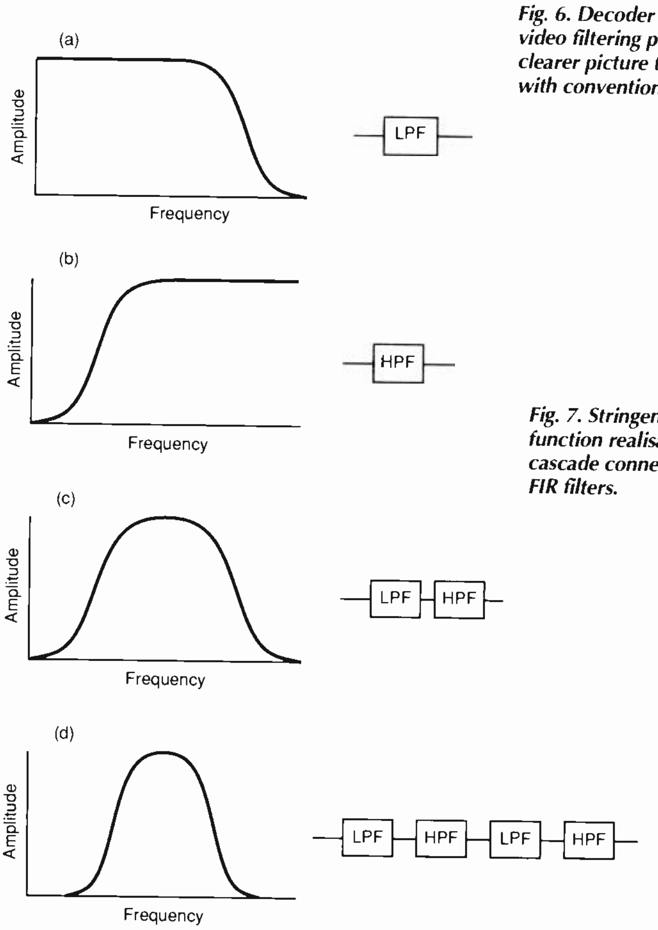


Fig. 7. Stringent FIR transfer function realisation by cascade connection of simple FIR filters.

ELECTRONIC ENGINEERS REFERENCE BOOK

6th Edition
F F Mazda

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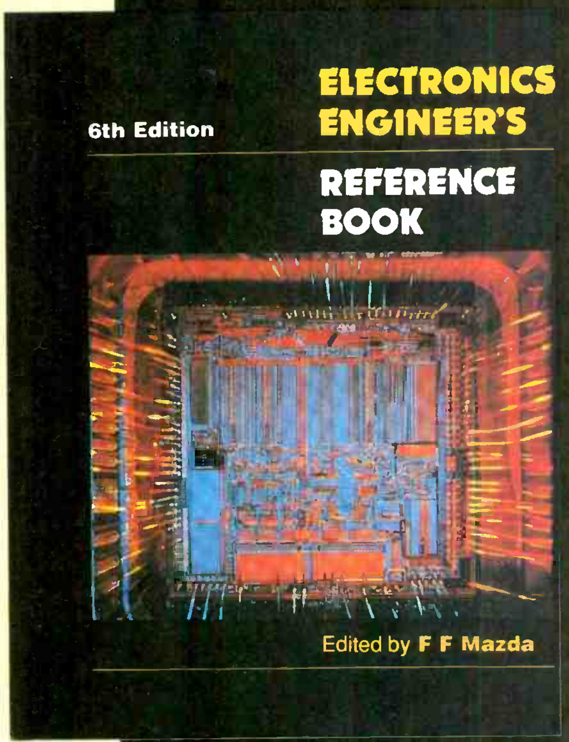

Fraidoon Mazda has worked in the electronics and telecommunications industry for over twenty years, and is currently Product and Operations Manager, Generic Network Management, with Northern Telecom. He is the author of six technical books (translated into four languages) and the editor of the Communications Engineers Reference Book published by Butterworth-Heinemann.

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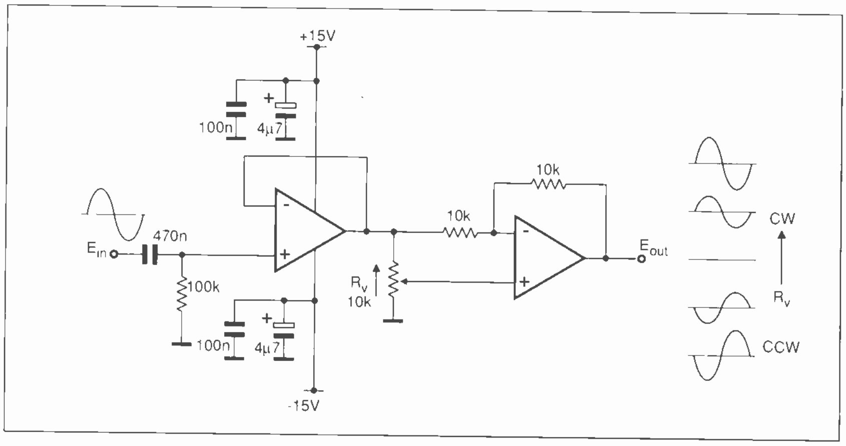
Single-pot polarity and gain adjustment

Using the 2:1 gain difference between inverting and non-inverting operational amplifiers, one control varies the output between a replica of the input and an inverted version, both being at unity gain.

Its uses include audio processing in which, used with a bucket-brigade delay element, it will allow the positions and amplitudes of the peaks and nulls of a comb filter to be adjusted.

Ben Sullivan
Waterlooville
Hampshire



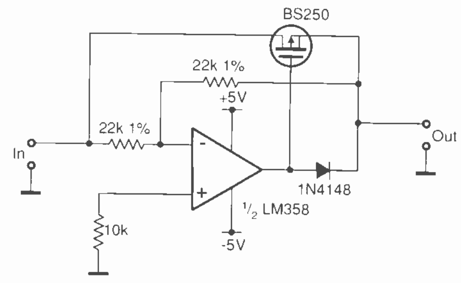
Potentiometer R_v varies output from inverted to non-inverted version of the input at unity gain.

Full-wave rectifier uses single amplifier

Incorporating an op-amp, one diode and a fet, this rectifier is linear down to less than 1mV, assuming that the op-amp's input voltage offset is compensated. Performance depends on input amplitude, being around 20% down at 1kHz and 50mV input.

On a negative half-cycle, the p-channel fet is open-circuit and the result is a unity-gain inverter. With a positive half-cycle, the fet conducts in reverse mode due to the presence of the large open-loop signal on its gate and the signal appears at the output.

Giuseppe Faini
Milan
Italy



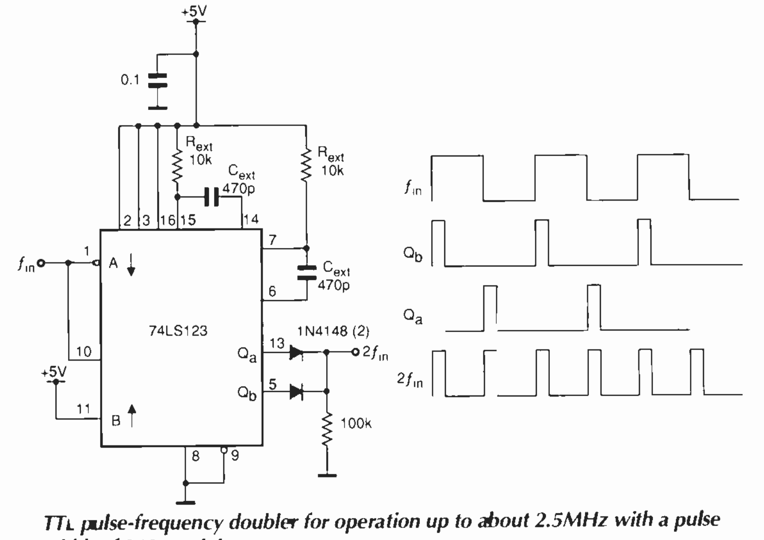
Simple, but accurate full-wave rectifier, linear down to less than 1mV.

Frequency doubler

TTL square waves at the input to the *74LS123* dual monostable multivibrator produce output pulses at double the input frequency. *1N4148* switching diodes Or together the outputs on pins 5 and 13, the two square-wave inputs having triggered the two monostables on opposite edges. Output pulse width is 3.3μs with 10kΩ and 470pF timing components according to the timing equation, pulse width = $0.7CR$. Since the package introduces 50pF residual capacitance and since R must lie between 5kΩ and 260kΩ, the practical minimum pulse width is around 350ns.

The process can be continued by driving a similar circuit with this output.

Mike McGlinchy
Mount View
California
USA



TTL pulse-frequency doubler for operation up to about 2.5MHz with a pulse width of 350ns minimum.

Low-level crossover is maximally flat

High and low frequency outputs from this low-level crossover network, **Fig. 1**, are in maximally flat, two-pole Butterworth form with 12dB/octave final slopes.

Output curves shown indicate that the amplifier is active mainly around the crossover frequency – in this case, 1kHz. Values shown are in ohms and farads, the crossover being at ω=1. Since input and output of the amplifier are dc-blocked, the practical circuit can use a single supply with the non-inverting input biased to half the supply voltage. A snag is the lowish input and high output impedances, but a practical circuit would include means of correcting them.

Analysing the circuit of **Fig. 2** shows the basic section, which gives 12dB/octave, but which cannot be maximally flat. **Figure 3** is more or less the same circuit with gain; resistors and capacitors are of equal value.

Since the desired transmission is

$$T = \frac{1}{p^2 + p\sqrt{2} + 1}$$

where $p = j\omega$, assume a 1V output and work backwards to make the input

$$e_{in} = p^2 + p\sqrt{2} + 1.$$

From Fig. 3

$$e_{in} = 1 + pA + A[p + p(1 + pA + pB)]$$

$$e_{in} = 1 + 3pA + p^2(A^2 + AB)$$

so

$$3A = \sqrt{2} \quad \text{and} \quad A^2 + AB = 1$$

or

$$A = \frac{\sqrt{2}}{3} \quad \text{and} \quad B^2 = \frac{7}{3\sqrt{2}}$$

Amplifier output is in bandpass form with its centre frequency ω=1. The high-pass circuit using that voltage is added to the low-pass section to form the complete crossover.

Actual component values for the circuit are calculated as follows, and given f=1kHz and C=4n7

$$R_1 = \frac{3}{\sqrt{2}} \times \frac{1}{2\pi fC} = 71.83\text{k}\Omega$$

$$R_2 = \frac{7}{3\sqrt{2}} \times \frac{1}{2\pi fC} = 55.87\text{k}\Omega$$

$$R_3 = \frac{\sqrt{2}}{3} \times \frac{1}{2\pi fC} = 15.96\text{k}\Omega$$

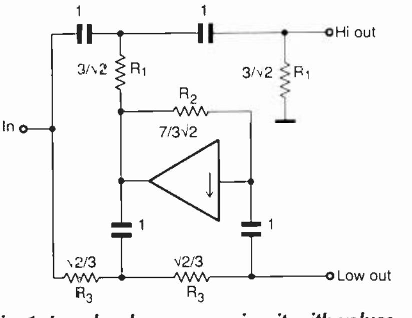


Fig. 1. Low-level crossover circuit with values normalised to ω=1, components in ohms and farads.

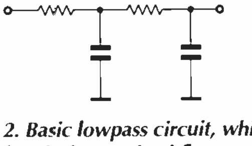


Fig. 2. Basic lowpass circuit, which needs gain for maximal flatness.

The circuit was built using an *LF356* op amp although a better device should be used for serious applications. The response departs from the theoretical only by the component tolerances used in the actual circuit.

McKenny W Egerton jr
Owing Mills
Maryland
USA

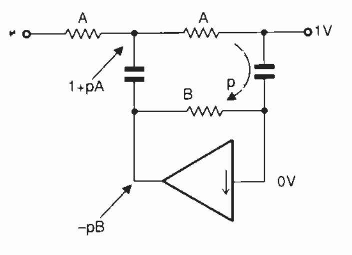


Fig. 3. Resulting lowpass section.

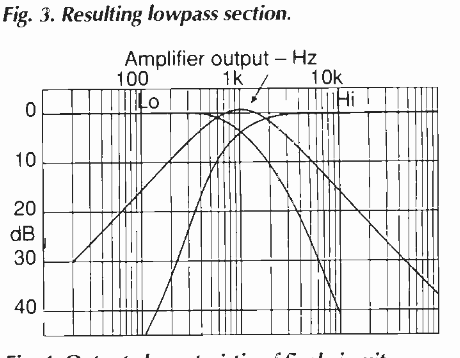


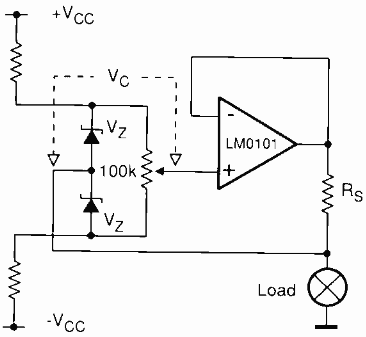
Fig. 4. Output characteristic of final circuit. Amplifier takes part mainly around crossover frequency.

Grounded-load current pump

Improvements on the Howland current source proposed by Steele and Green[1] demand close resistor matching to avoid gross error[2]. This circuit needs no matching.

The source uses two zener diodes as voltage regulators, the control voltage V_c thereby generated riding over the feedback voltage. Output voltage is the sum of load voltage and V_c, which gives a constant V_c across the sense resistor R_s, load current therefore being V_c/R_s. Lower-voltage zeners are preferable in this circuit.

Ashwani Karnal
Bhaba Atomic Research Centre
Jammu
India



Current source poses no resistor-matching problems.

References

1. Steele, J & Green, T. Tame, *Those Versatile Current-Source Circuits. Electronic Design*, Oct. 15, 1992.
2. Baker C B. *Current Source Saves Resistors. Electronic Design*, June 25, 1992.

Linear optocoupler cancels errors

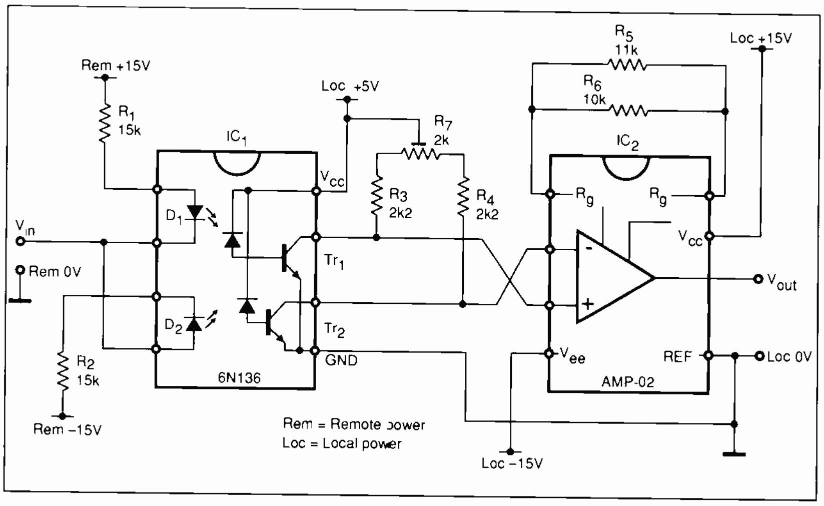
Using a dual optocoupler and an instrumentation amplifier, this isolation amplifier is linear and operates well above the cut-off frequency of much more expensive commercial devices.

To achieve a linear transfer function, the leds in the optocoupler are run in series at 1mA, the resulting differential signal appearing at the collectors of the transistors $Tr_{1,2}$ with virtually all errors cancelled out. Variable R_7 corrects for any offset. A useful feature of the *6N136* is that it has a screen between input and output.

The required circuit gain is obtained by setting R_g, gain being $1+50/R_g$. As shown, the total circuit gain is unity, but optocouplers do vary.

A triangular waveform at 500Hz showed no measurable distortion at up to 2.5V and 2% at ±8V. As regards frequency, a ±1V input gave a –3dB point of 150kHz. Input impedance is around 8kΩ, but a buffer could easily be used.

CJD Catto
Cambridge



Low-cost, but linear and wide-band, optically isolated amplifier. Running leds differentially cancels out errors.

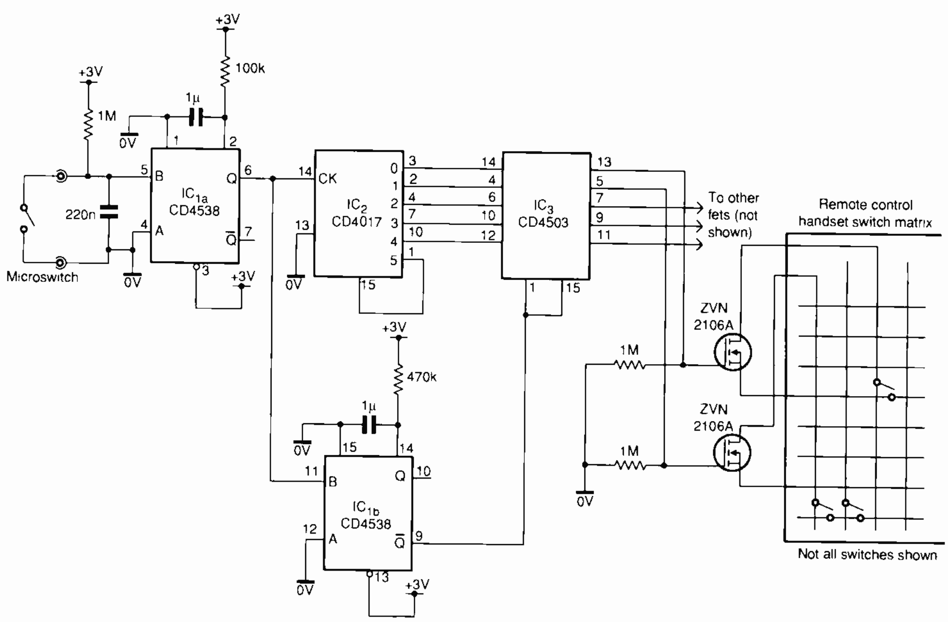
Disabled person's television remote controller

Adapting a remote control unit for use by someone with limited hand movement, this circuit steps through the four channels and standby in response to pressure on an elbow-operated microswitch. Output fets are in parallel with the controller buttons, so that the controller is still operable in its normal mode.

Half the dual monostable IC_{1a} debounces the microswitch, clocks the *4017* decoded decade counter and triggers the second monostable to produce a 0.5s pulse, simulating a button press. Decoded outputs from the *4017* are buffered in the tri-state *4503* hex. buffer and used to drive the button fets.

Since power for the adaptor comes from the controller, decoupling must prevent current spikes from the transmitter led clocking the counter.

C Drinkwater
Wotton-under-Edge
Gloucestershire



Disabled person's tv controller adaptor, operated by microswitch to step through channels and standby.

Current-biased Class-B output stage

Since, unlike the old germanium types, silicon power transistors do not need low-impedance sources to take account of thermal leakage effects, it seems strange that modern designs still appear to provide the low impedance. Indeed, the use of voltage sources can give rise to trouble due to amplification of thermal effects. This design uses current sources.

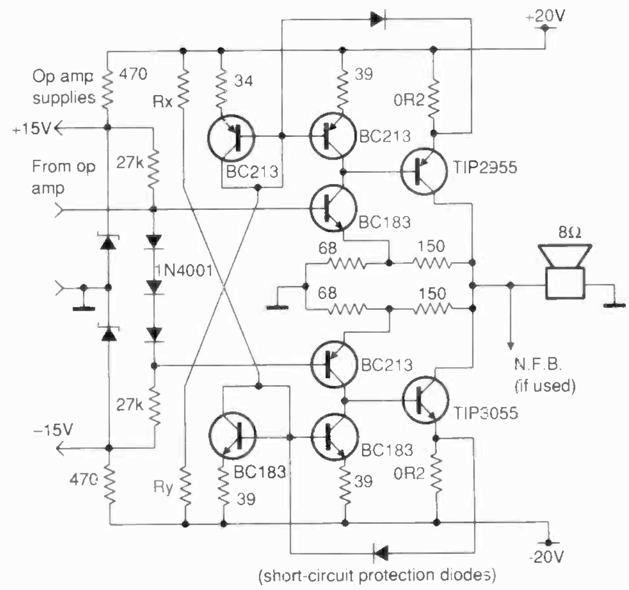
No bias adjustment is needed and R_x/R_y are chosen to suit the supply rails, in this case ±20V to give 20W into 8Ω. It is interesting that quiescent current increases as the power supply falls, reaching a maximum of 120mA at ±12V, which is small enough to avoid latch-up. This effect may eliminate crossover effects as output current peaks. Short-circuit protection is applied to the *BC213/183* current mirrors by means of the diodes.

The circuit possesses some voltage gain, so that an op-amp or even the output of a personal stereo will drive it to full power.

The type of Darlington output device with built-in bias resistance is unsuitable for this application.

Anon
Rotherham
South Yorkshire

Would this contributor please send his full name and address to our editorial offices? Ed.



Silicon audio power output stage, using current drive to the output transistors and needing no bias adjustment.

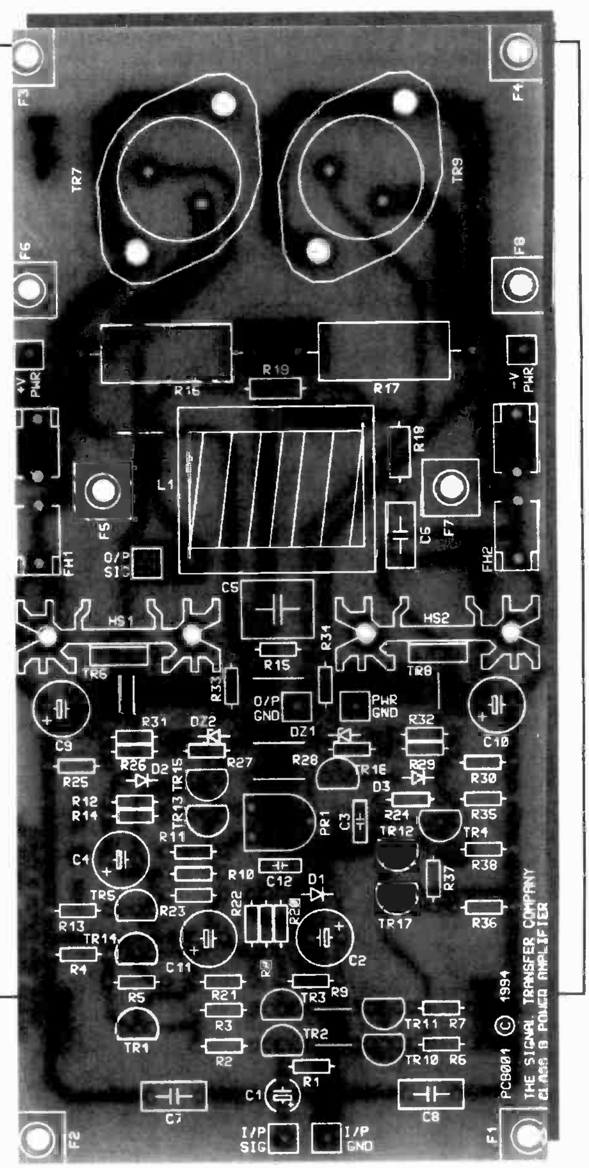
PCBs for Douglas Self's power amplifier series

Circuit boards for Douglas Self's high-performance power amplifier are now available via *EW+WW*.

Detailed on page 139 of the February issue, Douglas Self's state-of-the-art power amplifier is the culmination of ideas from one of the most detailed studies of power amplifier design ever published in a monthly magazine. Capable of delivering up to 100W into 8Ω, the amplifier features a distortion figure of 0.0015% at 50W and is designed around a new approach to feedback.

Designed by Douglas himself, the fibreglass boards have silk-screened component IDs and solder masking to minimise the possibility of shorts. Sold in pairs, the boards are supplied with additional detailed constructional notes.

Each board pair costs £45, which includes VAT and postage, UK and overseas. Credit card orders can be placed 24 hours on 081-652 8956. Alternatively, send a postal order or cheque made payable to Reed Business Publishing to EW+WW, The Quadrant, Sutton, Surrey SM2 5AS.



LETTERS

Critical response

I cannot agree with Graham Maynard's rather vague analysis of class A and B amplifiers (*Letters*, June). Apart from anything else, I don't care for the way he, at a critical point in the argument, turns "conceivably" into allegedly solid fact in the next sentence. Back emfs from reactive loads do not cause detectable intermodulation[1] and even if they did, a higher feedback factor would surely reduce rather than increase the effect. As far as I know, no-one has ever argued that nfb does not effectively reduce amplifier output impedance. To go on and state as a another fact that higher levels of nfb cause "audible confusion" (whatever that is, presumably some sort of distortion) is simply wrong.

Using nfb to correct an amplifier with second-harmonic distortion creates third and higher-order products that did not exist before. Baxandall[2] brilliantly demonstrated this, showing that these higher products have a maximum value for very low amounts of nfb (say 2-10dB) and that as nfb is increased to more practical levels (like 30dB) the level of all the higher products decreases. The result is surely a powerful argument against the use of low nfb factors.

As for the sliding-bias scheme, I can hardly comment as Maynard's diagram seems to have been misdrawn: Tr_8, the driver for Tr_9, has disappeared, so the drive to Tr_9 is now in the wrong phase. Worse still, Tr_9 is turned hard on via the two diodes and the only result of building the circuit as shown will be a loud bang as two power transistors lay down their lives to protect the ht fuses.

Turning to Marcel van de Gevel's comments (*Letters*, May) concerning control of quiescent currents, I agree completely. But I would mildly point out that I specifically mentioned "harmonic AB", which gives the long, smooth device-to-device transition he describes, in Part 5 of the series. I also gave a reference to the intriguing work of F A Thus, and I do have well-thumbed copies of all the other references he quotes. The trouble with these schemes as they currently exist is that they depend on IC techniques such as matched and collector-area-scaled bjts. A discrete component version would no doubt be possible, but it would need to be designed from the ground up and would represent a serious piece of research work – not something that can be dealt with as an adjunct to a larger article. Anyone care to have a go? (An approximation to harmonic AB can be achieved by eliminating the usual output emitter-resistors; crossover gain variations are less, but still exist, and there is a clear and present danger of thermal runaway.)

I am sorry that Chris Bulman (*Letters*, May) finds my class-A quiescent regulator inelegantly intrusive; surely it can't be too bad, having only three transistors, and not even being in the direct signal path? At least it is effective.

As far as I can determine from re-reading the 1969 article, John Linsley Hood's psu is essentially constant-voltage, though with that voltage rather uncertainly determined by transistor ß, as indeed is the amplifier quiescent current (I_q). If this psu really does play an important role in protecting against thermal runaway in the amplifier then I would suggest that the amplifier design could possibly be improved. It hardly counts as an effective method of I_q control.

Since the matter has been aired, I may as well report my own experiences with this design. With an I_q of 1.25A, the thd at 10W/8Ω and 1kHz was 0.25%, which I found rather discouraging. The output transistors are not driven so as to give pure push-pull action, and the sum of output-device currents (which in true push-pull is virtually constant over a cycle) varies by more than 40% (see *PSpice* plot).

The logic behind dc-coupled amplifiers is not obscure. Reservoir capacitors are simply storage elements, and have no direct effect on the amplifier output. That is (or should be) in the iron grip of a hefty negative feedback factor, and immune to ht rail influence. In contrast, an output capacitor handles the signal after the nfb has controlled it, and so any capacitor-distortion it may generate at lf is passed on untouched to the loudspeakers.

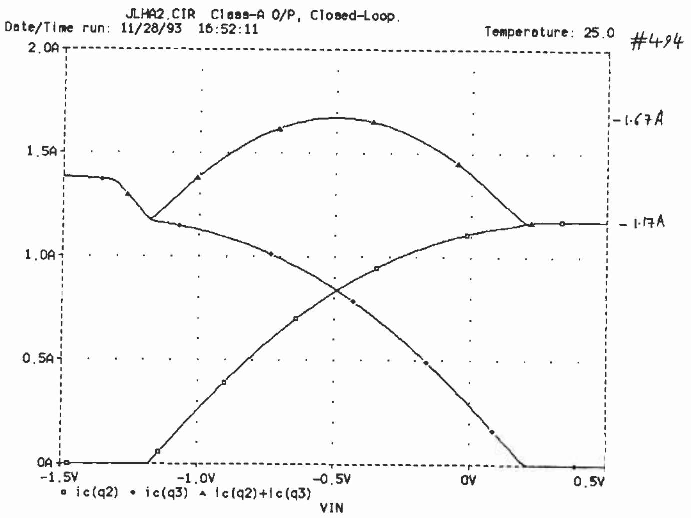
Any dc-coupled amp must have a dependable offset-protection system.

I can only admire the breath-taking audacity of Mr Bulman's scheme for cooling output transistors by making them boil water. One might object on the pedestrian grounds that the junction-case thermal resistance of even the heftiest TO3 transistors is a 0.7°C/W and so observing a 200°C junction temp maximum reduces the permissible power dissipation from 250 to 70W. I might also point out that elevated temperatures degrade reliability – however this is to detract from the grand sweep of the concept.

Since the transistor cases appear to be in actual contact with the water, there are presumably four electrically-isolated boiling tanks, two for each channel, each with their own condensers and circulating pumps. The arrangement will make for an impressive-looking piece of equipment, reminiscent of the engine-room of a medium-sized steam-tug. Deionised water could be used to minimise conduction – but this is not the ideal basis for a good cup of tea. I cannot resist speculating on the software used to test the system. Handel's Water Music is an obvious choice; tapes of steam radio might be even more appropriate.

Douglas Self
Forest Gate
London

PSpice plots of output device currents for the complete closed-loop amplifier. The plots in Part 8 of Self's *Distortion in Power Amplifiers* series were all done open-loop.



References

1. Cordell, R. *Interface Intermodulation in Amplifiers, Wireless World*, Feb 1983, p.30.

2. Baxandall, P. *Audio Power Amplifier Design: Part 5, Wireless World*, Dec 1978, p.56.

Beach buff

Alan Dyke's criticism of audio mysticism makes sense – except for his reference to sand-filled speaker stands, of which I have just bought a pair.

Producing stands with hollow vertical supports is presumably cheaper than using solid stock which would be better and heavier. So filling them with something to facilitate coupling the base of the cabinet to floor, in view of the powerful magnet and coil action of the main drivers, makes sense.

Liquid, especially water, would actually transmit sound vibrations faster than in air. Sand is freely available at low cost – especially if you are near a beach as I am.

Spikes through the carpet pile to floor complete the set-up, and are necessary when using the new D Self amp, of which mine must be one of the first examples.

Hugh Haines
Sunderland

Small world

H G Groenevelt seems not to appreciate the difference between a theory and a theorem. There is all the difference in the world. Maybe that accounts for his confusion.

R H Pearson
Lincs

Amnesty appeal

Mu'taz Qutlabi is an electrical engineer with Aleppo Electricity Company. He is 38, married with one child, and this July sees his 12th year of imprisonment as a Syrian Prisoner of Conscience.

Mu'taz, detained in Damascus, is just one of thousands of political prisoners held in Syria without charge or trial, for their peacefully held beliefs. Many are detained under emergency legislation in force since 1963, suspending all constitutional protection against abuse by the state. Others have been held for more than 20 years, face torture – routine in Syria – and are denied the chance to challenge the legality of their detention.

Mu'taz is being held at Fir'Amn al-Dawla prison, and Amnesty International is calling upon the Syrian authorities to release him immediately. Every letter of support helps. Please write saying you have read about his case. Courteous letters in English should be sent to: His Excellency, President Hafez al Assad, Presidential Palace, Abu Rummaneh, Al-rashid Street, Damascus, Syrian Arab Republic.

Paul Horvath
Rochdale Amnesty
Lancs

Hidden meaning

Unfortunately a short paragraph is missing in my article *Optoelectronics by design* (*EW + WW*, May 1994, pp. 364-369). The box titled 'Units and meanings' should conclude with two equations relating to the presented drawings (*Fig. d*, middle and bottom).

The equations regarding proximity sensors should read:

$$U_r = S.R.G.I_e[\tan(\omega)/d]^2 \qquad ...(4)$$

where ω is the transmitter diode half power angle and d is the distance to a large object, and,

$$U_r = S.R.G.I_e[r/d^2]^2 \qquad ...(5)$$

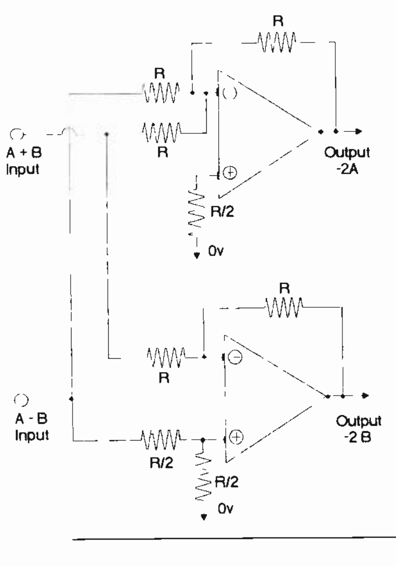
where r is the radius of a small object at distance d.

The final part of the article refers to a calculation based on equation (4).

Tore Arne Nielsen
Denmark

Graphic example

In my article on Arnold Sugden (*EW+WW*, pp. 486-487) a small error appeared in the published version of the matrix. Please would you publish a correction so my colleagues at Keele will continue to talk to me?



Incidentally, as one who is still climbing the steep learning curve of *Windows* after no problems whatsoever with the delightfully simple gui of the Atari, why on earth didn't Bill Gates produce a fully integrated dos with gui? Now, of course, he cannot; it would kill *Windows* stone dead.

Reg Williamson
Staffs

Gates didn't, but another company did. It's called the Amiga... Ed.

IC – can you?

Can any *EW+WW* reader help me find a superseded integrated circuit?

I have a Rohde & Schwarz test unit called an smpu, manufactured about 17-18 years ago. It is a radio-set test assembly for testing mobile radio transceivers up to 500MHz. I use it for testing radios provided by my club for communications at motor sport events.

No payment is received for the work which is really part of my hobby. So I can not justify spending large amounts of money on repair. The unit is a fine piece of test equipment and most components are bog-standard garden-variety. Unfortunately, the microprocessor has died. It is an Intel *4004*, a very early four-bit micro which does not appear to be available at traditional Australian component suppliers.

Rohde & Schwarz can supply the device – for A$450! While I can understand the need to charge this amount to cover the cost of storing this chip for 20 years, I still can not afford it.

What I am hoping for is that somewhere out there is a hoarder like me, who has a box full of Intel *4004s* just gathering dust.

If any one can help I would be most appreciative.

David Ireland,
77 Vahland Avenue
Riverton
Western Australia, 6148.
Tel: 61 9 457 2733
Fax: 61 9 326 6669

Decoding US actions

I found Robert Schifreen's article on the DES data encryption standard (*Big Brother's protection racket*, *EW+WW*, May, p.388) very interesting. Back in October 1998 I recall an *IEEE Review* claiming that the US National Security Agency was withdrawing its support for the NBS Data Encryption Standard – introduced in 1977 because: "The tools of attack are that much stronger now". It makes the proposed US legislation to outlaw DES quite puzzling.

TJ Wynn
Newport

Calling cable suppliers...

I cannot help but think that there is rather more involved than logic and mere engineering principles when discussing sound reproduction.

My hearing is bad, cutting off at around 8kHz, and I am by no means musical so I will leave the esoterics of subjective testing to those who believe themselves so qualified. But I would suggest that those really interested should perform proper double-blind tests on a reasonably large number of people, as psychologists do. Whether imperfect electrolytics and other components cause audile distortion or not seems a moot point.

Over the last 25 years I have designed something like 3000 separate items: from a 25kA rectifier to a vhf synthesised transceiver, and including a few audio systems. When involved with the audio projects, I remember looking at response curves of the loudspeakers, as published by manufacturers. All those that I saw resembled a cross-section of the Himalayas.

While I well appreciate the difference between amplitude variations and (harmonic) distortion and non-linearity, I cannot really see the point in aiming for an amplifier chain THD of 0.0001% with existing transducers. Perhaps someone could comment on the linearity, or lack thereof, of a coil moving in the magnetic field of a loudspeaker?

My guess is that the flux density is not totally uniform across the entire gap: so the deflection of the coil will not be a perfect reflection of the drive current. This of course will lead to distortion, most likely a reduction in amplitude of the peak of the wave (assuming the flux decreases towards the edge of the gap).

I have no idea what happens to the cone of the speaker when driven by complex waveforms, but would imagine that it would make an interesting study. What I would like to know is the figure for transducer linearity, both at the microphone and loudspeaker ends of the chain.

With these figures, one could then determine what level of amplifier distortion was acceptably small, and what was ridiculous.

As electronic design is a creative art, there is always the drive to achieve the ultimate, the perfect, or in the case of audio circuits, zero distortion. But, as all of those involved in design know, one must maintain a sense of reality, and know when it is time to shoot the engineer.

Perhaps an *EW+WW* reader familiar with loudspeaker and microphone performance could respond with some hard facts concerning transducer linearity?

Until such time, I don t think I'll be buying any silver loudspeaker cables.

Charles Frizell
Harare

...to prove a difference

Gold-plated speaker cables, silver-plated, platinum-plated etc... isn't it apparent that we professional engineers are starting to behave like amateurs. Neither the Yes nor No fraternity can produce irrefutable evidence that their beliefs are correct. I say 'beliefs' because without proof, that is what they must remain.

The continuing dialogue echoes the heated discussions that took place, many years ago, when the particular attributes of moving-iron and balanced-armature loudspeakers were compared. But that was resolved with the introduction of the moving-coil loudspeaker which quickly killed off the two warring factions.

There might be individuals who honestly believe they can detect a difference when special cables are used. But there are also far more audiophiles who declare that claim to be a load of rubbish.

I accept that 'might' does not necessarily mean 'right' any more than it did for the flat earth proponents of years ago. But until we have proof one way or the other, we are all just flogging a dead horse.

Logic tells us there should be no difference. Perhaps our logic is wrong, although I thought we were involved in a science more exact than most. Ramblings on the subject suggest we are actually a bunch of dabblers.

We should drop the perhapses and maybes and shut-up until someone produces proof either way. Could a supplier help by publishing details of test procedures? Come-on all you special cable manufacturers out there. Tell us the secret of your product. Or did I hear you say you do not have patent cover, or a registered design, or have not been granted a Kitemark?

RL Tufft
Thirsk

USING RF TRANSISTORS

Combined efforts bring power pay-offs

Splitting a signal to provide multiple inputs for several amplifier modules then recombining the result is an obvious way of boosting power output. Norm Dye and Helge Granberg describe different possible combiner/splitter configurations and show why they all have a place in the designer's portfolio.
***From the book* RF Transistors: principles and practical applications.**

When the required power output exceeds the capabilities of a single power amplifier, multiple stages or "modules" can be combined to produce the required result.

A splitter – simply a lower-powered version of a combiner used in reverse – divides the input signal into multiple equal amplitude outputs. These are applied to the inputs of each module. The power combiner then recombines the module outputs ready for feeding into a single load.

Combiners are closely related to wide-band transformers in design and construction – the main difference is how the lines or windings are connected (power splitters have the same configuration, so need not be treated separately).

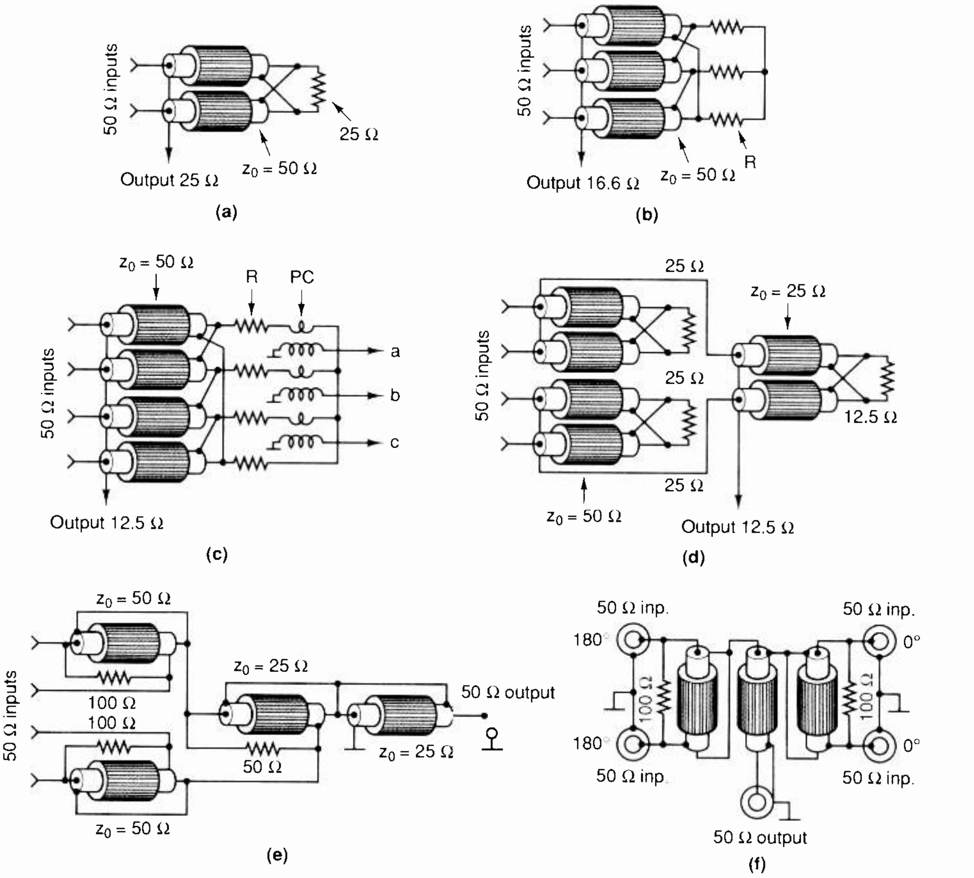


Fig. 1. a), b) and c) represent straight in-phase combiners. Even and odd numbers of input ports are possible with them. d) and e) are staggered or "totem pole" structures, which are adaptable only to even numbers of ports. f) is a 180° combiner that requires no step-up transformer into 50Ω. Note that step-up transformers are not shown for a), b), c) and d).

Wideband power combiners need to have low insertion loss over the required bandwidth; provide isolation (minimum coupling) between the input ports, and show low return loss at the input ports over the required bandwidth.

Their operating bandwidth should be wide too – or wider than the amplifier modules or they will restrict the overall bandwidth of the combined amplifiers.

Several different types of power combiners are available, each having advantages regarding frequency spectrum, bandwidth and other preferred features.

Four main combiner types

Zero degree device: ideally there will be no phase shift between the input ports and the combined output. The device can be designed for even or odd number inputs up to a practical limit. Practical frequency range is up to about 500MHz. These combiners are most common at lower frequencies because of their versatility and straight forward design.

180° device: the two input ports or sets of input ports are 180° out of phase. The device is only applicable to even numbers of inputs: eg 2, 4, etc. Out-of-phase characteristic must be taken into account when designing the input power splitter. Practical frequency range is up to about 100MHz.

90° hybrid: two port unit having, by nature, a narrower bandwidth than the above configurations. The device is applicable from low frequencies to microwaves in proper design configurations.

"*Wilkinson combiner*": relatively narrow bandwidth characteristics, but simple and inexpensive. The practical frequency range can be up to microwaves (1-2GHz).

The input power-splitter must be of the same type as the output combiner. An in-phase splitter used with a 180° combiner for example would produce zero combined output power, since the outputs from each amplifier module would cancel.

In-phase and 180° combiners

Transmission line techniques are commonly used in in-phase and 180° power combiners to produce lowest losses and widest bandwidths.

One primary function of an rf power combiner is to provide port-to-port isolation – 30dB is typical and acceptable. By this, the output of one amplifier module will be sufficiently isolated from the others. If a failure in one occurs, the remaining amplifiers will not be affected and will still be operating into the original load impedance. Amplitude unbalance is usually created by a completely disabled module (the "source" for the combiner) or, sometimes, more than one disabled source. Power output with various numbers of disabled sources can be calculated as:

$$P_{out} = (P/N)N_1$$

where P is the total power of operative sources, N is the total number of initial sources and N_1 is the number of operative sources.

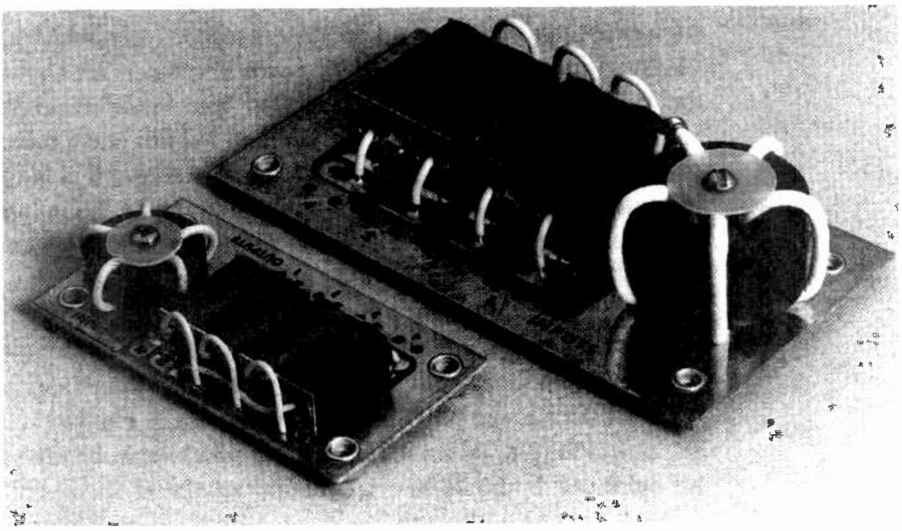


Fig. 2. Commercial four-port splitter-combiner for use at 1.6-20MHz and power levels up to 1kW. The device is of the type shown in Fig. 1c but the combiner lacks the power pick-up coils. The 50Ω lines are realised by metal tubes with a proper size of insulated conductor threaded through the tube, reducing construction costs. Note the compactness of the units, where the step-up transformer is right next to the combiner structure.

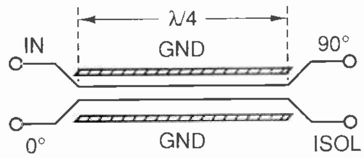
For example, in a four-port system designed to deliver 1kW with 250W modules, with one module disabled power output would be (750/4)3 = 562W. The difference power of 188W would be dissipated in the balancing resistors, dividing according to the type of combiner. In a straight zero-degree combiner (**Figs. 1a**, **1b**, and **1c**) all resistors are of equal value and the power dissipated in each would be (188/3) = 62.5W in a four port system.

In a two-port combiner system having a maximum output power P_{max} if one amplifier fails the output power will decrease to a value 6dB below P_{max}. Half the power loss is due to lack of power from the disabled module and an additional 3dB is lost because the power from the remaining module now divides equally between the balancing resistor and the output load.

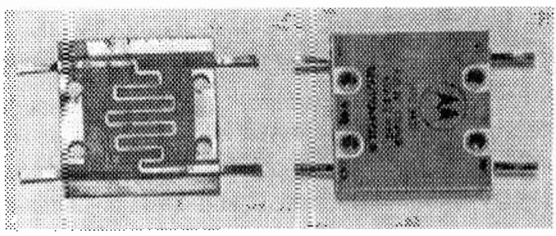
In this case the balancing resistor must dissipate 1/4 of the original P_{max}.

Values of the balancing resistors depend on the number of combiner ports and how many ports are assumed disabled at one time (as can be seen in the expression for P_{out} given above). Sometimes these resistors, which must be of the non-inductive type, are referred as dump loads since power due to phase- or amplitude-unbalance is directed to them. In most cases even if one module fails, the system is forced into a shut-down mode. So the balancing resistors do not have to dissipate significant power.

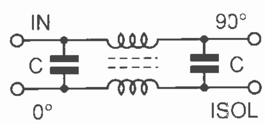
In the failure detection method using pick-up coils (Fig. 1c) the signal pick-up coils (pc) can be small toroids wound with multiple turns of wire. These can form the secondaries for rf voltage step-up transformers whose primaries are the leads of the balancing resistors



(a)



(a)



(c)

Fig. 3. The line hybrid, a) and b), is widely used at vhf and uhf since it can be made compact by folding the lines. Its lumped constant equivalent is shown in 3c. The complete uhf coupler has dimensions of approximately 38mm square and 3.5mm thick.

(eg, two-way carbon type) threaded through the toroids.

RF voltages in the secondaries – generated by the unbalance due to a module failure – can then be rectified and processed to operate the shut-down circuitry. If the *a*, *b* and *c* outputs (Fig. 1c) are kept separated, each one can be made to operate an indicator, showing which module has failed. Resistor (R) values are not critical, since the power will be shut-off within a millisecond or so making the load mismatch unimportant.

But, for operation under reduced power conditions, the balancing resistors must handle continuous power levels. Here high-power resistors must be used that can be heat sunk, and that are "floating", with the resistor ele-

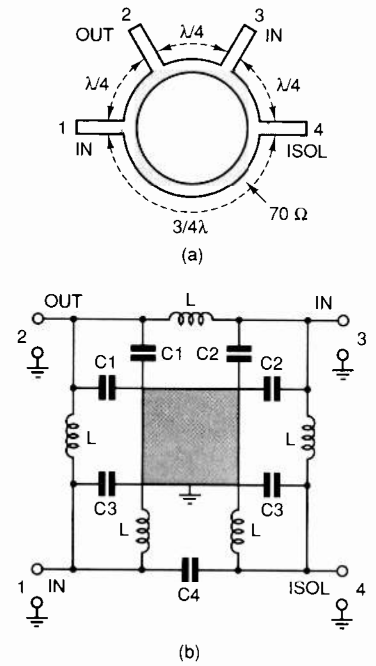


Fig. 4. Ring hybrid (a) and its lumped equivalent (b). In b), C1 and C2 can be paralleled into single units twice the value.

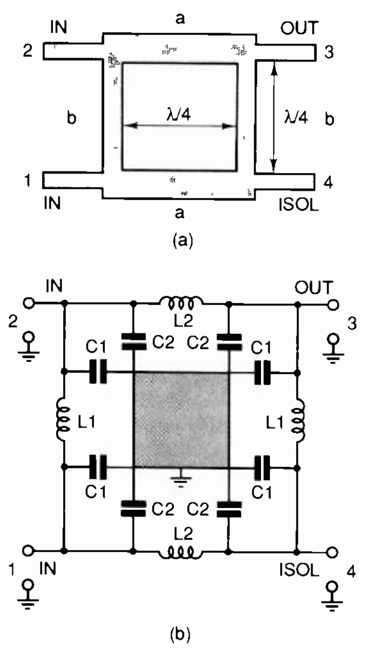


Fig. 5. The branch line hybrid (a) and its lumped equivalent (b). As in the ring hybrid equivalent, the paralleled capacitors (C1 and C2) can be combined into single units.

ment electrically isolated from its mounting structure.

In the event of a failure of one module (which is the most common case), the balancing resistor values can be determined from the formula:

$R = Z_{in}/N$

where N is the number of input ports. Values of 25, 16.6, and 12.5Ω result for two, three and four-port combiners (Figs. 1a, 1b and 1c).

Examples of 180° power combiners include a "totem pole" structure (**Fig. 1e**) and a system consisting of a pair of two port hybrids and a balun (**Fig. 1f**).

Comparing the 180° and in-phase combiners shows that the 180° type can be superior in input vswr, but that the latter has better port-to-port isolation characteristics.

In cascaded or totem pole structures, the balancing resistor values follow the values of the two-port hybrids and transmission line transformers.

It is possible to design a system that does not require a step-up transformer (Fig. 1f). Its simplicity is attractive compared to other four-port combiners. The number of transmission lines can be halved and the lack of the 4:1 step-up transformer means bandwidth characteristics are enhanced – though this combiner is covered by a US patent.

Impedance ratios such as 2:1 and 3:1 can be implemented with a 4:1 transformer wound with coaxial cable where the corresponding taps are made to the coax braid. Improved performance over wider bandwidths is possible with fractional-integer equal-delay transformers.

In a commercially available four-port splitter-combiner (**Fig. 2**) intended for use at 1.6-30MHz and up to power levels of 1kW, port-to-port isolation figures of 27-40dB have been measured on both the in-phase and 180° combiners over a bandwidth of several octaves.

Several manufacturers make high power combiners for use at frequencies up to uhf, but most units are housed in enclosures with connectors and do not suit many designers because of physical outlines or cost.

90° hybrids

One class of hybrid that is a variation of the two-port is the quadrature, and quadrature-couplers can be realised in several different forms.

A quadrature combiner usually refers to a passive device with one input port, two output ports (or vice versa) separated by 90° phase, and an isolated port.

Some networks that can be considered as quadrature couplers are:

- Line hybrid (**Fig. 3**);
- 3λ/2 ring coupler (commonly known as rat-race, **Fig. 4**);
- branch coupler, widely used in small signal mixer circuitry but also applicable to use as a power combiner (**Fig. 5**);
- Wilkinson combiner, consisting of a series of λ/4 transmission lines (**Fig. 6**). Though not a quadrature coupler like those mentioned above, it is included because all four types are based on the principle of a delay generated by a quarter wavelength transmission line.

Line hybrids

The line hybrid (**Fig. 3a**), one of the most common combiners used in the vhf and uhf frequency ranges, consists of two transmission lines (microstrip) of λ/4 in length separated by a dielectric. In addition the structure is sandwiched between two ground planes, again separated by a dielectric.

The mutual impedance between the two lines is designated as Z_{even}, whereas the impedance from the lines to ground is designated as Z_{odd}. Z_{even} is the impedance controlling the coupling coefficient between the lines and is typically $Z_{in}/2$. Then Z_{odd} must be calculated for a value that gives:

$\sqrt{(Z_{even} \times Z_{odd})} = Z_{in}(50\Omega)$.

If $Z_{even} = 25\Omega$ as in this case, then Z_{odd} must be 100Ω for $Z_{in} = \sqrt{(2.5 \times 10^3)} = 50\Omega$. The Z_{even} and Z_{odd} values can be modified for greater isolation or extended bandwidth, but the $\sqrt{(Z_{even} \times Z_{odd})}$ relationship is always valid.

Typical bandwidths of these couplers are about 15% with 1:1.5 vswr and port-to-port isolation ranging from 20 to 30dB.

In the past, dielectric materials were Teflon-fibreglass (dielectric constant around 2.5) or epoxy-fibreglass (5). Low values of dielectric constants limited the couplers to a lowest practical frequency of approximately 175MHz, though even at these frequencies the couplers were bulky.

Today, advances in dielectrics have brought materials with dielectric constants of 10 or higher, making lower frequency couplers more practical.

To realise any kind of a practical size or shape factor, the lines are usually folded several times (**Fig. 3b**) and the total electrical length calculated for *l*/4 according to the dielectric constant of the medium.

A variation of the line combiner, designed with lumped constant elements for use at low frequencies, would behave much like its counterpart, the stripline quadrature hybrid. In a true representation of the stripline design, capacitors (*C* in *Fig. 3c*) should be split and their centre taps grounded to simulate Z_{odd}. Z_{even} would be determined by the mutual line impedance and the electrical length of the line, which should also be λ/4.

In practice, the transmission line can be made of twisted enamelled wires or two lengths of low impedance coaxial cable with their braids connected together and floating. The centre conductors are left to form two symmetrical lines.

Another adaptation, especially suited to low frequency use, is Fisher's hybrid (resembling more closely the hybrid shown in Fig. 3c). Very tight coupling between the lines is required, and line physical lengths need not be λ/4 since they are electrically lengthened by the presence of the magnetic medium. *C* depends on the line coupling coefficient and may not be necessary at all in some cases.

The phase relationship is the same as in the stripline hybrid and port isolations are comparable too. Bandwidth achievable depends on line inductance – and so properties of the magnetic core – as well as the line mutual capacitance (coupling). But using the tech-

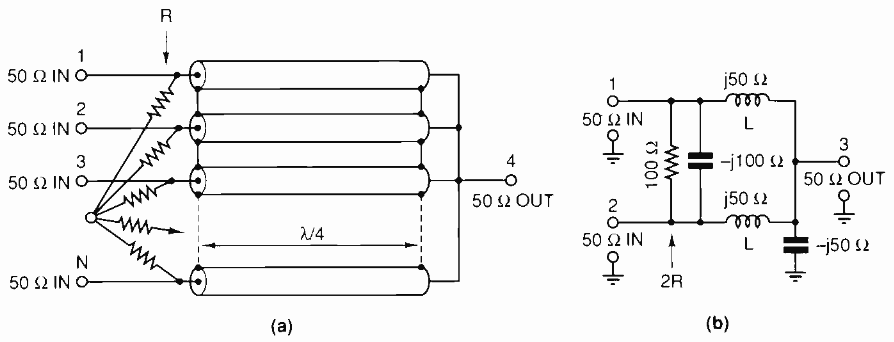


Fig. 6. Wilkinson combiner (a) and its lumped equivalent (b). In practice b) may not be feasible for more than two input ports.

niques described, it may be possible to develop quadrature hybrids for wider bandwidths than the 10-15% normally attributed to this type of coupler.

Ring hybrids

The hybrid ring, (3/2)λ hybrid or "rat race" (**Fig. 4**) as it is commonly called is a directional coupler that can be used to sample rf power travelling in different directions. As a result, it is also adaptable for mixer, power splitter, and combiner applications.

Construction is usually with microstrip transmission lines, a technique that limits use to about 1000MHz and above. Coaxial cable designs can extend the frequency range down to 200-300MHz, but such structures become bulky.

A simple hybrid ring consists of a transmission line in which input, output and isolation ports are connected in four places. If the Z_{in} impedances of these ports are 50Ω, characteristic impedance of the transmission-line ring is:

$$Z_0 = \sqrt{(Z_{in}^2\,N)}$$

where N is the number of active ports is 2, then,

$$Z_0 = \sqrt{(5 \times 10^3)} = 70.7\Omega$$

Each of the four ports is separated by quarter-wavelength sections making the remaining part of the ring three-quarter wavelengths – so the total circumference is 1.5 wavelengths.

The hybrid ring is commonly used as a power splitter or combiner. If a signal is applied to port 1 (**Fig. 4**a), power will be equally divided between ports 2 and 4 and their phase relationship will be 180°.

On the other hand, power incident at port 2 will also be equally divided between ports 1 and 3. But the two output signals will be of similar phase.

When port 2 is used as the input in a combiner, any power reflected at output port 3, due to a mismatch, arrives at the other output port 1 by two paths.

One signal travels a half-wavelength in a counter clockwise rotation from port 3, and the clockwise signal appears at port 1 delayed by a full wavelength. The half wave difference in arrival, and equal path loss results in cancellation of the two signals at port 4, with total cancellation resulting in highest port-to-port isolation. The reflected signal from any mismatch at port 3 arrives at port 4, in phase from both circular paths, where it is dissipated. This port is designated as the isolation port, where a termination absorbs any power due to the unbalance between output ports.

The input signal from port 2 cancels at port 4 because the two-way paths differ by a half wavelength. Advantages of a ring hybrid include simplicity of construction and a reasonable tolerance for variations in line impedance. In addition, the power at the output ports can be adjusted by varying the impedances of the interconnecting lines.

A simple hybrid ring can provide a good match and excellent isolation over a 8-10% bandwidth.

Ring hybrids can also be designed with lumped constant elements – extending low frequency response to audio frequencies.

Deriving a lumped-circuit equivalent is simple and entails replacing each length of transmission line by its pi equivalent (**Fig. 4b**). Three pi networks represent the $\lambda/4$ sections, namely C_3-L-C_1, C_1-L-C_2 and C_2-L-C_3. The lowest branch – L-C_4-L – forms the 3/2 λ section.

Since all capacitors will be of equal value, C_1 and C_2 can be paralleled into a single unit, $2 \times C_1$ and $2 \times C_2$ respectively. All branches must have a Z_0 of $Z_{in}\sqrt{2}$ or 70.7Ω as the line sections in the microstrip design. All inductors in the circuit are also of equal value, calculated as:

$$L = (Z_{in}\sqrt{2})/2\pi f_0 \text{ and } C = 1/(2\pi f_0 Z_{in}\sqrt{2})$$

For example, designing a lumped constant ring combiner for a frequency of 100MHz. Then:

$$L = 70.7/628 = 0.11\mu H \text{ and}$$

$$C = 1/(628 \times 70.7) = 22.5pF.$$

Capacitors C_1s and C_2s combined would then be 45pF each. The isolation and insertion loss characteristics are greatly dependent on component tolerances, loss factors (Q), and symmetry of the total structure. Typical numbers for 100MHz are 20-25dB and 0.4-0.45dB respectively.

Testing splitters and combiners

Multiple port configurations make the performance of splitters and combiners difficult to test. But a possible method is to terminate all outputs except one and make the measurements between the one unterminated output and the single input port. Each output port would be sequentially tested, revealing the amount of insertion loss, return loss, and the phase angle over the desired frequency spectrum.

To test isolation characteristics between the output ports, the input and all output ports but two should be terminated. Isolation can then be measured between all output ports by sequentially switching terminations and active ports.

One quick test would be to connect all the multiple output ports together (back-to-back), leaving two open ports compatible with a standard 50Ω system – though this would not provide information on port-to-port isolation characteristics of the output.

These test methods are applicable to most types of splitter or combiner. Testing car be carried out with a network analyser or an rf test stat on consisting of a signal source, load. and appropriate means for measuring forward and reflected power. The rf test station is probably the more accurate of the two methods, since the subject devices can be driven with a more realistic level of rf. In either case, the amount of return loss, insertion loss and also, in the case of the network analyser, the phase relationship can be found

Phase error can be as much as 25°, but this would still create only a 0.22dB loss. Phase errors of around 10° – typical maximums – have negligible effect.

Influence of source amplitudes are also exaggerated. In a two port system, a 50% difference in source power outputs produces only an approximate 0.2dB loss in the combined output.

Figures may vary depending on the type of combiner, but we can conclude that, in practice, the effect of amplitude unbalance between the input ports of a combiner (especially with a failed source) is far more dominant than the phase relationship.

Branch line couplers

Branch line couplers are the easiest types to construct – most often as two-branch, 3dB units (**Fig. 5a**). The main transmission line is coupled to a branch line by two quarter-wave lines, each spaced a quarter-wave apart. Simple microstrip configurations can produce coupling values from 3 to 9dB. Used as a power combiner, the device acts as a quadrature coupler with output signals 90° out of phase.

Overall bandwidth of branch-line couplers is rather narrow compared to the hybrid rings. Bandwidth can be increased by adding more branches, though this is not popular because additional branches have higher impedances, and as well as increased losses, the microstrip tolerances become critical.

The use of transformer sections in the main lines results in characteristic impedances of the branch lines equal to those of the input and output arms. The impedance of these transformer sections (*a* in Fig. 5a), also called main lines, can be figured as:

$$Z_0 = \sqrt{(Z_{in} \times Z_{out}/2)}$$

For a 50Ω system:

$$Z_0 = \sqrt{(50 \times 50/2)} = 35.4\Omega.$$

Impedance of the branch lines (*b* in Fig. 5a) will be 50Ω too.

Both the hybrid ring and the branch line coupler are very versatile devices. With the ring coupler, impedance transformation is possible by varying the characteristic impedances of the $\lambda/4$ line sections.

Similarly the transformer or the main line impedance can be varied in a branch-line coupler, producing impedance transformation between the input and output ports. Though this function may only be necessary when devices are used for non-power-splitting combining applications such as mixers.

The branch-line coupler is physically smaller than the hybrid ring since its loop periphery measures only one full wavelength. But its practical low-frequency limit is in the low microwave region. Construction using coiled coaxial lines would bring the lowest practical frequency into the 150-200MHz region. Lower frequency versions can be realised with lumped element designs (**Fig. 5b**) where the principle is the same as with the hybrid ring: each $\lambda/4$ transmission line section is replaced by a πLC network. In the hybrid ring equivalent, there are three such networks. The branch-line equivalent requires four, two for 35.4Ω impedance and two for 50Ω. Using the formula to calculate component values for 100MHz, the 35.4Ω branches (*a*) will be:

$$L = 35.4/628 = 56\text{nH}$$

and:

$$C = 1/(628 \times 35.4) = 45\text{pF}.$$

Similarly the 50Ω branches (*b*) are $L = 80$nH and $C = 32$pF. The combined capacitances ($C_1 + C_2$) will be 45 + 32pF = 78pF. Component tolerances are critical.

Wilkinson couplers

One other power combiner that uses quarter wave transmission lines is the Wilkinson hybrid, though it should not really be called a hybrid since that describes a device with two input or output ports.

The Wilkinson coupler is a reciprocal network and sums *N* coherent sources to a common port, all in one step. Any number of ports can be designed for (**Fig. 6a**) and so it is commonly called an *N*-way coupler.

In a similar way to other types of combiners, the device can be used for power splitting as well as combining. But unlike quadrature hybrids, all inputs are in equal phase *relative to the output*, but create a delay of 90°. Since the line length between any two ports is $\lambda/2$, the power arriving at each port is 180° out of phase with the power from any other port and so cancels, providing isolation between inputs.

The port-to-port isolation and vswr are theoretically perfect at mid-band (where the lines provide a 90° phase shift) and degrade at frequencies away from band centre. In practice the amount of isolation and the vswr are primarily dependent on the phase relationship (line impedance and length) of the transmission lines.

Typical numbers for the isolation are 20-25dB and for the vswr, 1.2:1. The balancing resistors (*R*) have an important role in that they serve to help isolate and match the input ports. Normally they would not dissipate power, but any unbalance between input ports would result in power dissipation by them.

With two-port combiners, if the power to one input port is completely lost due to a failure of an amplifier module, then half of the power output from the remaining module will be dissipated in the balancing resistor. This is the case with all two-port combiners described. (Refer to straight in-phase combiners and Fig. 1c where a system shut-off under the condition of an amplifier failure is described).

Perfect output-port amplitude-balance (due to its symmetry) is a benefit of the Wilkinson coupler, with equal phases at all of its input ports.

Chief advantage for high power applications stems from the series of combined balancing resistors. Perfect isolation requires completely non-inductive resistors, which must be heatsunk.

Overall, the Wilkinson *N*-way combiner – constructed with coaxial, stripline, or microstrip transmission lines – offers a relatively low cost method of combining several signals with a fair amount of isolation and low vswr.

Bandwith of operation is relatively narrow, compared with 90° hybrids. For bandwidths wider than 15-20%, multiple lines of appropriate lengths can be cascaded, with stepped characteristic impedances for an optimum design with Chebyshev response. But the method will add to line *IR* losses, and may only be practical up to three-four sections. For *N* number of input ports, the line impedances are calculated as: $Z_0 = \sqrt{(Z_{in}{}^2 N)}$. So the Z_0 for a two-port system is 70.7Ω; for a three port system, 86.6Ω; for a four port system 100Ω, and so on. The balancing resistor values (which are equal) can be obtained as: $R = Z_0{}^2/NZ_{in}$.

Operation of a lumped equivalent of a two-port Wilkinson power combiner (**Fig. 6b**) is based on the same principle as the ring and branch-line coupler equivalents – that the $\lambda/4$ sections are simulated with π networks. The two 50Ω balancing resistors have been combined into a single 100Ω unit because the centre tap is "floating."

Similarly, the *C*s between the input ports have been combined to a $-j100\Omega$ reactance, since it is not necessary for their centre tap to be grounded. The reactances shown can be converted to required values of capacitance and inductance for the centre of the operating frequency band.

The lumped constant version of the Wilkinson combiner works well, with comparable isolation and vswr characteristics to its transmission line counterpart. But component tolerances must be kept within 1-2%. ■

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Measurement of rf power at the milliwatt level

It is sometimes necessary to measure rf power at the milliwatt level. This usually requires the use of expensive equipment and the achieved accuracy depends on the standard of calibration. In this note, Nick Wheeler describes an inexpensive thermal device which can be calibrated with dc and has good overload capability.

The Analog Devices *AD590* is a two-terminal current regulator with the property of passing a current which increases linearly at the rate of 1μA/°C, independent of supply voltage, over range of 4 to 30V.

The device is in a TO-52 case similar in size to a TO-18 miniature metal can. This case has a gold-plated base through which the two active leads pass, together with a third lead connected to the case.

The *AD590* is specified for storage up to 155°C and for 10s lead soldering temperature at 300°C. Typical low melting-point solder melts at 179°C.

It proved easy, and apparently non-destructive, to solder two 100Ω chip resistors to the gold-plated base. The gold plating facilitates this. These form a relatively non-inductive 50Ω load when connected in parallel, which is fed by length of RG-402 semi-rigid microwave coaxial cable.

The outer conductor of the coax is soldered to the case pin of the *AD590*. The two leads of the *AD590* need to be decoupled by chip capacitors as close in as possible, to prevent rf pickup.

At typical laboratory temperature, the device will draw about 295μA; the specification is 298.2μA at 25°C. On test, a sample drew 291.2μA at an ambient level of 20°C. This corresponds to a nominal 18.7°C. As the maximum calibration error is ±1°C and the measurement was at the 1% accuracy level, this was satisfactory. On applying one volt to the 50Ω load, the current increased by 4μA. A volt across 50Ω dissipates 20mW.

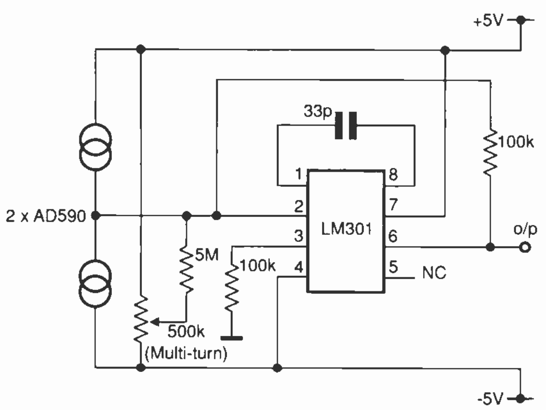
This simple form of the instrument measures small difference between two large quantities. This is bad practice.

By using two sensors in a differential circuit derived from the Analog Devices databook, and applying the signal to one of them, measurements may be made to the milliwatt level. The devices should be made as similar as possible and mounted in a draught-proof enclosure, thermally insulated from each other but otherwise identical in relation to the environment. I used foam polystyrene to achieve this.

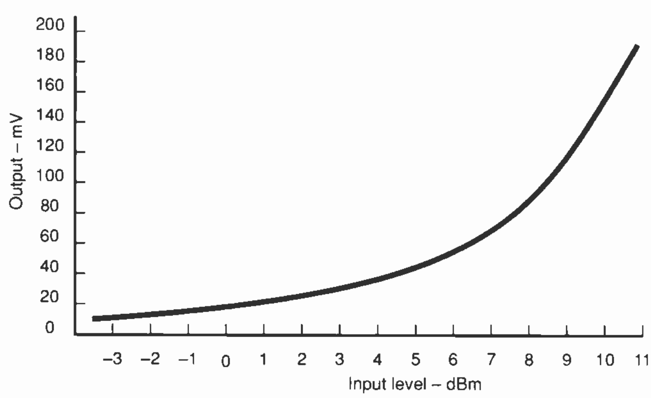
On transferring the signal input from one sensor to the other, a large output of reversed polarity should result. It is most unlikely to be identical. Both devices should be calibrated with dc at inputs up to 200mW. The resulting calibration curves are non-linear, falling off at higher inputs. This is because the heat loss from a heated object is proportional to the square of its temperature above ambient.

This is not a quick-response system. About 90% of the final reading is reached in a minute. Also note that a series of readings should start with the lowest input. Otherwise time must be allowed for cooling off between measurements.

Also note that when measuring rf, this is a true rms measurement. Measurements made at 50MHz using a good 100MHz oscilloscope to determine the applied voltage corresponded to the dc calibration within a few percent. ■



Apart from ensuring that the two sensors are mounted in a similar thermal environment, attention should be paid to decoupling of the leadouts with chip capacitors to avoid measurement rf rectifying in the junctions.



RMS rf power meter – typical calibration curve for differential circuit.

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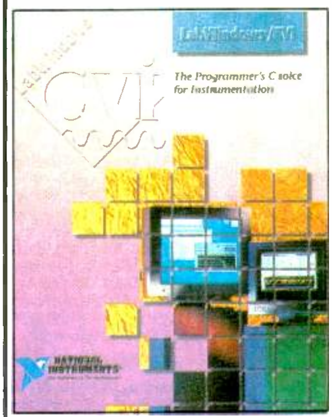

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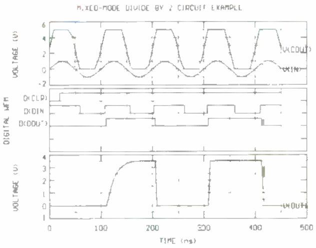

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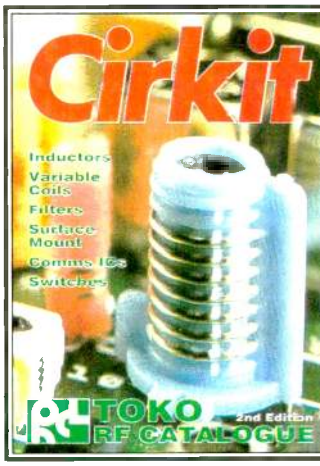

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